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Items are listed according to the monthly column in which they appeared and/or under any category(s) that generally describes them. Examples: "The Boom-Excited Antenna" is listed under Miscellaneous Technical and Antennas. "Reciprocal-Operating Countries" is listed under Happenings, Regulations and in the Mini Directory Index at the end of this index.

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OUR COVER

A group of vhf'ers gathers annually at W2SZ/1 atop Mt. Greylock in Western Massachusetts for the ARRL VHF contest (see p. 82). WA2AAU makes the final adjustments to equipment on the 1296/2304-MHz dish. *(June 1981 photo)*

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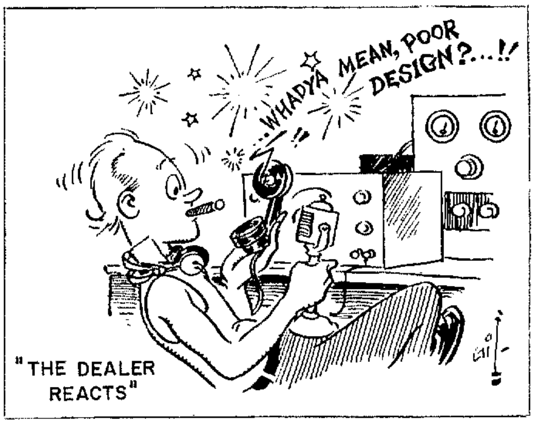
Modern Receivers and Transceivers: What Ails Them?



As the technology advances, some performance characteristics improve, while others worsen. Assorted ailments result from poor design and fundamental limitations; others are products of economic measures and modern technology.

By Doug DeMaw,* W1FB and Wes Hayward,** W7ZOI

How does your new super hf transceiver stack up against previous equipment you've owned? Chances are you're proud of that physically small, sophisticated and often costly transceiver that serves as the center piece of your operating desk. But, what is there about it that doesn't compare with some of the ham gear you bought or built earlier? Most users could prepare a list of faults on a moment's notice.

This paper will examine some of the common performance shortfalls that are peculiar to much of today's amateur equipment. Our main emphasis will be on receivers and the receiver portions of hf transceivers. Some comments will also be made regarding transmitter performance.

The Good Features

Manufacturers of amateur equipment have made noteworthy advances in circuit design and operating features in the past decade. Transmitter spectral purity (except for noise) has improved generally with each new model released. Harmonics and other coherent spurious emissions are typically very low with virtually all models of current equipment.

Receiver performance is similarly improved. Two-tone dynamic range has increased gradually, and for most equipment it is adequate for amateur needs.[1,2,3] Values in excess of 90 dB are common for both cw and ssb bandwidths. Oscillator stability and frequency-readout resolution have taken ham gear out of the Model A car class. This refinement appears not only in traditional designs with mechanically tuned oscillators and dial mechanisms, but in the later designs based upon digital technology. The latter include equipment with built-in counters and frequency synthesizers. Filter technology has advanced, offering significantly improved receiver selectivity. Modern designs provide more than adequate sensitivity while not severely compromising other performance factors. Image rejection and i-f rejection are also quite good, especially in those designs using a first i-f at vhf.

These fundamental improvements have been accompanied by operator-convenience features. Many of these are quite worthwhile. They include receiver incremental tuning — RIT (in transceivers), frequency memories, the ability for computer interface, various types of band-pass tuning and continuously variable bandwidth schemes, plus self-contained, tunable peak or notch filters. But, what have we traded off to obtain all of these marvelous improvements?

Some Not-So-Good Trends

Some lost performance may be attributed to economic considerations. Other losses result from fundamental limitations; the demands placed on equipment in today's crowded bands are severe. Another reason for degraded performance can be traced to design engineers unfamiliar with the requirements of amateur operators: Some designers aren't active amateurs and are not familiar with operating reality. One U.S. manufacturer of "ham" equipment refused for many years to include RIT in its transceivers. The rationale was that "RIT is used only by those manufacturers who produced rigs with unstable local oscillators"!

In the following paragraphs, we outline some of the problems we have observed. We will attempt to differentiate between fundamental items and those that result from economic considerations. Both reasons may apply.

Receiver Audio Quality

The receivers of the vacuum-tube era were typified by an abundance of available audio output. Things have changed, but not for the better. Many present receivers lack sufficient audio output power .This is most evident during mobile operation (owing to high ambient noise levels) — a paradox, for the state of the high-fidelity audio art permits undistorted audio output at levels up to 100 W (Fig. 1)! The problems are easily solved during design. Manufacturers could improve overall performance through the use of more efficient

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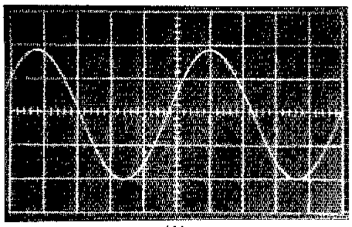
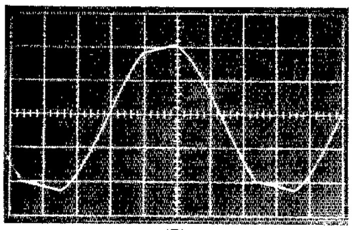
[1]Notes appear on page 16.



(A)



(B)

Fig. 1 — Oscillographs of (A) an undistorted sine wave at the output of a receiver audio system and (B) a distorted waveform resulting from inadequate audio-output power to overcome the ambient noise during mobile operation. Distorted audio is difficult to read, especially when automobile and wind noises are present.

speakers. Higher supply voltages, beyond the usual 12-V systems, would allow greater undistorted output power. Finally, the usual amplifiers, with one end of the speaker grounded, could be replaced with a pair of amplifiers (a "stereo" IC) that drives the speaker differentially, affording a 6-dB improvement in output power.

Attention needs to be paid to the frequency response of receiver audio systems. A number of receivers (new and old) have more or less flat responses out to 10 kHz. Communications information rarely contains components beyond 3 kHz. Receiver audio amplifiers should be shaped accordingly. The shaping should be changed as the i-f bandwidth is altered (switching from ssb to cw).

Agc Systems

A deficiency in many receivers is poor agc. There was a time in Amateur Radio when agc was used only during a-m phone operation. The agc was defeated for cw and, later, for ssb reception. Present operating practices are different: Most of us use agc during the reception of any mode. The agc system should be suitable for any condition that might be encountered.

The greatest agc failing is the attack characteristic. A loud signal, perhaps a cw station located within the first skip zone for a given band, appears within the i-f passband. An agc voltage is detected and *starts* to reduce the receiver gain. However, unless the bandwidth of the detection system is large with respect to that of the i-f, and detection occurs quickly, there will be a loud thump or pop. This is an annoyance to the operator, and it is fatiguing.

Designing a suitable agc response is not a simple matter of adjusting some R-C time constants. Rather, it is a system design problem, a detail that must be accounted for during the overall receiver design. A fast-attack agc is realized only if elements in the system that introduce a time delay are eliminated between the point where agc is applied (usually the i-f amplifiers) and the point of detection. The most common delay element encountered is a crystal or mechanical filter. Another potential delay element is the audio-amplifier chain. This precludes the use of audio-derived agc in a receiver for other than casual applications.

In spite of the difficulties, it is possible to design a suitable agc system. Such a system, when observed with an oscilloscope or with a critical "calibrated" ear, shows that the initial leading edge of a recovered audio signal has the same amplitude as that found with a steady tone. Some present equipment exhibits good agc performance. There are alarming exceptions, though! There are at least two agc-deficient commercial transceivers that DeMaw has tested. Both employ audio-derived agc (design rationale unknown) for which the attack times are faulty. Although the extremely loud pop in the receiver audio is barely tolerable when using a speaker, it is unbearable when using headphones during cw reception. Owing to a fairly short decay time, the agc can take hold many times during a single transmission by the station being worked, and the slower the cw is being sent the more times the pop will be heard in the phones.

Recovery time is the other critical parameter of an agc system. This is easily controlled by a suitable R-C time constant. The system should, however, be one that allows for operator control over the recovery time (fast or slow) and one that has a recovery time that is substantially independent of the strength of the incoming signal.[4]

Agc is an important part of present-day receiver technology, and is a preferred feature among most operators. The dynamic characteristics are important. There is no reason why the agc should not function smoothly and without popping, clicking or pumping.

A related deficiency in many receivers is the calibration of the S meter. It is not uncommon to find high-priced transceivers that show a difference between a displayed S1 and S9 signal of only 10 dB. The meter readings then take on greater meaning for stronger signals, although the calibration is rarely accurate. This deficiency could be eliminated through more careful design of the gain-controlled stages.[5] It may not be important for the manufacturers to develop a communications receiver that is suitable for making signal measurements. On the other hand, we suspect that a manufacturer who markets equipment with a calibrated S meter might be able to enhance sales with this feature.

I-F Filtering

An unfortunate fact is that some commercial receivers and transceivers contain essentially good cw and ssb filters, but leakage around the filters prevents them from providing the narrow response for which they were designed. There are other cases in which the filters themselves are deficient. While they may have excellent shape factors when the 6- and 60-dB down points are examined, the filters cease to provide much filtering action at the 80- or 100-dB points or more. Considering the state of front-end design, where it is possible to aim for essentially spurious-free responses over a 100-dB dynamic range or greater, it makes little sense to degrade the overall receiver response with poorly designed i-f systems. Some filter responses are shown in Fig. 2.

Poor stopband attenuation is observed readily in many receivers. The desired response is a dominant audio output of the proper pitch when a strong signal is encountered, with a total absence of signal-related output when the receiver is tuned away from the signal. It is common to hear a strong response for 10 kHz on either side of the signal frequency. A gradual decrease in strength as one tunes into the skirts of the filter is, of course, acceptable. The presence of a response completely outside the filter passband response is not!

A good ssb filter, alone and not in an i-f system, can have a stopband attenuation

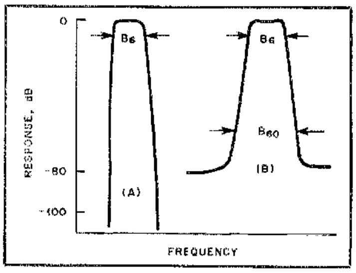


Fig. 2 — Passband responses for two crystal filters. That at A is a more-or-less ideal response. The skirts are very steep, and the filtering action continues for attenuations in excess of 100 dB. The filter at B is more typical of the performance found in a receiver i-f amplifier. While the skirt response is suitable with a good 6- to 60-dB shape factor, the stopband attenuation is limited to about 80 dB.

of 100 dB or greater. This is possible with relatively simple 8-pole designs. Even a 5-crystal cw filter with a bandwidth of 300 to 500 Hz is capable of more than 100 dB of stopband attenuation if the filter is built with care. These figures are probably acceptable if actually achieved in a receiver. This does not always occur. Actual filter characteristics are compromised through poor shielding of the i-f system and ground loops, and by inferior filter-switching circuits.

One solution is to use a multiplicity of crystal filters. This is more common in commercial equipment, although the intent of such a design usually has been for other reasons, the most common being a passband-tuning feature. A cw receiver we built has a 9-MHz i-f system using multiple crystal filters, each with a cw bandwidth.[6] This system exhibits a stopband attenuation in excess 130 dB. Listening to such a receiver can be a unique experience.

There are many virtues to designs that contain cascaded filters. It is easy to achieve 120 dB of stopband attenuation with two isolated filters, even if the individual filter characteristics are less than ideal. Moreover, the use of extra filters in an i-f amplifier will ensure that minimal i-f generated noise reaches the detector. Care must be exercised in such a design, for it can complicate the agc system by introducing the time delays described earlier.

Another problem with some receivers, related to the i-f filters, is the time-domain response. The ideal filter shape for communications is usually thought to be one with a flat top, with perhaps a slight ripple, and extremely steep skirt response, much like that shown in Fig. 2A. A practical approximation to this shape is a Chebyshev filter response.[7]

While the frequency response of a Chebyshev filter is close to ideal, there is a major drawback with such a design. This is in the time-domain response. Such a filter will show severe ringing. An impulse, either from a keyed signal or from a noise pulse, will cause an output that lasts much longer than the original excitation. Other filter types provide much better impulse characteristics. Examples are the Gaussian, the Maximally Flat Delay and the Linear Phase with Equiripple Error responses.[8] The compromise is in the frequency responses of these filters — the skirts are not as steep as those of a Chebyshev.

These fundamental problems are most evident with narrow cw filters. Chebyshev filters are usually acceptable for ssb applications. The best cw filters will have a peak response that is rounded rather than flat at the top.

Gain Flatness

There is barely adequate overall gain in some receivers. This leads to a receiver that sounds "quiet." In the absence of a signal, there is little noise output. Such a receiver is pleasing to use under normal operating conditions, but it is not adequate in other situations.

It is common with some receivers to find that the overall gain is lower on the higher bands. This is observed by replacing the antenna with a 50-ohm resistor and advancing the gain controls so the receiver noise is heard easily. Then, the band switch is changed. More often than not, the noise at 80 and 160 meters is significantly higher than it is in the 10-meter band. This is just opposite the gain distribution desired. Antenna noise is almost always the lowest at 10 meters. The problem is one of gain rather than noise figure, for even on 10 meters it is still common to hear an increase in receiver noise when an antenna is connected in place of a 50-ohm resistor. Receivers should have sufficient i-f gain to provide receiver noise that can be heard easily on any of the related bands.

Noise Blankers

Virtually all modern transceivers havew noise blankers. The performance can be spectacular for high amplitude noise pulses of short duration (automotive ignition noise). Unfortunately, the blankers are less than effective for other forms of noise. The most dramatic example is the over-the-horizon radar, or "woodpecker." Some builders have constructed special blankers that virtually eliminate interference from such additional sources.[9,10]

As we look to the future, we can expect that noise blankers will become much more sophisticated. Specifically, the blankers will be integrated with adaptive detection circuitry that will allow them to accommodate a much wider range of impulse widths and amplitudes. The key to such designs will be in implementing the blanking function without compromising the receiver close-in two-tone dynamic range.

Front-End Dynamic Range

Advances have been made recently in the design of receiver front-end sections. Ten years ago, many solid-state receivers were severely lacking in dynamic range (the ability to hear very weak signals in the presence of many strong ones). Today, many if not most receivers have two-tone dynamic range values of at least 90 dB.[11] Some equipment pushes the 100-dB figure, even in a ssb bandwidth.

It is interesting to examine the technology that is used to achieve the present-day designs. Surprisingly, most of the methods are *not* new. Diode-ring mixers have been available for over a decade. Power bipolar transistors suitable for strong rf and i-f amplifiers have been around for similar periods. The advances have been in the form of system analysis and specification, and in the realization that better performance is really needed.

This is not to say that device technology has not had an impact on front-end performance. Notable recent devices of interest are the VMOS FET mixers and monolithic integrated-circuit mixers for receiver front ends.[12]

In spite of the advances, commercial equipment with poor front-end design is being introduced. There is no excuse today for a receiver having less than an 80-dB two-tone dynamic range. Such performance is achieved easily, even with inexpensive components.

Noise in Local Oscillators and Synthesizers

Frequency synthesis has brought us a marvelous improvement in frequency stability — long overdue. Ironically, this has brought with it problems we do not need — notably, synthesizer noise and spurious responses. This is a recognized fundamental compromise.[13] The problems, while worse in synthesized equipment, are present in some equipment not using synthesis. Moreover, the noise problems are also evident in transmitted signals for which a synthesizer or noisy local oscillator is used for frequency control.

The basic nature of noise in an oscillator has been reported in the literature.[14] The problem is not generally appreciated by the amateur community. The reader is urged to review the literature to gain an understanding of this phenomenon.

The casual assumption is that the signal from an oscillator is "clean." After all, when observed with an oscilloscope, it appears as a sine wave. Usually, when observed with a spectrum analyzer, it looks like a rather pure tone, producing a trace resembling that of the filters in the analyzer. However, neither is a complete or accurate representation. Just because an oscillator has achieved limited level, this does not mean that the noise has vanished; the noise is still present. A representation of the output spectrum of an oscillator is shown in Fig. 3. The noise appears as a pair of sidebands, with modulation on the desired carrier. A broadband noise floor is also shown. The carrier-to-noise ratio is the vital parameter. The broadband ratio of the noise floor to carrier power is related directly to the transistor (or whatever) characteristics and to the power stored in the resonator during oscillation. The best (quietest) oscillators will be those with the maximum operating power level. The width of the noise pedestal is directly related to the *loaded* resonator Q of the operating oscillator. A well-designed oscillator will often have a loaded Q of about half of that of the Q_{μ} (unloaded Q) value of the resonator. All of the parameters may be controlled by proper design.

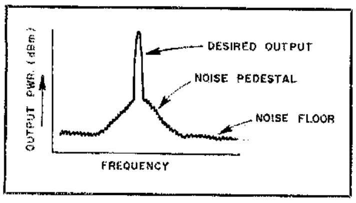


Fig. 3 — Nature of the output spectrum of a real oscillator. Desired output is actually the shape of the filter in the analyzer used for measurement. The noise floor is broadband in nature and is related to the operating power of the circuit. Noise-pedestal width is dictated by the loaded Q of the oscillator resonator.

How does this affect the performance of a receiver? The dominant phenomenon is termed "reciprocal mixing." Example: Assume that your neighbor has a very clean transmitter operating in one of the cw bands. You tune your receiver to his or her signal and note that the S meter needle is against the pin. As you tune away from the keyed cw signal, it would, ideally, disappear. You would then be able to copy the weak DX station that was only a kilohertz or two away. In reality, if your receiver local oscillator has excessive noise sidebands, you will hear a keyed hiss or noise as your neighbor keys the transmitter. Similar "splatter" occurs with ssb.

What if your receiver has a very clean LO system? Will the problems go away? They may or may not. The problems will cease if your neighbor's transmitter is really clean. If the rig is one with a similarly noisy local oscillator, however, the noise sidebands will be transmitted and you will hear them, even if you have cleaned up your own LO system.

How clean is clean enough? Consider that you wish to copy signals that are 100 dB below that of your neighbor, and that you are willing to tune at least 10 kHz away from him. Your receiver bandwidth is 1 kHz. Assume also that the front end is free of IMD and gain compression (blocking) for signals 100 dB more than the minimum detectable signal (MDS).

Recall that the problem is *noise*. Unlike a simple signal, the noise power will grow as you increase the bandwidth used to observe that noise. Hence, the noise sidebands must be 100 dB lower in your 1-kHz bandwidth, or 130 dB lower per hertz of bandwidth. Otherwise, your receiver LO will cause you to hear your neighbor instead of the desired weaker DX station. The situation becomes worse as you tune closer to your neighbor's signal.

The example has dealt with a simple L-C oscillator. The phase-noise value used in the example, −130 dBc/Hz (decibels with respect to carrier in a 1-Hz bandwidth) for the noise-sideband characteristic at a 10-kHz spacing, represents many rigs using L-C oscillators. Most are better, while a few are much worse! What happens when the local oscillator is replaced with a frequency synthesizer?

The largest problem related to synthesizer design, at least for hf applications, is that the oscillator must be voltage controlled. This is usually done with a Varactor® diode replacing the traditional mechanical variable capacitor. This imposes limitations. First, the diode must be operated in such a way that it *never* becomes forward biased by the rf signal in the oscillator. Should this occur, the tank Q will be degraded severely. This widens the oscillator noise pedestal. Also, this limits the amount of energy that may be stored in the resonator, thus degrading the broadband noise floor. For example, a receiver oscillator built by one of the writers had a voltage across the resonator of over 50, peak-to-peak. The measured noise was −154 dBc/Hz at a 10-kHz spacing. However, the high voltage required to achieve that noise performance would complicate tuning that circuit with a Varactor® diode. A Varactor could have been used to tune a very small range by using padder capacitors, keeping the rf voltage across the diode to a small value. However, the frequency range would have been too restricted.

The nature of the synthesis process will also lead to other potential spurious outputs. The typical synthesizer divides a sample of the voltage-controlled oscillator (VCO) down to a low frequency, usually in a programmable divider. The result, often at 1 kHz or 10 kHz, is then applied to a phase detector. A reference at the same frequency is also applied to the phase detector. The output is a chain of pulses that must be filtered. The dc component of that pulse chain forms the voltage that controls the VCO. The pulse nature of the detector output is removed with an active filter, the so-called "loop filter."

However, the filter can only be reasonably exotic. This is *not* merely a practical consideration. Excessive filtering can cause stability problems with the resulting phase-locked loop formed by the system. Hence, there will generally be a small amount of the reference frequency present at the output of the loop filter. This is applied directly to the VCO. This component leads to frequency modulation of the VCO at the reference frequency. The result is reference sidebands — signals on either side of the carrier displaced by the reference frequency. Careful system design will keep these sidebands low in amplitude with respect to the carrier. The sidebands may occur at many frequencies corresponding to harmonics of the reference frequency. This is the nature of fm. These sidebands can lead to reciprocal mixing, just as does the noise from a VCO.

Reciprocal mixing was measured on a popular synthesized transceiver. The performance was reasonable; it is one of the better amateur transceivers reviewed to date. The VCO phase noise at ±10 kHz was −142 dBc/Hz — rather good performance. This was degraded 8 to 10 dB by the presence of numerous reference sidebands. The performance was adequate, but not quite state of the art. This was a ham-band-only design.

The synthesis problems are complicated further in a general-coverage receiver. Performance is compromised by the wide tuning range required of the VCO — in excess of 30 MHz. The VCO is even more sensitive to extraneous voltages on the control line, including op-amp noise, power-supply noise and hum, and reference frequencies. One recent offering in the general-coverage arena sidesteps the problem partially by using not one, but six VCOs to cover the 30-MHz span. The even-more-restricted designs, where the VCO must only tune over a 500-kHz range, are better suited to the severe requirements of amateur communications. More sophisticated methods may be used to build extremely clean synthesizers. However, they often use a large number of phase-locked loops and become quite expensive.

The best performance the writers have obtained to date in a simple synthesizer is phase noise measured at −145 dBc/Hz. That synthesizer was restricted to just over 100 kHz of tuning range. We feel that phase-noise performance of LO systems is the most significant fundamental limitation on the advancement of receiver performance.

ARRL laboratory tests for receiver dynamic range have been complicated by the appearance of synthesized equipment. Often, the phase noise is so severe that blocking from gain compression can't even be measured. To this extent, phase noise may lead to what might be termed "early blocking," the condition where the increase in receiver noise output from reciprocal mixing dominates the response. The noise and spurious responses from synthesizers are sometimes so bad that it is difficult or impossible to identify front-end IMD products for two-tone dynamic-range measurements. This negates design efforts toward an improved front-end!

Other Problems

We have covered some of the major deficiences in modern equipment. There are other problems, as well as additional features that might be included in an "ideal" transceiver or receiver. Some of these are subjective and represent personal preferences. They do not reflect fundamental deficiencies. Others are more general and important.

One of the greatest problems relates to the complexity of much of the modern equipment. As features are added and fundamental performance is extended, the parts count increases and reliability suffers.

Reliability improves with enhanced quality. We are seeing new pieces of amateur gear that are inoperative or intermittent when it is first removed from the shipping cartons. This "infant mortality" has been traced to unsoldered joints, loose components and even wiring errors.

Another measure that can improve long-term reliability significantly, especially infant-mortality problems, is a burn-in. That is, once the equipment is completed and calibrated by the manufacturer, it is run for an extended period. It is then rechecked and calibrated prior to shipment. At least one manufacturer has instituted such a program with all of its equipment.

Many of the service manuals for existing equipment are poor. Moreover, some of the equipment is so sophisticated that it can't be serviced with the equipment found in a *well equipped* amateur shop. Few of us have spectrum analyzers or high-quality, wide-range signal generators. It would be helpful if service procedures were written that would allow for most of the routine maintenance with simpler equipment.

Shielding is poor in some equipment. It is, however, improving — partially because of the extensive amount of digital circuitry contained in the equipment. Not only should the signal coming from the antenna terminal be clean and free of spurious products, but the power lines, key and microphone lines, and whatever other interface cables might be used, should all be protected against outgoing and incoming radiation.

The Ideal Rig

As the cost of amateur equipment continues to rise, we should be able to expect advances in the design and performance of the units we buy. Price and quality do not always create a satisfactory blend today. That is, some rigs of modest cost seem to equal the top-of-the-line equipment in some performance areas. An understanding of the amateur's needs in equipment performance and features should be a fundamental rule among design engineers. The designer should assume that the product will be used in areas of dense amateur activity. This requires attention to receiver dynamic range, LO noise, transmitted noise, reduced spurious emissions, low IMD in transmitted ssb, and clickless cw waveforms (Fig. 4). Shielding and filtering should be adequate.

On the other hand, what is ideal for one amateur may be totally inadequate for another. That is, the devoted contest enthusiast needs more performance than the amateur who only wants to have something on hand for an occasional QSO. Don't go for more performance than you really need. Be aware that some features that may seem appealing are accompanied by potential performance compromises, or much higher cost and decreased reliability, or both. An example is the phase-noise compromise related to a general-coverage, synthesized transceiver.

It is difficult to generalize about where minimum performance specifications should be set. As mentioned above, it will depend upon the intended application. Still, realizing the subjectiveness, we would suggest the following:

Necessary Characteristics

1) Noise figure: 10 to 12 dB maximum (MDS may be calculated on the basis of the receiver bandwidth).

2) Blocking from gain compression: 110 dB above the MDS in a cw bandwidth for casual applications, 125 dB for contesting and DXing.

3) Two-tone dynamic range: 80 dB in a cw bandwidth for casual applications; 95 dB or more for competitive contesting and DXing.

4) All transmitter spurious responses down by 60 dB or more from the cw output.

5) Transmitter IMD on ssb greater than 30 dB below each tone of the two-tone test signal for casual applications; 35 dB or more for competitive use.

6) LO noise: Carrier-to-noise ratio in excess of 125 dBc/Hz at 10 kHz separation for casual applications; 140 dBc/Hz for competitive application.

7) Cw waveforms: 5 ms rise and fall times with a dot length equal to that injected at the key jack.

8) Agc: Attack characteristics such that the overshoot of the audio output from keying a carrier into the antenna jack, 60 dB over the MDS, does not exceed the steady-state value by more than 3 dB. Decay characteristics: 100 ms or less for "fast" recovery; 0.5 second or more for "slow" recovery; no more than a 100% difference for signals of various amplitudes. Agc threshold: − 100 dBm or less, but at least 20 dB over the MDS.

9) Receiver audio output: At least 1 W of output into the design load, with less than 10% total harmonic distortion (THD).

10) Frequency drift: 300 Hz or less warm-up drift in the first 30 minutes of operation in a "room-temperature-stable" environment; 100 Hz per half hour or less thereafter (should be much better in a synthesized system).

11) Frequency resolution: 100 Hz in a counted system; 1 kHz in an analog system.

12) Frequency accuracy: Within twice the resolution quoted above.

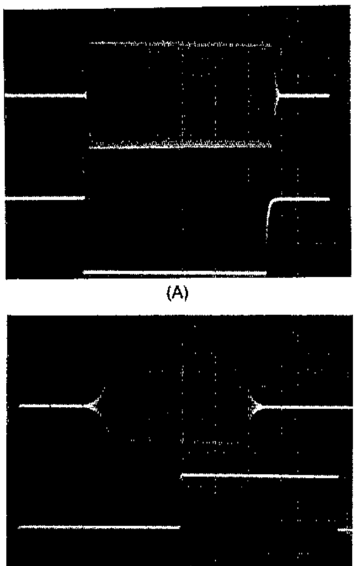
13) Coherent receiver spurious responses: None to exceed the MDS by more than 10 dB.

14) Reliability: Mean time before failure (MTBF) of three years, assuming four hours of use per day.

15) Receiver i-f stopband attenuation: Greater than 100 dB.

16) Image and i-f rejection: Equal to the two-tone dynamic range in a ssb bandwidth, or better.

17) Power supply: The 50/60-Hz, 117- or 234-V ac power supply should be capable of operating within its ratings without overheating and failure. Some accessory power supplies cannot be operated safely at 50 Hz, especially in tropical regions, with the cabinet cover in place. Most units have inadequate louver area for the high temperatures of the transformer core under such conditions. Adequate venting and an accessory cooling fan should be available to the purchaser.



(A)

(B)

Fig. 4 — Oscillographs of (A) an improperly shaped cw waveform and (B) a waveform that closely approaches the desired 5-ms rise and fall times. The waveform at A causes severe key clicks and is common to a number of commercial and homemade transmitters today. The example at B should provide clickless keying without being too "soft."

How to Buy a Rig

Now that the "ideal" rig has been specified, we come to making a suitable purchase. There are a variety of requirements, depending on the application. First, study the specifications in advertisements; these rarely tell the whole story. Most manufacturers specify a piece of equipment so that virtually any model making it through the manufacturing process will meet those specifications. Hence, they are rarely meaningful. Reviews in *QST* and other amateur literature are often more informative. But, they suffer from being the result of investigating only one or two samples of the unit in question. Also, owing to the high rate of introduction of new equipment, the prospective buyer may not want to wait until a review is published.

The best test is to *use* the rig. A possible way to do this would be to operate the transceiver, ideally in conditions under which the rig will be used. If possible, ask the dealer if you can try the receiver or transceiver for a day or two before a purchase. The answer will most likely be "No," but it's worth a try. Otherwise, look for a friend who owns the type of rig you're interested in. Ask for his or her opinion or, better still, ask if you can operate the rig.

There is another alternative that could be productive for the aggressive dealer — that of equipment rental. A dealer could allow a rig to be rented for a short period at a reasonable fee, later applicable against a purchase.

If your ideal rig cannot be found, you can build your own! You should end up with the specifications that you really need.

Conclusions

The manufacturer that can provide the features listed earlier *(Necessary Characteristics)* at a fair price is sure to obtain a significant segment of the amateur market. These are reasonable goals for the homemade-gear devotee as well.

Although we have taken some well-meant "pot-shots" at the manufacturers, we have tried to base our judgments on realistic observations. We can conclude that there are many areas of deficiency. On the other hand, we cannot help but be impressed with the high performance usually being offered at a reasonable price by most manufacturers. Ultimately, the equipment will be what the market demands.

Notes

[1]W. Hayward, "Receiver Dynamic Range," *QST*, July 1975. Also see Technical Correspondence, Nov. 1979 *QST*.

[2]W. Hayward and D. DeMaw, *Solid State Design for the Radio Amateur* (Newington: ARRL, 1977), Chapter 6.

[3]*The Radio Amateur's Handbook*, any recent edition.

[4]W. Hayward, "A Competition-Grade CW Receiver," *QST*, March and April 1974.

[5]W. Hayward, *Introduction to Radio Frequency Design* (Englewood Cliffs, NJ: Prentice-Hall, 1982), Ch. 6. There has long been an informal "standard" that S9 is equivalent to a signal level of 50 μV, with each S unit below that level 6 dB weaker. In 1978, the IARU Region 1 Division adopted this standard for hf receivers, with 5 μV equivalent to S9 in vhf/uhf receivers. See J. Lindholm, *QST*, March 1982, p. 83, for the other specifications of the Region 1 standard.

[6]See note 4.

[7]H. Blinchikoff and A. Zverev, *Filtering in the Time and Frequency Domains* (New York: Wiley, 1976).

[8]A. Zverev, *Handbook of Filter Synthesis* (New York: Wiley, 1967).

[9]Nicholls, "Blanking the Woodpecker," *Ham Radio*, Jan. and Feb. 1982.

[10]U. Rohde, "Increasing Receiver Dynamic Range," *QST*, May 1980.

[11]See note 5 for details on the measurement of MDS, noise figure, two-tone dynamic range, input intercept, gain compression and LO noise (Chapter 8).

[12]D. DeMaw, and G. Collins, *QST*, Jan. 1981.

[13]W. Sabin, "Use of Mixers in HF Up-Conversion Receivers/Exciters," *IEEE WESCON Conference*, San Francisco, 1982.

[14]Oscillator noise is covered in more detail in Chapter 7 of note 5.

New Books

☐ *Computers and the Radio Amateur*, by Phil Anderson, WØXI. Published by Prentice-Hall, Inc., Englewood Cliffs, NJ 07632. Hardcover, 7-1/4 × 9-1/2 inches, 208 pp., $18.95.

This slim volume is an innovation. Being the first to market a new device, a new product, or, in this case a new book, has advantages and disadvantages. An obvious plus is the opportunity to start fresh with a clean sheet of paper in the typewriter or on the drawing board and, not to be taken lightly, the possibility of profiting from being first out of the starting blocks. On the other hand, a product brought too quickly to market may well suffer from a number of "birth pains." That seems to be the case with *Computers and the Radio Amateur*. The *concept* of the book is an excellent one: a book (as the preface puts it): "for radio amateurs who have had little or no exposure to computers — how they function and how one programs them and attaches them to other equipment."

But, there seems to have been a number of complications, some serious, that developed between setting down the concept of the book and turning that concept into a workable manuscript.

For example, on page 3 there is a subsection headed "RTTY or ASCII terminal" yet the all-too-brief material under that section deals with Baudot code Teletypewriter signals and there is no mention, at that point in the text, of ASCII at all! (By the way, as far as I know, "Teletype" remains a registered trademark of the Teletype Corporation although it is not so identified in the book.)

Chapter 2, which is a brief history of computers, is *so* brief (less than one page takes us from 1890 to post World War II!) that it might as well have been omitted.

Chapter 4 satisfactorily explains the essential elements of BASIC programming. The author is liberal with the use of REM statements, which is a good programming style, especially during the debugging phase of writing programs. If the reader should wish to actually use the sample programs, he or she will likely wish to delete the REMs, especially those that, unaccountably, give the "example number" from the text.

Chapter 5 is devoted to assembly language programming. In the earlier chapter on BASIC, the particular dialect often used is from the TRS-80® Radio Shack level II language utilized by the Tandy Corporation. In Chapter 5, however, the examples are drawn from the 6502 microprocessor, which is used in a number of personal computers including the Atari but *not* in the Radio Shack machine. While Chapter 5 is interesting it will hardly make one an assembly programmer. (Actually, I wish I knew what *would*!)

The remaining seven chapters are devoted to interfacing the computer to amateur equipment and to some examples of computer programs you can use in your own shack. Many of the suggested programs are excellent examples of the ways the computer can be useful in Amateur Radio applications. Included are examples of contest log programs, code-reading programs and the like. As programming skills are sharpened, however, the reader might wish to try rewriting the program on page 182 to try to calculate the length of a dipole in fewer than 19 program lines. (Hint: There is only one really operative line — line 100, L = 234/F.)

If you have no knowledge of computers or how they might be useful in Amateur Radio, this book may be a good addition to your radio shack bookshelf. It is a relatively thin book that won't give you all the answers, but perhaps it will help you recognize what questions you need answered. — *Larry Price, W4RA*

Next Month in QST

- With action heating up on 10 MHz, you'll want to join in as quickly as possible. If you have a TS-820 or '820S, help is on the way. An article in February *QST* will show you how to add any or all of the WARC bands.
- Thinking about getting on line? The February On Line column will solve the mystery of how one type of BASIC is translated to another.
- 1982 was a terrific year for Amateur Radio! If you don't believe it, read all about it in the "year in review" article.

Automatic Control for the Alliance HD-73 Heavy-Duty Rotator

Tired of your manual rotator control? Try these modifications and enjoy "hands-free" operation!

By Joe Mehaffey,* K4IHP

For rotating large antennas, the Alliance HD-73 is vastly superior to the standard "TV" rotator motor used in the Alliance C-225 system. However, the manufacturers provide only a manual (push-to-turn) control box with the HD-73 system. After having enjoyed the fully automatic positioning feature of the C-225 control unit, I was very disappointed to have to hold my hand on a control lever while watching a meter slowly move to the correct position. With my C-225 unit, I simply rotated the control knob to the desired position, and the control box automatically shut off at the proper position. After using the HD-73 system for a few years, I set out to investigate what it would take to mate the C-225 control box with the HD-73 motor unit.

System Differences

Basically, the HD-73 and C-225 rotator systems vary in the design of the direction-sensing and display circuitry. However, in both cases a potentiometer is used in the motor housing to reference the antenna position back to the control box. In the HD-73, a 75-ohm potentiometer is used as the feedback device, and in the C-225 a slide-wire potentiometer of approximately 520 ohms is used. Additionally, the C-225 motor has the capability of rotating in excess of 360°, whereas the HD-73 motor stops prevent rotation beyond 360°.

To make the HD-73 motor operate with the C-225 control box, it is necessary to only make small modifications to both the rotator and control unit. The basic procedures are outlined here:

Rotation Angle: The HD-73 motor assembly must be modified to give it a small amount of overtravel at both the clockwise and counterclockwise directions of rotation. This is accomplished easily by physically modifying the limit stop lever within the HD-73 motor drive unit.

Potentiometer: A precision servo potentiometer must be substituted for the inexpensive wire-wound unit presently used in the HD-73. It is necessary to replace this unit for two reasons. The first is that standard potentiometers (such as the original) have proved to be so noisy and low in resolution that they cannot serve adequately in servo-positioning applications. The second reason is that the

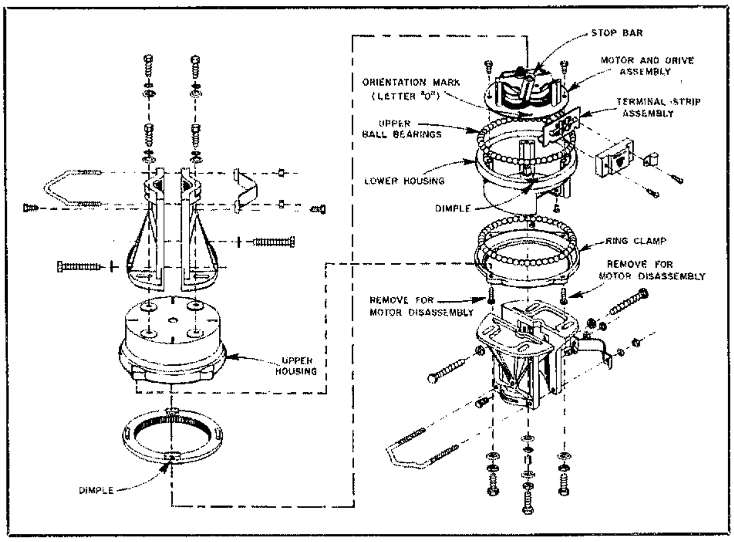


Fig. 1 — Exploded view of the HD-73 motor unit (diagram courtesy of the Alliance Manufacturing Company).

*6950 Hunter's Knoll, Atlanta, GA 30328

C-225 control-unit bridge is designed for a potentiometer in the 500-ohm range.

Bridge Balance Resistor: The third modification required is in the C-225 control unit, and involves changing the 511-ohm bridge resistor to a 464-ohm, 1% resistor. This change is necessary because the 360° rotation angle of the HD-73 motor results in a total resistance change of only about 430 ohms, as opposed to the total 500-ohm range of the servo potentiometer.

Rotation Range Adjustment: The last change required to complete the conversion is the selection of a rotation-range resistor, to be added in parallel with the 520-ohm, wire-wound strip resistor within the C-225 control unit.

The previously listed mechanical and electrical modifications cost about $20. Most of the work is involved with calibration and adjustment of the motor control unit. Much of my time in the original conversion was spent figuring out how to adjust the potentiometer in the HD-73 motor unit without spilling the ball bearings over the floor! Once this procedure was perfected, the calibration went very quickly.

Motor Disassembly

This procedure involves the disassembly of the HD-73 motor unit, and modification of the mechanical stop lever to permit the HD-73 rotator approximately 5° of overtravel in both directions. This is necessary to compensate for electrical tolerances that could direct the rotator to travel either in excess of slightly less than the desired 360° range.

To begin disassembly, please refer to the exploded view in Fig. 1. Invert the motor unit and clamp it in a vise with the screw heads facing upward. (This entire procedure should be performed in a clean area and executed very carefully, because of the likelihood that the ball bearings will become dislodged and fall out of the rotator motor assembly.) Remove the four screws from the bearing ring-clamp assembly. Lift the bearing-clamp assembly straight up off the motor unit. Remove all of the ball bearings from the upper bearing race. Carefully lift the upper housing assembly away from the lower one. No force should be required, but it may be necessary to wiggle the lower housing assembly as it is pulled away so that the bearings are released and remain in the upper housing race. The motor and drive assembly are installed inside the lower housing. This is where the first two modifications must be made.

Remove the stop bar from the mounting shaft and grind both sides as shown in Fig. 2. Smooth and reinstall the stop bar. In my case, it was necessary to destroy the lock washer (a type of speed nut) in order to remove the stop bar. I found an exact replacement at my local hardware store. If one cannot be found, a press-on axle nut for a toy tricycle should make an adequate substitute, and is available at most large hardware stores. Do not use excess force when installing the washer, or the casting may be damaged. The modified stop bar should have additional travel against the upper-housing stop tang. Don't forget to apply a small amount of bearing lubricant to the stop bar prior to installation.

Potentiometer Replacement

A Spectrol model 132-0-0-501 potentiometer was purchased to replace the original. This 500-ohm, wire-wound device is a precision semi-sealed unit with high resolution and long life expectancy in this type of service. This unit does not have stops at either end of its travel. A potentiometer with internal stops can be used. However, one runs the risk of damaging the stops if the potentiometer was initially adjusted incorrectly. To install the potentiometer, remove the motor and drive assembly from the lower housing by removing the four no. 10-24 hex-head screws holding the drive assembly to the lower housing assembly. The mounting plate will now swing freely on the connecting wires to the potentiometer and motor. Care should be used to bend the delicate connecting wires a minimum amount to prevent breakage.

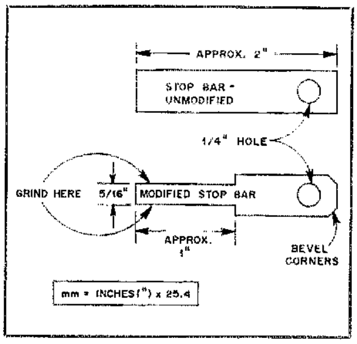


Fig. 2 — Illustration of stop bar prior to and after modification.

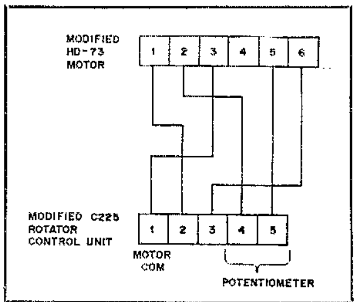


Fig. 3 — Interconnections for the modified rotator system.

Using an Allen wrench, loosen the gear at the top of the potentiometer shaft, and then loosen the mounting nut — freeing the potentiometer from the motor and drive assembly. Install the new potentiometer in its place. At this point, orientation of the potentiometer shaft relative to the body is not important, but use care that the keying peg goes into the keying hole in the motor drive unit. It may be necessary to enlarge the keying hole in the motor and drive assembly base slightly to accommodate the peg on the potentiometer. Be sure that no drilling debris remains within the motor assembly. After securely mounting the potentiometer and the associated drive gear, move the three wires from the old potentiometer one at a time to the Spectrol unit. Carefully observe which wire goes to which terminal, and put the wires on the new potentiometer exactly as they were on the old.

Refit the motor and drive assembly into the lower housing, and replace the hex-head screws. Use care not to pinch any wires as the drive assembly is reinstalled. As a preliminary setup, place an ohmmeter between pins 2 and 5 on the terminal strip of the motor unit. Rotate the potentiometer until the resistance is approximately 35 ohms. On the outside casing of the lower housing, mark the position of the potentiometer gear for future calibration reference. While viewing the motor and drive assembly as oriented in Fig. 1, move the stop bar all the way clockwise against the stop tang of the motor and drive assembly. Install the bearings in the race of the upper housing assembly. Invert the lower housing unit and mate it to the upper housing unit with the stop bar positioned just to the left of the stop tang. (The stop tang protrudes from the interior of the upper housing.) At this point, reinstall the bearings in the bottom of the lower housing race, and fasten the ring clamp using only two of the hex-head screws. These screws should not be tightened securely as they will have to be removed several times during adjustment.

Some experimentation is necessary to ensure that the HD-73 motor operates normally and that the newly installed potentiometer changes resistance as the motor rotates. To perform this test, connect the HD-73 to the C-225 control unit as shown in Fig. 3. Plug in the C-225 and move the control knob to the middle of its operating range. The motor should turn to approximately mid-position and automatically stop, which then turns off the red indicator light in the C-225 control unit.

Assuming the motor turns, rotate the control knob fully clockwise and counterclockwise, and note that the motor runs into the mechanical stops but does not turn off the control unit at both rotational extremes. Unplug the control unit as soon as the motor drives into the stop. Carefully mark both extremes of motor travel by making scratches on the motor housing. These will be used for reference during the final adjustments.

If the preliminary tests were successful, proceed to modify the control box as described below. Failure at this point might be traced to either reversed potentiometer wiring or reversed motor wiring. Be sure that when the HD-73 is upright, a clockwise rotation of the control knob causes a clockwise rotation of the motor as viewed from above.

Control Unit Modifications

Change the 511-ohm, 1% resistor (item 101-3 in the C-225A control unit, or item 101-11 in the C-225 control unit) to a 464-ohm, 1% half-watt unit. You should have no trouble locating this resistor, for it is the only precision resistor in the control unit. Now locate the two ends of the 520-ohm strip resistor that is a part of the control-knob assembly. This strip resistor has a slide wiper that moves around the spiral — first to one end, then to the other as the control knob is rotated. Carefully solder two flexible wires approximately 1 foot long to each end of the slide-wire unit.[1] Do *not* overheat the rivets or they may become loose, and intermittent operation of the slide-wire potentiometer may result. As part of the calibration process, several resistors may have to be connected across these newly installed wires.

Counterclockwise Stop Calibration — Motor Unit

Rotate the control knob fully counterclockwise until the rotator motor stops with the control light still illuminated. Quickly unplug the control box to prevent burnout. Remove the wire from screw terminal no. 2 of the motor unit and connect an ohmmeter between this terminal and no. 5. Remove the two screws from the ring clamp assembly and carefully rock the lower housing away from the upper housing. This maneuver should be executed similar to the opening of a clam shell — so that a crack just wide enough to insert a small screwdriver appears directly in front of the previously marked potentiometer-gear location. Using a small screwdriver, move the potentiometer gear until the ohmmeter reads approximately 450 ohms. Gently reclose the motor housing and reinstall the two locking screws into the ring clamp assembly. Next, replace the wire on terminal no. 2, and note whether or not the motor moves against the counterclockwise stop. The objective is for the motor to stop in a position approximately 3/8 to 1/2 inch away from the mechanical stop.

If the motor continues to run into the mechanical stop, the assembly must be reopened and the potentiometer setting increased by approximately 5 ohms. If the motor stops with a gap appreciably greater than 1/2 inch, the potentiometer should be lowered in resistance by approximately 5 ohms and retested. This procedure should be repeated until the counterclockwise stop is properly adjusted. When this adjustment is completed, work on the motor assembly is now finished and all four ring-clamp mounting screws may be secured.

Clockwise Stop Calibration — Control Unit

To begin, connect a 3.3-kΩ, 1/2-W, 5% resistor to the ends of the two wires previously soldered in the C-225 control unit. Rotate the direction control fully clockwise and wait for the motor to reach the end of its rotation. If it rotated less than 360°, slightly increase the value of the 3.3-kΩ resistor. If it rotated more than than 360°, decrease the value of the resistor. The proper value for this resistor should be between 2.5 and 5.0 kΩ. When this adjustment is completed, place the resistor safely within the control unit so it cannot short to the chassis or other internal parts. Make sure the wires are dressed properly and do not interfere with the operation of the slide-wire assembly.

A purist may wish to remove the wires and solder in the resistor. If this is done, much care is required to avoid damage to the slide-wire assembly. Be sure to unplug the control box when making any changes, to avoid electrical shock.

Operation

The modified Alliance HD-73 motor and C-225 control unit combination have been in use for several months at my location. No problems have been experienced. Some may observe that the padding method used on the control-box potentiometer will cause a slight nonlinearity between the actual motor position versus control-box indication. This error will be maximum at the middle of the rotation range. The error is only a few degrees, however, and while measurable on the bench it is not perceivable to the eye. The angular error is very small compared to the beamwidth of most rotatable antennas, so there is nothing to worry about.

Acknowledgments

I would like to thank the Alliance Manufacturing Company for their assistance with this project. Their donation of service manuals and cooperation is greatly appreciated.

References

Alliance Tenna-Rotor Service Manual. Alliance Manufacturing Co., Alliance, Ohio.
HD-73 Heavy-Duty Rotator Owner's Manual. Alliance Manufacturing Co., Alliance, Ohio.

[1] m = ft × 0.3048, mm = in. × 25.4.

Strays

TALLYHO THE FOXHUNT

☐ Sure, I know what a foxhunt is. That is, I thought I did until I got involved in Amateur Radio and happened to tune across the 40- and 15-meter bands one weekend in September and heard "CQ Foxhunt, this is WB1EMT," followed by a pileup. No, they weren't looking for that elusive furry red animal, and no direction-finding equipment was involved. The word "foxhunt" comes from its sponsor, the Foxboro Company (Massachusetts) Amateur Radio Club.

The idea was conceived two years ago as an activity to get as many club members as possible active on the hf bands. During the first year's contest, WB1EMT was a portable operation, as part of an emergency communications demonstration at the Bellingham Fair. Besides being a "second Field Day" for the club, it got some spectators interested enough to join our Novice class.

The object of the contest is to first work club station WB1EMT, which operates on 40 and 15 meters and issues a log number with an "F" prefix. Now the hunt begins, for you must then look for other club stations operating on the band using their own call who will, in turn, issue an "O" log number, and then an "X," when they are found and worked. When you have caught the F-O-X, you qualify for a certificate, providing your log numbers check out. Sound too easy? Although frequencies are coordinated among club members by way of the club repeater, K1HF/R 147.975/375, the operating frequencies are not made known to the participants. Trying to find another "CQ Foxhunt" indeed becomes a search. For variation, this year we also had a cw station available to hunt, and we did have stations trying to earn our award on both bands. — *Frank S. Jasinski, W1XA, Bellingham, Massachusetts*

Foxboro Company ARC member WA1UMA operates from the club van during the annual Foxhunt Contest on 40 and 15 meters.

The Inverted L Revisited

City dwellers, don't despair. Here is a *good* 160-meter antenna that should fit on your lot!

By John F. Lindholm,* W1XX

I'll never know what inspired me to make a few contacts in the ARRL 160-Meter Contest. My antenna was made by tying together the open-wire feeders of my 80-meter dipole. The performance was not fantastic, but it was the first step in getting me "hooked" on the "gentlemen's band." Working three dozen European stations from a friend's house got my interest up. My friend has a good 160-meter antenna system — and *lots* of property to fit it on!

Returning to my 60- × 150-foot lot made me feel depressed.[1] I suffered all winter while listening to the others working VKs, ZLs and even JAs at daybreak. What could I do to improve *my* signal?

Many hours the following summer were spent trying to figure out how to cram a 160-meter antenna within the confines of my small lot. Space restrictions dictated that my wire be no longer than a standard 80-meter dipole. I began to consider alternative antennas.

Shunt feeding my 50-foot tower was investigated but dismissed for various reasons. This arrangement would require disconnecting the shunt feed when cranking down the tower, and all guy wires would have to be broken up into nonresonant lengths with insulators. Additionally, my cables would have to be rerouted to ground level. I then considered a full-length dipole originating in a neighbor's yard three houses to the east, crossing my property and finally terminating in the yard two houses to the west! This idea was rejected. The legal negotiation fees would have run in six figures! After much head scratching, I settled on the inverted L, an antenna made popular by the grand master of 160-meters, Stew Perry, W1BB. I credit Stew for coming to my rescue!

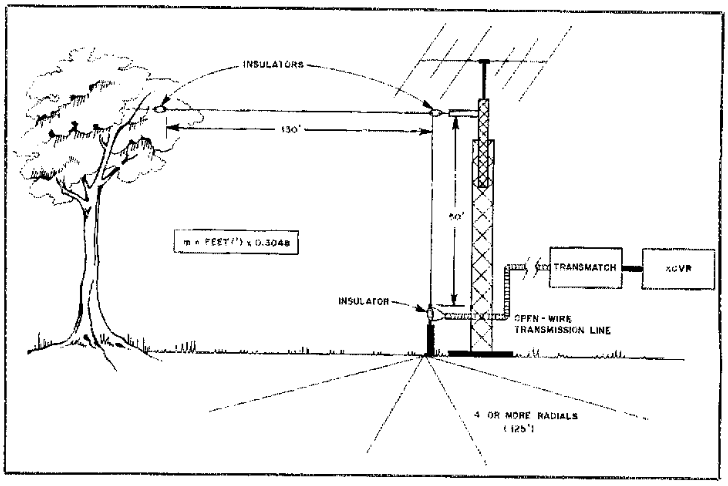


Fig. 1 — The W1XX inverted L is arranged in this manner.

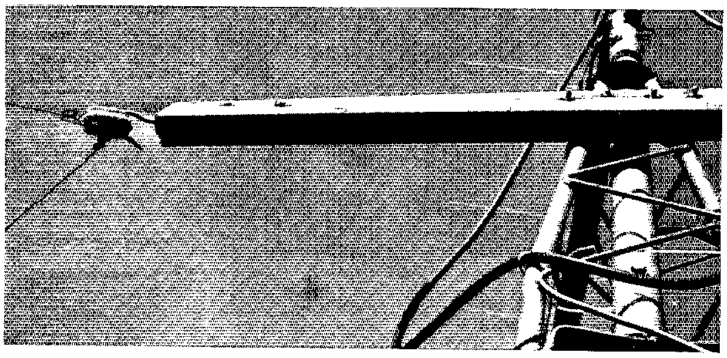


Fig. 2 — Detail of how the antenna is mounted to the top of the tower. TV-mast clamps are used to secure the wooden insulator.

The Inverted L

The inverted L was selected because it requires no more space than an 80-meter dipole and I could utilize my 50-foot crank-up tower for attachment. The vertical part is 50 feet long, and the horizontal part measures 130 feet, for an overall length of 180 feet (Fig. 1). This makes the antenna approximately 3/8 λ, with the horizontal part providing top-loading. W1BB advises making the vertical section as long as possible (depending on tower height), and that an overall length of 160 to 180 feet works well.

Construction

With a bow and arrow, I successfully attached the far horizontal end of the antenna to the top of a 60-foot fir tree. From there, I ran the wire back to the top of the tower, where I bracketed a 30-inch-long two-by-four with an insulator screwed in the end (Fig. 2). No. 14 Copperweld® wire is used for the horizontal section, and no. 10 copper wire is used for the vertical section, which is spaced about 2

[1]Notes appear on page 21.

*Communications Manager, ARRL

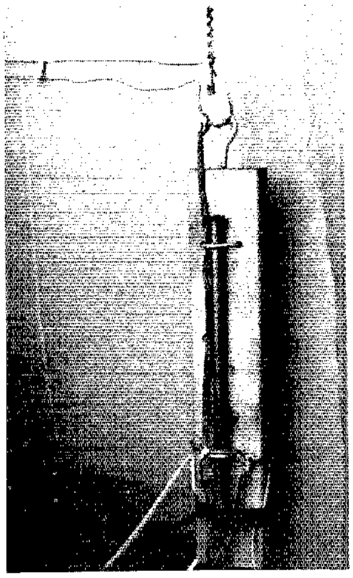


Fig. 3 — A ground rod at the tower base serves as a physical support for the insulator, and as an rf ground. Radial wires are connected to the rod by means of a ground bus, as described in the text.

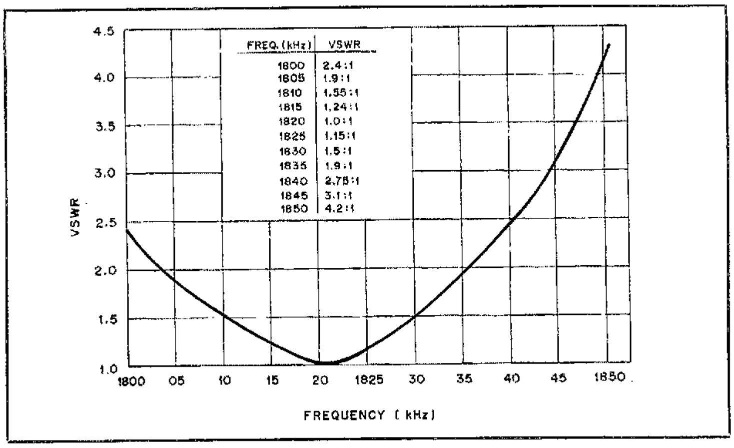


Fig. 4 — SWR curve for the inverted L, with the Transmatch adjusted for an SWR of 1 at 1820 kHz.

feet from the tower. These two wires are soldered together at the insulator.

A plumb line was used to locate a point on the ground directly below the 90° bend in the "L." At this point, a 10-foot copper-clad ground rod was pounded into the earth, leaving 18 inches sticking out. A two-by-four is clamped to the ground rod by means of TV-mast U bolts. The lower end of the vertical wire is attached to an insulator that is screwed into the top of the wooden block (Fig. 3). Next, I stripped some coaxial cable (RG-8/U) of its outer braid and used this to make a ground bus around the antenna base. One side of the open-wire transmission line is soldered to the base of the vertical antenna element, and the other side to the bus, which is attached to the ground rod with a clamp. This bus also serves as a connection point for the radials. So far, less than an afternoon of work had been invested.

Radials — the More the Better

The next day, my objective was to install radial wires, which are necessary because the inverted L is essentially a top-loaded vertical radiator. Previous meditation convinced me that several radials would fit on my lot. I'd *make* them fit! All radials were cut to 1/4 λ (125 feet), using scrap wire. About 300 feet of surplus telephone ground wire provided a good start. Stripping some old coaxial cable with a single-edged razor blade produced two radials from one length of wire (outer and center conductors). My technique for radial installation consisted of creasing the earth with a spade and shoving the wire in; afterwards the turf was pressed back into place with my heel.

With space a problem, it may be impossible to place all your radials in a straight line; don't worry, because it is not necessary. My installation followed a zigzag path to avoid fixtures like the house and driveway. W1BB advises putting some radials under the horizontal part of the antenna. Unfortunately, the location of my garage prevented this. Initially, only four radials were planted, but more were added later. As with all vertical antennas, the more radials you can put in the ground, the better the performance!

Matching System

Voltage-fed antennas approximately 1/2-λ long, such as the inverted L, will have a fairly high feed-point impedance. They will also exhibit inductive or capacitive reactance, depending on whether the antenna is slightly longer or shorter than 1/2 λ. Since the inverted L does not have a 50-ohm impedance, a matching system is needed. To make tuning adjustments easier, I opted for locating the Transmatch *in* the shack, using open-wire line to the antenna. Being able to adjust the antenna match conveniently is recommended, since the SWR climbs rapidly as you shift frequency. For example, adjusting the Transmatch for a 1:1 SWR at 1820 kHz produces an SWR of 2.4:1 at 1800 kHz (see Fig. 4). With the Transmatch located in the shack, you can easily adjust for a 1:1 SWR no matter where you operate in the band. My Transmatch consists of a plug-in, link-coupled coil and variable capacitors — all scrounged at flea markets. Any of the configurations found in the *ARRL Antenna Book*[2] should work well.

Performance

Does the antenna work? Having no comparison antenna, my conclusions are subjective. But I've been on the air enough to know when, as they say, "it plays." With only four radials in place, my first night of operation yielded plenty of U.S. contacts, plus a Caribbean DXpedition on the first call. Subsequently, many European stations have been worked from my northeast location with good signal reports. Contest activity has yielded some respectable scores, including many QSOs with the Caribbean, and South and Central America — even Antarctica!

By adjusting my Transmatch, I made limited tests with the L on 75-meter ssb. Comparisons were made to a 75-meter dipole at 50 feet. For signals close in, the L was down by some 3 to 5 dB, but equal or superior to the dipole for signals from eastern Europe. Apparently, the 160-meter inverted L also provides a low angle of radiation on this band too. On bands higher than 75 meters, the radiation angle will be tilted *upwards*, rendering the antenna inefficient for DX work. This phenomenon is explained in the *ARRL Antenna Book*.[3]

Giving up the 160-meter band for lack of sufficient real estate is unwarranted. With the inverted L, you can work Top Band from your urban lot. Installation is a breeze, and the performance is admirable. *Now* what's your excuse for missing out on the excitement of 160 meters?

Notes

[1]m = ft × 0.3048.
[2]*The ARRL Antenna Book*, 14th ed. (Newington: ARRL, Inc., 1982), pp. 4-1 through 4-8.
[3]*Ibid.*, pp. 2-23 through 2-24.

Beverage Antennas for Amateur Communications

Contrary to popular belief, the Beverage antenna can be used as an effective receiving *and* transmitting antenna for frequencies up to 30 MHz.

By John S. Belrose,* VE2CV, John Litva,*, G. E. Moss* and E. E. Stevens,** VE3CYO

A Beverage antenna is a broad-band aperiodic antenna that can be used over a frequency range of 2 to 30 MHz. It consists of a long wire stretched horizontally over the ground. In essence, it is a lossy transmission line with the ground acting as an imperfect conductor (Fig. 1). The antenna is terminated in the characteristic impedance of 400 to 600 ohms through a ground screen, with the received signal taken from the other end through a matching transformer that has one side connected to ground. This transformer is used to match the feed impedance of the antenna to 50-ohm coaxial cable. Dimensions of a typical Beverage antenna would be about 360 feet long and 6 feet above the ground.[1]

Beverage antennas can be used for receiving (single element) or transmitting (multiple-element array) applications. While no detailed study has been made on the optimum size of ground screens, the hf Beverage antennas we built employed radial ground screens comprised of 16 radials, 50 feet long, staked at the ends by 2-foot-long rods. Multielement arrays employ smaller ground systems — in our case, six radials, each 20 feet long. The connections to the ground side of the matching transformer and the terminating resistors were made with three, 3/4-inch-wide copper mesh ground straps.

Previous Work

The initial Beverage antenna work was carried out by H. H. Beverage and associates, who tested the antennas on a transoceanic circuit using very long waves of 12 to 42 kHz, and at medium frequencies of 250 to 1500 kHz.[2,3,4]. They found antenna lengths of approximately 1 wavelength were effective in reducing interference and static because of the directive nature of the antenna and its low response to radio noise. Wait and Mousseau calculated vertical-plane field patterns for horizontal traveling-wave antennas.[5] Travers and associates did extensive theoretical and experimental research with hf Beverage antennas from 1961 to 1967.[6] Their work is documented in a series of reports that have had limited distribution. A brief account of their work appears in a paper by Martin.[7] More recently, Litva and Rook gave a detailed description of experimental and theoretical results obtained from an extensive study of hf Beverage antennas.[8] Their theoretical work, supported by extensive experimental measurements, provides comprehensive Beverage antenna engineering data in a readily accessible format for the communications engineer.

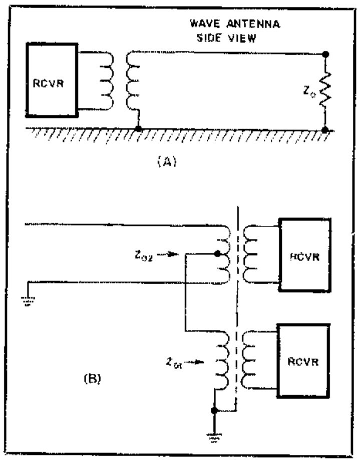


Fig. 1 — Alternative forms of wave antennas: (A) simplest form for unidirectional reception; (B) two-wire wave antenna for separate reception from reciprocal directions.

Beverage antennas are shown to be effective as "building blocks" for large hf arrays: When multiple Beverage elements are employed, the antenna system has sufficient gain at hf to be efficient in both receiving and transmitting. Of course, the directivity and size of such an array would dictate bidirectional point-to-point operation.

Two-wire wave antennas for reception in reciprocal directions have been described by Laport and Misek (Fig. 1B).[9,10] This bidirectional wave antenna is unusual in that it simultaneously possesses two directivity patterns. The wave field impinges simultaneously upon the two wires, and equal currents flow in both wires in the direction of wave travel. These currents continue to flow until they reach the far end of the antenna, where reflection occurs by grounding one wire and leaving the other open-circuited. This balances the current received from the right, but has no effect on the unbalanced current received from the left. To obtain sufficient balances in the transformers, an electrostatic shield is required. The two receivers for simultaneous reception in the two reciprocal directions are matched to Z_{01} and Z_{02}. Z_{01} designates the characteristic impedance of two wires unbalanced to ground, and Z_{02} is the balanced characteristic impedance between the wires. Basically, the bidirectional wave antenna is an aperiodic antenna whose terminations have been transposed to the receiving end. This provides greater flexibility in controlling the antenna with simple switching and phasing circuits.

Comparison Between Theory and Experiment

Detailed measurements were made for Beverage antennas of various lengths and heights, with a typical length of 360 feet. Also, various grounds were tested. The received signal strengths were measured at

[1]Notes appear on page 27.

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several frequencies, and airborne transmitters were used for signal generation. An aircraft towed a vertically polarized transmitting dipole, called XELEDOP (Transmitting Elementary Dipole of arbitrary Polarization).[11]

To develop azimuthal patterns with XELEDOP, the antenna must be towed concentrically around the antenna under test (AUT) at various heights and distances. A cut of the vertical pattern can be obtained by flying across the AUT at a given height (1.9 miles, typically) along the boresite, correcting the results for spatial attenuation, and for the changing angle between the XELEDOP dipole and the AUT ray direction.

Several antenna pattern measurements were made using a dipole suspended from a balloon. These measurements were made on a radius of about 1600 feet from the AUT, with the dipole coming within a quarter wavelength of the ground. Some examples of the various results using both balloon and XELEDOP techniques are shown in Figs. 2, 3 and 4.

In Fig. 2, a comparison is drawn between theoretical and experimental azimuthal patterns measured by the XELEDOP technique at 18 MHz. The vertical pattern (corresponding to the azimuthal pattern in Fig. 2C) is shown in Fig. 3. Measurements were also done using a 1/4-wave ground-plane antenna as a reference to measure the gain of the Beverage ground wave.

The results of detailed measurements of the vertical pattern at the azimuth of maximum gain are illustrated in Fig. 4. These results deserve some discussion. In this test, the AUT was a Beverage pair (two elements of a rosette array); therefore, the theoretical gain should increase 3 dB over a single Beverage element. The balloon measurements for low elevation angles ($\psi < 3°$)clearly illustrate the response of the antenna to ground waves. The skywave lobe was a maximum at $\psi_N \sim 16°$.

Clearly, there is a consistent agreement between theory and experiment in the main lobe, but a fairly large discrepancy in the side- and back-lobe levels. Typically, the side lobes of theoretical patterns are 25 dB lower than the main beam, whereas measured values are normally only about 15 dB below the level of the main beam.

Theoretically Derived Parameters for Beverage Antennas

Since the agreement between experiment and theory was, except in some instances, reasonably good, it follows that theoretically derived parameters can be used with confidence in Beverage antenna design. Since theory reveals design trends better than experimentation, we shall consider the effects of length, height, frequency and ground conductivity on the Beverage antenna.

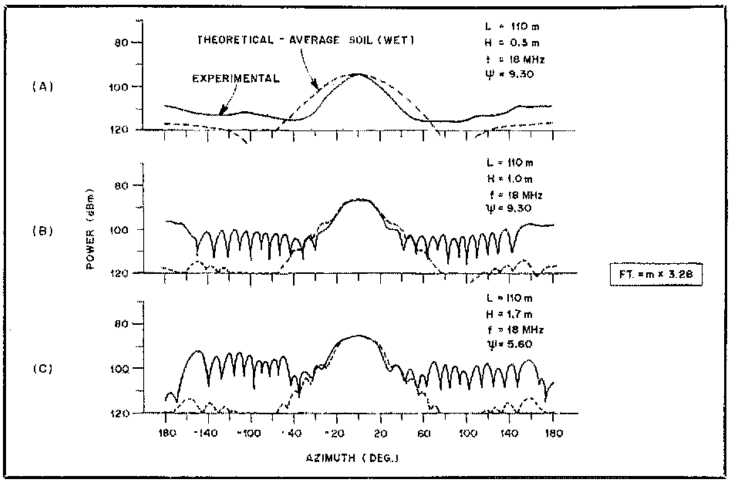


Fig. 2 — Comparison of theoretical and experimental azimuthal patterns for a Beverage antenna at 18 MHz.

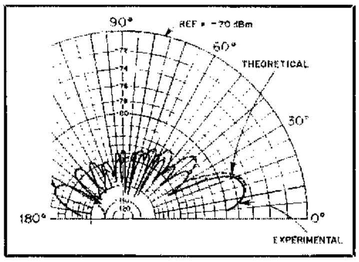


Fig. 3 — Comparison of theoretical and experimental vertical pattern in the bore-site direction (at 18 MHz) for an antenna length of 360 feet and a height of 4 feet (azimuthal pattern is shown in Fig. 2C).

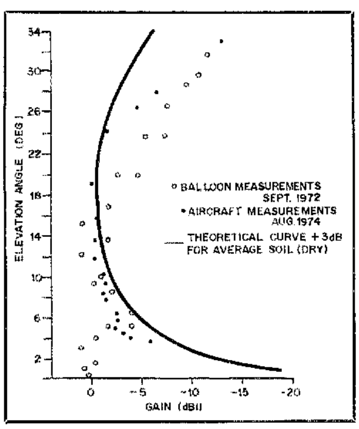


Fig. 4 — Vertical radiation pattern for a Beverage-pair antenna in azimuth of the main lobe at 9.5 MHz.

The length of a Beverage antenna should be greater than 1 wavelength, and since we are concerned with frequencies of 1.8 to 30 MHz, an antenna 656 feet long should give satisfactory performance at 1.8 MHz. Figs. 5, 6 and 7 are examples of design parameters — calculated values of gain in the main lobe (G_N), azimuthal beamwidth (BW_A), launch angle (ψ_N) and vertical beamwidth ($BW\psi$) for Beverage antennas 328, 656 and 984 feet long, all 6.5 feet high over average dry soil.

Figs. 8, 9 and 10 illustrate how the gain (G_N) varies with antenna length, antenna height and soil conductivity. Fig. 8 illustrates that the optimum length for a Beverage antenna mounted 6.5 feet above average dry soil is about 4 λ. In Fig. 9, it is shown that the gain increases at all frequencies as the height increases from 12 inches to 10 feet. The change is especially marked at the higher frequencies. The curves in Fig. 10 show the dependence of gain on earth conductivity. At 2 MHz, the gain decreases as earth conductivity increases, whereas at higher frequencies this variation has the opposite trend: Gain increases as conductivity increases.

Beverage Antenna Arrays for Transmitting Applications

While Beverage antennas are good for receiving because of their directivity and low noise, their efficiency is rather low. Compare, for example, the gain figures for the Beverage antennas discussed above with that of a quarter-wave antenna over a perfectly conducting earth, which is 5.16 dBi. For hf-receive applications, discrimination or directivity gain is the important factor, as the system signal-to-noise ratio is determined by ambient radio noise, not the internal-equipment noise as at vhf. Thus, a directive receiving antenna can be many decibels below isotropic reference and still be useful. Consequently, the lack of efficiency in the Beverage antenna is of no importance for receiving. During transmission, however, poor efficiency represents power wasted, since radiation efficiency is a measure of the power radiated divided by the power of the transmitter.

Beverage antennas can be operated as linear phased arrays, employing several closely spaced Beverage-antenna elements fed in phase. The power gain of such an array increases by about 3 dB each time the number of elements is doubled, provided the elements are independent. The effect of interaction among elements, which limits the efficiency achievable by such arrays, has been determined. For element heights on the order of 6.5 feet, the

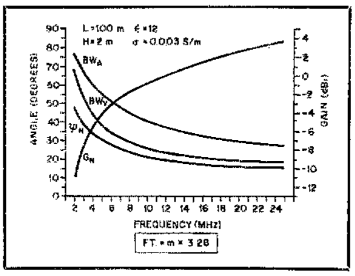


Fig. 5 — Design parameters for average dry soil: H = 6.5 ft; L = 328 ft.

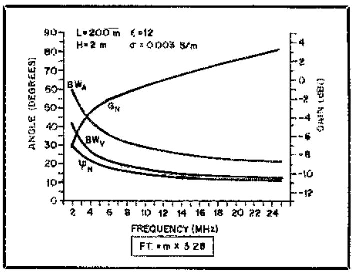


Fig. 6 — Design parameters for average dry soil: H = 6.5 ft; L = 656 ft.

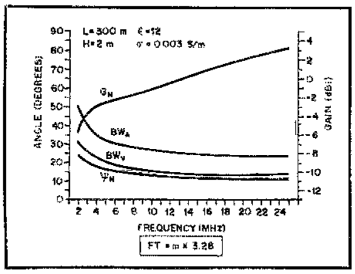


Fig. 7 — Design parameters for average dry soil: H = 6.5 ft; L = 984 ft.

interelement spacing can be as small as 20 feet. If there is sufficient land to mount a Beverage antenna in its length, only a modest increase in complexity could provide a power-gain increase of 3 or 6 dB by

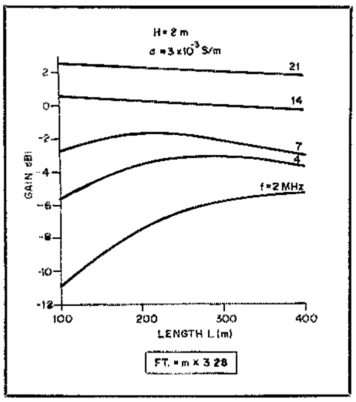


Fig. 8 — Gain of a Beverage antenna element as a function of frequency and length (H = 6.5 ft, average dry soil).

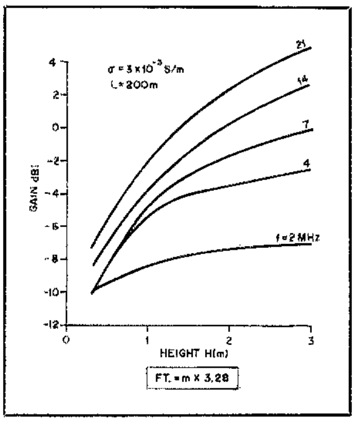


Fig. 9 — Gain of a Beverage antenna element as a function of frequency and height (L = 656 ft, average dry soil).

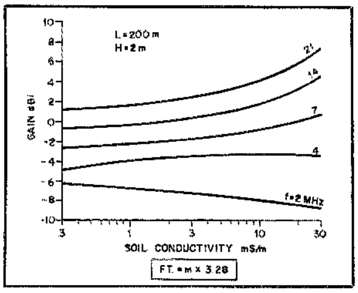


Fig. 10 — Gain of a Beverage antenna element as a function of frequency and conductivity (L = 656 ft; H = 6.5 ft).

employing 2 or 4 elements in a linear phased array.

An example of a Beverage array is shown in Fig. 11. It consisted of eight 500-foot elements mounted 6 feet above

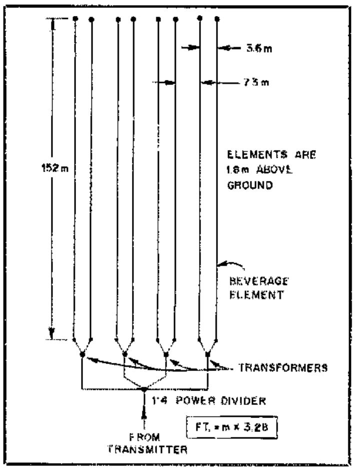


Fig. 11 — Plane view of an experimental Beverage array. See text for details.

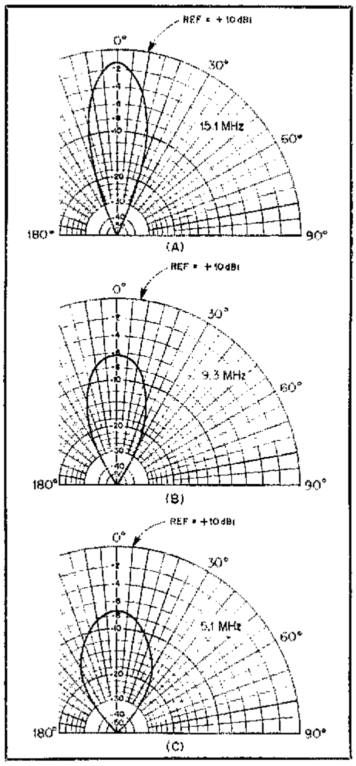


Fig. 12 — Measured azimuthal patterns for the experimental Beverage array shown in Fig. 11. Note that the 0 dB reference corresponds to + 10 dBi.

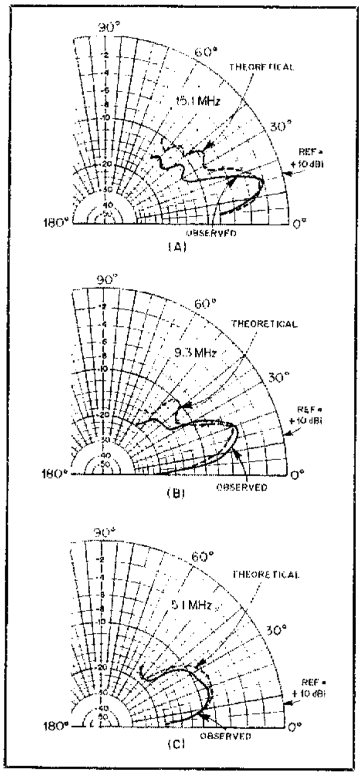


Fig. 13 — Measured and theoretical elevation patterns for the experimental Beverage array shown in Fig. 11. Note that the 0 dB reference corresponds to +10 dBi.

the ground, with the elements arranged in pairs. Each pair was coupled directly to a matching transformer, which was connected through a length of coaxial cable to a 1:4 power divider. The measured azimuthal and elevation patterns, expressed in dBi, are shown in Figs. 12 and 13, with those for 5.1, 9.3 and 15.1 MHz shown in A, B and C. The theoretical plot in Fig. 13 demonstrates the close agreement between practical results and theoretical calculations of the main lobe. The pattern results were repeated for summer (3-foot-high grass, damp ground) and winter (6-inch dense snow cover, partially frozen ground) with no change in the results.

Conclusions

The Beverage antenna and arrays of Beverage elements present an inexpensive alternative for transmitting and receiving on long-range, point-to-point communications circuits. The disadvantage, from a radio amateur's viewpoint, is that gain is realized in only one or two reciprocal directions and a large amount of real estate is required for a Beverage array on the lower hf bands.

Appendix

The Beverage antenna is analyzed as a transmission line in which current has been induced by the impinging electromagnetic field parallel to the Beverage wire. The characteristic impedance and propagation parameters derived (or measured) from transmission-line theory are therefore important elements in analyzing the performance of the antenna. The simplest formula for the characteristic impedance of a single wire over a ground is

$$Z_o = 60\ ln\ \frac{2h}{a} \qquad \text{(Eq. 1)}$$

where

h = height of the Beverage wire above ground

a = radius of wire

n = antenna current-wave propagation factor

Although a more complicated analysis of the impedance of a wire over lossy ground is discussed by Litva and Rook, the detailed analysis does not improve the agreement with the experimental data. The impedance and propagation characteristics of the transmission line can be determined experimentally by measuring the input impedance of the line with the far end open and short circuited; or, by measuring the current on the antenna by means of a small probe that is excited by an rf generator. The characteristic impedance is calculated from the open and closed circuit impedance measurement:

$$Z_o = \sqrt{Z_{oc} \times Z_{sc}} \qquad \text{(Eq. 2)}$$

where Z_{oc} = open-circuit line impedance at a given frequency (F_G)

Z_{sc} = short-circuit line impedance at F_G.

The input impedance is generally complex. The propagation constant $\gamma = \alpha + j\beta$ is another important parameter:

$$\gamma = \frac{1}{l} \tanh^{-1} \sqrt{\frac{Z_{sc}}{Z_{oc}}} \qquad \text{(Eq. 3)}$$

The real and imaginary parts of this parameter are the attenuation (α in nepers/meter) and the phase constant (β in radians/meter), which are needed to calculate the current wave that traverses the wire.

The velocity of propagation can be obtained from the current measurements, or from the open (or short) circuit impedance measurements, since the input is high (or low) when the electrical length of the line is an integral multiple of a half (or quarter) wavelength.

In Fig. 14, the average characteristic impedances of a 360-foot antenna about 4 feet (mean height) over average wet soil (for soil conductivities and types see Table 1) is shown

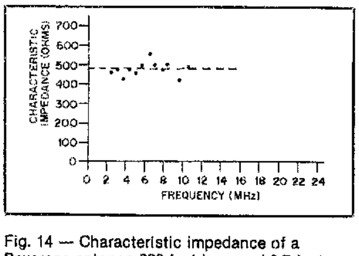


Fig. 14 — Characteristic impedance of a Beverage antenna 360 feet long and 3.7 feet high situated over average wet soil.

Table 1

Soil Conductivities and Types	*Conductivity S/m*	*Dielectric Constant*
Sea Water	5	81
Average soil (wet)	10×10^{-3}	10-25
Average soil (dry)	3×10^{-3}	10-15
Poor soil	1×10^{-3}	10
Poor soil (dry)	3×10^{-4}	8
Dry sand	1×10^{-4}	5

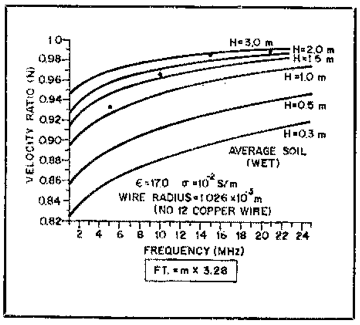


Fig. 15 — Theoretical and experimental values of current-wave velocity factor as a function of frequency for average wet soil (the experimental values were for an antenna height of 3.7 feet).

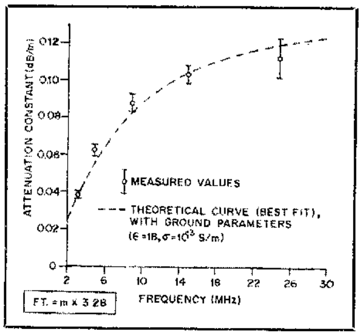


Fig. 16 — Attenuation as a function of frequency for a 6-foot-high Beverage antenna over poor soil.

for the range of 2 to 24 MHz. The mean value is 480 ohms, but clearly a single value for a terminating resistor will not result in absence of reflection from the end of the antenna. It should be noted that Z_o (calculated) equals 462 ohms. The Beverage wire was no. 12 copper-clad steel, for which a = 0.0043 in. Using Eq. 1:

$$Z_o = 60\ ln\ \frac{2\ (4.13)\ (100)}{0.0043} = 462 \text{ ohms} \qquad \text{(Eq. 4)}$$

In Fig. 15, we compare measured and calculated values for the current-wave velocity factor. The theoretical values have been calculated for various heights above average wet ground. The measured values were determined from current measurements. In Fig. 16, we show a comparison between theoretical and measured current-wave attenuation (decibels/meter) of another Beverage antenna, which appears to have a poor soil ground.

Theory of Operation

A Beverage antenna responds to vertically polarized waves in that it responds to the horizontal component of the vertically polarized ground wave, owing to the tilt of the wavefront, and to the sky wave, because of the tilt of the downcoming wave front. A vertically polarized ground wave at the surface of the earth will have a forward tilt, the magnitude of which depends on the conductivity and permitivity of the earth. This slight tilt forward, in the direction of propagation, is responsible for a small vertical downward component, sufficient to furnish the power dissipated in the earth over which the wave is passing, and it is the horizontal component parallel to the Beverage wire antenna that induces a current on it (Fig. 17A). This current flows in the direction of the wave travel, which is toward the receiver. All portions of the antenna collect energy from the impinging wave field in space, so long as the phase of the wave in the antenna does not differ greatly from the exciting field. The gain of a Beverage antenna, relative to an isotropic radiator, for surface or ground waves has been derived by Litva and Rook:

$$G = \frac{377\pi \sin^2 \delta}{Z_o \lambda^2} \left| \frac{1 - \varepsilon^{-\ell\Gamma}}{\Gamma} \right|^2 \quad \text{(Eq. 5)}$$

δ = tilt angle of surface wave

$$\delta = \tan^{-1} \left\{ \frac{\left[(\varepsilon_g - 1)^2 + \left(\frac{\delta_g}{\varepsilon_o \omega}\right)^2\right]^{1/4}}{\left[\varepsilon_g^2 + \left(\frac{\delta_g}{\varepsilon_o \omega}\right)^2\right]^{1/2}} \right\} \quad \text{(Eq. 6)}$$

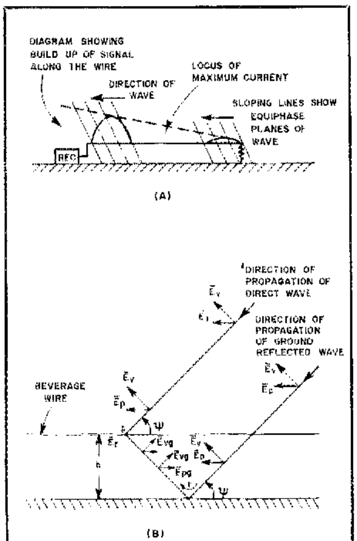


Fig. 17 — The wave or Beverage antenna illustrating: (A) current buildup along the wire for reception of ground wave; and (B) parallel voltage element induced in Beverage wire because of a vertically polarized downcoming skywave.

where
- $\omega = 2\pi f$
- ε_o = permittivity of free space
- ε_g = relative dielectric constant of the earth
- δ_g = conductivity of the earth
- $\Gamma = \gamma - j\beta_o \cos\delta$
- $\gamma = \alpha + j\beta$
- $\beta_o = 2\pi/\lambda$
- λ = free space wavelength
- ℓ = length of antenna
- α = current-wave attenuation on antenna (nepers/meter)
- $\beta = \beta_o/n$

For skywaves, wave tilt is provided by the arrival angle of the downcoming plane wave (see Fig. 17B), and the magnitude of the induced voltage will depend on the parallel component of the vertically polarized electric field. Again, this induced voltage will cause a current wave to traverse the wire, in the direction of propagation, toward the receiver. The report by Litva and Rook derives expressions for the skywave gain of the Beverage antenna and a computer listing program to calculate the necessary design parameters. Here, we consider the analysis in outline only. The magnitude of the induced voltage on the wire is calculated from the resultant field parallel to the Beverage wire $\bar{E}_r$, where

$$\bar{E}_r = \bar{E}_v \sin\psi \left(1 - \rho_v e^{-j\frac{2\pi}{\lambda}2h \sin\psi}\right) \quad \text{(Eq. 7)}$$

where ρ_v is the reflection coefficient for vertically polarized ground waves:

$$\rho_v = \rho e^{j\phi} = \frac{\sqrt{\varepsilon_c}\sin\psi - \sqrt{\varepsilon^2_c - \cos^2\psi}}{\sqrt{\varepsilon_c}\sin\psi + \sqrt{\varepsilon^2_c - \cos^2\psi}} \quad \text{(Eq. 8)}$$

where ε_c is complex and is given by:

$$\varepsilon_c = \varepsilon_r - j\frac{\delta_g}{\omega\varepsilon_o} \quad \text{(Eq. 9)}$$

where
ε_r is the ratio of the dielectric constant of the ground to that of free space
δ_g is the ground conductivity in mho/meter
ε_o is the permittivity of free space (8.85×10^{-12} farads/meter)

The negative sign in the equation for $\bar{E}_r$ indicates that the horizontal components of the direct and ground-reflected wave are oppositely directed in space.

Since $\bar{E}_r$ is parallel to the Beverage wire, a potential gradient results, and the voltage induced in the line can be calculated. This elemental voltage gives rise to an elemental current, and the total current, as a function of the elevation angle ψ and azimuth angle θ, is obtained by integrating over the length of the line. The resulting expression is a complicated function dependent on the transmission-line characteristics, the properties of the ground, the height of the antenna above the ground, and the length of the Beverage wire.

The derived expressions follow. The power gain of a Beverage antenna referred to that of an isotropic radiator is

$$P_G = \frac{4\pi\mu_o c \, | I_T(\psi,\Theta) |^2 \, \mathrm{Re}(Z_o)}{\lambda^2} \quad \text{(Eq. 10)}$$

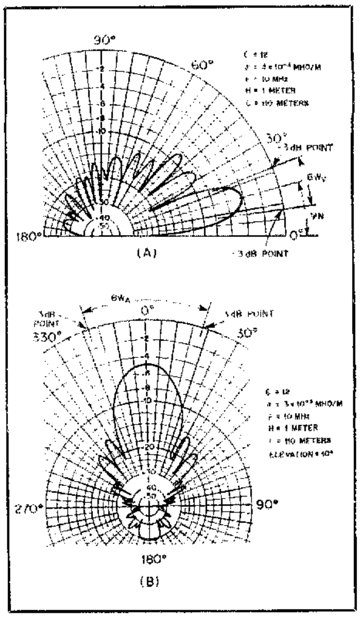


Fig. 18 — Theoretical elevation and azimuthal patterns for a Beverage-antenna element situated over average dry soil. The length is 360 feet and height is 3 feet.

(assuming E_v = unity)
where
- μ_o = permeability of free space (4×10^{-7})
- c = velocity of light
- Re (Z_o) = real part of complex characteristic impedance of Beverage wire
- $I_T(\psi, \Theta)$ = resultant current at receiving end of antenna, which is

$$I_T(\psi,\Theta) = \frac{E_r}{2Z_o} e^{-\frac{\gamma\ell}{2}} \ell \cos\Theta \times \left[\frac{\sinh\left[\frac{\gamma_1 \ell}{2}\right]}{\frac{\gamma_1 \ell}{2}} + P_L e^{\gamma\ell} \frac{\sinh\left[\frac{\gamma_2 \ell}{2}\right]}{\frac{\gamma_2 \ell}{2}} \right] \quad \text{(Eq. 11)}$$

where PL is the reflection coefficient, since in general Z_L will not equal Z_o at all frequencies:

$$PL = \frac{Z_L - Z_o}{Z_L + Z_o} \quad \text{(Eq. 12)}$$

- Z_L = terminating resistance
- $\gamma_1 = \alpha + j\beta(1 - N\cos\psi\cos\Theta)$
- $\gamma_2 = \alpha + j\beta(1 + N\cos\psi\cos\Theta)$

$$N = \frac{\mu}{c}$$

where
- μ = velocity of propagation of current on the wire
- c = velocity of light

A typical theoretical elevation and azimuthal pattern for a Beverage antenna over average dry soil is shown in Fig. 18.

Notes

[1]m = ft × 0.3048; mi = km × 1.6
[2]H. H. Beverage, C. W. Rice and E. W. Kellogg, "The Wave Antenna, a New Type of Highly Directive Array," *Trans. A.I.E.E.*, vol. 42, Feb. 1923.
[3]I. Herlitz, "Analysis of Action of Wave Antennas," *Trans. A.I.E.E.*, 1923, vol. 42, pp. 260-266.
[4]H. H. Beverage, "A Wave Antenna for 200 Meter Reception," *QST*, Nov. 1922, p. 7 (see update of 1922 article in *QST*, Jan. 1982, p. 11).
[5]J. R. Wait, and J. E. T. Mousseau, "Calculated Field Patterns for Horizontal Travelling Wave Antennas," Radio Physics Laboratory, Project Report No. 19-0-2, Jan. 15, 1953.
[6]D. N. Travers, P. E. Martin and W. W. Sherrill, "Use of Beverage Antenna in Wide Aperture High Frequency Direction Finding (Pt. IV Theory)," Interim Report for Contract NObsr-89345, Southwest Research Institute, March 23, 1964.
[7]P. E. Martin, D. N. Travers and R. Lorenz, "Circular Arrays of Beverage Antennas for High Frequency Direction Finding," paper submitted for technical program, Southwestern IEEE, April 1965.
[8]J. Litva and B. J. Rook, "Beverage Antennas for HF Communications, Direction Finding and Over-the-Horizon Radars," Communications Research Centre Report No. 1282, Ottawa, Aug. 1976.
[9]E. A. Laport, *Radio Antenna Engineering* (New York: McGraw Hill Book Co., Inc., 1952), pp. 55-60.
[10]V. A. Misek, *The Beverage Antenna Book* (Hudson, NH: V. A. Misek, 1977).
[11]C. Barnes, "XELEDOP Antenna Pattern Measuring Equipment, 2 to 50 MHz," Stanford Research Institute, Menlo Park, CA, July 1965.
[12]G. E. Moss, N. Muirhead and R. W. Jenkins, "The Use of Multiple-Element Beverage Antenna Arrays for HF Transmission," Communications Research Centre Report No. 1318, July 1978. QST

New Products

MOTOROLA BIPOLAR PROM

Motorola has introduced a 16-K Schottky TTL PROM, the MCM76161, with a 70 ns (maximum) address-access time. This fully decoded, high-speed, field-programmable ROM is organized as 2K, 8-bit words and is pin compatible with standard PROMs and ROMs. Three-state outputs and chip-enable inputs allow expansion to larger arrays.

The PROM is manufactured with a logical 1 (output high) at each location, and can be programmed selectively for logical 0 (output low) by raising the 5-V supply to 12-V with a rise time greater than 1 μs. The programming delay is typically 100 μs.

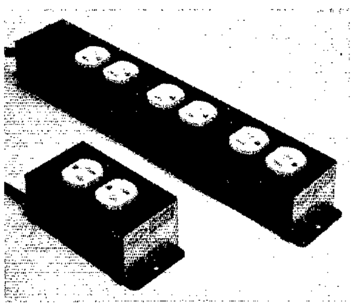
In a family of PROMs ranging in density from 2K to 16K, the MCM76161 is the highest-complexity product. Typical access times are 45 ns and 30 ns for address and chip-enable. The TTL-compatible outputs can sink 16 mA and source 2 mA. Supply current is typically 130 mA. In quantities of 100 to 999, the price is $18.85 each. For further information contact Jim Miele at Motorola Semiconductor Products Inc., P.O. Box 20912, Phoenix, AZ 85036. — *Paul K. Pagel, N1FB*

R. L. DRAKE MODEL LF2 AND LF6 LINE FILTERS

☐ These MOV-protected low-pass ac line filters may be of interest to both Amateur Radio operators and computerists. They are designed to restrict rf signals from traveling through the ac power line to and from equipment used.

The LF2 contains one grounded duplex outlet, while the LF6 has three duplex outlets. Each duplex outlet has its own hash filter and transient suppressor. The components comprising these filters are mounted on glass-epoxy pc board material and are attached securely to the outlet.

Transient protection is afforded by means of an MOV (metal oxide varistor) connected across the duplex outlet terminals. The General Electric V22OMA4B



MOV used is rated to handle voltage spikes having an 8 μs rise time and duration of 20 μs with a maximum spike current of 100 A. The rf line filters consist of ferrite-core inductors and disc-ceramic capacitors arranged in a T configuration on each side of the ac line.

Black-anodized aluminum enclosures are used for the filter assemblies. A no. 10 bolt is affixed to one end of the case. This grounding stud should be connected to a nearby station ground point. It is internally connected to the case, the duplex outlet ground and the ground wire of the ac line cord.

A lighted ON/OFF rocker switch on the LF6 indicates when power is supplied to the outlets. This filter is also supplied with a 15-A line fuse; the LF2 is unfused. A three-conductor ac line cord attaches each filter assembly to the ac source.

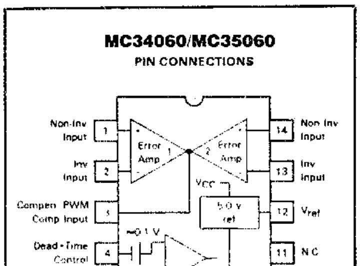
The LF2 measures 5.4 × 2.4 × 1.7 inches (LWD) and the LF6 is 16.2 × 2.4 × 1.7 inches (LWD) in size.[1] Price class of the LF2 is $40 and $70 for the LF6. They are available from the R. L. Drake Company, 540 Richard St., Miamisburg, OH 45342. — *Paul K. Pagel, N1FB*

[1]mm = inches × 25.4

MOTOROLA SINGLE-ENDED SWITCHMODE® CONTROL CIRCUITS

☐ A series of single-ended Switchmode® pulse-width modulation control circuits, MC34060/MC35060, has been introduced by Motorola. Designed specifically for Switchmode® power supply control, the MC34060/MC35060 is a low-cost, single-ended version of the Texas Instruments TL494 push-pull control circuit. Its electrical specifications are identical to the TL494 for all common characteristics.

These devices feature complete pulse width modulation control circuitry, adjustable dead-time control, and an uncommitted output transistor which can source or sink 200 mA. The MC34060/35060 has on-chip error amplifiers, a 5-V reference and an oscillator with master or slave operation. They're available in 14-pin plastic and ceramic packages with an operating temperature range of 0 to +70° C. The plastic MC34060 and ceramic MC34060 are $1.85 and $2.15 each, respectively, in quantities of 100 to 999. A full-temperature-range (−55 to +125° C) version is available in the ceramic package only (MC34060L) at $7.85 each in 100 to 999 quantities. For further information, contact Bob Benzer at Motorola Semiconductor Products Inc., P.O. Box 20912, Phoenix, AZ 85036. — *Paul K. Pagel, N1FB* QST



CATV Leakage: A Two-Way Street for Interference and Cooperation

Does it seem like you'll never win? Just when you thought you had the TVI situation in hand, the roles are reversed — ITV! Now you're the victim!

By Peter S. Carr,* WB3BQO

Interference has been a problem for radio amateurs since the early days. From the time of spark transmission to color television, amateurs often found themselves cast in the role of villain. Two recent developments have brought about a change in the amateur's role in interference cases.

The first development was the rapid growth of cable television (CATV). From a humble beginning in 1948, the CATV industry developed rather slowly until about 1972. At that time, changes in FCC rules, coupled with improved coaxial cables and CATV amplifiers, led the way not only to an increase in the number of CATV systems, but also in the number of channels used in the systems. Cable operators were originating programs locally and "importing" distant stations on microwave links. The mid-1970s ushered in the age of satellite communications for CATV operators. The rules and technology made more channels available, and the satellites delivered the programs to fill them. Today, the RCA SATCOM 1 satellite carries 24 channels of programming designed specifically for cable TV systems. The subscribers love it! Little wonder then that the number of CATV systems with over 29 channels more than doubled during 1980.

Coaxial cable shielding and the stronger signal voltage delivered to the TV antenna terminals by the CATV systems proved to be an advantage to the Amateur Radio community. Properly installed and maintained, cable systems meant a lot less TVI resulting from amateur hf operation. No one lamented the passing of corroded TV antennas and easily overloaded preamplifiers! So everything was looking better — right? Well, not quite.

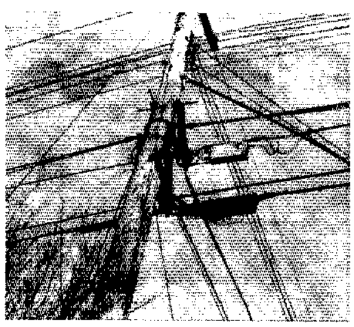


Fig. 1 — A typical utility pole with a CATV amplifier station mounted on a 1/4-inch steel support wire. Most CATV lines are mounted above the highest telephone cable and below the secondary electric wires.

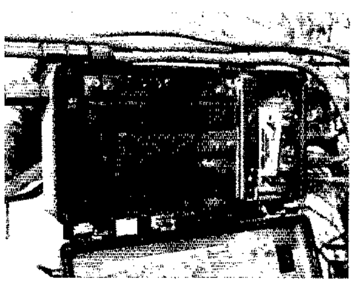


Fig. 2 — Interior of a CATV amplifier. Incoming trunk signals enter the housing at the upper left and exit at the top right of the housing. The vertical module is a high-level distribution amplifier that feeds four smaller cables going to houses in the immediate area. Both amplifiers are powered by 17-V dc, which is derived from an ac voltage that is fed down the coaxial cable.

Another development has occurred in the last decade: the explosive growth of amateur vhf-fm repeaters. Most parts of the country are served by at least one repeater, and a variety of choices is frequently the rule. A network of fixed, mobile and hand-held radios operating through a repeater has become the standard communication system for local emergency communications. The channelized operation of vhf fm, combined with telephone interconnection, has helped make this mode a favorite for portable and mobile operation. Today, the 2-meter fm transceiver, in its many forms, is the most commonly found piece of Amateur Radio communication equipment.

How has all this changed the amateur's role in interference? To provide all those extra channels, CATV systems legally use the full range of frequencies from 54 to 456 MHz — this includes amateur frequencies. Only a fraction of a microwatt is delivered to the subscriber TV antenna terminals. In the bigger, better shielded cable, power levels may approach 1 W. Even if a video carrier falls in the 2-meter band (145.25 MHz typically for cable channel "E") there should be no problem, at least theoretically.[1]

Leaks and Leakage

A corroded connection, a physical

*329 Little Ave., Ridgway, PA 15853

[1]Many new, larger systems use the HRC (harmonically related carriers) system. In that scheme, the channel E visual carrier is at 144.0 MHz. For a variety of reasons, carriers are sometimes slightly offset from nominal frequency. For a listing of CATV channels, see February 1982 *QST*, p. 14.

break or other damage to the system line or connectors, as well as poor quality components, can cause a signal leak. Where leaks exist, signals inside the cable escape and are radiated. (The lower the frequency, the more noticeable the radiation from the leak.) If the cable system uses channel E, there will be a 145.25-MHz visual carrier and sidebands being radiated in the vicinity of a leak. Roles are reversed as the radio amateur's reception is interfered with.

The role reversal is not complete, however. The leak that lets signals out of the cable also lets signals from outside enter into the cable; you might say it is a two-way street. The amateur hears interference in his receiver; when he transmits, it enters the cable and disrupts channel E reception. The neighbors complain! Will the cable company repair the leak?

Legal and economic reasons compel responsible CATV operators to repair leaks in their systems. Part 76 of the FCC rules requires CATV operators to check their systems for leaks at least once per calendar year. They are required to keep a log of leakage problems, causes and remedial measures taken. The rules (§76.610) limit allowable leakage in the 54- to 216-MHz range to 20 μV per meter at a distance of 3 meters (10 feet). In addition, harmful interference to other services is not permitted (§76.613).

For obvious economic reasons, the CATV operator wants to deliver high-quality signals to his subscribers. Leaks mean signal loss, and that means poor reception. Subscribers are not willing to pay for poor reception. That is especially true when the interference is to a premium program service. Channel E and the other mid-band channels in the 108- to 174-MHz range typically carry the movies and entertainment specials that provide added income to the cable company. Subscriber complaints are usually the first indication of a cable leak.

Not all cable systems use amateur frequencies inside the cable. If the system in your area does, and if you experience leakage problems, then you may want to know how to locate leaks on your own.

Begin by enlisting the aid of a couple of ham friends with rotatable antennas. Use the method of triangulation to narrow down the leakage location to a few square blocks. Place a battery-operated TV set in a vehicle and a pair of "rabbit ears" on top. Select a channel that is used in the cable but is not used for broadcasting in your area. With one person to drive and another to watch the TV screen, take a trip through the suspect area.

As you begin to approach a fault location, the first indication will be faint sync lines running either vertically or horizontally across the screen. As signal strength increases, faint images will appear, then fade, roll and return; this is caused by multipath signals reaching the antenna. When the picture stabilizes and becomes snow free, it indicates that the leak is very close. A return to snowy pictures indicates that the leak has been passed.

For ease of maintenance, most all cable equipment is mounted on the steel support wire close to a utility pole. The large majority of leaks that occur in the aluminum cable system center near these devices. For that reason, you should locate the pole nearest the spot where snow-free pictures were received. You can inspect the area from the ground, if you wish. For safety and liability reasons, make no attempt to climb the pole for a closer look. Copy the identification number on the pole and report your findings to the cable company.

Fig. 3 — Signal-level meter used to adjust amplifier levels. Continuously tunable from 54 to 300 MHz, the meter can be used for testing mid-band and super-band channels. It is battery powered and can be used with a small antenna to isolate cable leaks. Sensitivity is not great because the meter has broadband response.

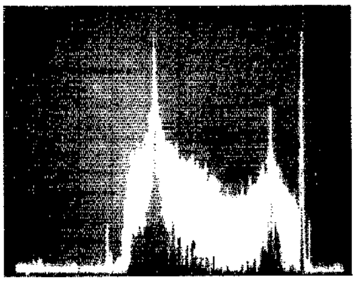


Fig. 4 — Spectral display of a channel E television signal. Horizontal divisions are 1 MHz; vertical divisions are 10 dB. At 145.25 MHz, the visual carrier (on the left) is the highest energy-level component of the composite TV signal. Notice that the chroma subcarrier and the aural carrier (on the right) are both outside the 2-meter band.

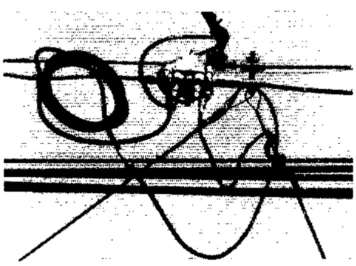


Fig. 5 — Directional tap unit installed in the aluminum distribution cable. Smaller drop lines are connected to the tap; the one at the left has a pay TV trap installed in series. Each connector may allow water to enter the cable, corroding the shield and causing radiation leakage.

A hand-held transceiver can be used to trace high-level interference to 2-meter communications. Remove the antenna and tune to a frequency around 147 MHz. You should be able to pick up sync and video sidebands when you are near a leak. With the antenna connected, the transceiver becomes a very sensitive leak detector. The sound carrier is 4.5 MHz above the video carrier. Less ambiguity will result from tuning to the sound signal, when possible.

Other Observations

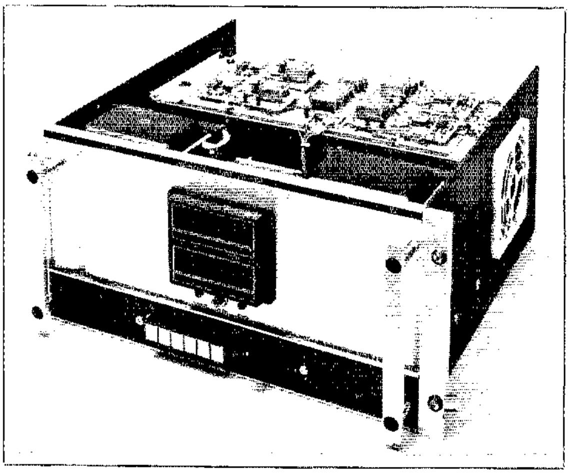
Not all leaks are located along the main cable. Cuts, breaks and corrosion on the subscriber drop cable are other sources of leakage. The good news is that the intensity will be less because signal levels are lower there. The bad news is that it can be hard to locate the source when several drop lines leave the cable from a single point.

Faults that are not located along the main cable near the poles are found mostly inside the subscriber homes. Lengths of twin-lead used to connect other TV sets or fm stereo receivers to the system are common offenders. Improper connection of video games and video cassette recorders will frequently cause leaks and will usually result in poor reception as well. I even heard of a fellow who connected his outside antenna in parallel with the cable at the TV set antenna terminals. Now that was a leak!

In the past, the amateur has looked upon CATV as a friend that offered a measure of protection against RFI to consumer electronic devices at no cost to the amateur. Today, it is time for hams to work actively with cable system technicians, a large number of whom are also hams, to control the CATV leakage problem. Each Amateur Radio operator is responsible for the proper operation of his station — emitting clean signals, reducing harmonics, using only enough power to communicate effectively and comply with the other FCC rules. CATV systems must also live up to the intent, not just the letter, of the FCC rules. Cooperation among all concerned will bring about the greatest benefit to all. QST

Peter S. Carr has been a licensed radio amateur since 1974, upgrading to Advanced class in 1978. His electronics background includes radar, telephone and cable-television systems operation and maintenance. Since 1972, he has been employed as chief technician and manager of Tele-Media, Inc., of Ridgway, Pennsylvania — a 3400-customer cable TV system. Pete also maintains the W3IE 2-meter repeater (the antenna is halfway up the 500-ft CATV tower). He has written a number of articles for Amateur Radio and R/C modeling magazines.

MOSFET RF Power — An Update



Part 2: The general application information provided in this series offers hard proof that power FETs have graduated beyond infancy. This concluding installment describes filtering methods and other modern amplifier concepts.

By Helge Granberg,* K7ES/OH2ZE

In Part 1 of this article (December 1982 *QST*), the author described a number of design and performance features for his 2-30 MHz, broadband MOSFET linear amplifier. Part 2 provides a wrap-up and offers a variety of design principles that can be applied to other solid-state power amplifiers.

Output Low-Pass Filters

Chebyshev five-pole, constant-k low-pass filters are used in the amplifier output (Fig. 5). Shunt elements have been added to provide an elliptic function. These elements usually are designed to resonate with the filter inductances at the harmonic or other undesired frequencies. This provides notches that give much higher close-band attenuation than can be obtained from a simple five-pole filter. The filter skirt is also much steeper, but at the cost of degraded far-band attenuation. However, in an application such as this in which the filters are used at a multiplicity of frequencies, they can't be optimized in this respect. The worst case (−35 dB) for the third harmonic (which is of the most concern) is with the low-frequency filters. This probably is caused by the low Q of the inductors.

The total harmonic attenuation at 7 MHz and below is approximately 50 dB or better, which meets the FCC requirement of 50 mW maximum for spurious emissions. The 1.8 to 30 MHz range can't be covered by the six filters without resultant gaps; a minimum of eight filters with sharper cutoff characteristics would be required for continuous coverage.

Filter Construction

Each filter is assembled on a separate pc board of similar layout. Only the component values are different. The most critical parts are the capacitors — often a stumbling block for practical high-power filter design. At these power levels, they must withstand peak rf voltages up to 800 (even higher in the event of a mismatch). The rf-voltage rating of ceramic capacitors is only some 30% of the nominal dc value. Hence, capacitors with ratings of 2000 to 3000 V are required. Also, they must be able to handle rf currents up to several amperes. These filters contain several inexpensive disc-ceramic capacitors in parallel. This provides current capacity and permits capacitor combinations that yield nonstandard values, as needed.

The distance from each filter input or output terminal is 2-1/2 inches to the corresponding terminals of the adjacent filter. This amounts to a distance of 12-1/2 inches from the first to the last filter.[5] Sections of 50-ohm strip line (Z1, Z2) are used to connect the T-R switch and the power amplifier to the filters. These are made of copper-clad pc-board material that is attached to the chassis below the filter boards. The lowest-frequency filter is located nearest to the input and output ends of the line, while the highest-frequency filter is at the far end. Thus, when the filters are switched, the remaining piece of unterminated line causes minimum VSWR in each case. Although the filters were tested individually with the amplifier, L1, L4, L7 and L10 had to be added to restore the filter characteristics in the composite circuit. This was necessary because the mechanical arrangement and the added capacitance from attaching the relay contacts to the 50-ohm lines affected the filters.

Filter-Relay Driver

Each relay field coil is driven by the circuit of Fig. 7. It is basically a programmable BCD-to-decimal decoder. The gates were added to accommodate the manual-switching feature. When the BCD input plug is disconnected, the manual switch is activated automatically. MMH0026 MOS clock drivers (U10-U12)

*Motorola Semiconductor Products, Inc., Phoenix, AZ 85062

[5]Notes appear on page 33.

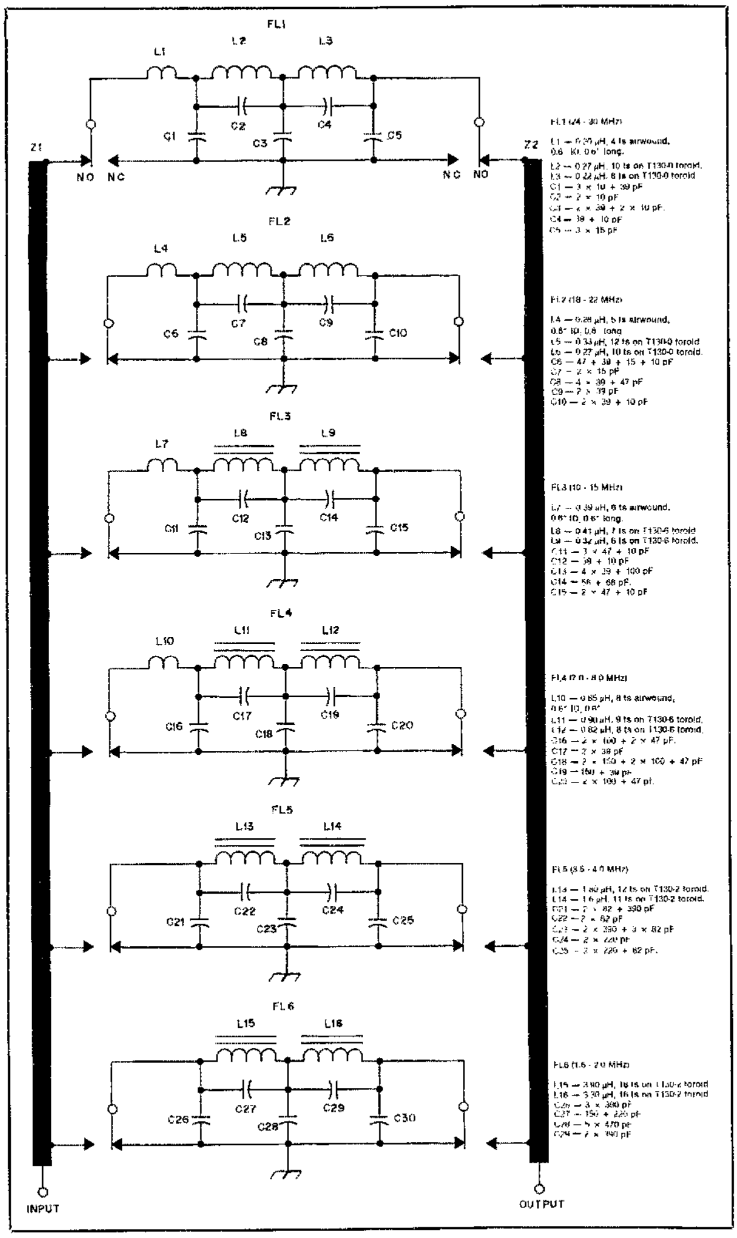


Fig. 5 — Low-pass filter circuit and component values for 1.6-30 MHz operation. Standard capacitor values have been placed in parallel to obtain nonstandard values. Toroid cores are Amidon, Palomar or Micrometals Corp. powdered-iron units. The designator "×" means "times," E.g., 2 × 47 + 10 pF means two 47- and two 10-pF capacitors, all in parallel to provide 114 pF. Wire type for all inductors is no. 14 enameled. Capacitors are RMC 3-kV disc ceramic, except those units that are 390 pF and higher, which are RMC 2-kV units. Relays are Deltrol 20693-83/405 with 12-V coils.

are used because they can be operated directly from a +12-V supply. Also, they have internal clamping diodes for suppressing inductive spikes.

T-R Relay and Timing Circuit

Fig. 8 shows the T-R relay driver and timing circuit. The input-output timing of the T-R relays must be precise. If the output relay is switched with full rf power, the relay lifetime will be very limited. Thus, the output relay must be switched on before the input relay is. Similarly, the output relay has to be switched off after

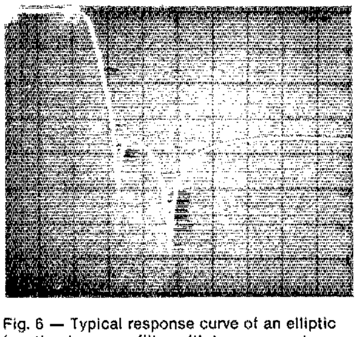


Fig. 6 — Typical response curve of an elliptic function low-pass filter with two resonant shunt elements.

the input one is deactivated. Simple R-C and diode networks are used to generate the delays.

A CMOS hex inverter performs the amplifier pulse-shaping functions. The upper channel, which controls the input relay, delays the leading edge of the keying pluse and shortens it by a few milliseconds. The lower channel controls the output relay, and delays the trailing edge of the keying pulse, which is lengthened from the original as shown in the timing diagram. Both delays are adjustable from 2 to 12 ms by means of R2 and R4.

Other Circuitry

The circuit in Fig. 9 performs the following functions:

1) Output VSWR detection.
2) Linear alc control.
3) Alc shut down.
4) Bias switching between standby and operate.
5) Bias temperature tracking.

The VSWR reflectometer senses the reflected power caused by load changes (50 ohms nominal).[6,7] Capacitor C is approximately 1 pF in value and must be capable of handling high rf voltages. It can be made from a piece of coaxial cable, such as RG-58/U, with the braid stripped to a length of 0.4 inch. The section of center conductor can be used as the primary of T1 and for connections to the amplifier output and relay terminal.

The rf meter (normally seen in place of D4) is replaced with a diode for rectifying the rf energy. The resultant voltage is filtered by means of C6 and R31. The level of 0.2-0.3 V for a 50-ohm load. It increases to about 1.5 V for 3:1 VSWR. This voltage is used for the linear alc function and is fed to the alc amplifier, an LM307. Controls R26 and R30 can be set, for example, so that an output VSWR of 2 will have no effect, but a 3:1 condition will reduce the power output by 3 dB or more. The alc shutdown operates also from the dc developed in the VSWR bridge. During a complete mismatch, such as an open load, the instantaneous voltage is greater than 10, thereby saturating the

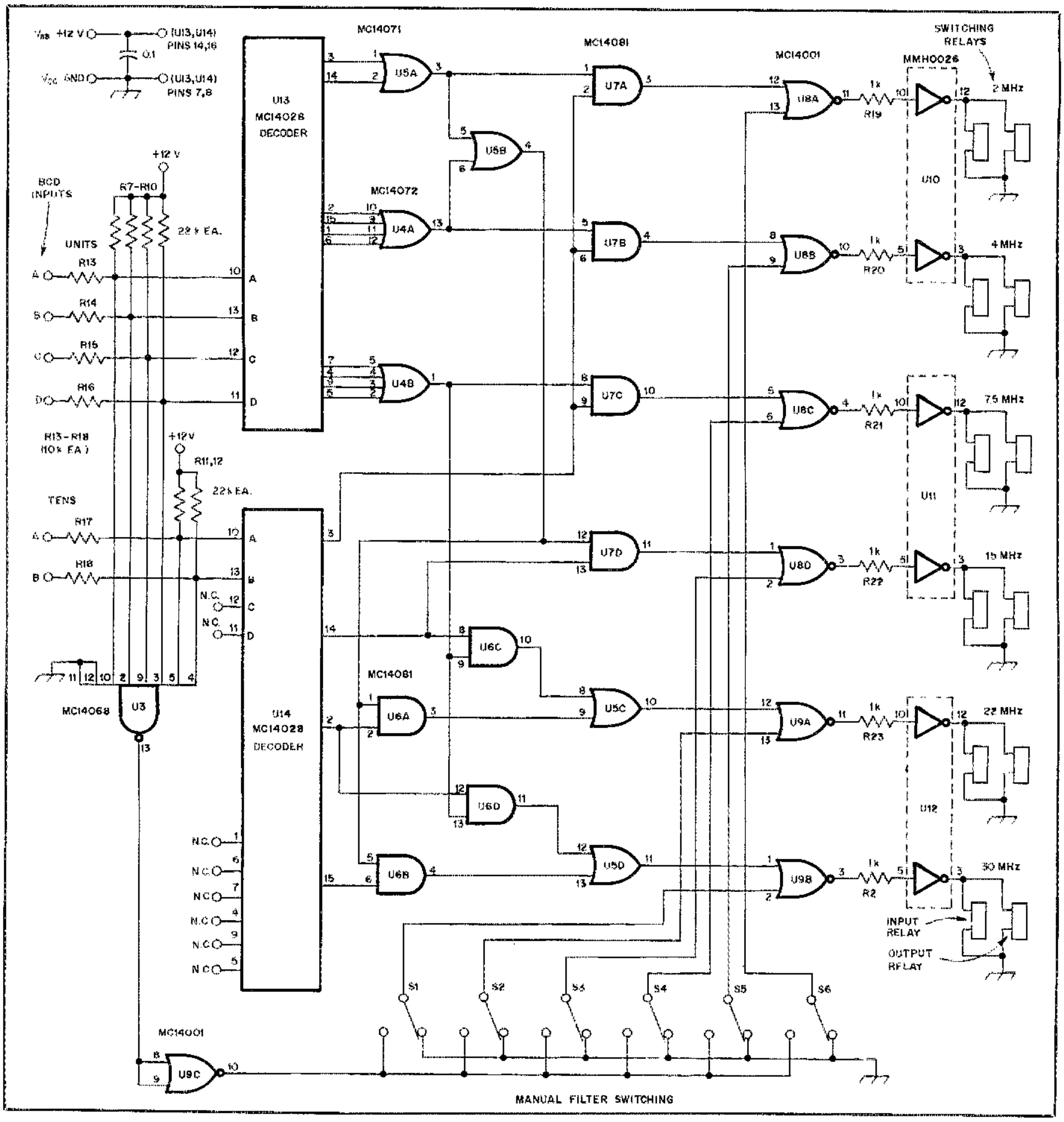


Fig. 7 — Output filter select and drive circuit indicator lights are connected in parallel with the relay coils, which are at the far right of the diagram.

LM307. After the alc loop has reacted and the power output has been reduced by a factor of 10, the voltage settles to 5 or 6 and the alc amplifier remains saturated. This holds the output power at the reduced level.

Automatic bias switching is initiated when a ground to +12-V signal is brought from the T-R relay driver to the same (−) input of the LM307. This overrides the control signal coming from the VSWR sensor and, when positive, turns the bias completely off. D5 makes this independent of the alc function.

Each of the rf amplifiers (Fig. 4, Part 1) and their associated circuits draw approximately 15 mA of bias current. The combined current is 120 mA, which is too much for the alc amplifier to handle. Therefore, a TIP31 has been added for use as a buffer/driver. Its input voltage is also controlled by a thermistor (R34), which is connected thermally to one of the main heat sinks. R34 controls the idling current of the power FETs by lowering the bias voltage during periods of elevated temperature, and vice versa. With bipolar transistors, this is normally done with forward-biased diodes, in which the diode voltage drop versus temperature closely follows that of the base-emitter junction. Both types of transistors have a positive temperature coefficient (for a constant bias current or voltxge, the idling current increases with temperature). Although the g_m (transconductance) of a MOSFET

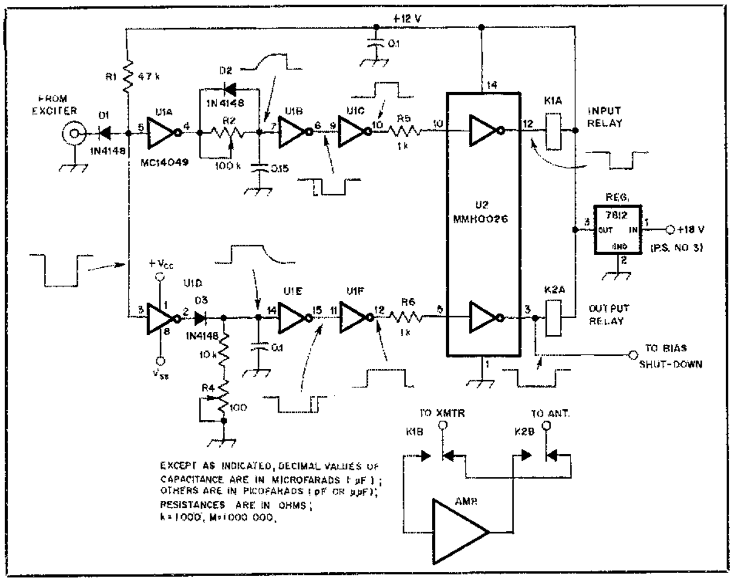


Fig. 8 — T-R switch drive and timing circuit. The approximate wave forms for the input and output functions are shown near the related ICs. Fig. 9 provides details of the relay connections.

decreases under these conditions the gate threshold voltage goes down, and usually has a dominant effect. There is a reversing point at higher drain currents. For the MRF150, this is around 5 A. The exact value depends on the g_m and the initial temperature. This is why power MOSFETs do not go into thermal runaway, provided the dissipation ratings are not exceeded.[4]

Conclusion

Although solid-state, high-power rf amplifiers are not in competition with vacuum-tube units, an increasing number of them appear on the commercial market regularly. Their feasibility versus output power is a question to be answered, at least until high-voltage rf transistors are available.

Power FETs are high-voltage, low-current devices, and some switching transistors for use up to several hundred volts are now on the market. In rf applications, however, several problems exist. Because of the internal structure of a transistor, rf arcing occurs (internally) easier at high voltages because the impedance levels are high. Transistor packaging techniques must be improved for this reason, and also for thermal considerations. The vertical-channel power FET (VFET) technology is fairly new and will certainly undergo many advances in the years to come.

This amplifier has been field-tested since late 1981, at the author's home. Switching from band to band is an absolute delight. Since no tuning is required, it should be a contest-operator's dream! Thus far, the only failures have been a jammed thermal switch that operates the fans, plus a burned low-pass filter board, caused by a loose piece of solder.

If you haven't worked with power FETs thus far, perhaps it's time you became involved, but maybe on not so grand a scale as is represented by this paper. Certainly, there are many advantages in the use of power FETs over bipolar transistors.

Notes

[1]mm = in. × 25.4
[2]*The Radio Amateur's Handbook*, 59th edition (Newington: ARRL, 1982).
[3]W. Orr, *Radio Handbook*, 18th edition (Indianapolis: Howard W. Sams & Co.).
[4]MRF150 data sheet, plus appropriate linear IC and CMOS data sheets.

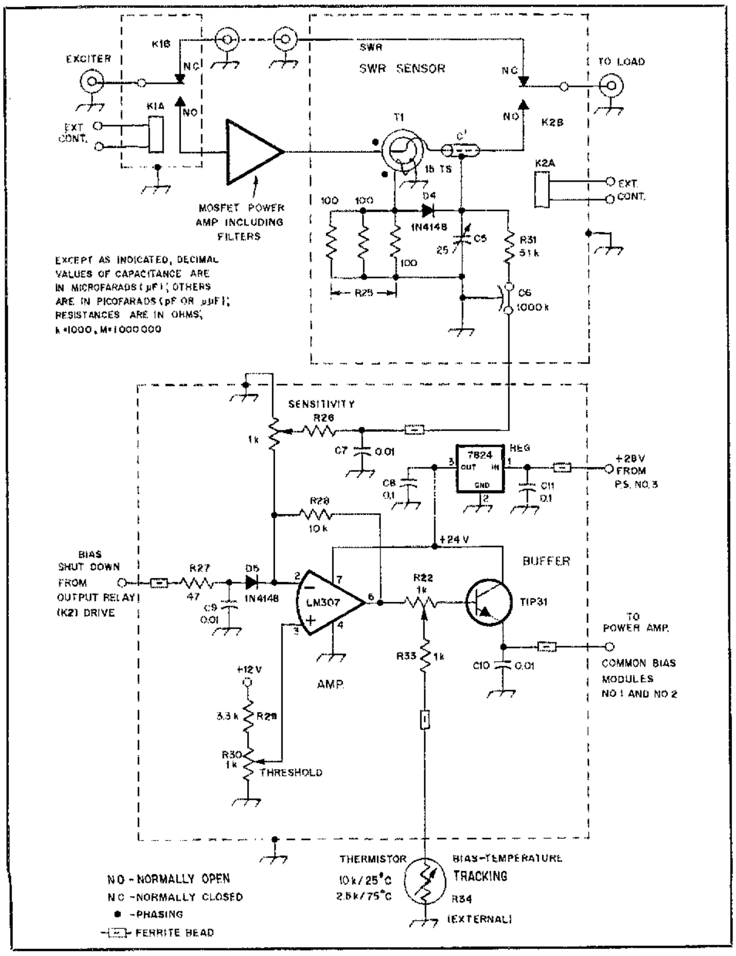
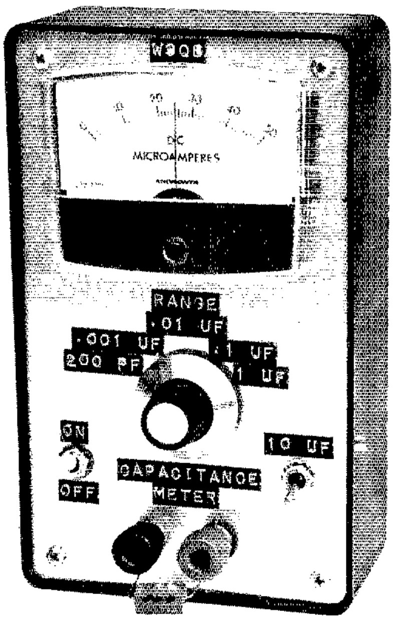


Fig. 9 — VSWR sensor, alc and bias circuits. Note ferrite-bead chokes in all alc input and output lines. They help prevent rf energy from getting into the alc amplifier.

A Simple Capacitance Meter You Can Build



This weekend project can save you time and money!

By Harry M. Neben,* W9QB

How many capacitors do you have that seem useless because they are marked in some strange hieroglyphics? If you're like most experimenters, you probably have a sizeable collection of such unknown capacitors. Here is a simple-to-build meter that you can use to decipher those strange markings and quickly find the right capacitor for the job at hand.

This meter has a reasonably linear scale and has sufficient accuracy to satisfy the needs of most experimenters. The scale linearity is approximately ±5%, and the accuracy is about the same. So the purists can stop reading now! If you are looking for a "ball park" capacitance value, however, this inexpensive meter will do the job.

The Circuit

The capacitance-meter circuit (Fig. 1) is similar to one described in *Test Equipment for the Radio Amateur.*[1] Modifications were made to improve the calibration and the scale linearity, however.

Six capacitance ranges, from 200 pF to 10 µF full-scale, are provided on the meter. While experimenting, I found that the characteristics of the 555 IC used in this circuit varied with temperature and from one IC to another. This resulted in problems when using a 100-pF, full-scale range. As a compromise, I selected 200 pF full-scale as the lowest range. The other ranges are in decimal relationship to one another.

To improve the meter accuracy, range adjustment trimmer resistors were included on the 200-pF and the 0.001-µF ranges. Fixed-value resistors may be substituted for these trimmers if you can tolerate the decreased accuracy.

In this circuit, the 555 is used in the monostable mode. One side of the unknown capacitor is switched between the positive and negative supply terminals by the 555 at a rate determined by R_A, R_B and C1. When connected to the negative terminal, the capacitor charges to a value near the supply voltage. When switched to the positive terminal, the capacitor discharges. M1 indicates the average discharge current value. It can be shown that

$$I_{ave} = \frac{V \times C_X}{(R_A + 2R_B)\,C1} \times K \qquad \text{(Eq. 1)}$$

where

V = voltage to which C_X is charged

K = a constant, depending upon the IC characteristics and the charge and the discharge time of the 555 circuit.

In this formula the internal resistances of the 555 IC are included in R_A and R_B. Therefore, I found it advisable to use trimmer potentiometers for meter calibration on the two lowest ranges.

Because the calibration is voltage-sensitive, a 6.2-V Zener diode is used to provide a constant voltage to the measuring circuit. While this places an additional current drain on the battery, the total current required is less than 10 mA, so the battery has a fairly long life.

Construction

All the components for this meter were purchased as stock items from a local parts store. [A complete parts kit is available from Circuit Board Specialists, P.O. Box 969, Pueblo, CO 81002. — Ed.] There are no specialty components to frustrate the builder. The meter is a 50-µA unit. All fixed-value resistors are 1/4-W, 5% carbon types. As the value of C1 determines the range and the accuracy of the meter, this capacitor should be as stable as possible and, therefore, of the highest available quality.

Front-panel layout is conventional, with the meter, the range-selector switch and the measuring terminals in line from the top to the bottom of the panel. The OFF/ON switch is mounted to the left of the range-selector switch and the 10-µF range switch is mounted to the right. The IC and most of the other components are mounted on a small piece of perf board or on the range switch.

Before completing the meter wiring, it is well to determine the resistance

[1] H. L. Gibson, *Test Equipment for the Radio Amateur*, 2nd. ed. (London: The Radio Society of Great Britain, 1978), pp. 7.5-7.7.

*151 Fairway Dr., Dunedin, FL 33528

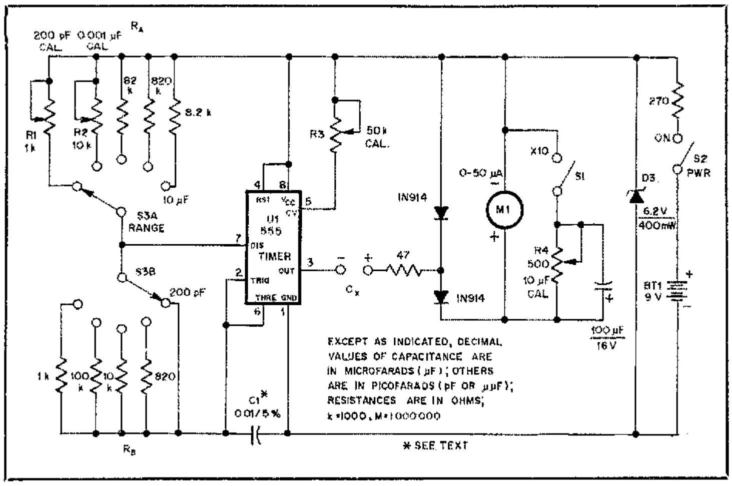


Fig. 1 — Direct-reading capacitance-meter schematic diagram. Resistances R_A and R_B discussed in the text correspond to the resistors selected by the range switch (S3). S3A selects the value of R_A, while R_B is selected by S3B. All fixed resistors are 5%, 1/4-W carbon types. Polarized capacitors are electrolytic. Numbered components not listed below are for text reference only.

BT1 — 9-V transistor-radio battery.
R1 — 1-kΩ pc-mount trimmer.
R2 — 10-kΩ pc-mount trimmer.
R3 — 50-kΩ pc-mount trimmer.
S1, S2 — Spst toggle switch.
S3 — 2-pole, 5-position rotary switch.
U1 — 555 timer IC.

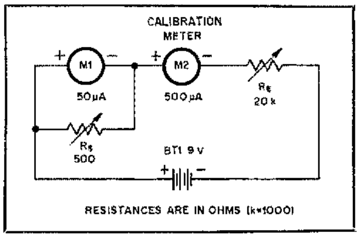
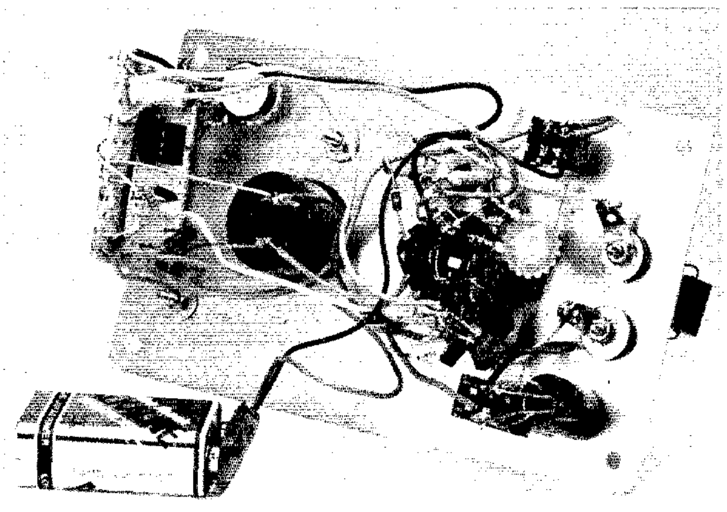


Fig. 2 — Suggested circuit for determining the value of the 10-µF range shunt resistor. After R_S is set to the correct value, it is used as R4 in the capacitance-meter circuit. The procedure for setting R_S is discussed in the text.



This interior view of the capacitance meter shows the parts placement used by the author.

necessary to produce a 10-µF range. This resistance (R4) shunts the 50-µA meter, increasing the full-scale reading to 500 µA.

To establish this resistance, another meter, such as a VOM (volt-ohm-milliammeter), must be used. Use the suggested circuit (Fig. 2) to determine this resistance. Set R_S to the lowest possible resistance before you start this procedure, to avoid damaging M1. Connect R_S across the terminals of M1. Connect this combination in series with the calibrating meter, R_E and the battery. Adjust R_E until the calibrating meter reads 500 µA. Adjust R_S until M1 indicates 50 µA. Readjust R_E until the calibrating meter again reads 500 µA, and adjust R_S so that M1 reads full scale. You may have to repeat this procedure several times to obtain a full-scale M1 reading when the calibration meter reads 500 µA. Once R_S is set to the correct value, it can be used as R4 in the capacitance-meter circuit (Fig. 1). Now the wiring of the capacitance meter may be completed.

Calibration

You can calibrate the meter using capacitors you have on hand. The more known-value standard capacitors you have, the easier the job is to do. Yet, you can calibrate this meter with only a 0.01-µF capacitor. Well, almost!

Begin calibration by placing the range switch in the 0.01-µF position. Attach your most accurate 0.01-µF capacitor to the C_X terminals. Turn on the meter and adjust R3 so that M1 reads full scale. This adjustment calibrates the 0.01 µF, the 0.1 µF, the 1.0 µF and the 10-µF ranges.

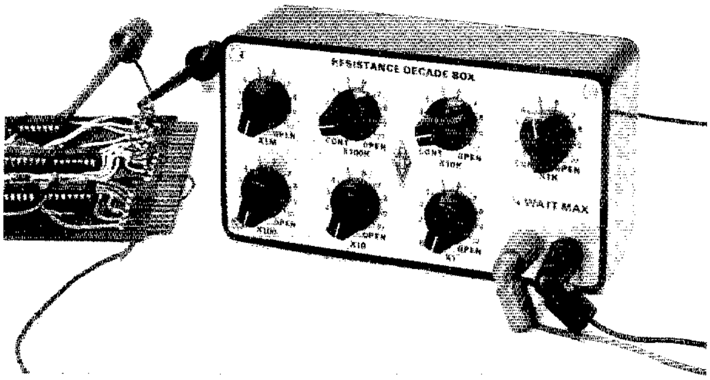
To calibrate the 0.001-µF range, leave the 0.01-µF calibrating capacitor connected to the C_X terminals. Place the range switch in the 0.001-µF position and place the 10-µF range toggle switch in the 10-µF position. Adjust R2 until M1 reads full scale. When the 10-µF range switch is returned to the OFF position, the meter will read 0.001 µF full scale.

The 200-pF range is calibrated by first selecting a 200-pF or less (preferably 200-pF) capacitor. Measure this capacitor using the 0.001-µF range. Place the range switch in the 200-pF position and adjust R1 so that the meter reads the value you found while using the 0.001-µF range. This completes the calibration.

Remember, the meter scale reads 0 to 50. So, unless you redraw the scale, you must use a scale factor on each range. This should not be a problem for most experimenters. I think you'll find that using this meter will make your next search through that sea of unmarked and unknown capacitors *much* less frustrating.

BEGINNER'S BENCH

Down Through the Decades



This is one project we think you'll find hard to resist!

By Bob Shriner,* WAØUZO and Paul K. Pagel,** N1FB

How much would 11,111,110 1-ohm, 1/4-W resistors weigh? Think you could pick up the pile? Sure you can! Fact is, you can (in effect) do it with one hand — if you've built the unit we're about to describe.

A Black Box

Experimenters and troubleshooters often need to substitute resistor values in a particular circuit during the course of bench work. While you could continuously insert and remove single resistors in the process of searching for the correct value, it would be tiresome and time consuming to do so. A much better approach is to use a resistance decade (substitution box). With this "black box," you can select a vast number of resistance values by simply rotating one or more switches. A pair of clip leads connected between the box and the circuit at hand makes resistance substitution an easy matter. Such a unit is shown in the accompanying photographs. The circuit appears in Fig. 1.

Circuit Description

A set of 12-position rotary switches is used to progressively select one or more of a number of fixed-value, 1/4-W resistors from each of seven decades. The CONT position of each switch provides a means of bypassing all of the resistors in that particular decade. Note that each unused decade must have its related switch set to CONT to provide continuity between the binding posts and the selected resistor(s). If any one of the switches is placed in the OPEN position, the resistance is effectively disconnected from the binding posts.

As the resistors are selected, they are placed in series with one another and across the binding posts (J1, J2). Clip leads, or a pair of leads equipped with banana plugs at one end and alligator clips at the other, connect the chosen resistance value to the circuit. Simple, isn't it?

Assembly

Gather the required parts first.[1] You must ensure that the enclosure you intend to use has enough height to accept the full length of the switches behind the panel without crushing the resistors attached to the switch lugs. The box we used measures 2-1/4 × 3-1/8 × 5-7/8 inches (HWD).[2] Sufficient panel and box space are available without component crowding.

The aluminum panel supplied with the box has been relegated to the "junkbox." A piece of 1/16-inch-thick pc board is used instead. (Although the pc board we used has copper on both sides, this is not a requisite. We used what we had on hand.) This material afforded greater structural strength than the flimsy piece of aluminum, and permitted us to etch the control markings indelibly on the top surface (front panel) of the unit. Of course, rub-on lettering or even tape labels could be used, but the aesthetics of the panel shown proved more appealing. Once the panel has been drilled and labeled, a coat or two of clear acrylic or polyurethane spray may be applied to add to the durability.

When the front panel is finished, mount the switches and binding posts. Make sure

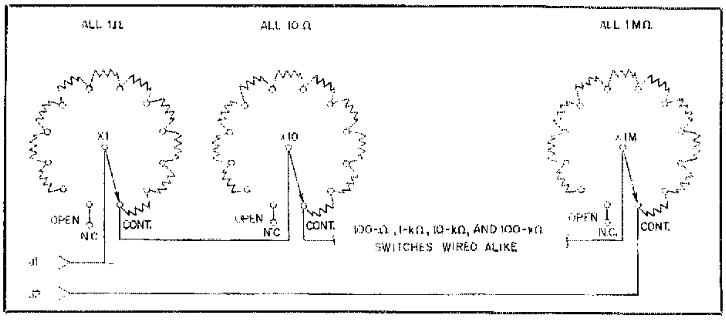


Fig. 1 — Circuit diagram of the resistance decade box. All switches are similarly wired. Ten resistors of equal value are soldered to the lugs of their respective switches.

R1-R70, incl. — 10 each of the following values of 1/4-W, 5%-carbon composition or film resistors: 1 Ω, 10 Ω, 100Ω, 1 kΩ, 10 kΩ, 100 kΩ and 1 M Ω.

[1]Notes appear on page 37.

*P.O. Box 969, Pueblo, CO 81002
**Assistant Technical Editor, QST

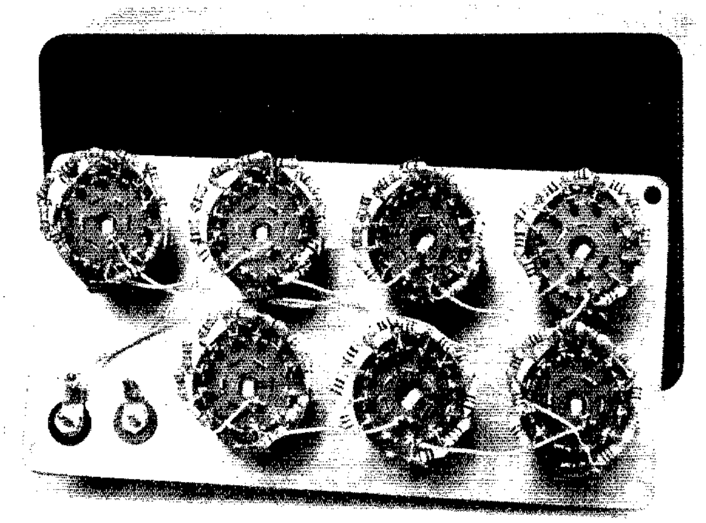


Fig. 2 — This inside view of the decade box shows the resistors surrounding each switch like ants playing "follow the leader."

the binding posts are insulated from the copper foil. Check the switch orientation and identify the first lug of each switch.

It's a good idea to check each resistor value with an ohmmeter prior to soldering it in place. It'll keep you from having to mess things up later, should there be a "baddie" in the bunch.

Then begin soldering the resistors to the switch lugs, forming a ring around the periphery of each switch. If you progress from left to right, bottom then top, it should be easy to keep track of the switch interconnections. When soldering, use a low-wattage (25 W or so) iron. Use of small-diameter (0.032-inch) solder helps things move along swiftly and smoothly.

After the resistors are in place, connect an ohmmeter to the binding posts and check each decade for proper operation. Once that's confirmed, you can plop the top panel assembly onto the bottom cabinet shell and secure it with a machine screw at each corner.

Summary

Remember that the power rating of the resistors used is only 1/4 W. Don't try to use the box as a dummy load for the transmitter or as a bleeder resistor for a kilowatt power supply! We hope you've enjoyed this "simple but engaging" project. It should make a welcome addition to your work bench.

Notes

[1]A complete kit of parts is obtainable from Circuit Board Specialists, P.O. Box 969, Pueblo, CO 81002.

[2]mm = in. × 25.4.

Strays

QEX: THE EXPERIMENTERS' EXCHANGE

Wonder what you've been missing by not subscribing to *QEX*, the ARRL newsletter for experimenters? Among the features in the December issue were:

- "RTTY Send Program & Interface for the Sinclair/Timex ZX81," by Brian Davis, W9HLQ
- A new, bimonthly "VHF + Technology" column by Geoff Krauss, WA2GFP
- "Nicads for the Azden PCS-2000," by Joe Pettengill, N2BC
- The popular "Components" column by Mark Forbes, KC9C.

QEX is edited by Paul Rinaldo, W4RI, and is published monthly. The special subscription rate for ARRL members is $6 for 12 issues; for nonmembers, $12. There are additional postage surcharges for mailing outside the U.S.; write Headquarters for details.

HAMS ATTEND NATO CONFERENCE IN PARIS

□ Recently, members of the Electromagnetic Wave Propagation Panel of the NATO Advisory Group on Aerospace Research and Development met in Paris. Hams who participated are Jack Belrose, VE2CV, as chairman; Dave Davidson, W1GKM, who gave a paper on earth-space propagation at 19 and 29 GHz; and Dick Grantham, VE1AI, who gave a paper on over-the-horizon salt-water path. VE2CV and W1GKM are ARRL technical advisors.

SOLID-STATE 6146s?

□ There was a time when some hams joked about having a miniature, solid-state equivalent to the 807 or 6146 vacuum tube. That once-jestful concept is now a reality, for Siliconix has been at work developing a device that almost fits the description. The DV1260T MOS power FET measures approximately 1 × 7/8 × 3/16 inch. The operating voltage is 12.5 (V_{DS} maximum = 45) and the power output is 60 W up to 175 MHz. The device dissipation (maximum) is specified as 240 W. Maximum drain current is 12 A.

For those who aren't familiar with power FETs, they are triode devices that have input impedances similar to vacuum tubes (1 megohm or greater). The typical drain impedance is low, on a par with that of power bipolar transistors. Furthermore, the IMD characteristics are much better than one can obtain with bipolar devices. In fact, the IMD products are as low in level as those from vacuum-tube amplifiers.

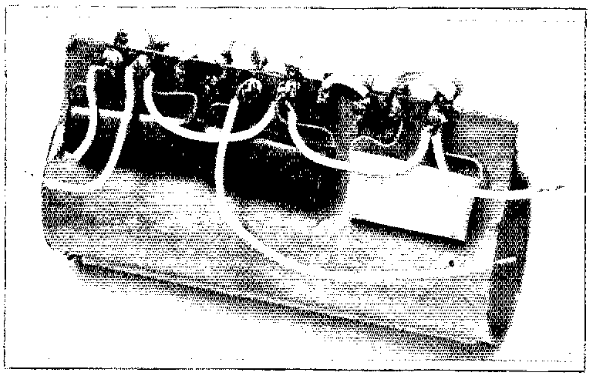
Siliconix and Motorola are both making significant advances in the power-FET technology. The writer, while recently working with a pair of Motorola MRF 138 power FETs, obtained 60 W of rf output at 30 MHz with only 288 mW of driving power. Sound like a tube? Yes, indeed! — *Doug DeMaw, W1FB*

Washington Governor John Spellman (seated) signs a proclamation designating September 13-19, 1982 as Amateur Radio Week. Looking on (l-r) are WA7RWK, K7RS, W7WLX, K7LAY and W7CKZ.

88-mH Inductors — A Trap!

Was the inductance of your modified 88-mH inductor correct? No? Here's the reason why, and some other useful filter information as well.

By Harold T. Mitchell,* NØARQ



Ed Wetherhold has designed a number of excellent audio-frequency band-pass filters that use surplus 88-mH toroidal inductors.[1,2] One of his latest designs is a simplified version of the Crud-O-Ject (COJ) cw filter.[3] It achieves greatly improved skirt selectivity using five resonators (tuned to 592 Hz), instead of three as in the COJ.

Fig. 1 depicts the simplicity of this single-stack filter that employs five standard-value capacitors wired to the existing solder terminals of the toroid stack. An end inductor is removed from the stack, modified to a value of 36.4 mH, then reinstalled in the COJ. Schematic and wiring diagrams and termination information are also given in Fig. 1.

The improved attenuation and skirt selectivity of the filter as compared with that of the original COJ is shown in Fig. 2. An excellent shape factor of 3.7:1 is indicated.[4] A 6/60-dB shape factor could not be calculated for the original COJ because it is so broad, relatively, that 60-dB attenuation data points do not appear on the plot.

Toroids Under Scrutiny

From 1968, when Wetherhold published his information on inductance and Q of surplus inductors for the amateur community,[5] tables of inductance versus turns removed have been used, which require interpolation for obtaining intermediate inductance values. Then, only one core permeability had been recognized.

*3M Electronic Products Division, 3M Center 207-1W, St. Paul, MN 55144

[1]Notes appear on page 39.

While modifying a number of 88-mH toroidal inductors for use in a different filter, I realized I had obtained different values of inductance using the toroids on hand, although the same number of turns had been removed from each toroid. One toroid had an inductance that was 18.3% lower than the required and expected value of 43.2 mH. This difference was traced to the fact that the toroid cores are made from two types of magnetic material having significantly different magnetic permeabilities.

Proper modification of the inductors can avoid large errors by taking into consideration the core type. The doubly wound (two-separate windings) toroid most commonly found is wound with green magnet wire on a light-blue (some might call it blue-green) core, and each winding start has a light-yellow tubing over the wire ends. Another type of two-separate-winding toroid is wound with green wire on a white core of higher permeability. Black sleeving is found on the ends of each start wire of that toroid type.

Bifilar-wound toroids are also available. These have red and green wires wound on a smaller core that is either brown or white in color. Tests of both bifilar-wound toroid types indicate that the cores have the same magnetic permeability. Perhaps the core colors are used to differentiate between manufacturers or core finishes.[6,7]

The Trap!

These audio-frequency toroids are beautifully analytic. It is possible to model these inductors precisely with equations that will give the number of turns to remove to obtain any desired inductance value. Since these modified 88-mH toroids have two equal windings connected in series aiding, I have written the equations in terms of the number of turns to remove per winding (turn pairs for bifilar-wound cores) and series-aiding inductance. These formulas are given in Table 1.

In addition to the gross error encountered as a result of the use of different core materials, there is a lot-to-lot magnetic permeability variation that is estimated to be ±8%. This variation is normally compensated for by adding or subtracting turns during the manufacturing process to obtain an estimated inductance value of 88 mH ±2%. An error analysis yielded a maximum error of ±6.7% at 43 mH for a modified toroid. Less error than this would almost always exist. In my estimation, an error within ±5% can be expected. The equations given in Table 1 are accurate for inductance values of 8 to 88 mH.

How many turns should be removed to obtain an inductance value of 36.4 mH for the improved COJ filter? From the equations: 132 turns per winding (264 total) for the blue core, 107 turns per winding (214 total) for the white core, or 133 turn-pairs (266 total) for the bifilar-wound toroid. If 132 turns were to be erroneously removed from the white-core toroid, the inductance value would end up being 27.7 mH, 23.9% low. This is the trap that exists. To avoid the trap, just identify the wire and core color and use the correct equation!

Summary

An s.a.s.e. (two stamps on a 4-1/2 × 9-inch envelope) sent to Ed Wetherhold[8] will bring a wealth of information on con-

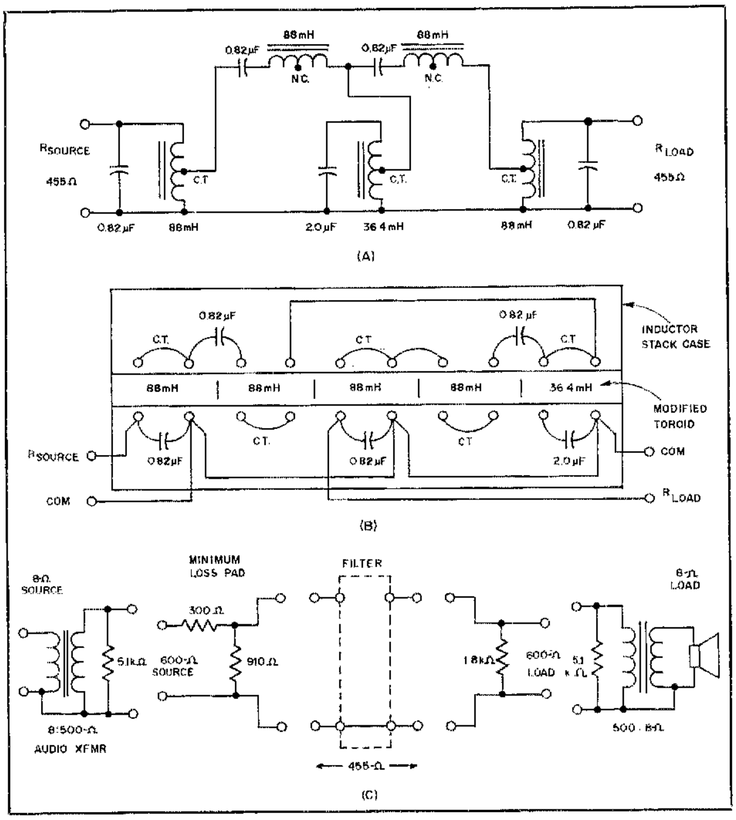


Fig. 1 — At A, the schematic diagram of the improved Crud-O-Ject cw filter. A pictorial wiring diagram is shown at B. One end toroid of the five-inductor stack is modified as discussed in the text. The impedance-matching methods shown at C should be employed to ensure that the filter is operating with the proper load impedances. Metallized Mylar capacitors with 100-V ratings may be used. Resistors are 1/2-watt, 10% units.

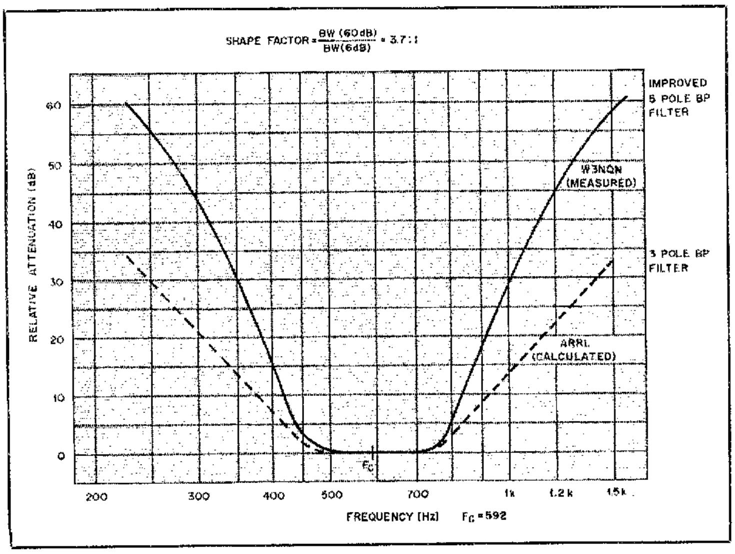


Fig. 2 — Comparisons of the attenuation characteristics of the original Crud-O-Ject filter and the improved Wetherhold version.

Table 1

88-mH Toroid Modification Equations

Two-separate-winding, blue-core, green-wire, yellow-sleeve toroid, 87.4 mH typical inductance:

$$T_{wr} = 373 - \sqrt{1592 \times L_{sa}}$$

$$L_{sa} = \frac{(373 - T_{wr})^2}{1592} \qquad \text{(Eq. 1)}$$

Two separate winding, white-core, green-wire black sleeve toroid, 87.3 mH typical inductance:

$$T_{wr} = 302 - \sqrt{1045 \times L_{sa}}$$

$$L_{sa} = \frac{(302 - T_{wr})^2}{1045} \qquad \text{(Eq. 2)}$$

Bifilar-wound toroid, 88.1 mH typical inductance:

$$T_{pr} = 372 - \sqrt{1571 \times L_{sa}}$$

$$L_{sa} = \frac{(372 - T_{pr})^2}{1571} \qquad \text{(Eq. 3)}$$

where

T_{wr} = turns removed from each winding.
T_{pr} = turn pairs removed.
L_{sa} = series-aiding inductance in mH. and 373, 302 and 372 are the number of original turns on the inductor (T_o).

struction and termination of a number of advanced cw, speech and RTTY filters, along with details of his offer to ship free surplus inductors to radio amateurs. These inductors are made available through the courtesy of the Chesapeake and Potomac Telephone Company of Maryland, and delivery is contingent on availability. For a nominal fee, Ed will not only ship inductor stacks to filter builders, but also will supply matched capacitor sets and impedance-matching audio transformers. These filters like to "see" and work into resistively terminated sources and loads of the proper value. A 10 to 20% impedance mismatch can be tolerated, however, depending on the reflection coefficient of the filter.

I gratefully acknowledge the assistance of Ed Wetherhold. The stimulation of his designs have led me to this interesting work. In addition to Ed's review of this article, I am indebted to Joseph A. Gutowski of EWC, Inc., for his assistance.

Notes

[1]E. Wetherhold, "Modern Design of a CW Filter Using 88- and 44-mH Surplus Inductors," *QST*, Dec. 1980, and Feedback, *QST*, Jan. 1981, p. 43.
[2]E. Wetherhold, "High Performance CW Filter," *Ham Radio*, April 1981.
[3]J. Hall and R. Myers, "The CRUD-O-Ject," *Weekend Projects For the Radio Amateur*, Volume I (Newington, ARRL, Inc., 1979) and *QST*, Feb. 1972.
[4]D. DeMaw, "Beating Rotten QRM — CW Filtering for the Beginner," *QST*, Oct. 1981 and Feedback, *QST*, Dec. 1981, p. 55.
[5]E. Wetherhold, "Inductance and Q of Modified Toroidal Inductors," *QST*, Sept. 1968.
[6]"Molybdenum Permalloy Powder Cores," Catalog PC-104H, Arnold Engineering, Marengo, IL 60152.
[7]"Molypermalloy Powder Cores," Catalog MPP-303T, Magnetics Inc., Butler, PA 16001.
[8]E. Wetherhold, W3NQN, 102 Archwood Ave., Annapolis, MD 21401.

APPENDIX

Since this article was written, a fourth type of 88-mH toroid has been found. Using inductance data obtained from the toroid by means of a precision RLC impedance bridge, equations were developed. The equation development method and the equations will be given here. A simple inductance-measuring technique will also be described.

Using the initial inductance value and any other inductance value measured after removing at least 20% of the toroid turns, the T_w^2/L_{sa} quantity (turns of wire per series-aiding inductance in mH) can be accurately calculated for various assumed values of T_o (number of original turns of wire). If the assumed value of T_o is less than actual, T_w^2/L_{sa} will decrease when turns are removed. If the assumed value of T_o is too large, the term will increase. For some T_o values, the calculated quantity will remain essentially constant, and the two then can be used to form the desired equations.

As shown in Table 2, four data points were obtained for the purposes of illustration, although two would have been sufficient in practice. When all turns were removed, 359 were counted for each winding, which agreed exactly with the assumed value of T_o.

Table 2

Data For the Two-Separate-Winding, White-Core, Red-Wire, Yellow-Sleeve Toroid

Measured Data		*Calculated Data* $T_w^2/L_{sa} = (T_o - T_{wr})^2/L_{sa}$		
T_{wr}	L_{sa}	T_o = 358	T_o = 359	T_o = 360
0	88.0	1456	1465	1473
50	65.1	1457	1467	1476
100	45.8	1453	1465	1476
200	17.25	1447	1466	1484
		(Decreasing)	(Constant)	(Increasing)

Therefore

$$T_{wr} = T_o - \sqrt{\frac{T_w^2}{L_{sa}} \times L_{sa}}$$

$$= 359 - \sqrt{1466 \times L_{sa}} \qquad \text{(Eq. 7)}$$

and

$$L_{sa} = \frac{(T_o - T_{wr})^2}{\frac{T_w^2}{L_{sa}}} = \frac{(359 - T_{wr})^2}{1466} \qquad \text{(Eq. 8)}$$

An Inductance Measuring Method

For purposes of equation development or resonating an L-C circuit at a frequency of interest, the simple circuit shown in Fig. 3 may be used. If a close-tolerance capacitor (or one measured with a precision capacitance meter) is used, the corresponding inductance at resonance can be calculated from the equation

$$F_r = \frac{1}{2\pi\sqrt{LC}} \qquad \text{(Eq. 4)}$$

where

F_r = resonant frequency in hertz
L = inductance in henrys
C = capacitance in farads

For example, if C = 0.512 μF, and F_r is 750 Hz

$$L = \frac{1}{C\,(2\pi\,F_r)^2} = \frac{10^6}{(0.512)\,(2\pi \times 750)^2} \qquad \text{(Eq. 5)}$$

After inductor turns have been removed and counted, if resonance at 738 Hz is established with a 1 μF capacitor, then the second inductance value is calculated:

$$L = \frac{10^6}{(1)\,(2\pi \times 738)^2} = 46.5 \text{ mH} \qquad \text{(Eq. 6)}$$

In the preceding example, about 27% of the turns were removed ($\sqrt{46.5/88}$ = 0.727), thus assuring accuracy when writing the other equations.

When using the test setup shown in Fig. 3, vary the generator frequency to obtain a maximum voltage indication on the VTVM, and record the resonant frequency. Calculate the inductance using Eq. 5. If the counter is accurate to ±1 Hz and the capacitor has a tolerance of ±0.5%, the calculated inductance should be accurate to within ±1%. For best accuracy, the resonating capacitance should be more than one thousand times the isolation capacitance. If the resonating capacitance is too low, the isolation capacitance will be significant enough to effect the accuracy of the calculations. QST

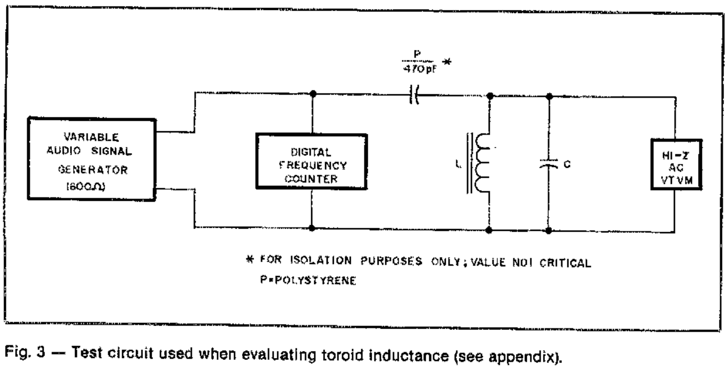


Fig. 3 — Test circuit used when evaluating toroid inductance (see appendix).

Strays

ARRL President W4KFC (center) poses with 50-year members Vice President KØGA (left) and K6EA, of Bemidji, Minnesota, at the Moorhead, Minnesota, convention in September.

TECHNICAL PUBLICATION FOR THE BLIND

□ Blind and visually impaired amateurs who are interested in current technology can subscribe to the *Smith-Kettlewell Technical File*, a low-cost publication tailored to their needs. Produced by the Rehabilitation Engineering Center of the Smith-Kettlewell Institute of Visual Sciences, the *Technical File* is published quarterly in Braille, large print or talking book form, and provides independent access by the visually handicapped to state-of-the-art devices, circuits and adaptive techniques. For more information and a free first issue, write to or call William A. Gerrey, Editor, Smith-Kettlewell Institute of Visual Sciences, 2232 Webster St., San Francisco, CA 94115, tel. 415-561-1619.

You're never too old to become an amateur. Ed Rheault, a white-caner from Kenora, Ontario, is 72 and now on the air. From left to right: Bill Lowe, VE3MOS; Ed, VE3MOX; Ed's sponsor, Phil Mosher, VE3JJF; and Milt Sizer, VE3JJO, of Lake-of-the-Woods ARC. *(Kenora Miner and News photo)*

Product Review

Conducted By Paul K. Pagel,* N1FB

Cushcraft Corporation A4 Triband Yagi

☐ I have neither the desire nor the tower capability to place aloft one of those huge, super-dandy, drooping-monster hf-band Yagis at my QTH. My modest 50-foot, unguyed spire (Rohn 25G) will accommodate any reasonable size beam antenna,[1] but there are weight and wind-loading limits that must always be considered. Therefore, I have worked with a succession of 3-element triband Yagis for operation on 20, 15 and 10 meters, and the results have been entirely satisfactory for my DX appetite (modest).

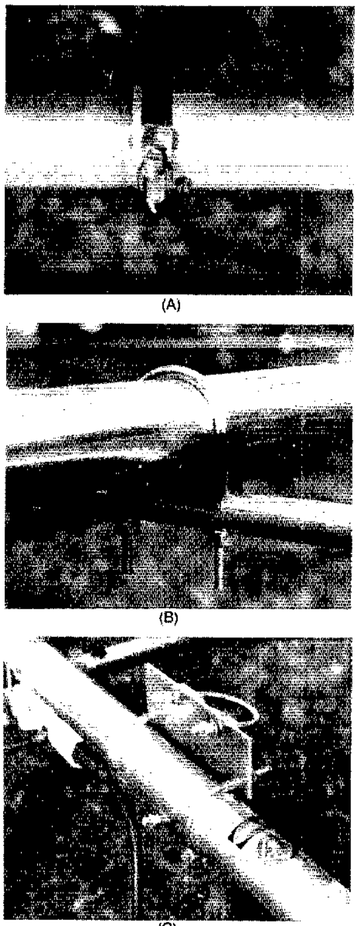
There must be an acknowledged trade-off between size and efficiency when using a trap type of antenna, for there are always some losses in the system. But, the difference between a full-size 3-element Yagi and a shorter one with traps would yield a performance difference that would be difficult to measure. Practically, a well designed triband trap Yagi will provide good performance for all-around hf communications.

A major consideration, once you've chosen to erect a tribander, is the physical makeup of the antenna. That is, it needs to be rugged enough to withstand wind and ice loading (and vibration) without generating a shower of metal parts on your house or lawn. (I had that unwanted experience with one brand of beam antenna some years ago, owing to poorly designed element clamps.)

Based on comments by Paul Pagel, N1FB, I'd concluded that the Cushcraft A4 Yagi should be as rugged as his A3 version, so it was selected for use at W1FB. He was correct, for the system is nicely designed in terms of physical structure. Fig. 1A shows how the element sections are held together by means of steel compression clamps. Fig. 1B illustrates the Cushcraft method of joining the elements to the boom in a plastic cradle through which a U bolt is inserted. The boom-to-mast plate and U bolts are shown in Fig. 1C. This hardware is situated just behind the driven element feed point. A Bencher balun is shown in photograph C. It was tried experimentally and replaced later by the specified eight-turn decoupling choke made from RG-8/U cable.[2]

All of the aluminum stock for the A4 is polished and of high quality. No defects caused by the extrusion process were noted (always check for cracks or thin areas before assembling an antenna).

The A4 is rated at 2000-W PEP and has a balanced feed-point impedance of 50 ohms.



(A) (B) (C)

Fig. 1 — View of the steel hose clamp that joins the element sections (A). The boom-to-element mounting hardware is shown at B. Illustration C shows the boom-to-mast mounting method.

The longest element dimension is 32 feet. An 18-foot boom is used. Boom diameter (OD) is 2 inches. A turning radius of 18 feet is required for the A4. The theoretical gain for a 3-element Yagi of this spacing is approximately 8 dB, with an average front-to-back ratio of approximately 25 dB being typical.

Although the assembled antenna appears to be a 4-element type, it is not — at least with regard to electrical traits. There are four *physical* elements, but in operation on a given band there are but three elements. The fourth element is a reflector for 10 meters. The larger reflector behind the 10-meter one is used during 15- and 20-meter operation. This eliminates the need for an extra set of traps in the main reflector element.

Construction

Owing to an inherent lack of mechanical dexterity, it probably took me longer than most amateurs to assemble the A4. I devoted about four hours to the task. Another hour was spent getting the antenna mounted on the tower (courtesy of KC1V and AK4L). The only glitch in the process occurred during erection (at dusk): The Yagi was attached to the mast "upside down," and nobody noticed it. The undesired effect was readily apparent two days later when the SWR skyrocketed following a rain storm! The trap drain holes were of course pointing upward, and each trap had become a small reservoir! A 180-degree flip-flop followed, and once the traps dried out I was back in business. This illustrates clearly the need to make certain the trap holes are pointing downward after assembly, as indicated in the instructions.

As a preventive measure after assembly, I placed a generous blob of noncorrosive RTV sealant over each bolt head and nut. Cushcraft supplies zinc-plated steel hardware, which should last a long time. But, an added measure of protection will help prevent oxidation and aid one's peace of mind when contemplating antenna longevity.

The ends of the elements and boom are fitted with plastic caps to prevent accumulations of moisture. This also lessens vibration from the wind and prevents the elements from "singing."

Performance

My A4 is fed with 50-ohm, aluminum-

[1] meters = ft × 0.3048, mm = in. × 25.4.

[2] The Bencher balun was tested in the interest of compactness and physical convenience. However, the VSWR readings of Fig. 2 were obtained while using the 8-turn, 6-inch diameter coil of RG-8/U cable specified by Cushcraft. It is likely that the coaxial-cable decoupling choke is somewhat less lossy than the trifilar coil balun. The balun shifted the antenna resonance somewhat lower which would have required readjustment of the element lengths.

*Assistant Technical Editor

Cushcraft Corporation A4 Triband Yagi

Manufacturer's Claimed Specifications	*ARRL Evaluation*
VSWR: 1.2:1 typical (resonance).	Confirmed (see Fig. 2).
Power rating: 2000 W PEP.	Confirmed.
Feed impedance: 50 ohms.	Confirmed.
Material: 6063-T832 seamless tubing.	Confirmed.
Weight: 37 pounds (16.8 kg).	As stated.
Wind surface area: 5.5 sq. ft (0.51 sq. m).	—
Longest element: 32 feet.	As stated.
Boom length: 18 feet.	As stated

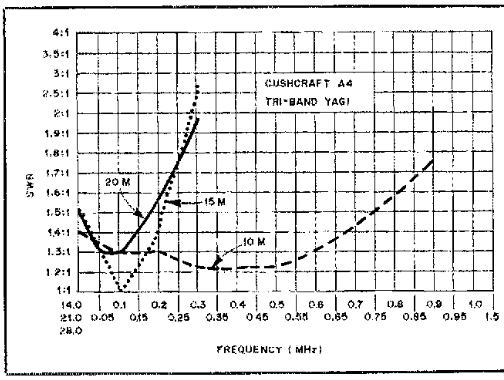
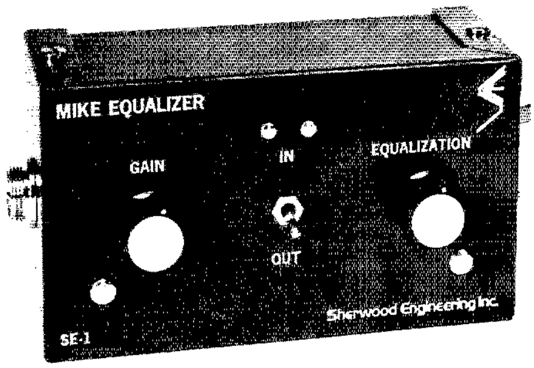


Fig. 2 — SWR curve of the Cushcraft A4 tribander.



jacketed Hardline. Overall transmission-line length is 60 feet. A Bird Thruline wattmeter was connected between the lower end of the feed line and an FT-101ZD signal source. The A4 was adjusted for optimum operation in the cw portions of the three bands, in accordance with the dimensions given in the instruction sheet. The resultant SWR curves are shown in Fig. 2.

$$\mathrm{VSWR} = \frac{1 + \sqrt{\frac{P2}{P1}}}{1 - \sqrt{\frac{P2}{P1}}} \qquad \text{(Eq. 1)}$$

where P1 is the forward power in watts and P2 is the reflected power in watts.

I find it necessary to use a Transmatch to disguise the SWR when operating in the phone bands (which I seldom do). Operation by that means is satisfactory. Those wishing to strike a compromise between phone and cw operation may wish to adjust the antenna for midband resonance. A chart is given in the instruction sheets to provide dimensions for three segments of each band — phone, center and cw.

On-the-air performance has been excellent, with suitable front-to-back and front-to-side ratios evident. I have had no difficulty working DX worldwide, using 100 W and 1000 W of dc input power to the transmitter PA. My batting average for breaking pileups on cw has also been satisfactory.

The A4 has endured several severe wind storms (up to 70 mph) and one ice storm without mishap. Based on my experiences with the antenna during the past year, I would give it a top rating for a small, moderately priced beam antenna. Those wanting four-band operation with the A3 and A4 Cushcraft antennas may purchase the 40-meter conversion kit and install it in the driven element. The A4 price class is $330; A744 (7- or 10-MHz) adapter kit, $90; A4SK stainless-steel hardware kit, $55. Manufacturer: Cushcraft Corporation, P.O. Box 4680, Manchester, NH 03108. — *Doug DeMaw, W1FB*

SHERWOOD ENGINEERING SE-1 MICROPHONE EQUALIZER/ PREPROCESSOR

□ Recently, amateurs have begun to take greater notice of their ssb signal fidelity. Traditionally, a "speech processor" has been a device that alters the signal peak-to-average amplitude ratio. Such systems invariably alter or degrade the audio characteristics of the voice — usually causing it to sound "bassy" and "rough."

Theoretically, the audio response of a microphone amplifier circuit in a communications phone transmitter should be flat across the voice band (300 to 3000 Hz). Often, this is not the case. The microphone response should be flat across this band, also. Again, frequently this is not the case. As a result, we have come to expect ssb signals to sound "Donald Duck-y."

Recording studios use equalizers to correct for distortion introduced by electrical and acoustical factors. An equalizer has controls that permit the operator to adjust the phase and level of several segments of the audio spectrum. The Sherwood SE-1 represents one attempt to introduce this kind of processing into Amateur Radio transmitters. It won't increase the average power level of your signal, but it may make it easier for operators on the receiving end to understand what you are saying.

The SE-1 is housed in a heavy, steel black box, measuring 3 × 5-1/4 × 2 inches (HWD).[1] Internally, the device is a "black box" also. A single transistor, one IC, a toroid and a few other components are mounted on a glass-epoxy pc board. All markings have been removed from the active devices. Sherwood does not supply a diagram of the circuit, but does provide minimal circuit description. In fact, the only paperwork supplied with the SE-1 are installation and operating instructions on two sides of one sheet of paper. The installation instructions are complete and should be adequate for anyone who is moderately comfortable when using a soldering iron and pliers.

A toggle switch places the circuit in line and applies power (9-V transistor-radio battery). Two controls, GAIN and EQUALIZATION, can be adjusted from the front panel. Adjusting GAIN simply changes the output signal level while changing the EQUALIZATION setting alters the tonal components of the signal. Advancing the control clockwise *appears* to enhance the higher-frequency components. Unfortunately, the exact functioning of the control is not specified in the literature.

I called Bob Sherwood and asked him about the EQUALIZATION control. Bob told me it controls the "tilt" of a processing stage. He went on to tell me what he meant by "tilt." With the control fully counterclockwise, a 2800-Hz tone through the device will have a 2-dB advantage over a 300-Hz tone. As the control is rotated to the fully clockwise position, the advantage increases from 2 to 20 dB. This description matches the subjective appraisal I received from operators on the air.

Bob stated that the microphone preamplifier stage consists of a single FET and a gain control. The high impedance of the FET permits the circuit to be used with any microphone, without danger of the circuit loading the microphone. Beyond that, Bob was reluctant to discuss the circuit.

On-the-air reports indicated that the SE-1, *properly adjusted,* made my ssb signal sound more natural — more like the audio extracted from an fm circuit. I believe it made it easier for other operators to understand what I was saying. If you are concerned about the quality of your signal and if you want to be understood as well as be heard, then you should give the SE-1 serious consideration. The SE-1 is available from Sherwood Engineering Inc., 1268 South Ogden St., Denver, CO 80210. Price class: $100. — *Peter O'Dell, KB1N*

DAIWA AF-606K ACTIVE AUDIO FILTER

□ Audio filters are in abundant supply these days. They come in many sizes, shapes and price classes, but the Daiwa unit has a distinctive look — akin to that which characterized post WWII military gear. I was impressed with the clean, snappy appearance of the filter when I extracted it from the box. The panel appears to be black anodized aluminum, and the lettering at each control is off-white. A dark gray case provides a two-tone contrast.

But, appearance is not the primary consideration when buying a new piece of equipment for the shack. How does it "play"? That's the question asked by a smart buyer, and rightly so! Bells and whistles (if I may use the vernacular) seem to have a biasing effect on today's purchaser of new apparatus. But some of the fancy gee-gaws being offered could just as easily be omitted in the interest of keeping the unit cost within the reach of the common man or woman. The AF-606K has one feature that might be classed as a frill (more on that later), but it otherwise is a pretty basic audio

[1]mm = in. × 25.4

filter with variable bandwidth and a notch function.

The Daiwa unit does not have a built-in power supply. A low-current, 12-V external power supply is required for operation. *Beware!* The outer ring of the 12-V jack is the positive one. The center pin is for the negative or ground lead of the supply. This is not the U.S. convention, so don't let habit get you in trouble when you hook up your unit.

Installation requires a patch cord from the phone jack of your receiver or transceiver to the input phono connector on the rear of the AF-606K. A built-in speaker permits monitoring the filter output. Alternatively, the operator can attach an external speaker, or may elect to connect headphones by means of a front-panel jack. I found it best to use phones, since the speaker function did not produce room-volume audio at a level that was comfortable during weak-signal reception. Attempts to increase the output level by turning up the receiver audio gain resulted in distortion from overdriving the audio filter (a normal experience with outboard audio filters when the maximum tolerable excitation limit is reached).

Five controls are located on the front panel of the unit. Left to right are NOTCH, PLL, BAND PASS, MODE and POWER. The NOTCH control is variable from approximately 500 to 2500 Hz, with some overrun at each end of the control. I measured the notch depth (at 700 Hz) as 33 dB. I'll discuss the PLL control later on.

The BAND-PASS control is used to peak the audio-filter response for the cw pitch the operator prefers (500 to 700 Hz in my case). A band-pass response is provided in this unit, which yields audio roll-off above and below the desired frequency. The pass band is variable from approximately 400 to 1200 Hz, with some extra range at each end of the control.

MODE selection is accommodated by the near-right control. It enables the user to choose a notch condition, three ssb bandwidths (1.5, 2.0 and 2.5 kHz), three cw widths (80, 110 and 140 Hz) and PLL. The POWER ON-OFF switch is at the far right on the panel. Directly below it is the PHONE jack.

Now comes the "hell" or "whistle," whichever word you may prefer. The PLL function enables the operator to tune in a cw signal and listen to it via a keyed tone that is generated within the AF-606K. In effect, the cw signal from the receiver is detected, then routed to a control circuit, which actuates a tone generator. An LED on the front panel of the audio filter illuminates when the PLL frequency control is set to the pitch of the cw signal, as heard when the filter is turned off. Under this condition the PLL is considered in the "lock" mode. The purpose of the PLL function is to eliminate QRM and band noise. Effectively, all you will hear is a single cw note coming from the audio filter.

Various schemes of this type have been contrived and tested for a number of years. None of them proved to be spectacular. The major limitations are that very weak signals do not trigger the tone oscillator in a reliable manner, which leaves gaps in the cw message. Also, noise pulses will key the tone generator, causing false blips, "stuttering" and incoherence. These problems were noted while testing the Daiwa filter. When a strong cw signal was used to lock the PLL, I noticed the effect of excessive "weighting" on the cw characters. There was also a clicky characteristic to the tone-generated cw note. For the most part, coherence was far superior without the PLL function in use. If the buyer does not wish to have the PLL feature, he or she can purchase the model AF-406K, which is minus the tone decoder.

Filter performance is otherwise excellent. I found no evidence of ringing, and audio output from the AF-606K was very clean within the normal listening range while using headphones. Certainly, the filter did a fine job of "laundering" the receiver output with respect to reducing QRM, annoying heterodynes and receiver wide-band noise. Weak signals were "lifted" nicely out of the noise, providing Q5 copy when copy was not possible without the filter in the line.

This filter and others with similar performance characteristics can spell the difference between success and failure when copying weak cw signals, such as one encounters on 160, 80 and 2 meters, where noise is a universal foe. Sideband operators will find that a good audio filter will reduce adjacent-frequency splatter and rumble. In many instances, the audio filter will give the same effect as a speech processor when it is actuated by the person you are listening to.

Dimensions are 6 × 6 × 2-1/2 inches.[4]

[4]mm = in. × 25.4

Price class is $121. Distributed by MCM Communications, 858 E. Congress Park Dr., Centerville, OH 45459, tel. 513-434-0031. — *Doug DeMaw, W1FB*

HAMLOG/APPLECODER

☐ I have some good news, some bad news and some more good news. The good news is that HAMLOG, a log keeping and maintenance program for the Apple II® computer, does everything it says it will do. The bad news is that there is so much that it will not do, frustratingly so, at times. The other good news is that those of you who feel compelled to use a $47 program and $2000 worth of computer equipment to do what a pen and a box of 5 × 7 index cards can do nearly as well won't be disappointed.[5]

HAMLOG fails to take sufficient advantage of the Apple II system power. Users are constantly restricted by the program limitations. That is, they must adapt to the program needs, rather than the other way around. Those who have seen a demonstration of or used such programs as Visicalc or Visidex know how "user friendly" a program can be. HAMLOG is a casual acquaintance at best, and a relatively expensive one at that.

Using HAMLOG

Initialization. When first using HAMLOG, you must initialize the system with your call sign, various modes that you expect to use, and your station-setup data. The latter demonstrates the limitations of HAMLOG rather quickly. Station-setup data is limited to seven characters. For example, the author suggests that a station consisting of a "standard" transceiver, wire antenna and linear amplifier be designated QRO-W-L. You can have several such station setups, so your log can show which particular configuration you were using when you worked a particular station. The brevity of a seven-character designation does have its advantages when it comes to displaying or printing out log data (this brings up another defiency of HAMLOG, which will be discussed later). Nevertheless, the restriction is too severe. Users, after all, can always choose brevity; it need not be forced upon them.

Log Entry: After initializing the program, you can run it. The program is menu-driven. One item on the main menu is "log entry," which is pretty straightforward, although it is definitely geared toward casual operating and not contesting. You enter one log entry item at a time: call, name, band, etc. However, as you enter the items, the previously entered items "scroll up" and quickly disappear from the screen. By the time you've entered the other station's signal report, the call and name have disappeared, and you cannot see them until you finish entering the entire log entry and select the "display log entry" item from the main menu. This is a major inconvenience. As it takes some time to step through the log entry menu, I found myself frequently writing down the other station's call, name, etc., so I wouldn't forget them in case he turned the QSO back to me before the log entry was completed (initial exchanges can be rather quick). This pretty much defeats the purpose of using a computer.

[5]H. Smith, "A Speedy QSO File," *CQ*, June 1962, p. 62.

Change Log Entry. Another item on the main menu is "change log entry." Unfortunately, when you use this item the old log entry does not appear on the screen; unless you have printed out or written down the log entry so you can look at it to know what you're changing, your memory is put to the test. It is a virtual necessity for the old log entry to be displayed while you are changing it. The fact that HAMLOG doesn't allow this is sheer shortsightedness.

Why would you want to change a log entry, anyway? Well, there are lots of reasons, but one very good reason would be to add some comments. You guessed it: According to the HAMLOG User's Manual, "you cannot add text that wasn't there, and you may not increase the number of text lines in the entry you are altering."

Search Log. A useful feature is the "search log" mode, which allows you to enter a call sign, a QTH, a date or a combination thereof, and causes the computer to display or print all log entries with the appropriate parameters. It takes about 3-1/2 minutes to search approximately 500 records, but fortunately you can set "core pointers" to optimize the search routine, and a search by call sign through the same 500 or so records is reduced to 15 seconds. Not bad! (Of course, with my file of 5 × 7 index cards I can perform the same feat in less than 5 seconds.) The core pointers have to be set each time you use the program, though. You can set pointers for call sign, QTH or date, but you can only have a single set of pointers at one time.

The search routines are limited, too, by the maximum number of log entries permitted on a log diskette. Typically, a single diskette can store data for 2000 contacts, *if* you do not append comments to the log entries. With comments, the maximum number of contacts on a diskette is reduced accordingly. For very practical reasons, you can only search *one* log diskette at a time, which seriously inhibits the utility of the search routines after you've been on the air for a while and have filled several log diskettes.

Output Format. HAMLOG formatting to the screen or printer is rather ho-hum. Log entries are printed out line by line, and the items within the log entries are separated by spaces. There is no serious attempt to format the screen or page, which would be easy to do, wouldn't slow down the program noticeably, and would make the entries so much easier to read. In fact, HAMLOG would be a whole lot better if entries were displayed on the screen while you are entering or changing them, with a "bouncing cursor" controlled by appropriate keys. This would allow you to move from item to item, make the necessary entries or changes, and show you exactly what you have entered as you enter it (and after you enter it) without its being blanked from the screen. The technology that permits this exists, but HAMLOG doesn't use it. Perhaps the second version of HAMLOG will. I suggest that you wait for it.

Applecoder

Applecoder is a Morse code teaching program for self-teaching or for the classroom. It, too, is menu driven. The author states that it needs no significant written instruction, and he's right. It is a cinch to use.

Applecoder sends random word groups utilizing the Apple II speaker. You select a "seed" number, which determines the random word groups and allows you to regenerate the same sequence at will. If you have a printer, you can generate "hard copy," which is useful for checking results for yourself or for a class full of students. The word groups are displayed on the screen, but you can always turn the screen off if you are tempted to peek.

Applecoder permits you to change the code tone (200-2000 Hz), speed (5-40 wpm) and character group length (1-9), and will also allow you to exclude certain characters (punctuation marks, numerals, etc.) at your discretion. By varying the random number seed, you never run out of "fresh copy."

Code-speed adjustments may need some fine tuning. The 5-wpm setting was accurate enough, but the 20-wpm setting was, by my calculation, closer to 25 wpm. Sound from the Apple II speaker tends to be muffled and tinny, but removing the top of the computer enclosure improves things considerably. Still, a large group might have difficulty, unless an outboard speaker were installed.

Applecoder uses the ARRL-accepted procedure for teaching and learning the code. At the slower speeds, the characters are sent fast, with additional space between words. This prevents the student from counting individual dots and dashes. All in all, Applecoder is a useful product. — *Hal Steinman, K1FHN*

HAMLOG

Requires 48K + DOS 3.3
Author: Dr. Keith W. Reiss, CECO, Inc.
Price: $47 (disk + manual)

Applecoder

Requires 16K + DOS 3.3
Author: Dr. Keith W. Reiss, CECO, Inc.
Price: $19.95 (disk + instructions)

Both: $58

THE FIST FIGHTER

To a cw operator's ears, there's nothing sweeter than the sound of a smooth fist batting out Morse code. The art of sending good cw has come a long way since the invention of the straight key. Electronic keyers of many different types are prevalent, and with computers and dedicated terminals becoming so much a part of the ham shack, it seems there's not much left that is done away from a keyboard and video monitor. However, there are those among us who still like the feel of the ol' straight key between their fingers.

Description

The Fist Fighter was designed by the Blacksburg Group to help straight-key advocates send perfectly timed Morse code. There are nine CMOS ICs and four transistors (and associated components) contained within the package to do just that. The front-panel mounted VOLUME/ON/OFF control functions as a sidetone level adjustment and power switch. Sending speed is varied by the adjacent SPEED potentiometer that provides a range of adjustment from about 3 to 30 wpm. There's an auto tune-up feature that is jumper-selectable. This provides a constant key-down condition during transmitter tune-up.

The rear panel supports the KEY jack, GRID (−300 V) and DIRECT (+300 V) key line outputs, and a 1/8-inch jack for connection of an external power source. A lantern battery or ac-operated supply delivering 5- to 15-V dc is required to power the unit. A built-in speaker is mounted on the inside of the top cover.

How Does It Work?

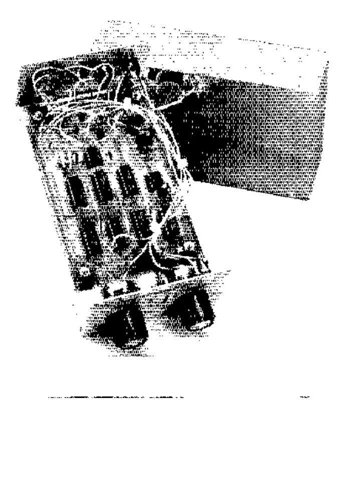
The space/dot/dash ratio has been designed for a 1:1:3 relationship. When you close the key contacts, the Fist Fighter begins sending a dot at the speed you've selected. If you release the key before the end of the dot, the element is completed automatically. If you keep the key closed for a period longer than the dot time, the unit "recognizes" this as an attempt to send a dash and lengthens the element. After each element a minimum off period (equal to a dot length) is enforced by the Fist Fighter. This prevents the dots and dashes from being sent too close together. A key closure during this period will be "remembered" and keying will start immediately after the forced space.

If the auto tune-up feature is wired on the board, a prolonged key closure will result in the sending of one code element followed by a constant key closure. This might lead initially to some confusion when you first try to use the keyer. You might try using it in both configurations to determine which is best for you.

Argument Pro and Con

The Fist Fighter does exactly what the manufacturer says it will do. It will make you send perfectly timed Morse code when using a straight key. However, this does not mean that it will teach you how to send good code with a hand key without the assistance of the Fist Fighter. In use, you must yield on a code element sooner than you would otherwise. This means the actual key closure will be less than it should be if you were sending Morse with only the hand key in the line. But . . . one may argue that such is the case *with all* types of electronic keying devices. For instance, if you've developed some of the characteristic sending habits of those who use iambic keying methods, would you be able to put down the squeeze key and keyer and send code as well with the old bug that's been relegated to the operating desk drawer? Probably not. The upshot of all this is that once you've learned to use the Fist Fighter effectively, the chances are that you will have to stay with it as you would with any electronic keying device. (The Wouff Hong should dispel any thoughts of using this device on SKN!)

The Fist Fighter is available from The Blacksburg Group, Box 242, Suite 100, Blacksburg, VA 24060. Size: 2-1/8 × 3-1/2 × 7-1/2 inches (52 × 89 × 190 mm) (HWD). Price class: kit form, $60; wired and tested, $80. — *Paul K. Pagel, N1FB*



Hints and Kinks

Conducted By Larry D. Wolfgang,* WA3VIL

75-W AMPLIFIER FOR 10-METER FM

☐ Converted CB rigs are popular for 10-meter fm operation.[1] When conditions were good, these rigs provided many contacts, but with declining sunspot activity the 5-W power limit may be too restrictive.

I modified an old, homemade cw rig to serve as a Class C amplifier. The original circuit came from the 1962 *Radio Amateur's Handbook*.[2] The amplifier works well with my converted Hy-Gain CB board, and has a power input of 60 to 75 W. A Class C amplifier is nonlinear and is not suitable for use with an a-m or ssb rig, because it will distort the signal. A power supply capable of providing 600-V dc under load for the 1625 amplifier tube and 12-V ac for the filament is required. The original project included a power supply, but the tube rectifiers could be replaced with solid-state units.

I removed the 6AG7 buffer/multiplier circuitry and added an impedance-matching network at the input of the 1625 tube. The schematic diagram of the amplifier is shown in Fig. 1. The variable capacitors were taken from old broadcast-band receivers. A neutralizing capacitor made from several turns of twisted wire is used. Be sure to use wire with good high-voltage insulation. You will have to experiment with the number of turns in order to stabilize the amplifier.

[1]Bob Heil, "Experience 10-Meter FM," *QST*, Aug. 1981, p. 22.

[2]"An Inexpensive 75-Watt Five-Band Transmitter," *The Radio Amateur's Handbook* (Newington: ARRL, 1962), pp. 174-177.

*Assistant Technical Editor

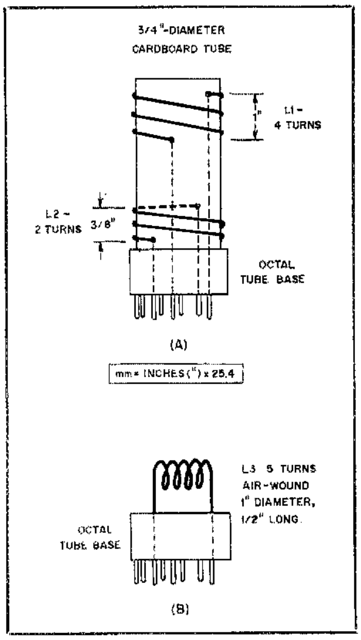


Fig. 2 — Coil-winding information for L1 and L2 is given at A, and for L3 at B. Octal tube sockets and bases were used in the original circuit to simplify band changes. This will not be necessary if you are building the amplifier instead of modifying an old transmitter.

Coil-winding information is given in Fig. 2. Use no. 14 wire for L1, L2 and L3. L3 should be air wound.

To use the amplifier, tune C1 and C2 for a transceiver output equivalent to that obtained when using a 50-ohm dummy load. Next, tune C3 and C4 to obtain the maximum output from the amplifier with a plate current of about 100 mA. This should result in 60 W or more of dc input power. One source of 1625 tubes is Fair Radio Sales, P.O. Box 1105, 1016 E. Eureka St., Lima, OH 45802. The cost should be but a few dollars each. — *Dick Gulatsi, AC3D, Devon, Pennsylvania*

INEXPENSIVE CTCSS TONE GENERATOR

☐ A common frequency for continuous tone-coded squelch systems (CTCSS) in many parts of the country is 100 Hz. I have a Sears Model 3880 2-meter rig, and was faced with the prospect of spending $20 or more for a commercial tone generator. In an effort to save that money, I built the circuit shown in Fig. 3. The circuit divides the 10-kHz synthesizer reference signal down to 100 Hz in a pair of 4018 CMOS divider chips.

The first stage can be any CMOS divide-by-10 chip, but the second stage must be a 4018 IC connected as a "walking ring" or Johnson counter. Four resistors form a digital-to-analog converter, which produces a fair approximation for a sine wave at 100 Hz. The 0.22-µF capacitor provides additional filtering. You should be able to use 39-kΩ and 22-kΩ resistors instead of the 1% values shown in the drawing. Just match them carefully with your ohmmeter.

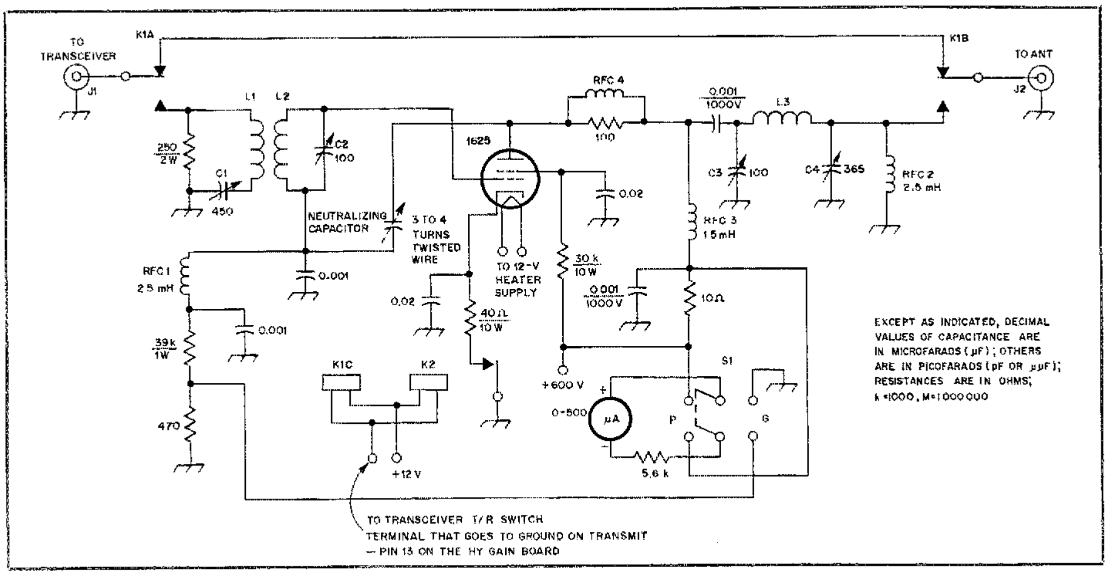


Fig. 1 — Schematic diagram of the revised cw transmitter used as a Class C amplifier for 10-meter fm operation. See the text for a description of how to build the "gimmick" type of capacitor used for neutralization.

K1 — Dpdt relay with 12-V, 160-ohm coil, such as Radio Shack 275-206.

K2 — Spst relay with 12-V, 1200-ohm coil, such as Radio Shack 275-003.

RFC4 — 8 turns of no. 18 enam., close wound on 100-Ω resistor.

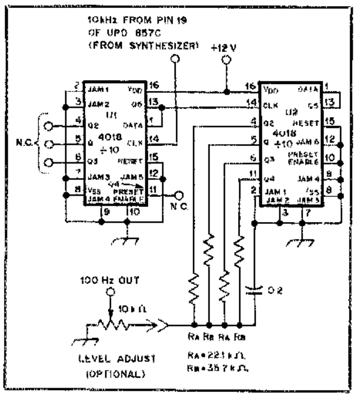


Fig. 3 — Schematic diagram of the 100-Hz tone generator built by WB8NBS. The level-adjust control need not be included if your rig has a built-in method of adjusting the tone level. Values of R_A and R_B are not critical, but R_B should be 1.62 times R_A for proper D/A conversion.

I glued the two ICs to the circuit board in my rig (upside down) and soldered the wiring directly to the pins. This construction may be ugly, but it is fast.

My rig was made by Yaesu and has the same basic synthesizer as the FT-227 "Memorizer." This circuit will work with that rig, and should also work with any radio that has a 10-kHz synthesizer reference signal available. Total parts cost for my unit was under $3. — *Jim Harvey, WB8NBS, Redford, Michigan*

REPEATER INTERFERENCE FROM HF OPERATION

☐ I was monitoring our club repeater when it suddenly came to life with what sounded like an ssb signal. I detected what I thought was an "N" prefix just before the interference stopped. A call to Albert Lankford, N4BIT, verified that he had been operating on 3965 kHz. Further tests revealed that his hf transmitter was keying his 2-meter rig. Albert uses a Touch Tone® mike with an auto-keying circuit for his 2-meter rig. He found that placing this mike on top of the radio instead of allowing it to dangle in front of his Transmatch cured the problem. — *Alton Erdman, W4CNQ, Montgomery, Alabama*

TS-830S 60-Hz HUM

☐ While having an ssb QSO with a friend, I learned that the signal from my TS-830S had a hum. It seemed to come and go in an unpredictable manner. I checked for loose connections or a broken ground wire, but everything appeared to be okay.

Finally I tried using the MONITOR function, with headphones. A few minutes of experimenting disclosed that the hum was coming from the magnetic field generated by the power transformer. My microphone is a low-impedance type, and when the mike was within 6 inches or so of the front panel the hum was present.[1] Moving it to about 1 foot from the transceiver was sufficient to reduce the hum to a negligible level. Perhaps others who have had this problem will find a similar solution. — *James Thurston, W4PPB, Clemson, South Carolina*

[1]mm = in. × 25.4; m = ft × 0.3048.

TRACKING CATV LEAKS

☐ When a CATV leak causes interference to your Amateur Radio reception you can use your portable or mobile equipment to locate the leak. But what do you do if the interference is only in the other direction and your transmissions disrupt TV reception? There may be a simple solution to this problem, too.

Many CATV systems also carry fm broadcast stations. If the cable system in your area does, and if one of those stations is beyond car-radio range, there is indeed a simple answer. Tune your fm car radio to a station that can be heard exclusively on the cable. (Some systems also carry a warbling tone or similar signal in the fm band specifically for this purpose.) Drive along the cable while listening to the radio. As you approach a leak, the signal level will increase considerably. When you locate a source of leakage, it should be reported to the CATV system operator. Give a street address or the number of the nearest utility pole. — *Robert V. C. Dickinson, W2CCE, Berkeley Heights, New Jersey*

TRACKING DOWN A LINE-NOISE PROBLEM

☐ The San Diego Gas and Electric Co. spent several days tracking a 3570-kHz signal that was being carried for miles on the power line. After localizing the area that the signal was originating from, they began to cut power to each house in order to pinpoint the source. Inspection of the offending house turned up a bell transformer with an arcing secondary. The coil winding was just right to make a spark-gap transmitter on 3570 kHz. The signal was being coupled back to the power line through the primary winding. This signal had been causing interference for several months before it was tracked down and eliminated. — *Ed Marriner, W6XM, La Jolla, California*

INCREASED CONTROL VOLTAGE FOR THE HEATH REMOTE ANTENNA SWITCH

☐ Most commercially available remote antenna switches (and antenna rotators) are designed to be used with 100 feet or less of control cable. Heavier conductors could be used for longer cable runs, but even that may be insufficient.

Recently, Craig Wheeler, KC9T, came to me with such a problem. His tower is located some 175 feet from the shack, and his Heath SA-1480 "Remote Coax Switch" would not work properly. The voltage drop in the control line was too great, even with larger conductors.

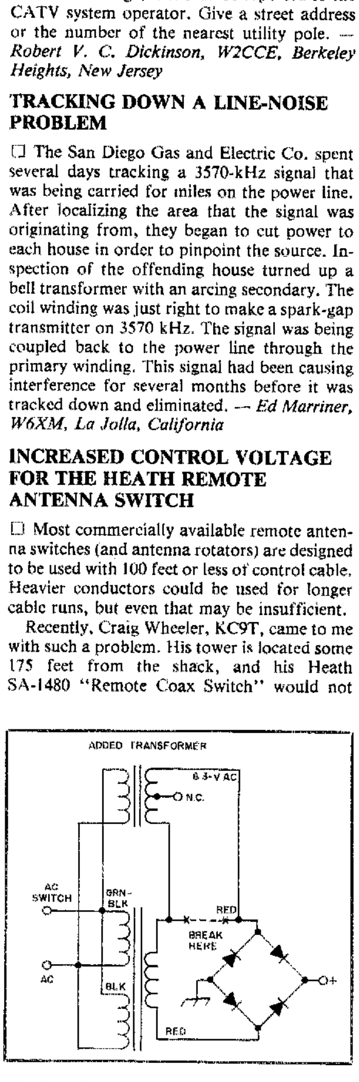


Fig. 4 — Wiring diagram for adding a transformer to increase the control voltage of a remote antenna switch to compensate for the voltage drop in a long run of control cable.

We decided to increase the power-supply voltage in the control box. We added a 6.3-V ac filament transformer in series with the original 25-V winding. Fig. 4 shows the transformer connections. Remember that the primary leads are wired in parallel and the secondary leads are wired in series. The phasing of the secondary windings is correct if the output from the bridge rectifier is about 42-V dc with no load. If the output is about 26-V dc, the phasing is reversed (wrong).

There was just enough room inside the control box to mount the additional transformer in Craig's unit. A 1- or 2-A rating on the new transformer should be sufficient. — *Lawrence Stark, K9ARZ, St. Charles, Illinois*

MICROPROCESSORS IN CARS, AND RFI

☐ I was procrastinating the job of installing an hf rig in my 1982 Buick Regal. The ritual of bonding all sections of the body and frame is one that few mobilers enjoy. In addition, the new cars with microprocessor controls could present some extra problems!

A new ICOM IC-730 provided the motivation to tackle the job. I was pleased to find that the car was clean. All of the usual shielding and bonding may not be necessary after all.

My car seemed to be developing an ignition problem, and it took a few trips to realize that the trouble occurred only when I was transmitting. The car engine would seem to miss on ssb voice peaks.

A trip to the Buick dealership led to a cure and an explanation of the problem. The service manager had seen the problem before, because he immediately pegged me as a "CBer with a kicker." I assured him that I was a licensed radio amateur and that my 100-W rig was perfectly legal.

The solution is to shield a four-wire cable that runs from the distributor to the microprocessor, grounding it at both ends. The processor in GM cars is located on the kick panel (fire wall) on the passenger side. The cable bundle leading from it contains many wires. Four of these go to the distributor, and control ignition, timing and advance. One or more of these lines was being affected adversely by the rf.

To shield these leads, I cut the four-prong connector off at the distributor and removed these wires from the larger bundle. I snaked them through a piece of braid from some old RG-11 cable and replaced the connectors. This solved my problem completely. — *David Aukamp, W4MJB, Dunedin, Florida*

STICKY VOX RELAY ON DRAKE T4X

☐ After many hours of operation, the VOX relay in my Drake T4X transmitter started to "hang up." I discovered that the armature of the relay had become magnetized. Using a tape-head demagnetizer, I slowly approached the armature with the demagnetizer turned on. I rotated the probe in a circular motion as close to the armature as possible for a few seconds. (Touching the armature will not harm anything.) Then I slowly backed the probe away, turning it off after it was clear of the transmitter. Of course, you should turn off the transmitter and unplug the power supply before performing this operation. — *John Chiuchiolo, W2LWB, Brooklyn, New York*

Technical Correspondence

Conducted By
Dennis J. Lusis,* W1LJ

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UHF TUBE OPERATING PARAMETERS

☐ Fred Merry's article, "Phase III With a Tetrode UHF Amplifier," (Aug. 1982 *QST*) is an excellent one, and the author is to be congratulated on the fine design of his amplifier. However, the operating conditions he proposes for the amplifier are ambiguous. A user of this device may be puzzled as to the correct operating parameters for the 4CX250B or 8930 tetrodes when used in grid-driven service. Tables 1, 2 and 3 contain suggested operating parameters.

Electron transit time in the vhf/uhf region causes back-heating of the cathode. Electrons leaving the cathode return and raise its surface temperature. To compensate for this, filament voltage should be dropped when the tube is used in continuous service (fm, RTTY, etc.) to reestablish the correct cathode temperature. Back-heating is less predominant during intermittent service (cw or ssb), so filament voltage may be maintained at the nominal rated value in this case. In either case, the rated heater voltage should be maintained during tune-up and standby periods.

The 4CX250B and 8930 tube types require a nominal heater voltage of 6.0, with a tolerance of ±0.3 V. It is recommended that an rms-responding (iron vane) meter be used for monitoring. For continuous service at 435 MHz, a recommended value of heater voltage that will compensate for back heating is 5.6 V. — *William I. Orr, W6SAI, Menlo Park, California*

*Assistant Technical Editor

Table 1

Operating Parameters for the 4CX250B in Class AB_1 (Ssb) Service at 435 MHz

Plate voltage	2000-V dc
Screen voltage	400-V dc
Grid Voltage[1]	−85-V dc
Zero-signal plate current	70 mA
Single-tone plate current	250 mA
Single-tone screen current (max.)	30 mA
Power output[2]	250 W

[1]Adjust to specified zero-signal plate current.
[2]Approximate value

Table 2

Operating Parameters for the 4CX250B in Class C (Cw or Fm) Service at 435 MHz

Plate voltage	2000-V dc
Screen voltage	300-V dc
Grid voltage	−90-V dc
Plate current	250 mA
Screen current[1]	10 mA
Grid current[1]	10 mA
Heater voltage[2]	5.6 V
Power output[3]	290 W

[1]Approximate; varies with drive and loading.
[2]See application note.
[3]Approximate value.

Table 3

Operating Parameters for the 4CX250R or 8930 in Class AB_1 (Ssb) Service at 435 MHz

Plate voltage	2000-V dc
Screen voltage	400-V dc
Grid voltage[1]	−80-V dc
Zero-signal plate current	70 mA
Single-tone plate current[2]	380 mA
Single-tone screen current[3]	5 mA
Power output[4]	470 W

[1]Adjust to specified zero-signal plate current.
[2]For tune-up only; maximum continuous dc plate current is 250 mA.
[3]Varies with loading and drive conditions.
[4]Approximate value

Note: The 4CX250R and the 8930 have slightly larger cathode areas than the 4CX250B, which permits a higher value of peak plate current. In addition, the 8930 has a maximum plate dissipation of 350 W, as contrasted with the 250-W rating of the 4CX250B and the 4CX250R.

ENERGY STORAGE, BANDWIDTH, AND TANK CIRCUIT Q

☐ I have just read the article by O'Hara, "The Care and Feeding of Linear Amplifiers for ATV" (August 1982 *QST*), and found it very interesting. However, there are statements under the subtitle "50-Watt Triode Amplifier" that are correct only under special conditions. I am referring to the bandwidth of an amplifier utilizing 1/4-λ versus that of a 1/2-λ plate line. If a 1/4-λ coaxial or strip-line plate circuit is designed to *minimize* stored energy, the bandwidth will be *greater* than with a 1/2-λ line. This is the opposite of what O'Hara has stated.

To minimize the stored energy in a transmission-line resonator loaded with capacitance (e.g., the output capacitance of a 2C39), the Z_o of the line should be made as high as possible. Thus, the line is made short, and there is little transmission-line capacitance to store additional energy. This principle applies to all coaxial lines, strip lines and parallel lines.

One might jump to the conclusion that in order to shorten the tuned line it could be loaded with capacitance. Unfortunately, there is no "free lunch." This technique *increases* the amount of stored energy!

For a number of reasons, the key to success in uhf output-network design is to minimize the loaded Q (Q_L) of the circuit. First, reducing the loaded Q improves the circuit efficiency.

$$\text{efficiency} = 1 - \frac{Q_L}{Q_O} \qquad \text{(Eq. 1)}$$

where
Q_L = loaded Q
Q_O = unloaded Q

Second, reducing Q_L makes circuit detuning effects caused by heat less noticeable. Improved circuit efficiency results in less circuit heating, thus causing less detuning. In addition, because of the wider bandwidth with lowered Q_L, the detuning on the resonance curve is less. The effect is somewhat of a square function.

To get wider amplifier bandwidth, you can minimize the Q_L by: (1) loading the output circuit heavily — run the tube at maximum rated plate current; (2) Not adding capacitance to the plate circuit — only the tube output C should be across the plate tuned circuit; (3) Using an L-C output matching circuit if possible — that is, an inductor that resonates with the tube output capacitance; (4) Using a 1/4-λ resonator if the L-C circuit is not feasible. Striplines with a Z_O up to 150 Ω are practical, as are coaxial circuits up to 120 Ω; and (5) Using a tuned coupling circuit (double-tuned circuit) to couple power out of the plate tuned circuit, just as O'Hara has done with his 2C39 amplifier.

As a practical example, the 2C39, with an output C of 2.01 pF, has an X_C of 183 Ω at 432 MHz. A 1/4-λ coaxial cavity can be built with a Z_O of 114 λ. This cavity could have an outer conductor ID of 5 in. and an inner conductor OD of 3/4 in.[1] It should be 58° long, which at 432 MHz is 4.40 in. The resulting amplifier bandwidth would be about 60% of that provided by an L-C circuit.

A 1/4-λ strip-line matching circuit would be very practical and could easily be made with a Z_O of 150 Ω. In this case, the line would be 50.66° long, which at 432 MHz is 3.84 in. The resulting bandwidth would be about 70% of that provided by an L-C circuit.

I have built a 432-MHz 2C39 amplifier using an L-C tank circuit. At 100-W input (3.2 W of drive), it gives an output of 60 W. This is Class C of course, not Class B (linear) service. I mention this to show that an L-C tank circuit *is* possible at 432 MHz (and would have the widest bandwidth of all).

I hope that the gist of my comments will be understood. It is evident that the concept of additional stored energy in resonant-line circuits is not well-known. — *Raymond F. Rinaudo, W6ZO, San Mateo, California*

SIMPLIFIED FORMULA

☐ The inductance formula appearing in all recent ARRL *Handbooks* appears to be taken from Terman's *Radio Engineer's Handbook* (McGraw-Hill). Although they may seem to be small points, two features of this formula disturb me: (1) the use of radius rather than the more convenient diameter, and (2) symbols that bear no evident relationship to the dimensions. The equation may be more simply expressed in this way:

$$L\ (\mu H) = \frac{d^2 n^2}{18d + 40\ell} \qquad \text{(Eq. 2)}$$

where
d = coil diameter (in.)
ℓ = coil length (in.)
n = number of turns

According to Terman, the accuracy is within 1% for coils longer than 40% of their diameter. — *Frank Noble, W3MT, Bethesda, Maryland*

IMPROVEMENTS FOR THE UPCONVERTING RECEIVER

☐ Helfrick's article, "A Modern Upconverting General-Coverage Receiver," (December 1981 *QST*), was of particular interest to me. I have long been interested in

[1]mm = in. × 25.4.

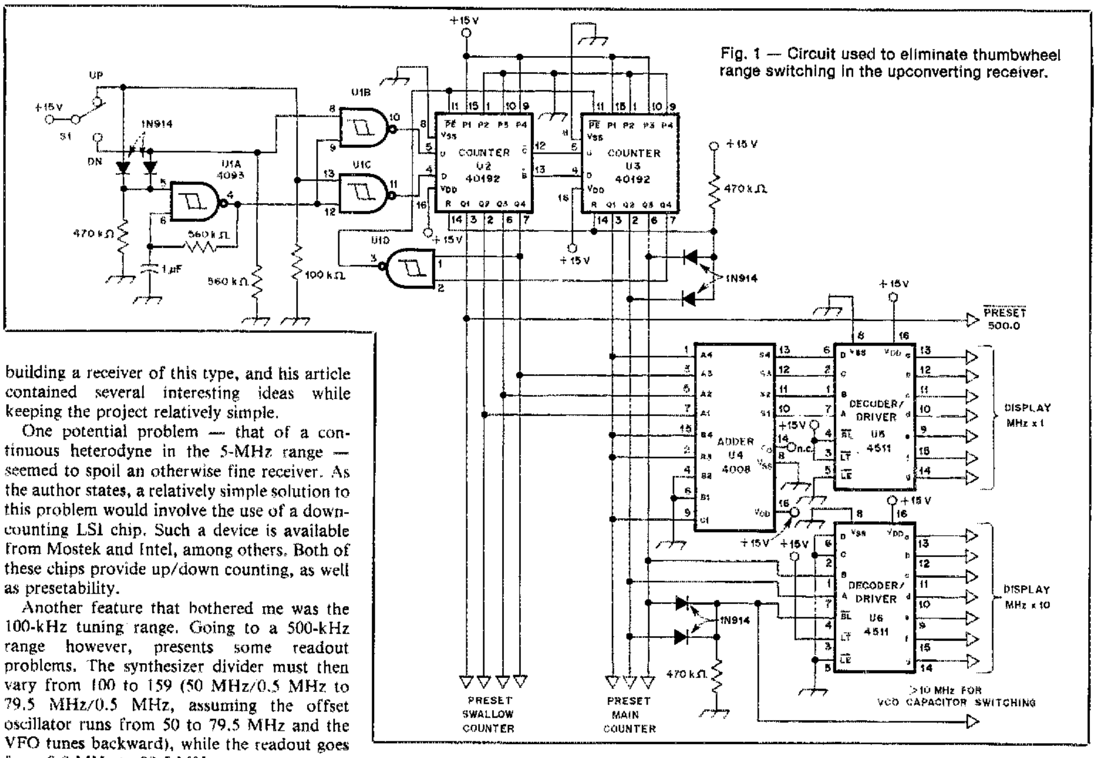

Fig. 1 — Circuit used to eliminate thumbwheel range switching in the upconverting receiver.

building a receiver of this type, and his article contained several interesting ideas while keeping the project relatively simple.

One potential problem — that of a continuous heterodyne in the 5-MHz range — seemed to spoil an otherwise fine receiver. As the author states, a relatively simple solution to this problem would involve the use of a down-counting LSI chip. Such a device is available from Mostek and Intel, among others. Both of these chips provide up/down counting, as well as presetability.

Another feature that bothered me was the 100-kHz tuning range. Going to a 500-kHz range however, presents some readout problems. The synthesizer divider must then vary from 100 to 159 (50 MHz/0.5 MHz to 79.5 MHz/0.5 MHz, assuming the offset oscillator runs from 50 to 79.5 MHz and the VFO tunes backward), while the readout goes from 0.0 MHz to 29.5 MHz.

To eliminate the thumbwheels for range-switching, it is relatively simple to provide an up/down counter from which the outputs preset the synthesizer counter. An examination of these outputs reveals a solution to the readout problem. It will be noted that only two decades are required; the hundreds divider may be hardwired to preset to one.

According to Table 4, we may take the least significant bit as the indicator to preset the LSI counter to 500.0 (remember the VFO is tuning backward, so this occurs for the even megahertz ranges). The next four bits (BCDA') will produce the unit megahertz if three is subtracted from the binary number whenever bit A' is one. Bits B'C' indicate the tens of megahertz. The circuit in Fig. 1 shows how this may be done.

Schmitt trigger U1A forms an astable multivibrator with a period of approximately 0.5 s. This time may be adjusted to suit individual preferences. Up/down counting is accomplished with S1, a momentary-contact spdt switch. The RESET and PRESET inputs of the two 40192 counters are used to limit the counting range from 00 to 59. The connections shown to the 4008 adder (U4) will subtract 3 each time bit A' is one. — *T. A. Bergstrom, W1IQW/DJ0QT, Munich, Federal Republic of Germany*

Table 4

Up/Down Counter Sequence

Count	*LSD* ABCD	*MSD* A' B' C' D'	*R/O*	*Count*	*LSD* ABCD	*MSD* A' B' C' D'	*R/O*
00	0000	0 0 0 0	00.0	.	. . . .	. . . .	.
01	1000	0 0 0 0	00.5	.	. . . .	. . . .	.
02	0100	0 0 0 0	01.0	43	1100	0 0 1 0	21.5
03	1100	0 0 0 0	01.5	44	0010	0 0 1 0	22.0
04	0010	0 0 0 0	02.0	45	1010	0 0 1 0	22.5
05	1010	0 0 0 0	02.5	46	0110	0 0 1 0	23.0
06	0110	0 0 0 0	03.0	47	1110	0 0 1 0	23.5
07	1110	0 0 0 0	03.5	48	0001	0 0 1 0	24.0
08	0001	0 0 0 0	04.0	49	1001	0 0 1 0	24.5
09	1001	0 0 0 0	04.5	50	0000	1 0 1 0	25.0
10	0000	1 0 0 0	05.0	51	1000	1 0 1 0	25.5
11	1000	1 0 0 0	05.5	52	0100	1 0 1 0	26.0
12	0100	1 0 0 0	06.0	53	1100	1 0 1 0	26.5
13	1100	1 0 0 0	06.5	54	0010	1 0 1 0	27.0
14	0010	1 0 0 0	07.0	55	1010	1 0 1 0	27.5
15	1010	1 0 0 0	07.5	56	0110	1 0 1 0	28.0
.	. . . .	. . . .	.	57	1110	1 0 1 0	28.5
.	. . . .	. . . .	.	58	0001	1 0 1 0	29.0
.	. . . .	. . . .	.	59	1001	1 0 1 0	29.5

Feedback

☐ Please make this correction to Fig. 6 of "Electrical Antenna Null Steering," October 1982 *QST:* The 343-pF capacitor should be labeled C3, and the 487-pF unit is C5. The series resonance of the L1/C2 combination is 1.025 MHz, and that of L2/C4 is 1.604 MHz. Author Webb also points out that L1 and L2 consist of 24 and 25 turns, respectively, of no. 26 wire on Micrometals T37-2 or T50-6 cores. In Fig. 5, no. 32 wire may be used for T1. C1 through C4, inclusive, should be increased in value to at least 1000 pF for 3.5-4 MHz operation, and to 2000 pF for 1.8-2 MHz operation. The four capacitors should be of equal nominal value.

☐ Dick Schellenbach, W1JF, has informed us of an error in Fig. 3 of November 1982 Technical Correspondence. Pin 4 of the NE570 alc chip should be grounded, rather than not connected (as shown).

☐ The March 1982 installment of Technical Correspondence indicates that copies of government regulations on phone-line interfacing are available from FCC. Dick Richardson, N8CDH, has indicated that these regulations are now out of print, and are available only at your library.

☐ Fig. 2 of the H. Granberg power FET article, Part 1, for Dec. 1982 *QST*, contains an error. The submitted drawings had decimal points rather than commas in the filter-capacitor designations for power supply no. 2. The correct value is 13,500 µF, rather than the 13.5-µF value indicated.

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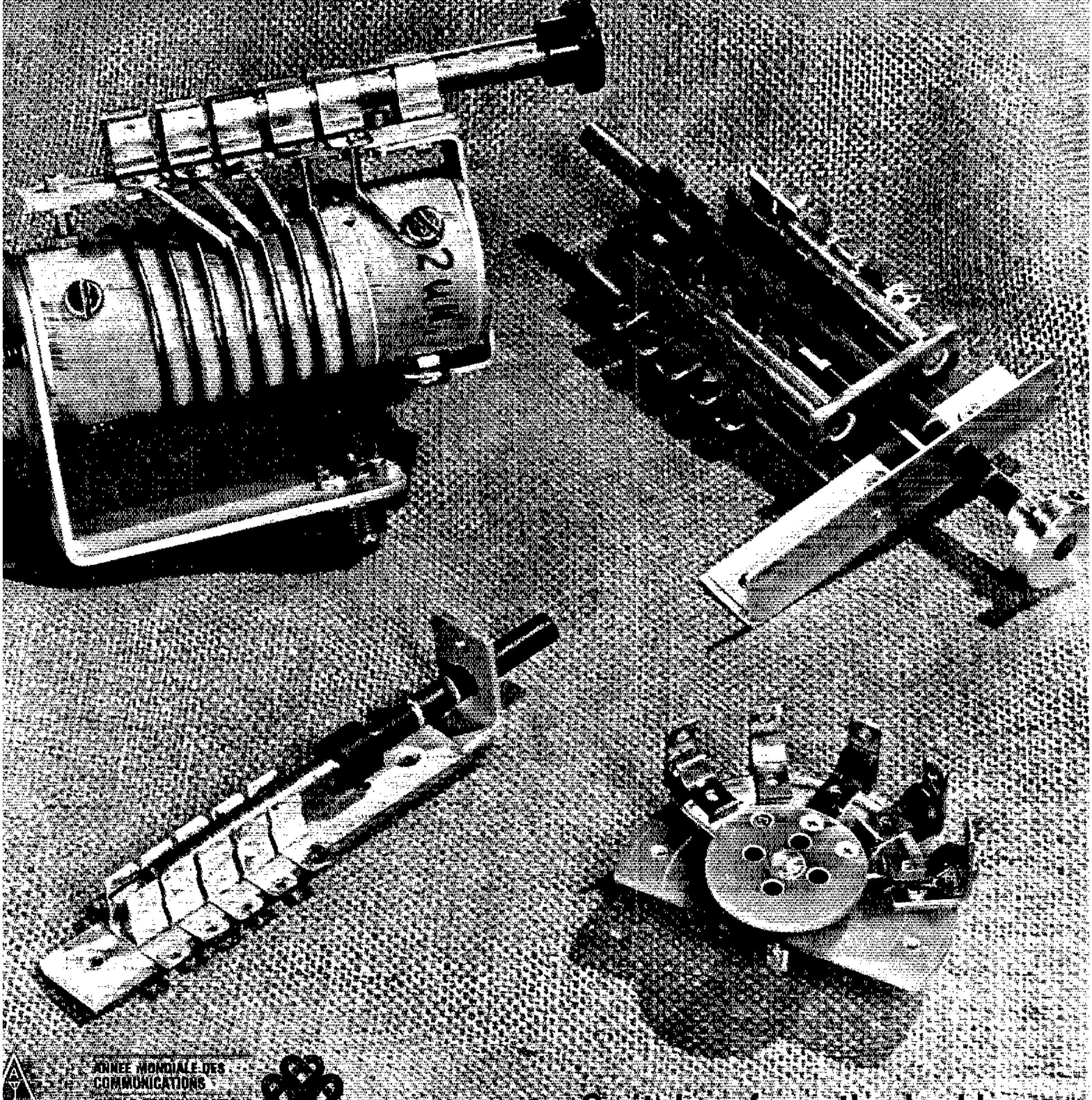

ANNEE MONDIALE DES COMMUNICATIONS
WORLD COMMUNICATIONS YEAR
AÑO MUNDIAL DE LAS COMUNICACIONES
1983

Switches from the junkbox

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We've all built an antenna or two, but how many hams have put together an assortment of switches? W7KBE tells us why he'd rather do it himself! See page 18.

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Amateur Capsule 1982

This was the year the League, under new leadership, gave new vitality to ARRL field volunteers, a landmark Amateur Radio bill became law, and U.S. and Canadian amateurs finally got on a new WARC band.

By Andrew Tripp,* KA1JGG

The year 1982 will be remembered for many things, but perhaps first and foremost will be the changes that occurred — events that undoubtedly will have a tremendous impact on Amateur Radio for years to come. In March, the Board of Directors selected a new president and a new general manager. At the same time, ARRL began laying the groundwork for the reorganization of its volunteer field organization. From section-level appointments to club affiliation, roles and responsibilities would be redefined and expanded, making the field organization more effective and developing closer ties between local amateurs and Hq. Coupled with this, ARRL Advisory Committees took on a new look, with increased membership and better representation, giving them a greater voice in League planning.

With the signing into law of a long-awaited "Amateur Radio bill," amateurs gained a valuable foothold in their efforts to solve some nagging problems, such as RFI, and a more active role in shaping the future of ham radio. Amateur Radio gains from WARC-79 began to bear fruit with FCC approval for limited use of the 10-MHz band, followed by Senate ratification of the treaty in the waning weeks of 1982. On the state and local levels, amateurs continued to tackle restrictive antenna ordinances, interference problems and leaky cable-television systems.

Internationally, a new president took the helm of the IARU, amid plans to restructure the 57-year-old organization. ARRL added a staff member to help the IARU Secretary expand and improve activities on behalf of amateurs worldwide. And China, absent for many years, was back on the air — testimony that Amateur Radio is alive and well in that country, but developing ever so slowly.

1982 was not without controversy. But despite these issues — or perhaps because of them — optimism prevailed, and there was widespread confidence that the winds of change would bring with them a breath of fresh air. Evidence of this was the marked increase in on-the-air activity, with public service and contesting leading the way, and the renewed interest of local amateurs to take part in — not just be a passive part of — Amateur Radio history. It promised to be a banner year; amateurs were not disappointed.

Setting Up Housekeeping

At its semiannual meeting in March 1982, the ARRL Board of Directors chose a new president and a general manager, setting the stage for the year's events. Vic Clark, W4KFC, was elected president, succeeding Harry Dannals, W2HD, who stepped down from that volunteer position after serving for the past 10 years. Harry retired to private life, but is certain to remain active in Amateur Radio affairs. Vic, an ARRL vice president for six years and an active participant at the 1979 ITU World Administrative Radio Conference (WARC-79), brought a wide breadth of experience to his new role.

ARRL past President W2HD (left), incoming General Manager K1ZZ (center) and newly elected President W4KFC at the March Board meeting.

For ARRL general manager and secretary, the Board selected David Sumner, K1ZZ, to fill vacancies created by the retirement of Richard L. Baldwin, W1RU, from paid employment at ARRL Hq. General manager since 1975, Dick began his employment at Hq. in 1948. He had the key staff role and was the "executive officer" for the ARRL at WARC-79.

Dave, assistant general manager since 1976, has worked at ARRL Hq. since 1972 and was an active volunteer from the time of his first license at age 13, in 1962. Also a member of the Amateur Radio team in Geneva, Dave has traveled extensively on behalf of the Amateur Radio Service.

In other Board action, W1RU was elected vice president for International Affairs, replacing Noel Eaton, VE3CJ, who retired. IARU member-societies later approved Dick's candidacy for IARU President.

With personnel matters decided, the League began setting in motion its two-year plan to breathe new life into ARRL sections. With recommendations from the Long-Range Planning Committee, ARRL initiated a major overhaul of the section-level organization. This reorganization (which began on January 1, 1983) is the most significant change in the section-level structure in more than 50 years.

*QST Features Editor

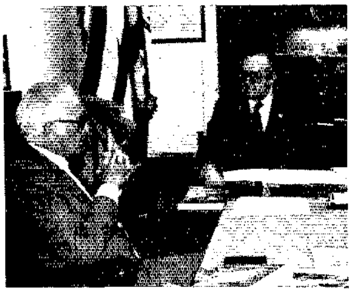
Traditionally, section-level responsibilities have covered on-the-air activities, particularly traffic handling and emergency communications. This role has now been expanded into other areas, such as public relations and government liaison. In addition, the elected post of Section Communications Manager (now called Section Manager) has been redefined and given greater authority to delegate responsibility to other volunteers.

To augment this restructuring, the Board adopted new rules and regulations for ARRL Advisory Committees. The committees now are organized on the basis of one member per division per committee, a departure from the one-member-per-call-area representation of the past. Also, the Special Services Club program, approved by the Board in 1982 and scheduled for implementation in 1983, will improve amateurs' effectiveness on the local level.

Several specialty newsletters saw their debut in 1982. Among them was *New Ham News,* an updated version of a publication that first appeared in the mid 1970s. Sporting a lively new format, *NHN* is mailed automatically to newly licensed amateurs. In late October, the first issue of the *ARRL Letter,* a biweekly newsletter designed to disseminate late-breaking Amateur Radio news in a timely manner, appeared. True to its purpose, the first issue announced our newly acquired access to the 10-MHz band.

Capitol Hill: A Landmark Law

On September 13, 1982, radio amateurs nationwide had good reason to rejoice: President Reagan had signed the Goldwater-Wirth Amateur Radio bill into law, capping a 10-year struggle by ARRL to get federal action on solving radio-frequency interference. Perhaps the most significant federal legislation for radio amateurs since the Communications Act of 1934, P.L. 97-259 amends the Act to give FCC the authority to (1) establish RFI-rejection standards for home electronics equipment; (2) enlist the aid of volunteers in monitoring the amateur airwaves for rules violations, and in preparing and administering amateur examinations; and (3) extend the license term to 10 years. The law also exempts amateur communications from the secrecy provisions of the Communications Act. Ratification of the Radio Regulations and Final Protocols, Geneva, 1979 (the formal name for the WARC results), remained stalled in the Senate for most of 1982. But things took a turn for the better when, on October 28, FCC approved limited use of the 10-MHz band — one of three new hf bands allocated for amateur use at WARC-79. Then, On December 21, the Senate ratified the Final Acts, thus removing the last major hurdle for implementation in the U.S. of the WARC-79 agreements.



ARRL Washington Area Coordinator Perry Williams, W1UED (right), meets with Senator Barry Goldwater, K7UGA, coauthor of the Amateur Radio bill signed into law.

Regulatory Matters

Cable-television interference dominated the regulatory scene in 1982. In January, ARRL filed a formal rule-making request (RM-4040) calling for the prohibition of cable-television operation on frequencies allocated to the Amateur Radio Service, particularly on the 144- and 220-MHz bands, where much amateur public service is performed. ARRL pointed out that cable interference is the direct result of inadequate shielding, poor-quality components and inferior installation procedures of cable companies. In calling for the action proposed in RM-4040, ARRL said the problem of CATVI cannot be resolved in the context of the present rules.

ARRL's action stirred the National Cable Television Association, which requested a meeting with ARRL officials. ARRL agreed to an extension of the deadline for comments on RM-4040, provided that NCTA would exhort its members to clean up their cable system operations. NCTA had until September 1, the new comment deadline, to show some improvement, but was unable to do so. ARRL followed with Comments on its own petition, reiterating the League's position that there be "adequate safeguards in the form of Federal regulation and enforcement so that legitimate users of the radio spectrum are protected from the insidious and detrimental effects of radio frequency interference both from and to these (CATV) ostensibly closed systems."

NCTA countered with its claim that the interference problem is "not as great as ARRL would have the Commission believe," and called for the dismissal of the ARRL request to ban cable operations from frequencies allocated for amateur use. ARRL, in its reply, urged the FCC to act favorably on its petition and to provide the relief sought on either a blanket basis or "on a case-by-case basis automatically upon receipt of a verified complaint of interference to or from an amateur station." In the meantime, the ARRL RFI Task Group continued in its role as legislative and regulatory watchdog, monitoring actions on Capitol Hill and advising the Board on RFI matters, particularly CATVI.

In what was to turn out to be one of the controversial issues of 1982, FCC instructed its staff to prepare and release to the public proposals for the creation of a new, no-code amateur license. This action prompted immediate protest from the amateur community. After much discussion on a number of alternatives, the Commission directed the Private Radio Bureau to prepare a Notice of Proposed Rule Making (NPRM) on two choices: a codeless Technician class license, and a Canadian-style digital license.

Meanwhile, ARRL was on record as opposing a no-code license. At its meeting on November 20, the ARRL Executive Committee decided that, ". . . in light of the impending transfer of amateur license examination responsibilities to the amateur community and its attendant workload, the possibility of deferring for at least 18 months the issuance by FCC of a Notice of Proposed Rule Making seeking to establish a new codeless license be explored with the Commission." Unless the FCC heeds ARRL's request, the NRPM should be released early in 1983.

On October 28, FCC delighted the amateur community by releasing temporarily the 10.100-10.109 and 10.115-10.150 MHz portions of the 10-MHz band for General, Advanced and Extra Class licensees. The band is available on a secondary, noninterference basis. The 10.109-10.115 MHz segment was not made available to amateurs because of its daily use by a priority government radio service.

ARRL had urged the FCC to make the frequencies available to U.S. amateurs as quickly as possible, arguing that amateurs would better be able to fulfill their public service obligations to foster international goodwill and to enhance their ability to provide reliable communications. FCC agreed with ARRL's objectives, saying, "Although the ratification process for the Final Acts of WARC-79 is not yet complete, we believe that the public interest supports amendment of our rules to allow amateur operators in the United States temporary use of the frequencies."

In other action, the FCC issued an NPRM in response to an ARRL proposal to expand the 20-meter telephony subband, in an attempt to ease serious overcrowding. In general, the idea of expanding the subband from 14.200-14.350 to 14.150-14.350 MHz received a favorable response from the amateur community, although the FCC version varied slightly from the ARRL plan. For General class operators, the ARRL argued, the FCC proposal would split the frequencies available, thereby creating two sets of band edges and increasing the possibility of rules violations. Instead, the ARRL contended, its original plan (reported in April 1981 *QST*) would be the best course of action: the lower 25 kHz for Extra, the next 50 kHz for Advanced and Extra, and the upper 125 kHz for General, Advanced and Extra, keeping the bands contiguous.

In a related item, ARRL also commented on an FCC Notice of Inquiry (NOI) regarding the future expansion of hf telephony subbands on 80, 40, 15 and 10 meters. League comments, based on member input solicited in *QST* last April and May, said that the additional expansion (1) for the most part, would *not* have a detrimental effect on domestic telegraphy operations and (2) should be contiguous with existing subbands. The League urged early FCC issuance of an NPRM on these proposals.

Effective October 28, FCC permitted the use of any digital code on amateur frequencies above 50 MHz, except those frequencies on which only A1 emission is permitted (50.0-50.1 and 144.0-144.1 MHz). As early as October 1980, ARRL had petitioned FCC for a rules change to permit greater experimentation with digital techniques, citing a "compelling need" if radio amateurs are to continue as the primary contributors to the advancement of the radio art. FCC's major concern in allowing the experimental frequencies was its enforcement capability. However, the inclusion of special provisions and amateurs' desire to protect their frequencies against unauthorized use, FCC agreed, would provide adequate protection against misuse of the frequencies.

By year end, proposals to cut some logging requirements and to change the rules for maximum authorized power were introduced. In addition, FCC approved new beacon rules. Most important was authorization of automatic control on specific frequencies for amateur stations making one-way beacon transmissions to detect unusual propagation conditions and to check out and adjust receiving equipment.

Legal Actions

State and local governments continued to introduce restrictive antenna ordinances, in some cases severely hindering amateur activities. With ARRL support, local amateurs challenged these restrictions.

In Burbank, Illinois, amateurs filed a class-action suit against the city. Backed by ARRL-pledged financial support, James C. O'Connell, W9WU, the plaintiffs' attorney, sought a declaration by the Court that the city's ordinance restricting antenna height is an unconstitutional abridgement of their rights. The complaint also asked for an order temporarily restraining the City of Burbank from bringing any legal action against anyone erecting or maintaining an amateur or CB antenna.

An important case is shaping up in Oklahoma City, where a ruling by a federal district judge in 1982 opened the door for a legal precedent favorable to Amateur Radio. Charles M. Guschke, N5SW, sued the city in June 1981, charging its zoning ordinances were too restrictive in not allowing him to maintain his 78-foot-high antenna tower. ARRL filed a friend-of-the-court brief in support of Guschke's case, emphasizing that federal interests in Amateur Radio were being thwarted by overly restrictive Oklahoma City ordinances. In denying the city's motion to dismiss the case for lack of federal jurisdiction, the judge ruled that the suit filed by Guschke "states a substantial federal question of the application of the Communications Act of 1934. . ." The judge's ruling means that Guschke will have a trial on the merits of his claim that Oklahoma City has deprived him of his constitutional rights of free speech and privacy, and of his liberty and property without due process of law.

Communications Highlights

Public service communications took the lead in 1982 when, in January, radio amateurs in Washington, DC were suddenly called into action. A commercial airliner, just departed from National Airport, had crashed into the rush-hour traffic on the 14th Street Bridge and then into the frozen Potomac River. Amateurs' response was swift, as they provided communications for rescue and salvage operations for four bone-chilling days.

In the same month, ARES groups in Alabama and Mississippi were called out when those states were besieged by more than a week of severe weather, beginning with heavy rains, continuing with a tornado and ending with heavy snowfall. Similar weather ripped through northern California in March, again generating exemplary public service from amateurs in the face of extreme conditions. The year closed out with outstanding amateur action during Hawaii's hurricane Iwa.

Field Day 1982 continued its tradition of being the most popular operation of the year. The technology may have changed over the years, but the purpose has remained the same: to test radio amateurs' preparedness to deal with any situation during an emergency. Amateurs came out in droves again this year, with nearly 30,000 participating in the 46-year-old event.

Carl Henson, WB4ZNH, sparked quite a controversy among DXers when he charged that operating from a "list" or "DX net" is poor operating ethics, and attempted to get DXCC Rule 12 amended to say so. His correspondence campaign to hundreds of DX enthusiasts prompted an avalanche of letters from a polarized DX community, thus catching the attention of the ARRL DX Advisory Committee.

In 1981, the VHF/UHF Ad Hoc Contest Committee was established to investigate and make recommendations on how to better meet the objectives of v/uhf contesting. In February, and in an update in November, the Committee reported on its latest actions. The availability of a single-band award, first seen in the June ARRL VHF QSO Party, proved to be an overwhelming success. This was followed by the moving of the EME contest to more favorable dates and the institution of scoring to reward more-distant contacts in the August UHF Contest. The Committee closed out the year considering a host of other matters, among them the ultimate best time frame for each contest and the use of grid squares for both contest and award purposes.

At ARRL Hq., W1AW invested further in the computer age with the installation of a second Heath H89 computer; the results can be heard on any of the scheduled W1AW transmissions. All W1AW code practice and cw, Baudot and ASCII bulletins are now computer generated.

Rounding out communications activity in 1982 was an operating record. A new 10-GHz DX record was set by Nicola Sanna, I0SNY, who worked I0YLI and IW0BFZ at distances of 1101 and 1166 km, respectively. These are the first-ever contacts over 1000 km on 10 GHz.

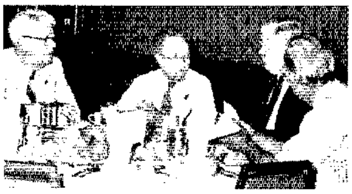
Technical Developments

While other amateur concerns were grabbing the headlines, amateur technology also moved along at an industrious pace, improving on old techniques and breaking new ground. Throughout the year, *QST* reported on several such developments, publishing an article on a power MOSFET kW amplifier, believed to be the first on the subject to appear in an amateur publication; articles on converting existing equipment to cover the new WARC bands; a construction project for a high-powered amplifier for the 900-MHz band, in anticipation of the future use of that band; and new articles by Beverage and Kraus, two authors prominent for decades in antenna research and development.

Meanwhile, others, such as the Amateur Radio Research and Development Corporation (AMRAD) in Washington, DC, were exploring other frontiers. With temporary authorizations from FCC in hand, these experimenters were busy making household words of such new amateur communications techniques as AMTOR and spread spectrum, exemplifying the amateur pioneer spirit.

International Affairs

In August, the International Amateur



One last time together (l-r): IARU President VE3CJ, General Manager W1RU, ARRL President W2HD and Vice President WØBWJ at the March Board meeting.

Radio Union took on a new president and the beginnings of a new structure. Noel Eaton, VE3CJ, the sixth IARU president and the first from outside the U.S., retired and was succeeded by Richard L. Baldwin, W1RU, retired ARRL general manager and the new Vice President for International Affairs. Noel's distinguished career as IARU president, which began in 1974, includes heading the IARU team in Geneva, Switzerland, for WARC-79.

During his term as president, he formed a Restructuring Committee to examine the IARU Constitution and make recommendations on how it could better serve the membership. This year, after three years of study and membership contact, the Restructuring Committee found a common view that IARU policies ought to be decided by an Administrative Council, comprised of officers of IARU and two representatives from each of the three Regions. The concept was put to the Union membership in June and was adopted in November.

Throughout the year, representatives of IARU Regions 1, 2 and 3 attended meetings at various locations around the world. Among the subjects participants discussed or acted on were the election of new directors for the Region 3 Association, implementation of the three new WARC bands, and IARU constitutional reform.

The big international operating news of 1982 has to be the reappearance of China on the amateur airwaves. Long dormant, China took a giant step toward active participation in international Amateur Radio when, on March 29, on-the-air operations began from BY1PK. Although Amateur Radio is now a club-station activity and no individual licenses were issued during the year, amateurs are confident that, in time, Amateur Radio in China will expand.

The Canadian contingent to the March Board meeting (l-r): IARU past President VE3CJ, Director VE3OT and Counsel VE2VW.

CRRL Matters

Citing an unexpectedly heavy schedule and a feeling that he would be unable to do justice to both his job and his League responsibilities, Mitch Powell, VE3OT, resigned as Canadian Division Director after three years in that position. Succeeding him was Thomas B. J. Atkins, VE3CDM, who became Canadian Division Director, effective September 20. Tom had served as vice director since 1978 and was the Canadian representative at the IARU Region 2 Conference in Lima, Peru, in 1980. Harry MacLean, VE3GRO, was appointed by ARRL President Clark as the new Canadian Division vice director. Harry is the conductor of Canadian NewsFronts in *QST*, and has been active in CRRL and ARRL affairs.

In October, several changes to the CRRL By-Laws were approved in a General Meeting in Waterloo, Ontario. Among them were (1) the addition of two new directors to the CRRL Board, (2) the CRRL president and the ARRL Canadian Division director becoming one and the same, and (3) the CRRL vice president and the ARRL Canadian vice director becoming one and the same.

On May 21, DOC released the 10.1- to 10.15-MHz band to Canadian amateurs. Amateurs are the primary users of the band in Canada, but DOC is treating amateurs as secondary users from an international standpoint. A1 emissions for Amateur class licensees and A1 and F1 for Advanced amateurs are permitted. DOC did not indicate any power restrictions or whether Amateur class licensees with "10-meter endorsements" would be permitted to use F1 emissions; CRRL requested clarification of these points.

Satellite News

Amateurs' high expectations for UoSAT-OSCAR 9 were temporarily allayed when, in April, a command-system glitch left both of the satellite's telemetry beacons gated on, causing UO-9, in effect, to QRM itself to "deaf." The UoSAT team at Surrey, England, tried to solve the problem, but to no avail. If nothing could be done, UO-9 would be useless.

In September, after repeated attempts to get UO-9 to respond to commands, a team of amateurs at SRI International, Menlo Park, California, took a shot at the bird. With 42-dBd gain of a 150-foot dish antenna pointed skyward, the SRI team blasted about 10 MW of 70-cm erp toward UO-9. The beacons fell silent; UoSAT-OSCAR 9 had been saved!

During the year, radio amateurs at AMSAT were busy putting the finishing touches on another spacecraft — AMSAT Phase IIIB, the latest in amateur satellites. Having passed its final testing, Phase IIIB now awaits a launch, scheduled for April 1983, aboard an ESA Ariane rocket.

Beginning with the December issue of *QST*, experienced amateur satellite enthusiasts and newcomers had a new source of information in a new monthly column, Amateur Satellite Program News. Using news gleaned from a variety of sources, ASPN will give readers a monthly update on amateur satellite activities.

Silent Keys

Reported along with the year's events was the passing of many memorable radio amateurs, enthusiasts who have contributed so much to the Amateur Radio Service. Among them was Francis Edward Handy, W1BDI, affectionately known by his ham community as "Mr. Amateur Radio." His 42-year service to ARRL and Amateur Radio is peppered with achievements, which include the writing of countless *QST* articles and the first two editions of *The Radio Amateur's Handbook* ("Handy's Handy Handbook"), and the founding of Field Day, the ARRL section-level field organization and other on-the-air activities and traditions.

Robert M. Booth, Jr., W3PS was another silent key during 1982. Bob's most illustrious contributions to ARRL and Amateur Radio were accomplished through his role as general counsel, a position in which his optimistic outlook, eye for detail and stick-to-itiveness were often called upon to turn the tide of a regulatory battle in favor of Amateur Radio.

The list goes on with such names as Jesse Bieberman, W3KT, an active amateur for more than 60 years, many of which were as a director or vice director of the ARRL; Morton B. Kahn, K4KR/W2KR, Hudson Division director (1958-64) and an SCM in the 1920s; Clark C. Rodimon, W1SZ, managing editor of *QST* prior to WW II; Dr. James L. Lawson, W2PV, noted *QST* author and DXer; Philip E. Haller, W9HPG, Central Division director for 13 years and long-time EC for Cook County; and Richard C. Spenceley, KV4AA, famous DXer.

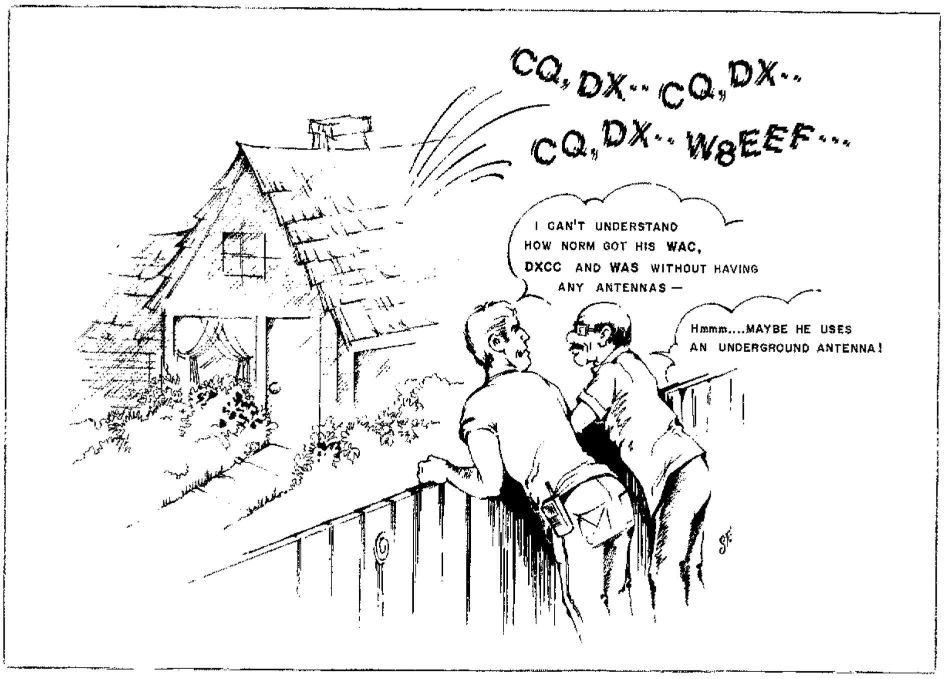
Looking Ahead

Much happened in 1982, and amateurs can take pride in the gains they have made. But the surface has only been scratched. The new Amateur Radio law is only a beginning; now comes the task of putting its provisions to work. In the coming months, section-level reorganization will get into full swing; amateurs' vigorous support and participation is needed if it is to succeed. Many pressing regulatory and legal issues need to be examined and resolved, and new technological advances must be developed and explored. If the past year's events are any indication of what is to come, Amateur Radio, with ARRL in the lead, is indeed embarking on a new era.

Antennas for Those Who Can't Have Antennas!

Radio amateurs don't engage in covert activities, but there are times in all of our lives when hidden or "invisible" antennas are necessary if we are to get on the air.

By Doug DeMaw,* W1FB



The unfortunate fact of the matter is that some radio amateurs dwell where antennas are prohibited. In other situations the operator may not want to erect outdoor antennas for fear of neighborhood opinions that he or she is destroying the beauty of the residential area. We amateurs don't regard our antennas as eyesores; in fact, we almost always regard them as works of art! But there are occasions when having an outdoor or visible antenna can present problems, especially for those who live in apartments.

When we are confronted with restrictions, self-imposed or otherwise, we can take advantage of a number of options toward getting on the air and radiating at least a moderately effective signal. In this context, a poor antenna is certainly better than no antenna at all! A number of techniques enable us to use indoor antennas or "invisible" antennas out of doors. Many of these systems will yield good to excellent results for local and DX contacts, depending on band conditions at any given time. *Don't erect any antenna that can present a hazard (physical or electrical) to humans, animals or buildings. Safety first!*

Invisible Antennas

In some areas, clotheslines are attached to pulleys (Fig. 1) so that the user can load the line and retrieve the laundry from a back porch. Laundry lines of this variety are accepted parts of the neighborhood "scenery," and can be used handily as amateur antennas by simply insulating the pulleys from their support points. This calls for the use of a conducting type of clothesline, such as heavy-gauge stranded electrical wire with Teflon or vinyl insulation. A high-quality, flexible steel cable (stranded) is suitable as a substitute if one doesn't mind cleaning it each time clothing is hung on it.

A jumper wire can be brought from one end of the line to the ham shack when the station is being operated. If a good electrical connection exists between the wire clothesline and the pulley, a permanent

*Senior Technical Editor, ARRL

connection can be made by connecting the lead-in wire between the pulley and its insulator. A Transmatch can be used to match the "invisible" random-length wire to the transmitter and receiver.

Invisible "Long Wire"

In reality, an antenna is not a classic "long wire" unless it is one wavelength (or greater) long. Yet, many amateurs refer to all relatively long spans of conductor as "long wires." For the purpose of this article we will assume we have a fairly long span of wire, and refer to it as an "end-fed wire."

If we use small-diameter enameled wire for our end-fed antenna, chances are that it will be very difficult to see against the sky and neighborhood scenery. The higher the wire gauge, the more "invisible" the antenna will be. The limiting factor with very fine wire is fragility. A good compromise can be realized by using no. 24 or no. 26 magnet wire for spans up to 130 feet (m = ft × 0.3048). Lighter-gauge wire can be used for shorter spans, such as 30 or 60 feet. The major threat to the longevity of fine wire is icing; also, birds may fly into the wire and break it. Therefore, this style of antenna may require frequent service or replacement.

Fig. 2 illustrates how we might install an invisible end-fed wire. It is important that the insulators also be lacking in prominence. Tiny Plexiglas blocks work well, as do small-diameter, clear plastic medical vials. Some amateurs simply use rubber bands for end insulators, but they will deteriorate rapidly from sun and air pollutants. They are entirely adequate for short-term operation with an invisible antenna, however.

The invisibility of the antenna can be carried even further if one is willing to use camouflaging techniques. This can be achieved by spraying the antenna wire with green, tan, brown, black and light blue paint at 1-foot intervals. In some instances, a single layer of gray or medium-blue paint will help to disguise the antenna. The wire must be free of grease and dirt if paint is applied, and the paint should be of "exterior" grade. This camouflaging effect can also be realized by dipping sections of the wire into cans of paint of the appropriate colors, assuming that spray paint is not available or desired.

Rain-Gutter or TV Antennas

A great number of amateurs have taken advantage of standard house fixtures when contriving inconspicuous antennas. A very old technique is the use of the gutter and downspout system on the building. This can be seen in Fig. 3: A lead wire is routed to the shack from one end of the gutter trough. We must assume that the wood on which the gutter is affixed is dry and of good quality in order to provide a reasonable insulation factor. The

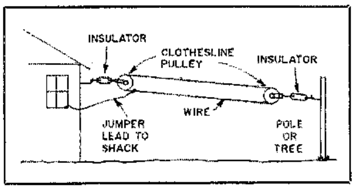


Fig. 1 — One form of hidden antenna is shown in this drawing: The antenna performs double duty as a radiator and a wash line. Large-diameter insulated wire (flexible) can be used as the line, or a clothesline with a wire center may be employed. The pulleys are insulated from the house and the pole.

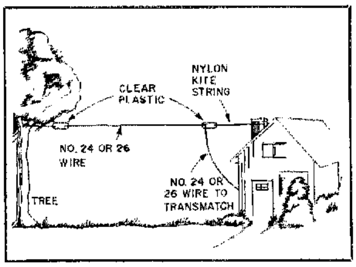


Fig. 2 — Light-gauge enamel-covered wire works nicely as a hard-to-see antenna. Although a bird may knock it down from time to time, re-erecting it beats having no antenna at all!

rain-gutter antenna may perform quite poorly during wet weather or when there is ice and snow on it and the house roof.

We need to ensure that all joints between gutter and downspout sections are bonded with straps of braid or flashing copper to provide good continuity in the system. Poor joints can cause rectification and subsequent TVI and other harmonic interference. Also, it is prudent to insert a section of plastic downspout about 8 ft above ground. This will prevent humans from receiving rf shocks or burns while the antenna is being used. Improved performance may result if the front and back gutters of the house are joined by a jumper wire to increase the area of the antenna.

Fig. 3 also shows a TV or fm antenna that can be employed as an invisible amateur antenna. Many of these antennas can be modified easily to accommodate the 144- or 220-MHz bands, thereby permitting the use of the 300-Ω line as a feeder system. Some fm antennas can be used on 6 meters by adding no. 10 bus-wire extensions to the ends of the elements and adjusting them for a VSWR of 1:1. If 300-Ω line is used it will require a balun or Transmatch to interface the line with the station equipment.

For operation in the hf bands we can tie the TV- or fm-antenna feeders together at the transmitter end of the span and treat the overall system as a random-length wire. If this is done, the 300-Ω line will

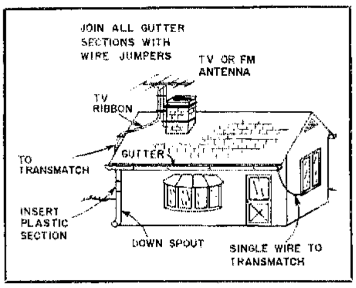


Fig. 3 — Some amateurs have had success when using the rain gutter as a random-length, end-fed radiator. The lower portion should be insulated from the remainder of the gutter and downspout to prevent rf hazards to animals and people. The TV or fm antenna-lead wire also can be used as a random-length antenna, as shown here.

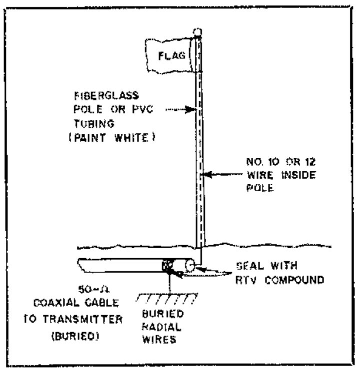


Fig. 4 — One can be patriotic while using a disguised antenna. The flagpole also serves as a vertical antenna in this illustration. Insulation should be used over the lower part of the antenna to prevent shock hazard to people and animals.

have to be on TV standoff insulators and spaced well away from phone and power company service-entrance lines. The TV or fm radio must of course be disconnected from the system when it is used for amateur work! Similarly, masthead amplifiers and splitters must be removed from the line if the system is to be used for amateur operation.

Flagpole Antenna

We can exhibit our patriotism and have an invisible amateur antenna at the same time by disguising our radiator as shown in Fig. 4. The vertical antenna is a wire that has been placed inside a plastic or fiberglass pole.

As shown, the flagpole antenna is structured for a single amateur band, and it is assumed that the height of the pole corresponds to a quarter wavelength for the chosen band. The radials and feed line can be buried in the ground as shown. In a practical installation, the sealed end of the coax cable would protrude slightly into the lower end of the plastic pole.

If a large-diameter, fiberglass pole were available, we might be able to conceal a four-band trap vertical inside it. Alternatively, we might use a metal pole and bury at its base a water-tight box that contained fixed-tuned matching networks for the bands of interest. The networks could be selected remotely by means of a stepping relay inside the box. A 30-ft flagpole would provide good results in this kind of system, provided it was used with a buried radial system. At least one commercial antenna (from Delta Corp.) is used in this manner, but with an elaborate, continuously adjustable matching network (and VSWR indicator) that is operated remotely.

Still another technique uses a wooden flagpole. A small-diameter wire can be stapled to the pole and routed underground to the coax feeder or the matching box. The halyard could by itself constitute the antenna wire if it were made from heavy-duty insulated hookup wire. There are countless variations for this type of antenna, and they are limited only by the imagination of the amateur.

Other Invisible Antennas

Some amateurs have used the metal fence on apartment verandas as antennas, and have had good results on the higher hf bands (20, 15 and 10 meters). We must presume that the fences were not connected to the steel framework of the building, but rather were insulated by the concrete floor to which they were affixed. These veranda fences have also been used effectively as ground systems (counterpoises) for hf-band vertical antennas that were put in place temporarily after darkness fell.

One New York City amateur used the fire escape on his apartment building as a 40-meter antenna, and reported high success in working DX stations with it. Another apartment dweller made use of the aluminum frame on his living-room picture window as an antenna for 10 and 15 meters. He worked it against the metal conductors of the baseboard heater in the same room.

There have been many jokes told over the past decades about "bed-spring antennas." The idea is by no means absurd. Bed springs and metal end boards have been used to advantage by many apartment dwellers as 20-, 15- and 10-meter radiators. A counterpoise ground can be routed along the baseboard of the bedroom and used in combination with the bed spring. It is important to remember that any independent (insulated) metal object of reasonable size can serve as an antenna if the transmitter can be matched to it. An amateur in Detroit, Michigan, once used his Shopsmith craft machine (about 5 feet tall) as a 10-meter antenna. He worked a number of DX stations with it when band conditions were good.

A number of operators have used metal curtain rods and window screens for vhf work, and found them to be quite acceptable for local communications. You'll have best results with any of these makeshift antennas when the "antennas" are kept well away from house wiring and other conductive objects. QST

Strays

N9CGC (left), WB9MVC (center) and KB9MU of the McHenry Co. (Illinois) Wireless Association helped members of the Schaumburg RC monitor a 10-km race sponsored by the Lake Zurich Jaycees. During the race, the hams radioed runners' times and medical information to race officials from checkpoints along the course. *(KA9HAO photo)*

W4KFC TO SPEAK NATIONWIDE ON REPEATER NET

☐ Radio amateurs are invited to listen and speak to ARRL President Vic Clark, W4KFC, when he discusses "The Future of Amateur Radio" on the Teleconference Radio Net, March 3, at 7:15 P.M. CST. Amateurs can participate on the following repeater frequencies:

Phoenix, AZ — 147.36
Avon, CT — 224.78
Roswell, GA — 145.47
Wichita, KS — 146.82
Minneapolis, MN — 146.64
Long Island, NY — 147.375
Beaverton, OR — 147.32
San Antonio, TX — 146.70
Los Angeles, CA — 224.04
Washington, DC — 147.21
Chicago, IL — 147.15
Billerica, MA — 147.12
Cherry Hill, NJ — 147.375
Rochester, NY — 145.11
Dallas, TX — 146.97
Madison, WI — 146.76

If you're not within range of one of these repeaters, it is still possible that a repeater in your area will be tied into the net. Watch for local publicity or check the local repeaters at net time.

Also, mark June 2, 1983, 7:15 P.M., on the calendar. Joe Reisert, W1JR, a nationally recognized expert on EME communications, antennas and TVI, will be the featured speaker on TRN.

The Teleconference Radio Net uses the latest multipoint teleconferencing technology to tie together repeaters across the U.S. The objective is to allow hams to listen and talk to leaders and experts from the Amateur Radio, scientific and electronics communities. The net is organized by the Honeywell ARCs of Minneapolis, Billerica and Phoenix as a service to all amateurs. For further information, contact Rick Whiting, WØTN, national net manager, 4749 Diane Dr., Minnetonka, MN 55343, tel. 612-870-2071.

ANOTHER CLAY HEARD FROM

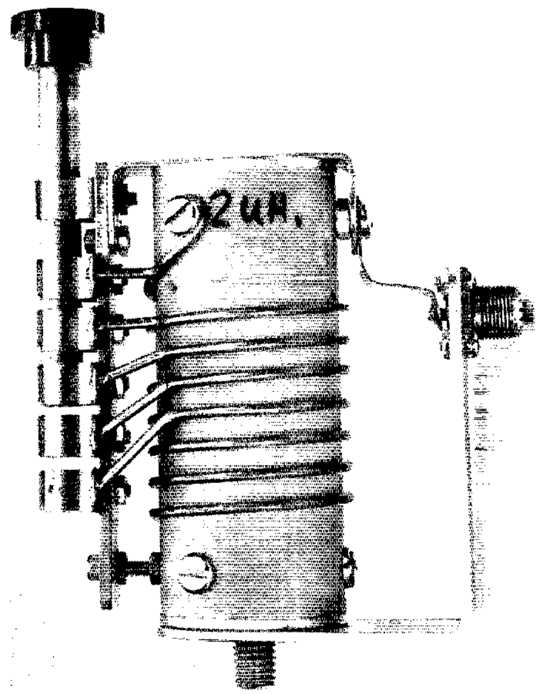
☐ In November 1982 *QST*, there is a Stray about the Clay family of Connecticut with five hams. Well, we are the Clay family of Louisiana with *seven* hams: Roy, Jr., KB5IG; Annie, N5AYU; Sacha, N5DUX; Roy, III, WB5HVS; Eugene, WD5HGD; Timothy, KA5DBK; and Robert, KA5DFJ.

Next Month in QST

Ah, spring, when hams from Las Vegas to Labrador begin to think antennas. March *QST* will bring you details of X Beams and Delta Loops, along with a beginner's look at impedances and an in-depth discussion of baluns. In addition, you'll read about

- an easy-to-build receiving converter for 435, just in time for the Phase IIIB launch, and
- AMSAT-OSCAR 8, perhaps the most successful OSCAR of all, as it celebrates its fifth anniversary.

"Be Switched," Easily



No, not a hickory switch, an rf switch! Don't punish your home-built equipment with a store-bought switch. Next time, build your own!

By Paul Johnson,* W7KBE

Switches for high-powered rf linear amplifiers, tuners and other equipment may be hard to find. The surplus market is drying up, and new switches are expensive. Unless your junk box yields a switch for your next project, the only choice may be to fabricate what you need.

Two requirements must be met. Rf switches have to handle high currents. This dictates large contact surfaces with very low resistance. Second, the use of good insulating materials is essential to prevent arcing and to reduce losses.

Several switches are described here, but they employ two basic designs. These switches can be fabricated with a vise, electric drill, hacksaw and other small hand tools. The materials used are easily obtainable. One type uses fuse clips containing beryllium-copper (silver-plated ones are best). Many sizes can be found in surplus houses. Round brass rod is available at metal supply houses. Glass-epoxy board, 1/8-inch thick, is excellent insulation and is easy to work with.[1] The second design uses hardware from a dpdt knife switch. One with a 2- × 4-inch porcelain base is a good choice. Electrical supply houses carry them as a stock item, but electronic flea markets are another source.

Slide Switches

The title-page photo shows a five-position, single-pole, shorting-type switch used to tap a matching coil at the base of a mobile whip. This switch has a push-pull action, shorting out coil turns to obtain the correct inductance to match 50-ohm coaxial cable on different bands. Large clips are used to carry the heavy current present at the base of a mobile antenna. The clips are mounted on a 1-inch-wide strip of glass-epoxy board. A 1/2-inch-diameter brass rod acts as a shorting bar and it slides up and down in the clips to change the inductance. Solder lugs, in the form of thin copper strips, are placed under the clips. The top clip acts as a guide when the sliding rod is in the upper position.

A single-pole, five-position shorting type of switch is shown in Fig. 1. Fuse clips and a piece of brass rod are used to short sections of a tapped coil. Small clips and a 3/8-inch rod are used. The fuse clips are mounted on a 1-inch-wide glass-epoxy strip. Careful alignment of the clips is necessary so the contacting surface areas are maximized and the sliding action is easy. Thin copper strips under the fuse clips serve as solder lugs.

I built an antenna-matching network that uses a two-pole, three-position, shorting type of switch.[2] Fig. 2 shows the construction details. Here again, fuse clips and brass rod for shorting the tapped coil are used. To obtain two poles on one sliding rod, two pieces of brass and two pieces of Bakelite® are used. This assembly is made by drilling and tapping the ends of the rods for no. 8-32 screws, and the rod is then fastened with studs. The fuse clips are mounted on a 1-inch-wide glass-epoxy strip, and are aligned carefully.

Another switch, using fuse clips, was designed for a linear amplifier. In this case the output circuit was a Pi-L network. Therefore, a five-position, two-pole, shorting type of switch was necessary. The amplifier operates on five bands. Fig. 3 is a picture of the switch. Two sliding brass rods are used in parallel to make two poles. A third pole, in the center, places another capacitor across the main tuning capacitor when the switch is all of the way out. This pads the main tuning capacitor for 80-meter operation. The fuse clips are mounted on a 2-1/4-inch-wide glass-epoxy strip. Again, careful alignment is important to obtain easy push-pull action.

A piece of phenolic rod protrudes through the front panel. Grooves are cut in this rod, and different colors of paint can be applied in each one. The grooves line up with the front panel to indicate the position of the switch. With the knob against the front panel, all sections are shorted, putting the amplifier on 10 meters. A shaft coupler on the inside prevents the rod from being pulled out of the guide clips. When the rod is at the stop, all the way out, the amplifier is on 80 meters.

Rotary Switch

A rotary type of switch is shown in Fig. 4. Here the knife-switch parts are used. The copper parts are removed from the

[1]Notes appear on page 19.

*10817 Brookside Drive, Sun City, AZ 85351

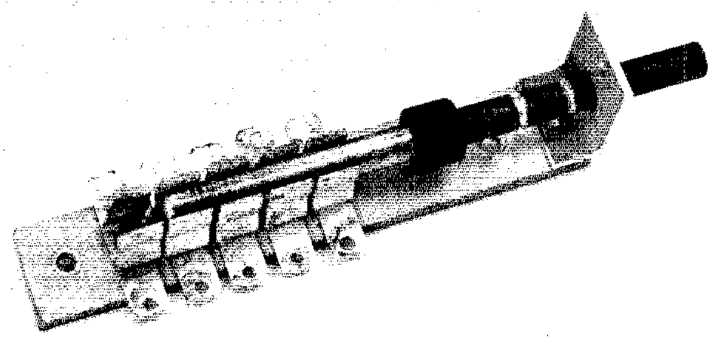


Fig. 1 — Construction details of a single-pole, five-position shorting type of switch are shown. Note the grooves in the Bakelite® rod. They are painted various colors and used to indicate the switch position on the front panel.

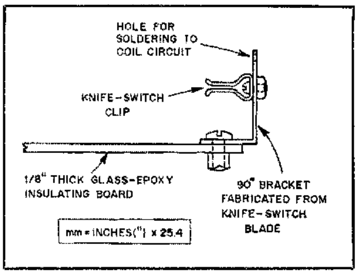


Fig. 5 — Construction details of the rotary-switch contacts.

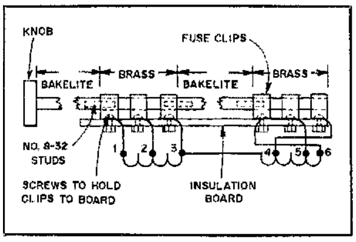


Fig. 2 — Construction and wiring of a two-pole, three-position switch for use in an antenna-matching network.

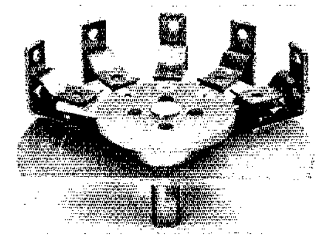


Fig. 4 — Photo of a rotary type of shorting switch. Knife-switch parts are used to make the contacts.

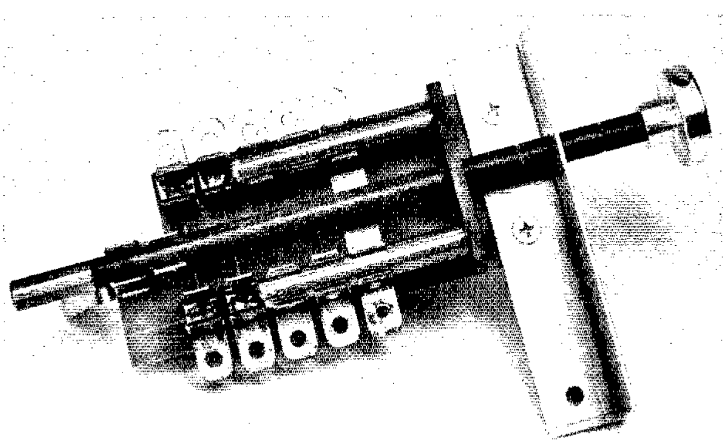


Fig. 3 — Photo of a switch built for use in the Pi-L network of a linear amplifier. Note the third set of contacts in the center. They are used to add an extra capacitor in the 80-meter position.

porcelain base. Bend the knife blades to form 90° brackets (Fig. 5). These angle brackets support the knife clips and also act as soldering lugs. This switch is a single-pole, five-position shorting type, useful in a linear amplifier Pi network. The 90° brackets, when bolted to the knife clips, are mounted on a circular piece of glass-epoxy board about 3-1/2 inches in diameter. The clips are mounted in a semicircle, 30° apart. A portion of the circular insulating board is cut off to save space.

A rotary segment, which engages the clips to short sections of the tapped coil, is fabricated from 0.062-inch thick brass sheet. It was first cut into a circle, the holes laid out, then recut to make the proper size segment. To insulate the rotor from the rest of the switch, it was mounted on a circular piece of glass-epoxy board with three screws. This assembly has a flanged 1/4-inch brass shaft fastened to the circular board by means of four screws. The 1/4-inch shaft and rotor assembly run in a 1/4-inch panel bearing. It is possible to gang two (or more) such switches. A large indicator knob is used to turn this switch, and the switch positions are marked on the front panel for reference.

Conclusion

You should consider silver plating all switch parts before assembling, if it can be done reasonably. The plating should be fairly heavy, but does not require polishing. Plating will help keep the contact resistance low, and soldering will be easy. The general appearance of a plated switch adds a professional touch to a home-built project.

Dimensions are given only as guides to switch construction. The switch sizes depend on available materials and the intended application. Alignment of the clips is important. This is accomplished by making the mounting holes a little oversized. The rod or swinger is engaged into the clips and the screws tightened, giving correct alignment and providing maximum surface contact.

The number of physical variations for switches using these designs is limited only by your imagination. Many new applications will be found. Will you be one of the innovators?

Paul A. Johnson, 67 years of age, retired in 1974 from the Carnegie Institution of Washington, DC (a nonprofit research organization) as the model-shop manager. He has done considerable field work in radio astronomy and seismology throughout the U.S. and South America. In addition to his present call sign, W7KBE, Paul has held W8EET, W8MPF, W3KTK and W3EET.

Notes

[1]mm = inches × 25.4.

[2]L. Aurick, "The 'Lowbander's' One-Antenna Farm," *QST*, Feb. 1982, pp. 23-24.

Efficient Ground Systems for Vertical Antennas

Elevated ground systems for vertical antennas have been a bit of a mystery for the past 60 years. This report of an extensive study reveals some startling results.

By Archibald C. Doty, Jr.,* K8CFU, John A. Frey,** W3ESU, and Harry J. Mills,*** K4HU

On the evening of November 27, 1923, the first two-way transatlantic radio communications took place between two Amateur Radio stations in the United States and a fellow amateur in France. One of the American stations was operated by John L. Reinartz, "the father of short-wave radio."

The antenna installation of that station was novel. The success of the Reinartz station was attributed to a new principle of antenna construction: A counterpoise was used as the second part of the balanced antenna system, rather than the capacitive ground used previously. After that night, the counterpoise (an elevated array of one or more insulated wires placed under the antenna) was used commonly in amateur and commerical antenna systems for more than 10 years. As the counterpoise was developed, its usual configuration became a symmetrical pattern of radial wires used as the "artificial ground" under electrically short (one quarter wavelength or less) vertical antennas, as shown in Fig. 1.

Development of the counterpoise was, by necessity, empirical in nature, as sophisticated instrumentation and standardized antenna testing procedures were not available in those early days of radio. But after 1937 the counterpoise was almost forgotten because of a landmark research program on another type of ground system — the buried radial.[1] This research, conducted by the legendary George H. Brown, was so detailed and authoritative that its results were used to develop the Federal Communications Commission's requirements for a-m broadcast stations. These requirements, which still apply today, specify that each station shall be equipped with a ground system consisting of at least 120 radials, each being at least 1/4 wavelength long.[2]

As is often the result of rigid regulation, interest and development of possibly superior ground systems, including the counterpoise, has languished for the past 40 years. The same situation holds true for the "first cousin" of the counterpoise — the elevated ground screen. This is an elevated ground system that may be physically identical to the counterpoise, except that the radial wires are connected to the earth. The ground screen is little used these days, and the literature provides a very limited amount of information on its operating characteristics.

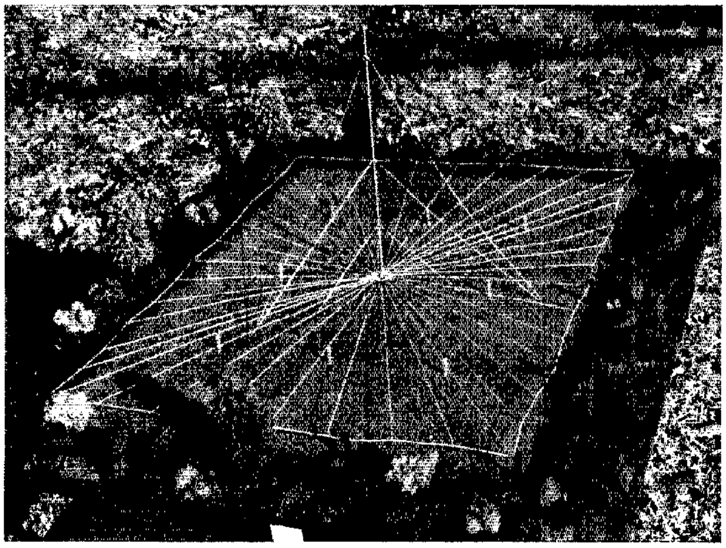


Fig. 2 — This scale model of the test antenna system was built before construction of the full-size version was undertaken. The model measures 3 by 5 feet, while the actual antenna system is 200 by 300 feet.

*Jackson Rd., Rte. 3, Fletcher, NC 28732
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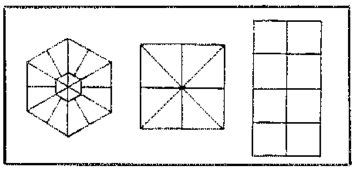

Fig. 1 — Typical arrangements of wires for counterpoise or ground-screen systems.

[1]Notes appear on page 25.

The Test Program

In 1979 a rather elaborate program was begun in Fletcher, North Carolina, to determine the effectiveness of the counterpoise and elevated ground systems. These tests were conducted at 1.8 MHz. The vertical antenna used for the portion of the test program reported here was a 45-degree (about 65-foot) top-loaded section of tower.[3] The base of this antenna was 4 feet above the ground and was insulated from the earth. The antenna was deliberately offset slightly from the center of the radial system so that no two radials would be of exactly the same length. The radial wire system used in the tests covers an area of 200 × 300 feet (slightly less than 1.4 acres). This system was used first as a counterpoise and then, after it had been grounded at several locations, as a ground screen. The longer dimension was selected on the assumption that the optimum radial lengths determined by Brown for buried radials would also be correct for a counterpoise and an elevated ground screen (a correct assumption, as it turned out!).[4] The smaller dimension was selected to be less than the optimum length found by Brown.

Tests were conducted with 6, 12, 18, 24, 36, 40 and 48 radials in place. Total lengths of wire in the various systems tested were from 1821 feet to 7386 feet. The radial wires are from 6 to 8 feet above ground. Before construction of the antenna system was undertaken, the 3- × 5-foot scale model shown in Fig. 2 was made and evaluated for mechanical considerations.

Ground Conductivity

Antenna literature is full of references to the importance of ground conductivity to the efficient operation of vertical antennas. Our first chore, therefore, was to determine the ground conductivity of the test site. There have been few mentions in the literature of actual measurements of conductivity, and no convenient system for such measurements was offered until recently. In the Fletcher test program, ground conductivity was measured directly by the relatively simple method described by Sevick.[5] At first we experienced trouble from ground currents. Thus, we modified Sevick's circuit by adding an isolation transformer, and that cleared up the problem. Fig. 3 shows the results of four months of daily tests of the ground conductivity at a "standard" fixed location adjacent to the base of the antenna. As is shown by this data, the conductivity varied by a factor of approximately 2:1 over the period covered.

Data has now been accumulated for a total of 16 months, and conductivity has remained within the limits shown. The long-term variations in conductivity correlated, as expected, with the variations in temperature and rainfall. In addition, however, severe variations in conductivity were found at different locations in the test site. As illustrated in Fig. 4, ground conductivity varied by a factor of 7.5:1 within less than 10 feet.

In theoretical analyses of vertical antenna radiation patterns, previous literature indicates that the ground conductivity was assumed to be homogeneous throughout the area covered by the antenna ground system (usually buried radial wires). Our tests indicate that it should *not* be assumed that the ground conductivity will be homogeneous. Past usage of an area (such as agricultural, where fertilizers or chemicals have been applied), past conditions (as on cleared land previously covered with brush or trees, which could accumulate salts or other chemicals in their root systems), and microtopography (smaller ruts or lower areas that may contain electrolytes washed from higher areas) can be expected to produce considerable local variations in ground conductivity. If you have built a vertical antenna and have found that its radiation pattern is unusual in any respect, perhaps the ground conductivity under it was the reason!

RF Current Measurements

The tests reported here (and those conducted subsequently) required the taking of literally thousands of rf current measurements. Currents were repeatedly measured in each radial wire at various distances from the antenna, for the various test conditions. As soon as the earliest measurements were made in the

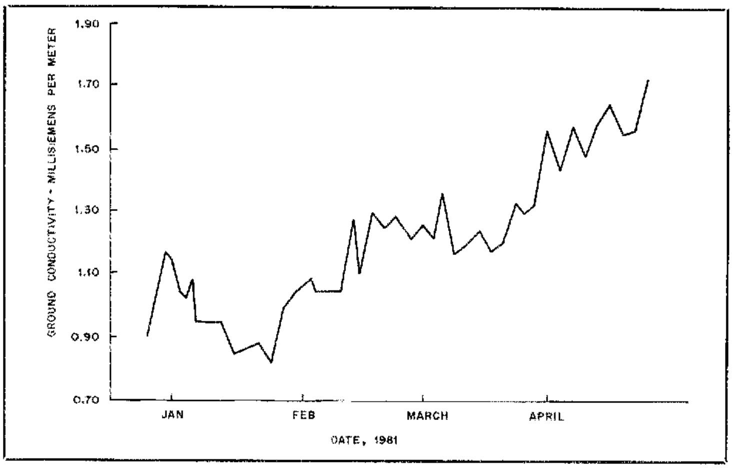

Fig. 3 — Ground conductivity at one fixed location in the test site at Fletcher, North Carolina. Measurements, taken daily, are shown here for the first four months of 1981.

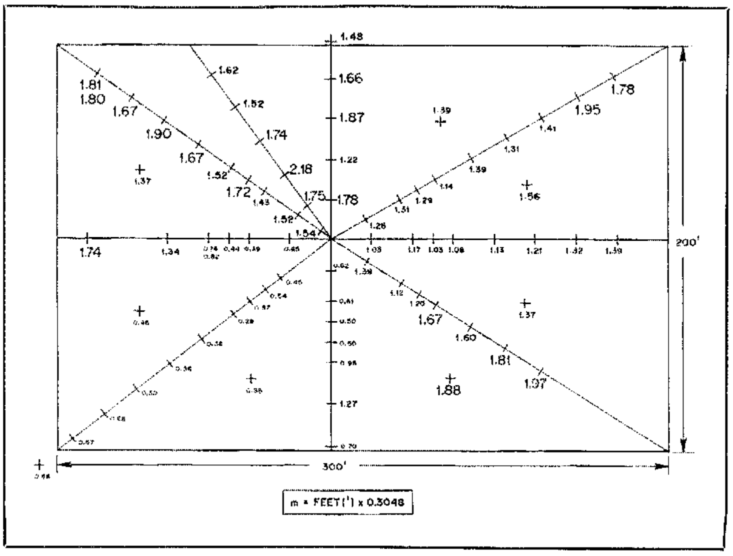

Fig. 4 — Ground conductivity along different radials at the test site. Note that there is a difference ratio of as much as 7.5:1 within a distance of 10 feet.

test program, it was evident that the most sensitive rf ammeter commonly available (0-100 mA) would not be satisfactory. Readings of less than 1 mA were commonplace at the power level that we wished to feed to the antenna. Accordingly, a magnetometer of the design published by Carr was constructed and used throughout the test program, Fig. 5.[6]

Although it is not evident in Fig. 5, two pieces of sheet plastic, spaced 1/8 in. apart, are mounted on the top of the magnetometer. The radial to be measured was placed in the slot between the pieces of plastic, thus locating the wire precisely. It was found that the meter was capable of reproducible measurements of rf currents of less than 10 microamperes, and of indicating the presence of levels in the area of 1 to 3 microamperes. The magnetometer was calibrated by averaging readings from a number of sensitive thermocouple-type instruments. The extreme sensitivity of the meter allowed it to indicate clearly the presence of a small aluminum ladder 5 feet from a radial!

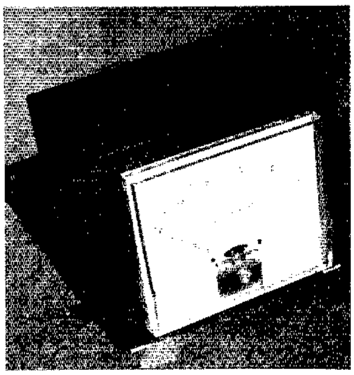

Fig. 5 — The magnetometer used for rf current measurements, after Carr.[6] The basic meter movement is a 5-inch, 50-μA instrument, selected for its sensitivity and readability.

In use, the instrument was placed against the radial wire and positioned so the wire would "snap" into place in the slot between the plastic sheets. This enabled repeatable measurements to be taken rapidly. In other tests, rf currents were measured in buried radials by cutting small rectangular holes in the earth beyond the depth of the radials. This permitted inserting the magnetometer into the holes with the buried wires positioned in the slot.

Current Distribution in Counterpoise Radials

No record has been found in past literature of specific data on the the distribution of current in the wires comprising any type of artificial ground system — buried radials, counterpoise or ground screen. What data is available indicates that there is uniform distribution of current in the radial wires of such systems. The thousands of measurements made in the Fletcher tests clearly indicate lack of uniformity of current distribution in the radial wires of a counterpoise. Fig. 6, showing the result of 800 data points, illustrates the wide variation of current flow found in the radial wires of a 48-radial counterpoise.

The distribution shown was a considerable surprise. Every indication found in the literature of the past 60 years[7-11], and in modern commentary by experts in the field, indicates that the counterpoise was thought to act as one plate of a large capacitor. The earth was thought to act as the second plate. It was believed that the return currents flowing back to the base of the antenna through the radials should be better distributed than with buried radials. It was also believed that the currents should show some concentration (as the result of the "edge effect" of a capacitor) along the periphery of the counterpoise system, rather than being concentrated near the base of the antenna, as is found with buried radial systems. The Fletcher tests showed an entirely different distribution of current flow in the radials of a counterpoise from that predicted in the literature.

A number of tests at the inception of the program indicated that there was a direct relationship between the level of current in the counterpoise radials and the conductivity of the ground under them. The most striking illustration of this correlation comes from comparing Fig. 4, which shows the conductivity of the ground at the test site, and Fig. 6, showing the currents in the radials. The comparison of these two illustrations offers an intriguing possibility. Shouldn't it be possible to modify the current pattern in the radials, and thus the radiation pattern of the antenna, by artificially modifying the conductivity of the ground under the radials? Preliminary tests by the authors indicates that this is definitely possible!

As radials were added to the counterpoise, a record was kept of the current supplied to the antenna, and the return current collected by the radials. The results are given in Table 1. This data indicates that an elevated counterpoise of less than 50 radial wires provides a very efficient ground system. Previous tests have indicated that more than 100 radials are required to provide the same level of performance if buried radials are used.

As previously mentioned, it has been thought that an elevated counterpoise acts in conjunction with the earth as a giant capacitor, with the counterpoise radials being one "plate" and the earth the other. The Fletcher tests clearly indicate that a counterpoise operates in a more complex

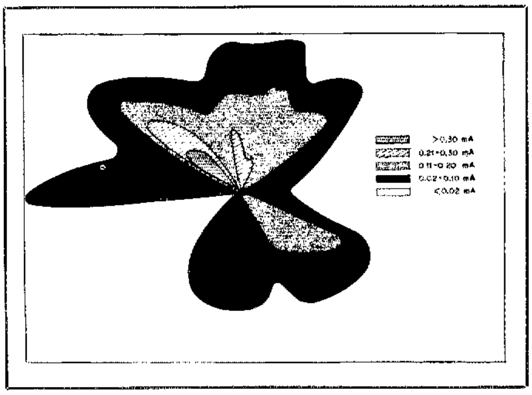

Fig. 6 — The range of current distribution in the wires of a 48-radial counterpoise is shown over the area of the test site. This drawing is based on approximately 800 data points. As one would assume, the current tends to be greatest at the base of the vertical radiator. But the surprising fact emerged that the current was nowhere near being equal from one radial wire to another.

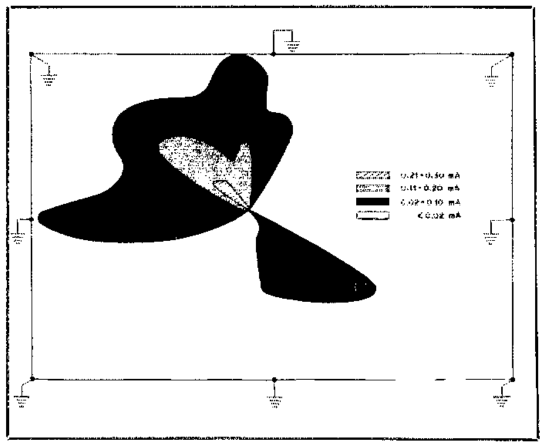

Fig. 7 — The range of current distribution in the wires of a 48-radial elevated ground screen. The wires were grounded at the center, and at eight points on the periphery of the system.

Table 1

Current Flowing in the Antenna and in the Radials of a Counterpoise

No. of Radials	*Antenna Current, mA*	*Return Current, mA*	*Return as % of Ant. Curr.*
6	350	220	63
12	390	260	66
20	495	445	90
28	495	450	91
36	495	455	92
40	495	490	99
48	495	495	100

manner than has been thought for the past 60 or so years, and that the wires in a counterpoise system definitely do not operate in concert as the plate of a large capacitor of the type usually considered. Rather, the counterpoise appears to operate efficiently as the result of a number of factors, including those itemized below.

1) Each radial of a counterpoise is individually capacitively coupled to the earth under that radial.

2) The magnitude of the currents carried by the radials of a counterpoise vary as the conductivity of the ground under the radials — the higher the conductivity, the higher the current. Current distribution in counterpoise radials is *not* similar to that found in buried radial systems.[12]

3) The counterpoise provides an efficient, low-loss path for antenna return currents, as do buried radial systems. The small amount of current in the counterpoise radials which is not directly induced from the antenna is in the form of displacement currents flowing from the earth as "charge-discharge" currents. However, the return currents in a buried radial must flow through high-resistance, high-loss earth to get to those radials. What this means in practice is that a counterpoise operates more efficiently — with less loss — in collecting return currents and guiding them to the base of the antenna than does a buried radial ground system.

The Elevated Ground Screen

The elevated ground screen used in the tests was the same physical array of radial wires as that for the counterpoise tests. For the ground screen tests, however, the radials were grounded at the center of the "spiderweb," and at eight locations around the peripheral wire that connected the far ends of the radials. Previous literature provides no specific data as to what distribution of current might be expected in the radial wires of an elevated ground screen. The only information we could uncover indicates the belief that the currents in the radials of a buried ground system should be equal from radial to radial. The measurements made at Fletcher leave no doubt that there was a pronounced lack of uniformity of current distribution in the radials of the ground screen tested. Fig. 7, which portrays several hundred data points, illustrates the wide variation of current flow found.

The radial wires in the ground screen vary in length from 96 feet to 173 feet. Each radial is grounded at its end near the antenna, and through the grounded peripheral wire at its far end. The various current levels in the radials are not related to the radial lengths. No rational explanation could be found for the unexpected, uneven, current pattern until the ground conductivity factor, discussed above in relation to the counterpoise, was discovered.

From the test data, a direct correlation was found between the level of current flow in the radials and the conductivity of the ground under those radials. Fig. 4 shows the ground conductivity at numerous locations of the test site, and Fig. 7 shows the level of the return currents in the radials of a 48-wire elevated ground screen over that ground. A comparison illustrates this correlation.

Additional testing showed that the correlation holds true even if a radial is placed on the surface of the ground, and connected to ground stakes every 20 feet. This was verified by testing a radial laid on the ground at two different locations, one having higher ground conductivity, and the other lower conductivity. Fig. 8 shows the considerable difference in radial current found. As shown, the current was substantially higher when the radial was on the ground with higher conductivity.

As each set of radials was added to the ground screen, a record was made of the current being supplied to the antenna, and the return current collected by the radials. The results are given in Table 2. As was also found in the case of the counterpoise, it appears that a ground system of fewer than 50 elevated radials may be as efficient as more than 100 buried radials.

Table 2

Current Flowing in the Antenna and Radials of Elevated Ground Screen

No. of Radials	*Antenna Current, mA*	*Return Current, mA*	*Return as % of Ant. Curr.*
0*	440	380	86
6	415	370	89
12	380	330	87
20	410	370	90
28	490	450	92
36	495	470	95
40	495	490	98
48	495	493	99

*Two 8-foot ground rods at base of antenna used to collect this data.

The lack of research data in the past literature on the operational characteristics of the counterpoise extends to the ground screen. The ground screen is described in some detail by Laport[13] and others, and its use is recommended in conjunction with buried-radial-wire ground systems to reduce ohmic losses near the base of a vertical antenna — losses caused by the concentration of return currents in that area (i.e., I^2R losses). There is, however, no specific data on elevated ground screens operating alone, without additional buried radials. It can only be deduced that the ground screen has been considered to be identical in its operating characteristics, regardless of whether the radial wires were buried or elevated.

The Fletcher tests indicate that an elevated ground screen operates with excellent efficiency as the result of a number of factors.

1) The ground screen operates as a low-resistance, low-loss path for return currents.

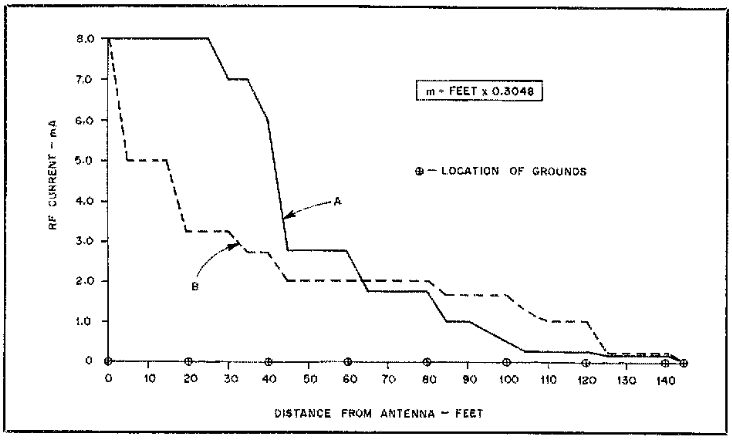

Fig. 8 — This drawing shows how ground conductivity affects the current flowing in a radial wire placed on the surface of the earth. Both wires were grounded every 20 feet. The wire of curve A was placed over soil with conductivity averaging 1.66 mS/m, while that under curve B averaged 1.17 mS/m.

2) Even though it is grounded at each end, each radial of the ground screen tested was capacitively coupled to the ground beneath it.

3) Return currents carried by the radials of an elevated ground screen vary with the conductivity of the ground under the individual radials — the higher the conductivity, the higher the current.

Summary of Findings

Our findings may be summarized briefly as follows: (a) There can be large variations in surface ground conductivity within the area covered by the ground system of a vertical antenna. (b) If there are variations in ground conductivity in the area occupied by a radial system, there will be corresponding variations in the magnitude of the currents in the various radials. These variations will persist regardless of whether the radials are elevated and insulated, as in a counterpoise, or grounded, as in a ground screen. (c) Elevated ground systems — counterpoise or ground screen — will collect return currents primarily as displacement currents induced directly by the field of the antenna. Currents will also be collected as charge-discharge currents induced from the ground below.

The majority of the return currents carried by elevated ground-system radials are collected directly from the antenna, or from the ground through an air dielectric. There is no necessity for these currents to pass for any considerable distance through high-resistance earth (as is the case when buried radials are used). As a result, these systems are highly efficient. Initial indications (i.e., a comparison of the data presented here with Brown's very thorough research on buried radials[14]) are that elevated "ground" systems used with electrically short vertical antennas may require substantially fewer radials than needed by a buried wire system of equal efficiency. (These indications have not yet been confirmed by direct comparison.)

Future Research

The research project described in this article was designed to help fill the gap of technical knowledge about the counterpoise and the elevated ground screen. Our findings, to a considerable degree, do not agree with the technical writings of the past 60 years, and they may therefore be controversial. In this regard we would like to point out that in the 45 years since the last comprehensive work was done on radial-current measurements, there have been great advances in instrumentation and techniques for measuring both ground conductivity and rf currents in wires. Thus, today's researcher, using relatively simple equipment, can make infinitely more accurate measurements than could be done in the past with a whole laboratory full of instrumentation.

We hope this research will encourage others to continue the investigations we have started, and that Amateur Radio operators, who have always led the way in the area of practical experimentation, will obtain operational evaluation of counterpoises and ground screens from on-the-air tests. To assist efforts in this direction, we have examined the results of our research and have designed a "Minipoise" ground system that can be constructed in almost any backyard. As shown in Figs. 9 and 10, this antenna system comprises a 30-foot-high vertical antenna and a 100- × 100-foot counterpoise.

You might think the 30-foot vertical is short for use on 160 meters. However, a check of the literature shows us that ". . . an antenna of infinitesimal length, subject to no losses, yields a field strength which is only 4.25 percent less than the field from a quarter-wave antenna."[15] This makes a 30-foot-high antenna, if used with a really efficient ground system, begin to look more attractive!

The 100- × 100-foot radial system, again based on past practice with buried radials, might seem too small to fill the requirement of a "really efficient ground system" for a 1.8-MHz antenna. But a look at Fig. 6 shows that by far the majority of the return currents in the radial wires of a counterpoise are collected within 100 feet of the base of the antenna. And if this is so, why should the radials be longer?

At present, we are testing an extensively instrumented version of the system shown in Fig. 9, and have designed a rigorous test program to check its characteristics. We are making our design widely available, in the hope that others will join us in operational testing of this antenna. If impetus is required to get others interested, we might mention that our original antenna has been providing some rather remarkable results. From our location in the mountains of North Carolina — 400 miles inland — we have, with no difficulty at all, worked all continents on 160 meters. Our contacts included ZL, 4X4, OA, LZ, ZS, many European countries and all 50 states. And 99% of these contacts have been on ssb! We hope that many amateurs will join us in further testing of the counterpoise or ground screen systems.

Acknowledgments

From the inception of our test program

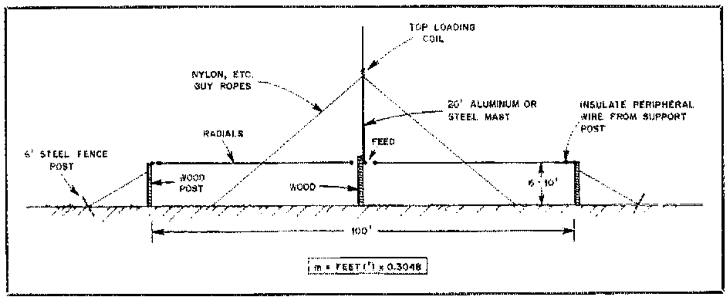

Fig. 9 — The Minipoise antenna system with a short, top-loaded vertical radiator, for 160-meter operation. See text.

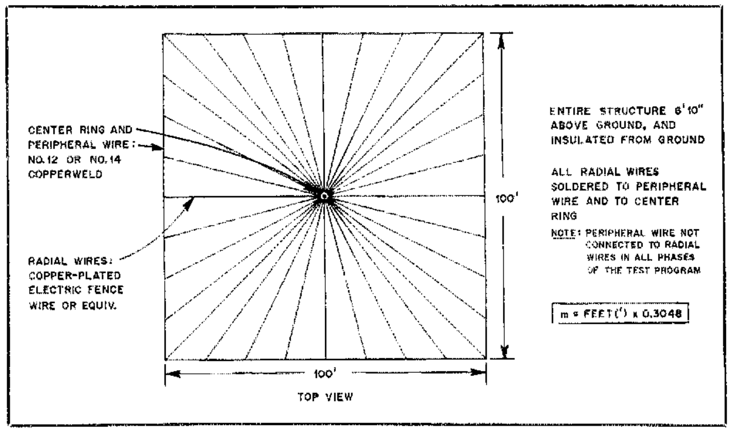

Fig. 10 — The Minipoise (miniature counterpoise) radial system is 100 feet square. This antenna was developed to promote interest in furthering the tests begun by the authors.

we have had a great amount of help, suggestions and encouragement from a number of our friends. Thus, we would like to express our very sincere thanks to those who gave their personal time to help this project: to Edmund A. Laport, without whose encouragement this project would never have been finished; to Barry A. Boothe, W9UCW, Earl W. Cunningham, K6SE, Paul R. Engle, K9QLL, and Richard B. Frey, K4XU, for their incisive and expert editing of the full technical report, from which this article was excerpted.

The complete (75-page) report of the full (15,000-measurement) test program includes the results of extensive additional tests not reported here. These included investigation of (a) capacitance of elevated radial wires to ground, (b) dielectric anomolies in the test area, (c) radial current vs. number of radials, and (d) effect of artificial modification of ground conductivity on radial current. Also, 27 references, in addition to those appended, were used in the design of the test program. The authors will be glad to share this material with anyone planning to undertake a definitive test program on the counterpoise or the ground screen. QST

Notes

[1] G. H. Brown, R. F. Lewis, and J. Epstein, "Ground Systems As A Factor in Antenna Efficiency," *Proceedings of the IRE,* June 1937, pp. 753-787.

[2] FCC Rules and Regulations, Part 73, Subpart A, 1972.

[3] Meters = ft × 0.3048.

[4] See note 1.

[5] J. Sevick, "Measuring Soil Conductivity," *QST,* March 1981, pp. 38-39.

[6] K. F. Carr, "Ground Currents Measuring," *Ham Radio,* June 1979, pp. 46-47.

[7] K. Henny, *Principles of Radio,* 3rd ed. (New York: John Wiley and Son, 1938), p. 461.

[8] S. A. Schelkunoff and H. T. Friis, *Antennas, Theory and Practice* (New York: John Wiley and Son, 1952).

[9] A. R. Nilson and J. L. Hornung, *Practical Radio Communication* (New York: McGraw-Hill, 1935).

[10] *Antennas and Radio Propagation,* Dept. of the Army, 1935.

[11] Schure, *Antennas* (New York: John R. Rider, 1957), pp. 23, 73.

[12] E. A. Laport, *Radio Antenna Engineering* (New York: McGraw Hill, 1952), pp. 48-54, 115-117, 187.

[13] See note 12.

[14] See note 1.

[15] See note 1.

References

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Electrical Communications System Engineering Radio, Dept. of the Army, TM 11-486-6, 1956.

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"Some Thoughts on Spark Operation," January 1921 *QST.* See the secton on "Grounds," pp. 6, 7.

Spencer, L., "John L. Reinartz," *Ham Radio,* Aug. 1981, pp. 10-18.

The ARRL Antenna Book, ARRL, 1968, pp. 200, 201.

New Books

☐ *Amateur Radio Operating Manual,* R. J. Eckersley, G4FTJ, editor, published by the Radio Society of Great Britain, London; second edition, 1982. 204 pages, soft-cover, 7-1/4 × 9-3/4 in., $10. Available from ARRL Hq. or your radio book counter.

There are lots of new books on Amateur Radio subjects floating around these days, many of them with inflated price tags and few of them worth the asking price. Shelling out upwards of $10 for a poorly researched and sloppily edited publication appears to be what some publishers expect radio amateurs to do these days.

One book that definitely *is* worth the money, if you're at all interested in viewing on-the-air operating from a fresh perspective, is the *Amateur Radio Operating Manual,* published by the Radio Society of Great Britain. In the brand-new second edition of this popular publication, editor R. J. Eckersley, G4FTJ, combines the efforts of some 50 contributors into a highly informative volume of more than 200 pages. Much of this material is original and unique to this publication; you probably won't find it anywhere else on this side of the Atlantic.

While written primarily for the British amateur and shortwave listener, most of its contents are of universal interest. Just about all aspects of Amateur Radio operating are covered in sufficient depth to be instructive to the old-timer but the text is written at a level that should appeal to the newcomer. In fact, a non-radio enthusiast could pick it up, read the opening chapter defining the Amateur Radio Service, and continue from there, developing a pretty good (although vicarious) working knowledge of what Amateur Radio is all about. This is in contrast to most other Amateur Radio literature, which seems to presuppose a large amount of "inbreeding."

The 10 chapters and 7 appendices include 11 pages on how to set up a station, 17 pages on operating practices, 52 pages of information on DXing (at hf *and* vhf/uhf), 10 pages on amateur satellites and 19 pages on contesting (did you know that most contesters drink Coke, Pepsi, black coffee or tea?), and so forth. The book is lavishly illustrated throughout, with high-quality charts and graphs, diagrams and very detailed maps. Of particular interest, for both the armchair traveler and the DXpeditioner, are the individual call area maps for each country that has been assigned an ITU call sign block, along with the addresses of the licensing administration and the national Amateur Radio society in each country.

I have the utmost admiration for Editor Eckersley and his associates for the painstaking work involved in compiling all the information that appears in the RSGB *Amateur Radio Operating Manual.* However, I would comment that the *ARRL Operating Manual,* although not nearly as comprehensive, may be more readable to amateurs on this side of the pond, owing to its somewhat more informal prose style. Between the two of them, no questions about Amateur Radio operating — one of life's spirited adventures — will be left unanswered. — *Robert Halprin, K1XA*

☐ *EMP Engineering And Design Principles,* by Bell Laboratories, copyright 1975, second printing. Soft-bound, 8-1/2 × 11 in., 151 pp., $14.95 plus $2 shipping. Available from Clayton Survival Services, P.O. Box 1411 Mariposa, CA 95338.

In the August 1981 issue of *QST,* a major article featured the subject of EMP — electromagnetic pulses generated by a high-altitude detonation of a nuclear device. A 250-mile-high nuclear detonation over the central USA would create an intense pulse of tens of kilovolts within nanoseconds over a frequency range from a few cycles into vhf. The radius of this pulse would cover the entire continental USA. Consequently, EMP represents a potential threat to the integrity of all solid-state devices in general, and to Amateur Radio equipment in particular.

This book, although written at the engineering level, is an excellent source that enables amateurs to better understand EMP and take precautionary measures against it. The book is directed toward telephone company buildings, installations and equipment. However, most of the information is applicable to radio communications.

An opening chapter covers EMP generation and characteristics in detail, and is excellent in providing for an understanding of the EMP phenomenon.

A chapter on coupling to exterior structures deals with EMP conduction by power lines, cables, towers, waveguides and other metallic objects into the interior of a building.

The chapter on shielding is of limited interest to amateurs, since it mostly involves shielding an entire building from EMP by means of reinforcing steel rod loops installed during the construction of concrete buildings.

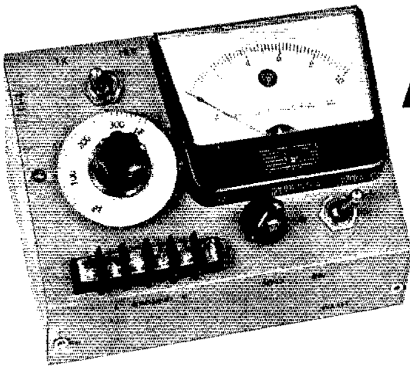
Coupling inside the building is the subject of a chapter that deals with methods of preventing EMP from reaching sensitive devices, principally by multiple bonding and grounding outer shields of conductors, but also through the use of EMP-effective lightning arrestors.

Component susceptibility and electrical devices for EMP mitigation are two chapters of real interest and practical value to radio amateurs — particularly to designers and builders of solid-state equipment. Detailed factual data is given on various types of component vulnerability to EMP, and various methods are outlined on how to protect these components.

The final two chapters concern EMP testing and personal safety. This information is interesting, but few (if any) hams have this kind of high-voltage testing capability.

EMP Engineering And Design Principles is one of the few available sources of detailed information on EMP. I highly recommend it for reading and application by all radio amateurs, and designers of Amateur Radio equipment.

With new Amateur Radio equipment now virtually 100% solid-state, our hobby, which serves in the public interest, can ill afford to fall behind the state of the art with regard to EMP. We must be able to provide radio communications in the event of a nuclear emergency situation. — *Fred Huntley, W6RNC* QST



A Simple L-C Meter

Build this circuit and unlock the secrets of unmarked components

By Frank Noble,* W3MT

The instrument described in this article minimizes pitfalls that tend to discourage the prospective builder: It is relatively cheap to build, the set-up is fast, the L/C calculations are easy and the accuracy is good. A miniature crystal-controlled oscillator, in conjunction with the unknown coil, forms the "heart" of the device. Voltage across the coil is indicated on a meter; it will be maximum when the coil is resonant. The coil is resonated with a calibrated variable capacitor. Operating frequencies for the crystal oscillator are chosen so that the unknown inductance is an integer power of 10 times the reciprocal of the resonant-tank capacitance. This simplifies the math involved. It is easily shown that the frequencies required will have the number sequence either $1/2\pi$ or $\sqrt{10}/2\pi$.[1] I used 5033- and 15,915-kHz crystals with a 365-pF broadcast variable capacitor in the tank circuit. This results in an inductance range of 0.286 to 28.6 μH and a capacitance range of 10 to 330 pF — values commonly used in the range of 2 to 30 MHz.

The Circuit

Fig. 1 is the schematic diagram of the meter. The positive battery terminal and one end of the coil are directly grounded, which simplifies the metering circuit. A Pierce crystal oscillator is used because it does not require a tuned circuit. Other oscillators require tuned tank circuits that must be switched, increasing complexity and cost. The oscillator FET source resistor value was found experimentally. This value produces the cleanest waveform; it is free from harmonics that can produce false meter indications. A capacitive voltage divider is used as an interstage coupling network, employing a small input capacitor to minimize oscillator loading and a large output capacitor to reduce amplifier drive. With the amplifier operating in the linear mode, the chance for spurious meter indications is further reduced. An unbypassed source resistor in the amplifier stage reduces distortion and increases the effective drain resistance, reducing the load on the output circuit. This increases the tuning sharpness. A 1N34A germanium diode rectifies the rf energy, which is then displayed on a sensitive μA meter.

Measuring Inductance

To measure inductance, the unknown coil is connected, the instrument is energized, and the variable capacitor is tuned until the coil resonates; this is indicated by a peak in the meter reading. If it does not peak, the crystal is switched and the unit is tuned again. At resonance, the inductance in microhenrys is $1000/C_{pF}$ when using the 5033-kHz crystal; and $100/C_{pF}$ for the 15,915-kHz crystal.

Instrument accuracy is limited by crystal tolerance and the variable capacitor calibration. Crystal frequency accuracy far exceeds the dial calibration accuracy, so low-tolerance crystals make economic sense.[2] The specified capacitor has semi-circular plates and a capacitance range from 10- to 365-pF, which is close to 2 pF-per-degree of shaft rotation. Accuracy adequate for amateur purposes may be obtained by dividing the dial in increments of 10 pF, or 5 mechanical degrees, by means of a protractor. Accent the 50-pF points and label the 100-pF points to make the dial easily readable.

A correction for amplifier output capacitance, including the rectifier diode and strays, may be made by resonating a standard 5-μH coil, using the 1000 switch position and setting the variable capacitor to read 200 pF.[3] Since the capacitance variation is linear with shaft rotation, calibration will be correct throughout its range.

Measuring Capacitance

To measure capacitance over the range of 10 to 333 pF, set the crystal switch to 1000 and resonate a 3-μH coil with the variable capacitor; note the dial reading.[4] The unknown capacitor is then connected across the coil, and resonance is established by tuning the variable capacitor. The difference between the two capacitor-dial readings is equal to the unknown capacitance, within the dial calibration accuracy. Frequency has no effect as long as it is constant; with crystal control this is no problem.

Construction

The mechanical layout of the circuit is not critical. As with all rf gear, use short, direct leads, with the output circuits separated from the input to prevent spurious oscillations. I used Vectorbord®, supported by the capacitor frame, to mount the rf circuits. The crystals were "epoxied" to the board with the pins protruding through, and wires were soldered directly to the pins. Amplifier output capacitance should be minimized by keeping leads short and away from grounded objects. I installed the circuit, including a surplus 3-1/2-inch meter and battery, in a steel sloping-front cabinet, as shown in Fig. 2.

Conclusions

There are a number of practical uses for

[1]Notes appear on page 27.

*10004 Belhaven Rd., Bethesda, MD 20817

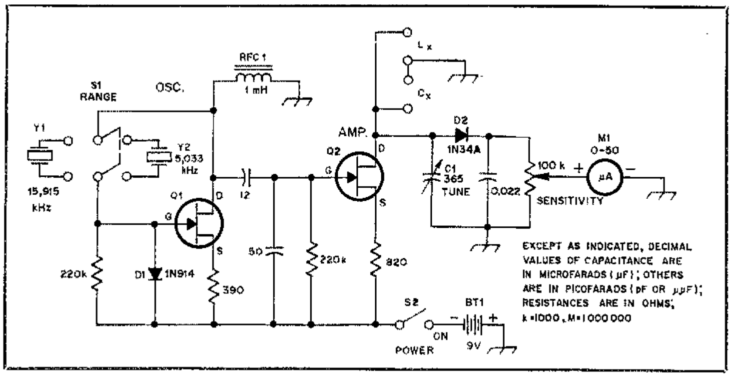


Fig. 1 — Schematic diagram of the L-C Meter. Resistors are carbon composition, 1/4- or 1/2-W; capacitors are disc ceramic, 100-V units.

C1 — 10- to 365-pF variable capacitor. See text for details.
D1 — 1N914 silicon switching diode.
D2 — 1N34A germanium point-contact diode.
M1 — Small dc meter, 50 µA.
Q1, Q2 — Silicon JFET, Radio Shack 276-2035 or equiv.
Y1* — 15,915-kHz quartz crystal, 32-pF loading, HC6/U holder, ICM type 434115 or equiv.
Y2* — 5033-kHz quartz crystal, 32-pF loading, HC6/U holder, ICM type 433115 or equiv.

*The exact crystal frequencies may be found by using the following equations:

$$Y1 = \frac{10^5}{2\pi}\ \text{kHz}$$

$$Y2 = \frac{10^4\sqrt{10}}{2\pi}$$

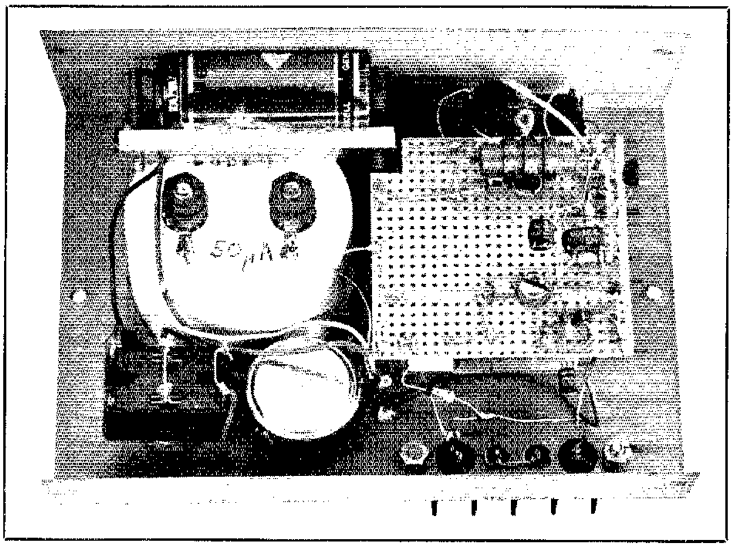


Fig. 2 — Interior view of the L-C meter.

a fast and simple inductance meter. Rather than fiddle with inductance formulas or calculators, it is much easier to wind too many turns on a form and measure the inductance, then remove turns until the desired value is reached. This is especially true for slug-tuned and toroidal coils when the permeability is not known.

Capacitors are usually labeled with their value, which negates the need for a capacitance meter. With time, however, labels deteriorate and capacitors open, leak or short. If the builder uses old or surplus stock, measurement may be the only way to go. QST

Notes

[1]From the resonance formula

$$L = \frac{1}{\omega^2 C} \quad \text{(Eq. 1)}$$

we require that

$$L = \frac{10^{-N}}{C} \quad \text{(Eq. 2)}$$

where N is any integer.

From the above

$$\omega^2 = 10^N \quad \text{(Eq. 3)}$$

so that

$$\omega = 10^{N/2} \quad \text{(Eq. 4)}$$

If N is an even number, N/2 is an integer, which we will call M. Then

$$f = \frac{10^M}{2\pi} \quad \text{(Eq. 5)}$$

f will have the number sequence of $1/2\pi$; the decimal position will be determined by M.

If N is an odd number, N − 1 is an even number that we will call P. Then

$$\omega^2 = 10^P \times 10^1 \quad \text{(Eq. 6)}$$

$$\text{and } \omega = 10^{P/2} \times 10^{1/2} \quad \text{(Eq. 7)}$$

But P/2 is an integer, which we will call K. Then

$$\omega = 10^K\sqrt{10} \quad \text{(Eq. 8)}$$

and

$$f = \frac{10^K\sqrt{10}}{2\pi} \quad \text{(Eq. 9)}$$

f will have the number sequence of $\sqrt{10}/2\pi$; the decimal position will be determined by K.

[2]The crystals are general-purpose (GP) types available from International Crystal Mfg., 10 North Lee, Oklahoma City, OK 73102.

[3]For details describing the 5-µH coil, see recent editions of the *ARRL Handbook* under Dip Meter, Measuring inductance and capacitance.

[4]I used a home made, 3-µH coil, consisting of 11 turns of number 20 tinned wire, with a total length of 5/8-inch. The exact value of this inductor is not critical, but the tank should resonate within a dial reading of 300 to 350 pF.

Strays

I would like to get in touch with . . .

☐ anyone having an instruction manual and circuit diagram for an OS-8C/U Navy Department oscilloscope, manufactured by Jetronic Industries, Inc., Philadelphia. Leo H. Hansen, WBØTDU, Star Rte., Box 2032, Virginia, MN 55792.

☐ other amateurs who are involved in the emergency services field, particularly EMTs and paramedics. Jeff Howell, WB9PFZ, P.O. Box 463, Madison, IN 47250.

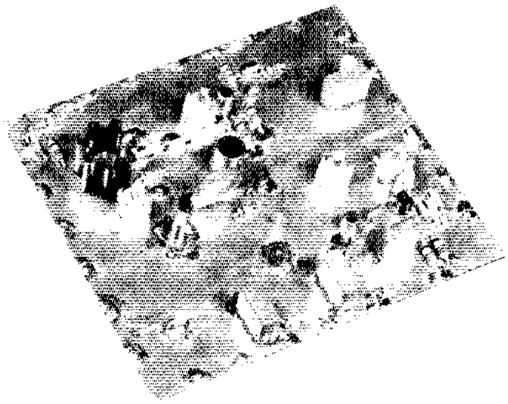
☐ University of Texas alumni who are hams, to join the UTARC as associate members. Lee Murrah, WD5CID, 5303 Scenic View Dr., Austin, TX 78746.

☐ any hams who were cadets at the Florida Military Academy in St. Petersburg. H. Vandergrift, WA4WME, 2308 Zinnia Ct., Killeen, TX 76541.

***QST* congratulates . . .**

☐ Frank Thornburgh, WA6GYR, of Pleasant Hill, California, on being appointed as a communications specialist at the Federal Emergency Management Agency (FEMA) Region IX Headquarters.

A High-Quality UHF Source for Microwave Applications†



Need a state-of-the-art signal source for the high bands? Build this one! It has what you may be looking for.

By The RSGB Microwave Committee*

This article describes a unit that delivers a minimum of 100 mW from a 12-V supply, at an output frequency of 360 to 440 MHz, using a fifth-overtone crystal in a range of 90 to 110 MHz. The output is intended for multiplication up to microwave frequencies for use as a local oscillator or transmitter. This signal still has a good quality note at 10 GHz. Members of the RSGB Microwave Committee developed it from the Plessey AMETS transmitter board.

Overview

The design uses a Butler crystal oscillator, eliminating frequency-multiplier stages. This circuit has a very low noise output compared with the commonly used single-transistor variety. Crystal loading is less, allowing a higher working Q to be realized. Leadless disc capacitors are used on double-sided circuit-board material for efficient decoupling.

The two amplifier stages are operated in Class A, contributing to the low-noise performance and making instability problems very unlikely. All inductors and lines are printed on the board; only four adjustments are needed to align the unit. We made provision for on/off keying of the output stage, or for applying fm or fsk to the oscillator. Several dozen of these boards have been built and used in a variety of microwave applications, giving excellent results.

Construction

The unit is constructed on 0.063-in.-thick, double-sided, copper-clad fiberglass epoxy board (dielectric constant is approximately 5) with ground plane on the component side of the board.[1] Using materials with different dielectric constants may result in incorrect resonant frequencies for the tuned circuits; you should avoid this. The circuit diagram is given in Fig. 1.

Construction is generally straightforward; the layout of components on the board is shown in Fig. 2. It is important that the component values be adhered to and in particular that only new, branded semiconductors be used for Q1-Q4, inclusive; surplus types usually do not perform satisfactorily. The grounded ends of the components are soldered to both top and bottom ground planes. All components should be mounted with absolute minimum lead lengths. Q2, Q3 and Q4 should be pushed down flush with the top side of the board. Solder the cans of Q2 and Q3 to the ground plane. Q4 is fitted with a TO-5 heat sink.

Take care mounting the 1000-pF leadless disc-ceramic capacitors. They are located in slots in the circuit board, made by drilling several holes close together and then joining them to form a slot. You can do this quite easily using the tool shown in Fig. 3. Break off the end of a hacksaw blade and cut or file the end to a taper. The teeth should be oriented as shown. The point is thin enough to fit into one hole at the end of the row. Hold the hacksaw blade quite close to the board, and use it gently to prevent it from snapping off. If it breaks, the blade can easily be repaired by filing it to a point again.

Another method uses a no. 60 (1 mm) drill bit in a vertical drill as a mill. Hold the drill bit in the chuck so only 1/8 to 1/6 inch protrudes. Move the board sideways so that the edge of the drill bit cuts the slot, joining the holes. Hold the board against a straight edge to ensure a straight slot. If the drill breaks, it can still be reused. Clean the edges of the slot with a flat needle file or nail file. Do not make the slot too wide; otherwise, it will be difficult to solder the discs, which should be positioned and soldered exactly as shown in Fig. 4. Ensure that the solder flows properly on the metal of the capacitors. Various "surplus" leadless discs gave trouble in this respect on some of the prototypes. We found the most reliable types to be those made by Steatite[2], which are actually trapezoidal (coffin-shaped) rather than round. They proved extremely easy to solder in place, and are recommended.

For other types, a useful technique is to heat the edge of the slot first, insert the disc, touch the soldering iron to the disc (not the board), and then apply solder to form a neat fillet. If silver-loaded solder can be obtained, then this is also worth using.

The ground planes on both sides of the board are joined at many points by short lengths of no. 20-24 wire (e.g., scrap component leads) soldered in all the remaining

†Adapted from an article of the same title in *Radio Communication* (RSGB), Oct. 1981.

*Radio Society of Great Britain, Alma House, Cranborne Road, Potters Bar, Herts. EN6 3JW, England

[1]Notes appear on page 32.

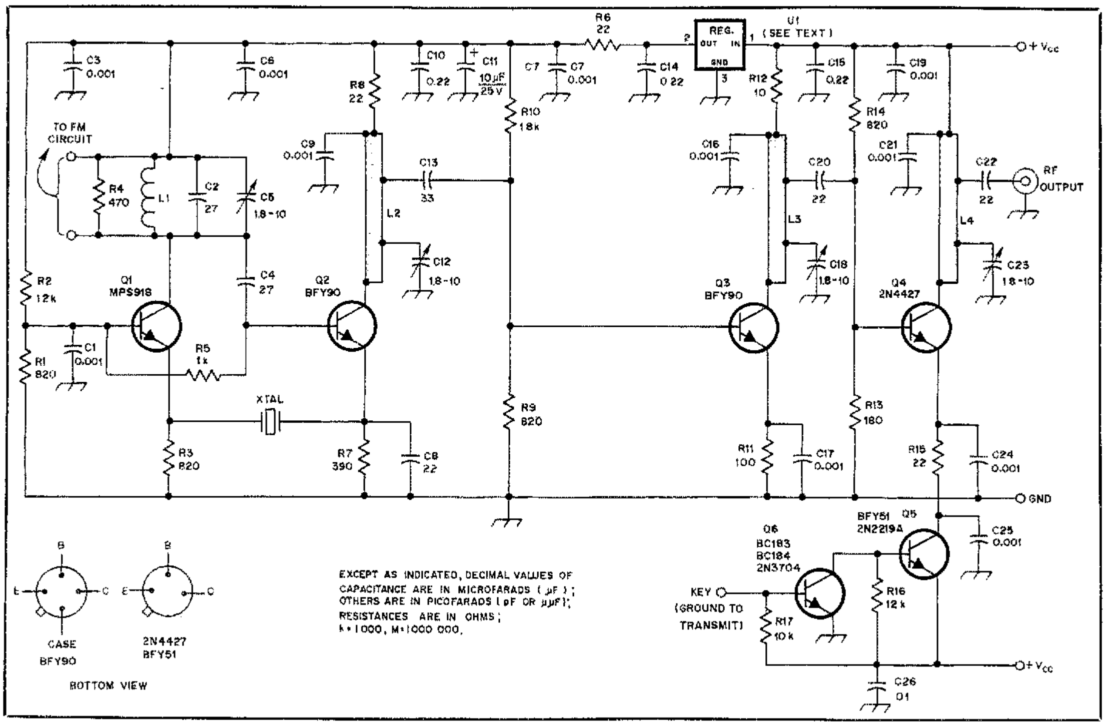


Fig. 1 — Circuit diagram of the signal source. For best results, construct using the board layout found in the Hints and Kinks section of this issue. All fixed-value resistors are 1/4-W, carbon composition type. Variable resistors are 1/4-W pc-board mounting type. Capacitors are subminiature disc ceramic, unless otherwise specified.

C1, C3, C6, C7, C9, C16, C17, C19, C21, C24, C25 — 0.001-μF leadless disc, Steatite 0.001 μF/80/20 TEFK7 400 V or equiv.
C5, C12, C18, C23 — 1.8-10-pF film trimmer, Mullard 809-05002 or equiv.
C10, C14, C15 — 0.22-μF polyester.
C11 — 10-μF, 25-V electrolytic.
Q1 — Npn, silicon rf/i-f amplifier, 250 mW, MPS918.
Q2, Q3 — Npn, silicon, rf amplifier, 1 W, BFY90.
Q4 — Npn, silicon, power amplifier, 1 W, 2N4427.
Q5 — Npn, silicon, power amplifier/switcher, 1 W, 2N2219A or equiv.
Q6 — Npn, silicon, general purpose, 500 mW, BC183 or equiv.
U1 — Voltage regulator IC, see text.
Xtal — HC18/U or HC25/U fifth overtone.

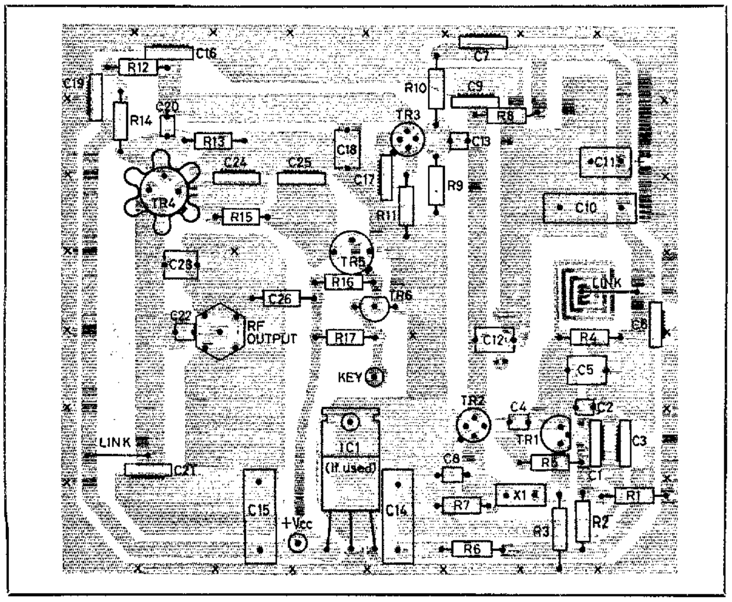


Fig. 2 — Parts-placement guide for the signal source. Parts are placed on the ground plane side of the board. (The etching patterns for both sides of the board appear in the Hints and Kinks section of this issue.) X indicates jumpers joining top and bottom ground planes. [Several parts labeled on the diagram are not in *QST* style — Ed.]

holes that are not cleared on the upper side of the board. These are located all around the edge of the board, as well as in a number of other places. It would be possible to replace the wires around the edges of the board with thin copper foil, folded and soldered along the edges of the board, connecting the two ground planes.

Do not forget the two wire links on the board connecting L1 and L4 to the positive supply. Holes are provided for either the 2- or 3-pin types of trimmer capacitor, which should be mounted so that their movable plates are connected to minimize stray capacitance coupling and loading effects when trimming tools are used.

The output can be taken by small-diameter coaxial cable directly from the board, with the cable braid soldered carefully to the top ground plane (Fig. 5). Or, you can use an SMB, SMC (Conhex), or SMA socket on the board.

We made provision on the board for a TO-220-style voltage regulator for the oscillator supply. This reduces the chirp on the note considerably during keying. These regulators generally require an input voltage at least 2.5 V greater than the regulated output voltage. If a 12-V one is used (e.g., 7812), the supply to the board should be at least 14.5 V. If only a 12-V supply is available, use an 8-V regulator (7808) — though this may reduce the output slightly — or omit the regulator altogether and insert a wire between the two outermost regulator holes.

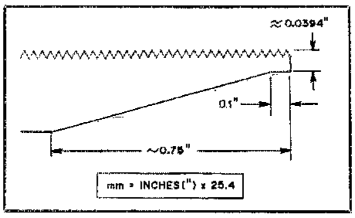


Fig. 3 — Slot-cutting tool.

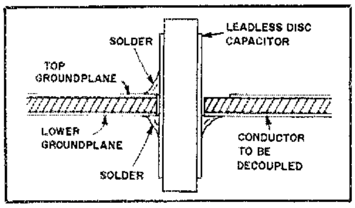


Fig. 4 — Soldering of leadless disc capacitors.

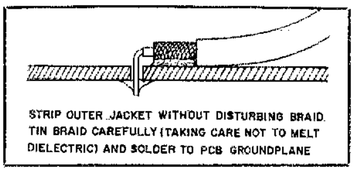


Fig. 5 — Soldering of output cable to board.

Modulation Facilities

The PA stage can be on/off keyed by Q5 and Q6; however, this pulls the oscillator frequency slightly. At 10 GHz, the keying chirp is unacceptable unless the oscillator voltage regulator (U1) is used, and even then it is still noticeable. Frequency-shift keying (fsk) is the preferred method. If you do not plan to use A1 keying, Q5, Q6 and associated components should be omitted. Ground the end of R15 previously connected to Q5.

Frequency modulation or fsk can be produced using the circuit in Fig. 6. A varactor diode varies the capacitance across L1. Reduce the value of C2 to maintain the total capacitance at 27 pF. The BB105B (C = 7 pF at −5 V), with C2 = 18 pF, should give a deviation of about 2 kHz/V at 10 GHz. Wider deviation can be obtained by using a higher capacitance diode, such as the BB110G, or a hyper-abrupt type, which has a larger capacitance swing (e.g., the Alpha DKV6520)[3].

The circuit has inputs for audio at low impedance (source impedance less than 1 kΩ), or for cw where grounding the input shifts the carrier about 1 kHz at 10 GHz; the shift can be altered by changing the 47-kΩ resistor.

The components to the right-hand side of the dashed line should be mounted directly across the tuned circuit. The rest can be mounted in place of the A1 keying circuit if this is not required, but in any case they should be mounted on the board to avoid ground-loop problems. Efficient filtering is necessary on both inputs to prevent any stray af or rf noise causing unwanted sidebands on the output.

Alignment

The tune-up for this unit requires the use of some form of output indicator. An hf/vhf SWR meter connected to the board via a short length of coaxial cable can be used, set initially on the most sensitive range, with a 50-Ω load or resistor on the output socket. We recommend, however, that the power meter described below be used, because it is a simple, yet remarkably accurate, way of measuring power in the range 100 mW to 2 W, up to at least 500 MHz. Constructional details are given in Fig. 7.

The meter indication, V, is the peak rf voltage across the load minus the forward voltage drop across the diode. This is divided by $\sqrt{2}$ to give the rms value, and the power is calculated from this using the equation

$$P = \frac{(V_{rms})^2}{R} \quad \text{(Eq. 1)}$$

For germanium diodes, such as the OA47, the relationship will therefore be

$$P = \frac{(V + 0.25)^2}{100} \quad \text{(Eq. 2)}$$

while for silicon diodes, such as 1N914, 1N916 and 1N4149, it is

$$P = \frac{(V + 0.7)^2}{100} \quad \text{(Eq. 3)}$$

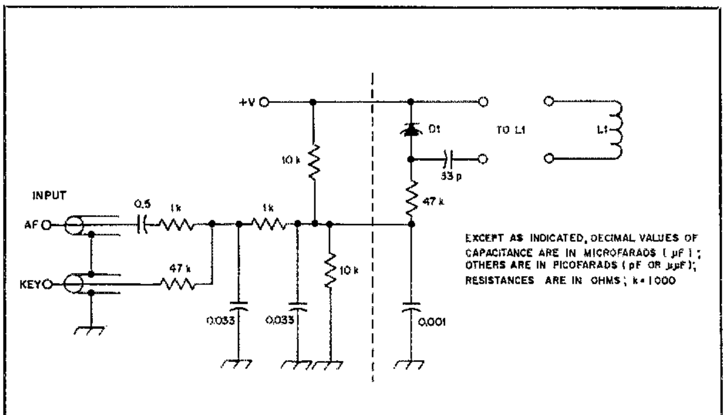


Fig. 6 — Schematic diagram of network for fm/fsk modulation. Resistances are in ohms; k = 1000. Resistors are 1/4-W, carbon-composition types. Capacitances are in microfarads, except those with whole-number values, which are in picofarads. All capacitors are disc ceramic. D1 is discussed in the text.

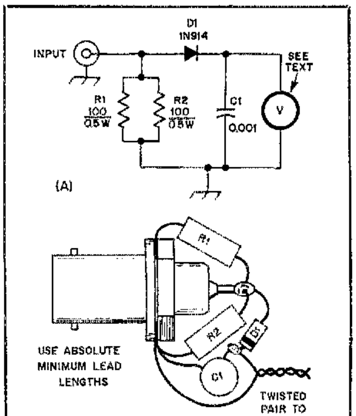


Fig. 7 — At A, schematic diagram of simple power meter used for aligning signal source. See text for discussion. At B, construction details for simple power meter.

The results from this meter using a 1N914 at 500 MHz are compared with the actual power in Fig. 8, and show the high accuracy obtainable.

Apply power to the board, and check the current drawn (approximately 150 mA). For a 96-MHz crystal, preset the trimmers to the following positions: C5 about 50% meshed, C12 about 40% meshed, C18 about 80% meshed and C23 about 50% meshed. Some output should be evident; maximize it by peaking the trimmers.

You should not be able to tune up on the wrong frequency using the component values specified, but if in doubt, check them with an absorption wavemeter or frequency counter. The position of C5 should not be too critical with regard to output power, and may be used to trim the frequency. If crystal frequencies outside the range 90 to 100 MHz are used, it may be necessary to alter the value of C2. You should be able to adjust C5 over about half its travel, with the power output staying constant and the frequency pulling smoothly. If necessary, the frequency can be raised slightly by putting a 5- to 20-pF trimmer in series with the crystal. This may reduce the power output or stability; it is preferable to put up with a small frequency offset (which can be calibrated out) rather than compromise on stability.

When the unit is tuned up it should produce a comfortable 100 mW of rf output. It may be possible to increase this slightly by optimizing certain components (e.g., R10, C8, C22). If more power is needed, it is preferable to add an extra stage, rather than risk detracting from the stability, low-noise sidebands and reliability of the oscillator board.

Decoupling, Shielding and Supply Regulation

The board is designed to fit in an STC (ITT) 6-1/2 × 3-1/2 inch diecast box (part number 46R CS00 043 A00) with the internal ribs filed to size. Mount the board with bolts in the corners, about 1/4 inch off the bottom of the box. This avoids excessive damping of the lines by the box, and ensures that the heat sink on Q4 does not touch the lid of the box. This, or similar shielding, is recommended for best results, because any feedback from subsequent stages to the oscillator can degrade the signal. It is essential to prevent any rf or af noise from reaching the board through the power-supply leads. This can cause either a noisy carrier or sidebands on the output. All connections (e.g., supply, keying, etc.) must be thoroughly decoupled. Use a feedthrough capacitor where the leads pass through the box. Install a small 10- to 100-μH rf choke outside the box, in series with the supply. If possible, check with a wide-bandwidth oscilloscope for oscillating voltage-regulator circuits — possibly at quite a low level. This can occur anywhere from audio to tens of megahertz.

Ground the board to the box at several places. Secure the lid firmly in place, because any intermittent contact here may cause frequency jumps. Ground the outer conductor of the rf output where it passes through the box by mounting a socket on the box. In general, the amount of effort that must be put into these precautions depends on the factor by which the frequency is to be multiplied.

The on-board regulator, U1, is probably only worth using if you intend to use

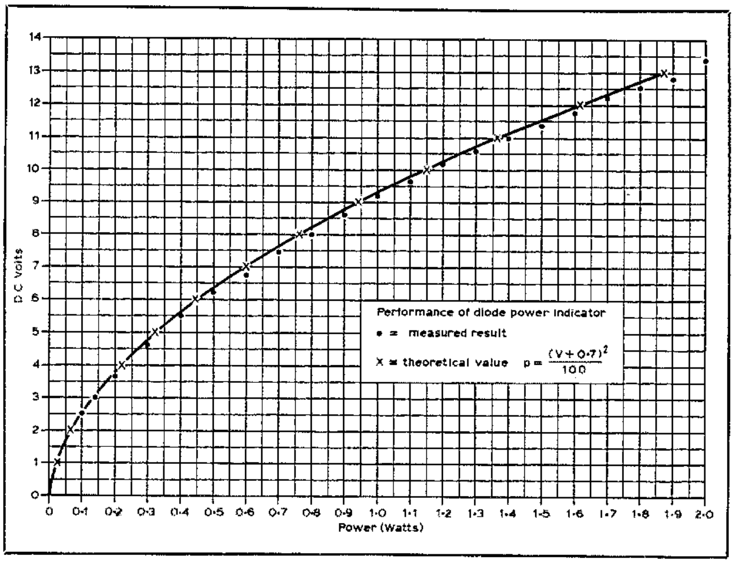


Fig. 8 — Comparison of expected values versus measured results for the power meter.

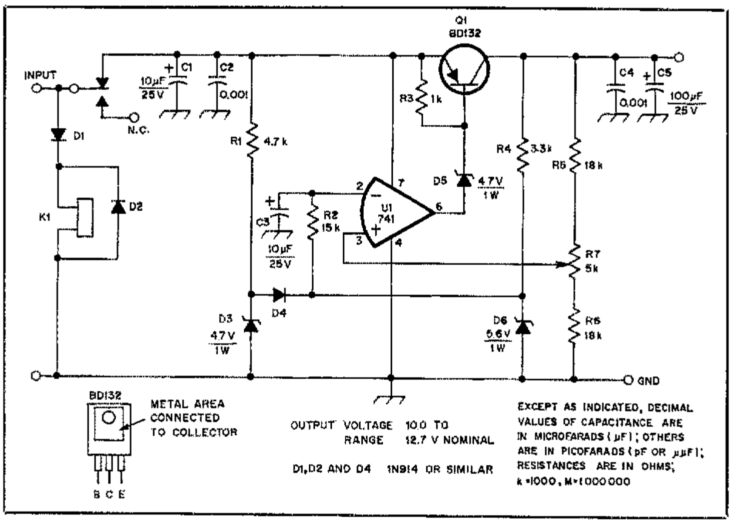


Fig. 9 — Schematic diagram of the low voltage/drop regulator circuit. All fixed-value resistors are 1/4-W, carbon composition types. Variable resistor is 1/4-W, pc-board mounting type. Capacitors are disc ceramic, unless otherwise specified.

C1, C3 — 10-μF, 25-V electrolytic.
C5 — 100-μF, 25-V electrolytic.
D1, D2, D4 — Silicon, small-signal diode, 1N914 or equiv.
D3, D5 — Zener, 4.7 V, 1 W.
D6 — Zener, 5.6 V, 1 W.
K1 — Spdt relay, contacts rated for 1 A at 28 V, Radiospares 348-510 or equiv.
Q1 — Silicon, pnp, power, 90 W, BD132 or equiv.
U1 — Operation amplifier, 500 mW, type 741 or equiv.

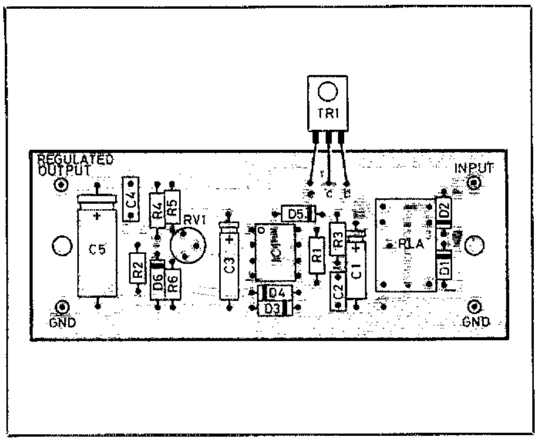


Fig. 10 — Parts-placement guide for voltage-regulator circuit. Parts are placed on the nonfoil side of the board; the shaded area represents an X-ray view of the copper pattern. (The etching pattern appears in the Hints and Kinks section of this issue.) [Several parts labeled on the diagram are not in *QST* style — Ed.]

Table 2
Frequency Stability (at 10 GHz)

Variation of frequency with supply voltage
without U1: 5 kHz/V
with U1: 100 Hz/V (once regulating)
with regulator of Fig. 10: hardly detectable

Variation of frequency with temperature
2 kHz/°C (depends on crystal)

Table 3
RF Output frequencies and Crystal Frequencies for Various Applications of the Board

Application	*RF Output Freq. (MHz)*	*Crystal Freq. (MHz)*
Microwave cw/fsk/fm tx	384	96
1.3-GHz converter/transverter (144-MHz i-f)	384	96
2.3-GHZ converter/transverter (144-MHz i-f)	360	90
3.4-GHz converter/transverter (144-MHz i-f)	368	92
5.7-GHz converter/transverter (144-MHz i-f)	374.4	93.6
10-GHz converter/transverter (144-MHz i-f)	378.666	94.666
24-GHZ converter/transverter (144-MHz i-f)	381.714	95.4286

Table 1
Output Power vs. Supply Voltage

Supply Voltage (V)	*Output Power (mW)*
11	225
12	270
13.5	325
15 (without U1)	350
15 (with U1)	330

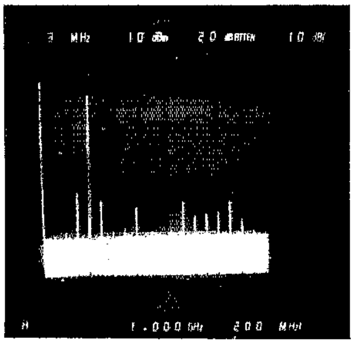


Fig. 11 — Spectral display of rf output from the signal source. Vertical divisions are each 10 dB; horizontal divisions are each 200 MHz.

the A1 keying facility, or if a + 15-V supply is available. Without the regulator, the circuit will operate satisfactorily from a supply as low as 11 V, with little drop in output.

For best frequency stability, the whole circuit must run from a regulated supply. This should be the same voltage whether for portable or fixed operation, so that the frequency calibration is maintained. This is particularly important when using car batteries, as the voltage varies somewhat according to the state of charge, current drain, etc. (perhaps as much as 11.5 V to 16 V), and the same is true of some other battery types. The circuit shown in Fig. 9 makes the oscillator insensitive to supply voltage variations, and provides over-voltage and reverse supply protection. It will regulate down to a minimum voltage drop of approximately 300 mV (governed by the transistor V_{ce} and saturation). Fig. 10 shows the parts layout for the etched board.

The relay used for reverse supply protection is rated for 1 A at 28 V. We highly recommend this circuit for all portable gear needing a regulated supply up to this rating. R1 sets the regulated output voltage; 11.2 V is a reasonable value that will guarantee a regulated output even at the lower supply voltages. Q1 is bolted to a suitable heat sink, with the usual insulating washer.

Performance

Table 1 gives typical performance figures measured on the board shown in the photograph. The output spectrum is shown in Fig. 11. This varies little with supply voltage. Unwanted products are better than 35 dB down from the main carrier. Table 2 shows the frequency stability at 10 GHz under varying conditions.

Applications

Most of the units built so far have been used on 10 GHz to provide a local oscillator for the G3JVL 10-GHz transverter[4] in conjunction with the G8DEK step-recovery multiplier design.[5] Most of our comments refer to measures required to give a good note at 10 GHz. For this application, a 94.666-MHz crystal is used to give an output at 378.666 MHz, for a final local-oscillator frequency of 10.224 GHz. An output power of about 2.5 W is needed, so some form of additional amplification is required. This amplifier should not be mounted in the same box as the board.

Amplifiers that have been used with success include the Mullard BGY22 module (which requires an attenuator[6] at the input to reduce the power from the board to approximately 50 mW) and the Wood & Douglas MD3PA. While the former is very simple and convenient, the latter has better spectral purity. We recommend the Mullard BGY22/BGY23 combination or the Wood & Douglas MD10PA[7] as higher power amplifiers for transmitter-type applications. Crystal frequencies and rf output frequencies for various applications are given in Table 3.

Conclusion

Although the unit is quite simple to construct and align, there is no compromise in performance. The built-in versatility should eliminate all problems of constructing a new local-oscillator strip design for each new microwave project.

Notes

[1]mm = in. × 25.4.
[2]Steatite-Roederstein Ltd., Hagley Rd., Birmingham B16 8QW, England.
[3]Alpha Industries, RMC House, Station Rd., Witney, Oxon OX8 6BP, England.
[4]"G3JVL 10 GHz Transverter," *Rad Com*, Jan. and April 1979, pp. 41, 342; April 1980, pp. 372-5; Feb. 1981, p. 146.
[5]"G8DEK BXY41E 10 GHz step-recovery diode multiplier," *Rad Com*, March 1976, p. 202.
[6]"Design and construction of simple attenuators," *Rad Com*, March 1979, p. 239.
[7]"Equipment review — Wood & Douglas 384 MHz MD05T microwave drive source and MD10PA power amplifier kits," *Rad Com*, June/July 1980, pp. 650-652.

Tracking the Terrible TVI

"If I knew what the problem was, I'd fix it!" Here is some practical help toward reducing troublesome TVI.

By C. L. "Chuck" Hutchinson,* K8CH

Brass pounder, Charley the Clown,
Kills TVs all over the town,
When he dahs and he dits,
The screens have three fits,
And there isn't a trace of a sound.

TVI! Oh what will I do?
For I really haven't a clue,
Do I start here or there,
Or just stay off the air,
Until it's a quarter of two?

A neighbor just called and he's mad,
The TVI is getting quite bad,
The pass high and low,
I have given a go,
So now I am feeling quite sad.

TVI! he cried in despair.
I think I'll pull out my hair!
*But he bought him a book,***
And he gave it a look,
And now he's back on the air.

**the ARRL *Radio Frequency Interference* book

Trouble with TVI? If your Amateur Radio transmissions are interfering with TV reception in the neighborhood, you are not alone. On a recent Friday, 25 ARRL members called the Technical Information Service (TIS) desk. Every one of them had a TVI problem; there were no other calls that day.

Why do you have TVI? Where do you start to track down the source? What do you do to cure it? This article will help you answer these questions.

The usual reason for TVI is a lack of compatibility between your transmitter and the affected TV receiver. Harmonics or parasitics from the transmitter can result in unwanted rf energy in the TV band. Conversely, a TV set or amplifier may not be capable of rejecting strong out-of-band signals; the resulting overload condition causes interference to be generated in the set itself. There are various ways for rf energy to get into a TV set; not all of them may be obvious. To further complicate matters, interference may be generated in any nonlinear junction and then picked up by the TV receiver.

Rf energy can be propagated by radiation (as from an antenna), induction (as in a transformer) or conduction (as along a wire). Does all this begin to sound pretty complicated? Don't despair! By a process of elimination you should be able to locate the source and minimize the TVI that plagues you.

Where to Start

Your TVI tracking efforts should begin at home. It is easier to coordinate tests when the transmitter and receiver are in the same house — it's not necessarily easier to solve the problem, just easier to coordinate your efforts. What you have learned at home will better prepare you to deal with TVI at your neighbor's house. If you have no TVI on your own set, congratulations are in order. Without taking preventive measures, most of us are not that fortunate.

The following suggestions are for you to use on your own TV set. Be careful of what you do to your neighbor's set. It is not your responsibility to make modifications or additions to his equipment. If you and your neighbor agree to make circuit changes then you are free to do so. Be prudent and exercise your best judgment before you get involved.

FCC experience indicates that in most TVI cases, significant improvement would result if TV receivers were properly designed for RFI rejection. That was recognized with the recent enactment of P.L. 97-259.[1] This new law should help the situation after the implementation of appropriate rule changes.

Some manufacturers have taken a responsible approach to the problem and are willing to help. A list of these manufacturers, with addresses, can be found in May 1981 *QST*, or in Chapter 8 of the second edition of *Radio Frequency Interference*.

It is important that you contact the manufacturer. Any TVI or RFI problem involving consumer entertainment equipment should be put "on the record" by writing a letter. If you call on the telephone, be sure to follow up with a letter to the manufacturer. Send copies to the FCC, the EIA and the ARRL RFI Task Group.[2] The company should have

*ARRL Assistant Technical Editor

[1]Notes appear on page 35.

the opportunity to correct any deficiency and to hear your complaint. Copies to the other agencies are vital to ensure that the magnitude of the problem is understood by all. Your silence is an indication that no problem exists!

Fig. 1 is a trouble-shooting flow chart designed to help you track down TVI in a logical, step-by-step process. To begin, transmit into a dummy load and check your TV set for interference. If you see any, check transmitter tuning, neutralization and operating voltages — especially bias. If these are okay, the fault may be caused by rf traveling into the TV set on the ac line, improper shielding and bonding of the transmitter cabinet or improper grounding.

To eliminate rf energy on the ac line, you can use a brute-force ac line filter as shown in Fig. 2 (mount the filter at or in the equipment). Alternatively, you can wind most or all of the line cord around a 7-inch ferrite rod.[3] Rod permeability (U_i) should be at least 800. Sometimes a 0.01-μF 1400-V disc ceramic capacitor mounted in a power plug can help squelch rf on the ac line, when it is plugged into an outlet. You might make up two or three of these bypass plugs to experiment with. Placement is usually critical.

Don Noble, NR4V, reports an interesting CATVI problem. A neighbor experienced TVI when Don ran a kilowatt on 40-m cw. Unplugging the HBO converter box cleared up the problem. A brute force line filter did no good. Don solved the problem by wrapping the full length of the power cord on a ferrite rod. A short extension cord was added to reach the outlet. Problem cured and case closed!

Even the best transmitters generate internal harmonic and other unwanted signals. As long as these signals are confined inside the cabinet, they pose no problem. Improper shielding and bonding of the transmitter cabinet can cause TVI by chassis radiation through openings in the cabinet. In addition, it can allow unfiltered rf currents to flow on the outside of the coaxial cable braid.

Because of skin effect, a good quality coaxial cable has three rf conductors: the inner conductor, the inside of the outer conductor and the outside of the outer conductor. Currents flowing on the outside of the braid will flow unimpeded around low-pass filters and traps. Fig. 3 illustrates how to stop these unwanted currents by using a shield choke.

Obtaining an effective ground frequently requires a bit of experimentation. The normal rules of thumb apply; drive a copper-coated ground rod and keep the lead to it as short and wide as possible. If that works — fine! Don't despair if it does

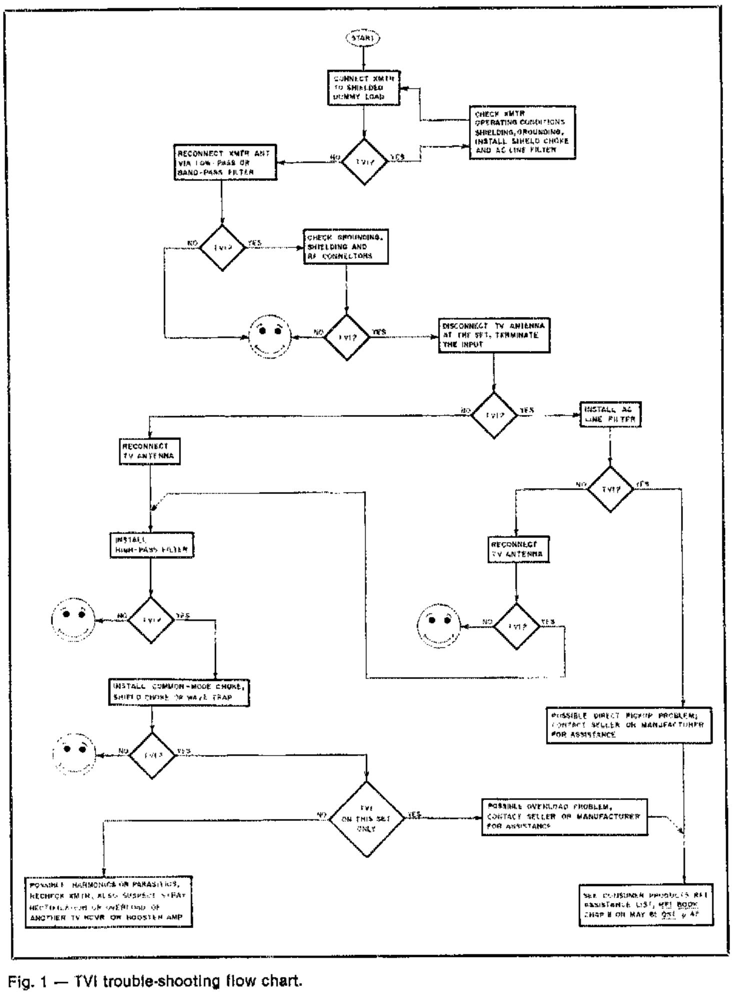


Fig. 1 — TVI trouble-shooting flow chart.

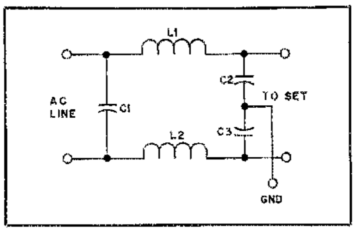


Fig. 2 — "Brute-force" ac line filter. The values of C1, C2 and C3 are not critical; capacitances from 0.001 to 0.01 μF at 1400 V can be used. L1 and L2 can be a 2-inch winding of no. 12 enameled wire (no. 18 for receivers) on a 1/2-inch diameter form. Make sure that there are no exposed conductors to offer a shock hazard.

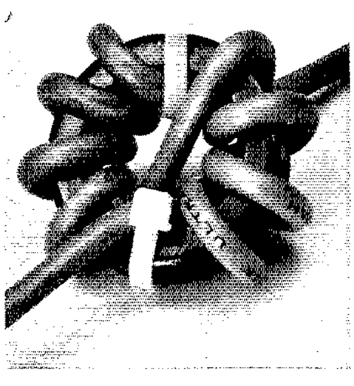


Fig. 3 — A shield choke can be formed by winding the cable on a ferrite toroid. Reversing the winding as pictured allows more turns with less shunt capacitance. The most important property of the cable is complete shielding — avoid "bargain" cable having less than 95% braid coverage.

not. You may have to put in more than one ground rod, each with its own strap connecting to the equipment. Fastening 1/4-wavelength radials to your equipment can provide a pseudo ground. Use at least two for each band you use. Bonding to water pipes and heating ducts may or may not help lower the Q and increase the effectiveness of your ground system. (Doug DeMaw, W1FB, once used a Transmatch to "tune" the ground wire that ran from his second-story ham shack.)

When the transmitter causes no TVI while it is operating into a dummy load, you are ready to proceed to the next step. Buy or borrow a low-pass or band-pass filter, and install it in the antenna feed line. If TVI persists, check grounding, shielding and the tightness of the rf connectors again. Uhf and even BNC type connectors can "leak" a lot of rf energy; type N connectors are best. If you still have TVI, it's time to go to work on the TV receiver.

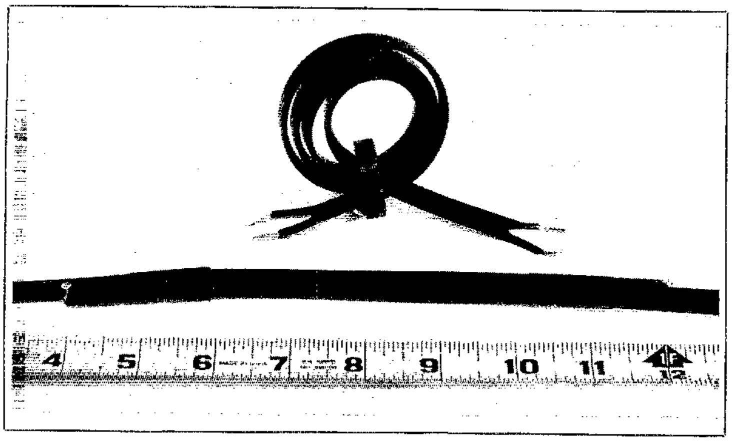


Fig. 4 — Common-mode chokes. At the top, several turns of twin-lead are taken through an FT-114-43 ferrite toroid; best results will be obtained if the twin-lead has an oval cross section. At the bottom, the twin-lead has been cut near the TV set. The conductors have been shorted at the resulting ends and the ends have been overlapped approximately 9 inches. Shrinkable tubing or tape is used to hold the two pieces together.

The TV Receiver

Unwanted rf can get into the TV set by way of the antenna terminals, the ac line cord and by direct pickup. Eliminate the antenna terminal route as the source by disconnecting the antenna and replacing it with a resistor (300 ohms for balanced, 75 ohms for coaxial, inputs; keep the leads as short as possible). If the interference persists, suspect the ac line as the source of entry.

The same ac line filtering techniques described earlier are used for the TV set. When you see no interference, reconnect the TV antenna to the set via a high-pass filter.

High-pass filters can be purchased from a variety of sources. Not all manufactured filters give satisfactory performance. Effective, low-cost filters were described in "Practical 75- and 300-Ohm High-Pass Filters," by Ed Wetherhold in February 1982 *QST*. Of particular interest is the 300-ohm elliptical filter design.

There are two situations in which a high-pass filter is ineffective. The first is when interference-causing rf is flowing on the outside of a coaxial feed-line braid. That can be cured with a shield choke. The second situation occurs when using balanced feed line. Antenna currents are push-pull or differential-mode signals; this type of signal exists as a voltage difference between the feed-line conductors. Frequently a TV feed line will act as an antenna at hf. When this happens, there is no voltage difference between the feed-line conductors; the signal exists between the conductors and ground. This is called a common-mode signal; regular high-pass filters will not stop it. A wave trap will, or you can use a common-mode choke, as shown in Fig. 4.

If you have gone through all of the preceding steps and are still plagued with TVI, try replacing the TV set with another. If the replacement is clean, then the first set is evidently at fault; if not, some other possibilities should be checked out. The first is to inspect the TV antenna and feed line. Look for loose or broken wires and corroded contacts. These can act as rectifiers, generating harmonics. (Don't forget to check your transmitting antenna, too.) Poor contact between metallic objects can be the source of stray rectification. Guy wires, pipes, rain gutters and down spouts are potential culprits. One ham found the source of cable TVI to be his old TV antenna and feed line — when they were removed, the problem went away!

Mast-mounted preamplifiers have been a source of trouble. Channel Master has a new series with improved out-of-band rejection. If a preamplifier is needed, one of these or a similar unit is recommended. Removing power from an amplifier will not eliminate it as a source of stray rectification — it may make the problem worse. An unused preamplifier and antenna combination in a neighbor's attic sent Rick Smith, WB4MRW, on a month-long TVI search. This search began after the FCC imposed quiet hours despite the clean bill of health they gave Rick's transmitter.

Conclusion

TVI can be difficult to tame even though FCC studies indicate that most cases can be cured by installing a low-pass filter on the transmitter and a high-pass filter on the TV set. A logical step-by-step approach should lead you to the interference source; the troubleshooting flow chart in Fig. 1 should help. Further information can be found in *Radio Frequency Interference* and in the interference chapter of *The Radio Amateur's Handbook*.

Most of the time you'll be able to resolve TVI problems by applying the simple remedies described here or in the ARRL publications mentioned above. If you are a League member who has tried these remedies but cannot solve the TVI problem, you can write Hq. for assistance from the TIS staff. When you write, be sure to include your name, call sign, membership expiration date and a stamped, self-addressed, business-size envelope. Describe the nature of the TVI complaint as completely as you can. Detail what steps you have taken and the results obtained.[4]

Local radio amateurs and club groups can be of help, and they are close at hand. Don't give up in despair! Seek out information and assistance, if you need help. Play the role of the great detective. Find and use your clues. Above all, be persistent! Don't stop at a dead-end — go back and recheck what you have done. It may not be easy, but you can minimize or eliminate that TVI.

Notes

[1] W. Clift, "RFI Bill Becomes Law," *QST*, November 1982.

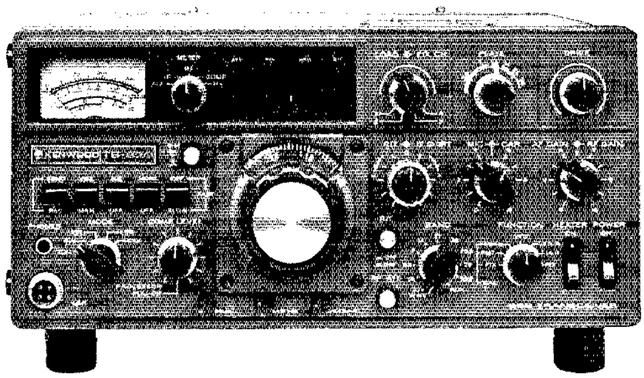
[2] A sample letter can be found in the ARRL book *Radio Frequency Interference*. Copies of the letter should go to:
Federal Communications Commission
1919 M St., N.W.
Washington, D.C. 20554
Electronic Industries Association
2001 Eye St., N.W.
Washington, DC 20006
American Radio Relay League
225 Main St.
Newington, CT 06111
(ATTN: RFI Task Group)

[3] mm = in. × 25.4

[4] See December 1982 *QST*, page 23, for a complete list of TIS guidelines.

WARC Bands for the TS-820(S)

Now you proud TS-820(S) owners have one more reason to hang on to your favorite rig. These modifications will update your transceiver for operation on the WARC bands.



By Robert C. Cheek,* W3VT/ex-JF1YBU

If you purchased a Trio-Kenwood TS-820 or TS-820S shortly before WARC 79, you knew something would be missing when use of the new bands was authorized.[1] All the major manufacturers' current hf transceiver models now include the new bands

Trade or Modify?

I am reluctant (and you probably are, too) to trade in a practically new transceiver just to get the WARC-band capability. I began studying the TS-820(S) circuits to see what it would take to allow it to operate on at least one of the new bands, preferably 10.1 MHz. To make a long story short, my TS-820S is now ready to operate on all three! It retains the original band capabilities except for the 200-kHz slice from 29.5 to 29.7 MHz — a frequency range in which I don't usually operate, and don't expect to.

The modifications are simple and straightforward. They require no panel or chassis drilling and the external appearance of the unit is unchanged. You can make these modifications for any number of the new bands, for receiving only or for full transceiving capability. If you add all three bands the 24-MHz addition will require some minor surgery on the coil pack pc board. However, all the modifications are reversible, so the unit can be restored to original condition at any time.

I used the WWV position (15 MHz) of the BAND switch for the 10-MHz band. This permits reception of WWV at 10 MHz and full transceive operation in the 10.1- to 10.15-MHz segment. For 18 MHz (18.068 to 18.168 MHz), I used the AUX position and for 24 MHz (24.890 to 24.990 MHz), the 29.5-MHz switch position.

Fig. 1 shows circuit changes made in the final amplifier compartment. Fig. 2 details the coil pack board changes. PLL system changes are not shown in schematic form because they involve only direct component substitution or addition on the PD and VCO PLL system boards.

Parts and Tools Required

Table 1 lists the parts required. PLL system PD unit crystals in HC-18/U holders will have to be ordered from a crystal supplier since Kenwood does not supply crystals except for direct replacement purposes.[2]

The coils used must be similar physically to the originals. They are common pc-board mounting types. All the coils used have five mounting pins. The 10- and 18-MHz drive coils (unshielded) are wound on 5/16-inch diameter phenolic forms with tuning slugs.[3,4] The four corner pins are on a 9/32-inch square. The VCO coil for 18-MHz is in a 9/32-inch square shield can, and the terminal pins are on a 3/16-inch square. Antenna and mixer coils for 18-MHz are in 3/8-inch square shield cans with corner pins on a 9/32-inch square. Many of the coils found in the coil-assortment packs available from Radio Shack stores are wound on identical forms, with and without shield cans.

If you use forms from such an assortment, be sure they are coils that contain cores useful at hf. A GDO is essential for "home brewing" coils anyway, and the coil Q can be judged roughly when you are checking the coil tuning range. A pronounced dip should be obtained when the parallel combination of a coil and resonating capacitor is probed. Don't use 455-kHz i-f transformers, those with internal capacitors and windings of many fine-wire turns. On the other hand, 10.7-MHz i-f coils in 3/8-inch square shield cans, often found in such assortments, are ideal. Remove any link winding and check the tuning range (using a GDO) with a 100-pF capacitor across the main winding. If 18 MHz is well within the range, the coil should tune satisfactorily in the antenna and mixer circuits at 18 MHz. If not, remove some coil turns and try again. Likewise, the driver coils for 10 MHz and 18 MHz should be tunable to those frequencies with a 100-pF capacitor in parallel. The 18-MHz VCO coil should tune to 27 MHz with a 10-pF capacitor across it.

For all coils except the VCO coil, the active terminals are the outside two of the three in-line terminals. For the VCO coil, the active terminals are the left and center pins as the coil is viewed from the bottom with the three in-line pins in the bottom row.

Although I wound my own coils on some forms salvaged from surplus pc boards, a more direct approach would be to order coils handled as replacement parts for the TS-820(S). Table 1 lists the part numbers of those coils that will tune satisfactorily in the circuits shown in Figs. 1 and 2. The 8.830-MHz trap coil has no TS-820(S) replacement part equivalent and must be homemade.

The relay used for final amplifier coil and capacitor switching on 10 MHz is a 12-V, 4pdt unit (one pole is unused) that is identical to that used in the TS-820(S) for antenna switching. If you can't find one

[1]Notes appear on page 42.

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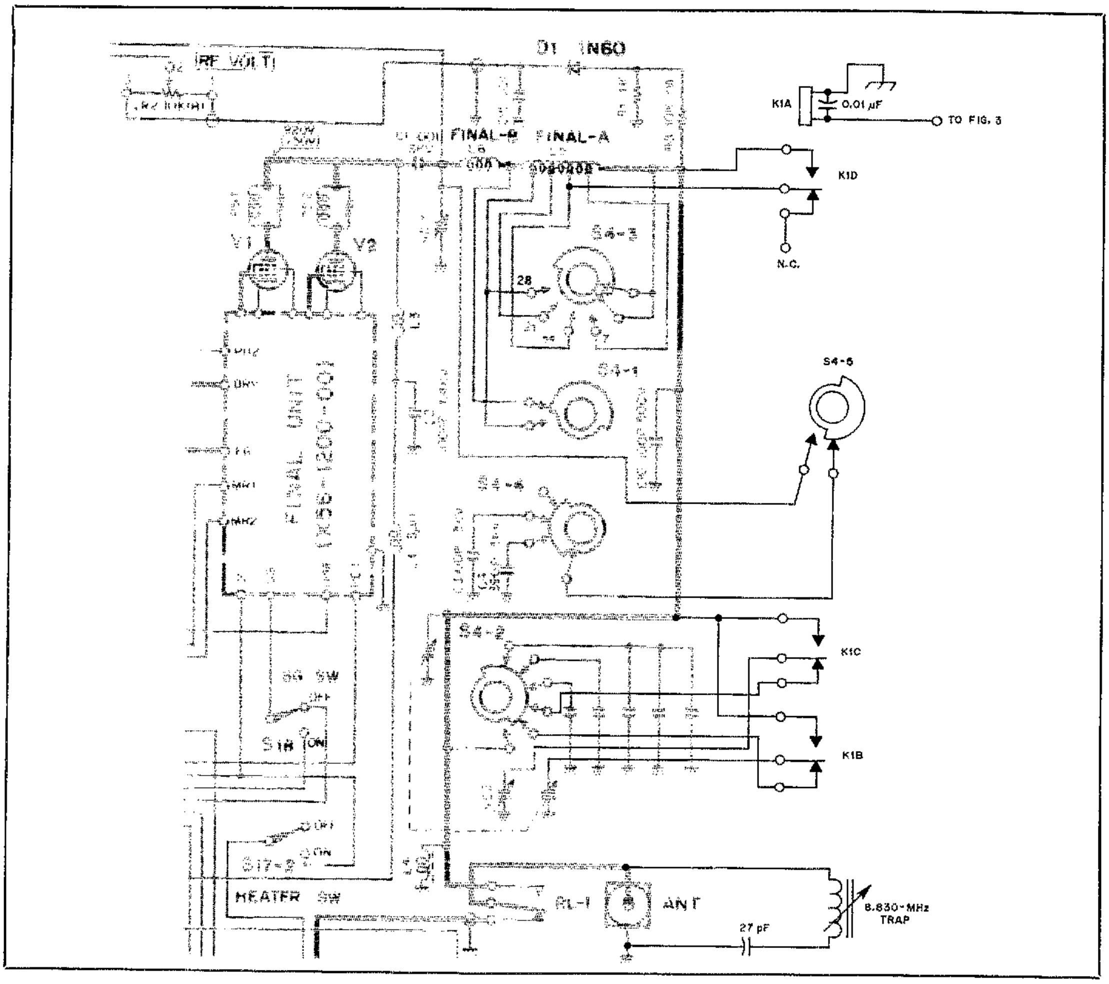


Fig. 1 — Schematic diagram of the TS-820(S) final amplifier after modification. The changes include the addition of a relay (K1) and the use of an existing but previously unused BAND switch wafer. The shaded portion of the diagram is reproduced from the service manual.

locally, the TS-820(S) replacement part given in Table 1 may be used.

A GDO, VTVM and a general-coverage receiver that will tune to 8.830 MHz (and is equipped with an S meter or BFO) are the only test instruments required. If you have a TS-820 without the digital frequency display, a frequency counter will be useful (but not absolutely necessary) for tuning the VCO coils. The TS-820S digital display can be used for this purpose, negating the requirement for a separate frequency counter.

10-MHz Modifications

To permit tuning the final-amplifier circuit to resonance at 10 MHz with the BAND switch in the WWV position, several stumbling blocks must be overcome. Refer to Fig. 1. First, the entire plate coil (used on 1.8 MHz) is in the plate circuit in this position. Second, two plate tuning padding capacitors are inserted in parallel with the variable capacitor (as on 1.8 MHz). One of these is switched out at 3.5 MHz, and the other at 7 MHz and above, by S4-4. Third, a succession of loading-capacitor combinations, both fixed and variable, are switched in and out by means of S4-2 in the 1.8- to 14-MHz positions, with only one section of the variable loading capacitor remaining in the circuit at 21 MHz and above. This one section is also the only loading capacitance in the circuit in the WWV position, and more than this amount is needed at 10 MHz.

There is one unused switch wafer in the final-amplifier compartment. It is not shown in the TS-820(S) export model operating manual schematic diagram.[5] In the domestic (Japanese) model, it is designated as S4-5 and is wired to reduce the final ampifier screen voltage during 28-MHz operation. This keeps the output power at approximately 50 watts as required by Japanese regulations. The switch wafer has only two terminals. Both are open in the WWV position, closed from 1.8 to 21 MHz, and open in the 28- to 29.5-MHz and AUX positions. This switch wafer can be used to remove the tuning capacitor padders from the plate circuit in the WWV position by placing it in series with the wiper of S4-4. Since S4-5 is closed at 1.8 and 3.5 MHz, the tuning capacitor padders are inserted and switched normally for those bands.

The plate tank coil and loading-capacitor switching is done by means of a relay. One relay pole short circuits the coil turns from the 7-MHz tap to the end, as is done by S4-3 for 7-MHz operation. Two remaining relay poles switch two variable loading capacitor sections (normally removed from the circuit by S4-2 in the WWV position) into the output circuit to provide adequate loading capacitance for 10-MHz operation.

The relay is controlled by a simple AND circuit that operates only when the BAND switch is in the WWV position and when a transmit condition is called for. Thus, the relay is activated only during 10-MHz transmission periods, and the final-amplifier plate-circuit configuration is otherwise unchanged.

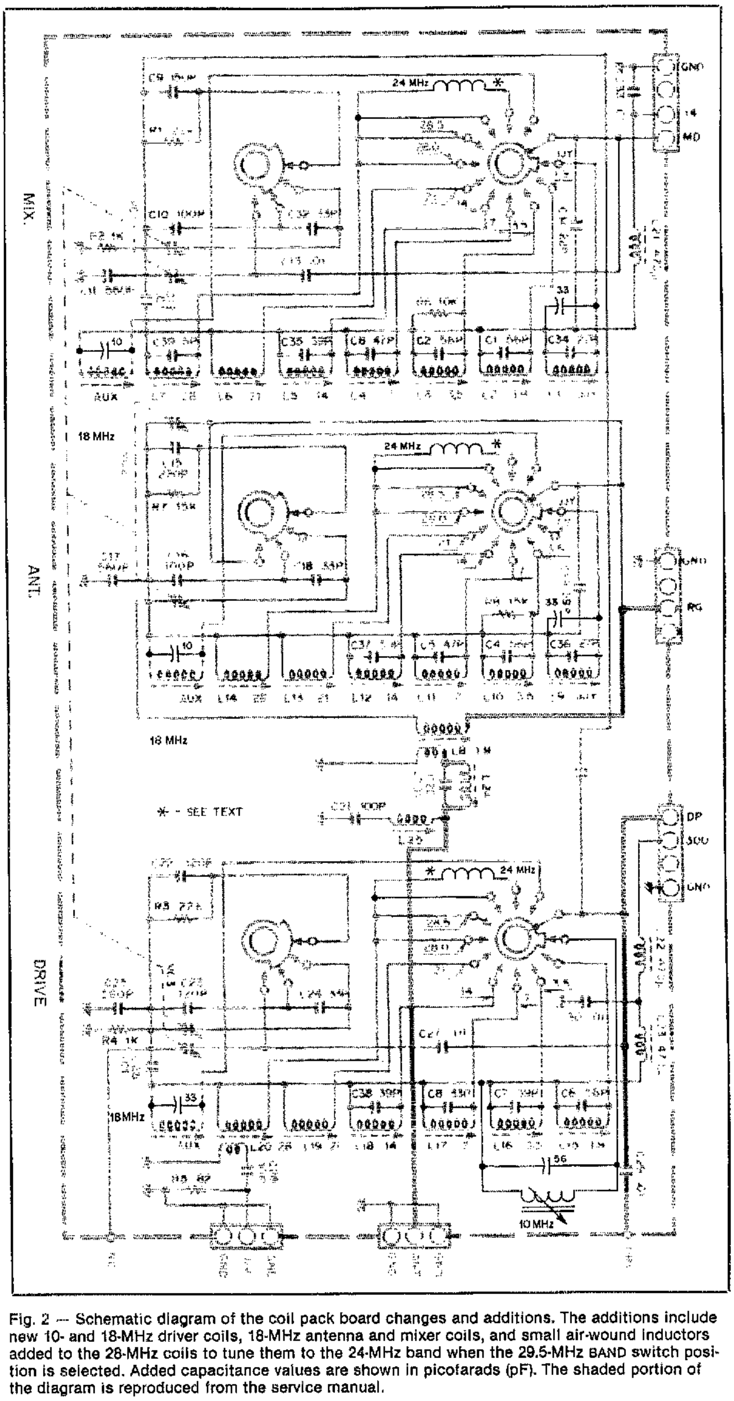


Fig. 2 — Schematic diagram of the coil pack board changes and additions. The additions include new 10- and 18-MHz driver coils, 18-MHz antenna and mixer coils, and small air-wound inductors added to the 28-MHz coils to tune them to the 24-MHz band when the 29.5-MHz BAND switch position is selected. Added capacitance values are shown in picofarads (pF). The shaded portion of the diagram is reproduced from the service manual.

One other item needs consideration. Because 10 MHz is close to the TS-820(S) 8.830-MHz i-f, there is a potential for i-f leakage through the mixer and driver stages to the final amplifier. Listening tests with a general-coverage receiver confirmed that the 8.830-MHz signal was considerably stronger during 10-MHz operation than on the other bands. Use of an antenna-matching network with any 10-MHz antenna will alleviate the problem to some extent. I also added a series-tuned 8.830-MHz trap directly at the antenna output connector in the amplifier compartment. This reduced the 8.830-MHz signal output to a level lower than that which was previously present when the amplifier was operated at 7 MHz.

One anticipated problem turned out to be no problem at all. In the TS-820S counter mixer unit is a 10-MHz oscillator from which counter timing frequencies are derived. I expected the oscillator signal to be rather strong during 10-MHz WWV reception. The oscillator signal is audible when the antenna is disconnected, but it is so weak it does not move the S meter. With the antenna connected, it totally disappears under the WWV signal.

Final Amplifier Modifications

Before starting the modifications, remove the amplifier tubes and the fan assembly. The relay location is shown in the final-amplifier compartment photograph, Fig. 3. The relay is mounted by means of its plastic cover. Remove the cover and drill a hole through the top of the cover large enough to pass a no. 4 self-tapping screw. Make sure the hole is located so that the screw head will clear the operating parts of the relay. Mount the cover on the side of the amplifier compartment in the position shown. Slip the relay into the cover with the coil terminals pointed toward the fan

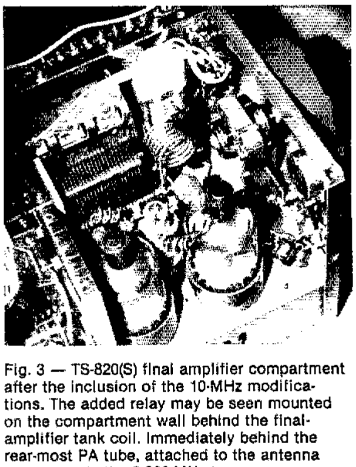


Fig. 3 — TS-820(S) final amplifier compartment after the inclusion of the 10-MHz modifications. The added relay may be seen mounted on the compartment wall behind the final-amplifier tank coil. Immediately behind the rear-most PA tube, attached to the antenna connector, is the 8.830-MHz trap.

Table 1
Parts Requirements

	For Receiving Only	*For Transmitting*
10 MHz	PD board: 1 - 15.5-MHz crystal (X1). VCO board: 1 - 15-pF, 50-V disc ceramic capacitor. Coil pack board: 2 - 33-pF, 50-V disc ceramic capacitors.	Final amplifier compartment: 1 - 4pdt, 12-V dc coil relay, Kenwood part no. S51-4031-05 or equivalent (Radio Shack 275-214 suitable). 1 - 8.830-MHz trap coil, a 5/16-inch winding of no. 32 enam. wire close-wound on a 5/16-inch dia. slug-tuned form. 1 - 27-pF, 500-V disc ceramic capacitor 1 - 0.01-μF, 50-V disc ceramic capacitor. Coil pack board: 1 - driver coil, 1/4-inch long winding of no. 28 enam. wire, close-wound on a 5/16-inch dia. slug-tuned form, or Kenwood part no. L34-0554-05. 1 - 56-pF, 500-V disc ceramic capacitor. Relay control unit: 2 - 2N4400 npn transistors. 1 - 2N2907 pnp transistor. 4 - 4.7 kΩ, 1/8-W resistors. 1 - 560Ω, 1/8-W resistor. 1 - 10Ω, 1/8-W resistor. 1 - 100-PIV, 0.5-A power rectifier.
18 MHz	PD board: 1 - 23.5-MHz crystal (X01) 1 - 3.3-kΩ, 1/8-W resistor (R01). 1 - 5.6-kΩ, 1/8-W resistor (R02). 1 - 150-pF, 50-V disc ceramic capacitor (C01). VCO board: 1 - VCO coil, Kenwood part no. L32-0197-05 or equivalent. (T11). 1 - 5-pF, 50-V disc ceramic capacitor (C01). 1 - 18-pF, 50-V disc ceramic capacitor (C02). 1 - 27-pF, 50-V disc ceramic capacitor (C03). 1 - 15-pF, 50-V disc ceramic capacitor (C04). 1 - 33-pF, 50-V disc ceramic capacitor (C05). Coil pack board: 2 - Antenna/mixer coils, Kenwood part no. L34-0546-15 or equivalent. 2 - 10-pF, 50-V disc ceramic capacitors.	Coil pack board: 1 - driver coil, 8 t no. 28 enam. wire, close-wound on a 5/16-inch dia. slug-tuned form, or Kenwood part no. L34-0555-05. 1 - 33-pF, 500-V disc ceramic capacitor.
24 MHz	PD board: 1 - 30.0-MHz crystal (X10). Coil pack board: 2 - Air-wound, self-supporting coils, 8 t no. 24 enam. wire, 3/16-inch ID.	Coil pack board: 1 - Air-wound, self-supporting coil, 8 t no. 24 enam. wire 3/16-inch ID.

Note: All crystals are available from International Crystal Manufacturing Co., Inc., 10 North Lee St., Oklahoma City, OK 73102. Crystals are third-overtone, GP-1 types in HC-18/U (FM-1) holders. When ordering, simply use catalog number 471165 and specify the frequency desired.

Remove the wire between the tuning capacitor stator and the wiper (common) terminal of BAND switch wafer S4-4 immediately beneath the capacitor. From the underside of the chassis, remove the two screws securing the MARKER unit, and move the MARKER unit aside. This will expose a rectangular cover plate that should be removed to uncover an access hole to the unused switch terminals of wafer S4-5. This wafer is nearest the compartment front. Using no. 18 or 20 hook-up wire, connect an insulated lead from S4-4 (from which the capacitor lead was removed) to either terminal of S4-5. Connect a wire from the other S4-5 terminal to the capacitor stator terminal (from which the lead to S4-4 was removed). Replace the access cover plate and the MARKER unit.

Access to the variable loading capacitor stator terminals is difficult unless the neutralizing capacitor is moved out of the way. Unsolder the lead from the neutralizing capacitor at the driver variable capacitor frame on the coil pack board. Remove the driver tube to get at the two neutralizing-capacitor mounting screws. It may be necessary to remove the screw securing the top bracket of the rf unit and to push the rf unit aside slightly to reach the neutralizing-capacitor mounting screws.

Taking care not to disturb the capacitor setting, bend the remaining lead to move the capacitor sufficiently out of the way to gain access to the stator terminals at the two front sections of the loading capacitor. Disconnect the leads at these two terminals. *Do not* disconnect the lead from the rear section, the one nearest the plate-circuit coils. Splice a few inches of insulated wire to the end of each disconnected lead, slide some insulation tubing over the joints and reroute the leads along the compartment wall and around the plate tank coil to the relay. Be careful not to flex the wafer switch terminals unduly, as they may break off. Solder each lead to a NC relay contact as shown in Fig. 1. Remount the neutralizing capacitor and reconnect the lead to the capacitor frame on the coil pack board. Replace the driver tube and the rf unit bracing screw.

Install insulated leads from the stator terminals at the top of the variable loading capacitor sections, including the rear section, to the appropriate relay contact terminals. Keep these leads close to the capacitor frame to ensure adequate clearance for the compartment cover when it is replaced.

Connect a lead between the bottom end of the plate tank coil to a NO relay contact terminal, and another lead from the operating-arm terminal of that contact set to the 7-MHz coil tap (second from the coil bottom). The NC contact is not used.

Relay operating voltage is supplied from the relay control unit (to be described) through a piece of small-diameter shielded wire or coaxial cable (RG-174/U). The cable should be long enough to reach from the relay to the AND board mounting position when routed through the hole where the other wiring enters the compartment. Bare a short length of the shield in the vicinity of the entrance hole. Attach a ground lead from the shield to the nearby chassis ground pin and connect a 0.01-μF capacitor across the relay coil terminals.

Before it is installed, the 8.830-MHz i-f trap should be adjusted. Connect the coil in parallel with a 27-pF capacitor and adjust the coil for a GDO dip at 8.830 MHz. Listen to the GDO signal on a well-calibrated general-coverage receiver while making the adjustment to be sure the frequency is set accurately. The trap coil is mounted by soldering one active terminal pin and one unused pin directly to the antenna lead at the antenna output connector just inside the amplifier compartment. The coil should be oriented at an angle to permit the slug to be reached by a plastic alignment tool passed through the wide vertical slot in the final amplifier compartment cover. Connect the 27-pF trap capacitor between the other active pin and the nearest available ground point. Replace the tubes, fan and compartment cover.

Relay Control Circuit

The AND circuit used to operate the relay is built on a piece of perf-board (approximately 1-1/2 inches wide and 3/4 inch long) using point-to-point wiring. See Fig. 4. The board is attached to small right-angle brackets secured to the chassis by the same screws that hold the COUNTER assembly unit forward corners. The 9-V input signal is taken from the BAND switch WWV position terminal. Fortunately, this terminal is at the chassis bottom directly behind the front panel and is reached easily. It can be identified by operating the BAND switch and noting which contact is engaged when the switch is moved to the WWV position. Identification can be confirmed with a VTVM by noting that this terminal has 9 V applied to it when, and only when, the switch is in the WWV position.

The RL input can be obtained from the RL line at the external VFO socket, terminal 5. Voltage exists on this line only during transmit. The relay operating voltage is obtained from the 14A line at the input to the filter choke. This choke is accessible when the 5V-AVR unit is moved aside temporarily after removing two mounting screws. The choke input terminal is the one with a wire connected to the 5V-AVR unit.

A relay control unit ground connection is made to a chassis ground tab near the end of the COUNTER assembly. All relay control unit leads should be threaded under some cable harness ties so that they become part of the harness.

PLL System Modifications

The PD and VCO boards make up the PLL assembly, which is attached to the chassis by means of four screws. Two screws are at the outer side of the chassis and two beneath a cable harness at the underside. Remove these four screws, unplug the connectors and remove the unit. Take off the bottom cover plate and remove the four unit VCO mounting studs and two screws. The VCO board can now be unplugged from the through-connector pins which connect it to the PD unit. Remove the PD unit cover and the five screws mounting the board to the shield enclosure and remove the board.

Locate the 20.5-MHz crystal, X1. Using a low-wattage soldering iron, remove the crystal and replace it with a 15.5-MHz crystal. (If you are making the other WARC band modifications simultaneously, refer to the following sections for other changes and additions to the PD unit.)

Identify T1, the 15-MHz oscillator coil, on the VCO board. Next to T1 is a pair of convenient mounting holes for placing a fixed-value capacitor across the coil, although no capacitor is used in the original configuration. Mount a 15-pF capacitor in these holes.

If you are performing the 10-MHz modifications only, replace the PD and VCO boards in their enclosure. Otherwise, refer to the additional VCO board changes. This completes the PLL system modifications for 10-MHz, but the unit should not be replaced until the coil pack board modifications are completed.

Coil Pack Board Modifications

Locate the WWV antenna and mixer stage coils (L9 and L1), and solder a 33-pF capacitor across the active terminals of each. All the capacitors mounted on the board bottom should be located flat against the board to leave clearance for the PLL and CARRIER units.

If you are modifying for transmit, remove the two screws securing the CARRIER unit assembly to the chassis, unplug the connector and set the unit aside. This will expose the entire foil side of the coil pack board. The board need not be removed from the chassis. Refer to Figs. 2, 5 and 6. To mount the new driver-stage coil, drill three small holes (about no. 60 size) 9/64 inch apart, between the AUX coil mounting space and the driver section ground foil border. Clip off entirely the two inactive pins opposite the two active corner pins of the coil and insert the three remaining pins in the new mounting holes, with the coil body outboard of the holes to leave the AUX coil space clear. Clip short the 56-pF capacitor leads, loop them around the two outer pins, snug against the board, and solder the connections. Using a piece of hook-up wire, connect one outer pin to the foil lead which is common to all the driver-stage coils, and connect the other unused WWV switch terminal on the pc board.

If you are performing only the 10-MHz modifications, these units can now be replaced. When replacing the connectors, be careful to line up the plugs to the pins. It is easy to misalign them by one position, and the result might be disastrous when the set is energized.

Tuning and Alignment for 10-MHz Operation

The first tuning and alignment step is adjustment of the JJY/WWV VCO coil (T1 in the PLL assembly) to the proper free-running frequency. This can be done by connecting a frequency counter to TP5 (ground is at TP6) in the PLL assembly. Alternatively, a VTVM connected to TP4 can be used for an approximate adjustment. If you own a TS-820S with the digital frequency display, it can be used as an indicator of accurate coil adjustment.

Rotate the BAND switch to the WWV positon and set the main tuning dial to mid range (250 on the dial scale). If you are using a frequency counter or the TS-820S frequency display, set the PLL unit TUN-NOR switch to the TUN position. There will be no frequency indication if the VCO is oscillating out of the PLL range. Adjust the core of T1 until a frequency reading appears. Continue the adjustment until the frequency-counter reading is as close as possible to 19.080 MHz, or the TS-820S display reads approximately 10.250. Return the switch to the NOR position. The frequency indication should become stable and it should change smoothly as the dial is moved from one end of the range to the other. (Note: If you are using the TS-820S frequency display, the high-order digit (1) that appears correctly at 10.000 MHz and above *will not disappear* when you tune below 10.000 MHz. In other words, 9.9000 MHz will appear as 19.9000 MHz. This is because the initial digit is fixed by the band-switching circuit and is not a function of the frequency being measured.)

To use the voltmeter method of setting the VCO frequency, keep the PLL assembly switch in the NOR position, with other conditions as described earlier. As the core of T1 is turned, there should be a range where a voltage appears at TP4 that varies with the core adjustment. The correct core setting is that at which the voltmeter indicates 3.2 volts. After finding this setting, move the tuning dial back and forth over the entire range. The voltage reading should vary smoothly above and below 3.2 V in proportion to the dial setting.

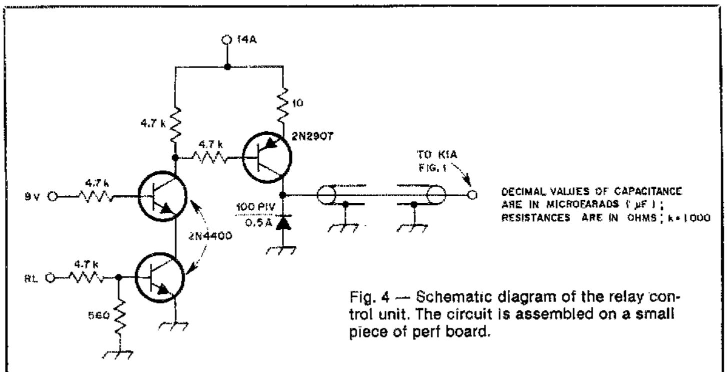


Fig. 4 — Schematic diagram of the relay control unit. The circuit is assembled on a small piece of perf board.

Connect a 50-ohm resistor or dummy load to the antenna output terminal. Turn the FUNCTION switch to the 25-kHz position and tune in the marker signal at 10.250 MHz. Set the DRIVE knob to mid-position (12 o'clock). Adjust the cores of the WWV mixer and antenna coils on the coil pack unit for maximum S-meter indication.

The adjustments are now complete for receiving in the range from 10.000 to 10.500 MHz. Check the receiving capability by connecting an antenna and tuning for signals, including WWV at 10.000 MHz.

To tune the new driver coil, place the rear panel SCREEN-GRID switch in the OFF position and connect a 50-ohm dummy load to the antenna terminal. Set the main tuning dial to 10.125 MHz and peak the marker signal with the DRIVE control. Set the MODE switch to TUN, and the meter switch to ALC. Advance the CARRIER control to about mid-scale. Place the SEND/REC switch in SEND and adjust the driver coil core until an alc meter indication begins to show. As the indication increases, reduce the CARRIER control setting so the meter shows only enough deflection to identify the peak. Without changing the DRIVE control setting, touch up the mixer coil adjustment to make sure its peak and that of the driver coil coincide exactly at 10.125 MHz. Place the screen grid switch in the ON position, switch the meter to RF, and perform the normal carrier, plate loading and tuning adjustments for maximum rf output indication with the MODE switch in the TUN position.

Last, touch up the adjustment of the 8.830-MHz trap in the final amplifier compartment. Connect a clip lead or a short piece of wire (about 3 feet long) to the center ("hot") terminal of the dummy load. Use a similar length of wire as an antenna for a general-coverage receiver tuned to 8.830 MHz. Place the two wires side by side, put the SEND/REC switch in the SEND position and accurately tune in the 8.830-MHz signal. Shorten or separate the wires as necessary to obtain a mid-scale S-meter reading or a moderately strong BFO beat note on the general-coverage receiver, avoiding receiver overload. Adjust the trap coil for minimum received signal. Be careful not to turn the slug too far from the initial setting, because if you tune it too close to 10 MHz, the coil will probably overheat and burn out during transmision. The trap will not completely remove the 8.830-MHz signal emission because there is considerable signal pickup in the receiver caused by radiation from other areas of the transceiver, particularly with the top and bottom covers off.

I did not find it necessary to change the neutralizing adjustment after completing the final-amplifier modifications. If there is any evidence of instability when the amplifier is tuned on any band, follow the operating manual procedure to re-neutralize the final stage.

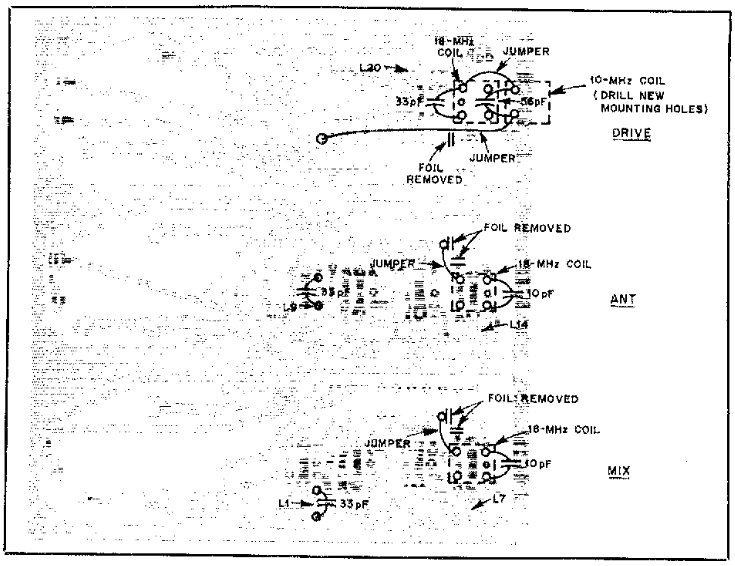


Fig. 5 — Coil pack board circuit pattern showing the physical location of the additional components. This view is from the bottom side of the board as shown in the service manual.

18-MHz Modifications

The 18-MHz band modifications uses the AUX position of the BAND switch. Mounting provisions for all the required new components for auxiliary frequency coverage already exist in the TS-820(S), and the circuit boards are marked to show their locations. The PLL unit must be disassembled as described for the 10-MHz modifications to add the new 23.5-MHz crystal and associated resistors and capacitors on the PD board, and the new VCO coil and capacitors on the VCO board. These components are listed in Table 1 with designations corresponding to their locations as marked on these boards

One additional step is required with a TS-820S before the PLL unit is reassembled. On the VCO board, connect a jumper from the wire-wrap post marked AUX to the post marked B2. This connection establishes a 1 as the first digit of the frequency display and selects a suitable band-pass filter in the counter mixer when the AUX position of the BAND switch is in use.

As described for the 10-MHz modifications, remove the CARRIER unit and install the new driver, antenna and mixer coils, and associated capacitors on the coil pack unit. Then reinstall the PLL and CARRIER units.

No circuit modifications or additions are required in the final amplifier compartment. The 28-MHz section of the plate coil is selected by the BAND switch in the AUX position and is used for transmitting on 18-MHz. With a 50-ohm load, the plate tuning capacitor is about two-thirds meshed at 18-MHz, and the L-C ratio of the plate circuit is rather low. This reduces the rf output slightly because of losses caused by the resulting heavy circulating tank current. Squeezing the turns of the 28-MHz plate coil somewhat closer together will raise the L-C ratio and reduce these losses, slightly increasing the output at 18-MHz. After squeezing the turns closer together, be sure the plate circuit will still tune to the highest transmitting frequency you intend to use in the 28-MHz band.

The alignment procedure is basically the same as described for 10-MHz. For 18-MHz, with the main tuning dial set at 250 and the BAND switch at AUX, the core of T11 should be adjusted to display 27.080 MHz if you are using a separate frequency counter, or 18.250 MHz if you are using the digital display of the TS-820S. When the PLL TUN/NOR switch is returned to NOR, the frequency should stabilize and follow the main tuning dial settings. The alternative VCO adjustment procedure using a voltmeter is exactly as that described for the 10-MHz modification.

To adjust the antenna and mixer coils, place a 50-ohm resistor or dummy load across the antenna output terminal, turn on the 25-kHz marker and tune in the signal at 18.250 MHz. Set the DRIVE control to the 12 o'clock position and adjust each coil core for a peak S-meter reading.

Tune to the 18.125 MHz marker signal and peak the S-meter reading with the DRIVE control. Switch the meter to read alc and place the SCREEN GRID switch on the rear panel to OFF. With the CARRIER control turned a little past half scale and the MODE switch at TUN, put the SEND/REC switch in the SEND position and adjust the AUX driver coil core, backing off on the CARRIER control as an alc indication appears, keeping the alc indication at a level sufficient to determine the peak setting. Return the SCREEN GRID switch to the ON position and proceed with normal tune-up adjustments.

24-MHz Modification

Remove and disassemble the PLL unit as described earlier and replace the 34.5-MHz crystal (X9) on the PD board with a 30.0-MHz crystal. No other changes are required in the PLL system. Before replacing the PLL assembly, remove the CARRIER unit to provide access to the foil side of the coil pack board.

Study Fig. 5 carefully and refer to Fig. 2. In the driver circuit section, locate the four in-line switch terminals that are connected together and to L20, the 28-MHz driver coil, by the foil trace. With a sharp knife or razor blade, score the foil on each side of the small segment between the third and fourth switch terminals (29.0-and 29.5-MHz positions). Remove the foil between the scored lines, isolating the 29.5-MHz terminal.

On the antenna- and mixer-coil sections of the board, isolate the 29.5-MHz switch terminals in a similar fashion. In these cases, two segments of foil must be removed, one just to the left and one just below the 29.5-MHz terminal, as shown in Fig. 5. With a short piece of bare hook-up wire, restore the connections between L14 and the three remaining in-line terminals in the antenna coil section, and between L7 and the identically connected terminals in the mixer-coil section of the board. Replace the CARRIER and PLL assemblies, following the precautions already mentioned involving replacement of the connector plugs.

The added inductance needed to tune each of the three stages to 24.9 MHz is provided by small air-wound, self-supporting coils mounted on the upper side of the board, directly between the 29.0- and 29.5-MHz switch terminals. These are visible in Fig. 6. The coils should be close wound when they are installed and will be adjusted by spreading the turns during the alignment procedure.

To tune the VCO coil (T10, previously used for the 29.5-MHz segment), follow one of the procedures described previously for the 10-MHz band (with the BAND switch set to 29.5 MHz). In this case, the VCO should be set to 33.580 MHz as measured by a frequency counter, or 24.750 MHz as indicated by the TS-820S digital display.

Set the BAND switch to 29.0 MHz and perform the alignment of the antenna, mixer and driver coils for the 28.0-MHz segment as directed in the operating manual. These coils may need a slight touching up because of the coil pack board changes.

With a dummy load connected to the antenna output terminal, set the BAND switch to 29.5 MHz (now 24.5 MHz) and tune to the marker at 24.750 MHz. Set the DRIVE control to 12 o'clock. With a pair of plastic core-alignment tools or other non-metallic devices, slowly pull apart the ends of the added antenna and mixer coils, adjusting the spread as necessary to obtain the peak S-meter reading as each coil is adjusted. Rock the DRIVE control back and forth to determine that the peaks created by the adjustment of each coil coincide at the 12 o'clock position of the control.

Tune to the marker signal at 24.950 MHz and peak the S-meter reading at this frequency with the DRIVE control. Place the SCREEN GRID switch in the OFF position, set the MODE switch to TUN, the meter switch to ALC, the close the SEND switch. Adjust the turns spacing of the added driver stage coil to make the alc indication peak without a change in the DRIVE control setting. Switch back and forth between SEND and REC to be sure the S-meter indication and the alc indications peak at the same setting of the DRIVE control. Adjust only the driver coil as necessary to make them do so. Return the SCREEN GRID switch to the ON position. This completes the 24-MHz alignment.

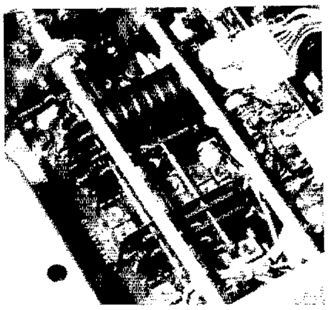


Fig. 6 — A view of the coil pack board with the 10-MHz driver coil in the outboard position beside the 18-MHz driver coil near the front wall of the PA compartment. Also visible are the small air-wound coils added directly to the BAND switch terminals in all three stages.

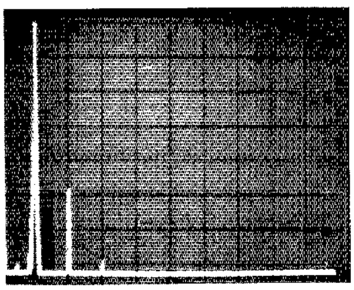


Fig. 7 — Worst-case spectral display of the TS-820S modified for 30-m operation by N1FB. Vertical divisions are each 10 dB; horizontal divisions are each 10 MHz. Output power is 135 W at 10.105 MHz. All spurious emissions are at least 48 dB below peak fundamental output. The modified TS-820S complies with current FCC specifications for spectral purity.

Conclusion

If you've made all the modifications described, your '820 is ready to go on the 10-, 18- and 24-MHz WARC frequencies. Some additional parts and a few well-spent hours have given obsolescence the boot, extended the versatility of your favorite rig, and provided you with the opportunity for more operating fun. Not only that, but think of the things you can buy with the money you saved by not buying that "other" rig! Now let's see — I wonder what a three-element tribander for the WARC bands would cost . . .

[Editor's Note: I installed the 10-MHz transceive modification in my TS-820S a day after the band was opened to U.S. amateurs. It worked the first time!

Because the Kenwood replacement relay was relatively expensive ($17), I sought a substitute and found the $5 Radio Shack relay (275-214) to be suitable.

My junk box produced a Miller 46A014-2 ceramic, slug-tuned form. This I wound with 20 turns of no. 28 enameled wire for use as the driver coil; two unused lugs of the form were cut off.

The AND circuit board is mounted just above the high-voltage filter capacitors by means of a single aluminum strap secured with a COUNTER unit enclosure screw. The 14-A line may be accessed at the foil side of the 5-V AVR unit pc board. You need not remove the unit to reach the filter choke terminal if this is done.

If you're anxious about using the 40-meter tank-coil tap for 30 meters, forget it! Using that tap, my TS-820S supplies 135 watts to a 50-ohm dummy load as measured with a Bird model 43 wattmeter. That's about 20 watts more than is obtained on 40 meters (which says something about the L-C ratio of the PA tank). The PLATE tuning control pointer is positioned at the 21-MHz mark when on 10 MHz. See you on 10 — MHz, that is!] QST

Notes

[1] I used the designations TS-820, TS-820S and TS-820(S) as I think Kenwood intended. That is, the TS-820 is the version without the digital display, the TS-820S is the one with the digital display, and the TS-820(S) refers to either of them.

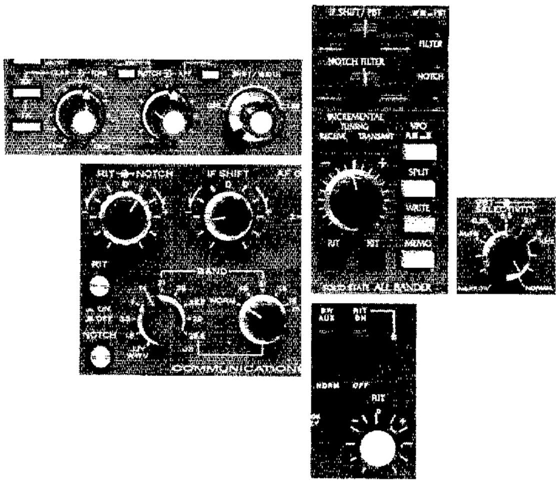
[2] International Crystal Mfg. Co., Inc., 10 North Lee, Oklahoma City, OK 73102.

[3] mm = in. × 25.4; m = ft × 0.3048.

[4] The actual inductance values and number of coil turns used for each coil are not available. Some of the coils I used were removed from surplus pc boards. A GDO should be used to check the resonant frequency of the coil/capacitor combinations. The Kenwood replacement coils will work in the circuits for which they are specified. I took the time to temporarily tune the stock coils in my TS-820S to the appropriate frequencies to make sure!

[5] [However, it is shown on the *service manual* diagram — Ed.]

BEGINNER'S BENCH



Receiver Features that Help You Beat Interference

Been plagued by "rotten QRM" lately? The right receiver features can help you copy the "weak ones."

By George Collins,* KC1V

What's the most important attribute of a modern receiver? Considering the crowded band conditions of today, most operators agree that the ability to copy signals in the presence of interference is at the top of the list. Modern receivers are provided with features to aid in this task — features that were not available a few years ago. Because most amateur stations are now equipped with a transceiver, our discussion of these features will center around the receiver section of a modern transceiver. Operation with a separate receiver differs in some ways, but much of this discussion also applies to "separates."

Crystal Filters

In one way or another, each feature we will be examining relates to *selectivity.* Basically, selectivity is the ability to reject interfering signals. In the past, receivers contained many stages of L-C (coil and capacitor) filtering to achieve high degrees of selectivity. This generally required multiple frequency conversions, which introduced problems with spurious responses and "birdies."

Selectivity in modern receivers is most often obtained through the use of crystal filters. Because these filters can be made for high frequencies (9 MHz is a common crystal-filter frequency), multiple conversions are not necessary. The crystal filter used in a receiver determines the basic selectivity. You can think of the filter as the first line of defense against interference.

Bandwidth

Receiver selectivity is specified in terms of bandwidth. A bandwidth of 2400 Hz means that the receiver will respond to signals spread over a 2400-Hz range. This range is often referred to as the passband of the receiver. Normally, the receiver bandwidth specified is that of the crystal filter being used.

Many (but not all!) transceivers are provided with at least two crystal-filter bandwidths. A filter of approximately 2500-Hz bandwidth is standard for ssb reception. For cw reception, a filter of about 500-Hz bandwidth can be selected (the cw filter is often an option). Some transceivers do not have provisions for a cw crystal filter. Instead, cw selectivity is provided through the use of an audio filter. This is a less-expensive method, and it can be satisfactory for the casual cw operator. The limitations of audio cw filters will be discussed later when we look at audio filtering.

Several newer rigs are equipped with provisions for more than the basic ssb- and cw-bandwidth filters. Some also have a 1.8- or 1.7-kHz filter for ssb reception and a 300- or 250-Hz filter for cw operation. Using these additional filters lets you select the optimum degree of selectivity for a given mode and the current band conditions.

Shape Factor and Ultimate Attenuation

In addition to bandwidth, there are other important filter specifications. One of these is the shape factor. It is the ratio of the filter bandwidth measured at a point of high attenuation on the filter response curve, to the bandwidth measured at a low-attenuation point (Fig. 1). Shape factor is a measure of the steepness of the filter skirts (or sides of the passband). Steeper skirts improve interference rejection and yield a *smaller* shape factor. Normally the −6 dB and the −60 dB response bandwidths are used to determine the shape factor.

Occasionally, however, the bandwidths at other points on the response curve are used for the calculation. Keep this in mind when comparing filter specifications, because the points chosen affect the value of the shape factor for a given filter. For example, if the −30 dB bandwidth is used for one filter and the −60 dB point is used for another, the two shape factors could have the same value, when in fact one filter has much steeper skirts.

Another critical filter parameter is ultimate attenuation. As you move away from the filter center frequency, you reach a point at which the attenuation no longer increases. At that point, the ultimate attenuation of the filter has been reached. The ultimate attenuation depends not only on the filter characteristics, but also on the way the filter is used. Poor circuit layout or improper filter placement can result in signals passing *around* the filter. Under such conditions, even the best filter will appear to have poor ultimate attenuation.

You can easily spot poor ultimate at-

*Basic Radio Editor

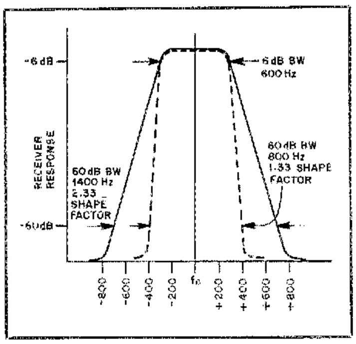


Fig. 1 — The shape factor of a crystal-filter passband is important to good receiver selectivity. A more rectangular passband provides better interference rejection and a smaller shape factor.

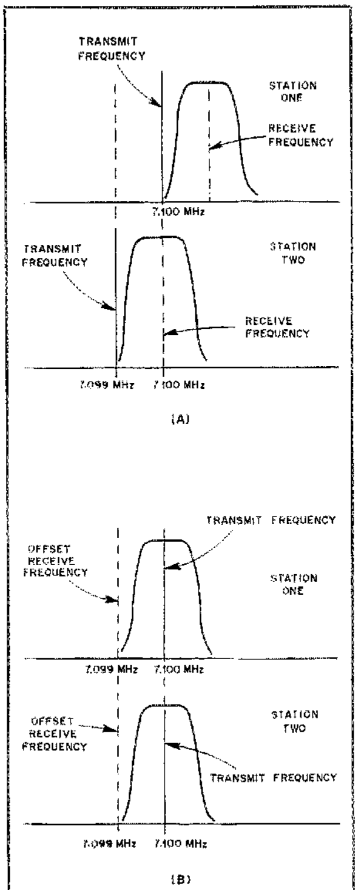


Fig. 2 — Without T-R offset, the transmit and receive frequencies differ by the audio beat note (A). Offset allows the transmitters to operate on the same frequency (B).

tenuation by tuning several kilohertz away from a strong signal. If you can still hear the signal (it will be high-pitched), the receiver suffers from leakage around the filter, or the filter has poor ultimate attenuation.

RIT and XIT

Receiver incremental tuning (RIT) has become an almost universal transceiver feature. While not changing the bandwidth, RIT helps you use the basic receiver selectivity to best advantage. RIT enables you to adjust the receive frequency without changing the transmit frequency. One important use of this feature is correcting for an operator who tunes to a beat-note frequency that differs from the transmit-receive frequency offset of his transceiver. In a transceiver, it is necessary to shift, or offset, the frequency when switching from transmit to receive. The amount of this shift is called the transmit-receive frequency offset, or offset for short. Without offset, tuning your transceiver to another station's cw signal would result in your transmitted signal falling outside the passband of the other station's receiver (Fig. 2A). You could call the other station all day this way and he would never know you were there! The solution to this problem is to offset the receive frequency of both transceivers. Now, by tuning your receiver to the other station's signal, your transmit signal automatically falls in his receiver passband (Fig. 2B).

Great — that fixes everything! Why do I need RIT?

Unfortunately, not all operators tune their receivers for a beat-note frequency equal to the offset frequency. If you adjust your transceiver tuning to obtain a desired beat note, it also causes your transmit frequency to change. Now the other operator must retune. This results in the stations "leap frogging" across the band. Having RIT enables you to avoid this situation. Once you have established contact with a station, don't use your main tuning control to adjust the pitch of the received signal. Instead, make any needed receive frequency adjustments with the RIT control. In a similar manner, RIT can be used to adjust the pitch of an ssb signal to your preference, without causing a change in your transmit frequency.

Some RIT controls will shift the receive frequency as much as 6 kHz. These controls can almost be used as a second VFO (variable-frequency oscillator), enabling you to work stations that are operating split (using different transmit and receive frequencies). DX stations often operate split to avoid having QRM on their transmit frequency cover their signal.

To set up your transceiver to work split, you first determine where the station you wish to contact is listening. Generally, this is easy, as you will hear others calling the DX station on his listening frequency. With the RIT switched off, tune to that frequency. This places your transmitter on the correct frequency. Then turn on the RIT and use it to tune in the DX station. Most transceivers have an RIT switch that you can use to turn the RIT on and off without changing the control setting. This is very useful because it enables you to easily and quickly check your transmit frequency.

A control that is related to RIT is XIT (transmitter incremental tuning). It functions in a manner similar to that of RIT, but shifts the transmit rather than the receive frequency. When a transceiver is equipped with XIT, both the XIT and RIT shifts are set by the same control. Separate on/off switches are used to select the desired function.

A potentiometer is normally used as the frequency-shift control in RIT circuits. The potentiometer is part of a voltage divider used to supply a control voltage to a varactor diode. Varactor diodes act like voltage-controlled capacitors. By placing the varactor in the VFO tank circuit, the operating frequency can be varied by changing the potentiometer setting. To turn the RIT off (and during transmit periods), the variable voltage is removed and a stable, fixed voltage is applied to the diode.

XIT circuits are the same except that the variable voltage is applied during transmit and the fixed voltage is applied during receive. Switching of the variable and fixed voltages is done by means of transistor switches or relays. A simplified RIT/XIT circuit is shown in Fig. 3.

I-F Shift

This useful receiver feature is called *passband tuning* by some manufacturers. To make things even more confusing, the term passband tuning is applied to a completely different control by others. To avoid confusion, we'll use the term *i-f shift* for the control we're discussing now. Under either name, the principle (and result) is the same.

By changing the effective center frequency of the receiver passband, without changing the frequency to which the receiver is tuned, i-f shift allows you to make maximum use of the receiver selectivity. A typical amateur receiver will have a cw bandwidth of 500 or 600 Hz. While this is fairly narrow, it is possible to have two (or more) cw signals fall within the passband.

This situation is shown in Fig. 4A. Here the desired signal has been tuned to the passband center frequency (f_0). The interfering signal, 300 Hz lower in frequency, is also in the passband. By using your i-f shift control, you can "move" the passband center frequency as shown in Fig. 4B. Now the interfering signal is outside the passband, where it can be removed by the filter.

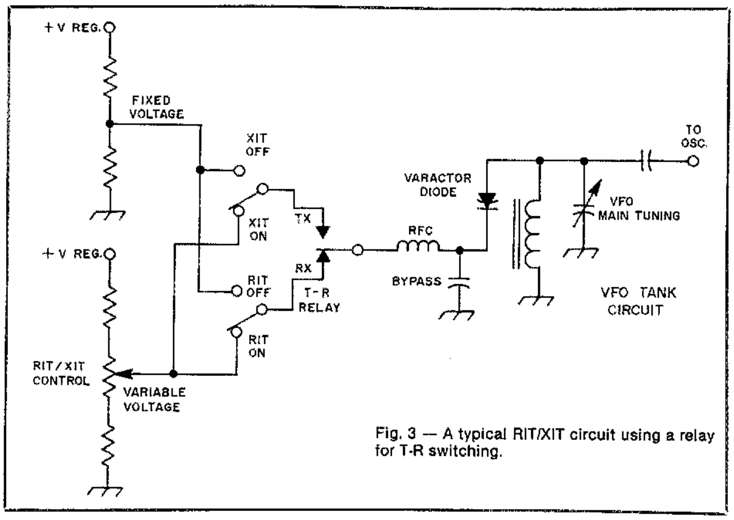


Fig. 3 — A typical RIT/XIT circuit using a relay for T-R switching.

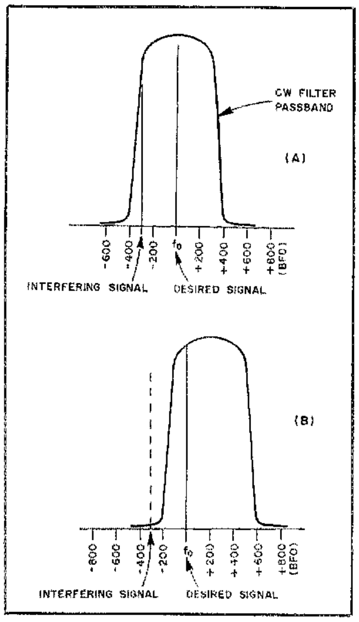


Fig. 4 — Interference from an undesired signal falling in the receiver passband (A) can be reduced by using the i-f shift control (B).

Note that this has been accomplished without adjusting the receiver tuning; thus, the pitch of the desired signal will remain the same. This is because the beat-note pitch, or frequency, is determined by the spacing between the signal frequency (f_o in this example) and the BFO (beat-frequency oscillator) frequency. In Fig. 4, the BFO frequency is 800 Hz away from the signal frequency; thus, the signal will produce an 800-Hz audio tone in the receiver speaker.

Why not just tune the receiver to move the interfering signal out of the passband and listen to a different beat note? That is a possibility, but the problem is that with a transceiver as you tune the receiver you also change the transmit frequency. It is possible to compensate for this by using the RIT control.

If you're receiving an ssb signal, you don't have the option of changing the receiver tuning to move an interfering signal out of the passband. When you're copying ssb signals, the signal frequency and the BFO frequency *must* have the proper relationship (unless you enjoy listening to Donald Duck!).

Before we examine the use of i-f shift while receiving voice signals, let's review some ssb basics. In the process of amplitude modulating a carrier signal, two sidebands are produced. One sideband is above the carrier frequency, and the other is below the carrier. In an ssb transmitter, one of the sidebands is removed, generally by filtering, and the carrier is suppressed. The resulting signal is shown graphically in Fig. 5. When you tune in an ssb signal, two things must be accomplished. First, the received signal must be positioned correctly within the receiver passband. The second requirement is that the signal and BFO frequencies have the correct relationship. If the signal is to be reproduced properly, the BFO frequency and the suppressed carrier frequency (after conversion to the receiver i-f) must be the same (Fig. 6). If they're not, you'll hear the characteristic "Donald Duck" sound of an improperly tuned sideband signal.

Now let's see how i-f shift can be used to reject interference while receiving ssb signals. If you are listening to a signal and another station begins transmitting on the *same* frequency, i-f shift won't save the day (nothing else will, either!). Fortunately, this type of interference is not the most common. More often, the interfering signal will be off to one side or the other, but close enough to overlap the desired signal. Such a situation is shown in Fig. 7A. You can see that a fairly large part of the interfering signal falls within the receiver passband. We can't change the receiver tuning to reduce the interference because that would move the desired signal away from the BFO frequency.

What can we do? Shift the passband! By using the i-f shift we can place the passband edge, or skirt, between the two signals (Fig. 7B). This can reduce the interfering effect of the undesired signal greatly. Shifting the passband will also cause part of the desired signal to be attenuated. The result is a reduction (in this example) of the higher voice frequencies. As long as we don't shift the passband too far, the voice will still be understandable. This puts a limit on how close the interfering signal can be and still be rejected. Naturally, if the interfering signal were lower in frequency than the desired signal, we would shift the passband in the opposite direction. In that case, we would lose some of the lower voice frequencies of the desired signal.

Earlier, we mentioned that the i-f shift and the RIT controls can be used together. The cw operator may find this useful to avoid the monotony of listening to the same beat note for long periods of time. Fig. 8A shows a "normal" tuning condition (no i-f shift or RIT): The signal is tuned to f_o, 800 Hz from the BFO frequency, and the passband is centered over the signal. If you prefer a lower-pitched beat note, the RIT control is used to tune the signal to the lower pitch. Then the i-f shift is used to reposition the passband center at the signal frequency (Fig. 8B). If you are using a fixed-frequency audio filter, you may want to use this technique to obtain peak receiver response at the af filter peak frequency. I-f shift can also be used to enable you to copy the standard RTTY (radioteletype) audio tones of 2295 and 2125 Hz while using a 500-Hz cw filter. Without i-f shift, these tones would be outside the cw passband.

How's It Done?

You can look at i-f shift as though it were changing the center frequency of the receiver i-f filter. Of course, it is not practical to make a variable-frequency crystal filter, so a mixing process is used to achieve the same result. In this process (Fig. 9), two mixers are used. Both receive a local-oscillator signal from the same variable-frequency oscillator. In the first mixer, the signal is converted from the i-f (455 kHz in this example) to the crystal-filter frequency. After passing through the filter, the signal is applied to the

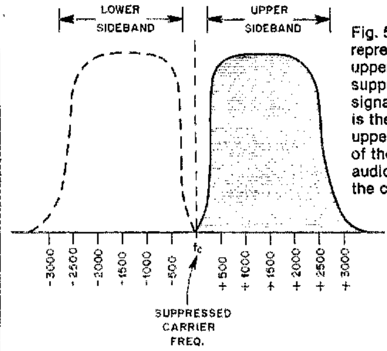


Fig. 5 — A graphic representation of an upper-sideband, suppressed-carrier signal. The sideband is the sum (for an upper-sideband signal) of the modulating audio frequencies and the carrier frequency.

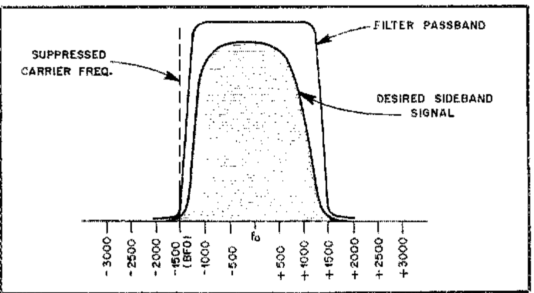


Fig. 6 — When tuned correctly, the sideband is centered within the passband, and the suppressed carrier and the BFO frequencies are aligned.

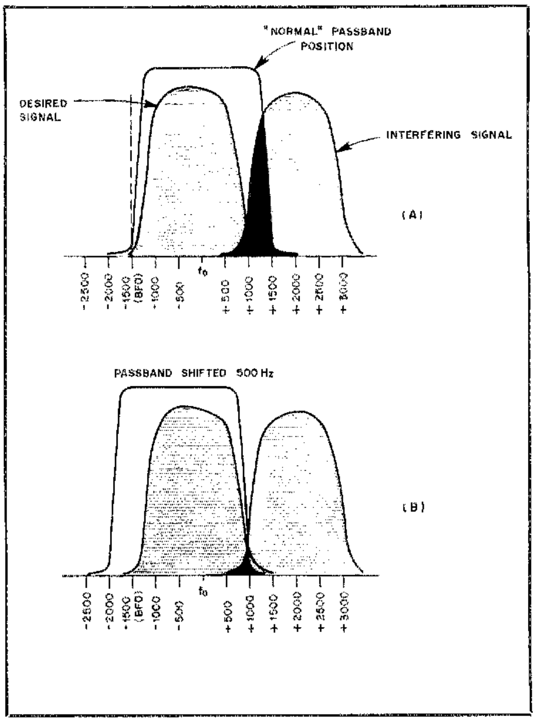


Fig. 7 — By shifting the passband, you can reject an interfering signal without disturbing the carrier and BFO frequency relationship.

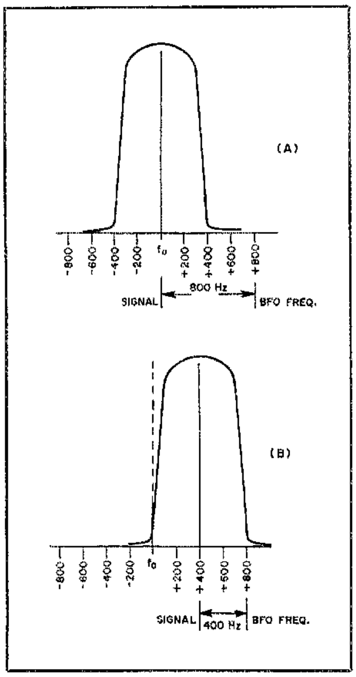


Fig. 8 — Using i-f shift and RIT allows you to select a different beat-note frequency without affecting the transmit frequency.

second mixer. Here it is converted back to the receiver i-f. If we change the oscillator frequency slightly, the input frequency (f_{sig}) that is converted to the filter center frequency also changes. The second mixer simply puts the filtered signal back where we started. The result is that we can select the frequencies that will pass through the filter without changing the receiver tuning.

As useful as i-f shift is, it is not a "QRM cure-all." We mentioned one limitation already: If the interfering signal is too close to the desired signal, i-f shift isn't going to be much help. Should there be *two* interfering signals, one above and one below the desired signal, i-f shift will again be ineffective in producing Q5 copy. In this situation, the only way out is to reduce the *width* of the passband. For example, a 1.8-kHz filter could be used for ssb reception under difficult conditions.

Variable-Bandwidth Tuning

To avoid the expense of equipping a receiver with a multitude of crystal filters, *variable-bandwidth tuning* (VBT) circuits have been developed. These circuits allow you to adjust the receiver bandwidth to fit the operating conditions.

One way of implementing VBT is to place a second filter at the output of the i-f shift circuit (Fig. 10). We now have two passbands through which the signal must

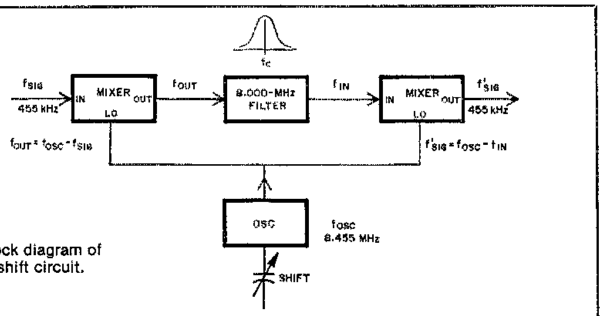


Fig. 9 — Block diagram of a simple i-f shift circuit.

pass. The first passband (that of FL1) can be "shifted" as we saw in our discussion of i-f shift. The second passband (FL2) is fixed. By adjusting the first passband so it is aligned with the second, we obtain the normal (maximum) bandwidth (Fig. 11A). If we shift the first passband, the overall bandwidth is reduced (Fig. 11B).

The maximum bandwidth obtainable with VBT is that of the narrowest filter in the system. If, for example, both filters have 2.7-kHz bandwidths, the maximum bandwidth will also be 2.7 kHz. Typically, the minimum usable bandwidth will be 100 to 300 Hz; it depends on the filters used and the design of the particular VBT circuit. Generally, cw filters will yield narrower minimum bandwidths than ssb filters. VBT is often used to achieve reasonable cw selectivity when a receiver is equipped only with ssb filters. This may prove to be very satisfactory for the operator who works cw only occasionally. When you are comparing features on different transceivers, be sure to keep in mind that VBT is called passband tuning on some rigs. If you have any doubt about the function of a control, check the owner's manual.

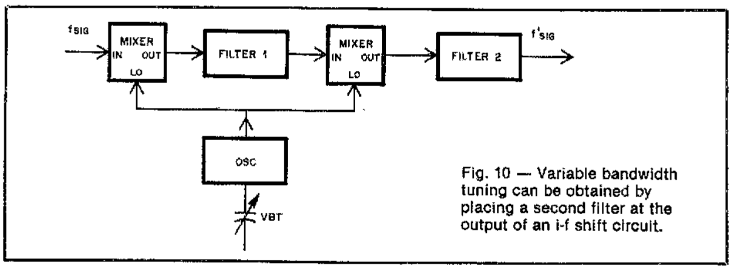


Fig. 10 — Variable bandwidth tuning can be obtained by placing a second filter at the output of an i-f shift circuit.

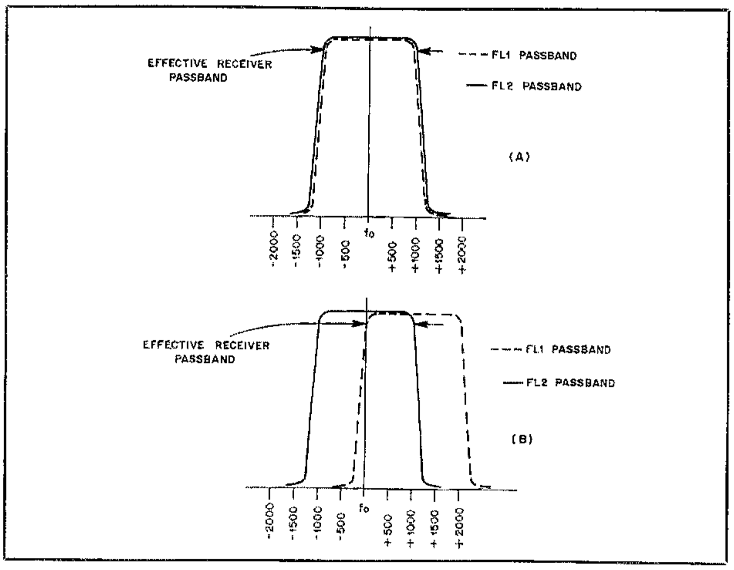


Fig. 11 — Variable bandwidth tuning requires the use of two passbands. By moving the center frequency of one passband, the overall bandwidth is reduced.

Notch Filtering

Another way of reducing interference is to "notch" the offending signal. A notch control, or filter, allows you to reduce the receiver response over a narrow band of frequencies. The notch frequency is adjustable, so you can tune it to the interfering signal. It is important that the notch be narrow; if it is too wide, it will reduce the desired signal amplitude along with that of the interfering signal.

Because the notch must be narrow, these filters are most effective in eliminating interference from signals of narrow bandwidths. A cw signal and the carrier of an a-m voice signal are examples of narrow bandwidth signals that can be handled by notch filtering.

Today, two types of notch filters are found in amateur transceivers. One type is the i-f notch. As the name implies, these filters operate at the receiver i-f. The other type is the audio notch filter. I-f notch filters are most often employed at relatively low frequencies, such as 455 kHz or 50 kHz. This is because it is easier to obtain the high circuit Q required for a narrow notch bandwidth at these frequencies. In single-conversion receivers, with an i-f in the 8- to 10-MHz range, notch filtering is generally provided by means of audio filters.

Both types of filters can be effective, although some audio notch filters do not provide as much attenuation, or "notch depth," as is afforded by a typical i-f notch filter. Typical values of attenuation range from 30 to 60 dB. If a filter has only a 30-dB notch, it will attenuate a 20-dB-over-S9 signal to a level equivalent to about S7. A 60-dB notch would reduce the same signal to approximately S2.

Audio Peak Filters

An af notch filter is one type of audio filter. Another type found in some receivers is the audio "peak filter" (APF). These filters are narrow-bandwidth bandpass filters. Usually, the bandwidth is a few hundred hertz (or less) and the center frequency is adjustable.

An APF can offer a significant improvement in overall selectivity, even in a receiver equipped with a narrow-bandwidth i-f filter. There are some drawbacks, however, when an APF is used in place of an i-f crystal filter for cw reception. Because the APF is located in the receiver audio section, a strong signal *outside* the APF passband can still be amplified to a very high level in the receiver i-f stages. In fact, the signal level can be so great that some receiver stages will be overloaded. Overload can produce distortion and a reduction in receiver sensitivity, making it impossible to copy the desired signal. A crystal i-f filter avoids this problem by removing the undesired signal before it reaches the i-f amplifiers. Even if the receiver is not overloaded, the undesired signal can activate the agc (automatic gain control) circuit, reducing the receiver sensitivity to the point at which a weak signal can't be heard. Sometimes this problem can be overcome by switching to manual gain control.

Closing Comments

Well, these features are all very nice, but I don't need *all* of them, do I? Probably not. Okay; which ones do I need? That's a question each of us must answer for ourselves. Our personal operating styles and preferences will have a lot to do with what we require in a rig. If you're considering purchasing a new rig, think about the features your present equipment doesn't have the next time you operate. Would RIT or a notch filter help you out? If so, look for those features when examining a prospective new rig. If you are looking for your first rig, understanding what the various transceiver controls do (and don't do) should help when you make your choice. QST

Product Review

Conducted By Paul K. Pagel,* N1FB

Yaesu FTV-901R VHF/UHF Transverter

It wasn't too many years ago that a glance behind the equipment installed in a ham shack — *any* ham shack — made one wonder how radio could ever have been called "wireless." A maze of interconnecting cables, external relays, and add-on boxes and gadgets of all descriptions seemed as unavoidable as QRM on 20 meters. Things began to change with the advent of the transceiver, and especially with the appearance of rigs with lots of built-in "goodies," such as the Yaesu FT-902DM. However, that only took care of the hf side of things; for vhf and uhf the rule has remained separate rigs or transverters for each band, with *ad hoc* cabling and switching layouts. It's not uncommon for hams to have different rigs for different *modes* on the same band. OSCAR satellite operation has made things even more complicated.

With the FTV-901R, Yaesu strikes yet another blow for the full-featured, integrated "wireless" station. Designed for use with the FT-ONE, FT-102, FT-101ZD and FT-901/902 series of transceivers, the FTV-901R transverter system extends station capability to the 50-54, 144-148 and 430-440 MHz bands. With the addition of an external hf receiver, a carefully thought-out switching scheme makes it possible to shift from conventional to OSCAR operation at the flick of a wrist. All this is accomplished by one small box and three cables! Even the repeater user isn't forgotten: The transverter automatically offsets the transmit frequency by the proper amount for access to 6- and 2-meter repeaters, assuming that the hf transceiver has fm capability. If you don't need all three bands right away, you can buy the basic unit with the 144-MHz rf module installed and add the other bands later.

With the FTV-901R, ssb and cw operating on vhf/uhf is as easy as on hf. All the features of the basic hf transceiver are retained except transmitter power output, which is rated at 10 watts on each band. An ac-operated power supply is built in; there is no provision for connecting the transverter to a 12-volt dc supply, although all three modules operate at that voltage.

Front-panel features of the unit include a meter which shows relative drive level and relative power output; two rows of red LEDs to indicate which band is in use for both transmit and receive, and whether repeater offset or an external receiver are in use; an RF GAIN control which can be used to reduce the gain of the active mixers in the 50- and 144-MHz receiving converters; and a 12-position band switch for selecting the vhf/uhf band segment or the mode of satellite operation desired. A TUNE control peaks the transmitting sections of the 50- and 144-MHz transverters for operation in the desired portion of each band, but it is not needed on 430 MHz. The control tunes broadly and is simple to adjust.

The circuits used in the individual rf modules are worth mentioning. Input and output are at 28 to 30 MHz. The 6- and 2-meter units are almost identical: in the receiving converter each uses dual-gate FETs for the rf amplifier and mixer, and for the transmitting mixer, an MC1496G IC. Most of the rf selectivity in the receiving converter occurs between the rf amplifer and mixer. As one might expect, the 70-cm module is quite different in design. A common doubly balanced diode mixer is used for both transmit and receive, with two stages of bipolar rf amplification preceding it on receive. Once again, front-end selectivity is provided after, not ahead of, the rf amplifiers. The 430-MHz unit has a Type N output connector instead of the UHF connector used on the lower-frequency units.

The instruction manual for the FTV-901R is excellent, and provides sufficient information to permit most servicing that might be required. Especially useful are the instructions

Yaesu FTV-901R VHF/UHF Transverter

Manufacturer's Claimed Specifications	*Measured in ARRL Lab*
Frequency coverage: 50-54 MHz (optional), 144-148 MHz, 430-440 MHz (optional).	As specified.
Modes of operation: Ssb, cw, a-m, fm.	As specified.
I-f: 28-30 MHz.	As specified.
Transmitter rf power output: 10 W, 50% duty cycle.	As specified.
Receiver spurious responses: Below 1 μV equivalent to antenna input.	See Fig. 1.
Converter Noise Figure (dB): Not specified.	430-440 MHz, 2.9, except for 432-434, 2.8. 144-146 MHz, 4.4; 146-148 MHz, 4.8. 50-54 MHz not measurable because of LO signal feedthrough.
Converter gain (dB): Not specified.	450 MHz, 20; 145 MHz, 20; 53 MHz, 14.
Converter third-order intercept point (dBm); not specified.	450 MHz, − 17.5; 144 MHz, − 11; 50 MHz, − 16.
Weight: 22 lb.†	
Size: 6.2 × 8.3 × 13.8 in. (HWD).	

†kg = lb. × 0.454; mm = inches × 25.4.

*Assistant Technical Editor

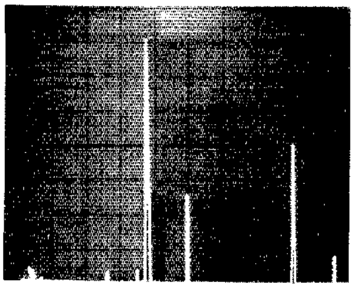


Fig. 1 — Spectral photo of the FTV-901R showing 22-MHz LO energy at the i-f port during reception on 50 MHz. The tall pip to the left of the display is the 22-MHz signal, followed by the 28-MHz signal at the center of the display. Three divisions to the right is the desired 50-MHz signal. Vertical divisions are each 10 dB; horizontal divisions are each 5 MHz.

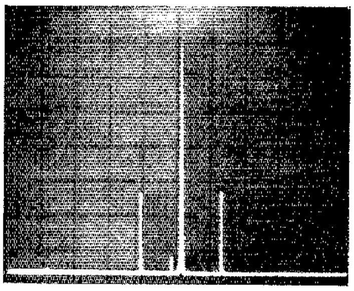


Fig. 2 — Worst-case spectral display of the FTV-901R operating at 50 MHz with a power output of 10 W. Vertical divisions are each 10 dB; horizontal divisions are each 5 MHz.

for adjusting the local oscillators to exact frequency, which is essential if the frequency display of the hf transceiver is to be an accurate indicator of the actual frequency.

One inexplicable lapse in the design of the FTV-901R is the method used to switch between NORMAL and TRANSVERTER operation of the companion hf transceiver. The FTV-901R cuts the *filament voltage* on the final-amplifier tubes in the FT-101 or FT-901/902 when the transverter is switched on. Unfortunately, if the transceiver is already warmed up, the transmitting tubes will retain some emission for as much as a minute afterwards, and *damage to the transverter can result* if the operator attempts to transmit during this period. Some sort of automatic time-delay feature is needed to protect the transverter, or, better yet, a change in the design so a QSY from hf to vhf/uhf can be made instantaneously.

The only other problem encountered in use came to light during noise-figure measurements. Initial tests showed a noise figure of more than 8 dB on 432 MHz — a result that was difficult to understand, given that in on-the-air checks we had heard K8WW in Ohio on a flat band with no preamp and a single Yagi! Some additional poking around revealed that pressure on the Type N antenna connector in turn put pressure on the internal T-R relay in the transverter, resulting in an intermittent contact. The relay design is such that this problem might occur in other units as well, which is worth bearing in mind if 430-MHz performance seems to be below expectations.

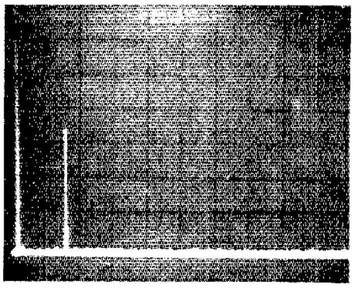


Fig. 3 — Spectral display of the FTV-901R operating at 144 MHz at a power output level of 10 W. Vertical divisions are each 10 dB; horizontal divisions are each 100 MHz.

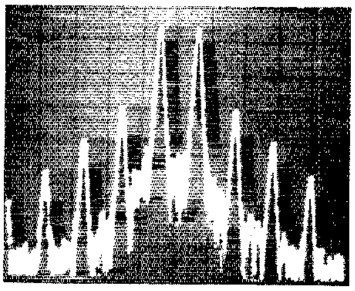


Fig. 4 — Spectral display of the FTV-901R output during transmitter two-tone IMD tests. Third-order products are approximately 30 dB below PEP and fifth-order products are about 39 dB down. Vertical divisions are each 10 dB; horizontal divisions are each 1 kHz. The transverter was being operated at rated input power on the 50-MHz band.

Otherwise, the transverter leaves little to be desired in actual operation. In terms of versatility, the FT-902DM/FTV-901R combination has no peer. While the transverter obviously is not intended for mobile operation, the addition of a 12-volt dc power jack would make the pair suitable for battery-powered vhf/uhf hilltopping. If there was anything else overlooked in the designing of the FTV-901R, I couldn't find it! The FTV-901R is available from Yaesu Electronics Corp., 6851 Walthall Way, Paramount, CA 90723. Price class: $390 (with 2-m module installed); 6-m module $110; 70-cm module $255. — *David Sumner, K1ZZ*

ICOM IC-3AT 220-MHz FM TRANSCEIVER

This rig is truly amazing, because it still works after *I've* used it for nearly three months! That may not seem like much to some of you, but you see, I have this problem. I bought a 2-meter hand-held transceiver, but I had to ship it back to the factory, because it drains the battery while turned off. I have an hf transceiver I'm testing, but it had to go back to the factory for repairs. I assembled a kit recently, but it had to go back to factory because I put a transistor in backwards. All this happened in a two-month period, but the IC-3AT just keeps on working.

First Cousin of the IC-2AT

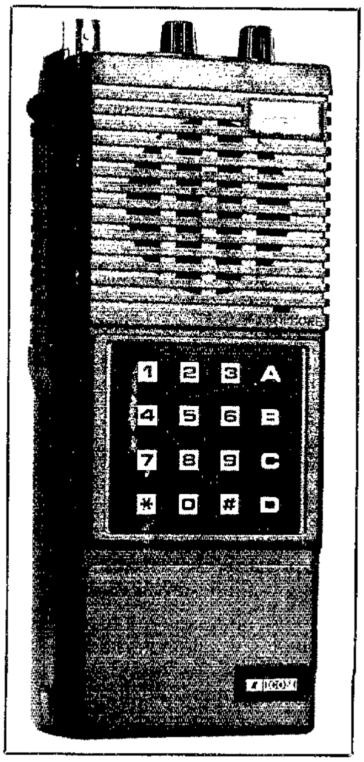
Above and beyond different operating frequencies, there are a few differences between the popular IC-2AT and the IC-3AT.[1] While the '2AT comes with a 12-button standard 2-of-7 tone pad, the '3AT has a 16-button 2-of-8 tone pad. Because of the standardized coherent bandplan for the 220-MHz band, there is no need for a switch to move the operating frequency up 5 kHz. ICOM, therefore, did not add this feature to the '3AT, but did leave the switch in place on the top panel. This switch is quite handy for adding a CTCSS pad or some similar device requiring an external ON/OFF switch. Because there is only one standard repeater offset in the 220-MHz bandplan, ICOM eliminated the third switch on the rear panel, which on the IC-2AT selects the + 600 or − 600 kHz offsets.

Operation of the '3AT is similar to that of the '2AT. The receive frequency is selected using the three thumbwheel switches on the top panel. Using switches on the rear panel, the operator chooses high or low power and simplex or duplex (repeater) modes of operation. The basic package is the same as for the '2AT; accessories are interchangeable, also.

On the Air

Operating the IC-3AT is a pure joy. In several months of steady use I have found nothing to complain about. Never once has there been an indication of malfunction. The receive audio is excellent; signal reports from others indicate transmitted audio is also superb. This is one radio the owner could repair, if there were a problem. The owner's manual contains voltage charts, block diagrams, a detailed schematic and an easy-to-

[1] "Product Review," *QST*, January 1981, p. 38.



ICOM IC-3AT, Serial No. 03376

Manufacturer's Claimed Specifications	*Measured in ARRL Lab*
Frequency range: 220.000 to 224.990 MHz.	As specified.
Mode of operation: Fm.	As specified.
Current drain at 8.4 V	
Transmitting (Hi): 550 mA	
Transmitting (Lo): 220 mA	
Receiving (max. audio): 130 mA	
Receiving (squelched): 20 mA.	
Size (HWD): 4.6 × 2.6 × 1.4 inches (without battery pack)†	
IC-BP3 dimensions: 1.9 × 2.6 × 1.4 in.	
Weight: 16.6 oz. (includes IC-BP3 and flexible antenna).	
Transmitter power output (at 8.4 V):	
HI — 1.5 W; LO — 0.15 W.	HI — 2.2 W; LO — 90 mW.
Spurious radiation: > −60 dB.	As specified. See Fig. 5.
Receiver 1st i-f: 16.9 MHz, 2nd i-f: 455 kHz.	
Receiver sensitivity: Less than 0.5 μV for 20 dB quieting.	0.24 μV for 20 dB of quieting.
Squelch sensitivity: Less than 0.4 μV.	0.26 μV.
Audio power output: More than 300 mW.	300 mW.

†mm = inches × 25.4; g = oz × 28.35.

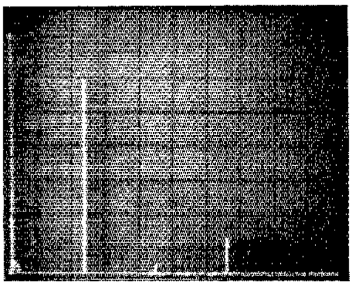


Fig. 5 — Spectral display of the IC-3AT. Vertical divisions are each 10 dB; horizontal divisions are each 100 MHz. Output power is approximately 2.2 W at 224.94 MHz. The fundamental has been reduced in amplitude approximately 20 dB by means of notch cavities; this prevents analyzer overload. All spurious emissions are approximately 66 dB below peak fundamental output. The IC-3AT complies with current FCC specifications for spectral purity.

read board-layout diagram. I think this is the kind of owner's manual that every manufacturer should supply with Amateur Radio equipment — few do, though. Heck, ICOM even provides a schematic for an external microphone — they sell one, too. The IC-3AT is available from: ICOM America, Inc. 2112 116 Ave N.E., Bellevue, WA 98004. Price class $300; IC-3A (without the tone pad) $270. — *Peter O'Dell, KB1N*

ADVANCED ELECTRONIC APPLICATIONS, INC. BT-1P CODE TRAINER

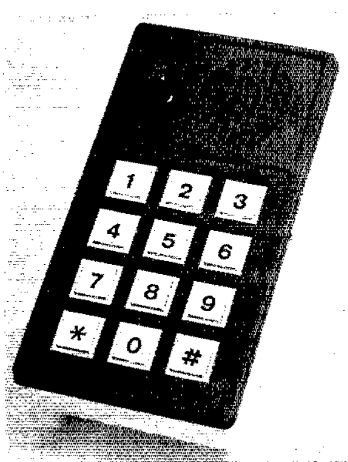
I was excited the day the AEA BT-1P Code Trainer arrived at ARRL Hq. for review! Finally, after many months of trying to increase my code speed, here was my big chance! Then I asked myself, how can this code trainer help? I have tried every conceivable code-learning method, with no luck.

How many of you have experienced the same thing? Well, now you can build up your code speed as I have done. With a previous code speed of 13 wpm, I can now copy 20 wpm with ease. I did it in eight weeks, practicing two hours per week. Even a beginner can learn the code with AEA's suggested two 20-minute sessions a day for four weeks.

Features

The BT-1P and its kin, the BT-1, are handy little gadgets that can easily be carried with you — they're about the size of a pocket transistor radio. The BT-1P is powered by means of internal NiCd batteries and is supplied with a 117-V ac-operated wall charger. The BT-1 operates from an external 12-V dc source or 117-V ac-operated ac/dc adapter. Both are computerized Morse code instructors that send random character groups and offer a total of 61 different characters. They are factory programmed to a minimum speed of 18 wpm and a maximum speed of 99 wpm. With the exception of the means of powering the units, they are identical.

The trainer is programmed by a 12-key pad to send one character continuously or in random groups of two to nine characters, with a pause between groups of 0.1 to 9.9 seconds. After you feel comfortable copying the code in groups, you then have the option of running the characters consecutively. When adding a new character, you have a choice of programming your BT-1P to have all characters equally probable, or have the new character occupy approximately 50% of the time, along with the previous characters.



The BT-1P power switch is on the volume control which may be adjusted to the listening level you prefer. The sidetone pitch can easily be increased or decreased by pressing and releasing a designated key several times until the desired pitch is reached.

I feel the AEA BT-1P Code Trainer would be valuable to instructors of Amateur Radio classes as well as to the individual who wishes to learn the Morse Code or increase his or her code speed. The BT-1P package includes a battery charger and an ear phone. Size (HWD): 2 × 2-1/2 × 4-3/4 inches; color: cream and brown; weight: 10 oz.[2] It is available from Advanced Electronic Applications, Inc., P.O. Box C-2160, Lynnwood, WA 98036. Price class: $80. — *Marian Anderson, WB1FSB*

TELEX HY-GAIN TH7DX BROADBAND SUPER THUNDERBIRD

What has over 700 parts, takes over 10 hours to assemble and attracts birds from all over the East Coast? If your answer is that it is just a TH6DXX with another element added, you are wrong, at least partially. Hy-Gain has added a second driven element, which significantly contributes to the broadband characteristics of the TH7DX. Adding the second driven element forced Hy-Gain to change element spacing and length, which caused some changes in gain and front-to-back ratio. It also forced them to change the matching network. Another improvement in the TH7DX over the TH6 is the switch to stainless-steel hardware.

Construction

The TH7DX comes from the factory in two boxes that are sized to permit UPS shipping — from the appearance and condition of the cartons, I surmised my antenna had arrived via the "Murphy Shipping and Demolition Service." Hy-Gain quickly replaced the few missing parts. A revised manual for the TH7DX suggests the builder read the manual three times before beginning construction. That is sound advice. With over 700 parts, the kit requires considerable care and attention during assembly. The manual could use more diagrams and improved explanations of some procedures.

According to the steps in the manual, the builder should first assemble the boom (four sections) and then construct the elements on the boom. If I were to assemble the antenna by the time AA2Z came to help with the erection, I would have had to leave the partially constructed beam in my yard for a few days. I chose to assemble the elements on my basement floor and then attach them to the boom as one of the last steps. This did not affect the assembly time or quality of work.

Hy-Gain instructs the builder to loosely assemble the beta match/phasing network to the boom before erecting the antenna. The driven elements straddle the mast; therefore, the phasing network is routed past the boom-to-mast clamp. Hy-Gain suggests rotating the network out of the way while attaching the boom to the mast. This is easier to write about than to do — particularly when you are

[2]mm = in. × 25.4; g = oz × 28.35.

clinging to a tower and wrestling with 70 pounds of swaying aluminum tubing.[1] I would like to see Hy-Gain modify the design so the portion of the phasing network located in front of the mast could be taken off easily during installation and removal.

Because of the broadband nature of the TH7DX, Hy-Gain provides only one set of element-length measurements. On each band (10, 15 and 20 meters) the SWR is below 2:1 from one band edge to the other. This eliminates the need to choose one mode of operation over another — it is also extremely helpful for those owning solid-state equipment.

AA2Z and I found it only moderately difficult to install the TH7DX. He placed a pulley on the mast and did the other tower work; I stayed on the ground to provide the "hauling muscle." Total elapsed time from flat on the ground to completed installation atop the 50-ft guyed tower would have been less than 20 minutes, if AA2Z had not experienced difficulty inserting two of the boom-to-mast-clamp bolts. The phasing network was in the way of the bolts. After several minutes he announced that the bolts were in. We then attached the coaxial cable to the BN-86 balun supplied with the kit.

Initial Difficulties

The first SWR curves bore no relationship to those that Hy-Gain suggested would be normal. I was up and down the tower several times in the next few days trying to discover the cause of the problem. Roger Cox at Hy-Gain thought the balun might be bad, so he sent a replacement. While on the tower to change the balun, I noticed the top boom-to-mast-clamp bolts were inserted from the wrong side. The bolt tips extended beyond the nuts, resulting in the bolts being closer to the matching network than the 3/4 inch specified by Hy-Gain. I loosened the matching network, turned bolts around and repositioned the network. Now the SWR checked closer to the specifications, but there still was quite a divergence.

After several phone conversations with Roger, and several futile attempts at curing the problem, I found out what was wrong with the TH7DX — nothing! I was using a moderately priced combination wattmeter and SWR bridge. Before resigning myself to the necessity of removing the TH7DX from the tower to check the problems, I decided to take a Bird Model 43 wattmeter home and check the SWR. With the Bird, the SWR curves were very close to those listed in the manual. I barely restrained myself from using a sledge hammer on my SWR meter!

Structure

The TH7DX is a seven-element beam antenna covering the 10-, 15- and 20-meter bands. Two of the elements are driven on each band and are fed with 50-Ω coaxial cable through a balun (included) and a combination beta match and phasing network. One reflector and two directors are active on 10 meters; on 15 and 20 meters, it is one reflector, two driven elements and one director. Boom length is 24 feet; diameter is 2 inches. The longest element measures 31 feet across, and the antenna has a turning radius of 20 feet. The TH7DX weighs 75 pounds and has a wind surface area of 9.4 square feet. It is rated to survive winds up to 100 mph; at 80 mph it presents 240 pounds of wind loading.

Hy-Gain is producing a conversion kit that will enable the user to convert a TH6 to a TH7. The kit consists of all new stainless-steel hardware and materials for the beta match/phasing network and the second driven element.

Performance

The TH7DX has been a superb performer during the months that I have used it. Good front-to-back and front-to-side ratios have been observed on local and DX signals. Received reports are also excellent. No "cold" numbers are available for these parameters, because ARRL does not have an antenna testing range to accurately measure them. Measurements performed at different station locations with the same antenna would likely produce varying results. My impression is that it performs at least as well as the TH6DXX that it replaced on my tower. In short, I am quite pleased with the performance and the broadbanded characteristics of the antenna.

Price class: TH7DX, $500; conversion kit, $200. Additional information is available from TELEX Hy-Gain, 9600 Aldrich Ave. South, Minneapolis, MN 55420. — *Peter O'Dell, KB1N*

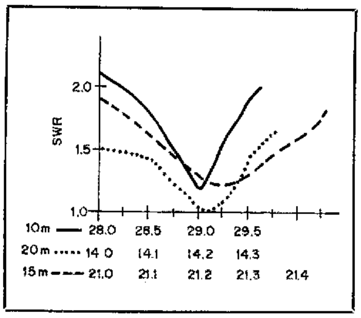


Fig. 6 — SWR curves for the Hy-Gain TH7DX installed at KB1N. The beam is mounted atop a 50-ft guyed tower. Inverted Vs for 40 and 80 meters are hung from the tower approximately 6 feet below the beam. Measurements were made with a Bird model 43 wattmeter.

[1]kg = lb × 0.454

New Products

ENCON INC. PHOTOVOLTAICS

□ The Solarex Corporation manufactured, tested and mounted the solar cells on four panels that power UoSAT-OSCAR 9. Supported by Solarex, ENCON assembles complete photovoltaic power systems for emergency and primary communications applications as well as residential, commercial and Amateur Radio packages. According to Paul DeNapoli, WD8AHO, a major priority had been to develop an emergency power supply package that ENCON is now offering to the Amateur Radio fraternity.

The SX series of photovoltaic panels offered by ENCON utilize semicrystalline silicon solar cells. These cells provide high reliability under shaded conditions because of high heat-dissipation capability and outstanding reverse-bias condition performance. Nickel-solder metallization provides excellent corrosion resistance and high-power performance. A patented tantalum-oxide anti-reflective coating ensures long-term stability and maximum cell efficiency.

Photovoltaic panels have no moving parts and require little maintenance. The Federal government offers a substantial tax credit for solar installations. Some states also offer such tax incentives.

A sample of the Amateur Radio oriented systems available (Systems I, II, III) shows capabilities of providing outputs of from 12-18 Ah to 36-54 Ah per day. Systems include the required number of solar panels, a voltage regulator, storage batteries and hardware kits. In the System III, meters and cables are included; these are optional items with Systems I and II. The price range for these systems runs from approximately $1770 for the System I to $4900 for the System III.

ENCON offers periodic free seminars on solar electricity. If you want to keep up with the latest advances in the photovoltaic industry, you might consider attending one of these seminars.

For further information, contact Paul DeNapoli, ENCON Inc., 27584 Schoolcraft Rd., Livonia, MI 48150. Tel. 313-261-4130. — *Paul K. Pagel, N1FB*

Technical Correspondence

Conducted By
Dennis J. Lusis,* W1LJ

AMPLIFIER OUTPUT CAPACITANCE AND GAIN

☐ Some amateurs who have duplicated *QST* solid-state amplifiers, but tailored them for frequencies other than those specified in the article, have reported such maladies as low output power and instability. The syndrome seems to be most prevalent among those builders who attempt to apply vacuum-tube techniques to semiconductors. They do not understand the characteristics of power transistors — notably the change in C_o (output capacitance) with frequency and the gain increase of the device as the operating frequency is lowered. Both parameters are highly significant in a design.

Fig. 1 illustrates clearly the change in C_o versus operating frequency. The curve is for a Motorola MRF475 power transistor. Other transistors have a different rate of capacitance change per octave, and the C_o may be entirely different from that shown in Fig. 1 for a specific frequency. Therefore, one should consult the manufacturer's data sheet before starting a network design for the amplifier. The C_o must be included in the design equation if the output network is to be correct.

As an example, let's assume that we want to use an MRF475 in the circuit of Fig. 2. Since a Q_L (loaded Q) of 4 is a good "ballpark" value for amateur work, let's use it. Assume in this case we are designing for a cw power output of 5 W and an operating frequency of 3.5 MHz. The collector impedance will be approximately $V_{ce}^2/2P_o$, where V_{ce} is the collector-to-emitter dc voltage and P_o is the output power in watts. Hence, for 13.6 V and a 5-W output the impedance is 18 ohms. Fig. 1 shows the C_o to be approximately 2300 pF (X_c = 19.7 ohms) at 3.5 MHz. These are equivalent parallel values and must be transformed to series equivalents by use of the following equations:

$$R_s = \frac{R_1 X_{co}^2}{R_1^2 + X_{co}^2} = \frac{(18.5)(19.8)^2}{(18.5)^2 + (19.8)^2} = 9.87\ \Omega \quad \text{(Eq. 1)}$$

$$X_s = \frac{R_1^2 X_{co}}{R_1^2 + X_{co}^2} = \frac{(18.5)^2 (19.8)}{(18.5)^2 + (19.8)^2} = 9.23\ \Omega \quad \text{(Eq. 2)}$$

From this we can apply the following equations to obtain correct values for the simple network we have chosen:

$$X_{L1} = Q_L R_s + X_s = 4 \times 9.87 + 9.23 = 48.7\ \Omega \quad \text{(Eq. 3)}$$

The required inductance for L1 (μH) can now be determined from

$$L = X_L / 2\pi f(\text{MHz}) = 2.22\ \mu\text{H} \quad \text{(Eq. 4)}$$

Next, X_{C2} is obtained from $A \times R_L$ where

$$A = \sqrt{\frac{R_1(1 + Q^2)}{R_L} - 1} = 1.535 \quad \text{(Eq. 5)}$$

*Assistant Technical Editor

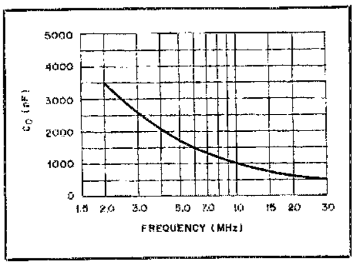


Fig. 1 — Curve showing the change in output capacitance versus operating frequency for an MRF475 power transistor.

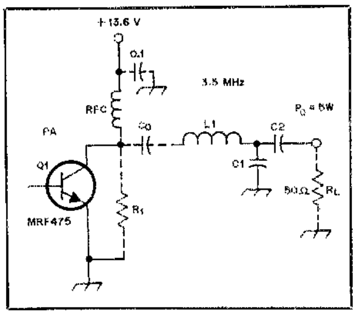


Fig. 2 — Example circuit of an MRF475 rf amplifier that contains a simple collector network to match 50-ohm antennas to the 18-ohm collector impedance. C_o must be included in the network design equations.

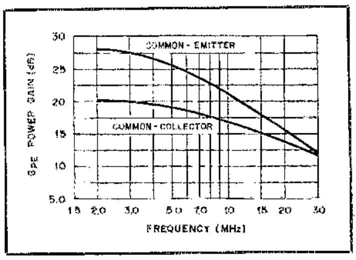


Fig. 3 — Curves showing the increase in transistor gain as the operating frequency is lowered. Common-emitter and common-collector curves are as shown.

Hence, X_{C2} = 1.535 × 50 = 76.74 Ω or 593 pF.

Finally, we can determine the value for C1 from the following:

$$X_{C1} = \frac{B}{Q - A} = 68\ \Omega, \text{ or } 669\text{ pF} \quad \text{(Eq. 6)}$$

where

B equals $R_s(1 + Q^2) = 168$.

The C_o of Q1 has now been absorbed into the network design, and a matched condition will prevail from 18 to 50 ohms. Note that at 28 MHz, for example, the C_o (Fig. 1) will be 500 pF, as compared to 2300 pF at 3.5 MHz. C_o can also present design problems when working with broadband transformers. An improperly designed collector network will result in an impedance mismatch, and, as a consequence, it will be impossible to realize maximum power transfer to the load, R_L. Mismatches can also cause amplifier instability.

Instability can be attributed to another factor — device gain versus operating frequency. This is seen in Fig. 3, where the theoretical gain of an MRF475 is 15 dB greater at 2 MHz than it is at 30 MHz. The gain figure is based on the transistor operating with forward bias in the linear mode, idling at 20 mA — yielding a PEP output of 12 W. This inherent gain increase explains why some solid-state amplifiers may be stable at 15 meters while being difficult to tame at, say, 1.8 MHz. All of this calls for great care in layout and bypassing. Also, less driving power will be needed for a given output power as the operating frequency is lowered. For this reason, an amplifier that covers several amateur bands must have some type of compensating network at the input to provide only enough excitation for the desired output power. Such networks generally consist of a combination of L, C and R components, to cause a gradual loss in driving power as the operating frequency becomes lower. — *Doug DeMaw, W1FB, and Phil Accardi, AJ1N, ARRL Hq.*

THE REALITY OF REFLECTED POWER

☐ From time to time, articles concerning transmission-line techniques appear in various Amateur Radio journals, promoting the erroneous notion that reflected power is fictitious.[1,2] Tragically, many amateurs are misled by authors who are misinformed on the subject, while specialists in wave and field propagation are dismayed at seeing this erroneous concept in print. The truth is, reflected power is as real as the power delivered to the line by the transmitter, or that radiated from the antenna.

While there are many analytical proofs that are unfailing, such proofs are often disbelieved by those who do not understand them. Here's one easy laboratory proof, with physical evidence that can't be denied: Energy reflected by a mismatched line termination can be entirely separated from the forward-traveling wave, and then be dissipated in a temperature-calibrated resistor and accurately measured as

[1] J. Kroenert, "What Your Wattmeter Really Reads," *QST*, Feb. 1981, p. 26.
[2] S. Gibilisco, "How Important Is Low SWR?" *Ham Radio*, Aug. 1981, p. 33.

I^2R heat. If the power associated with the reflected energy were fictitious, no heat would be generated! Within the accuracy of the measuring equipment, the power measured as heat in the resistor equals the reflected energy calculated from the known relationship between the impedances of the mismatched load and the line. The power measured in the load equals the source power minus the reflected power, giving further proof. I have performed these measurements many times.[3]

A semantic problem with the term "power flow" also fuels the erroneous belief that reflected power is fictitious. This brings us to the question: "Does power flow?" To help us understand the answer, let's examine an analogy that involves current.

When we talk about "current flow" we take the meaning for granted. However, does *current* really flow? The basic electricity sections of engineering textbooks (also the ARRL *Handbook*) say that current does not flow — *charge* flows. Current is defined as the quantity of charge flowing past a point per unit time. Once we leave basic electricity and move on to circuit analysis, the term "current flow" is used almost exclusively — and yet we know exactly what is meant.

The same problem applies to the term "power flow." Engineering textbooks define power as the "quantity of energy passing a point per unit time." Thus, power does not flow — *energy* flows. Except when reciting the definition of power, most textbooks on wave propagation use the term "power flow" interchangeably with "energy flow." As with "current flow," we know what is meant because of the common usage, which often overshadows the strict definition. Apparently unaware of the common usage of "power flow," Woods has misinterpreted its meaning in an attempt to show not only that reflected power is fictitious, but also to challenge several other proven textbook principles of transmission-line phenomena.[4] His definitions of power and energy are correct, but the remaining portion of his article is not.

Another reason many amateurs believe reflected waves contain no power is because they misinterpret certain characteristics of standing waves. First, let me state the facts: Because the voltage and current components of the forward waves are always in phase, and those of the reflected waves are always 180° out of phase, the resulting standing-wave plots of voltage and current are always separated on the line by a distance of 90°.[5] The distance between the plots conveys no phase information, but at every point where either the voltage or current standing wave is minimum the phase angle between voltage and current [of the reflected wave — Ed.] is exactly 0°. At points in between, the phase angle may vary between 0° and ±90°, depending on the degree of mismatch, but reaches 90° only when the mismatch results in total reflection of power (an infinite SWR). Now for the classic misconception: The 90° distance between the standing-wave plots is misinterpreted as the phase angle between the voltage and current of the reflected wave. At least two authors have drawn the erroneous conclusion that because cos 90° = 0, reflected waves contain no power![6,7]

I can further prove that reflected power is real by using the Poynting Vector Theorem. Consider an imaginary unit-cube with its sides parallel to the x, y and z axes, containing a quantity of energy, E. If this cube moves forward with a velocity v, the energy will be carried past a given point along the z axis at the rate of Ev joules per second, which is equal to the power in watts per unit area xy. In vector notation, $E_x \times H_y = P_z$ (watts per unit area). E_x and H_y are energy field vectors; E and H represent the magnitudes, and x and y the mutually perpendicular directions of the electric and magnetic fields contained in the cube. The Poynting Vector P_z represents the power passing the point in the z direction, at right angles to the xy plane formed by the E and H field vectors. When a flow of energy encounters a change of impedance (a mismatch), the magnitudes of both the E and H fields change, causing a motor-generator effect on the circuit elements, which converts the mismatch to a virtual source generator (of the reflected wave) in series with a matched load.[8]

According to Lenz's law on counter-electromotive force, in any energy not absorbed by (or transmitted past) the mismatch, the vector direction of either the E or the H field must reverse. (Which field reverses depends on the character of the mismatch.) When either the x or y direction reverses, the Poynting Vector of the unabsorbed energy must also reverse to the −z direction — launching the reflected wave. The energy now flowing in the −z direction is a portion of that flowing earlier in the +z direction, only its magnitude has changed and its direction reversed. An exact analogy is that of light energy reflected from a mirror. Wave energy reflected from a mismatch can be totally separated from the energy traveling in the forward wave. After separation, the Poynting Vector, P_{-z}, of the reflected wave is the cross product of the E and H field vectors of the reflected wave:

$E_{\pm x} \times H_{\pm y} = P_{-z}$ watts per unit area, again proving that reflected power is real power. — *Walter Maxwell, W2DU, ARRL TA, DeLand, Florida*

[3]W. Maxwell, "Another Look At Reflections — Part 3," *QST*, Aug. 1973, p. 36.

[4]H. Woods, "Exploding the Power Myth," *73 Magazine*, Dec. 1976, p. 120.

[5]See note 3.

[6]H. Woods, "Power In Reflected Waves," *Ham Radio*, Oct. 1971, p. 49.

[7]C. Drumeller, "Logic and Reflected Power," *73 Magazine*, June 1973, p. 65.

[8]Everitt, *Communication Engineering*, 3rd ed. (New York: McGraw-Hill, 1956), p. 330.

PACKET RADIO — A SOFTWARE APPROACH

☐ Packet Radio using a 1200-baud rate on 2 meters is thriving in western New York and southern Ontario. Most all the active stations are using the Vancouver Area Digital Communication Group's printed circuit board with the rather expensive ($50+) Intel 8273 SDLC (synchronous data link control) chip.

It is interesting to note that most anything done in firmware (the SDLC 8273 chip) may also be done in software. As such, I have written an assembly-language software program for the TRS-80® Model I microcomputer that simulates the 8273 SDLC chip, when used for synchronous packet-radio communications. A number of two-way contacts have been made with VE3MWM, VE3DSP, VE3DVV and others. The only external hardware required is an EXAR 2211 demodulator, a 1200-Hz/2200-Hz afsk generator and a port-zero decoder for the 48-k memory TRS-80.

No RS-232C interface is required, as all the parallel-to-serial and serial-to-parallel conversion is done with the software. Since the Model I TRS-80 utilizes a 1.774-MHz clock, the receive-mode software must be broken down into three steps to perform functions that the 8273 MSI chip performs simultaneously:

1) A software equivalent of a DPLL (digital phase-lock-loop) automatically "centers" each incoming serial bit for sampling and then stores the sampled value in memory. Interrupts are not utilized.

2) A short subroutine scans the input binary bits to determine the end of the opening flags, skips over any message less than 4 bytes long (illegal), and determines the memory location of the first bit of the address byte of the packet.

3) A subroutine converts the binary packet message to ASCII decimal, stores it in high memory and displays it on video, and finally checks the IBM CRC16 (modulo 1) checksum for an error-free packet.

The above items 2 and 3 take less than 90 milliseconds to execute, so the program can send an acknowledgment if the packet was received without error; if not, it can automatically request retransmission.

I have no intention of suggesting that the popular SDLC chips, such as the Intel 8273, Western Digital 1933 or Zilog 8035, be replaced with software, but I want to illustrate the fact that almost any firmware can be replaced with software if the data rates are relatively slow; i.e., this software program is only good up through 4800 baud. Faster data rates would require a higher-speed clock or, better yet, the previously mentioned dedicated SDLC chips that can handle data rates over 1000 times faster. — *Robert M. Richardson, W4UCH/2, Chautauqua Lake, New York*

Feedback

☐ A correction should be made to the "correct formula" in the December 1982 Product Review of the Instant Software Electronic Breadboard Program. Eq. 2 on page 48 should read

$$R_{EQ} = \frac{R_1 R_2}{R_1 + R_2} \quad \text{(Eq. 2)}$$

☐ In Part 2 of "Antenna Gain Measurements" by Fred Brown, December 1982 *QST*, Q1 and Q2 are indicated as the wrong type of transistors in Fig. 9, page 28. The correct type is 2N5179. Further, Fred's correct ZIP code is 92069.

☐ A sentence in the article by Clifford Appel, "Semi-conductor Testing — in or out of the Circuit," December 1982 *QST*, may be misleading. On page 25, the sentence is located in the center column, midway through the first complete paragraph. It should read: "Note the drop is 0.625 V, *about* what we expected."

☐ In the photo Stray on page 85 in December 1982 *QST*, the two coils in the W4DN receiver are Basketweave coils, not Reinartz coils.

☐ Although the photo accompanying "The Ultimate QSO" (Dec. 1982) may imply otherwise, no one at SRI International is involved in the NASA SETI program, nor are there any plans to use the SRI 150-foot dish in the program.

Hints and Kinks

Conducted By Larry D. Wolfgang,* WA3VIL

MOTOR-DRIVEN ROLLER-INDUCTOR LIMIT SWITCHES

☐ In my October 1982 *QST* article I recommended the use of a slip-clutch drive mechanism for motor-driven roller inductors.[1] Since writing that article, I have tried to use a coil that had too much drag for any of the slip clutches that were available. Others may have run into the same problem when they tried to build the remote mobile-antenna matching circuit.

Fig. 1 shows a simple method of adding limit switches to stop the drive motor when the roller contact comes to the end of the coil. The polarity at pins 1 and 2 on the tuner board is reversed to change the direction of the motor. The limit switch at each end will open the positive lead to the motor. When the polarity is reversed, the diode will bypass the now-negative lead around the open limit switch, allowing the motor to start.

If the direction of travel is such that the limit switch opens the negative lead, then just change the direction of both diodes. I am sure this principle can be applied to other problems as well. — *Don Johnson, W6AAQ, Esparto, California*

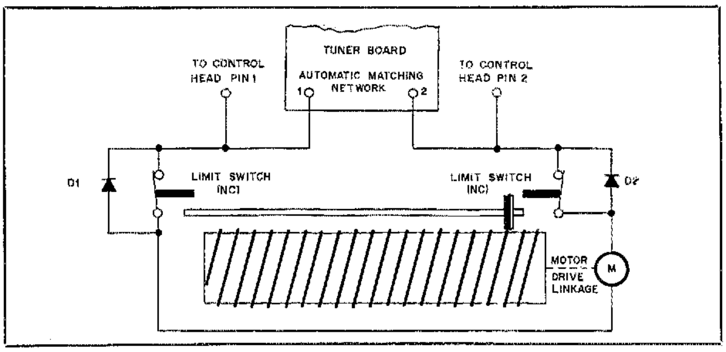


Fig. 1 — Sketch showing a method devised by Don Johnson to use limit switches with a motor-driven roller inductor.

TS-820(S) SIDETONE MONITOR MODIFICATIONS

☐ Once I became aware of the click in the sidetone signal of my Kenwood TS-820S, I couldn't bear to listen to it. Something had to be done! I also wanted to lower the normal sidetone frequency (about 850 Hz in my unit) to approximately 500 Hz. Both of these tasks were accomplished in less than an hour.

The sidetone-circuit components are located on the FIX-VOX (X50-1350-00) board. As shown in Fig. 2, the sidetone oscillator is a phase-shift type, with R48/C35, R45/C36 and R46/C37 determining the oscillator operating frequency. I used 15-kΩ, 1/4-watt, 5% resistors to replace the 10-kΩ units at R45, R46 and R48. The resulting sidetone frequency is now about 545 Hz.

To eliminate the sidetone click I tried different values of capacitance in place of C38. I found that a 0.25-µF unit resulted in a smoother-sounding sidetone, and the click disappeared. Capacitance values up to 0.25 µF have virtually no effect on the rf output waveform. As the capacitance is increased beyond this point, the decay time of the rf-output waveform increases. This causes the on-the-air signal and sidetone signal to sound soft.

If you wish to determine the replacement capacitor value for C38 experimentally before removing the FIX-VOX board, use a probe to make contact with the above-board lead of R47 that is connected to C38. Connect the capacitor(s) between the probe end and chassis ground. Use caution during the procedure, because power is applied to the transceiver, and high-voltage dc and ac line voltage are present at nearby component areas.

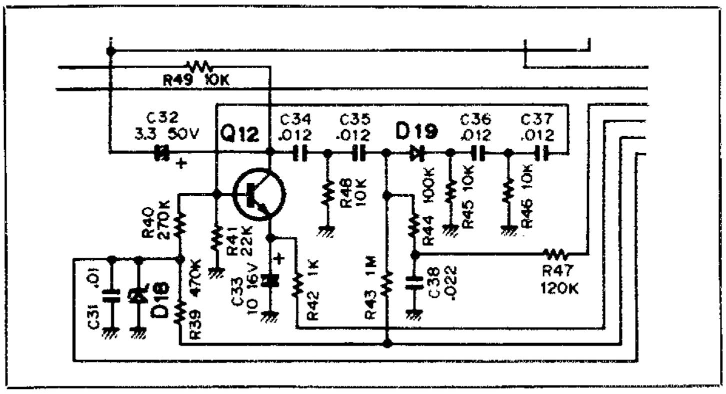


Fig. 2 — Partial schematic diagram of the TS-820(S) FIX-VOX board. This has been reproduced from the TS-820(S) service manual.

The physical size of the replacement capacitor(s) is large compared to the small 0.022-µF disc-ceramic unit originally used for C38. Board space is at a premium. I paralleled two 50-V disc-ceramic capacitors and a 50-V Mylar capacitor to achieve the required value. These capacitors are what I had on hand. The Mylar unit is mounted on the component side of the board in place of C38, leaving lead lengths sufficient to act as tie points for the two disc-ceramic capacitors, which are mounted beneath the board. (A viable substitute may be a Radio Shack 0.22-µF capacitor, part no. 272-1070.)

If you mount the capacitors beneath the board, be aware that there is a limited amount of room between the board and the bottom of the chassis. Also, the capacitor body cannot extend too far beyond the edge of the board, as it might interfere with a compartment shield that is nearby.

The FIX-VOX board is removed by extracting the four mounting screws, removing the on-board connectors and the two-pin connector attached to the shielded wire that connects to the I-F (X48-1150-00) board. Use a 25-W iron and solder wicking to remove the components from the board.

Operators who use audio filters with a fixed center frequency should note that changing the sidetone frequency may not be desirable. For instance, if the filter center frequency is 800 Hz, some attenuation of a 500-Hz sidetone signal will result. — *Paul K. Pagel, N1FB, ARRL Hq.*

HW-101 TROUBLESHOOTING CHART

☐ The troubleshooting chart in my HW-101 manual does not list a check of R316 on the audio circuit board when the sympton is loss of audio in the headphones and speaker. When I experienced this problem with my rig, the S meter indicated lots of signal in the receiver. A continuity check of R316 gave an approximate reading of 22 kΩ (the correct value for that

[1] D. Johnson, "Mobile Antenna Matching — Automatically," *QST*, Oct. 1982, pp. 15-20.

*Assistant Technical Editor

resistor). A voltage check at this point revealed that when power was applied the resistor heated up, the resistance increased and reduced the audio to almost zero. With a lighted magnifying lens I was able to see only a slight discoloration of the resistor. — *Dick Cromer, WD4MZK, Bedford, Virginia*

CHEAP, POLARIZED POWER CONNECTORS

☐ I have been using the top connectors from old 9-V transistor-radio batteries as inexpensive polarized power connectors. After removing the top from a battery, I solder the power leads from the equipment to the connector. Another connector is used on the leads from the power supply. Be careful to maintain the correct polarity. (Positive and negative leads go to the opposite sides of the connector compared to the one on the equipment. See Fig. 3.)

Connecting leads must be of a sufficient wire size to handle the required current. After the wires are soldered to the connectors, I apply a layer of epoxy over the wires to provide insulation and strength. — *Ron Zornow, N9AHU, Dundee, Illinois*

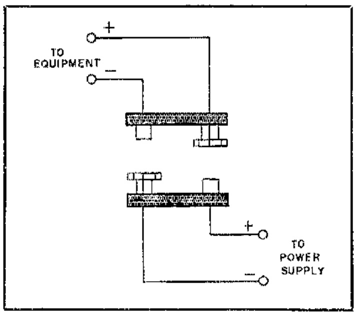


Fig. 3 — Tops from old 9-V transistor-radio batteries can be an inexpensive source of polarized power connectors.

COMPONENT REPLACEMENT ON PC BOARDS

☐ Have you ever had to replace a resistor or diode in the middle of a crowded pc-board circuit? Here is an easy way to get such a nasty job done.

Heat the wire where it comes out of the unit to be removed. The heat will flow down the wire and melt the solder. Lift that end of the component with a pliers, or use a knife blade to pry it up. Repeat the process on the other lead. A small drill bit will clear the solder from the hole. Then a new part can be soldered into the circuit. Don't use this method with a double-sided board that has plated-through holes. In this case a piece of dirty copper wire can be heated and pushed through the solder to avoid damaging the plating. This is much easier than trying to find the right place to unsolder from the bottom of the board. — *Lew Stapp, WØPHY, Hays, Kansas*

MEASURING ALTERNATING CURRENT: AN UPDATE

☐ I believe I have a better method for measuring alternating current with a VOM than the one described by Edwin Walker, WA4OFS, in the June 1982 Hints and Kinks column. My method, shown in Fig. 4, uses diodes in place of the resistor. The diode method introduces only about a 0.6-V drop in applied voltage, regardless of load power. The resistor method produces a voltage drop that varies directly with load power. This drop amounts to 7.1 V, or almost 6% of the applied voltage at the power limit of the resistor. The diode drop would amount to only 0.5% of the applied voltage.

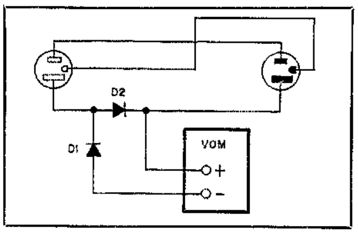


Fig. 4 — Diagram showing the method used by K2SE to measure alternating current with a VOM. Be sure the diodes and meter connect to the neutral side of the ac line for safety.

A second advantage of the diode method is that the power dissipation varies linearly with load power, but the resistor-circuit dissipation varies with the square of the load power. In a circuit drawing 7 A, the resistor will dissipate nearly 50 watts, but the diodes would only dissipate 1.9 watts each. The accuracy of the WA4OFS method depends on the tolerance of the resistance, but the accuracy of my method does not depend on the diode parameters.

The maximum load that can be handled by this method calculated to be 2.22 times the line voltage times the maximum dc current capability of the meter. My Radio Shack VOM can handle 2665 W on the 10-A scale. To find the required diode current rating divide the load power by 85. If you are measuring a 500-W load (even though the meter can measure a lot more) you will need 6-A diodes. The diode PIV rating need not be considered because the peak inverse voltage across either diode will be the forward voltage drop of the other diode.

One drawback of my method is that the meter does not read rms current directly. The meter reading must be multiplied by 2.22 to obtain the rms current in the circuit. For those who are curious where this "magic" number comes from, I will show a simple derivation. I'll skip the details of the integral calculus that derives equations 1 and 2, however.

$$I_{rms} = \frac{I_{pk}}{\sqrt{2}} \qquad \text{(Eq. 1)}$$

$$I_{pk} = \pi \times I_{dc} \qquad \text{(Eq. 2)}$$

Combining these equations we get

$$I_{rms} = \frac{\pi}{\sqrt{2}} \times I_{dc} = 2.22 \times I_{dc} \qquad \text{(Eq. 3)}$$

Please note that if this circuit is wired into an electrical box as suggested with the other circuit, you should not leave the load connected without the meter, since this results in half-wave rectified dc being supplied to the device under test! If the leads that should go to the meter are left dangling, one of them will be "hot" and would present a safety hazard. Plug the leads into your meter before plugging in the power cord. — *Edwin Solov, K2SE, Wayne, New Jersey*

[Editor's Note: George Woodward, W1RN, suggests that a full-wave bridge rectifier could be used, with the meter connected between the dc terminals of the bridge. In this configuration, the load would be disconnected from the line if the meter was not connected. The dc meter reading would have to be multiplied by 1.11 to obtain ac rms current. This would result in a 1.2-V drop in applied voltage instead of 0.6 V as with Solov's circuit. The diodes in this type of circuit should have a PIV rating of 400 V if the circuit is to be plugged into the power outlet with the meter disconnected.]

THE TRS-80® COLOR COMPUTER AND THE STATE-OF-THE-ART TU

☐ The State-of-the-Art Terminal unit[2,3] works well with the mechanical RTTY machines, and with only minor modification will also work with the TRS-80 Color Computer. For an RTTY beginner this TU seems to be an economical approach.

The modification to make the TU transmit with the Color Computer is the easiest to accomplish. Fig. 5 shows the circuit changes required. Remove the wire from pin 9 on U2, the XR-2206 tone generator. Add an spdt switch

[2]M. Di Julio, "State-of-the-Art Terminal Unit for RTTY," *QST*, Dec. 1980, pp. 20-22.
[3]R. Witmer, "Auto-start and Anti-space for the State-of-the-Art TU," *QST*, Nov. 1981, pp. 28-30.

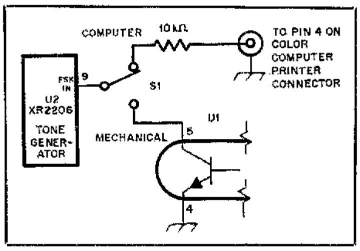


Fig. 5 — Schematic diagram of the transmit modifications for using the State-of-the-Art TU with a TRS-80 Color Computer. Additions are shown with heavier lines.

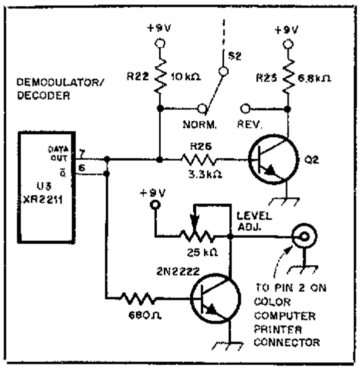
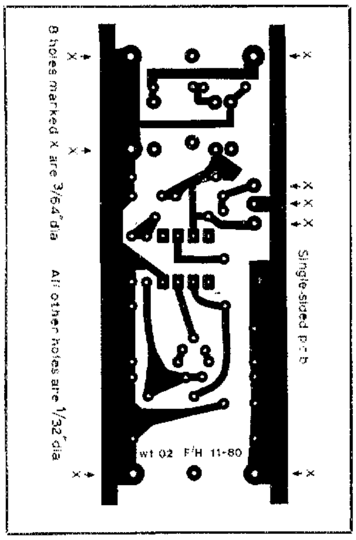


Fig. 6 — Schematic diagram of the changes to the State-of-the-Art TU for receiving with the TRS-80 Color Computer. Additions are shown with heavier lines.

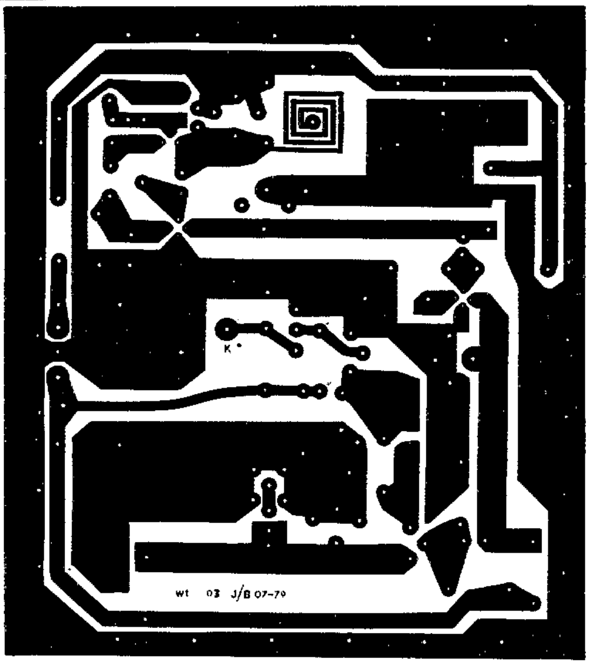
between this pin and pin 5 of U1. This gives you the option of using the TU with your computer or a mechanical device. The other switch pole connects to a 10-kΩ resistor in series with a lead to pin 4 of the printer connector on the back of the computer.

The receive modification is a bit more involved, and requires some planning. There are three components to be added to the existing board. If you are using one of the boards sold by the author of the original article, then you will have to solder these leads to the components on the top of the board. Find the proper locations for the added components shown in Fig. 6, and solder them into the circuit. A 680-Ω resistor connects between the base of a 2N2222 transistor and pins 6 and 7 of U3, the XR-2211 demodulator/decoder. The transistor emitter connects to ground, and the collector goes to the center terminal of a 25-kΩ, 10-turn, pc-mount potentiometer. One side of the potentiometer connects to the 9-V supply, and a wire is connected from the 2N2222 collector to pin 2 of the printer connector on the computer. I mounted phono jacks on the back of the TU for the connections to my computer.

Final adjustments are made after the station equipment is turned on and the RTTY program is loaded into the computer. Turn on the TU and tune the receiver to a clear frequency. Connect a VOM between the 2N2222 collector and ground. Adjust the 25-kΩ potentiometer for a reading of 2.6 V on the VOM. Now tune the receiver to an RTTY station. You should see the monitor begin to print. If not, check the keyboard to be sure the computer is in the receive mode, check all connections and the position of the new switch. With the receiver properly tuned on an RTTY signal adjust the potentiometer again until the printing starts. This completes the circuit adjustment. Here's looking for you on 20-meter RTTY. — *Bill Sepulveda, N5EIF, San Antonio, Texas*

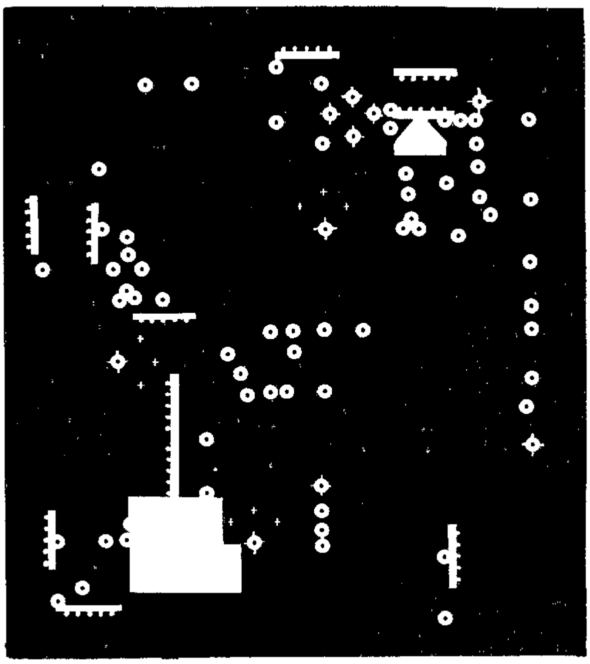


Circuit-board etching pattern for the uhf signal source voltage regulator (see the parts layout of Fig. 10, p. 32 of this issue). Black represents copper. The pattern is shown actual size from the foil side of the circuit board.



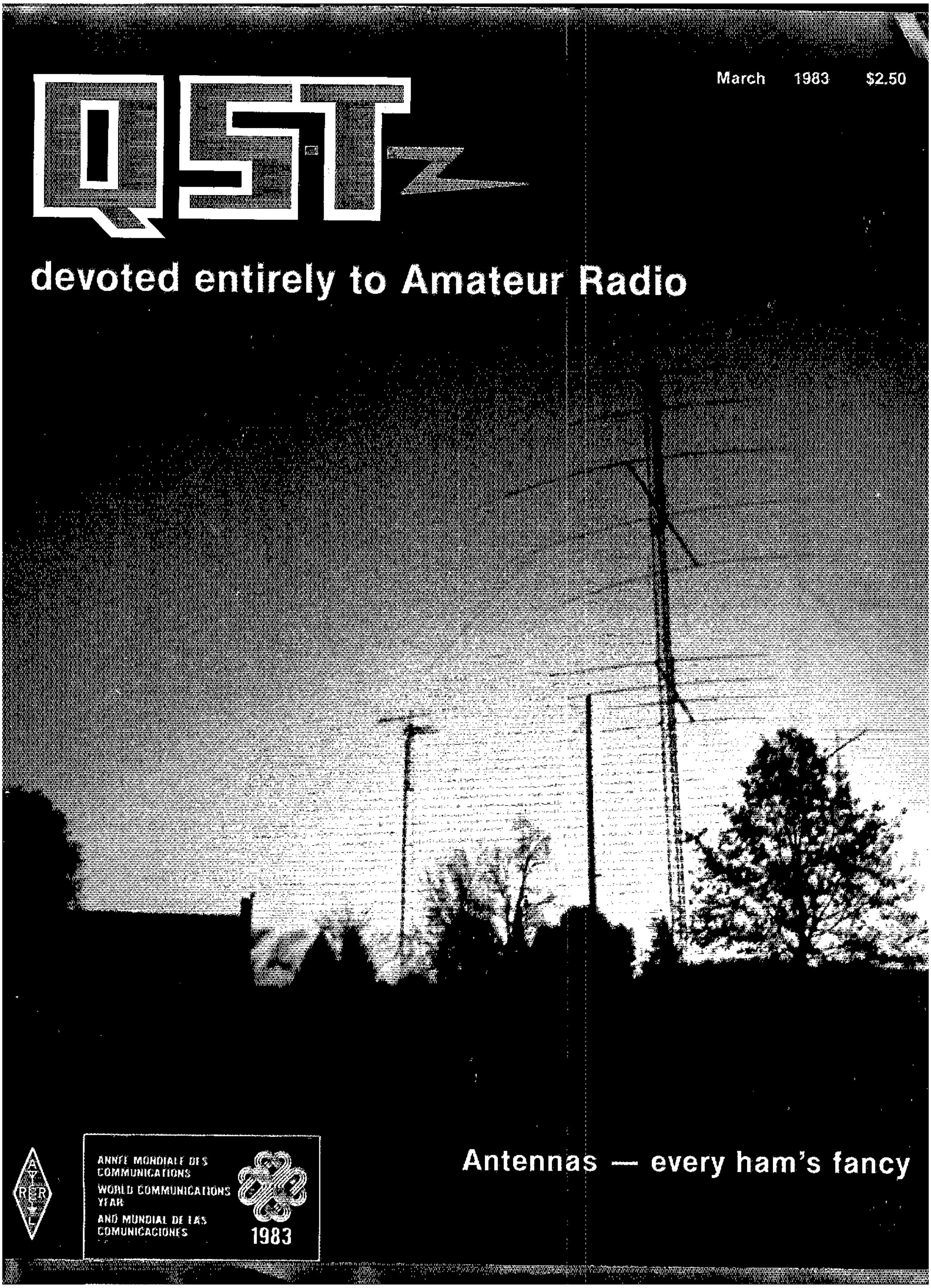
(A)

11 rows of white dots, 9 holes ■ and 10 holes ◆ are drilled 3/64" dia



(B)

Circuit-board etching patterns for the uhf signal source (see parts layout of Fig. 2, p. 29 of this issue). A is the pattern for the circuit side of the board. B is the pattern for the ground plane (component) side of the board. The patterns are shown actual size. Black represents copper.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CONTENTS

OUR COVER

Up early one morning, Associate Technical Editor K1TD came across this lovely sunrise over the W1AW antenna farm. Your antennas may or may not be as photogenic, but they're certainly worth paying attention to, especially at this time of year. The articles beginning on pages 16, 33, 36 and 38 provide some timely reading.

TECHNICAL

BEGINNER'S BENCH

NEWS AND FEATURES

OPERATING

DEPARTMENTS

Make Mine Modular: Easy-to-Build Receiving Converter and Test Equipment for 435 MHz

The amateur experimenter is not dead! Get in on the uhf excitement. Easily found parts yield high performance on a low budget.

By John C. Reed,* W6IOJ

A lot of interesting activities take place in the 70-cm band. I am particularly interested in the amateur satellite program. To operate OSCAR 8, Mode J, I needed a receiving converter for 435 MHz. I decided to build one.

There are two potentially discouraging hurdles to get over in a project of this kind. The first is where to buy the parts. I was able to put together a relatively high-performance converter using parts purchased mostly from Radio Shack. A mail order parts emporium carried the few remaining items that were not in my "junk box."[1]

The second potential discouragement is where to find the uhf test equipment for troubleshooting and alignment. Highly sophisticated equipment is not needed; you can construct the test equipment you need. This article will show you how.

I built my first 435-MHz converter on a 6-inch-square pc board.[2] It worked, but there were a number of compromising limitations. I also found it difficult to make changes. What to do? Start over and forget about the shape and size — make it easy to work with. The final version had three separate assemblies: oscillator, mixer and preamplifier. This new arrangement resulted in simple assemblies that are stable without shields — an important convenience during checkout and alignment.

Oscillator

This module is mounted on a 5 × 4-1/4 inch double-sided pc board (Figs. 1 and 2). The layout deliberately places the input and output circuits for each stage on opposite sides of the pc board. Isolation provided by this method allows stability without additional shielding. I use no-etch circuit boards. Construction is easy and the results are good.

I chose a 29.5-MHz-and-up converter i-f. Oscillator output at 405.5 MHz is the eighth harmonic of the 50.7-MHz crystal. Three criteria guided the design process. I wanted a simple circuit that had plenty of spurious-free output.

Circuit simplicity dictated the use of an FET for Q1. The piston-trimmer and feedthrough capacitors were purchased from Meshna. The 24-pF capacitor in the source lead of Q1 is actually two 47-pF, disc-ceramic capacitors in series. Being satisfied with their performance, I chose to use disc-ceramic capacitors throughout the project.

A simple pi network provides coupling between Q1 and Q2. The 100-pF capacitor should be mounted as close as possible to the base of Q2 to prevent spurious oscillations. Reasonable Q is maintained in the 202.7-MHz tuned circuit through the use of link coupling. Spurious signals at this point are at least 40 dB below the desired signal.

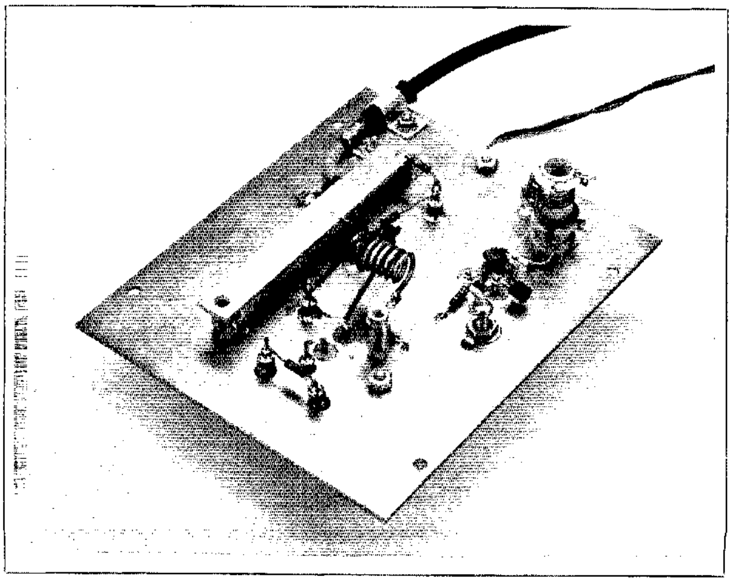


Fig. 1 — Photograph of the local-oscillator module.

[1]Notes appear on page 15.

*770 La Buena Tierra, Santa Barbara, CA 93111

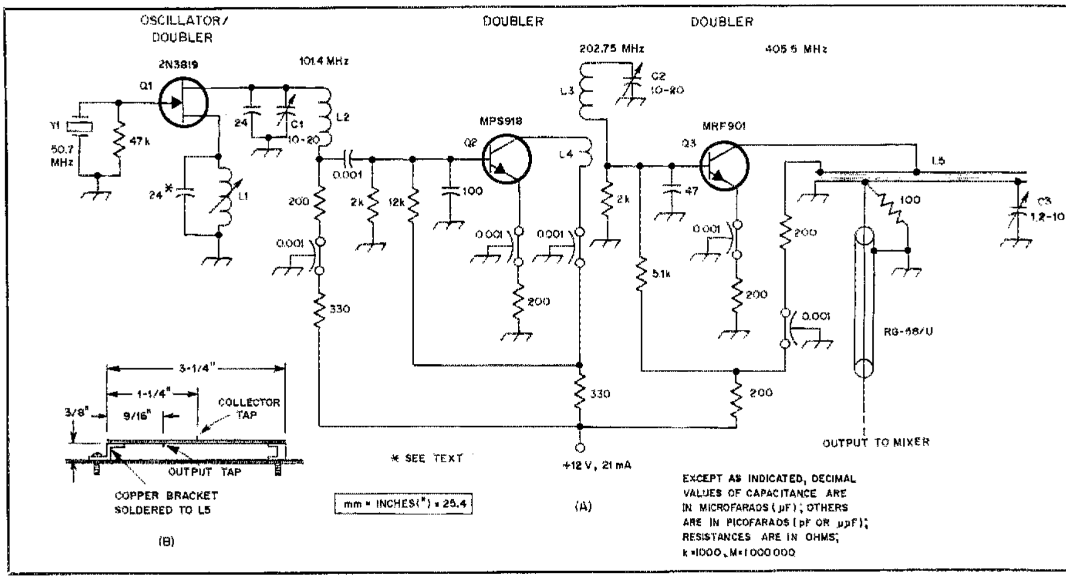


Fig. 2 — Schematic diagram of the local oscillator is shown at A; a detail drawing of L5 is at B. All resistors are 1/4 W; fixed-value capacitors are disc ceramic. Feed-through capacitors are from Meshna, part no. H-30.

C1, C2 — 20-pF piston trimmer (Meshna SP-109).
C3 — 10-pF piston trimmer (Meshna SP-109A).
L1 — 0.327 to 0.587 μH (Miller 40A477CB1).
L2 — 5 turns of no. 14 wire, 3/4 in. long and 1/4 in. in diameter.
L3 — 7 turns of no. 14 wire, 3/4 in. long and 1/4 in. in diameter.
L4 — 2 turns of no. 22 hook-up wire, 1/4 in. in diameter.
L5 — 3-1/4 × 3/8 × 1/16-in. double-sided glass-epoxy pc board. Q3 collector tap 1-1/4 in. from cold end. Output tap 9/16 in. from cold end.
Q1 — Radio Shack 276-2035.
Q2 — Radio Shack 276-2011.
Q3 — Radio Shack 276-4055.
Y1 — Third-overtone crystal (International Crystal Mfg. Co., no. 031081).

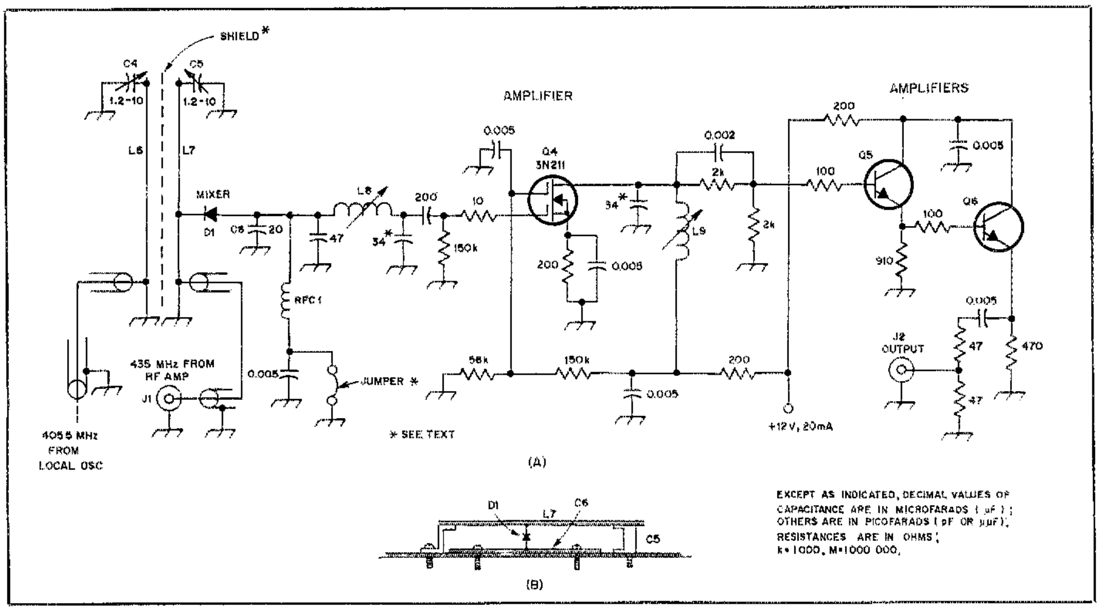


Fig. 3 — Schematic diagram of the mixer is shown at A; a detail drawing of L7 and C6 is at B. All resistors are 1/4 W; fixed-value capacitors are disc ceramic.

C4, C5 — 10-pF piston trimmer.
C6 — 2-1/2 × 3/8 × 1/16-in. double-sided glass-epoxy pc board. See text.
D1 — Schottky diode (Radio Shack 276-1124).
J1, J2 — UG-1094 BNC chassis connector (Radio Shack 278-105).
L6, L7 — 3-1/4 × 3/8 × 1/16-in. double-sided glass-epoxy pc board. L6 input tap 3/4 in. from cold end. L7 input tap 3/8 in. from cold end. D1 tap 1-1/4 in. from cold end.
L8, L9 — 0.63 to 1.02 μH (Miller 40A827CB1).
Q5, Q6 — Radio Shack 276-2033.
RFC1 — 42 turns of no. 34 wire wound on a 10-kΩ, 1-W resistor.

I have had some difficulties with the Radio Shack 276-2011 bipolar transistor (Q2). I bought eight of them and two were inoperable. Of the remaining six, only one had the lead connections listed on the package. I have never experienced this problem with other Radio Shack semiconductors.

As with Q2, mount the 47-pF capacitor as close to the base of Q3 as possible. A strip-line circuit, resonant at 405.5 MHz, is coupled to the collector of Q3. The collector tap position on the strip line was determined experimentally for best signal purity consistent with reasonable output.

All converter strip lines are made from double-sided glass-epoxy pc board. The necessary dc isolation is obtained by using both sides of the board. Compatibility with the piston-trimmer capacitors led to the choice of 3/8-inch-wide strip line, spaced 3/8 inch away from the pc mounting board. The strip lines, at their "cold" ends, are fastened to the mounting board by means of copper brackets (Fig. 2B).

Output is approximately four times that required by the mixer; it appears to be relatively free from spurious energy. Tuning the output strip line either plus or minus 50.7 MHz produces no perceptible output.

Mixer

I tried several active mixers using commonly available FETs. Rated to perform above 500 MHz, the devices did show considerable gain. However, noise-figure performance was better using a Schottky diode. It is a bit more difficult to make a single-ended diode mixer circuit perform optimally. Nevertheless, adjusting for optimum noise figure is not a formidable task when the local oscillator is clean and you have a gated noise source.

Two strip lines, mounted with their centers 7/8 inch apart, are used in the input of the mixer circuit (Figs. 3 and 4). Owing to the high oscillator output there should be very light coupling between L6 and L7. Coupling is controlled by an L-shaped shield of solid, light-gauge aluminum. A pair of screws through the foot of the shield mounts and holds it centered between the strip lines. I adjusted the shield dimensions for about 1 to 2 mA of mixer current. Mixer current can be read by substituting a milliammeter for the jumper from RFC 1 to ground.

Mounting details for C6 are shown in Fig. 3B. The capacitor is formed by the two sides of a piece of double-sided pc board. Be sure to remove copper from around the mounting-screw heads. That will prevent short-circuiting the capacitor.

A pi network is used to obtain an impedance match between the mixer diode and the low-noise MOSFET, Q4. The 34-pF capacitors at the input and output of Q4 are actually each a pair of series-connected 68-pF disc-ceramic units.

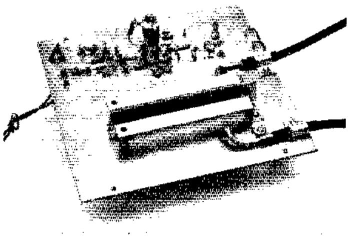


Fig. 4 — Photograph of the mixer module.

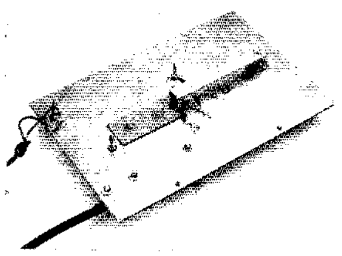


Fig. 5 — Photograph of the rf-amplifier module.

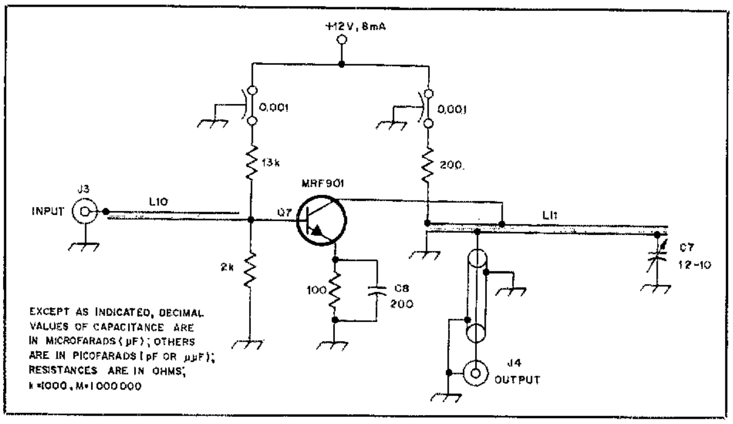


Fig. 6 — Schematic diagram of the rf amplifier. All resistors are 1/4 W; fixed-value capacitors are disc ceramic unless otherwise specified. Radio Shack part numbers are in parentheses.

C7 — 10-pF piston trimmer.
C8 — 1-3/4 × 3/4-in. pc board insulated from mounting board by Saran Wrap®. See text.
J3, J4 — UG-1094 BNC chassis connector (278-105).
L10 — 2-3/4 × 3/8 × 1/16-in. double-sided glass-epoxy pc board. Bias resistor tap is 1/2 inch from transistor end.
L11 — 3 × 3/8 × 1/16-in. double-sided glass-epoxy pc board. Collector tap 7/8 in. from cold end. Output tap 1/2 in. from cold end.
Q7 — (276-2044).

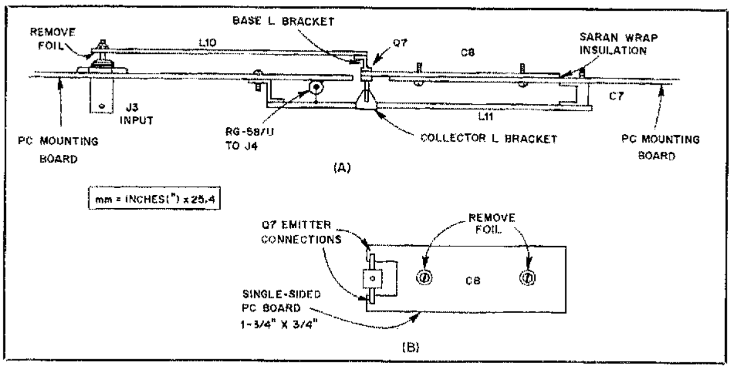


Fig. 7 — Rf amplifier module assembly detail is shown at A; details of C8 are at B.

Post-amplifier gain (approximately 40 dB) is much more than is usually required for the application. Emitter followers Q5 and Q6 were chosen for low output impedance, simplicity and stability.

The module is mounted on a 5-1/2 × 5-1/2 inch double-sided pc board. Input and output circuits are mounted on opposite sides of the board to provide isolation.

Preamplifier

It would be nice to use a GaAs FET preamplifier. However, the readily available and inexpensive MRF901, with a rated noise figure of approximately 1.5 dB and a gain of 20 dB at 450 MHz, is not a poor compromise. The transistor is a consistent and stable performer.

The preamplifier is mounted on a 6 × 4-1/2 inch double-sided pc board (Fig. 5). A broadband strip-line circuit is used for input coupling (Fig. 6). Double-sided pc board provides dc isolation between the antenna and the transistor base circuit. A special emitter-bypass capacitor is made from pc-board material insulated with Saran Wrap® (Fig. 7). Saran Wrap is a soft material, and exceptional care must be taken to ensure that the mating surfaces are smooth; soldering must be done before assembly. I have experienced no problems with the material after final assembly.

The position of the collector tap on the output strip line was determined experimentally. It was set for a near-maximum-gain condition, consistent with a reasonable Q. The output circuit is sharp enough to provide substantial rejection of the 376-MHz image.

Mode-J Filter

Energy from the 145-MHz uplink transmitter may overload or "desense" the preamplifier. A strip-line filter has proved effective in preventing this (Figs. 8 and 9). L12 and L13 are spaced 1/2 inch on centers. Insertion loss of the filter is slight, being barely perceptible with a noise generator and oscilloscope. Checks were made with the filter in and out of the line.

Test Equipment

You don't need a digital ohmmeter to check continuity — a battery and lamp or buzzer will do the job. Having a lab full of uhf test equipment is nice, but a few relatively simple devices will add little to the complexity of this project. Those devices, properly used, will virtually ensure successful performance.

Lecher Wires

How would you measure frequency in the uhf range? Would you try to borrow a uhf frequency counter? (You probably don't own one — they are expensive!) Lecher wires have been around almost as long as radio communication. They can be used to determine frequency with a fair degree of accuracy; they are also simple, inexpensive and easy to use. When the end-effect factor has been calculated while using a known frequency, further measurements can be made with 1% accuracy.

Simple Lecher wires are shown in Fig. 10. A one-turn loop is mounted permanently at the shorted end; a sliding short can be moved freely along the wires. Resonance of the Lecher wires is indicated by a null in the reading on the external VTVM. Using the known frequency output from the oscillator module, I was able to calculate end effect using the formula:

$$\ell \text{ (inches)} = \frac{5904}{f(\text{MHz})} \times \text{end effect} \qquad \text{(Eq. 1)}$$

The result, 0.96, was substituted into the formula, which was then used to calibrate the frequency scale opposite the sliding short.

Before the Lecher wires were calibrated, I used them to great advantage during the development of the oscillator. One time I was able to detect, and consequently eliminate, an 800-MHz spurious oscillation. Another time, I discovered that the final doubler stage was actually acting as a tripler. After retuning, it functioned properly. Later, when the Lecher wires had been calibrated, they were used to calibrate the uhf test oscillator.

Uhf Test Oscillator

I had two goals for the test oscillator design: It should have a wide frequency range and no major output level variations. The simple circuit shown in Fig. 11 is the result. The output frequency is varied through two ranges. With C12 set at maximum capacity, C11 tunes the oscillator from 340 to 420 MHz; with C12 at minimum, C11 tunes from 410 to 510 MHz. An L network output attenuator was chosen to minimize output variations as a function of frequency. An additional 20 dB of attenuation is used at J9.

A polystyrene block supports the

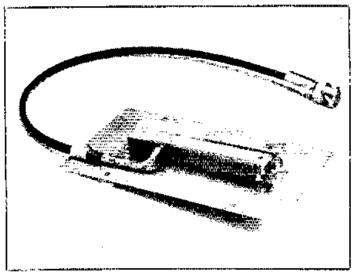


Fig. 8 — Photograph of the Mode-J filter.

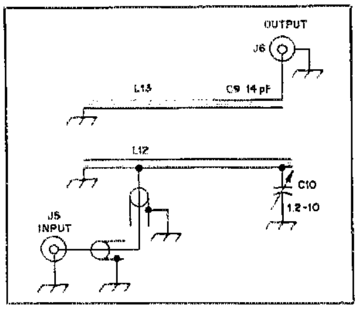


Fig. 9 — Schematic diagram of the Mode-J filter.
C9 — 14-pF pc-board capacitor (formed by removing 1-1/2 in. of foil from the top side, cold end of L13).
C10 — 10-pF piston trimmer.
J5, J6 — UG-1094 BNC chassis connector.
L12 — 3-1/4 × 3/8-in. glass-epoxy pc board. Input tap 7/8 in. from cold end.
L13 — 3-1/4 × 3/8 × 1/16-in. double-sided glass-epoxy pc board.

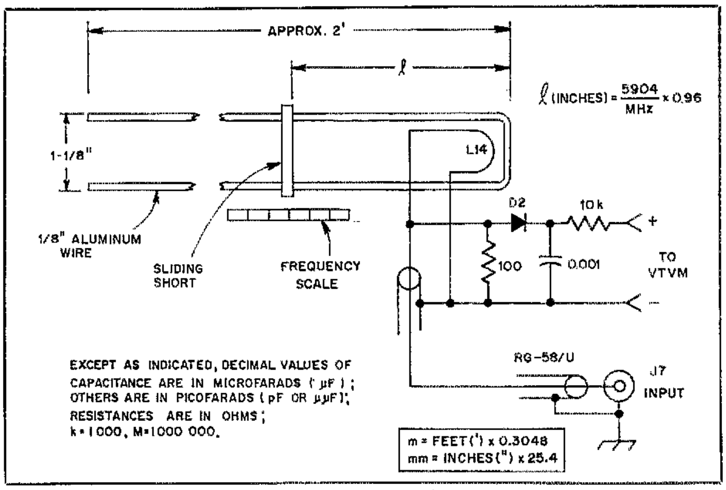


Fig. 10 — Diagram of the Lecher wires.
D2 — Schottky diode (Radio Shack 276-1124).
J7 — UG-1094 BNC chassis connector.
L14 — Single-turn loop of no. 14 wire, 1 × 9/16 in.

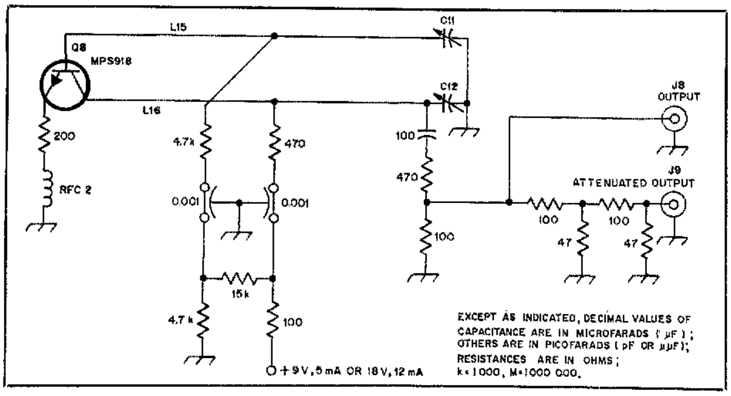


Fig. 11 — Schematic diagram of the uhf test oscillator. All resistors are 1/4 W; fixed-value capacitors are disc ceramic.

C11, C12 — 9-pF subminiature variable (Johnson 9M11 or equiv.).
J8, J9 — UG-1094 BNC chassis connector.
L15, L16 — No. 14 copper wire 5-3/4 in. long, tapped 2 in. from the hot (C) end.
Q8 — Radio Shack 276-2011.
RFC2 — 15 turns of no. 28 wire wound 3/4 in. long on a 1/4-in.-diameter plastic form.

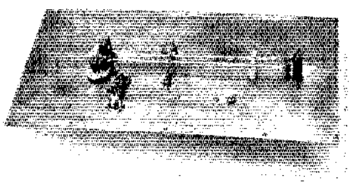


Fig. 12 — Photograph of the uhf test oscillator.

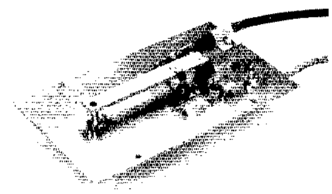


Fig. 13 — Photograph of the gated noise generator.

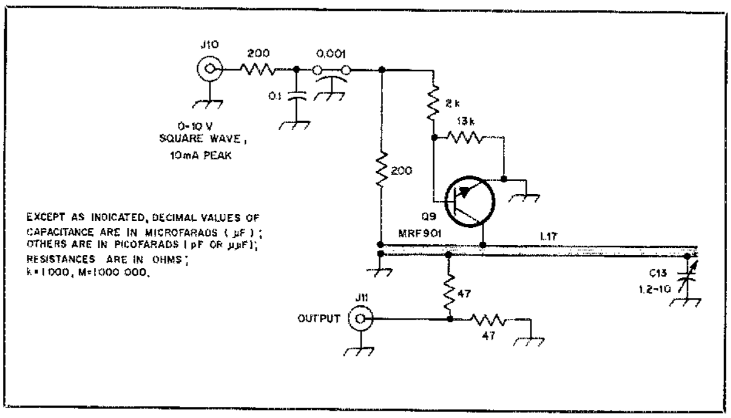


Fig. 14 — Schematic diagram of the gated noise generator. All resistors are 1/4 W.

C13 — 10-pF piston trimmer.
J10 — Phono connector, single-hole mount.
J11 — UG-1094 BNC chassis connector.
L17 — 3 × 3/8 × 1/16-in. double-sided glass-epoxy pc board. Collector tap 7/8 in. from cold end. Output tap 1/2 in. from cold end.
Q9 — Radio Shack 276-2044.

"cold" (transistor) end of L15 and L16; spacing between them is 5/16 inch. They are both spaced 3/8 inch above the pc mounting board (Fig. 12).

Frequency stability is poor, particularly as a function of input voltage. This makes the signal unsuitable for tests using a narrow-bandwidth receiver at the i-f. It can be easily used to check the converter if you add a diode detector at the mixer module output.

I used the test oscillator for two major purposes. First, it proved helpful in checking initial strip-line resonance. When the oscillator output is coupled into a strip line, several volts of rf will be present when the line is tuned to resonance. Rf voltage can be measured by means of an rf probe that uses a Schottky detector diode. Some loading will result from connecting the probe to the strip line; this can be minimized by moving the probe toward the cold end.

A second use of the oscillator was as a large-signal source for converter testing. This allowed me to check for spurious responses that might result from stray resonances or faulty local-oscillator injection. In addition, I was able to get an idea of converter image response. Owing to the lack of frequency stability, however, use of a narrow-band i-f was not possible. A simple diode detector at the mixer-module output works just fine.

Noise Source

Since the 1979 edition, *The Radio Amateur's Handbook* has contained a description of a gated noise source and the procedure for using it.[3] Although the procedure section is fine, I found that the output level was low at 435 MHz. The circuit shown in Figs. 13 and 14 overcomes that problem. Noise output is sufficient to permit using a 50-ohm terminating network to ensure proper loading. A strip-line tuned circuit limits noise output to the 70-cm region; this helps prevent tune-up on the image frequency. Noise output is a function of input voltage. I use an ancient square-wave generator to supply up to 10 peak volts of square-wave drive.

Final Comments

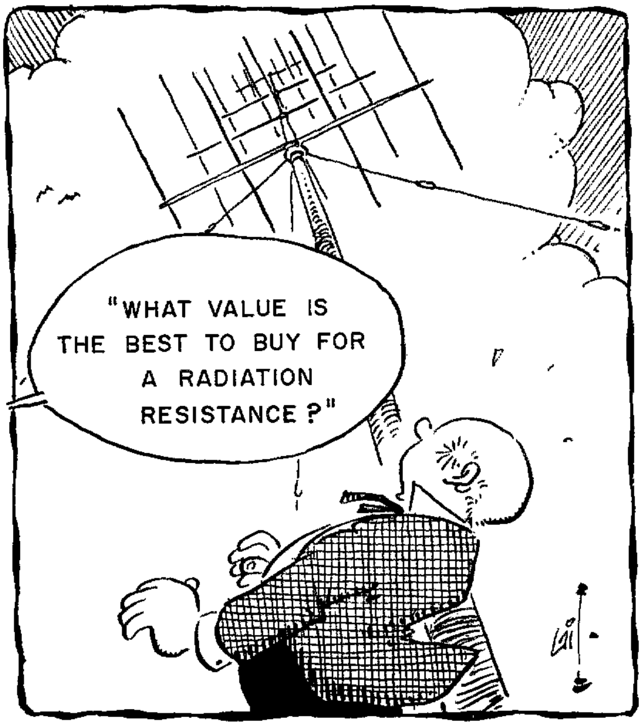
Perhaps you are now a little more inclined to try building a uhf converter — I hope so! Building your own gear and test equipment (and getting it to perform properly) is indeed a satisfying achievement. It is also great experience for hams who want to learn more about uhf.

My desire was to operate OSCAR 8, Mode J. The converter has performed in a respectable manner for me on Mode J. I have enjoyed QSOs with hams as far away as New England and as close as Los Angeles. My next plans for the converter are to use it in conjunction with the Phase III satellite on Mode L. See you there?

Notes

[1]John J. Meshna Jr., Inc., P.O. Box 62, E. Lynn, MA 01904.
[2]mm = in. × 25.4.
[3]*The Radio Amateur's Handbook* (Newington: ARRL, Inc.), Chapter 16.

BEGINNER'S BENCH



A Simple Approach to Antenna Impedances

Every active amateur owns at least one antenna. And every antenna has a property we call impedance. You can't see it and you can't feel it, but it's there.

By Jerry Hall,* K1TD

In July 1977 *QST*, we introduced two fictitious characters, Gus and Jack.[1] Gus was the seasoned amateur, the fellow who had tutored several folks into Amateur Radio and helped them progress to the higher classes of license. Jack was the new Novice who lived down the street from Gus. In 1977 Gus was helping Jack learn some of the theory he'd need for the General class exam. Well, he passed that exam with flying colors, and has since gone on to obtain his Amateur Extra ticket.

But enough about Jack. Our scenario now opens with Gus and Donna working on an antenna in Donna's backyard. Donna is a young housewife who had just received her Novice ticket a few days before. She has already made a few 80-meter contacts with a wire dipole antenna, but is itching to work some DX (foreign countries). Donna's OM bought her a new 15-meter beam, and she accepted Gus's offer of assistance in getting it up. (Her OM knows *nothing* about these things, she told Gus.) The installation is going smoothly, with Gus and Donna simply following the manufacturer's instruction sheet for the assembly of the array. But Donna wants to understand all the technical information in the instruction sheet and the brochures that accompanied the antenna. Let's listen as they converse.

"Gus," Donna is saying, "will my SWR curve be exactly like this curve here?" She was referring to the printed instruction sheet for the antenna they were assembling. Donna already knew that SWR stood for standing-wave ratio.

"Well," Gus began his careful reply. "That curve is typical for this antenna. We'll measure the SWR across the band later and see what we get, but it should be close to what they show there."

"You mean it won't be exactly the same," Donna inferred from Gus's reply. "Will that mean there's something wrong with my antenna, like maybe we didn't put it together right?"

"Oh, no, nothing like that," Gus was quick to say. "What I mean is that the impedance of any antenna will depend a lot on where it is installed. Its height above ground and being near other objects will have the greatest effect on the impedance."

"Yes, but what does that have to do with the SWR curve," Donna asked.

"Well, no two antenna installations are exactly alike. The impedances for your antenna may be a little different from

[1] J. Hall, "A Simple Approach to Complex Circuits," QST, July 1977, p. 35.

*Associate Technical Editor

those at the factory when they measured the SWR, even though the antennas are identical. And that could change the SWR curve a little bit," Gus explained.

"I guess I don't really know what impedance is," Donna confided. "How do different impedances change the SWR curve?"

Radiation Resistance

"Oh," exclaimed Gus, realizing he hadn't answered Donna's question as clearly as he could have. "Forget about impedance for a little bit, and let's talk about resistance. You remember what resistance is, don't you?"

"Yes," Donna answered, "resistance is something in a circuit that limits the flow of electric current." "That's right," said Gus. "When you're talking about an ordinary circuit, that kind of resistance consumes power, and it produces heat in the process. The consumed power can be calculated from the equation

$$P = I^2R \qquad \text{(Eq. 1)}$$

where
- P = power in watts
- I = current in amperes
- R = resistance in ohms

An antenna also has resistance, but . . ."

Donna interrupted, "But won't that resistance be *really* low? For my 80-meter dipole there's nothing but copper wire. And for this beam antenna it's all aluminum tubing. That can't have very much resistance, can it?"

"That's true," Gus agreed. "But the kind of resistance I'm talking about is called the *radiation resistance* of the antenna. It isn't a real resistor. In fact, you can't even measure the value of the radiation resistance with an ohmmeter."

"I don't understand," said Donna. "How can you have a resistance that you can't even measure?"

"Oh, you *can* measure it," Gus remarked, "but it takes a special kind of measuring instrument." [See the article by Priedigkeit elsewhere in this issue. — Ed.] "I just said you can't measure it with an ohmmeter. That's because it isn't the same as an ordinary resistor. Let's say you key your transmitter with your antenna connected. And let's say that 100 watts of power is delivered to the antenna terminals through the transmission line. What happens to that 100 watts?"

"Well, I guess it gets radiated," was Donna's reply.

"Yes, that's right. Most of it does get radiated, and that's the useful part of the energy. But current flows in the wire or tubing, and a small part of that energy is converted to heat. This happens because of the resistance you were thinking about. The heating losses are usually called ohmic losses, and the same old I^2R equation applies (Eq. 1, above). In the case of these heat losses the resistance is real; it's the resistance of the copper or aluminum conductors. But in the case of the power being radiated from the antenna, the resistance is only assumed to exist. Its value is the one that would consume the power being radiated by the antenna."

"Oh," said Donna, "what value is the best to buy for a radiation resistance?"

"You can't just go out and buy a value of radiation resistance to your liking," Gus explained. "Each different type of antenna has its own radiation resistance, and that value depends on the fundamental design of the antenna. For example, a half-wave dipole that is far away from the earth and any other objects would have a radiation resistance of about 73 ohms. We call this its free-space resistance. This resistance can go either up or down from that value when you put the dipole close to other objects or at a practical height above the ground. Are you with me so far, Donna?"

"I think so," she stated. "What about other antennas? What kind of radiation resistance will this 15-meter antenna have?"

"Good question," Gus commented. "This is a 3-element parasitic beam. It has one driven element and two parasitic elements. The driven element is like a dipole."

"Oh," chimed in Donna, "then will the free-space radiation resistance of my beam be 73 ohms?"

"No, it'll be much lower. That's because of the presence of the two parasitic elements. It'll probably be something like 20 to 30 ohms."

"Well, then, shouldn't we use something besides this 50-ohm coax to feed the beam?" Donna asked.

"No, we're okay on that. The manufacturer has included this matching arrangement to transform the low resistance up to 50 ohms, so you *can* feed it with 50-ohm line," Gus explained. "This kind of matching arrangement is called a beta match. Another popular type is a gamma match. But I won't go into matching arrangements just now. Maybe we can talk about them later."

"Okay," replied Donna. "A little bit ago you said a *half-wave* dipole had a radiation resistance of 73 ohms. What if it's not a half wave? Will the radiation resistance be different?"

"Aha!" Gus exclaimed. "You're thinking the resistance might change with antenna length, and you're right. If the dipole were made shorter, its radiation resistance would go to lower values. And if it were made longer, the resistance would go higher, up to a point. Now tell me, Donna, what happens to the radiation resistance if we have a fixed length for an antenna, but we use it for different frequencies, like your 80-meter dipole? You use that for several different frequencies in the Novice band, you know."

"Well," Donna mused, "the antenna is a half-wavelength long for 80 meters, so I guess if it were in free space it'd be 73 ohms for all the Novice frequencies. Is that right?"

"Not quite," was Gus's reply. "The antenna is really a half-wavelength long at *just one* frequency. As you move up or down from that frequency, the antenna is not an electrical half wave any more. So its radiation resistance changes as you move in frequency."

"Oh, I see!" Donna exclaimed. "That's why my SWR reading changes when I change frequency."

Antenna Reactance

"Partly true," said Gus. "But across the 80-meter Novice band the radiation resistance won't change very much — only a few ohms. And that by itself won't have much effect on your SWR. But remember, I mentioned the antenna is really a half-wavelength long at just one frequency. At that frequency the antenna is said to be resonant. And resonance means there is no reactance present. You remember what reactance is, don't you Donna?"

"Doesn't reactance come from circuit elements that don't consume any power?"

"Exactly right. And remember, Donna, only resistances can absorb power. Or in the case of the radiation resistance, you might say it can radiate power. You know what kind of circuit elements have reactance, don't you?"

"Oh, sure," Donna replied, "inductors and capacitors. And now I'll bet you're going to tell me that an antenna can have reactance, even though there aren't any inductors or capacitors there."

Gus exclaimed, "You took the words right out of my mouth! With a half-wave dipole, you get inductive reactance as you go above the resonant frequency. And you get capacitive reactance when you go below. It's exactly the same as a tuned circuit. At circuit resonance there is no reactance, and as you depart from resonance, you get either inductive or capacitive reactance, depending on which way you changed the frequency."

"I see," Donna said gleefully. "A little while ago you said the radiation resistance didn't change much over the band, and you said the change wouldn't have much effect on the SWR. Then it must be the reactance that makes the SWR go up when I change frequency."

"Exactly right," Gus confirmed. "How about if we take a short break from putting this antenna together, and I have you do some simple calculations on paper? You're very close to having a good understanding of antenna impedance."

"Good idea," Donna thought aloud.

Antenna Impedance

Inside, Donna fixed cool drinks to sip while she and Gus talked. "You know,

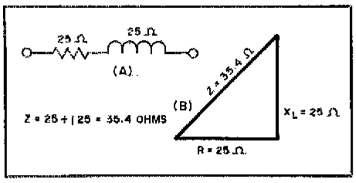


Fig. 1 — Electrical circuit and diagram representing an impedance of 25 + j25 ohms. The lower-case j is a shorthand notation indicating that the ohmic values cannot be added directly. The plus sign with the j indicates inductive reactance.

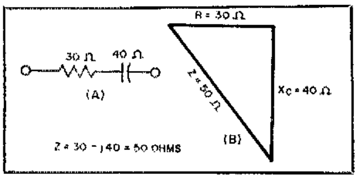


Fig. 2 — Electrical circuit and diagram representing an impedance of 30 − j40 ohms. The minus sign with the j indicates capacitive reactance.

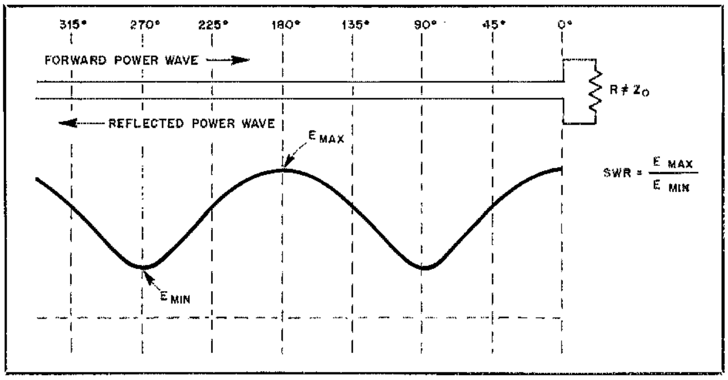


Fig. 3 — Voltage standing wave along a mismatched line, resulting from components of the forward power wave and the reflected power wave combining in various phase relationships at different points on the line. The power waves travel along the line, while the standing wave is stationary. The positions of the maxima and minima depend on the value of the load components with respect to the line impedance, but there will always be an electrical quarter wave-length of line between a maximum and a minimum.

Donna, an impedance is nothing more than what you get from a circuit that has both resistance and reactance in it. And as I've indicated, an antenna behaves just like a circuit." Gus explained further, "You've heard me talk about antenna impedance. Such an impedance is made up of some resistance and some reactance. The resistance and the reactance are both measured in ohms. What if we had an antenna with 25 ohms of radiation resistance and 25 ohms of inductive reactance? Can you tell me what the total antenna impedance would be?"

"I think it'd be 50 ohms," was the reply, "25 plus 25."

"Nope, sorry." Gus corrected, "It's more like 35 ohms."

"How'd you get that number?" Donna asked sharply.

"That's why I wanted to sit down with pencil and paper, so I could show you," said Gus. He drew the circuit shown in Fig. 1A, talking as he was drawing. "This circuit represents the impedance I was talking about." Then he drew the triangle shown in Fig. 1B. "And here's how we diagram that impedance. This horizontal line represents the resistance, R, 25 ohms. And this vertical line represents the inductive reactance, X_L. That's also 25 ohms for this example. It's customary to draw the inductive reactance line upward from the base line or resistance line, and to draw a capacitive reactance line downward from the base line. Now, the hypotenuse of the triangle represents the total impedance, and we represent that with the letter Z."

Recognizing the solution from earlier school days, Donna said, "Oh, yes, we can use the equation

$$Z = \sqrt{R^2 + X^2} \qquad \text{(Eq. 2)}$$

where

Z = total impedance
R = resistance (radiation resistance of the antenna)
X = reactance (may be inductive or capacitive)

and we get 35.4 ohms." Donna had used her household electronic calculator to figure the answer.

"Good," said Gus. "Here's how we indicate that same impedance without having to draw the triangle. It's a more convenient way when you are writing about impedances." At the bottom of the drawing he wrote, "Z = 25 + j25 = 35.4 ohms," as shown in Fig. 1. "That little letter j is a shorthand way of saying that you cannot add the resistive ohms and the reactive ohms directly. You have to use the process you went through (Eq. 2). Sometimes that is called vector addition."

Offering another example, Gus said, "How about if we had 30 ohms of resistance and 40 ohms of capacitive reactance? Draw the equivalent circuit and diagram the impedance for me. And tell me the impedance value." Donna drew the information shown in Fig. 2, and calculated the impedance to be 50 ohms.

"Exactly right," Gus said proudly. "I think you've got this antenna impedance business down pat."

Standing Waves and Standing-Wave Ratios

"Well I'm not sure," said Donna. "I still don't understand exactly how the SWR curve depends on the impedance."

"Okay," said Gus. "One important thing to remember is that the reactive part of the impedance does not absorb any power. Or in the case of an antenna, it does not radiate any power. But it can prevent some of the available power from being transferred to the resistance. The power that isn't radiated is then reflected back down the transmission line, and that's what creates a standing wave."

"If the antenna had no reactance, then you'd have no standing wave?" Donna asked. "And just what is a standing wave?"

"Well, let me answer your questions in the reverse order," Gus said as he began drawing what appears in Fig. 3. "A standing wave is developed on the transmission line any time you have power reflected from the antenna. The forward power and the reflected power waves are traveling in different directions along the line. If the line is long enough, at some points in the line the voltages in the two waves will be in phase with each other. At these points the voltages will add, and the total rf voltage there will be greater than the voltage from either power wave alone. And at some other points on the line the two waves will be 180° out of phase with each other. At these points the voltages will tend to cancel each other, and the total rf voltage will be less than from either power wave alone, as I've shown in this drawing."

He continued, "Now this wavy line represents the voltages that are developed at all the points along the mismatched transmission line. A lot of amateurs forget that at every point along the line the rf voltage will go through zero at some instant in time. You see, the rf voltage at any point on the transmission line is a sine wave. It's just that the amplitude of the sine wave is higher at some points than it is at others. And this is what that wavy line represents, nothing more than a plot of the rf voltage at all points along the line. That wavy line also represents the standing wave, which is merely the resultant of the forward power wave and the reflected power waves."

"Then how does that tie in with standing-wave ratio?" Donna asked.

"Simple. By definition, the standing-wave ratio, or the SWR, is the ratio of the maximum voltage to the minimum voltage in the standing wave." Gus indicated this information on his sketch, also shown in Fig. 3.

$$SWR = \frac{E_{MAX}}{E_{MIN}} \quad \text{(Eq. 3)}$$

where
SWR = standing-wave ratio
E_{MAX} = maximum amplitude in the standing wave
E_{MIN} = minimum amplitude in the standing wave

"Now to answer your first question, it's not always true that you have no standing wave when you have no reactance. No standing wave means you have no reflected power. A line operating under these conditions is sometimes called a flat line, because the wavy line showing the voltage along the line now becomes flat. And of course the SWR is 1:1, because the ratio of maximum to minimum voltage along the line is 1."

"That's easy to understand," said Donna.

Gus continued, "But you can still have a mismatched line without having reactance at the load. Say you have an antenna that has a radiation resistance of 100 ohms but no reactance. And say you feed it with 50 ohm-line. The line is not matched, so there would be some reflected power."

SWR vs. Antenna Impedance

"What would the SWR be if you did feed a 100-ohm antenna with 50-ohm line?" Donna asked.

"If you have a load that is resonant or purely resistive, it's easy to figure the SWR. Just divide the line impedance into the load resistance, or vice versa, whichever gives a number bigger than 1. In equation form, you can write it this way.

$$SWR = Z_0/R \text{ or} \quad \text{(Eq. 4)}$$

$$SWR = R/Z_0 \quad \text{(Eq. 5)}$$

where
SWR = standing-wave ratio
Z_0 = characteristic impedance of transmission line
R = load resistance (load must be purely resistive)

Use the equation that gives an answer greater than 1," Gus continued. "So your SWR would be 2:1. You get this by dividing 100 by 50. You'd also have a 2:1 SWR if you fed a 25-ohm resistive load with 50-ohm line."

"I see. And if you had an antenna with an impedance of 50 ohms and fed it with a 50-ohm line, your SWR would be 1:1. That figures, since 50 divided by 50 equals 1."

"Be careful," Gus cautioned. "That's true only if the load is purely resistive. You know, a lot of amateurs talk about *antenna impedance* when they really mean radiation resistance — or maybe radiation resistance plus loss resistance. That's what you just did. The word *impedance* sort of implies that there is some reactance present. See this example I gave you," Gus said as he pulled out the information shown in Fig. 2. "This impedance is made up of 30 ohms resistance and 40 ohms reactance. Yet the total impedance is exactly 50 ohms. But that'll look a lot different to a 50-ohm line than a 50-ohm resistance will."

"How much different? What would the SWR be?"

Gus had to do some brain searching to come up with an answer to that one, for it involved a set of equations he didn't use often. Finally he said, "Okay, here's how you can calculate it.

$$SWR = \frac{A + B}{A - B} \quad \text{(Eq. 6)}$$

where
$A = \sqrt{(R + Z_0)^2 + X^2}$
$B = \sqrt{(R - Z_0)^2 + X^2}$
R = resistance at load, ohms
Z_0 = characteristic impedance of feed line
X = reactance at load, ohms

"In this case, R equals 30 and X equals 40," Gus went on. Borrowing Donna's calculator, he said, "Okay, A is going to equal the square root of 30 + 50 or 80 squared plus 40 squared. That's the square root of 8000, which is 89.44. And B equals the square root of 30 − 50 or negative 20 squared plus 40 squared. That's the square root of a positive 400 plus 1600 or the square root of 2000, which is 44.72. Okay, A plus B is 134.16, and A − B is 44.72, so the SWR is 134.16/44.72, or 3.0 to 1. Donna, would you believe a 50-ohm *impedance* on the end of a 50-ohm line could give you a 3:1 SWR?

"No, I wouldn't have. Did you figure that right?" she asked, rechecking Gus's calculations. Coming up with the same answer, she exclaimed, "But I do now. It's still hard to believe, though."

"Well, we'd better get back to work on that antenna so we can get it done before dark," Gus said.

"Gus," Donna said, "I'm glad you came over to help me with this antenna. I guess almost anybody can put the parts together, but you explained it so well that I feel I really understand it. Thanks." Together they walked out the back door to the waiting hardware and tools.

Strays

Past Hudson Division Director Stan Zak, K2SJO (right), presents a plaque to *Time* senior writer Ed Magnuson, W2IJB, honoring him for his story (May 3, 1982) about the DXpedition to Navassa Island. An avid DXer himself, Ed participated in the DXpedition.

Francis C. Leonard, W2NPT, of Fairlawn, New Jersey, proudly displays the plaques he received after being named 1982 Elmer of the Year by the Northern New Jersey Chapter of the QCWA. The larger plaque will rotate annually, but the other will grace W2NPT's shack permanently. *(tnx N2XJ)*

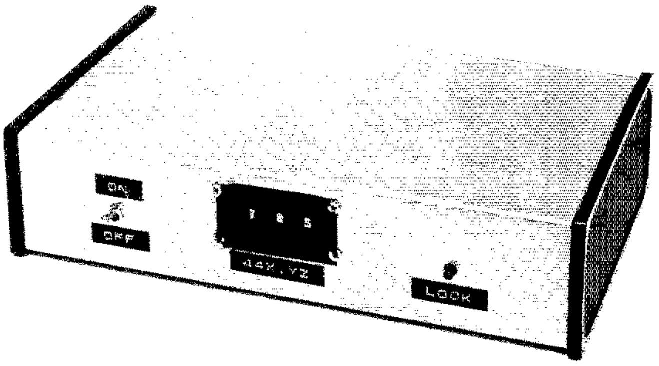
Next Month in QST

Now that the FCC has unveiled its no-code license proposal (see pages 9 and 49, this issue), it's time to digest the next piece of news from Washington: volunteers will be able to give code and theory exams. Details will appear in April *QST*, as will:

- news about a new League book explaining the ins and outs of the FCC Rules, Part 97, and
- two items of interest to microwave enthusiasts — a discussion of 10-GHz Gunnplexer theory and operation, and details on how to build a 1296-MHz slug tuner from easily obtainable parts.

Modifying a CB-Board Synthesizer for Amateur Use

If you've ever needed a synthesizer for a receiver or transmitter but weren't up to building a circuit with 10 to 20 ICs (and then trying to make it work), this is for you!



By J. Robert Witmer,* W3RW

The widely available Hy-Gain CB transceiver board can be adapted to provide synthesized frequency control of a receiver or transmitter.[1, 2, 3] Included here are some guidelines to follow when performing the conversion, along with some suggested applications. The modifications involved are simple. One elective (more complex) step provides for a high synthesizer-signal output level. This involves modifying the CB board to connect the output that is normally fed to the receiver section to the on-board transmitter predriver and driver stages for amplification.

The CB-board synthesizer section can be modified to provide a 35- to 41-MHz output range of frequencies by changing the PLL mixer oscillator crystal and retuning the VCO coil. The version I'm using with the higher drive-level modification supplies approximately 1.8 V of rf output to a 50-ohm load from 35 to 35.7 MHz. It has a very clean output waveform. Harmonics are about 45 dB down from peak fundamental output, and the close-in spurious signals are at least 60 dB down. Measured residual fm noise deviation is 0.003 kHz, which is not noticeable even after multiplication.[4]

Basic Synthesizer Operation

The unmodified CB-board synthesizer output-frequency range is approximately 37 to 38 MHz in 10-kHz steps. This output is used as high-side local oscillator (LO) injection for the receiver section (10.695-MHz i-f), and is mixed with a 10.695-MHz crystal-oscillator signal to obtain the required 26.9- to 27.4-MHz CB transmitter output frequencies. See Fig. 1. The described modification will ignore the 10.695-MHz oscillator and transmitter mixer functions and will concentrate on the 37- to 38-MHz synthesizer.

Applications Steps

The following steps will help you to modify the synthesizer for your purposes. Refer to Table 1 and Figs. 2 and 3. Details of my first conversion work, a 443- to 450-MHz synthesizer for a modified surplus fm receiver, are included as a guide. The required calculations are shown in equation form.

1) *Frequency Range.* Establish the frequency range you wish to cover. I required reception of frequencies between 443 to 449.95 MHz (Table 1, column 1).

2) *Determine The Equipment Heterodyning Scheme.* What is the multiplication factor of the multiplier string and of the first multiplier stage? It's also necessary (in the case of receiver applica-

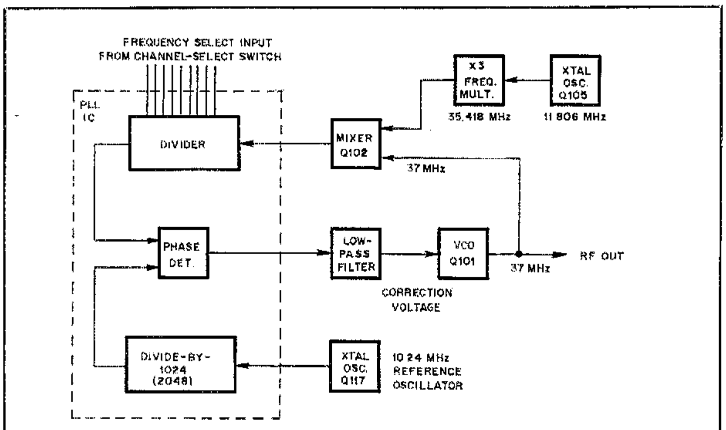


Fig. 1 — Block diagram of the Hy-Gain PLL synthesizer.

[1]Notes appear on page 24.

*79 Blaine Ave., Leola, PA 17540.

Table 1

Key Parameters For Various PLL Synthesizer Applications

1	2	3	4	5 (A)	6 (B)	7 (C)	8	9 (D)	10 (E)	11 (F)	12
443-449.95 receive (21.4-MHz i-f)	52.7-53.568	50	8/12	35.1333-35.7125	4.166	8.5333	139	256-395	1.0666-1.64383	11.3555 (34.0666)	2048
443-449.95 transmit	12.3055-12.4986	50	36/12	36.9166-37.495	4.166	8.5333	139	256-395	1.0666-1.64383	11.950 (35.850)	2048
222-224.98 receive (10.7-MHz i-f)	52.825-53.57	40	4/6	35.2166-35.7133	6.666	6.8266	75	256-331	1.7066-2.2066	12.3077 (36.9233)	1024
222-224.98 transmit	18.5-18.748	40	12/6	37.0-37.4966	6.666	6.8266	75	256-331	1.7066-2.2066	11.7644 (35.2933)	1024
222-224.98 receive (10.7 MHz i-f)	58.825-53.570	20	4/6	35.2166-35.7133	3.333	6.8266	150	256-406	0.85333-1.35333	11.4544 (34.3633)	2048
222-224.98 transmit	18.5-53.57	20	12/6	37.0-37.4966	3.333	6.8266	150	256-406	0.85333-1.35333	12.0488 (36.1466)	2048
26.965-27.405 (original configuration)	——	10	1	37.66-38.10	10.0	10.24	40	224-268	2.240-2.680	11.806 (35.418)	1024

Notes

Circled letters refer to corresponding letters in Fig. 2

1 — Application and frequency range (MHz)
2 — Original crystal frequency range (MHz)
3 — Channel spacing (kHz)
4 — Multiplication factor, old/new
5 — Synthesizer VCO range (MHz)
6 — Reference frequency (kHz)
7 — Reference oscillator crystal frequency (MHz)
8 — Number of channels
9 — Divide-by-N range
10 — Divide-by-N input frequency range (MHz)
11 — PLL oscillator crystal frequency (MHz)
12 — Divide-by-N number

tions) to know the i-f and whether upper- or lower-side LO injection is used. This can usually be obtained from the equipment schematic diagram or operating manual (Table 1, column 4).

As shown in Fig. 3A, my surplus 450-MHz receiver used a frequency multiplication factor of eight. The first stage of the multiplier string was a frequency doubler with output in the 106-MHz range. The i-f is 21.4 MHz with low-side LO injection.

3) *Channel Spacing.* Decide what sort of channel spacing you require. At 450 MHz, I felt that 50-kHz channel spacing was adequate (Table 1, column 3).

4) *Crystal-Range To Synthesizer-Range Conversion.* If you're performing this modification to use the synthesizer with an existing piece of equipment, find the original crystal frequency operating range of the unit and determine the needed synthesizer output frequency range. Calculate the input frequency range necessary for your desired band coverage. As shown in Fig. 3A, my fm receiver used a crystal frequency operating range of 52.7 to 53.56 MHz. Since the synthesizer output is about 35 to 40 MHz, I decided to use the first multiplier stage of the surplus receiver (which normally operated as a doubler) as a tripler. Therefore, I needed a 105.4- to 107.1375-MHz output from this first multiplier stage or a synthesizer output frequency range of 35.1333 to 35.7125 MHz. Thus, the new multiplication factor is 12. See Table 1, columns 2, 4 and 5, and Fig. 3B. More information concerning the selection of synthesizer operating frequencies is available.[5, 6]

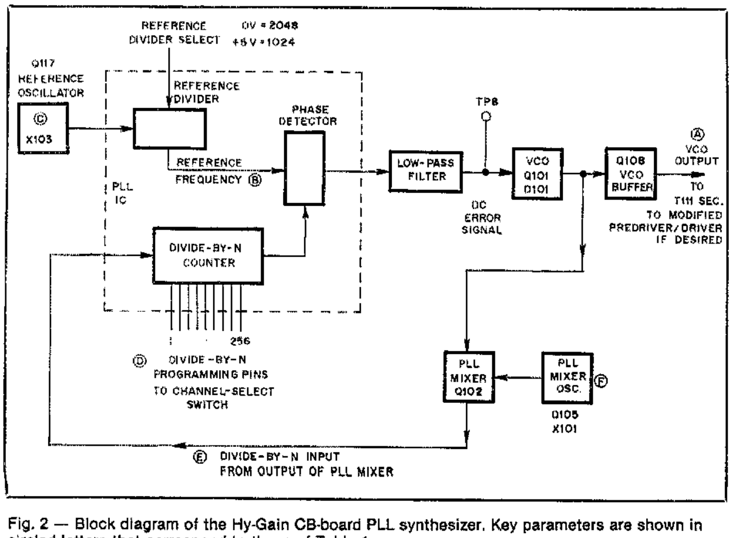


Fig. 2 — Block diagram of the Hy-Gain CB-board PLL synthesizer. Key parameters are shown in circled letters that correspond to those of Table 1.

5) *PLL Reference and Reference-Oscillator Crystal Frequencies.* The PLL reference frequency is equal to the selected channel spacing divided by the required frequency multiplication factor of the receiver or transmitter. In my application:

$$\frac{50 \text{ kHz}}{12} = 4.166666 \text{ kHz} \qquad \text{(Eq. 1)}$$

The MC145109 (IC101) on my CB board has selectable reference-divide ratios of 1024 or 2048 (pin 4 high or low, respectively). The reference-divide ratio of the CB board is 1024, but it can be modified easily to obtain a divide ratio of 2048. To determine the reference-oscillator crystal frequency, multiply the reference frequency of Eq. 1 by the two reference-divider ratios (1024 and 2048). Thus:

$$4.166666 \text{ kHz} \times 1024 = 4.26666 \text{ MHz} \qquad \text{(Eq. 2)}$$

$$4.166666 \text{ kHz} \times 2048 = 8.53333 \text{ MHz} \qquad \text{(Eq. 3)}$$

Since 8.5333 MHz is closer to the original reference crystal frequency of 10.24 MHz, I chose it and modified the PLL board wiring to use the 2048 divide ratio (ground pin 4 of IC101). See Table 1, columns 6 and 7.

6) *Number of Channels.* This is determined by dividing the frequency range (Table 1, column 1) by channel spacing (Table 1, column 3). In my case:

$$449.95 \text{ MHz} - 443 \text{ MHz} = 6.95 \text{ MHz} \qquad \text{(Eq. 4)}$$

$$\frac{6.95 \text{ MHz}}{50 \text{ kHz}} = 139 \text{ channels} \qquad \text{(Eq. 5)}$$

7) *PLL IC Divide-By-N Range.* The unmodified CB-board synthesizer has a divide-by-N range of 224 to 268 with the

divide-by-32, -64 and -128 select pins wired to V_{CC}. I freed these pins and made them available for divide-by-N programming.

At this point, the divide-by-N selection can be anything from 2 to 511. But the important factor to consider is that the maximum input frequency to the PLL divide-by-N input is 4 MHz. The highest frequency that will be applied to the PLL divide-by-N input may be determined by multiplying the highest divide-by-N number by the reference frequency. To simplify things, I wired the divide-by-256 pin to V_{CC}. This provides a divide-by-N range of 256 to 511 or a maximum of 255 channels to be selected by the use of eight spst switches connected to the 1 through 128 divide-by-N programming pins of the PLL IC. Since I needed only 139 channels for my receiver, the divide-by-N range is 256 to 395 (Table 1, column 9).

8) *Divide-By-N Input Frequency Range.* As mentioned earlier, this is found by multiplying the divide-by-N range (step 7) by the reference frequency (step 5). For my receiver, this is

$$256 \times 4.166\ \text{kHz} = 1.0667\ \text{MHz} \quad \text{(Eq. 6)}$$

$$395 \times 4.166\ \text{kHz} = 1.6458\ \text{MHz} \quad \text{(Eq. 7)}$$

See Table 1, column 10.

9) *PLL Oscillator Crystal Frequency.* The new PLL oscillator crystal frequency can be calculated by subtracting the low-end divide-by-N frequency found in step 8 from the corresponding low-end VCO output frequency, and dividing by three. In my case:

$$\frac{35.1333\ \text{MHz} - 1.0667\ \text{MHz}}{3}$$

$$= 11.3555\ \text{MHz} \quad \text{(Eq. 8)}$$

See Table 1, column 11.

Upon completion of these steps, you're ready to order your crystals and start the physical modifications of the CB board. My crystal oscillator circuits have 39-pF capacitors in series with the crystals, so I used a 32-pF load capacitance factor when I ordered new parallel resonant crystals.

CB-Board Modifications and Alignment

Before starting modifications, check to see that the PLL circuit is basically the same as that described (see Fig. 4). I've used Hy-Gain 75A080 boards that I purchased from Poly-Paks several years ago. CB units manufactured by Kraco, Lafayette, Midland and Pierce-Simpson use the basic Hy-Gain design. Avoid the remote-PLL boards (such as the Hy-Gain 750096).

Get the original synthesizer section working. One advantage of the CB-board synthesizer approach is that if the board requires troubleshooting, there usually is a service manual available to aid you. You don't usually have this type of help with home-made synthesizers.

Crystal Installation

Refer to Figs. 4 and 5. Remove the offset oscillator crystal (X102), the PLL reference oscillator crystal (X101) and C126. Install the new PLL crystal at X101. With power applied to the board, measure the reference oscillator frequency at pin 3 of the PLL IC. If you wish, you may add a trimmer capacitor to fine-tune the oscillator. PLL oscillator operation can be checked by loosely coupling a frequency counter to the PLL mixer transistor base (Q102) or to the emitter of Q105 through a 5-pF capacitor.

VCO Alignment

Connect a high-impedance voltmeter to pin 6 of the PLL IC and program the syn-

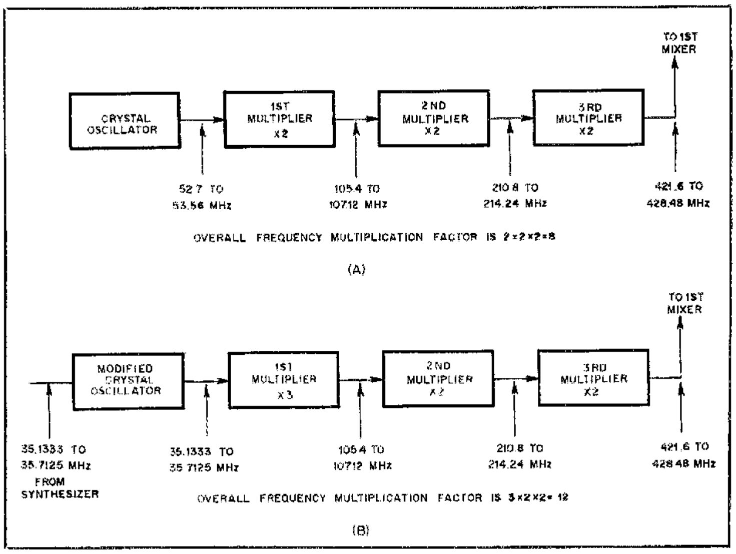


Fig. 3 — Block diagram of the original multiplier chain of the author's surplus fm receiver is shown at A. At B, the modified chain used with the synthesizer.

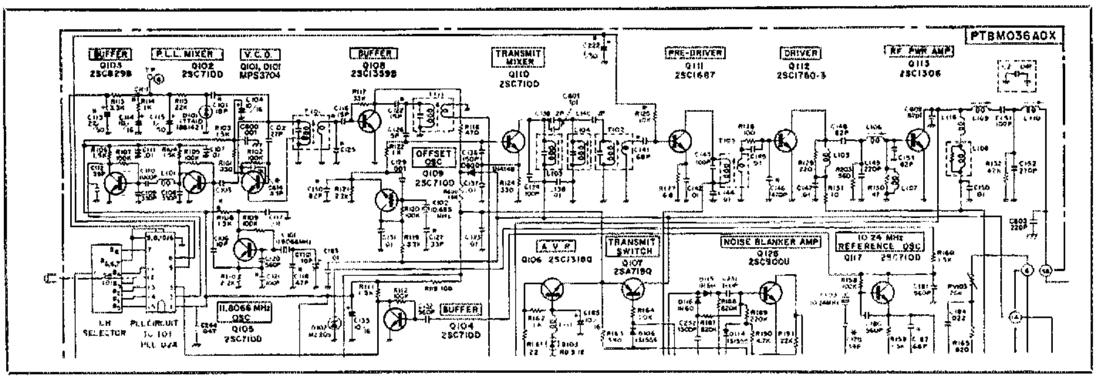


Fig. 4 — Partial schematic diagram of the CB board prior to modifications.

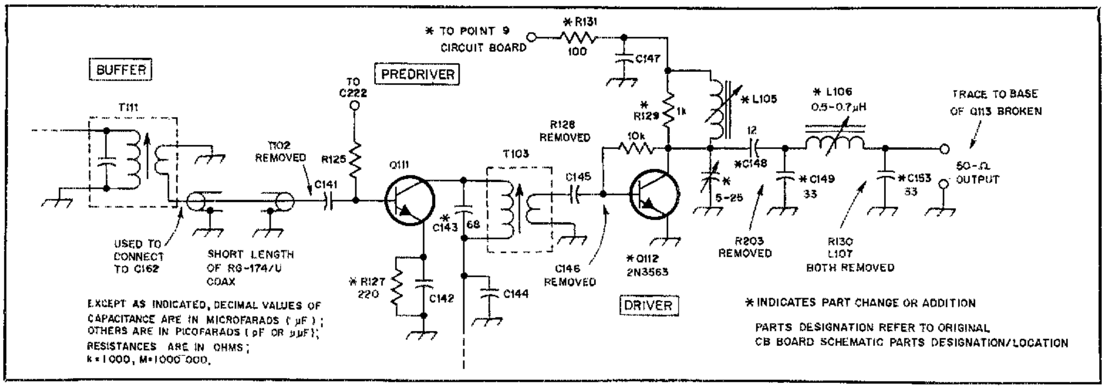


Fig. 5 — This schematic diagram shows the areas involved for the higher drive level modification.

thesizer divide-by-N for a synthesizer output frequency in the middle of the desired frequency range. If the voltmeter indicates approximately 5 V, the synthesizer is locked. If not, adjust T101 slowly until a lock is indicated by the presence of 5 V at pin 6. Once the synthesizer is locked, connect the meter to TP8 (R114) and adjust T101 so that a voltage range of about 1.5 to 4 (at the respective low and high ends of the synthesizer range) is obtained. Voltages below approximately 1 or above 4 may indicate the synthesizer is out of lock.

Reference Divider and Divide-By-N Programming Changes

If it's necessary, you should now change the PLL IC reference-divider programming. After this is completed, change the PLL divide-by-N programming input line wiring. Break the V_{CC} connections of pins 8, 9 and 10 and connect pin 7 to V_{CC}. Pins 8 through 15, inclusive, control the synthesizer frequency selection.

Final Touch-Up

You may want to recheck the VCO alignment using your intended frequency range. At this point, your synthesizer is complete except for coupling out the signal at a level sufficient to drive the oscillator or multiplier circuit of the particular receiver or transmitter. Synthesizer output can be obtained from the secondary of T111 (break the connection to C162). T111 should be then tuned for maximum output.

Higher Synthesizer Output Level

To raise the synthesizer output level, changes are needed in both the on-board transmitter predriver and driver stages. See Figs. 4 and 5 for before-and-after schematic diagrams.

PreDriver and Driver Changes

To change the predriver, connect a short piece of RG-174/U or similar 50-ohm coaxial cable from the secondary of T111 to the base circuit of predriver Q111. Remove T102 to make a connection point for the inner cable conductor. Replace the 100-pF capacitor across the primary of T103 with a 68-pF unit.

See Fig. 4 for driver changes. Carefully remove R128, R129, R130, R131, R203, C146, C148, C149, C153, L105, L106, L107 and Q112. Remove L109 and install it in place of L105. A 1-kΩ resistor is mounted where R129 was located. Install a 5- to 25-pF trimmer capacitor at the board location for C221 (Q112 collector). A 2N3563 is mounted at the Q112 location. Connect a 10-kΩ resistor from the base to the collector of Q112. At location C148, mount a 12-pF disc ceramic capacitor. Install a 100-ohm resistor from the junction of C147/R129 to the pad adjacent to connection point 32. Connect a wire from point 32 to point 9 (+12 V). Install 33-pF disc ceramic capacitors at locations C149 and C153. At location L106, mount a 0.5- to 0.7-µH coil or rf choke. Break the connection between the base of Q113 and the junction of L106 and C153. Connect the coaxial cable output lead across C153.

Tune-Up

You may find it helpful to preadjust the L105/C221 combination for resonance at the VCO frequency with a dip meter (GDO) prior to applying power to the board. When power is applied, adjust T111, T103, L105, C221 and L106 (if you've used a variable inductor) for maximum rf output across C153. I used a VTVM equipped with an rf probe as a tuning indicator. You may also monitor the voltage drop across R131 or the CB-board power-supply output current. In the latter case, adjust T111 and T103 for maximum current drain. As mentioned earlier, I have obtained 1.8 V of rf output across a 50-ohm load from this synthesizer. This should be more than enough to drive most oscillator circuits. Exact interfacing instructions for this synthesizer and your receiver or transmitter are beyond the scope of this article, but notes 5 and 6 contain excellent discussions of this subject.

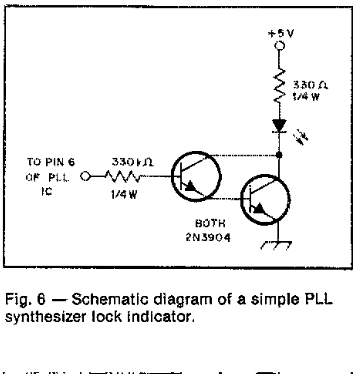
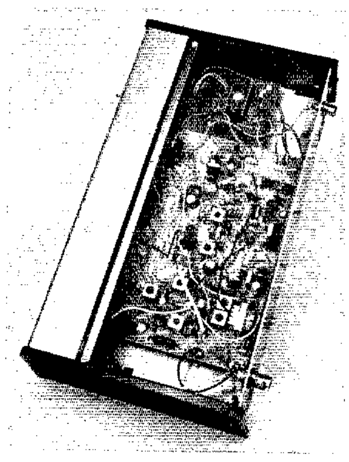


Fig. 6 — Schematic diagram of a simple PLL synthesizer lock indicator.



This photo and the title-page photo are pictures of the author's synthesizer mentioned in the text. BCD thumbwheel switches have been added to provide selection of various PLL IC divide ratios. The thumbwheel entries are sent to the PLL IC through a 2716 EPROM BCD-to-binary converter, which may be seen at the upper end of the enclosure.

Limitations

Reference Frequency. The CB-board circuit was designed originally for a

10-kHz reference frequency using a 10.24-MHz crystal in the reference oscillator, and wide deviations from these figures may not work. However, I've successfully tried reference-oscillator crystal frequencies as low as 3.7 MHz with resulting reference frequencies as low as 1.8 kHz. I've also used a variety of reference crystals at frequencies of 5.05 and 6.317, and a 25.66-MHz overtone crystal operating on its fundamental. This would seem to indicate that the circuit is quite tolerant of a wide range of crystal and reference frequencies.

VCO Operating Frequency. The lowest frequency at which I've operated the VCO is 34 MHz. Most 10-meter conversion articles I've seen operate the VCO in the 40-MHz range, so there appears to be a wide spread of VCO frequency ranges available. The original frequency spread of the VCO was from 37.66 to 38.1 MHz, or a width of 0.44 MHz. I've operated it from 35.1333 to 35.71666 MHz (a 0.58-MHz spread) without any trouble. Wider ranges should be possible at higher VCO frequencies. To optimize the VCO tuning for maximum bandwidth, monitor the voltage at TP8 (R114) with a high-impedance voltmeter. The voltage should range from approximately 1.5 at the low lock frequency to about 4 V at the high lock frequency. T101 should be adjusted to center the voltage swing within this range as much as possible. When the PLL is out of lock, the voltage at TP8 is usually greater than 4 V and the voltage at pin 6 of the PLL IC is about 0 (about 5 V when locked). A simple PLL lock indicator is shown in Fig. 6.

Summary

The Hy-Gain CB-board PLL synthesizer circuit is quite versatile. This step-by-step modification and the examples shown should enable you (with a few calculations and some circuit changes) to convert the original CB synthesizer output to a range that can be useful in many applications. It also allows you to learn about the operation of a PLL synthesizer with a minimum investment if you just wish to experiment.

I'll be happy to try to answer any questions you have about the synthesizer modification. Please enclose a large s.a.s.e. with your inquiry.

Notes

[1] H. Knickerbocker, R. Stielau and A. Wise, "CB-To-10 FM — Best Conversion Yet?," *73*, Jan. 1980.

[2] B. Heil, "Experience 10-Meter FM Operation," *QST*, Aug. 1981.

[3] Boards are available from Poly-Paks, P.O. Box 942, South Lynnfield, MA 01940. Also, Surplus Electronics Corp., 7294 N.W. 54 St., Miami, FL 33166. Be certain to request the single-unit style board, not the remote microphone version.

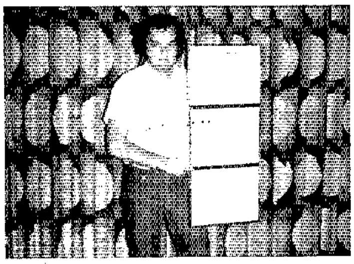
[4] ARRL lab spectral analysis of the author's synthesizer showed the second harmonic to be about 53 dB below peak fundamental output.

[5] B. Fanning and G. Grantland, "800 Channel 2-Meter Synthesizer," *Ham Radio*, Jan. 1979.

[6] GLB Channelizer instruction manual, GLB Electronics, 1952 Clinton St., Buffalo, NY 14206.

Strays

An airwound coil of 2.5 mH resonates a 10-meter whip antenna for a trial in the experimental 1750-meter band. One watt of input to a transistor final amplifier produced enough signal for a 10-mile mobile-to-fixed-station signal. *(W6HDO photo)*



Bruce Balla, VE2QO, of Quebec, shows the spoils of his DXing: cw, phone and RTTY DXCC awards. The RTTY DXCC award is number 12 in the world and number 1 in Canada. *(VE2BBP photo)*

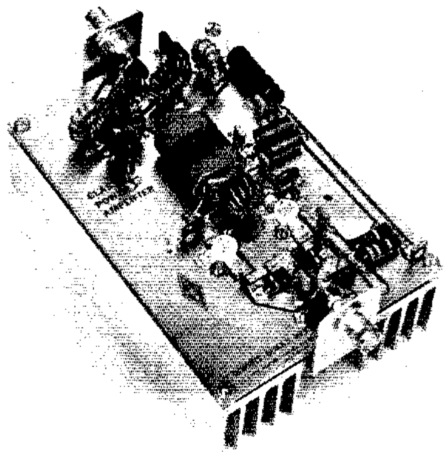
AMSAT NETS SCHEDULE

☐ The following AMSAT Nets meet regularly to disseminate information to newcomers and to keep regular satellite users in communication with one another:

Net Name	*Day/Time (UTC)*	*Frequency*	*Remarks*
HF Nets			
AMSAT Argentina	Sunday	3737	2100 Local Time
AMSAT Argentina	Sunday	14,277	2200 Local Time
AMSAT Asia/Pacific	Sunday 1100	14,305	
AMSAT Austria	Saturday 0900	7070	
AMSAT Australian	Sunday 1000	3680	During winter
AMSAT Australian	Sunday 1000	7064	During summer
AMSAT Canada			To be determined
AMSAT East Coast, U.S.	Wednesday 0200	3850	2100 EST Tuesday
AMSAT Espanol			Being rescheduled
AMSAT European	Saturday 1000	14,280	
AMSAT International	Sunday 1800	21,280	
AMSAT International	Sunday 1900	14,282	
AMSAT Mid-America	Wednesday 0300	3850	2100 CST Tuesday
New Zealand V.U.S.			Being rescheduled
AMSAT South Africa	Sunday 0900	7080	Lsb
AMSAT South Africa	Sunday 0900	14,280	
AMSAT South Pacific	Saturday 2200	28,878	
SESAT Southeast U.S.	Sunday 1300	7280	
AMSAT U.K.	Sunday	3780	1015 Local Time
AMSAT U.K.	Mon. & Wed.	3780	1900 Local Time
AMSAT West Coast, U.S.	Wednesday 0400	3850	2000 PST Tuesday
VHF Nets			
AMSAT Buenos Aires	Sunday	145.700	Fm
AMSAT Dallas/Ft. Worth	Wednesday	146.610	Fm 2000 Local Time
AMSAT Goddard	Wednesday 0200	146.835	2100 EST Tuesday
AMSAT London	Sunday	144.280	1930 Local Time
AMSAT Los Angeles	MTWThF	144.144	0730-0830 Local Time
AMSAT South Africa	Sunday 0900	145.650	Fm
AMSAT South Africa	Sunday 0900	145.725	Fm

A new "Satellite Basics" net is in the "talking and thinking" stage. Many requests to provide introductory-level material in an accessible format/frequency have been heard. Possibly, a new net will develop on 75 meters (3900-4000 kHz) or 40 meters to service newcomers. Several existing nets will be moving to Phase IIIB next summer, so other hf nets are now being looked at as "breeding grounds." Send your suggestions regarding AMSAT Nets to AMSAT Net Manager W8GQW, 1617 West McKaid Rd., Troy, OH 45373.

Go Class B or C with Power MOSFETs



The "solid-state tube" is here to stay! Power FETs offer similar performance to triode vacuum tubes, at comparable cost. Let's examine them in Class B and C amateur service.

By Doug DeMaw,* W1FB

How about a transistor that works well from 1.8 MHz to 175 MHz, delivers 30 W of output power, is virtually destruct-proof, has a gain of 15 to 18 dB and exhibits excellent two-tone IMD characteristics? The Motorola MRF138 and others of that power-FET family fit this general description. The new MRF series contains devices that will deliver as much as 100 W from a single unit (MRF150).[1, 2] Some of these transistors are in the same price class as a 6146B tube. Amplifier design is considerably less rigid when using power FETs than when power bipolar transistors (BPT) are employed.

This paper treats practical designs for 12- and 28-V power FETs. The amplifiers are highly efficient and provide a gain in excess of 20 dB. Information is given with regard to curing some of the problems that arise in power-FET design and operation — data that are kept a secret or ignored unintentionally by those who write application notes.

Power-FET Advantages

Immunity to damage from mismatching is a virtue of the power FET. It can withstand all load conditions at any phase angle without being damaged. This makes it unnecessary to include VSWR-protective circuits in one's amplifier. Thermal runaway, a specter that is ever-present with power BPTs, need not be a major concern with FETs. Also, the input capacitance (C_{GS}) and the output capacitance (C_{DS}) of the FET do not change with the operating frequency or drive level. This simplifies the design of fed-back power-FET amplifiers. The same is not true of BPTs. Still another advantage is the relatively flat gain characteristic of the power FET. The gain undergoes a gradual decline toward the f_T of the device, whereas the theoretical gain of a BPT is 6 dB per octave as the operating frequency is lowered. In other words, a specified power BPT might have a 10-dB gain at 14 MHz in a selected circuit, whereas the gain could increase to 28 dB at 1.8 MHz in the same circuit. This poses a threat to amplifier stability, especially at medium and low frequencies. To ensure a fairly constant gain bandwidth for a 2-30 MHz power BPT amplifier, it is necessary to include some form of R-C or R-C-L compensating network at the input port of the amplifier. This reduces the effective driving power as the operating frequency is lowered. Such is not the case with power-FET amplifiers; gain leveling is not needed.

Owing to the high impedance of FETs (1 megohm or greater at dc), design of the input circuit is uncomplicated. Shunt resistors can be used to establish the desired input impedance (usually between 50 and 500 ohms). The gate-to-ground resistor can serve double duty as part of a resistive divider for establishing the forward gate bias. This feature is helpful when using transmission-line broadband transformers of fixed-value integer ratios; a suitable gate resistor is chosen to provide an impedance match to the excitation source — usually 50 ohms. Bipolar transistors, on the other hand, have inherently low base impedances — typically 10 ohms or less in power amplifiers — which complicates the transformer design.

The high-order IMD of power FETs is on par with that of vacuum tubes. This ideal characteristic is not common to BPTs. Also, the efficiency of FETs is excellent, owing to the low internal resistance (R_{DS}) during conduction. This value is typically 0.25 to 1 ohm, which means that very little power is dissipated within the FET. This is especially noteworthy when using FETs that operate with high drain voltage (V_{DS}) and low drain current (I_D). Finally, the driving-power requirement for FETs is somewhat lower than for bipolar transistors. Only voltage (at a few microamperes) is needed to turn on the FET (30 V pk-pk maximum), which is easy to develop across a gate impedance of 100 or 200 ohms. Just 288 mW of driving power was required to obtain 60 W of output at 7 MHz from push-pull MRF138s in Class B service (see note 2). A simple model of a power FET is shown in Fig. 1.

Some Disadvantages

A cliche is in order: "Nothing is perfect," and this includes power FETs. Vhf self-oscillations haunt the designer. This is because most FETs are rated to 175 MHz for normal operation. The gain of an MRF-series FET is stated as 15 to 18 dB for 30 MHz, and it drops to only 10 dB at 175 MHz. So all of the encouragement

[1]Notes appear on page 29.

*ARRL Senior Technical Editor

needed for vhf parasitics is ever-present! The same kinds of preventive measures used with vacuum tubes are effective in stabilizing power FETs. That is, lowering the input impedance (swamping), bypassing the drains at vhf or using resistors in series with the gates will usually damp these unwanted oscillations. Careful layout (input-output isolation) is also necessary, as is effective bypassing of critical circuit points.

FETs can be damaged quickly — perhaps faster than one can destroy a bipolar device. The MOSFET is sensitive to excess gate or drain voltage. Too much drive, spikes on the supply lines or transients in the driving energy can dispatch an FET *instantly!* Self-oscillations, if of high magnitude, can create excessive peak voltages that will destroy an FET quickly. It's almost like the old "now you see it, now you don't" expression: The circuit can be working perfectly, when at the blink of an eye the FET is stone-cold dead! It's happened to the author more than once. Finding the cause is not always easy, because the damage occurs almost instantly. Reversed-parallel diodes from gate to ground (external to the FET) can be installed as protective clamps (Fig. 2), but they increase the effective input capacitance (C_{in}) of the circuit. They are recommended mainly for experimental circuits. Once stability is achieved, the diodes can be removed.

Similarly, Zener diodes can be used temporarily from drain to ground to prevent dangerous voltage peaks from causing damage in a circuit under development. They should have a voltage rating that is slightly greater than the anticipated peak drain-voltage under full excitation. Thus, for an FET rf amplifier that uses a 12-V dc supply for operation on cw or ssb, the peak swing at the drain would be on the order of 24 V. A 36-V Zener diode would be suitable.

The Self-Oscillation Syndrome

Taming an FET power amplifier is routine if preventive measures are taken. Most of the stabilization techniques used with bipolar transistors are applicable. The first order of business calls for careful layout of the circuit board. Input components should be isolated as much as practicable from the amplifier output components. Double-sided pc board is recommended for providing an effective ground plane on both sides of the board. The copper conductors of the etched circuit should be as short and as wide as possible to minimize unwanted inductive reactance. It must be recognized, however, that the etched conductors and the ground plane on the reverse side of the pc board form parasitic bypass capacitors. Capacitance values vary from, say, 5 pF to as much as 50 pF. Although this may be significant at vhf, it generally poses no problems in hf-band amplifiers. In fact, the added capacitance can be beneficial for discouraging vhf oscillations.

Fig. 3 illustrates stabilization methods that can be applied when self-oscillations occur. R1 and R2 establish the characteristic gate impedance for Q1 while serving as a bias-voltage divider. The lower the value of R2, the better the potential stability. R3, C1 and C2 form an R-C decoupling circuit to further aid stability. C1 is chosen as an effective bypass element in the mf and hf ranges, while C2 is selected for bypassing at vhf.

Z1, a small ferrite bead near the Q1 gate terminal on the pc board, may suffice as a vhf parasitic suppressor without R4 added. The bead or beads (two or three can be used) should have a high mu in order to be effective. An initial permeability (μ_i) of 125-900 is suggested. In stubborn cases of self-oscillation, a low-value resistor (R4) can be placed in series with Z1. Values from 10 to 27 ohms are effective.

C3 of Fig. 3 is suggested for stabilizing single-band amplifiers. The X_c should be four times or greater the characteristic drain impedance in order to prevent attenuation of the desired frequency. In other words, C3 is meant to be an effective bypass capacitor at vhf, but not at hf. Hence, if the drain impedance of the amplifier were determined from

$$Z_{D(ohms)} \doteq \frac{V_{DD}^2}{2P_o} \qquad \text{(Eq. 1)}$$

where P_o is the required power output in watts, and V_{DD} is the drain supply voltage, we might have an impedance of, say, 15 ohms. Thus, the X_c of C3 should be at least 4 × 15, or 60 ohms. For an operating frequency of 3.8 MHz, we would choose the nearest standard capacitor value to 0.00069 µF (690 pF), determined from

$$C_{\mu f} = \frac{1}{2\pi f X_c} \qquad \text{(Eq. 2)}$$

where f is the operating frequency in MHz and X_c is in ohms. A rule of thumb is to use only that value of capacitance which stops the self-oscillation (the smallest value practical). C3 will also help reduce the level of vhf harmonic energy. The C3

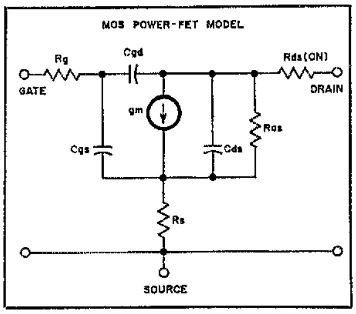


Fig. 1 — Simple model of a MOS power FET showing the intrinsic elements of resistance and capacitance.

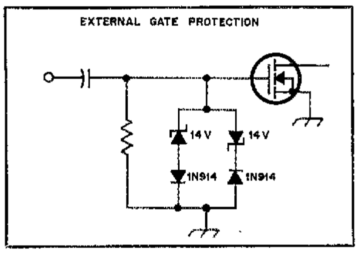


Fig. 2 — Arrangement for external gate protection of a power FET. Two 14-V Zener diodes are connected in series with two small-signal switching diodes to form a reverse-parallel network that will clamp the gate voltage at approximately 29 V pk-pk. This prevents damage from the application of excess gate voltage.

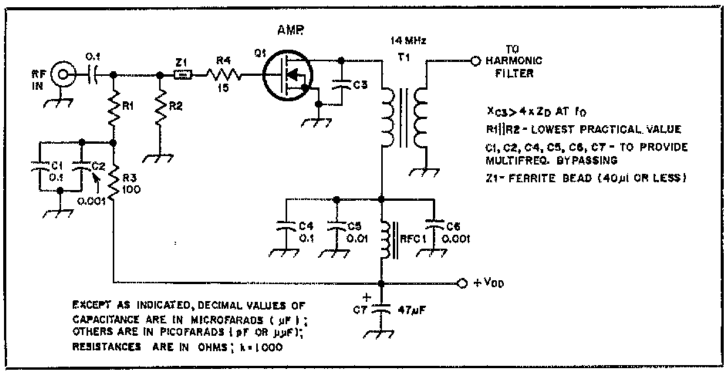


Fig. 3 — Techniques for preventing self-oscillations when using power FETs. C1, C2 and R3 comprise a decoupling filter in the bias line, while C4-C7 and RFC1 serve the same purpose in the drain supply. Capacitor values are chosen to provide effective bypassing from low frequency through vhf. C3 can be added to reduce vhf harmonic currents and aid stability. The X_c value should be four times or greater than the full-power drain impedance to prevent unwanted bypassing of the fundamental frequency. A high-mu ferrite bead in series with a low resistance (Z1 and R4) form an effective parasitic suppressor at vhf.

technique is not always practical in band-switched amplifiers, since the value of C3 might have to be changed along with the operating band.

C4, C7 and RFC1 form a decoupling network in the drain supply line. A ferrite choke of low dc resistance is suggested for RFC1. A few turns of no. 20 enameled wire on a 900-mu ferrite toroid core should suffice. Again, the capacitor values are selected to ensure good by-passing from vhf down to vlf.

The addition of negative feedback (drain to gate) will also aid stabilization. Linear power-FET amplifiers require feedback networks as part of the design procedure, but Class C amplifiers need not have feedback if stability can otherwise be achieved.

Power FETs at VHF

Impedance differences versus operating frequency are found primarily at the device input. Whereas an FET may present an almost infinite gate impedance at dc, the gate Z can approach or equal that of a BPT at vhf.[3] For operation at 150 MHz, the 45-W MRF171 FET has a large-signal input impedance of 1.94 − j4.59 ohms, as compared to the 45-W MRF315 BPT, which exhibits a characteristic of 1.2 + j1.0 ohms.

Noise figures under 3 dB are attainable with power FETs as vhf small-signal amplifiers in receiver front ends. The MRF134 5-W vhf FET exhibits a typical NF of 2 dB at 150 MHz (note 3) with a V_{DD} of 28 V and an I_D of 100 mA. Values as low as 1.5 dB have been measured. The vhf power FET is, therefore, excellent in receiver circuits that are designed for high dynamic range.

Agc for Power FETs

Another feature that sets the FET apart from BPTs is the agc capability. The output power of an FET is variable from full amplitude over a range of approximately 21 dB (note 3). This is possible without changing the amplifier excitation level. Manual or automatic gain control is effected by changing the value of gate bias. For example, an amplifier that can deliver 125 W of rf output can be adjusted to give only 1 W of output by shifting the forward gate voltage (dc) from +3 to −10. This makes the FET ideal for use in amplifiers that employ an output-level control. This feature is fine for cw and fm operation, but the gate bias must not be dropped below that value which ensures proper linearity for ssb service, lest the IMD quality be degraded.

Narrow-Band versus Broadband Operation

It is practical to operate a power FET in a narrow-band rf amplifier, just as we might utilize a triode vacuum tube in that type of circuit. The procedure is demonstrated in the transmitting chapter of the ARRL *Handbook* and in an IEEE *Preprint* paper.[4,5] The advantage in using narrow-band techniques is high efficiency. A single-ended 28-MHz Class C power FET amplifier built by the author yielded an efficiency of 85% (no gate bias applied). A Siliconix VN67AJ transistor was used.

The broadband push-pull FET amplifier of Fig. 4 exhibits an efficiency of 72.6% in Class B service (gate bias of 1 V). The efficiency of power FETs in broadband Class A linear service is typically 40-50%.

Practical 60-W Amplifier

A push-pull Class B amplifier for operation at +28 V is shown in Fig. 4. With an excitation power of only 228 mW, it is possible to obtain 60 W of out-

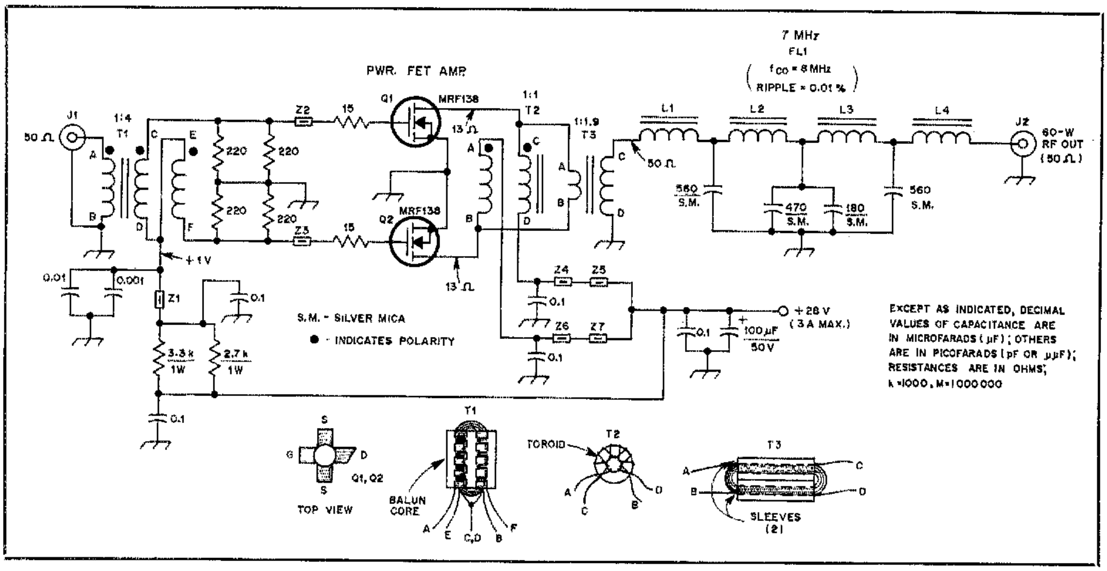


Fig. 4 — Schematic diagram of a practical power-FET amplifier that is biased for Class B service. Capacitors are disc or chip ceramic, 50 V or greater. The polarized capacitor is electrolytic (tantalum also suitable). Resistors are 1/2 W unless otherwise noted.

J1, J2 — BNC connector (builder's choice).
L1, L4 — 0.79-μH inductor. 13 turns no. 22 enam. wire on Amidon or Palomar T68-6 toroid core.
L2, L3 — 1.74-μH inductor. 19 turns no. 22 enam. wire on T68-6 toroid core.
Q1, Q2 — Motorola MRF138 MOS power FET.
T1 — 4:1 impedance ratio balun transformer (X_L = 200 ohms or greater). 10 trifilar turns of no. 28 enam. wire through Fair-Rite balun core no. 2843000302, type 1, 43 mix (950 μ_i) (available from Amidon Assoc.); or 12 trifilar turns of no. 26 enam. wire on an Amidon FT-50-43 toroid core.
T2 — Phase-reversal drain choke; 12 turns no. 22 enam. wire on two Amidon FT-50-43 toroid cores (stacked). The μ_i = 950.
T3 — Broadband conventional transformer. Primary has two turns of no. 18 plastic-insulated hookup wire. Secondary has three turns of the same type of wire. Both windings wound through the holes of two Amidon S43-621-1 ferrite sleeves. The sleeves are epoxy-cemented together, side by side (μ_i = 800). Two rows of six each T50-43 Amidon ferrite toroids can be substituted for the sleeves. Enameled wire can be used in place of the hookup wire, but will be more difficult to thread through the core material. Also suitable is the new Amidon jumbo balun core, part no. BN-43-7051.
Z1 and Z4-Z7, incl. — Jumbo 900-mu ferrite bead. Amidon FB-43-801. (See text.)
Z2, Z3 — Miniature 900-mu ferrite bead. Amidon FB-43-101. (See text.)

put after filtering. Peak drain current is 2.95 A, for an efficiency of 72.6% at 7 MHz. A gate bias of + 1 V is used. Under these conditions, the amplifier gain is 23 dB. Operation on the bands from 160 through 10 meters is possible by selecting an output filter for the desired frequency.[6]

T1, T2 and T3 are ferrite-core, broadband transformers. The gate resistors consist of four 220-ohm, 1/2-W units, arranged to present a gate-to-gate impedance of 210 ohms. This permits the use of 4:1 transformer (T1) to match the amplifier input to 50-ohm driving source. The gate resistors form part of a bias divider with the parallel pair of 1-W resistors (3.3-kΩ and 2.7-kΩ).

Parasitic suppressors are located at the gates of Q1 and Q2. They stop a strong self-oscillation that was observed at roughly 60 MHz. Stability is aided further by the decoupling networks in the bias and drain-supply lines.

T2 is a phase-reversal drain choke. The windings are returned separately to the V_{DD} supply line in order to enhance the decoupling of the transistor drains from other parts of the circuit. This procedure is recommended by ARRL TA Helge Granberg (K7ES), of Motorola.

T3 is a conventional broadband transformer that closely matches the 26-ohm drain-to-drain impedance to the 50-ohm harmonic filter, FL1. For practical reasons, it is not possible to obtain a precise match, owing to the lack of a common integer ratio. However, the 2.25:1 transformation ratio is close enough to the calculated 1.9:1 ratio. This results in a 1.5:1 turns ratio for T3 (two-turn primary and three-turn secondary).

FL1 is a 7-element Chebyshev design. It provides excellent harmonic attenuation. Harmonic energy is 70 dB or greater below peak cw output power (Fig. 5). The two-tone IMD was not measured, owing to the design being aimed purely at cw operation. However, the MRF family of MOSFETs is rated for an IMD of 35 dB or greater below peak power in Class A service (note 1).

Zero-bias operation with this amplifier was checked to determine the Class C performance. Without changing the drive level, the output power declined to 52 W and the I_D dropped to 2.75 A. A slight increase in drive level restored the output power to 60 W. The Class C efficiency was 67.5%. This indicates that the amplifier gain and efficiency are somewhat better when the FETs are biased into conduction. This is reasonable, since power FETs are enhancement-mode devices.

The question may arise, "What kind of performance can be expected when the V_{DD} is lowered to, say, 12 V?" This procedure is not recommended because the transistors will saturate rather quickly. The result is poor efficiency and greatly reduced power output. The devices should always be chosen for the operating voltage, or vice versa.

Construction

The amplifier of Fig. 4 is laid out for single-band operation. The filter is located on the circuit board, which makes band switching impractical. There is no reason why the pc board could not be

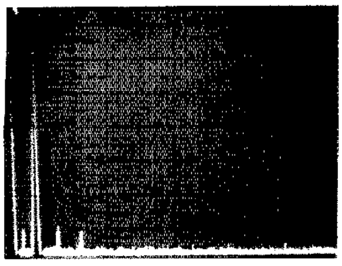


Fig. 5 — Spectral output of the 60-W FET amplifier showing the spurious energy at 70 dB below peak power output. The current FCC requirement for commercial amateur transmitters is −40 dB for all out-of-band spurious responses. This amplifier exceeds that level by some 30 dB. Vertical divisions are 10 dB; the horizontal scale is 10 MHz/div. The tall response at the far left is the spectrum analyzer reference. The full-scale response to the right of it is the 7-MHz desired signal.

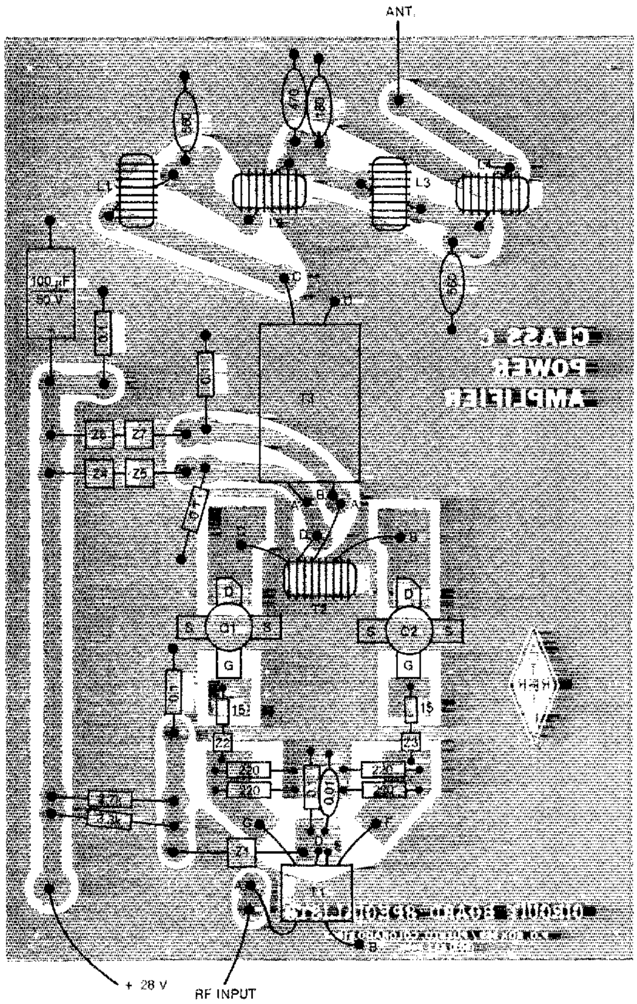


Fig. 6 — Parts-placement guide for the power-FET amplifier, shown in X-ray format from the component side of the pc board. The gray areas indicate unetched copper. Double-sided pc board material is used.

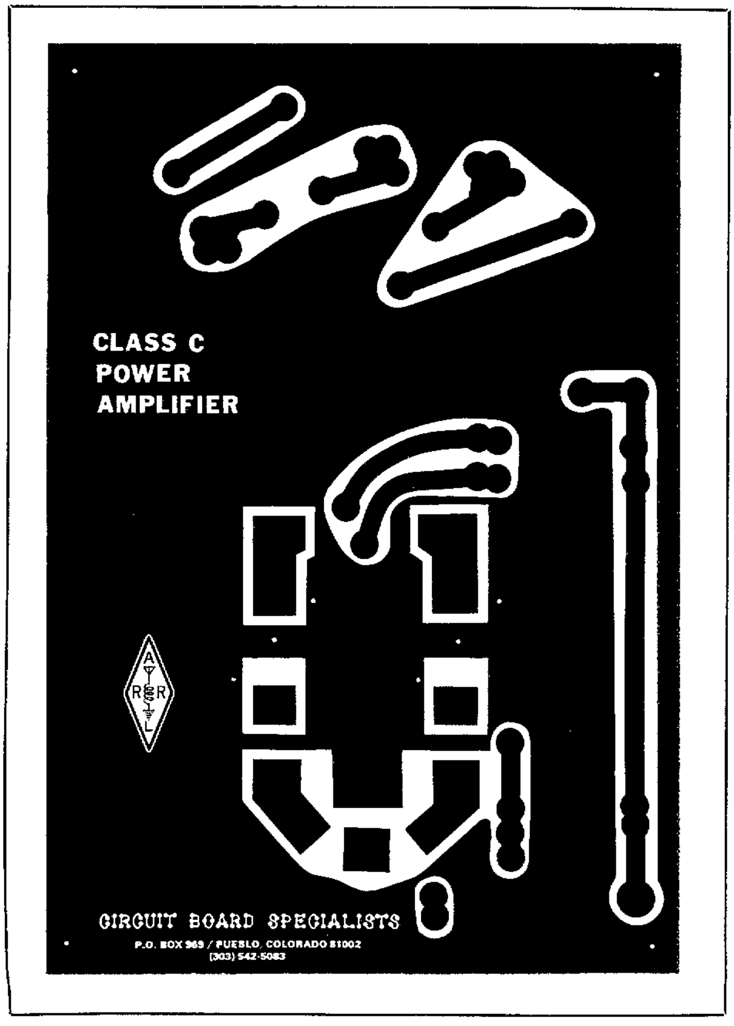
shortened to contain only the circuit up to the filter. Separate boards could then be employed to contain a bank of switchable filters.[7]

The pc board measures 6 × 4-1/4 inches.[8] It is mounted approximately 1/8 inch above the heat sink on metal standoff posts. This permits mounting the FETs with their strip-line tabs at right angles to the bodies of the transistors. This minimizes unwanted stress on the seals. *Always avoid soldering a transistor to a circuit board when the leads need to be bent up or down to mate with the pc-board pads.*

An extruded aluminum heat sink with cooling fins is suggested. It should be of the same length and width as the circuit board, or larger. If such a heat sink can't be located, one may be fashioned from heavy-gauge sheet aluminum, as illustrated in the construction-practices chapter of recent editions of the ARRL *Handbook.* The transistors and the heat sink should be no more than warm to the touch after a 3-minute key-down period at full output power. Under no conditions should the FET case temperature exceed the manufacturer's safe specifications. Be sure to use a thin layer of heat-conducting grease between the mating surfaces of the transistors and the heat sink. Tighten the transistors in place until the nuts are snug, but not super tight. Input and output coaxial connectors can be installed on aluminum L brackets at the ends of the amplifier module.

FL1 is designed for a cutoff of approximately 1.15 the highest desired operating frequency. In this example (Fig. 3), the f_{co} is 8 MHz. The center capacitance of the filter is a nonstandard value. Hence, there are two standard-value silver-mica capacitors in parallel to obtain the required value. Polystyrene capacitors may be substituted for the silver-mica units for frequencies through 14 MHz. Silver-mica capacitors should be used above 14 MHz. The leads of the capacitors need to be kept as short as practicable to minimize unwanted series inductance.

Z1 through Z7, inclusive, are formed by passing short lengths of bare bus wire through the holes of the ferrite beads. Miniature and jumbo beads are used in the amplifier.



Etched-foil side of power-FET amplifier circuit board to scale. Black areas are solid copper.

12-V Power FETs

We can obtain rf power FETs for operation at 12 V. Siliconix, Inc., has a DV1260 transistor that is rated for 60 W of output at 175 MHz. A 2-meter amplifier was built by Hq. staff member AK4L for the purpose of testing a sample of the DV1260. Operation was smooth and an output of 40 W was obtained easily. The power output would have been greater if the supply voltage had been set at 13.5.

Motorola also manufactures 12-V power FETs. The MRF128 is listed as the 12-V equivalent to the MRF138, and is rated for an output of 30 W at 30 MHz. These parts are aimed specifically at the land-mobile market. As production increases the price per unit should drop.

Notes

[1]H. Granberg, "MOSFET RF Power — An Update," *QST,* Dec. 1982 and Jan. 1983.
[2]Generally, power FETs in this class require a V_{DD} of 50 V.
[3]R. Hejhall, "VHF MOS Power Applications," *Preprint,* session 24, IEEE Midcon, Dallas, Texas, Dec. 1982.
[4]*The Radio Amateur's Handbook* (Newington: ARRL, 1980 and 1981).
[5]D. DeMaw, "Practical Class-A and Class-C Power-FET Applications at HF," *Preprint,* session 24, IEEE Midcon, Dallas, Texas, Dec. 1982.
[6]Filter tables with normalized values for various numbers of poles and ripple characteristics are available in the transmitting chapter of recent editions of *The Radio Amateur's Handbook.*
[7]Available from Circuit Board Specialists in both formats.
[8]Circuit Board Specialists, P.O. Box 969, Pueblo, CO 81002. Negatives, circuit boards and parts kit are available.

Strays

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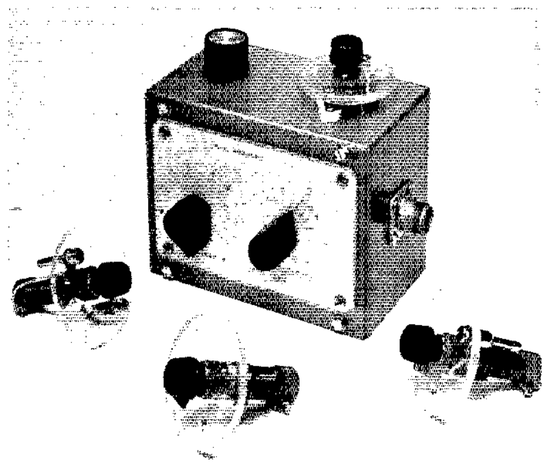
Measuring Impedance with a Reflection-Coefficient Bridge



Read the resistance and reactance components of an unknown impedance with this simple R-X adapter.

By Jack Priedigkeit,* W6ZGN

Voltage standing-wave ratio, or VSWR, is probably the most popular way of expressing the degree of impedance match between an rf transmission line and a load, such as the match between a 50-ohm coaxial cable and an antenna. However, VSWR does not provide sufficient information to design an impedance-matching network to correct mismatch. For example, a 4:1 VSWR in 50-ohm line can be caused by a resistive load of either 12.5 or 200 ohms, or, as may be seen on a Smith Chart, by a complex load impedance of many possible combinations of resistance and reactance.[1]

This article describes an R-X adapter for a reflection-coefficient bridge that permits both the resistance and reactance components of a complex impedance to be measured.[2] The measurement range of this adapter is extended, when necessary, by the addition of a known length of cable. A Smith Chart is then used to determine the impedance without the added cable.

The operational performance and accuracy of this simple R-X adapter should not be expected to be equivalent to that of more sophisticated (and expensive) test equipment, such as a vector impedance meter or a network analyzer.[3,4] Nevertheless, this adapter can provide useful information to the amateur interested interested in experimenting with antennas, transmission lines or rf circuits.

Fig. 1 shows the electrical schematic of the R-X adapter as a series circuit — tuned to resonance at the measurement frequency. In concept, the series-reactance component of the unknown impedance is canceled by adjusting C1 so that the reactance of L1 and C1 in series ($X_L - X_C$) produces a net positive, or negative, reactance that is equal and opposite to the reactance component of the unknown.

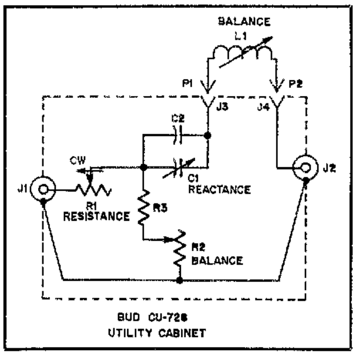


Fig. 1 — Electrical schematic for the R-X adapter.

R1 — 50 ohms, type "J" linear composition.
R2 — 1000 ohms, type "J" linear composition.
R3 — 220 ohms, 1/2-W composition.
C1 — 100-pF variable capacitor, modified (see text).
C2 — 10-pF silver mica.
L1 — Set of plug-in coils, National XR-50 form. See Table 1.
J1 — UG-1094/U or equiv.
J2 — 83-1R (SO-239) or equiv.
J3, J4 — Banana jack, H. H. Smith type 100.
P1, P2 — Banana plug, H. H. Smith type 256.

Assuming the resistance component of the unknown to be less than 50 ohms, the variable resistance R1 is adjusted so that the total resistance of the series circuit is exactly 50 ohms. This terminates the cable from the reflection-coefficient bridge — resulting in a null, or dip, in the S meter on the receiver used as the detector for the reflection-coefficient bridge.

The series resistive and reactance components of the unknown impedance ($R \pm jX$) are read from calibrated dials on R1 and C1. R2 is adjusted during an initial balance to compensate for the resistance of L1, which is in series with the unknown impedance.

It will not be possible to terminate the cable to the reflection-coefficient bridge when the resistive component of the unknown impedance is greater than 50 ohms. In this case, a known length of 50-ohm cable can be added between the adapter and the load to transform the load impedance to the measurement range of the adapter. A Smith Chart, Fig. 2, is used to find the impedance without the added cable.

Measurement Limitations of the R-X Adapter

The limits of the impedance measurement range, normalized to 50 ohms, are shown in Fig. 2. The normalized resistance range, zero to 50 ohms, is bounded by the outer circle of the chart and the $R = 1$ circle that passes through the center of the chart. The normalized reactance range is bounded by the outer circle of the chart and the $X = 1.7$ to $X = 13$ reactance circle, depending on the frequency.

Notes appear on page 32.
*441 Sherwood Way, Menlo Park, CA 94025

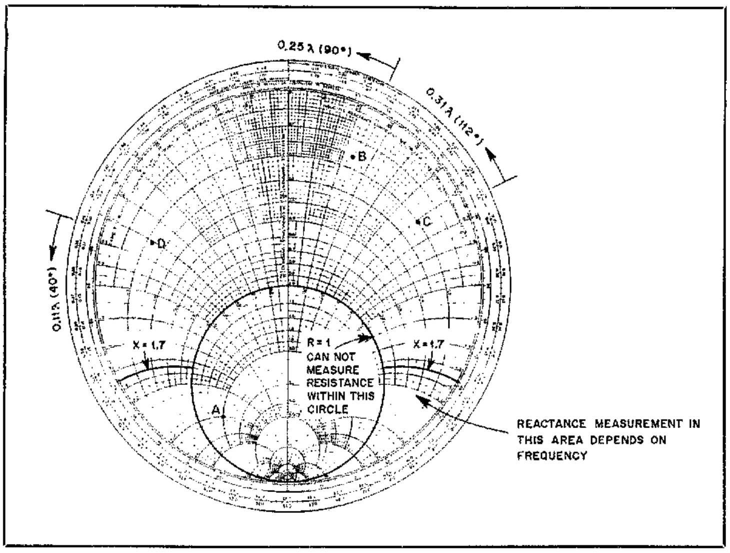


Fig. 2 — Smith Chart examples of using the R-X adapter. See text.

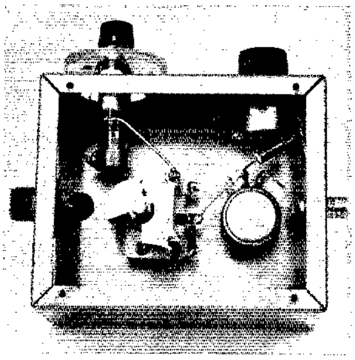


Fig. 3 — Rear view of the adapter with cover removed. Note the plates of C1 after modification, at the approximate center of the box as shown in this view.

Resistance and reactance values falling inside the R = 1 circle cannot be measured since the resistance exceeds 50 ohms. However, impedance values within this circle can be transformed, or rotated on the Smith chart, by adding a length of transmission line. For example, the addition of a 1/4-wavelength line (0.25 λ) will move the unknown impedance diametrically across the Smith Chart, as is illustrated by points A and B in Fig. 2.

Construction

The photos show the R-X adapter constructed by the author. A 3 × 4 × 5-inch (mm = 25.4 × in.) Bud model CU-728 utility cabinet is used. A BNC jack (UG-1094 or equiv.) is used for the cable connecting the adapter to the reflection-coefficient bridge. A type 83-R (SO-239) jack is used for the unknown impedance connector.

The component layout is not critical. However, type "J" molded composition variable resistors should be used for R1 and R2. Further, L1 and C1 must be insulated from the metal box. Point-to-point wiring with no. 16, or larger, solid wire is recommended. The rotor of C1 is connected to R1 and is isolated from the reactance dial with an insulated shaft extension to minimize the effects of hand capacitance.

C1 is a linear 10- to 100-pF capacitor that was modified by shaping the rotor plates on a belt sander. The modification decreases the rate of change of capacitance near minimum setting (see Fig. 3). This modification, together with the addition of the 10-pF fixed capacitor, C2, results in a nearly linear reactance range of approximately ±250 ohms at 10 MHz.

L1 consists of a set of five plug-in coils for the 80, 40, 30, 20 and 10-meter bands. These coils are each wound on a National XR-50 coil form, or equivalent, and plug into a hole cut in the side of the Bud box. Each coil is mounted on a 2-inch-diameter Lucite disc with two banana plugs (H. H. Smith type 100). The banana plugs mate with two plastic-insulated banana jacks (H.H. Smith type 256) mounted on the box. A small knob, for a 1/8-inch shaft, is installed on the slug adjusting screw for ease of tuning the coil to the measurement frequency. The winding details for the set of five coils are shown in Table 1.

Calibration

The resistance and reactance dials may be calibrated to be read in ohms or, as this author prefers, in values normalized to 50 ohms for use with the Smith Chart. The adapter is calibrated at 10 MHz by connecting known resistors, capacitors or inductors to the unknown connector, J2. The capacitance of the type 83-R connector used for J2 is approximately 5 pF, and this must be considered as it is in parallel with these discrete components used for calibration. The capacitance of J2 need not be considered when a 50-ohm coaxial cable is connected to J2 since, to the extent that the 83-R connector has a 50-ohm characteristic impedance, J2 is simply a very short extension to the length of the cable.

The calibration of the resistance dial changes slightly from band to band (see Fig. 4) as the rf resistance of the five plug-in coils differ. For most amateur work, however, it can be assumed that the resistance dial is independent of frequency.

The calibration of the reactance dial is frequency-dependent. The reactance at any frequency other than 10 MHz is

$$X = (\text{dial reading}) \times 10.0/f \qquad \text{(Eq. 1)}$$

where

f is the measurement frequency in MHz.

The actual capacitance and inductance values used to calibrate the reactance dial in 50-ohm increments are shown in Table 2. These values are corrected for the capacitance of J2.

Measuring A Complex Impedance

Initial balance: Fig. 5 shows connections for taking various measurements. (See earlier *QST* articles.[5,6]) Connect the R-X adapter to a reflection-coefficient bridge and tune the signal generator and bridge receiver to the frequency at which the measurement is to made. Plug in the appropriate coil for L1. Set the resistance dial to 1 (50 ohms) and the reactance dial to zero. Terminate J2 with a 50-ohm carbon-composition resistor and adjust L1 for resonance, indicated by a dip of the bridge receiver S meter. Alternately adjust

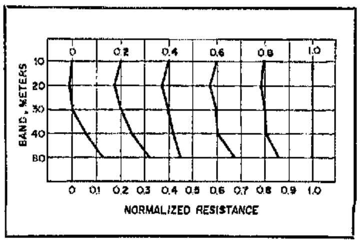


Fig. 4 — Resistance calibration vs. frequency.

L1 and R2 for initial balance — the lowest possible S meter reading.

Measuring an unknown within the range of the R-X adapter: Connect the impedance to be measured to J2. Adjust the resistance dial, R1, and the reactance dial, C1, for minimum reading on the S meter. *Do not* readjust the initial balance controls, L1 and R2.

The resistance and reactance components are indicated on the dials of R1 and C1. The reactance value must be corrected for frequency using the relationship:

$$X = (\text{dial reading}) \times 10.0/f \qquad \text{(Eq. 2)}$$

where f is the measurement frequency in MHz.

Measuring an unknown outside the range of the R-X adapter: It will not be possible to obtain a sharp, deep null on the S meter for impedance values outside the range of the adapter. In fact, for some impedance values, the S meter will change very little, or not at all, as the resistance and reactance dials are adjusted. In this case, add a known length of cable between the adapter and the unknown, and proceed as in the previous paragraph.

The electrical length of the added transmission line, in wavelengths, is calculated with the formula:[7]

$$N = Lf/984k \qquad \text{(Eq. 3)}$$

where

L = the physical length in feet
f = the frequency in MHz
k = the velocity factor for the particular cable type used

Some examples: Assume the normalized resistance and reactance dial readings for an unknown impedance measured at 7.2 MHz using a 1/4-wavelength line are 0.16 + j0.17. Correcting the reactance component for frequency, this impedance is 0.16 + j0.17 × (10.0/7.2), or 0.16 + j0.24 — point B on Fig. 2. Since a 1/4-wavelength line (0.25 λ) was used, the normalized impedance without the added line is located diametrically opposite point B, or point A on the Smith Chart of Fig. 2. The normalized impedance of the unknown without the added line is 2.0 − j3.0. This impedance in ohms is 50 × (2.0 − j3.0), or 100 − j150.

Had a 0.31-λ line been used in the above example, the normalized impedance shown by the dials on the R-X adapter would be 0.2 + j0.45. Correcting for frequency, this impedance is 0.2 + j0.63 — point C on Fig. 2. In this case, it is necessary to rotate point C through 112 electrical degrees (360 × 0.31) toward the load on the Smith Chart to find the impedance without the added line. This will move point C to point A on Fig. 2.

Similarly, if a 0.11-λ line had been used, the R-X adapter would show the normalized impedance to be 0.23 − j0.52. Correcting for frequency, this becomes 0.23 − j0.72 (point D on Fig. 2). The impedance without the added line is found by rotating point D through 40 electrical degrees toward the load, which again falls at point A of Fig. 2.

Table 1

Reactance Range and Coil Details

Band (Meters)	Reactance (Ohms)	Reactance (Normalized)	No. Turns	Wire Size	Inductance (μH)*
80	±670	±13	65	30	25-44
40	±350	±7	27	30	8.7-15
30	±250	±5	20	22	3.8-6.5
20	±175	±3.5	14	22	2.3-3.9
10	±90	±1.8	8	18	0.6-0.9

*Using a National XR-50 coil form or equiv.

Table 2

Inductance and Capacitance for Calibrating Reactance Dial

Reactance (Ohms)	Reactance (Normalized)	+jX Inductance (μH)	−jX Capacitance (pF)
50	1	0.78	313
100	2	1.5	154
150	3	2.3	101
200	4	3.0	75
250	5	3.7	59
300	6	4.4	48

These inductance and capacitance values include the correction for approximately 5 pF of capacitance in the "unknown" connector, J2.

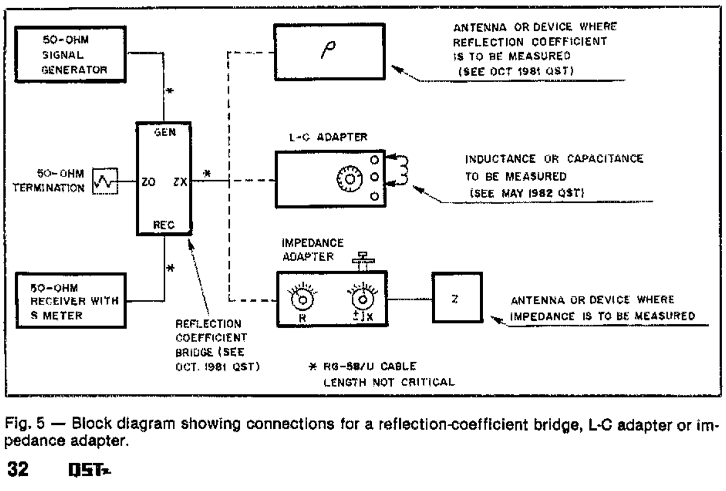


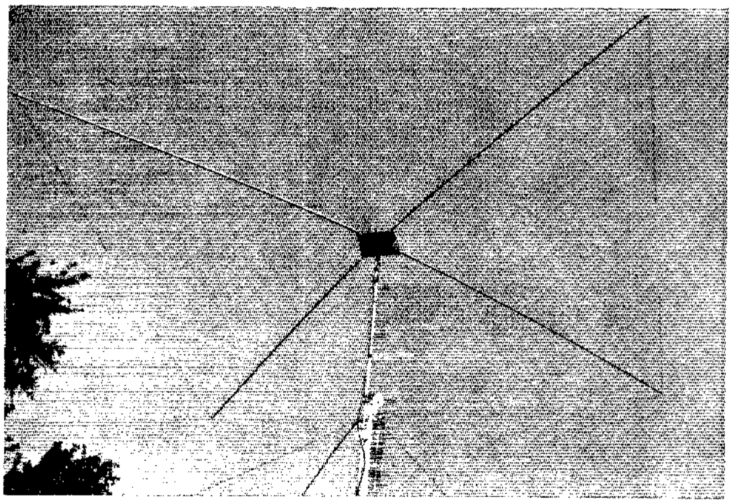
Fig. 5 — Block diagram showing connections for a reflection-coefficient bridge, L-C adapter or impedance adapter.

Jack Priedigkeit received his BSEE degree from the University of California in 1942. He was licensed as a radio amateur in 1947, and received his First Class Radiotelephone license in 1940. A registered electrical engineer in the state of California and a Senior Member of the IEEE, he retired in 1980 after more than 40 years of experience. He holds several patents related to research and development in the areas of instrumentation, antennas, propagation, avionics, communication and navigation systems, and position location.

Notes

[1] *The ARRL Antenna Book* (Newington: ARRL, Inc.; 1974 [13th ed.] or 1982 [14th ed.]); Chapter 3, section titled "Smith-Chart Transmission-Line Calculations."
[2] J. Priedigkeit, "A Reflection-Coefficient Bridge — Impedance-Matching Measurements the Easy Way," *QST*, Oct. 1981, pp. 18-20.
[3] Hewlett-Packard Model 4815A Vector Impedance Meter.
[4] Hewlett-Packard Model 8407 Network Analyzer.
[5] See note 2.
[6] J. Priedigkeit, "Measuring Inductance and Capacitance with a Reflection-Coefficient Bridge," *QST*, May 1982, pp. 28-29.
[7] See note 1.

Horizontal X Beams for 15 and 20 Meters



Try this design from "across the pond" and enjoy BIG GUN performance from a small package.

By Brice Anderson,* W9PNE

Are you a serious 10, 15 or 20-meter operator who lacks a beam antenna? How would you like to build a compact, lightweight beam that offers good performance and a low SWR, band edge to band edge? This description fits the horizontal X-beam antenna. I have built "X" beams for 15 and 20 meters, and find their performance impressive!

The X-Beam Antenna

An on-the-air acquaintance, Bob Norwood, W6FWL, built an X antenna and then sent me an article from *Break-In*, journal of the New Zealand Association of Radio Transmitters, by J. F. Harper, ZL2NH, that describes it. The article covers the theory and construction of a compact 2-element beam antenna using aluminum-tubing arms and wire loading "tails." The X beam is derived from a 2-element Yagi design in which the parasitic element is self-resonant and serves as a director. The design apparently originated in England, where G4ZU had built a number of antennas with the "X" format. The concept traveled to Australia and to New Zealand, where VK4RF and ZL2NH built and used it with enthusiasm.

Fig. 1A shows a typical 2-element Yagi antenna. Suppose the centers of each element were pulled inward until the arms were at right angles to each other. This would form an X-beam antenna, which is shown in Fig. 1B.

This X design eliminates the need for a boom, but the overall physical size is not reduced significantly because of the long element lengths. By shortening the elements, the physical size becomes manageable. Resonance is restored by adding wire extensions to the element ends. This is shown in Fig. 1C. The effective spacing between the elements is 0.05 and 0.10 wavelength. For 20 meters, the element arms may be as short as 12 feet.[1] I felt that the antenna might have greater bandwidth, however, if the elements were a bit longer and the wire tails shorter. I used arms 13 ft 9 in. long — a convenient dimension for the tubing I had on hand.

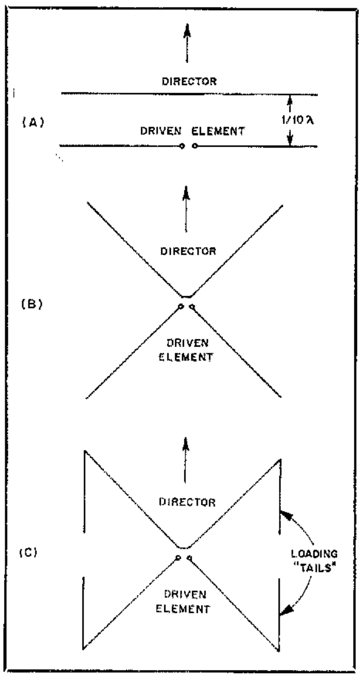


Fig. 1 — Derivation of the X beam from the Yagi design. See text for details.

Construction of a 20-Meter X

A trip to the lumber yard netted the necessary aluminum tubing and associated parts. I bought two 8-foot pieces of tubing in the following sizes: 1-inch OD, 7/8-inch OD and 3/4-inch OD. I also purchased 1 foot of 1-inch ID heavy-wall, clear-plastic hose; eight 1-1/4 inch pipe brackets; a 15-inch-square piece of 5/8-inch plywood; a 30-inch length of 2 × 4; a supply of lag screws; and some 1/4-inch stove bolts and nuts. The total cost was less than $30.

The ZL2NH design called for short pieces of angle iron, welded in the form of an X, with a stub mast welded on, to serve as the hub of the beam. Plywood is used as a hub in my design, with two 15-inch lengths of 2 × 4 wood bolted to the underside, spaced 2 inches apart. I used lag screws to fasten a 2-1/2 foot mast of 2- × 2-inch wood to the 2 × 4s. Fig. 2A shows the hub assembly. This arrangement gives me the option of vertical or horizontal polarization.

Assembly

Each piece of tubing is sawed in half, deburred inside and out and lightly polished with emery paper until the sections telescope together. Old TV-antenna elements can be used for the 5/8-inch-OD end sections on the arms.

The 4-foot pieces of 1-inch tubing are mounted on the plywood square in the shape of an X, with a separation of 3

*P.O. Box 14, Lancaster, IL 62855

[1] m = ft × 0.3048; mm = in. × 25.4.

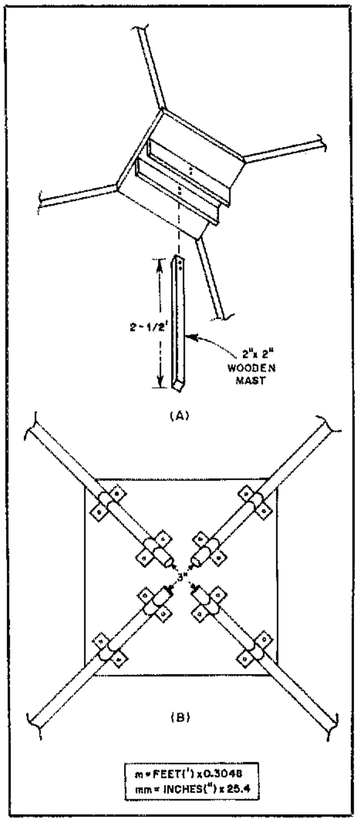


Fig. 2 — Details of the hub assembly. A shows the bottom view, and B details the top assembly.

inches between tubing ends at the center. Insulation for the tubing is provided by 1-1/2 inch pieces of plastic tubing, which slip over the aluminum tubing. Stove bolts hold the brackets to the plywood. This provides a very strong and well-insulated mounting as shown in Fig. 2B.

Insert the 7/8-inch tubing pieces 2-1/2 inches into the 1-inch tubing and secure with two no. 6 self-tapping screws. Likewise, the 3/4-inch tubing is fastened to the 7/8-inch tubing. Four pieces of 5/8-inch tubing, each 2 feet 4-1/4 inches, are fastened to the 3/4-inch tubing in the same way. The total length of each arm is 13 feet 9 inches. See Fig. 3 for details on the arm assembly.

An SO-239 type coaxial connector is mounted to the hub by means of short standoffs. Solder lugs are fastened to the ends of each arm, and to the inner ends of each arm at the hub, with no. 6 self-tapping screws. A piece of no. 12 wire connects the director arms together, with the wire bent into a 3/4 turn to aid in coupling during tuning. The coaxial connector is wired directly to the driven element with no. 12 wire.

Tail-wire length will vary with different

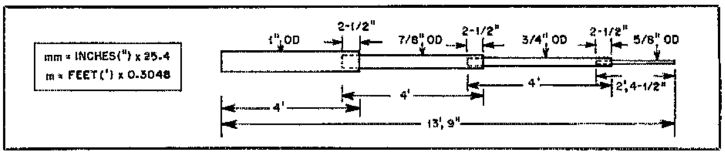


Fig. 3 — Details of the antenna-arm construction. No. 6 machine screws are used to hold the tubing together.

wire sizes. This length must be determined experimentally. Nylon cord runs between the ends of the arms to reinforce the assembly and to serve as a support for the tails. After finishing the mechanical construction, give the hub assembly several coats of aluminum paint. Be sure to mask the plastic insulators and coaxial connector with tape to keep them free of paint. The hardware is also sprayed with aluminum paint. The completed antenna is light enough to be carried up a ladder in one hand.

Tuning and Checking

It is important that the tails be longer than necessary at the start, and that each driven-element tail be 12 inches longer than each director tail. All four tails are trimmed 1 inch at a time until the director is resonant at the desired center frequency. This may be checked by mounting the antenna on a 10-foot mast, placing a 50-ohm load across the driven element feed point and checking the director resonance with a dip meter. After tuning the director, remove the 50-ohm load and trim the two driven-element tails until resonance is indicated on the dip meter. For the best frequency coverage, the antenna should be tuned for 14.100 MHz. At this point, the tail-wire lengths should be 7 feet 8 inches for the driven element and 6 feet 8 inches for the director. Fold back and tape the tail-wire ends to prevent ionizing effect of a sharp point.

Results

After tuning my antenna, I connected a Ten-Tec Argonaut 515® to the antenna and checked the SWR. The readings were

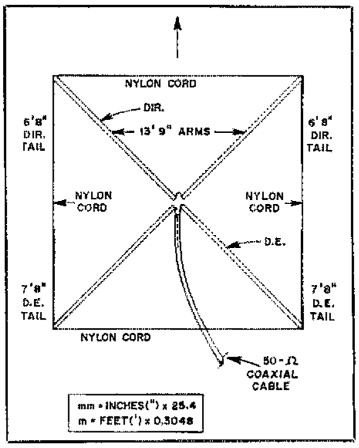


Fig. 4 — Layout of the 20-meter X. The nylon cord is used to hold the wire tails taut.

so low that another bridge was used to confirm the results. The antenna was then raised to its final height of 30 feet. As expected, the resonant frequency increased, but the SWR remained low across the band. The SWR curves for the X antennas are shown in Fig. 5.

On the Air

Well, does it work? As ZL2NH says, "She works, mate." I had three other antennas to compare with my X beam: A horizontal V with 160-foot legs, 50 feet high, a four-band trap vertical on top of a 20-foot pipe; and a dipole broadside to Europe, up 30 feet.

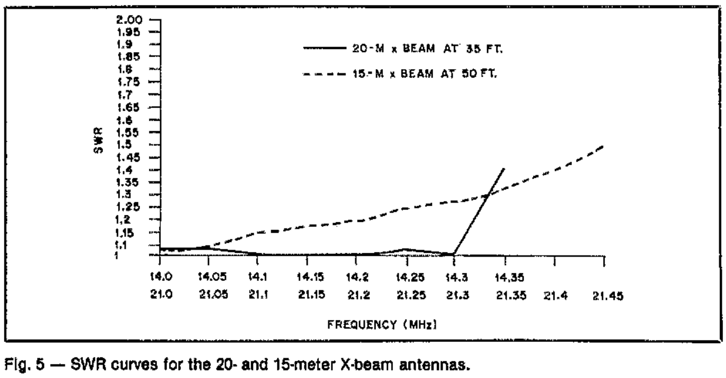


Fig. 5 — SWR curves for the 20- and 15-meter X-beam antennas.

Input power for my on-the-air tests was 5 W; there was no difficulty in raising DX stations with the X beam. I received many encouraging signal-level comparisons from U.S. and DX stations. In every case the X Beam produced the best signal reports, even though the inverted V was 15 feet higher. In general, the X Beam was 1 to 2 S units stronger than the V, and was always stronger than the trap vertical or dipole antennas.

A 15-Meter X

I decided to scale the 20-meter X to 15 meters and erect it next to a vertically polarized 4-element, 20-meter Yagi on my 50-foot tower. The 20-meter X was replaced by a 5-element commercial "tribander." Because of the size and weight of the tribander, it was raised to a height of only 30 feet.

Mechanical Changes

To change the 20-meter X to 15 meters, the arms are shortened to 9 feet 3 inches. Apply the same tuning procedure as before. The director tails should measure 4 feet 2 inches, and the driven-element tails 4 feet 10 inches. Prune the tails until resonance is indicated at 21.110 MHz. The SWR should be low across the band.

15-Meter X Performance

The performance of the 15-meter X beam at 50 feet has been spectacular! In normal operation, DX stations usually can be raised on the first call. During the CQ WW cw contest in November 1981, I ran the Argonaut 515 at an output of 4 W. During a 70-minute operating period, 23 stations were worked. Fourteen were raised on the first call and five on the second call. The X again proved its worth in the 1982 ARRL DX contests in the QRP division, resulting in 70 QSOs in 30 countries in the cw section and 53 QSOs in 30 countries in the ssb section.

A comparison of the performance of the X-beam antenna and the commercial tribander was made. Of course, the low height of the tribander (30 feet) was a disadvantage. The two antennas were equally good for medium distances, with the tribander a little better for short skip. However, on the long hauls, especially when the band was just opening or closing, the high X beam was superior. Transmitted and received signal-strength reports were at least 1 S unit better. Contacts made with the X beam were impossible to achieve with the tribander.

Conclusions

Since this article was written, I have found that the addition of a coaxial balun, mounted at the antenna feed point, will improve the azimuthal pattern considerably. Also, the director does not have to be insulated from the hub; I removed the element insulators and performance remained identical.

The X beam has a lot going for it. I wish the design were my invention! I am looking forward to building an all-metal "plumber's delight" version for 10 meters using inexpensive aluminum clothesline props and aluminum angle stock. I also hope to try a 40-meter version in which the arms would be only 25 feet long.

New Books

☐ *Three New Directories for Amateurs.* Ever wondered how many hams there are in your town? Unless you know everyone in town, you would probably be surprised at just how many hams there are. Or, how about hams with the same name as yours? There are many uses for this information, not the least of which is to build a list of other hams who might want to start a club.

One of three new directories, the *Amateur Radio Call Directory,* can make an otherwise impossible task easy. Every U.S. amateur is listed alphabetically by state, town, street address and call. It's interesting to pick a town, any town, and observe the number of hams who live there. For example, opening the book completely at random, and placing my finger blindly on a page produced the information that there are 55 hams in Cinnaminson, New Jersey, and 76 hams in Conyers, Georgia. Of the approximately 80 hams shown in Newington, including many staff members who use Hq. as their FCC address, more than 40 were unknown to me. Our club secretary will be contacting them about coming around to club meetings.

The second directory is more of a curiosity at the moment, though it will no doubt prove of value in the future. It indexes, by last name, every ham in the country. There are, for example, only two Auricks: my son in Pennsylvania and myself. Information is listed by first name, initial call and state. There are more than 28 columns of Smiths, and only one listing for Sneary. It's fascinating to see the diversity of names, and the duplication of middle initials of hams who are undoubtedly never had the faintest idea that there was another ham with exactly the same name.

The third directory is similar to the listing we are all familiar with one important exception: There is no break between the name of the community and the state and ZIP code. It appears to make for easier reading. Everyone is listed by call district, name and QTH.

The three directories are available from Buckmaster Publishing, 70 Florida Hill Rd., Ridgefield, CT 06877. Prices: *Amateur Radio Call Directory* (by districts and by call signs), $12.95; *Geographical Index,* $25; *Name Index,* $25. — *Lee Aurick, W1SE*

☐ *The 10 Meter FM Handbook,* by Bob Heil, K9EID. Published by Melco Publishing, P.O. Box 26, Marissa, IL 62257. First edition, 1980. Soft-bound volume, 6 × 9 inches, 80 pp., $4.95.

Many hams seem to be interested in 10-meter fm so it would appear that *The 10 Meter Handbook* might go a long way toward answering many questions. There have been articles about converting CB transceivers to operate on the 10-m band, and that seems to be the main idea behind this book. Almost half the book is devoted to describing the advantages of 10-m fm operations, what is involved, some of the problems and other introductory information.

The later sections are used to describe the steps in converting a surplus Cybernet (HyGain) CB transceiver board so that it can operate on 10-m fm. I was looking forward to the technical details, but was greatly disappointed by the coverage provided. The descriptions of the conversion steps are poorly written, and a great deal is left to your imagination.

For example, the first step appears to be a checkout of a surplus board on the citizens' band to be sure it is operating. This is a good idea, since you don't want to "convert" a defective board. However, here's a typical description: "All check out is done with frequency programming on channel 1, if it is a 23-channel unit or channel 20 if it is a 40-channel rig. This will become 29.5 when conversion is finished." This left me a bit confused, since there was no mention of *how* the unit was supposed to be "programmed" for these channels. Remember, you're supposed to be working with an open pc board, with no external controls. Other steps assume that you know exactly what parts are needed, and that you can identify solder connections on the pc board without a diagram. Eight pages later, a wiring diagram is shown, but it is dually labeled as the MELCO FM I.F. BOARD and the Cybernet board. It is also missing the resistance values for the volume and squelch controls and the current range of the meter.

Since the CB transceiver doesn't have the ability to process fm, an fm detector must be added. A detector circuit is shown, but component values are missing from both the diagram and the text. Likewise, the schematic diagram shows an LM3065 fm detector IC, while all the descriptions mention a 1358P IC. Over 20 pages later, the chips are mentioned, along with others, in a two-page chapter about fm detection systems. The two chips appear to be functional equivalents, but not pin-for-pin replacements for each other.

Diagrams and figures are hand-drawn, and the lack of complete information about parts and connections makes much of the information less than useful. The book suffers from lack of detail and from poor organization. It is difficult to see how this 80-page book can be called a handbook.

Since many of the needed parts, as well as the main CB transceiver board, are available from the publisher, the book can be best viewed as a "sell job" for their kits and assemblies. You shouldn't have to spend $4.95 for a short, incomplete introduction to 10-m fm. The book is also billed as a "must" for builders of the Melco kits. I hope the builders don't have to rely on the information in the book to assemble their systems. — *Jon Titus, KA4QVK*

The Two-Band Delta-Loop Antenna

An interesting antenna, indeed! But you may be more interested in the electronic switching for a remote impedance-matching network.

By Richard O. Gray,* W9JJV

This antenna provides excellent performance on the 40- and 80-meter bands. It requires a limited amount of space, and includes a pretuned, remotely located antenna-matching network. The design illustrates the use of resonant circuits instead of switches to select the proper matching circuit for each band. You can build a similar antenna, with variations in loop shape or with different impedance-matching circuits, if you follow the basic philosophy that I did.

Fig. 1 shows the arrangement I used in developing the matching networks. The open-wire feeder was connected to an 18-inch length of RG-8/U cable at the outside wall of my house.[1] This short length of coaxial cable served as a feed-through line, so the experimental work could be carried out in the comfort of my home.

Only two pieces of test equipment — an rf voltmeter and an SWR indicator — were used to adjust the matching network. An exciter with a 50-ohm output impedance was adjusted to produce just enough power for a full-scale reading on the SWR meter. The rf voltmeter was connected between the inner conductor of the short coaxial cable and ground. The rf voltage at this point can be 1000 or greater when operating at full power, so be careful! Alternatively, a field-strength meter can be loosely coupled at this point. Short pieces of coaxial cable connect the transmitter to the SWR meter, and the meter to the matching network.

The total length of wire in the triangular loop is 140 feet, which is close to being a full wavelength on 40 meters and a half wavelength on 80 meters. I am sure the loop is affected by the close proximity of the vertical portion to the steel tower leg.

The apex of my loop is mounted by means of an insulator tied to a nylon rope through a pulley at the top of my tower. The rope is a continuous loop from the ground through the pulley and back to the ground. Should the wires break, or I have to lower the antenna for any reason, I won't have to climb the tower.

My matching network was determined empirically by trying parallel and series tuned circuits, pi networks and various other circuits. You may find it easier to discover the best network for your particular case by using a commercial antenna-matching network that has most of the common combinations included. When a proper impedance match is obtained, you can build a pretuned network using the appropriate components and values.

All of the parts used in the experimental work came from my junk box. Inductors are wound on various tubing sizes that I had available. All of the wire for the loop, open-wire line and coils is no. 16 copper conductor.

To adjust the experimental matching network, I tuned the capacitors and changed the inductor tap points while watching the SWR meter and the rf voltmeter. Look for a decrease in SWR and an increase in the rf voltage. Patience and practice will lead to a low SWR and a fairly high rf voltage at your favorite operating frequency.

The End Result

In the final configuration, I moved the matching network to the outside wall of my house. Fig. 2 shows the final matching-network circuit. Since most of the components are above ground potential, I used a 12- × 12-inch piece of 1/4-inch-thick Lucite® as a base. The enclosure is made of 0.018-inch-thick aluminum gutter flashing, available from most hardware stores. I no longer needed the 18-inch piece of RG-8/U to feed through the house wall, but it is still a part of the circuit capacitance. I connected one end of the center conductor to the network output, and the shield braid to ground. The other end of the cable is wrapped with electrical tape to prevent shorting to any other part of the circuit. This is C5 in Fig. 2.

Next, I simplified the operation by eliminating the three switches. I did this by using tuned circuits to switch the signal

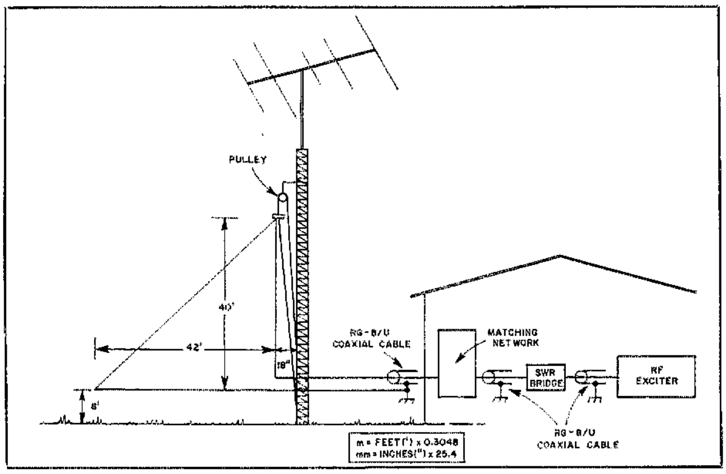


Fig. 1 — Construction details for the two-band Delta Loop. Notice that the matching network was located inside the house during the design stage. It was moved outside for the final arrangement.

[1]mm = in. × 25.4; m = feet × 0.3048

*3625 Saratoga St., Downers Grove, IL 60515

paths automatically. I determined that the impedance of L1′ and C1 is negligible on 80 meters when compared to that of L2′ and C2. S1 can simply be eliminated. The S2 mode must present a high impedance on 80 meters and be near zero on 40 meters, so I replaced the switch and its function with a series-resonant circuit tuned to 40 meters. This circuit presents a capacitive reactance on 80 meters and reduces the required value of C2. The S3 mode should provide a low impedance on 40 meters and a high impedance on 80 meters. S3 can also be replaced by a series-resonant circuit that is tuned to 40 meters. The refined circuit is shown in Fig. 3. Fig. 4 shows the internal layout with the aluminum cover removed from the box.

Circuit Analysis

On 40 meters, the resonant impedance of L4 and C4 is about 3 ohms, assuming a conservative Q value of 20. This is considerably less than 50 ohms, and so meets my requirements. On 80 meters, the impedance is about 70 ohms, which is not considerably greater than 50 ohms. This impedance is a capacitive reactance, however, and can be tuned out by the adjustment of C2.

The resonant impedance of L3 and C3 on 40 meters is approximately 30 ohms, again assuming a Q of 20. This is much lower than the anticipated impedance of the antenna system. On 80 meters, the impedance is equivalent to a 60-pF capacitor, which is in parallel with C2. The capacitance of C2 will have to be reduced accordingly.

Final Adjustment

C1 and C3 should be adjusted for the lowest attainable SWR on 40 meters. Next, adjust C2 for a low SWR on 80 meters. Any change in the setting of C3 on 40 meters will affect the setting of C2 on 80 meters. As a final step, switch between the 80- and 40-meter bands to be sure the performance is satisfactory on both.

The SWR curves for my system are shown in Fig. 5. If the frequency of lowest SWR is chosen properly, you should have an SWR of 2.5 or less across the entire 40-meter band, and for a 300-kHz segment of the 80-meter band.

Conclusions

My station is equipped with a 50-foot tower and is a triband Yagi antenna. The tower is grounded and is tuned for 40- and 80-meter operation by means of a gamma match. Tuning and switching are done remotely from the operating position. I used this as a vertical antenna for comparison with my Delta Loop.

The most noticeable difference between the antennas is in reception on 80 meters. The atmospheric and man-made local noise can be as high as S5 or more when using the vertical antenna. At times, I know there is a signal present, but I can't copy it through the noise. Upon changing to the Delta Loop, the station is "solid copy." This is what I was looking for.

I expected the loop to exhibit directional characteristics, with the maximum propagation being perpendicular to the plane of the loop. This was not borne out in the limited number of tests I conducted. The directivity might be affected by the tower, which is only 18 inches from the vertical side of the loop. It could also be affected by the aluminum siding on my house (10 feet away) or by the forest on the other side of the tower.

The receive sensitivity is about the same as with the vertical antenna. The Delta Loop may be 1/2 S unit better. On transmit, the loop results in better reports by an average of 1 S unit.

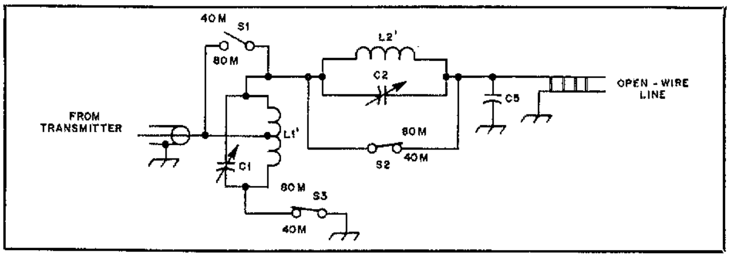


Fig. 2 — Schematic diagram of the matching network. Switches S1, S2 and S3 are shown in the 40-meter position.

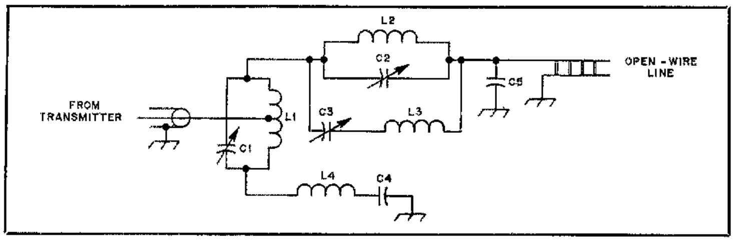


Fig. 3 — Schematic diagram of the matching network after the switches have been replaced with series-resonant circuits.

C1 — 10 to 80 pF, with 0.035-inch spacing between plates.
C2 — 21 to 100 pF, with 0.080-inch spacing.
C3 — 8 to 50 pF, with 0.080-inch spacing.
C4 — 400 pF, 1000 V.
L1 — 18 t, center tapped and tapped up two turns from bottom. The coil is wound on a 1-3/4 inch diameter form, and is 2 inches long (approx. 6.3 μH).
L2 — 19 t, tapped up five turns from input end. The coil is wound on a 1-3/4 inch diameter form, and is 2 inches long (approx 5.5 μH).
L3 — 20 t on a 1-7/8 inch diameter form, 1-5/8 inches long (approx. 12.0 μH).
L4 — 7 t on a 1-1/16 inch diameter form, 5/8-inch long (approx. 1.35μH).
All coils were wound with no. 16 copper wire.

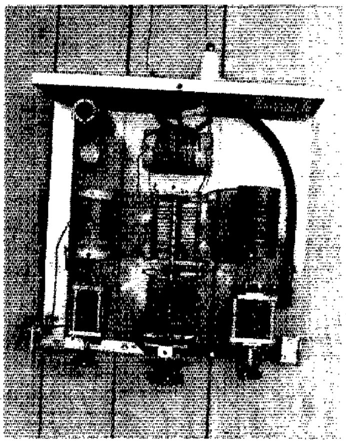


Fig. 4 — Construction details of the automatically switched matching network for the two-band Delta-Loop antenna.

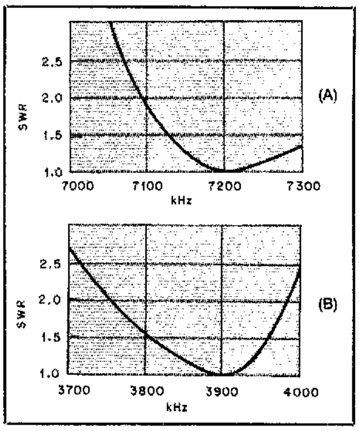


Fig. 5 — SWR curves for the W9JJV two-band Delta-Loop antenna. The 40-meter curve is shown at A, and the 80-meter curve is shown at B.

Some Aspects of the Balun Problem

Why all the mystery surrounding baluns? Here's some straight talk to dispel the rumors!

By Walter Maxwell,* W2DU, ARRL TA

The balun — to use, or not to use — is one of today's hottest topics in Amateur Radio. Because certain aspects of the connection between a coaxial feed line and a balanced antenna have been ignored, misunderstanding still exists concerning the function of baluns. Many commercial baluns embody some form of impedance transformer, promoting our tendency to misconstrue them as little more than a matching device, while their *primary* function is to provide proper current paths between balanced and unbalanced configurations.

To help clarify the misunderstanding, I will explain some of the undesirable effects that occur when a balun is not used, and some that occur when using baluns employing coupling transformers. (In many cases, these effects cause significant errors in measurements of antenna impedance and SWR.) I will also describe a simple and inexpensive method of loading the outside of a coaxial feed line with ferrite, which effectively produces a well-balanced, wide-band choke balun. Because this configuration eliminates the coupling transformer (with inherent impedance-transfer ratio errors), the accuracy obtainable in antenna impedance and SWR measurements is greatly improved. In addition, antenna-matching networks may be used with this choke balun, because no mismatch limits are imposed.

Transformer Accuracy

Using precision impedance bridges, (General Radio GR-1606-A and the Boonton 250A RX Meter), I have made measurements of transformer-type baluns that prove with a 50-ohm resistive load, the transformers in typical 1:1 or 4:1 baluns do not yield a true 1:1 or 4:1 impedance transfer ratio between input and output. This is because of losses, leakage reactance and less-than-optimum coupling; my findings have been substantiated by the work of John Nagle, K4KJ.[1] Furthermore, when we move away from the resonant frequency, the antenna becomes reactive and the impedance-transfer ratio degrades even further. This degradation of impedance transfer associated with baluns poses no serious operational problems. However, SWR curves plotted of an antenna using such a balun differ significantly from those plotted using a balun having no impedance-transfer error. Thus, when a precision bridge is used to measure antenna impedance (R + jX), the data will be erroneous with either a transformer-type balun in the circuit, or with no balun at all.

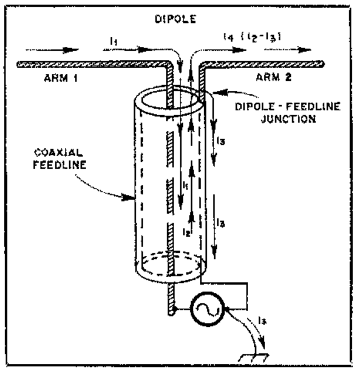


Fig. 1 — Illustration of the various current paths at a dipole feedpoint.

Should SWR Change with Line Length?

We know that the feed-line input impedance changes with line length when the load (antenna) is not matched to the line. Sometimes, trimming the length of our feed line helps to obtain a load impedance better suited to match the transmitter. Theoretically, SWR should not change with line length — except for a barely perceptible change because of line attenuation. Then why does the SWR sometimes change? If the SWR changes significantly with line length, it must mean that the load impedance is also changing. The *load* impedance changes with line length? Yes. If a balun is omitted when you feed a balanced antenna with coaxial cable, the load impedance will change, as will the SWR! To explain this, we must investigate how current flows in an antenna system.

Examination of Current Flow

To understand the functions of a balun, it is essential to be familiar with current paths at the dipole feed point. This is shown in Fig. 1. Because of their symmetrical relationship, the dipole arms couple energy of equal magnitude and opposite phase to the feed line, thus cancelling induced current flow on the outside of the feed line.[2] What is disturbing is the discovery that there are *three* paths for current flow in a coaxial feed line, instead of only two. How can there be three current paths in only two conductors? At rf, skin effect "divides" current between the inner and outer surfaces of the coaxial shield. This effect, which does not occur significantly at dc or low-frequency ac, prevents currents flowing on the inner braid surface from interacting with those on the outer surface, and vice versa.

While traveling within the transmission line, I_1 flows on the center conductor, and I_2 flows only on the inner surface of the outer shield. When antenna current is flowing from left to right (Fig. 1), I_1 flows out of dipole arm 1 onto the center conductor, and returns to the generator. I_2, being of opposite phase, flows along inside the feed line until it reaches the junction of dipole arm 2. At this junction, I_2 divides into two separate paths and forms I_3, which flows back down the *outside* surface of the feed line, and I_4, which equals $I_2 - I_3$ and flows onto dipole arm 2. The magnitude of I_3 depends on the impedance to ground provided by the outside surface of the coaxial shield.

If the effective path length to rf ground is an odd multiple of a quarter wavelength, the impedance will be very high, making I_3 negligible. In this case, I_1

[1]Notes appear on page 40.

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and I_4 will be nearly equal. On the other hand, if the rf path to ground is a multiple of a half wavelength, the impedance will be fairly low and current I_3 may be substantial, — resulting in unequal currents in the dipole arms and radiation from the feed line. In many instances, this rf path to ground includes the transmitter line cord and some house wiring, terminating at the power-line ground! Thus, the amplitude of I_3 varies with changes in feed-line length because of impedance effects.

Effects of I_3 on Antenna Impedance

It should be kept in mind that transmission-line currents I_1 and I_2 cannot produce radiation because their fields are not only of equal magnitude and opposite phase, but are confined within the shield of the coaxial cable. The field developed by I_3 *does* radiate, however, and thus the outer surface of the coaxial braid effectively becomes "dipole" arm 3, which is connected in parallel with arm 2.

To clarify this equivalent connection of radiators, I've simplified the circuit as shown in Fig. 2. Since I_1 and I_2 do not interact with any other currents, we may hypothetically place the rf generator directly between the input terminals of the antenna. Now that coaxial cable is no longer needed to transfer power from the generator to the antenna, the third conductor of the feed line (the outside surface) can be replaced with a single wire connected between arm 2 and rf ground. We have not changed the circuit electrically because I_3, which previously flowed on the outside of the coaxial cable, still flows to ground — but on the single wire.

We know that, depending on height, the impedance of a dipole (at resonance) is usually between 50 and 75 ohms, and is purely resistive. At frequencies above resonance, the resistance increases gradually and series inductive reactance appears; below resonance, the resistance decreases and capacitive reactance appears. The impedance of each dipole arm is one half of the total dipole impedance. Since the far end of arm 3 is at rf ground, its impedance behavior follows that of a short-circuited transmission line. Thus, when the length of arm 3 is an odd multiple of a quarter wavelength, its impedance is a parallel-resonant maximum, a high resistance typically of about 2000 or 3000 ohms. This high resistance in parallel with arm 2 has little effect on the total dipole impedance. However, as the effective length of arm 3 departs from a quarter wavelength (or odd multiples thereof) by changes in either its physical length or the generator frequency, the input resistance of arm 3 decreases, and reactance also appears in series with the resistance. This reactance is inductive when length decreases and capacitive when length increases. If the length of arm 3 is a multiple of a half wavelength, the resistance will be a series-resonant minimum value (but not zero, because of radiation). Thus, when arm 3 departs substantially from an odd multiple of a quarter wavelength, the net resistive and reactive components of the parallel combination of arms 2 and 3 are different than those of arm 1. Consequently, the dipole impedance is different than if arm 3 was not present.

Returning to Fig. 1, we can now see that, without a balun, changing the feed-line length is also a change in the antenna length, which in turn affects the impedance at the far end of the feed line. Therefore, the SWR measured at the transmission line input *changes with line length* when no balun is present to eliminate I_3. This phenomenon explains a point that is often puzzling for the amateur who uses no balun, and must trim his dipole each time the feed-line length is changed!

Function of the Balun

It is evident that, in coupling an unbalanced line to a balanced load (such as a dipole), the *primary* function of a balun is to block the current path between the inside and outside surfaces of the coaxial shield. With a balun in the circuit, I_2 will not divide at the end of the feed line to form I_3, but instead will flow only onto dipole arm 2. Thus, when I_3 is zero, $I_4 = I_1$, and the currents flowing on dipole arms 1 and 2 are balanced.

Although I pointed out this concept to Reisert for use in his balun article,[3] he apparently missed my point concerning the source of external current I_3. Consequently, his Fig. 2 and associated paragraph do not address the *principal* function of a balun. Contrary to his explanation of Fig. 2, when antenna currents on the feed line are caused by *asymmetrical coupling* to the antenna, a balun *will not* eliminate these currents, but will only change their phase and magnitude.

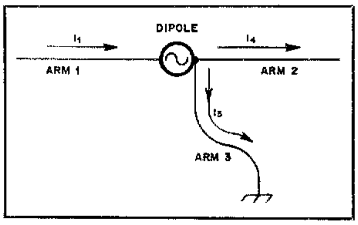


Fig. 2 — Simplified electrical representation of Fig. 1.

Effect of No Balun on Measurement Accuracy

It should be obvious that obtaining accurate impedance measurements of a dipole antenna is difficult. When a transformer-type balun is used to avoid errors caused by I_3, impedance-transfer errors obscure the true impedance at the antenna terminals. If the balun is omitted, the true impedance is obscured by the impedance of arm 3 shunting one half of the dipole. Since there is no practical way of determining the impedance of arm 3, the true antenna impedance and SWR cannot be calculated from the measured data.[4]

We should bear in mind that for any given physical length of feed line, the *electrical length* of the coaxial braid surface carrying I_3 is not the same as the inside conductors carrying I_1 and I_2. This is because the dielectric constants and the propagation-velocity factors are different. For example, the velocity factors for polyethylene and Teflon dielectric coaxial cable are 0.659 and 0.695, respectively. If the outside surface of the coaxial cable is bare, the velocity factor for the outer shield carrying I_3 approaches 0.95. However, the usually thin outer covering of polyvinylchloride (or sometimes Teflon) will reduce the antenna-current velocity factor to a value somewhat less than 0.95.

Effect of No Balun on Antenna Performance

From an operational viewpoint, I_3 itself is usually not detrimental to the perfor-

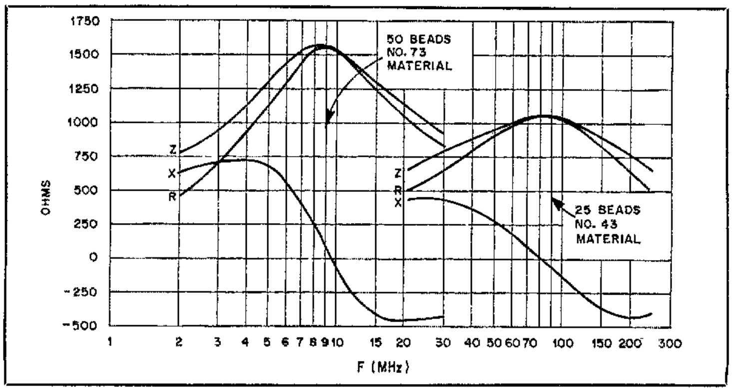


Fig. 3 — Graph of frequency vs. series impedance of coaxial-balun shield outer surface.

mance of simple dipoles for 160 through 40 meters. In addition, I_3 alone does not cause TVI, but radiation from external feed-line current can cause severe distortion in the radiation patterns of directive antennas, such as Yagis and quads. Unless a gamma match or other type of unbalanced input-matching scheme is used, all beam antennas with balanced input terminals require a balun if the optimum performance of the antenna system is to be achieved when fed with coaxial cable. For example, when a balun is not employed, the feed line and tower together become a separate, nondirectional antenna. This produces unwanted vertically polarized radiation that fills in the rearward null in the beam pattern, destroying the front-to-back ratio. The tower radiates along with the feed line, because currents are induced through coupling between the feedline and the tower.

The Choke Balun

Although many baluns embody some form of coupling transformer, an alternative is to insert an rf choke in the outer conductor of the feed line. This presents a high impedance to I_3 without affecting the internal currents. Advantages of this method are the lack of limitations on either maximum SWR or power handling. In addition, there is no impedance-transfer error that plagues transformer types of baluns (causing a skewing of SWR and impedance plots), because the choke balun has no coupling transformer — the feed line goes straight through to the antenna terminals!

The simplest choke balun is formed by coiling up a few turns of the feed line, starting where it connects to the antenna terminals. In the frequency range of 14 to 30 MHz, several turns of feed line coiled in an 8-inch diameter form an inductor with enough series reactance to minimize I_3 and practically eliminate feed-line radiation. Unfortunately, this form of choke (with an air core) is not practical below 14 MHz, because too much coiled-up feed line would be required to reduce I_3 to an acceptable level.

A word of caution is in order when the choke balun is used on tower-mounted antennas: The choke coil should be placed directly at the feed terminals of the driven element. If the coil is placed away from the feed terminals, any portion of feed line between the terminals and the coil is coupled to the boom or mast, which in turn is coupled to one arm of the driven element. The result — imbalance of currents in the driven-element, pattern skewing and tower radiation.

The frequency range of the choke balun can be extended to well below 2.0 MHz by using a core of high-permeability ferrite instead of air. With higher core permeability, the choke inductance increases dramatically, thereby retaining the high reactance needed to minimize I_3 at the lower frequencies. Of great importance, no core saturation occurs at high-power levels in the choke balun (a serious problem in transformer-type baluns), because the core excitation is low level, produced only by I_3 and not by the high internal current that feeds the antenna.

At my suggestion, Reisert made his choke balun with a Q1 material ($\mu = 125$ to 400) ferrite toroid, winding 9 turns of RG-141/U coaxial cable on the core for use from 14 to 30 MHz.[5] However, his 12-turn balun appears to provide marginal performance at 4 MHz. The problem stems from the toroidal winding arrangement. It is difficult to get a tight wrap of coaxial cable around the toroid, resulting in a coupling loss that makes it impossible to utilize the full value of the core permeability.

Balun Construction Using Ferrite Beads

I have obtained greatly improved choke-balun performance by placing several ferrite beads or sleeves of even higher permeability around the coaxial feed line.[6] For readers who wish to build this simple coaxial balun, bead materials of various size and rf characteristics are available that dramatically increase both the reactance and resistance of a conductor. (Adding resistance to the reactance in this circuit improves the operational bandwidth of the balun with no increase in loss.) In general, the impedance of the outer coaxial braid surface increases almost proportionately with the number of beads placed over it. A combination of 50-ohm teflon-dielectric RG-303/U cable (or RG-141/U, with the fabric covering removed) and ferrite beads having an ID of 0.197 in. and a length of 0.190 in., form a superb, compact, wide-band balun.[7] While the two inner conductors of the coaxial cable remain unaffected, the beads introduce a high impedance in series with the braid outer surface. This configuration effectively isolates the external output terminal of the feed line from that at the input end.

A test balun was made by slipping 300 no. 73 beads ($\mu = 2500$ to 4000) over a piece of RG-303 coaxial cable. The impedance of the outer conductor of the cable measured 4500 + j3800 ohms at 4.0 MHz; 15.6 + j13.1 ohms was measured utilizing a single bead. For practical baluns (less than 12 in. long, including connector) used from 1.8 to 30 MHz, use 50 no. 73 beads (Amidon no. FB-73-2401 or Fair-Rite no. 2673002401-0); for 30 to 250 MHz, use 25 no. 43 beads ($\mu = 950$ to 3000, Amidon no. FB-43-2401 or Fair-Rite no. 2643002401). No. 64 beads ($\mu =$ 250 to 375) are recommended for use above 200 MHz, but I have not yet experimented with them.[8] The coaxial cable need only be long enough to hold the beads, and to access the end connectors.

The graphs in Fig. 3 show the measured values of series resistance (R), reactance (X) and impedance (Z) versus frequency of the outer braid surface of a choke balun, for both the 25- and 50-bead types. With either balun, I_3 will be negligible. Using a balance-measuring technique learned from my RCA antenna-lab colleague, O. M. Woodward, the output terminal imbalance relative to ground of these baluns is undetectable using an HP-410C rf VTVM.[9]

At legal input levels, no power-handling problems will arise using these baluns, because the cw power-handling capability of the cable is 3.5 kW at 50 MHz, and 9 kW at 10 MHz.[10] Any suitable connector that will mate with the load end of your feed line can be used at the input of the balun, and the balanced-output terminals may simply be pigtails formed by the inner and outer conductors of the feed line. Methods for connecting the output terminals to the antenna are left to the ingenuity of the reader.

To emphasize simplicity, what vhf antenna buff wouldn't delight in dumping his unwieldy, frequency-sensitive, half-wavelength line balun? He can replace it by simply putting some ferrite beads on the last few inches of his coaxial feed line!

Notes

[1] J. Nagle, "RF Impedance Bridge Measurement Errors and Corrections," *Ham Radio*, May 1979.
[2] G. Hall, ed., *The ARRL Antenna Book*, 14th ed. (Newington: ARRL, 1982), Chapter 5, p. 5.
[3] J. Reisert, "Simple and Efficient Broadband Balun," *Ham Radio*, Sept. 1978, p. 12.
[4] W. Orr, "Multiple Dipole for Portable Use," *Ham Radio*, May 1970, p. 14.
[5] See note 3.
[6] D. DeMaw, *Ferromagnetic Core Design and Application Handbook* (Englewood Cliffs, NJ: Prentice-Hall, Inc., 1981), Chapter 4.
[7] mm = in. × 25.4.
[8] Ferrite bead materials are available from Amidon Associates, 12033 Otsego St., N. Hollywood, CA 91607, or Fair-Rite Products Corp., 1 Commercial Row, Wallkill, NY 12589.
[9] O. Woodward, Jr., "Balance Measurements on Balun Transformers," *Electronics*, Sept. 1953, p. 188.
[10] *RF Transmission Line Catalog and Handbook, No. TL-6* (Wallingford, CT: Times Wire and Cable Co., 1972).

Strays

QEX: THE EXPERIMENTERS' EXCHANGE

Wonder what you've been missing by not subscribing to *QEX*, the ARRL newsletter for experimenters? Among the features in the February issue were:

- Second ARRL Packet Conference Preregistration
- "PROM Programmer/Reader and Utility Software for the 2708 and 2716," by G. M. Palmer, K8LG
- "VHF + Technology," by Geoff Krauss, WA2GFP

QEX is edited by Paul Rinaldo, W4RI, and is published monthly. The special subscription rate for ARRL members is $6 for 12 issues; for nonmembers, $12. There are additional postage surcharges for mailing outside the U.S.; write Headquarters for details.

Hints and Kinks

Conducted By Larry D. Wolfgang,* WA3VIL

THE J² ANTENNA FOR 10 AND 24 MHz

☐ This J² antenna was developed to cover the 10-MHz band, but with an eye toward future operation on the 24.89- to 24.99-MHz band. The antenna provides omnidirectional low-angle radiation with a single feed point. Fig. 1 shows the antenna dimensions.

On 10 MHz, the J² is configured as a 5/8-λ vertical, which exhibits a theoretical gain of about 3 dB compared to a 1/4-λ vertical. At 24 MHz, the J² becomes two in-phase, half-wave J antennas. The antenna base should be mounted not more than 2 feet above ground for best performance.[1] You should provide a few 1/4-λ radials for 10-MHz operation (23 feet). No radials will be required for the 24-MHz band, when that one becomes available for amateur use.

Matching to the base of the J² can be implemented with either an open-wire transmission line and matching network in the shack, or by means of an L network at the base of the antenna. The feed-point impedance will be high on both bands (>1000 ohms).

The antenna can be suspended from the side of a tower or from the limbs of a tall tree. Remember that both ends of the antenna are high-impedance points, so the rf voltage will be high. Use good insulators to support the main vertical wire.

The 1/4-λ stubs are held away from the main wire by means of homemade Plexiglas spreaders. The length is not critical up to a maximum of about 6 inches. Position the spacers about 1 foot apart along the stubs, to maintain an even spacing.

This antenna is a little short of being 1/2-wavelength long on 40 meters. By switching in some additional inductance at the base of the antenna, you should be able to use the J² on that band also. Operation as a 1/4-λ vertical for 80 meters should also be possible, but that would require a much more extensive radial system. — *Richard Schellenbach, W1JF, Reading, Massachusetts*

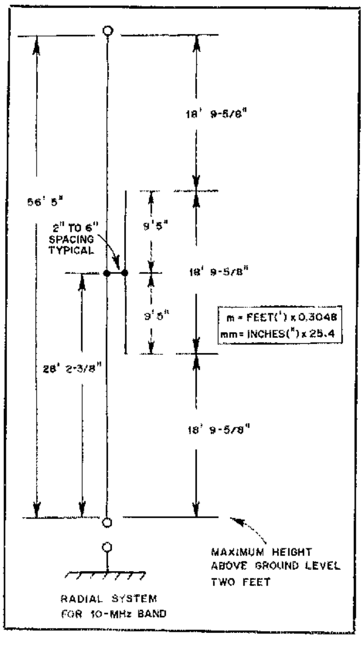


Fig. 1 — Dimensions and construction information for the W1JF J² antenna.

FIBERGLASS POLES FOR ANTENNA CONSTRUCTION

☐ John Williamson, WA2UTG, and George Smith, W4AEO, have passed along some information about locating fiberglass poles for use in building the YV5DLT Telerana antenna. Others interested in building this antenna or looking for fiberglass poles for quad antennas may find it useful. Each of the listed companies has expressed a willingness to sell to individuals. You should contact them for prices and specific sizes available at the time. Two of the companies handle materials specifically for Amateur Radio antennas: db+ Enterprises and Viking Instruments, Inc. They also have other antenna hardware available. The other companies have various sizes of tubing and solid poles. The lengths and diameters may depend on the size of the production-run stock at the time of your order. Table 1 summarizes the ordering information provided by each company. All shipments are made by truck, because of the pole lengths. You may have to pay a minimum charge set by the motor freight companies. — *Larry Wolfgang, WA3VIL, ARRL Hq.*

ALLERGIC REACTION TO JULY 1982 HINT & KINK

☐ I was alarmed to read the Hint by Charlie Burke, WA2SLK, in the July 1982 Hints and Kinks column. I have used this method of heating with a propane torch to remove components from surplus pc boards, but have two cautions for others who would use this method.

I donated some old pc boards to our high school. A student was using a propane torch to strip the parts in a well-ventilated area outside the building. He developed an allergic reaction that caused swelling and hives. As a result, he spent several days out of school. I'm not sure what caused the reaction, but others should be aware of this case.

The second caution is that resistor color bands may shift colors when heated (orange becomes red, for example). I find that a large soldering iron and an awl or probe used as a pry bar will remove resistors almost as fast. — *Robert Dixon, W8HGH, Rochester, Michigan*

LUBRICATION FOR THE HEATH REMOTE COAX SWITCH

☐ I have owned a Heath SA-1480 remote coax switch for several years. It worked fine in the warm months, but when the weather got cold the switch became unreliable and sometimes refused to operate until I would climb the tower and hammer on the assembly. Finally, I disassembled the unit and applied a liberal amount of CRC 5-56 spray lubricant on the solenoid end of the motor. This seems to have completely cured the problem. The lubricant I used is manufactured by CRC Chemicals, Warminster, PA 18974. It should be available in most auto-supply stores. — *Ed Kuebert, K3KA, Brookeville, Maryland*

END CAPS FOR ANTENNA ELEMENTS

☐ Don't discard the plastic or metal tops from spray paint cans! You may find that they will come in handy as end caps for antenna booms and masts. Not only are they functional, they will also add a bit of color to your antenna.

Many plastic paint-can tops will fit pipes of two different outside diameters, 1-1/4 and

Table 1
Fiberglass Pole Ordering Information

Company	*Material Size*	*Orders*
Advanced Composites P.O. Box 15323 Salt Lake City, UT 84115	1-1/4 and 1-1/2 in. OD 12- and 20-foot lengths.	Cash in advance; no minimum order.
db+ Enterprises P.O. Box 24 Pine Valley, NY 14872	1-1/16 in. diameter 13-ft lengths. Severe-duty cubical quads and accessories.	No minimum order.
Dynaflex Manufacturing Corp. Rte. 14, Box 370 Tallahassee, FL 32304	Three-section telescoping poles, 15-foot total length.	$19.50 each; minimum order four poles. Payment with order, plus $8 shipping and handling charge.
Sky-Pole Manufacturing, Inc. 1922 Placentia Costa Mesa, CA 92627	Vaulting poles and tubing of various sizes and lengths. 1 to 1-5/8 in. tubing in odd lengths; $1 per foot.	No minimum order. Payment with order, or C.O.D.
Viking Instruments, Inc. Kirk Electronics Div. 73 Ferry Rd. Chester, CT 06412	Arms for quad antennas, 9- and 13-foot lengths. Hollow and tapered, but reinforced at the 10, 15 and 20-meter drill points.	No minimum order. 9-foot arm, $17; 13-foot arm, $21 each.

[1]mm = in. × 25.4; m = ft × 0.3048.

*Assistant Technical Editor

2-3/8 inches. The 1-1/4 inch part is a firm fit over the end of a standard piece of steel TV mast. — *Paul Pagel, N1FB, ARRL Hq.*

THE CW PARROT

☐ One of the net control operators on the Idaho-Montana Net (IMN), Kevin Nathan, K7RX, is blind. Kevin is a very competent net control operator, and also serves as RN7 liaison. He uses a device, developed by several other net members, that allows him to retransmit cw messages without first having to transcribe them into Braille. Kevin tape records the traffic, then uses the CW Parrot to interface the recorder with his transmitter for playback.

Fig. 2 shows the schematic diagram of the basic unit. Additional features, such as peak, notch, high-pass and low-pass filters or a noise blanker, could be included to aid the operator. For transmitters with a positive keying voltage (up to about 30 V), the Parrot can be wired in parallel with the transmitter key line. It could serve as a "poor man's memory keyer," or to let others hear what their fists sound like.

All components are mounted on a small etched-circuit board (or perfboard) in a 2-3/4 × 2-1/8 × 1-5/8 in. aluminum box. Rubber grommets in the box protect the two shielded cables from damage. The construction details can vary to suit the needs of each individual.[1]

The cw signals are recorded directly from the receiver. When it is time to retransmit the message, simply plug the Parrot into the recorder earphone jack, and the tape will be reproduced faithfully over the air.

Ferrite beads and C1 are used for rf suppression. D1 and D2 provide full-wave rectification of the audio signal. This pulsating dc causes Q1 to switch on, keying the transmitter. — *Jim Voyles, K7JV, Boise, Idaho*

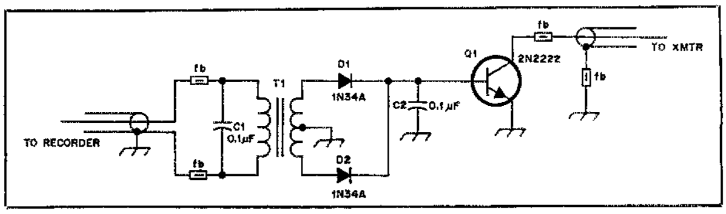


Fig. 2 — Schematic diagram of the CW Parrot, used to transmit a cassette-recorded message.
T1 — Audio transformer, 8-ohm primary, 1000-ohm secondary, ct, Radio Shack no. 273-1380.

HW-101 OSCILLATION PROBLEM

☐ My Heath HW-101 developed a sudden problem that had a simple cure. I noticed that the rig would start to oscillate on every band as I tuned it for maximum-rated output power. My first guess was that the driver or final-amplifier tubes were weak, but they proved to be okay. Further checking turned up a ground foil on the driver grid-switch board that was not making contact with the comb bracket. Fig. 3 shows a portion of detail 8-19A, found on page 99 of the HW-101 manual. My construction manual contains no mention of the need to solder this connection, but after I did, the oscillation was eliminated. Later manuals include this step. I suggest you check your transceiver to see if the connection is soldered. — *Robert St. Amant, N0AXK, Edina, Minnesota*

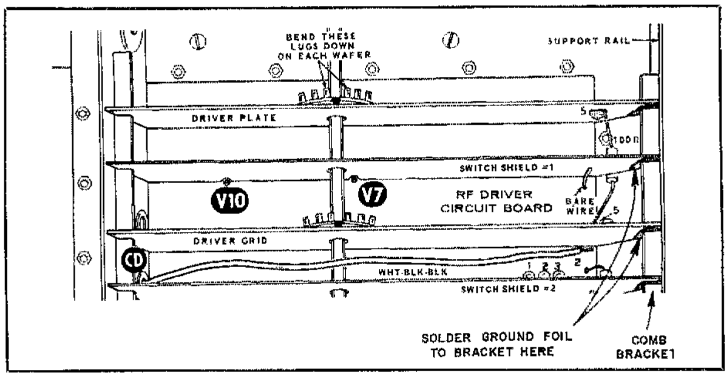


Fig. 3 — Pictorial diagram showing where N0AXK soldered the ground foil of the driver grid-switch board to the supporting comb bracket.

INEXPENSIVE BURGLAR ALARM

☐ Photoconductive cells are less sensitive to light changes than are phototransistors. For burglar-alarm purposes this can be an advantage. A very simple alarm system can be built on a piece of perforated board 2 × 3 inches or smaller. I mounted mine on the lid of a tobacco tin. A 12-V dc supply is fed directly to a 5- or 6-V DIP relay through a CdS cell, which has a dark resistance approaching 0.6 MΩ and a bright-light resistance of about 100 Ω. The complete diagram is shown in Fig. 4. S1 is used to select whether V + is fed through the normally open or normally closed relay contacts. This provides for beam-open or beam-closed operation. K2 has two sets of normally open contacts. One set is used to lock the relay closed when the alarm is tripped, and the other supplies voltage to a horn or other warning device. An LED across this line serves as a beam-reception indicator for alignment purposes, or as a remote indicator that the alarm has been set off.

With the CdS cell mounted in a cardboard tube with lenses from a toy telescope, the relay can be held "on" by a pocket-flashlight beam at about 25 feet. The alarm will be triggered by any object crossing the beam, including a person at a brisk run. This simple form of security fence is adequate for many purposes, and can be built for under $10.

Other warning devices can be used instead of the horn. An LM-3909 IC can be wired as an audio oscillator to drive a small speaker. An amplifier can be added for a louder signal. A 12-V bulb and reflector could be mounted alongside the photocell, and the beam reflected by a mirror.

There are many applications for such a simple device. Mine keeps unauthorized people out of the ham shack! — *Alex Comfort, M.D., KA6UXR, Santa Barbara, California*

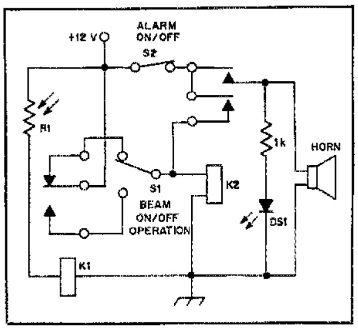


Fig. 4 — Schematic diagram of a simple burglar alarm, using a CdS photoresistor. K1 has a 5-V coil, and K2 has a 12-V coil. R1 is Radio Shack part no. 276-116 or equivalent.

ETCH-RESIST IDEA FOR CIRCUIT BOARDS

☐ Most hams have struggled with marker pens, stencils, rub-on patterns and other ways of tracing etching patterns onto circuit-board material. Each of these methods has certain drawbacks. For all but the most complex patterns I find it easier to draw freehand on the board. I use an inexpensive straight pen with replaceable points.

The ink I use is an etch-resist lacquer purchased in an electronics-parts store. Dilute the ink with the solvent listed on the label (toluol, in my case) so it has a consistency that works well with your pen. Nail polish diluted with remover (acetone) also works. Either of these inks will gum up the pen points, so be sure to wipe them clean when you're finished tracing the pattern. After etching, remove the lacquer with fine steel wool. — *John S. Mason, Jr., EA4AXW, Madrid, Spain*

[1]Parts for this project are available from RADIOKIT, Box 411, Greenville, NH 03048.

Technical Correspondence

Conducted By
Dennis J. Lusis,* W1LJ

The publishers of QST assume no responsibility for statements made herein by correspondents.

YAGI ELEMENT MOUNTING ADVICE

☐ The article by Lusis, "Go for the Gain, NBS Style," (Aug. 1982 *QST*) is valuable for pointing out the effect of element and boom diameters on Yagi tuning. As a Yagi is increased in length, element dimensions become more and more critical. Failure to take element and boom diameters into account can make the difference between full performance and a gain figure little better than a dipole. However, Lusis neglected to mention that Yagi performance is also influenced by the way in which elements are mounted to the boom.

These effects can be understood by considering the parasitic element as a parallel tuned circuit containing distributed inductance and capacitance (Fig. 1). As with any tuned circuit, the element resonant frequency will be determined by the values of inductance and capacitance.

The inductance of a straight conductor depends on diameter as well as length — the larger the diameter, the less the inductance. For the case of a half-wave antenna element, the diameter at the center has more influence on inductance than end diameter — because that is where current is maximum.

Capacitance of a cylindrical rod is approximately proportional to the diameter. Unlike inductance, capacitance is more influenced by the diameter at the ends of the half-wave element, where voltage is maximum.

If we uniformly increase the diameter of our element, we decrease the inductance, which raises the resonant frequency. We also simultaneously increase the capacitance, which lowers the resonant frequency. Unfortunately, the two effects do not completely cancel; diameter has more influence on capacitance, which means that a larger diameter element of given length will have a *lower* resonant frequency. Put another way, the larger-diameter element will need to be cut shorter for resonance at a given frequency.

The effect of boom diameter can also be understood by referring to Fig. 1. Where the element pierces a larger diameter boom, the boom in effect "shorts out" a small amount of inductance at the center of the element, and thereby raises the resonant frequency. The boom has a negligible effect on capacitance because there is a voltage minimum at the center of the element. Of course, the boom cross section also has some inductance, but since the diameter is greater than that of the element, the effect is to slightly diminish total inductance. We can conclude, therefore, that the influence of the boom will be much less when the element is insulated from the boom, as in Fig. 2A, than when the element is in electrical contact on both sides, as in Fig. 2B. The effect of the boom will also be much less when the element is mounted atop the boom, as in Fig. 2C.

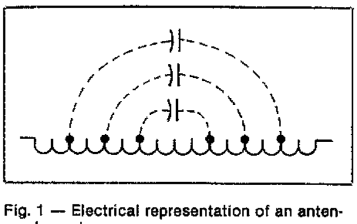


Fig. 1 — Electrical representation of an antenna element.

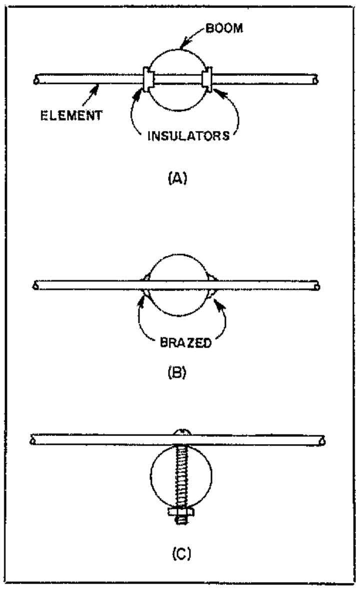


Fig. 2 — Various mounting techniques for Yagi elements.

Unless proper precautions are taken, corrosion can eventually result in poor electrical contact at the point where the element pierces the boom, changing the resonant frequency of the element. For this reason, the junction should either be soldered or brazed around the complete circumference of the element, or be completely insulated from the boom by the use of plastic sleeves.

When the element is mounted to the boom by means of a conducting plate, as in the W1LJ 6-meter antenna, the resonant frequency will be influenced much more by the size of the plate than by the diameter of the boom. — *Fred Brown, W6HPH, Lake San Marcos, California*

NBS BOOM-CORRECTION FACTORS

☐ I would like to make some points concerning the Lusis article, "Go for the Gain, NBS Style," (Aug. 1982 *QST*). The boom-correction factor in Viezbicke's preliminary data was based on elements passing *through* the boom. When the *NBS Technical Note 688* was published, *this point was omitted.* Mounting elements on top of the boom with a plate and muffler clamps requires a different correction factor, as pointed out by Lawson in "Yagi Antennas: Practical Designs" (Dec. 1980 *Ham Radio*). The length of the elements is so sensitive to the type of boom and the mounting procedure that Viezbicke's preliminary data show separate curves for elements passing through round and square booms — even though the final report states "round and square booms yielded similar results" (Fig. 3).

For a typical vhf Yagi, elements passing through the boom require a correction factor of approximately 60% of the boom diameter. Elements mounted on top of the boom may require a correction factor as small as 6% of the boom diameter, plus a correction due to the increased element diameter attributable to the mounting plate. Misapplication of these correction factors on a typical 2-meter Yagi can result in element-length errors of up to 1%. Viezbicke states that element lengths must be cut to an accuracy of 0.3% to maintain performance.

These boom corrections are relatively small, but at vhf and uhf they can make the difference between an optimum design or just another mediocre Yagi. Because of the mechanical variables inherent in mounting elements on top of the boom, I recommend that at 100 MHz and higher all NBS Yagis be constructed with the elements passing *through* the boom, using the boom-correction factor published in *NBS Technical Note 688.* — *John Brosnahan, WØUN, Boulder, Colorado*

ALTERNATIVE FILTER DESIGN

☐ John Webb suggested a fifth-degree elliptic high-pass filter design in his interesting article, "Electrical Antenna Null Steering" (Oct. 1982

*Assistant Technical Editor

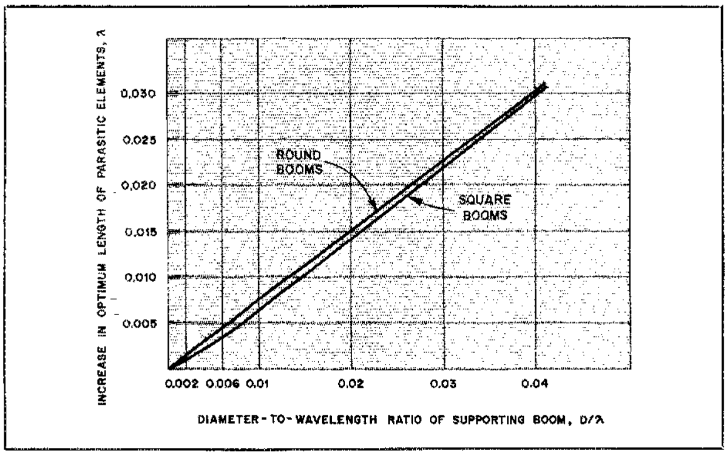


Fig. 3 — Viezbicke's early data for *NBS Technical Note 688* shows different correction factors for round and square booms.

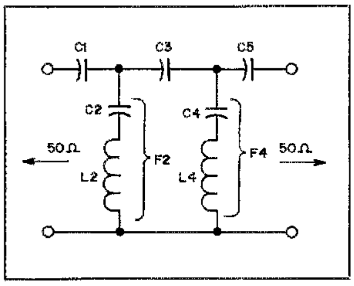


Fig. 4 — Schematic diagram of a fifth-degree elliptic high-pass filter. See Table 1 for component values.

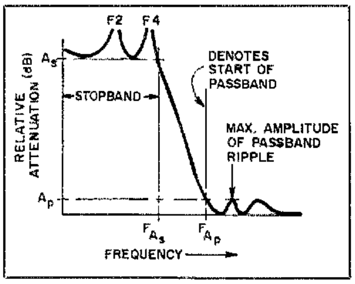


Fig. 5 — Response curve of a fifth-degree elliptic high-pass filter. See Table 1 for response figures.

Table 1

Comparative Parameters of W1ETC and W3NQN Elliptic High-Pass Filter Designs

Design and Response Parameters	*W1ETC*	*W3NQN*	*Filter Component Values*†	*W1ETC*	*W3NQN*
FA_s (normalized) =	0.49167	0.52754	C1 =	456	620
A_s (dB) =	70.3759	62.0793	C2 =	10,600	10,415
A_p (dB) =	1.03451	0.37677	C3 =	343	430
R.C. (%) =	46.04	28.83	C4 =	3900	3844
VSWR =	2.706	1.810	C5 =	487	680
F_{co} (MHz) =	3.4	3.423	L2 =	2.28	1.947
FA_s (MHz) =	1.67	1.806	L4 =	2.5	2.201
F2 (MHz) =		1.118			
F4 (MHz) =		1.730			

†All capacitance values are in pF; all inductance values are in μH.

QST), for attenuating broadcast-band signals below 3.4 MHz. Since the three series capacitor values in Webb's design are nonstandard, construction could be simplified by using a design employing standard values. I have completed such a design, which closely approximates the performance of Webb's. A comparative listing of the designs and performance parameters of the two filters appears in Table 1.

The shunt capacitors in my design are not standard values, but one close to (and below) the design value can be used with additional capacitance added in parallel to achieve the desired resonant frequency. Further fine tuning can be accomplished by squeezing or spreading the inductor windings.

For attenuation levels in excess of 50 dB, my personal preference is to use a seventh-degree elliptic design with a much lower VSWR than that used by Webb. Of course, this complicates the design by requiring additional components and more tuning.

I am completing tabulations of 50-ohm, 1-10 MHz, fifth-degree elliptic low-pass and high-pass filter designs similar to the one provided. When these tabulations are completed, standard-value capacitor filter designs should become popular because of their obvious convenience. Furthermore, choosing different cutoff frequencies may be done without the use of calculations, if input and output impedance levels remain equal to an integral power of 10 times 50. A relatively simple scaling procedure is used for other impedance levels. — *Ed Wetherhold, W3NQN, ARRL TA, Annapolis, Maryland*

RECONSIDERING ELLIPTIC FILTERS

□ Correspondence regarding my article, "Electrical Antenna Null Steering," (Oct. 1982 *QST*), led to further consideration of operating the device at lower frequencies, particularly regarding the input filters. The 3.4-MHz elliptic filter mentioned in my article, as well as other filters to be described, were taken from "Simplified Modern Filter Design," by Philip R. Geffe, John F. Rider Publications, 1963. The filter attenuation of 70 dB (below 1.6 MHz) and passband ripple of 1.0 dB were verified on a 2.0-MHz scale model. This ripple factor results in a VSWR of nearly 3.0:1; however, the attenuation at 1.6 MHz is slightly better than the amount provided by filters having less ripple. The filter is to be used in a receiving application, so the impedance mismatch is not a critical issue.

A five-pole elliptic filter would provide only 30 dB of attenuation at 1.6 MHz, with the cutoff frequency at 1.8 MHz to allow 160-meter operation. Thus, seven- and nine-pole filters were considered for 160 meters. Broadcast-band attenuation will be 45 dB for seven poles and 65 dB for nine poles. The passband ripple of these designs is 0.18 dB, giving a VSWR of 1.5:1. A nine-pole filter is being built on a circuit board, and the results will be reported in *QST* when more data is available.

My filters are built with standard-value parts that add to near the calculated values. Other designs that can be implemented with single, standard-value components may be available — hopefully with only small compromises in performance. As the number of parts in a filter increases, a goal of using single, standard-value components becomes increasingly difficult. I wish to thank ARRL TA Ed Wetherhold, W3NQN, and others for their contributions to discussion of the broadcast-band filter issue. — *John K. Webb, W1ETC, Amherst, New Hampshire*

Feedback

□ The schematic diagram of "A Simple Capacitance Meter You can Build" (*QST*, Jan. 1983, p. 35), shows the 10-μF range resistors as 8.2 kΩ and 1 kΩ. These resistors should be 8.2 MΩ and 1 MΩ, respectively.

□ Fig. 7, page 33, of the article by H. Granberg in Jan. 1983 *QST* has a minor error: R2 should be labeled R24. In Fig. 8, the adjustment control, R4, should be 100 kΩ, rather than 100 Ω.

□ Last month's In Training column recommended the ARRL book *A Course in Radio Fundamentals* for review in licensing courses. Unfortunately, this book is out of print and no longer available for sale.

Conducted By Paul K. Pagel,* N1FB

Cushcraft R3 Three-Band Vertical Antenna

When asked if I would be interested in reviewing the Cushcraft R3 10, 15 and 20-meter antenna, I was pleased to accept. I had moved from a "no antennas" townhouse to a single-family house just two weeks earlier. Outside antennas were now allowed, but space was limited. I needed an effective antenna system for these three higher-frequency bands; a trap vertical seemed like a good compromise.

The R3 is more than an ordinary trap vertical! It operates as a 1/2-λ radiator on the three bands, and comes equipped with a sealed matching-network assembly and a remote tuning capacitor. A 24-V motor drives the large variable capacitor to match the antenna impedance to 50-ohm coaxial cable. Coupled to the capacitor shaft is a potentiometer, which provides a voltage drop that is monitored in the control box. The tuning control unit meter face is calibrated to show for which frequency band the capacitor is set.

Assembly

I spent a little more than an hour taking inventory of the kit hardware and reading the assembly instructions. As I examined the capacitor tuning assembly, my curiosity caused me to remove the cover to see what was inside. The cover should not have come off as easily as it did! A ceramic feed-through bushing on the top of the unit has a short piece of wire soldered to a lug on the variable capacitor (Fig. 1). This joint was cold soldered and loose on the review unit! Reconnection required that I remove the feed-through insulator, solder the wire and then reinstall the insulator as the cover was being put back in place. Had I not found this problem before installing the antenna, it would have led to frustration and much wasted troubleshooting time.

Antenna assembly took about two hours. I usually work slowly, reading directions several times as I go. The instruction sheet that comes with the R3 refers you to four different drawings for assembly details. Even after careful study, I had a few nuts, bolts and washers left over at the end. As I looked at the drawings again, I saw where I had forgotten to install the parts. I had to disassemble some of my work to correct the errors.

The instruction sheet says the antenna is designed to fit "conveniently" over a 1-7/8 inch OD tube.[1] A check with local electronics shops and TV dealers led to a morning spent trying to locate a short mast on which to mount the antenna. I found a hardware store that could supply galvanized pipe of the proper dimensions, but at a price greater than $5 per foot! I found a 6-foot section of 1-1/2 inch copper pipe at home. With a wooden plug in the top (to prevent crushing), this has served as a mast. The antenna is mounted on a 4-foot-high extension attached to the back of my garage. This places the bottom of the antenna about 10 feet above ground.

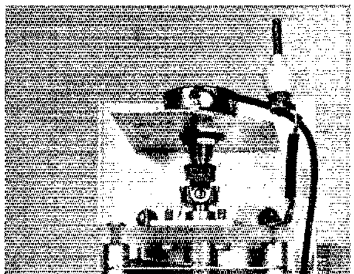


Fig. 1 — A look inside the capacitor tuning assembly. The short piece of wire between the capacitor and the ceramic feed-through bushing can be seen at the top right.

The instruction sheet cautions you to be sure the four-conductor control cable (not supplied) is connected in the same sequence at both ends. I found no indication of the correct sequence, so I assumed that as I looked at both terminal strips they would be the same from left to right. I connected a color-coded cable in this manner and turned on the control unit. The meter needle vibrated noticeably, and the TUNE control did nothing. I reversed the leads on the control unit and everything worked fine! A label on the terminal strips would be helpful. (As I prepared to write this review, I glanced at the instruction sheet again. Sure enough, the terminals *are* shown there, numbered — you guessed it — in reverse order from each other!)

Included with the literature is a pamphlet entitled, "How to Install your Outdoor Antenna Safely." It begins, "These safety recommendations apply to all Cushcraft CB, TV, Amateur and General-Purpose Communications Antennas." It contains a wealth of information on safe antenna installation procedures, stressing the danger of allowing the antenna to contact power lines. There is an inappropriate warning, however: "Remember that the FCC limits your antenna height to 60 feet." There is no indication that this applies only to CB installations.

Tables of voltage and resistance measurements taken at the control box and at the capacitor tuning assembly are included to aid you in troubleshooting the control circuit. A chart of troubleshooting checkpoints should also prove helpful if any problems develop.

Performance

SWR curves for the completed antenna are shown in Fig. 2. I built it to the dimensions suggested in the instructions. As you can see, lengthening the 10-meter portion would improve the SWR on the low end of the band, but I didn't think it was worth taking the antenna down to fool with it! Of course, the SWR curve doesn't tell the whole story. Even a dummy load should have SWR characteristics like this. The R3 is no dummy, though. When compared to using a 120-foot wire antenna and Transmatch, it provided two to three S units improvement in received signal strength at most times. On transmit, it also performed admirably. In my installation, the antenna is not high above the ground, and is within 20 feet of three very tall maple trees. In spite of this poor location, I was able to work plenty of European and West Coast stations and I received good signal reports.

The R3 requires no radial system, and so takes up almost no horizontal space at all. It is light in weight and mounts easily on a roof or to the side of a house. If you are looking for a good three-band antenna without the expense and effort of installing a tower and triband beam, the R3 could be for you. Leaving the tuning assembly and bottom section together, the antenna should be easy to take apart and put back together. This would make it an ideal

[1] mm = in. × 25.4; m = ft × 0.3048

*Assistant Technical Editor

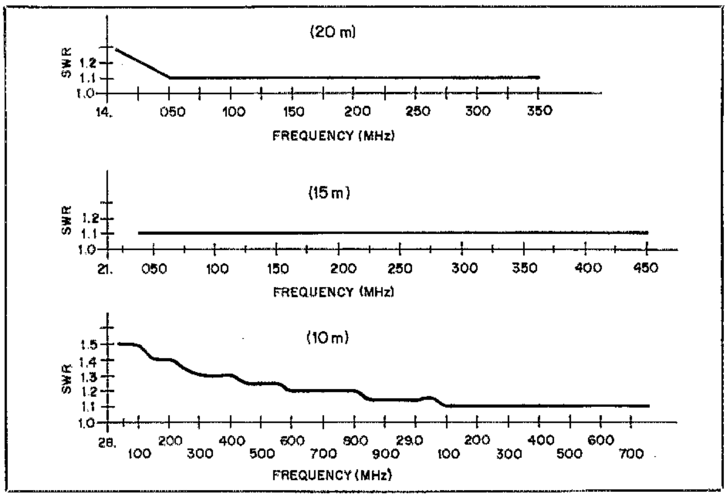


Fig. 2 — SWR curves for the Cushcraft R3 vertical antenna. No length adjustments were attempted after the initial construction.

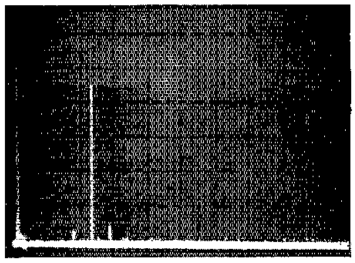


Fig. 3 — Worst-case spectral display of the Mirage C-22. Vertical divisions are each 10 dB; horizontal divisions are each 100 MHz. Output power is approximately 20 W at 220 MHz. The fundamental has been reduced in amplitude approximately 23 dB by means of notch cavities; this prevents analyzer overload. All spurious emissions are at least 64 dB below peak fundamental output. The C-22 complies with current FCC specifications for spectral purity.

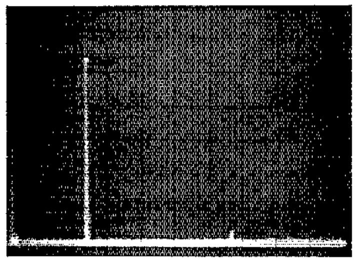


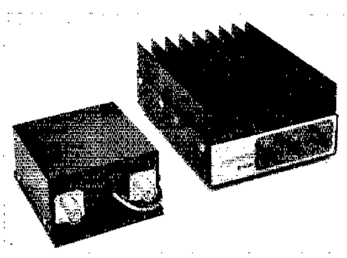
Fig. 4 — Worst-case spectral display of the Mirage C-106. Measurement conditions remain the same as for the C-22. The fundamental has been reduced in amplitude by about 15 dB to prevent analyzer overload. Power output is approximately 60 W at 220 MHz. The C-106 complies with current FCC specifications for spectral purity.

antenna for portable applications.

The R3 is available from Cushcraft Corporation, 48 Perimeter Rd., P.O. Box 4680, Manchester, NH 03108. Price class: $330. — *Larry Wolfgang, WA3VIL*

MIRAGE C22 AND C106 ALL MODE 220-MHz AMPLIFIERS

□ One of the brighter aspects of vhf fm operation is the wider range of products available for the 220-MHz enthusiast these days. Mirage has introduced two amplifiers ideally suited for use



with low-power hand-held transceivers. As with their amplifiers for the other bands, these are linear amplifiers and may be used for any mode of operation. Depending on input/output power requirements, one or both of these amplifiers might be ideal for the operators on the "low end" of 220 MHz.

The C22 is identical in appearance to the B23 reviewed earlier (May 1981). The resemblance to the 2-meter version goes even deeper than looks. Like the B23, the C22 uses an MRF240 for a single stage of amplification. At 220 MHz, the potential output power for the transistor is lower, but otherwise it functions quite well. Engineers from Motorola tell me that this is true for most of their line of modern "vhf" transistors. The C22 produces 20 W of output power with an rf input power of 2 W. A signal as low as 200 mW will key the rf switching circuit — which is identical to that of the B23. The first C22 failed shortly after it arrived, and Mirage replaced it. The second C22 has performed flawlessly during several months of mobile use. If you are looking for a compact "brick" to go along with your 220-MHz hand-held unit, consider the C22.

Mirage has provided for the mobile or base operator who wishes to boost the output of the typical 10-W transceiver to the 60-W level. The C106 is similar in appearance to the B108 2-meter amplifier, and it has the same features. Like the C22, the C106 can be keyed with as little as 200 mW. With a 2-W drive level, it will deliver approximately 20 W of output power, making it useful with "hand-helds." With 10 W of drive power, the output power climbs to 60 W or more. A single MRF247 provides this gain.

The C106 also has a built-in receiver preamplifier which provides 10-dB gain with a 2.5-dB noise figure. A front panel switch permits the user to turn the preamplifier on and off. A second switch allows the user to add a dropout delay to the antenna relay, facilitating use with an ssb transmitter. The third switch on the panel of the C106 applies power to the transistor driving the antenna relay. If the amplifier is located near the operating position, the user has complete control at this fingertips.

Of course, I wanted to mount the C106 in the trunk of my car. Mirage has very wisely taken care of that contingency. An option for the C106 is the RC-1 remote-control head. The RC-1 is a 1-1/2 × 3-1/2 × 1-1/2 inch (HWD)

Mirage C22 220-MHz Amplifier

Manufacturer's Specifications	*Measured in ARRL Lab*
Frequency range: 220 to 225 MHz.	As specified.
Power output: 20 W (nominal) at 13.6-V dc with 2-W drive.	As specified.
Input VSWR: Not specified.	Less than 1.5:1.
Spurious and harmonic output: Not specified.	−65 dB (See Fig. 3).
Size: (HWD) 2.25 × 4.75 × 4 inches (57 × 120 × 100 mm)	
Weight: 1.25 pounds (0.567 kg)	

Mirage C106 220-MHz Amplifier

Manufacturer's Specifications	*Measured in ARRL Lab*
Frequency range: 220 to 225 MHz.	As specified.
Power output: 60 W (nominal) at 13.6-V dc with 10-W drive.	As specified.
Input VSWR: Not specified.	Less than 1.5:1.
Spurious and harmonic output: Not specified.	−65 dB (see Fig. 4).
Receive preamp: Nominal 10-dB gain with 2.5 (±0.5) dB noise figure.	6-dB gain with 2.8-dB noise figure.

box with three switches and two LED status indicators that connects to the amplifier with a 20-foot-long cable (supplied). One simply sets the amplifier switches to POWER OFF, FM and PREAMP OFF; control of these functions is then accomplished with the RC-1. I've used the C106 and RC-1 with a Midland 13-509 transceiver for several months and am delighted with the performance. The combination of the receiver preamplifier and the extra output power is ideal for fringe-area usage. If you are on 220 MHz with a "barefoot box," you may want to consider the C106. It'll help you "hear" as well as "talk."

Price classes: C22, $90; C106, $200; and RC-1, $25. More information may be obtained from Mirage Communications Equipment, Inc., P.O. Box 1393, Gilroy, CA 95020, tel. 408-847-1857. — *Peter O'Dell, KB1N*

KANTRONICS CW TRAINING SYSTEM

☐ Contrary to the belief of some, cw is *not* dead! With this thought in mind, Kantronics developed a system that teaches cw to the beginner, and it also can be used by those who are already familiar with the Morse language to increase their proficiency. The system consists of a booklet and a computer code-practice program.

The Text

Morse Code, Breaking the Barrier, by Phil Anderson, WØXI, is a step-by-step instruction manual written to familiarize the beginner with the Morse code. The booklet contains five chapters. The subjects range from the structure of the Morse code right through copying off the air.

Chapter 1 introduces the "mechanics" of the code. The reader is made familiar with the elements that letters are "coded" from. Several examples are given, with the proper timing emphasized through "real-time" figures.

Chapter 2 defines the FCC regulations regarding the speed at which one must copy cw to obtain an amateur ticket. This chapter then presents a method for determining the speed of a tranmission.

Chapter 3 is entitled "Learning to Receive Code." The text in this chapter instructs the student in the proper method of learning "letter codes," with the intent of learning to copy the entire alphabet, a few basic procedure signals and punctuation.

Chapter 4 contains flashcards for initial memorization of letter sounds. The user is instructed to "read" a sound from the card and say the letter that corresponds to that sound. The correct letter, by the way, is printed on the back of each card.

Chapter 5, entitled "Special Codes/Copying Off the Air," introduces the aspirant to some Q signals and a few procedure signals.

In addition to the five chapters mentioned previously, the book contains separate appendices for symbols, numbers, abbreviations and Q signals. After reading these the user will be familiar with the jargon likely to be encountered in typical QSOs.

Software

The second portion of the training system is a computer program, "Hamsoft Code Practice: APPLE," written for use with a disk-based Apple II microcomputer system. The program is divided into six segments, ranging from beginner to actual QSOs (similar to the format used by the FCC). Each of the segments allows the user to select the desired speed, from 1 to 60 wpm.

The Beginner selection contains a choice of seven four-letter groups, with the XCAD group suggested as a starting point. Using this option, the student can practice the letters that are the most troublesome. After a slight pause, the previously sent letter is displayed on the screen.

"Letters" and "Numbers" choices send 18 five-character groups, and print them on the screen after the 18 groups have been sent (a handy tool for checking the accuracy of copy).

The "Calls" option sends a list of 15 call signs and displays them on the screen after the group is sent. This selection is handy for the experienced ham who is planning on going on a DXpedition and wants to practice call sign copying.

The program also contains an "Abbreviation" option, which sends 16 short words/abbreviations at a clip, displaying them on the screen.

"QSOs" allows the user to copy two FCC-type QSOs, and then check the copy by comparison to the "sent text" shown on the screen. I noted that the software could provide sending faster than 60 wpm in this mode, but after the text was sent, the computer "crashed" or locked up.

Comments

The program performed flawlessly during the time I was reviewing it. The training system is available from Kantronics, 1202 East 23 St., Lawrence, KS 66044. Price class: $30. — *Michael B. Kaczynski, W1OD, ARRL Hq.*

MACROTRONICS CODE CLASS

☐ Learning the Morse code can be difficult if one does it the wrong way (as I did). I memorized the code from a list of dots and dashes. When I copied the code off the air, I had to translate each character from a sound (di-dah) to the dot and dash equivalent, and finally to the alphanumeric equivalent. At 5 wpm, I could perform these mental acrobatics easily. As the speed climbed toward that enigmatic 13 wpm, however, I could not make the translation fast enough — getting it down on paper was impossible.

At that point, I discovered that I had to relearn the code by character sound and translate that sound directly into an alphanumeric format. The Macrotronics Code Class is a computer program for the Radio Shack TRS-80® Models I and III that teaches code in this manner. You never encounter a dot or a dash — only sounds and their alphanumeric equivalents. To hear the code, you must connect an audio amplfier and a speaker to the TRS-80 cassette audio input. The video display will "tell" you or "test" you on the code that is being sent.

Code Trainer

The code may be sent at speeds of 1 to 1000 wpm. Transmission speed is controlled by pressing the up-arrow and down-arrow keys on the computer keyboard. The program has five functions. A "code trainer" function consists of 11 lessons that drill you, four characters per lesson, in the following manner:

1) The program generates the code audibly for one character.

2) You press the key of the character that the sound represents.

3) If you are correct, the program tells you so both visually and aurally.

4) If your entry is wrong, you are visually and aurally informed and you hear what your guess actually sounds like (instant feedback). You are then retested for the incorrectly guessed characters. As you progress, you can include all the characters from previous lessons in the current lesson, or you can limit the drill to the four characters in that particular lesson.

Code Practice

Three of the other functions, the "code practice" portion, fill the video display with random words, random alphanumerics (in five-letter groups) or random call signs. The program generates the code for all of the characters on the screen at the speed you desire, and you attempt to copy the code. When the program has generated the full screen of characters, you can check your written copy against the display to see how well (or how poorly) you did.

The fifth program function allows you to send cw to the computer to check the quality of your sending. Whatever you send is displayed on the screen, allowing you to see how others copy your "fist." The program also displays your sending speed. (To use this fifth function, you must have Macrotronics interface — models M80, M83, CM80, CM83, TM80, TM83 or Terminall — connected to the TRS-80 to provide interface for your key, bug, etc., to the computer.)

Evaluation

The program runs flawlessly; I found no bugs. Documentation (a nine-page manual) is good. The program loaded from diskette perfectly (it is available in both disk and tape version for the TRS-80 Models I and III). This program teaches Morse code in a logical manner and can be adapted for classroom code learning.

Code Class is produced by Macrotronics, Inc., 5125 N. Golden State Blvd., Turlock, CA 95380. The cassette version costs $29; the disk version costs $39 (when ordering, specify Model I or III). — *Stan Horzepa, WA1LOU*

COMMUNICATIONS SPECIALISTS SS-32M CTCSS ENCODER

☐ When the ICOM IC-3AT came in for review and was assigned to me, I had only one question, "How am I ever going to get a CTCSS (PL) encoder inside that tiny little thing?"[2] Some of the people on W1NI/R had already purchased IC-3s and installed encoders, so I knew it was in the realm of the possible. Upon callng ICOM, I was told they did not have a CTCSS option for the IC hand-held series, but they did recommend the Communications Specialists SS-32M. Communications Specialists happily supplied us with an SS-32M for use with the IC-3AT.

The problem was the installation — a miniature encoder must fit exactly into a miniature space inside a miniature transceiver. Finding my medicine cabinet devoid of "dauntless technician" pills, I called Arnie Chase, WA1RYZ, and asked if he would help. Sure, he had already installed several in other IC-3s.

Arnie performed the surgery with minor assistance from me. The encoder nests in a

[2]Product Review, *QST*, February 1982.

hollow spot between the two circuit boards — it is critical that the encoder board be positioned properly. Instructions provided with the SS-32M give detailed installation information for any of the ICOM hand-helds. Three wires coming from the encoder must be routed to the proper circuit board areas. Communications Specialists suggests modifying the IC-3AT by adding a 4.7-μF capacitor (supplied) in parallel with C13 (2.2 μF) on the PLL. Arnie felt it was easier to remove C13 and replace it with a 6.8-μF capacitor. A subminiature 100-kΩ potentiometer (included) serves as a deviation control.

Circuitry and specifications of the SS-32M are similar to the Communications Specialists TE-64;[3] the SS-32M is designed for the 32 standard CTCSS tones only and does not have provisions for generating the audible tones. Frequency selection is accomplished by grounding pins on the single IC with solder-bridge jumpers. A programming chart provides information on which pins should be grounded to produce a specific CTCSS frequency. For example, the code for 100.0 Hz is 01011 (the sequence represents pins 10, 11, 12, 13 and 14); pins 10 and 12 are grounded with the jumpers, and pins 11, 13 and 14 are left unconnected. Any of the other 31 standard tones can be selected by grounding different pin combinations.

Before starting the installation, I estimated it would take about 20 minutes to complete the job. It took closer to 60 minutes, and Arnie said it was typical. If you are not intimidated by the thought of working on miniaturized equipment and you have some experience, it is reasonable to expect a similar installation time. If you are hesitant to tackle the installation, I suggest you turn the project over to the service department of a qualified dealer (if you don't have a friend like Arnie).

Price class is $30 (without installation). Additional information can be obtained from Communications Specialists, Inc., 426 W. Taft Ave., Orange, CA 92665. — *Peter R. O'Dell, KB1N*

[3]Product Review, *QST*, September 1980, p. 41.

Bearcat 100

Manufacturer's Claimed Specifications	*Measured in ARRL Lab*
Sensitivity for 12-dB SINAD:	Sensitivity for 20-dB quieting:
Low band — 0.6 μV	33.760 MHz — 0.85 μV
High band — 0.6 μV	138.150 MHz — 1.2 μV
Uhf — 1 μV	406.125 MHz — 1.8 μV

BEARCAT 100

☐ The Bearcat 100 is a 16-channel hand-held programmable scanner featuring coverage of the 30-50 MHz, 138-174 MHz and 406-512 MHz bands. It is about the size of a typical amateur hand-held transceiver 7 × 3 × 1-3/8 in. (HWD) and weighs approximately 1 pound[4]. It has a liquid crystal display and a two-second scan rate, and is powered by six AA NiCd batteries.

In addition to the scan feature, which will sample up to 16 discrete frequencies that you have entered into the unit's memory, the Bearcat 100 has a search mode. It will sample all frequencies within a specified range of frequencies to locate signals that would otherwise be undetectable. All you have to do is enter the lower and upper search limits.

A logical question is, "Why would an amateur want to have a portable scanner?" There are at least four good reasons, not all of which might apply to you. I live within range of at least a dozen 2-meter repeaters. With the Bearcat 100, all of them, plus simplex frequencies, can be monitored with ease. For those who travel and frequently find themselves in strange cities, the search feature can be used to locate active repeaters — even those not listed in the *Repeater Directory*. Police, fire and other public service channels can be monitored, potentially improving response time in emergencies. Of course, scanning can be fun, too.

One disadvantage of the Bearcat 100 is that the front-panel pushbuttons have a distinct lack of tactile feedback; they fail to give you a positive indication that the data has been successfully and accurately entered. You'd best keep your eye on the display. This can cause problems at night or when driving. The unit comes with a heavy-duty carrying case with a belt loop. You can't access the pushbuttons while using the case unless you use a razor blade to cut out a suitable opening. All in all, if you feel you have use for a portable scanner, the Bearcat 100 packs a lot into a small package. Price class: $450. Available from Electra Co., Div. of Masco Corp. of Indiana, 300 East County Line Rd., Cumberland, IN 46229. — *Hal Steinman, K1FHN*

[4]kg = lb × 0.454

New Books

☐ *Ferromagnetic Core Design & Application Handbook*, by M. F. "Doug" DeMaw. Published by Prentice-Hall, Inc., Englewood Cliffs, NJ. First edition, 1981. Hard-bound, 6 × 9 inches, 256 pp., $24.95.

There is hardly a circuit today — from the milliwatt QRP rig to the most complex computer — that does not employ a ferromagnetic-core device of one kind or another. Are you one of those who simply wondered what these modern-technology devices are all about? Or are you the experimenter and "homebrewer" who has collected folders full of manufacturers' data sheets for toroids, rods and pot cores, and who prayed for a comprehensive reference manual? Cheer up! Doug DeMaw, W1FB, who has contributed so much to the pages of *QST* as Senior Technical Editor, has written an outstanding book on the subject. It has all the ingredients to become the standard reference manual on ferromagnetic-core devices for amateur, technician, student and engineer alike in the years to come.

The 256 pages of this handsome hard-cover book, comprising five chapters and appendix, are well organized and illustrated. The author covers a seemingly complex subject in easy-to-understand language and superb writing style. He refers to basic formulas associated with ferromagnetic-core technology only when essential and keeps cumbersome textbook-type math to a minimum. This handbook does not seem to omit anything, yet is not technically overpowering — going from basic theory and application to proper selection and use of design concepts. For a layman such as I, it is especially gratifying that the practical aspects of ferromagnetic cores are covered so well for amateur and professional alike. For instance, the author's treatment of ferromagnetic-core baluns is most valuable and suitable to the amateur. Likewise, the reference table for standard core sizes of ferromagnetic material will be a real asset to anyone trying to wind a coil, choke or transformer. The circuit examples given are practical, lab-researched and proven.

The many topics presented in this book defy individual listing. Rods, bars, slugs, beads, sleeves and pot cores — all are dealt with in a "hands-on" manner. Sample circuits are added throughout. The appendix is full of valuable references, including IEC publications, magnetic-core symbology, manufacturers' names and locations, core size and selection charts — just to mention a few. Ferromagnetic-core technology is ably presented in this remarkable book. — *Hans J. Meurer, W2TO*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QST

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OUR COVER
Ulrich Mueller, DK4VW (left), and Gordon Hardman, ZS1FE/KE3D, work on the Phase IIIB satellite during spacecraft integration activities at AMSAT DL, Marburg, West Germany, October 1982. *(photo courtesy AMSAT DL)*

TECHNICAL

BEGINNER'S BENCH

NEWS AND FEATURES

OPERATING

DEPARTMENTS

Long Life for Your Transmitting Tubes

Here are some "housekeeping" rules that will help you achieve maximum tube life.

By William I. Orr,* W6SAI

Transmitting tubes represent a significant proportion of the money a radio amateur spends for his or her station, so it's prudent to reduce the cost-per-hour of operation by getting longer life from the tubes. Abuse of the power tube can be costly. Common sense tells the user to operate the tube in the manner recommended by the manufacturer. This article provides some insight into tube operation and provides information that will allow the operator to gain the maximum life from power tubes.

A data sheet covering operation of a specific tube type is available from the manufacturer.[1] It outlines the maximum ratings and provides typical operating conditions for the tube. Maximum ratings must never be exceeded, but some latitude exists in the typical operating conditions. The data sheet is a good starting point in the search for long tube life.

What is the Life of a Power Tube?

If a large number of power tubes of a given type are run in a life test, tube "death" will follow the same mortality curve as that used by insurance companies to determine policy rates. Thus, while it is impossible to determine when a particular tube will fail, it *is* possible to ascertain the *average* life of a number of tubes, which can be extended by proper care. The situation is akin to the Old-Timer who, after taking a physical examination, told the doctor, "If I knew I was going to live this long, I would have taken better care of myself!"

All things being equal, the better care an operator takes of a power tube, the longer it will last. It is possible to extend the life expectancy of an individual tube, just as clean living will help to extend *my* life expectancy (or so I have been told!).

Well-designed power tubes are very forgiving of abuse — more forgiving, for example, than TV type "sweep tubes" or rf power transistors. Nevertheless, the ultimate life of a particular power tube in a given piece of equipment depends on the care and expertise of the operator. Even in the best equipment, tube life can be shortened by the operator who believes in "all knobs to the right!" Tube life can be curtailed by the operator who does not read the instruction manual and doesn't know how to operate equipment properly. Heat is the enemy of long tube life, and an amplifier operated in an off-tune or overloaded condition can quickly damage the power tubes therein.

Filament-Voltage Management of Thoriated-Tungsten Tubes

Certain tubes, such as the 3-500Z, 4-1000A and others have a thoriated-tungsten filament. The filament is processed and heated in the presence of a hydrocarbon to produce ditungsten carbide on the surface of the wire. Life is proportional to the degree of carburization and the filament operating tempera-

[1]Information on most EIMAC tubes and other products may be obtained by writing: Application Engineering Department, Varian/EIMAC, 301 Industrial Way, San Carlos, CA 94070. Data sheets for the 3-400Z, 3-500Z, 3-1000Z and 8877 tubes may be obtained from Varian/EIMAC, 1678 South Pioneer Rd., Salt Lake City, UT 84104.

* c/o Varian/EIMAC, San Carlos, CA 94070

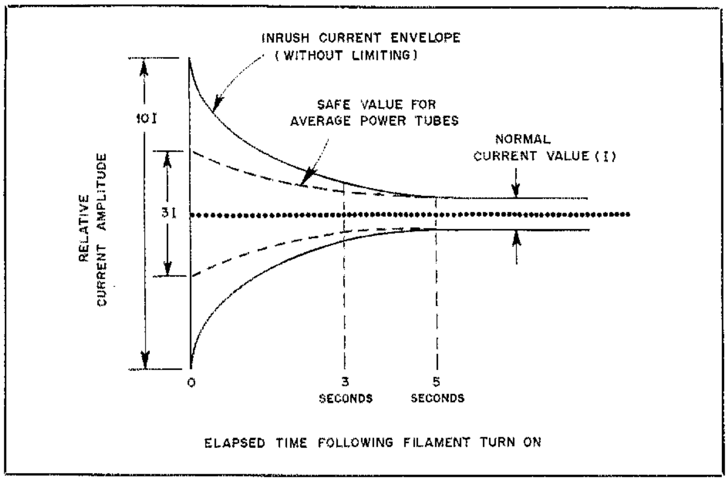

Fig. 1 — Graphic representation of filament inrush current. With no protection, filament inrush current may be as high as 10 times the normal value, for up to 5-seconds' duration (with tubes of the amateur power level). Most tubes are easily capable of withstanding three times the normal filament current; this is a good limiting value. Amateur type tubes can be run at 60% of nominal filament voltage for two to three seconds, after which the voltage is raised to nominal value. Filament voltage may also be brought up to nominal value by means of a variable-voltage transformer.

Table 1

Suggested Primary-Circuit Resistor for Limiting Filament Inrush Current

Tube Type	*Time Delay (Sec.)*	*Resistor*
3-500Z, 4-400A	2	50 Ω,50 W
2 × 3-400Z, 2 × 3-500Z, 4-1000A	2	25 Ω, 50 W
8873/4/5, 4CX250B	4	150 Ω, 50 W
2 × 8873/4/5, 2 × 4CX250B	4	75 Ω, 50 W
8877, 4CX1000A	4	75 Ω, 50 W

ture. As the tube is used, the filament slowly decarburizes and finally is no longer an effective electron emitter.

The key to extending the life of a thoriated-tungsten filament is temperature control, which is a function of filament voltage. For most power tubes a nominal value is given on the data sheet, plus recommended variation limits (usually ±5% of the nominal voltage). The danger of operating the filament in excess of the high limit is that when electron emission is accelerated, the filament decarburizes more rapidly and life is shortened. The danger of operating beyond the lower limit is that a "cool" filament is in danger of being "poisoned" by gas ions in the tube (no vacuum is perfect), or by contaminants forced out of the other tube elements. These can be released during a period of high-temperature operation brought on by an overload.

Operation of the filament within the ±5% voltage range cannot be done without an accurate filament voltmeter. If your amplifier has such an instrument, its accuracy should be checked. If no internal meter is present, the filament voltage can be checked with an external instrument, and a correlation made between line voltage and filament voltage. Once this is done, it is only necessary to check the line voltage to know the filament voltage. Holding the filament voltage within the recommended range will pay big dividends in extended tube life!

Filament Voltage Management of Oxide-Cathode Tubes

Some smaller transmitting tubes, such as those in the 4CX250 family, the 8873/4/5 and the 8877 employ an indirectly heated cathode emitter instead of a thoriated-tungsten filament. The cathode is coated with a barium-strontium emitter that is brought up to temperature by a heater located within the cathode can. These tubes require a period of time for the cathode to reach operating temperature, and potential should never be applied to the tube until the cathode temperature is stabilized. Warm-up time is noted on the data sheet for each tube type.

As with the filament type tube, it is important to hold the heater of the oxide-cathode tube within the voltage limits set by the manufacturer. In the majority of cases, it is ±5%. Exceeding the recommended heater voltage in these tubes will tend to reduce tube life, and falling short of it will severely restrict electron emission.

Filament Inrush Current Protection

As in the case of the common light bulb, the filament or heater of a cold power tube has about one-tenth the resistance of a hot filament. Thus, at the instant of turn-on, inrush current can be up to 10 times the normal amount, until filament temperature rises to the proper value (Fig. 1). This large current surge overloads the filament structure and also creates a strong magnetic field. This field *can warp the filament and grid structures* in a very large power tube! Filament inrush current can be limited to some extent by the filament transformer, which should never be larger than necessary to do the job.

Simple inrush current protection circuits are shown in Fig. 2. Any of these may be retrofitted into an amplifier. A variable-voltage transformer (Fig. 2A) in the primary circuit is a practical solution. Before the amplifier is turned on, the transformer is set at zero volts. Next, the amplifier is turned on and the filament voltage is brought up to normal. This simple operation need only take two to four seconds. With practice, the operation will almost become automatic.

Another effective inrush current protector is shown in Fig. 2B. A series-connected current limiting resistor is placed in the primary circuit of the filament transformer, or in the ac line. Once the power switch is thrown, the resistor is shorted out after two to four seconds. A time-delay relay or simple shorting switch may be used. The resistor should limit the filament voltage to about 70% of the nominal value (Table 1).

Simpler yet is a shorting type (make-before-break) rotary switch, connected as shown in Fig. 2C. In switch position 1, filament voltage is off. In position 2, voltage is applied through the limiting resistor. In position 3, full voltage is applied to the filament transformer. Passing through all the switch positions takes only a few seconds.

Line Voltage Regulation

A vexing problem to many radio amateurs is primary line voltage regulation. Line voltage can wander about during a 24-hour period, and may drop abruptly when a moderate load comes on the line. Unless the line is "stiff" and can withstand the full amplifier load, filament voltage will drop under peak input power conditions. This places a strain on the power tube filament, as voltage is lowest

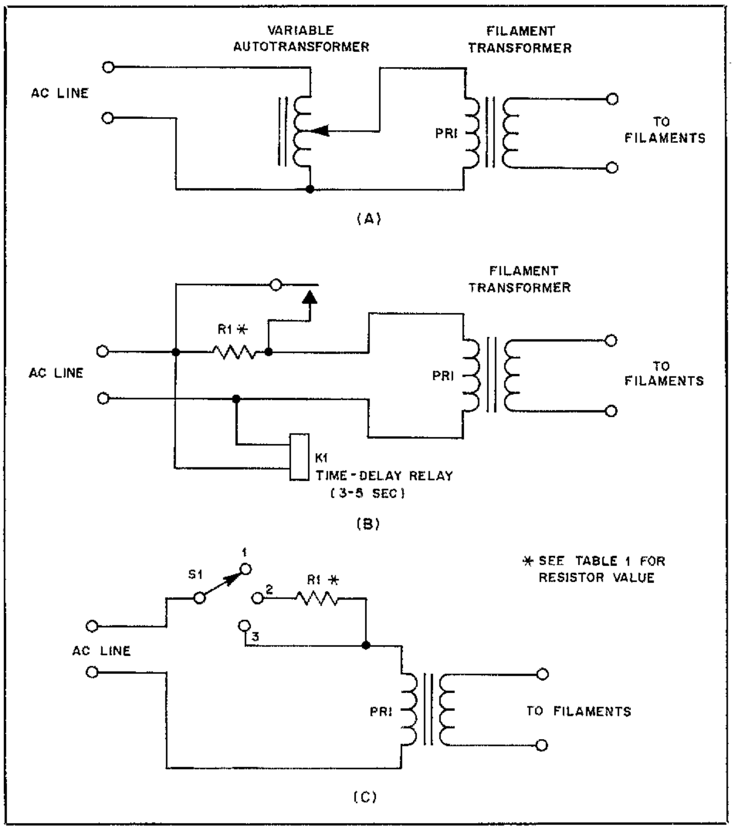

Fig. 2 — A variable-voltage transformer is used to bring the filament voltage up to nominal value over a period of two to three seconds (A). At B, series-connected resistor R1 is shorted out by time-delay relay K1 or a hand-operated switch. In circuit C, make-before-break rotary switch S1 is used to turn on the filament transformer. Switch position 2 introduces the current-limiting resistor.

when electron demand is highest.

On the other hand, filament voltage can rise during periods of low primary power utilization. Line-voltage fluctuations of 10 V or more are common; this is equivalent to a 0.4-V variation on a 5-V filament line powered from a 117-V primary transformer.

When voltage regulation is a problem, it is wise to set nominal filament voltage in the key-up (standby) condition, and permit less than a 5% drop in voltage at maximum power output. This can best be achieved by operating the amplifier on a 234-V line, as opposed to a 117-V line. When long-term voltage variations are pronounced, the only practical solution is to utilize a primary voltage control device, such as a rheostat or variable-voltage transformer, and maintain an "eyeball" check on the filament voltmeter.

Cooling

A high-power amplifier generates plenty of heat. Sources of this are the tube filament, plate and ohmic losses in the amplifier circuitry. Most amateur amplifiers of the 2-kW-PEP class incorporate a ventilation system to move the hot air out of the cabinet. The operator should make sure this system is in proper order and that air intake and exhaust vents are clear to allow free passage of air. Don't place books or pieces of equipment atop an amplifier if the air passes through this part of the cabinet. And don't push the amplifier back against the wall if the air intake is on the rear of the cabinet. Always allow free circulation of air around the amplifier. While manufacturers have been able to miniaturize equipment, nobody has been able to miniaturize the watt. The smaller the cabinet for a given size power amplifier, the more important it is to extract the heat from it. Make sure your equipment runs cool by providing the tubes with an unimpeded flow of cooling air. In a hot climate, or at an elevated altitude (Denver, for example), an additional cooling fan that forces air *into* the amplifier cabinet may help.

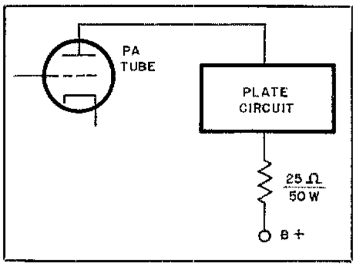

Fig. 3 — A series resistor in the B+ lead protects the tube and related components in the event of a flashover.

Arc Protection

It is possible for a power tube to internally flash over after a period of disuse. Similarly, other components in the plate circuit of an amplifier can flash over on occasion. Since the power supplies in many modern amplifiers use a large filter capacitor, a tremendous amount of energy can pass through a flashover, damaging the tube and associated equipment. A series resistor will dissipate a large portion of this energy, and can be placed in the plate lead to the power tube (Fig. 3). For retrofitting existing equipment, a 25-Ω, 50-W resistor in the B+ lead, *after* the power supply filter capacitor and before the plate rf choke and bypass capacitor, should do the job.

Finally, make sure that the socket contacts for your power tube are clean and make good contact with the tube pins. Also, check that the cooling system is clear of dust and lint. These simple "housekeeping" rules will help you achieve maximum tube life!

Acknowledgments

Thanks to William Barkley, Dick Rasor, W6EDE, William Sain, Bob Tornoe and Bob Sutherland, W6PO (all of Varian/EIMAC) for their help. Also thanks to Robert Artigo, KN6J, formerly of Varian/EIMAC, who prepared a paper on extended tube life for *Broadcast Management/Engineering Magazine* (April 1982). Some of this material has been extracted from that article.

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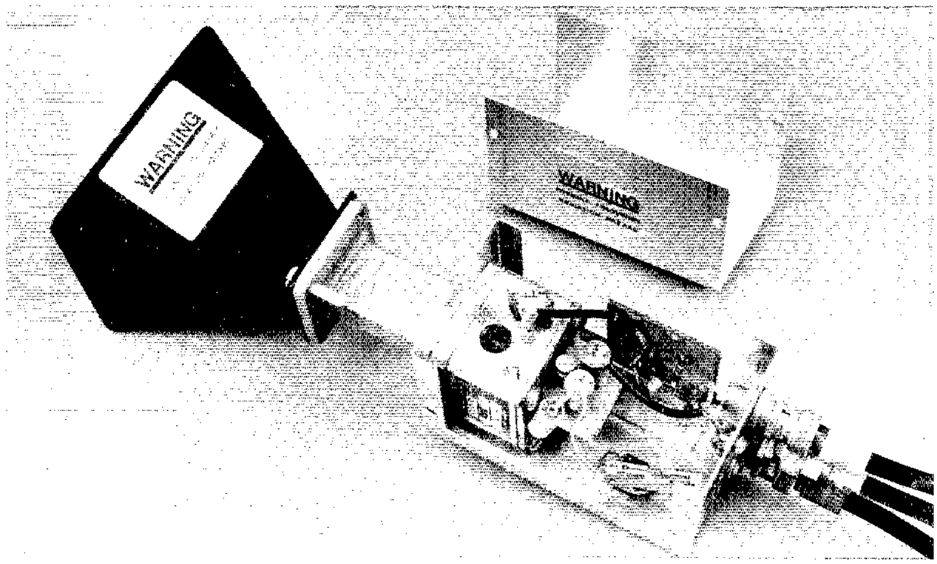
The Care and Feeding of Gunnplexers



Microwaves too difficult? Not so! We'll show you just how easy 10-GHz fm can be.

By Daniel N. Petersen,* WA6OIL

Do you have an adventurous spirit? Do visions of dish antennas pop into your head when you look at your portable barbecue? If so, you should consider "playing microwaves" with a Gunnplexer.[1] The Gunnplexer is a simple, effective way to jump on the 10-GHz bandwagon. In this article, I'll point out some of the pleasures and pitfalls of Gunnplexer operation.

What Is It?

The Gunnplexer, manufactured by Microwave Associates, is a solid-state microwave transmitter and receive converter. It consists of three sections: A Gunn source, a circulator/mixer and an antenna. The Gunn source contains the heart of the system — a Gunn diode. This diode, when placed in a waveguide cavity and subjected to a forward bias, will oscillate at a frequency dictated by the dimensions of the waveguide cavity and by the diode parameters. To put it simply, you apply 10 V dc to the Gunn diode and about 15 mW of 10-GHz "goo" will come out of that rectangular hole at the other end, see?

Also contained in the Gunn source section is a varactor diode. It is used to electrically change the frequency and to frequency-modulate the Gunn source. There is also a mechanical tuning slug that can be used for coarse frequency changes. I do not recommend "fiddling" with that control unless you have had some experience with microwave gear.

The center section is the circulator/mixer. It contains a ferrite circulator and a Schottky mixer diode. The circulator diverts a small portion of the transmit energy to the mixer diode, thus providing a local-oscillator signal for the mixer. A received signal mixes with the local oscillator and appears at the output port, which extends above the circulator/mixer section. The antenna is a 17-dB horn. This type of antenna provides a good amount of gain in a small package. Other types of antennas can be substituted, but for most experimentation the horn is more than adequate.

How Do You Use It?

The Gunnplexer requires very little in terms of operating voltages. A 10-V, 150- to 200-mA dc supply is required for the Gunn diode. This voltage goes directly to the diode with no series resistor. The diode provides the necessary current limiting. Gunn sources are not pillars of efficiency, usually being less than one percent efficient at these power levels. The important thing is that they work well without requiring special support equipment. The varactor tuning diode requires from 1 to 20 V for proper operation. Being reverse biased, the varactor draws virtually no current. You can use a potentiometer as a voltage divider, with a dial on the shaft calibrated in frequency. You can also use a dc voltmeter to indicate your frequency.

[1]Notes appear on page 18.
*Rte. 1, Box 11AA, La Center, WA 98629

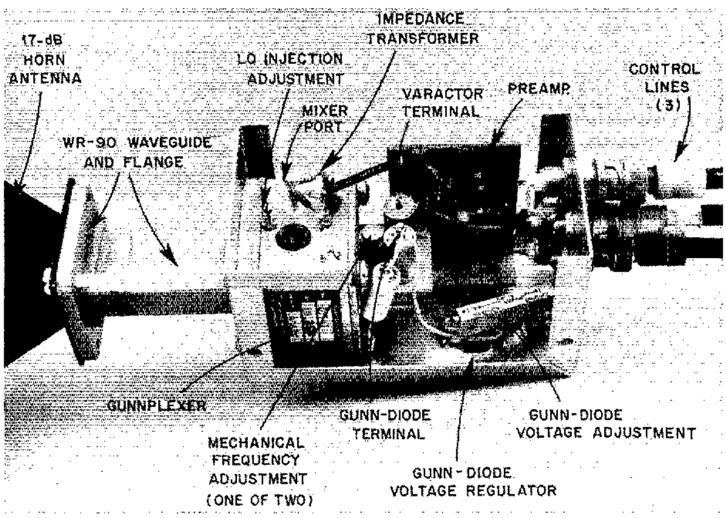


Fig. 1 — This Gunnplexer is housed in a small Minibox. Coaxial cables connect the unit to the i-f receiver and the control circuits. If you mount the horn antenna directly to the Gunnplexer enclosure, the waveguide section can be eliminated.

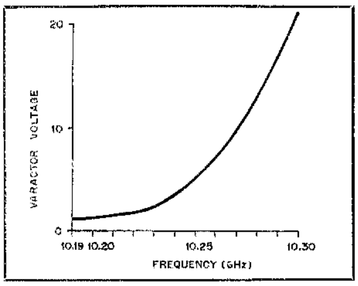


Fig. 2 — Gunnplexer frequency versus varactor-diode voltage. This graph is for a Gunnplexer that is factory tuned to 10.250 GHz at a varactor potential of 5 V.

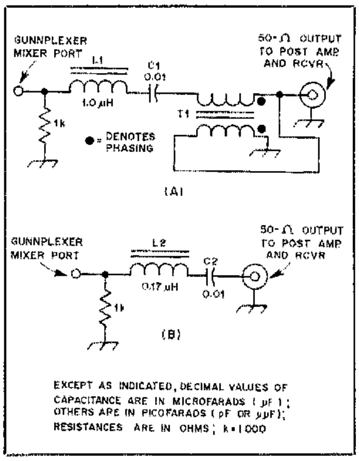


Fig. 3 — Gunnplexer i-f port matching circuits. The circuit at A is used to match the i-f port to 50 Ω at 30 MHz. Circuit B can be used to provide a 50-Ω output impedance in the 90-MHz range. T1 consists of 7 bifilar turns wound on an Amidon FT23-75 core.

Table 1
Mixer Port Impedances

Freq. (MHz)	*Z (ohms)*
10	355 – j 102 = 370 Ω
20	286 – j 165 = 330 Ω
30	214 – j 180 = 280 Ω
40	147 – j 164 = 220 Ω
50	123 – j 176 = 215 Ω
60	98 – j 157 = 185 Ω
70	75 – j 141 = 160 Ω
80	61 – j 126 = 140 Ω
90	56 – j 106 = 120 Ω
100	43 – j 106 = 115 Ω

I have been speaking "pleasure" so far, so here are some precautions. Treat that mixer port with a great deal of respect! The mixer diode is subject to damage from static electricity. Walk across a carpeted floor and touch the mixer port, and you could have a dead diode. Microwave Associates places a protection diode on the mixer port before shipment. Since this diode is removed before the Gunnplexer is placed in operation, you can't be too careful about avoiding static discharge. A good plan is to attach a clip lead to the mixer port and the Gunnplexer body while it is not being used, or when you are soldering to the mixer port.

Watch the Gunn- and varactor-diode power-supply polarities. I have inadvertently connected the Gunn diode backwards with no apparent damage to the diode (I hope), but I would not recommend doing it. Not only does it give you gray hair when you notice it's hooked up backwards, but having the Gunn diode changed is not cheap! As long as you have positive voltages on the Gunn and varactor ports, you'll be okay.

Mechanical Mounting

For better utilization, the Gunnplexer should be mounted in an enclosure. A 2-1/4 × 2-1/4 × 4-inch Minibox makes an ideal housing (Fig. 1). It provides plenty of room for the Gunnplexer, the Gunn-diode voltage regulator, an i-f preamplifier and connectors for signal and control lines. Because of the small size, the unit can be mounted on a tripod or at the focus of a parabolic dish, with cables going to the i-f and control circuits.

Another reason for mounting the Gunnplexer in a separate housing is that Gunnplexers tend to be somewhat microphonic. I discovered this while on an outing. I was making some adjustments in the Gunnplexer housing and as usual, I was talking to myself. My friend, at the receiving end, told me he could hear me just fine, even though the modulator was turned off. Tapping the case produced a healthy "bonging" noise at the receiving end. I would be interested to know if others observe this phenomenon.

Just When You Thought It Was Safe

Now that the preliminaries are out of the way, things get a bit more complex. As I said before, the varactor requires from 1 to 20-V dc. Before the Gunnplexer is shipped from the factory, a frequency is stamped on the body. This is the transmit frequency at a varactor voltage of 5 V. A change in this voltage produces a nonlinear change in frequency. The shift is much greater at low varactor voltages than it is at high voltages (Fig. 2). The output power also fluctuates with changes in frequency. I have measured a maximum variation of 0.6 dB. This variation should not cause a great deal of concern.

The next subject to be dealt with is the mixer port. Microwave Associates implies that it has an impedance of approximately 200 Ω. Since this declaration is somewhat vague, I connected the mixer port to an impedance analyzer. I found that the complex impedance varies quite a bit with frequency and that the mixer port is capacitive (Table 1). At 30 MHz, which is pretty much a standard Gunnplexer i-f, the transformation to 50 Ω is fairly simple. A matching network is shown in Fig. 3A: L1 cancels the capacitive reactance at the mixer port and dc blocking is provided by C1. T1 produces the required 4 to 1 impedance transformation. I tested the network with a Gunnplexer and came up with the results shown in Table 2. This network appears to work well with both of my Gunnplexers.

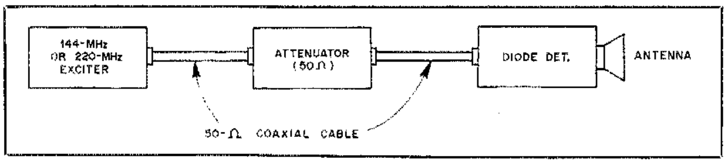


Fig. 4 — Frequency calibration setup. The attenuator should reduce the power applied to the detector diode to approximately 250 mW or less.

Table 2

Fig. 3A Matching Circuit Output Impedances

Freq. (MHz)	*Z (ohms)*
20	69.1 − j 11.0 = 70 Ω
25	60.7 − j 6.4 = 61 Ω
30	54.0 − j 0.94 = 54 Ω
35	48.4 + j 7.7 = 49 Ω
40	42.7 + j 17.2 = 46 Ω

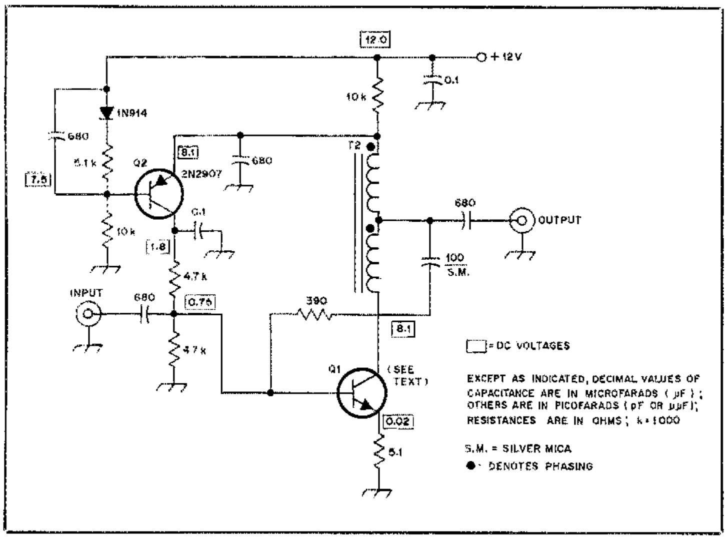


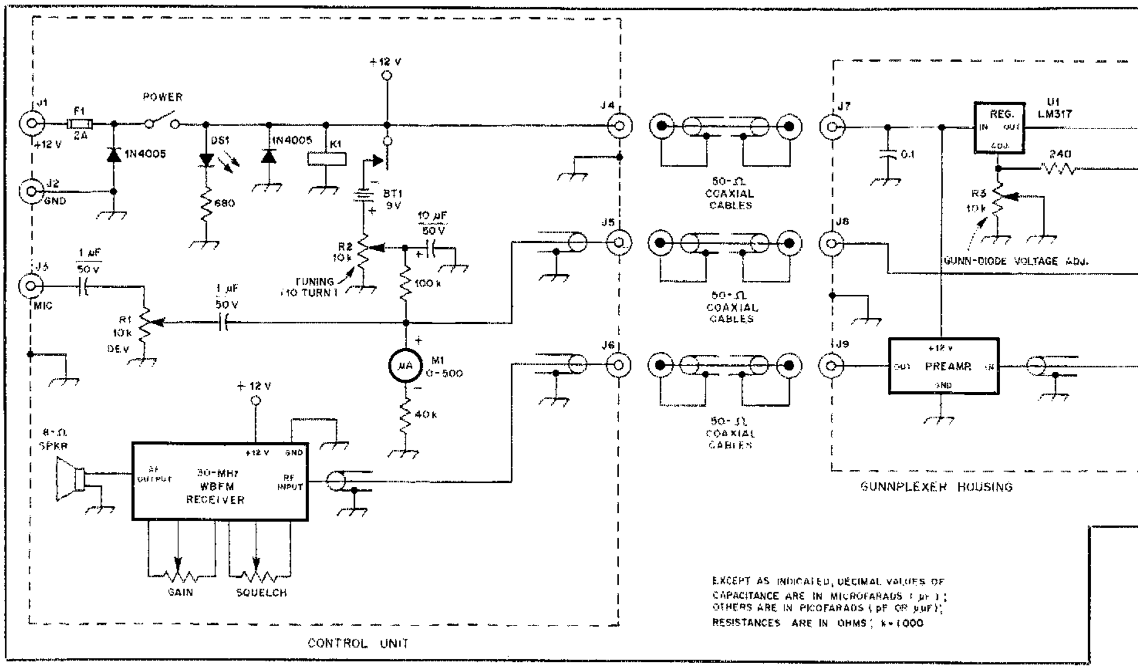
Fig. 5 — I-f preamplifier. All resistors are 5%, 1/4-W carbon types. Unless indicated otherwise, capacitors are disc ceramics. T2 consists of 7 bifilar turns wound on an Amidon FT23-75 core.

Tuning Considerations

The amateur X-band allocation is 10.0 to 10.5 GHz. Any modulation mode, with the exception of pulse modulation, is allowed. Gunnplexers can be ordered for operation anywhere in the band, but most hams prefer operating near 10.25 GHz. There are two reasons for this. First, sitting at the center of the band gives you lots of room to roam before you run into a band edge. The second reason is that calibrators, used to help you find your receive frequency, are simple to make for frequencies near 10.25 GHz.

So Where Are You?

The simplest calibrator is a waveguide type of diode detector (Fig. 4).[2] If you apply enough rf power to the diode, causing it to conduct, it will act as a multiplier. With a careful choice of frequencies you can place a harmonic at a favorable spot in the 10.25 GHz area. For example, the 70th harmonic of 146.52 MHz falls at 10.2564 GHz. A 220-MHz source can also



be used for harmonic generation. The 46th harmonic of 223.5 MHz provides a calibration signal at 10.281 GHz. I have used both a 144- and a 220-MHz rig for calibration sources. A word of caution, however: Please pad your excitation source. The idea is to turn the diode on and off with a volt or so of rf potential. Applying 10 W to a microwave detector diode will launch the diode into the "Twilight Zone" in a cloud of gaseous silicon. You should be able to detect the calibrator signal at a distance of several feet.

I-F Receivers

It has been generally accepted that a 30-MHz i-f is the standard for amateur 10-GHz work. If it is just you and someone else who are "X-banding," however, you can bend convention a bit. For your first experiments you can use a standard fm-broadcast receiver. Since Gunnplexers are easily modulated to produce wide-band fm, it seems only natural to use an fm broadcast set as a ready-made i-f receiver. The circuit shown in Fig. 3B can be used to approximately match the Gunnplexer mixer port to a 50-Ω line at frequencies within the fm broadcast band.

Commercial i-f receivers are available or you can "roll your own" receiver by designing it from scratch (a pain) or by P and M engineering (P and M, by the way, stands for Plagiarism and Modification). The ARRL *Handbook* contains an

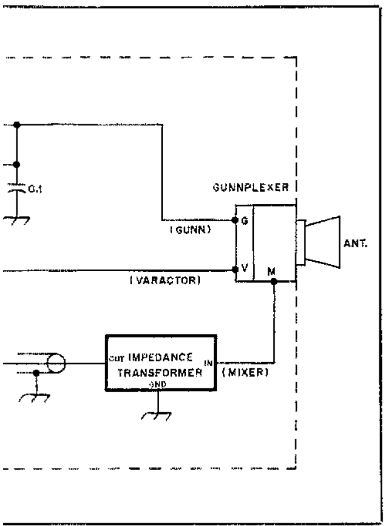


Fig. 6 — Gunnplexer control circuit. An interior view of the Gunnplexer housing unit is shown in Fig. 1. Any wide-band 30-MHz fm receiver can be used as the i-f receiver (see text). The remainder of the control-unit circuit supplies the varactor tuning voltage, Gunn-diode voltage and the modulation audio to the Gunnplexer.

fm receiver circuit that can be modified for 30-MHz wide-band operation.

The primary concern is the 10.7-MHz filter. For this mode, the filter has to be wide — in excess of 100 kHz or more is ideal. As luck would have it, Radio Shack has in their parts trove a 10.7-MHz ceramic filter that has a 280-kHz bandwidth for a low price. Why use such a wide filter? For tuning considerations mostly. If you had a narrow-bandwidth receiver, say 5 kHz, and were trying to tune in a modulated oscillator that drifts a bit, all you would hear is an occasional "pop" as you tuned across it or it drifted past your frequency. Also, you'll be impressed by the fidelity of wide-band fm. Your partner will sound exactly like your partner and not Donald Duck at the bottom of a well. [Later editions of the ARRL *Handbook* contain a 30-MHz wide-band fm receiver circuit designed specifically for use with Gunnplexers. — Ed.[3]]

Frequency Stability, or Do You Catch My Drift?

In the last section, I made mention of the Gunnplexer being "a modulated oscillator that likes to drift a bit." This can be construed as an understatement, for the Gunnplexer is an unstabilized oscillator. For example, the VFO in an hf rig should maintain a stability of one part in a million, resulting in a drift of 7 Hz in the 40-m band. The Gunnplexer, on the other hand, has a stability of one part in ten thousand (or 100 parts per million). This results in a drift of 1 MHz at 10,000 MHz. It may not sound like much, but an hf VFO with the same drift at 40-m would move 700 Hz, and that would be unacceptable. Don't despair, for we are not trying to receive cw or ssb with the unit — just wide-band fm.

Some people think that a Gunnplexer cannot be used without automatic frequency control (afc). It is my contention that, for experimentation over short distances, afc is not required. If you and your "partner" want to go hilltopping you should have no problem finding and tracking one another's signals without afc. I agree that for long-term reliability, afc or some sort of phase-lock system should be used.

The Gunnplexer drifts, but don't get the idea that it goes helter-skelter all over the band. When first fired up the Gunnplexer will drift at a terrific rate, slowing down after a minute or two. After five minutes of operation it will have settled down to a very slow drift, caused mostly by temperature changes. If both units are turned on at the same time they will both drift in the same direction at about the same rate. This will minimize the need to chase one another's signals. Some useful advice when operating without afc: Only one of you should tune. It's like hide and seek. One operator "hides" in one spot while the other one "seeks." The "hunter" shouldn't have any trouble finding his "prey." If you are only a few miles apart, tuning in your partner's signal will be as easy as tuning in a commercial fm broadcast station. It's easier, in fact, because there's only one station on the band.

Add an Afterburner

In this section we deal with an i-f preamplifier for your Gunnplexer. The circuit (Fig. 5) is essentially foolproof. It's easy to build and contains fairly common components. Q1 is the amplifier transistor while Q2 is part of the dc biasing circuit. A feedback network, extending from the base of Q1 to the output coupling capacitor, keeps the amplifier gain constant from below 10 MHz to over 90 MHz. The input and output impedances are near 50 Ω.

This circuit can be used in many applications. I find it to be unconditionally stable. If a low-noise transistor is used for Q1, the circuit should work very well. My preamplifiers, using N.E.C. (Nippon Electric) devices, have noise figures (NF) of about 3.5 dB.[4] That isn't great, but, with the Gunnplexer noise figure at about 12 dB or so, a 3.5-dB NF preamplifier will surely aid the system noise figure.

Controlling Your Gunnplexer

The Gunnplexer is simple to control. Fig. 6 contains a schematic diagram of the control system I use on my Gunnplexers. It's in two boxes: the control/receiver unit and the Gunnplexer housing. The control/receiver unit includes the main power control, varactor tuning circuit, modulator and receiver.

When the main switch is turned on, power is supplied to the Gunn diode, the modulator and the receiver. K1 is also energized, enabling the tuning circuit. The 9-V battery "rides" on top of the 12-V line to provide 21 V for varactor tuning. The modulator is a crystal microphone and gain control, with the output connected to the varactor control line. This impresses a small audio-modulation voltage on the varactor. I set the audio level by listening to the signal, adjusting R2 until over-deviation occurs. I then back off the setting of R2. It doesn't take much audio to *really* fm the Gunnplexer. [Some crystal microphones will require the use of an amplifier as shown in note 3 — Ed.]

I have left the receiver question open. You may use a bona fide 30-MHz receiver or a commercial fm-broadcast receiver. I have been using a homemade 30-MHz receiver that is still in the development stage. I hope to treat this receiver in a future article.

The second unit is the Gunnplexer housing. It contains the Gunnplexer, a regulator for the Gunn-diode voltage and an impedance-matching/preamplifier circuit. I have used an LM317 as the Gunn-diode voltage regulator with good success. It provides precise control of the diode voltage — an important parameter for good oscillator stability. Another reason for a local regulator is that it allows you to operate the preamplifier at 12 V. The preamplifier, although not necessary, is an advantage. I would not compromise the impedance-matching circuit. Without it, you would see, at 30 MHz anyway, a 4 to 1 mismatch to your 50-Ω system. The varactor line simply conveys the varactor dc and modulator audio to the varactor diode. In my units I use BNC connectors and coaxial cable for all three connecting lines. You should use coax for the i-f line, but shielded cable can be used for the dc lines. In fact, Fig. 6 is meant only as a guide. It is one ham's attempt to "ride herd" on the Gunnplexer.

An FM Pocket Radio I-F Receiver

For casual Gunnplexer experimentation, a low-cost fm-broadcast receiver can be modified easily for use as an i-f unit. One such receiver is the Realistic Model 12-714 sold by Radio Shack.

The first step in converting this broadcast set into an i-f receiver is to provide for a 50-Ω input. This will allow the i-f receiver to be connected to the Gunnplexer mixer through 50-Ω coaxial cable. Shown in the accompanying diagram is the input circuit of the Model 12-714. The fm antenna is connected to the input tuned circuit (C2 and L1) through a 10-pF capacitor. All you need to do to modify this circuit for use with 50-Ω cable is to remove the wire going to the fm antenna and replace the 10-pF capacitor with an 18-pF unit. The 50-Ω cable can then be attached to the two terminal lugs at the lower right-hand corner of the receiver circuit board. After the receiver has been attached to the Gunnplexer, this circuit can be adjusted for maximum sensitivity by spreading or compressing the turns of L1. The tuning is broad and the adjustment is not critical.

To reduce i-f feedthrough, the receiver should be enclosed in a metal box and all leads entering the enclosure should be filtered. A ferrite bead (Amidon FB101-43) and a 0.001-μF disc-ceramic capacitor on each lead will provide the needed filtering.

Note that this receiver is designed for use with a positive ground. That is, the positive battery terminal is the common circuit point. Because of this, it is difficult to power the receiver from the 12-V supply used to power the Gunnplexer. Using a separate battery (with the positive terminal connected to ground) to power the receiver avoids this problem.

To complete the Gunnplexer transceiver, an audio stage, to modulate the varactor voltage, is required. The simple circuit shown in note 3 will serve this purpose well.

Those wishing to experiment with afc will find a suitable voltage at the junction of C27 and R28 (refer to the schematic diagram that accompanies the Model 12-714). The afc voltage at this point swings approximately 0.4 V either side of the +9-V receiver reference level (the *transceiver* ground point). A simple op-amp circuit, powered from the Gunnplexer supply ($+V_{cc}$) and the receiver battery ($-V_{cc}$), can be used to provide the required amplification and level shifting.

Remember that while the exact i-f used is not critical, it is necessary that the transmitting and receiving units use the *same* i-f. The i-f receiver should be set to an agreed-upon frequency and the varactor-voltage control used for tuning. While the use of 30 MHz as the i-f is highly recommended for general Gunnplexer operation, this simple broadcast set modification will provide experimenters with a low-cost way to "try their hand" on the 10-GHz band.
— *KC1V*

Where are They Sold?

A few phone calls to the New England area revealed the following information: Microwave Associates does not sell Gunnplexers directly or disseminate technical information about them. They sell the Gunnplexer through a single distributor, Advanced Receiver Research.[5] The 10-mW unit is type MA 87141-1 and is in the $250 per pair price class. Advanced Receiver Research also sells support equipment for the Gunnplexer.

If you have trouble locating parts to build this project (or others), you might give the folks at Circuit Specialists, Inc. and Jameco Electronics a try.[6,7] I have done business with Circuit Specialists for years. They have been prompt in replying and are usually well stocked. Jameco Electronics has a complete catalog with a wide variety of parts.

I have found that local Radio Shack stores have some surprising items. For instance, they sell 10.7-MHz wide-band filters (200 kHz wide) that can be used in wide-band fm receivers. Two filters cost $1.99. Consulting the various Amateur Radio publications will net you some other sources, but those I've mentioned are my favorites.

I have found the Gunnplexer to be very satisfying to experiment with. The system and information presented here is meant to stimulate interest in Gunnplexer operation. Don't worry about afc or phase locking yet; get on the air first! Once you have the ship launched and your feet wet, you can add the bells and whistles. Happy Gunnplexing!

Notes

[1]Gunnplexer is a registered trademark of Microwave Associates, Inc., Burlington, MA 01803.
[2]S. J. Noll, "X-band Calibrator," *Ham Radio*, April 1981, pp. 44-50. [This article gives construction details for a waveguide-type calibrator using a 1N23 diode — Ed.]
[3]G. Woodward, ed., *The Radio Amateur's Handbook*, 60th ed. (Newington, CT: ARRL, Inc., 1983), pp. 14-24 to 14-26.
[4][An NE41632E-2 or a similar device is suitable — Ed.]
[5]Advanced Receiver Research, Box 1242, Burlington, CT 06013, tel. 203-582-9409.
[6]Circuit Specialists, Inc., Box 3047, Scottsdale, AZ 85257.
[7]Jameco Electronics, 1355 Shoreway Rd., Belmont, CA 94002.

Strays

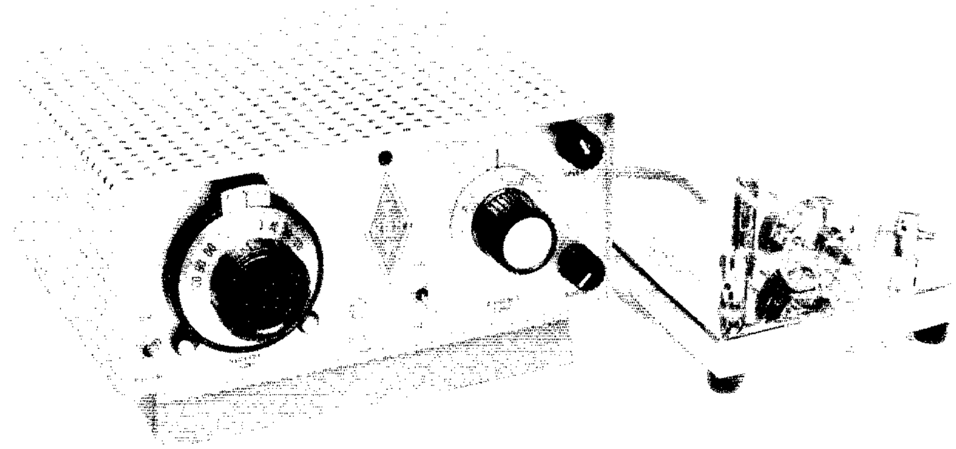
Terry A. Ketron, WA4RJN, of Boone, North Carolina, used a crude but effective means to dispose of a CB linear amplifier he tried to sell at the Shelby Hamfest. The Norfolk, Virginia, FCC office sent WA4RJN a citation warning him of the penalties and prohibitions regarding the sale of such devices, and ordered him to show proof the equipment was destroyed. This photo served as part of the proof. (*photo courtesy FCC*)

Next Month in QST

Those who read May *QST* will learn about

- Vic Clark, W4KFC, and Dave Sumner, K1ZZ, who speak out on a wide range of issues facing the Amateur Radio Service and ARRL.
- digital logic, and how it relates to electronics — a timely Beginner's Bench article.
- the weather conditions that can cause a routine 2-meter contact to make the pages of "The World Above 50 MHz" in *QST* and 10 new states.
- the rules governing Field Day 1983.

Putting the "8P6 Special Hamcation Rig" on 10 MHz



No gear for 30 meters? Here's your chance to get on the new WARC band with an inexpensive homemade trans-receiver.

By Doug DeMaw,* W1FB

I've been "hounded" for conversion information ever since the 20-meter "Hamcation Special" appeared in November 1982 *QST*. Amateurs have been asking how they might convert the 20-meter trans-receiver for use at 10 MHz. The changes are simple, and a kit of parts is available by mail.[1]

Transmitter Modification

Owing to the broadband circuits in the transmitter section of the portable station, there is little to do in order to convert the rf strip from 14- to 10.1-MHz operation. The crystal for the VXO is selected for the upper or lower segment of the band (remember that 10.109 to 10.115 MHz is an amateur "no man's land"; we can't operate there!). The VXO provides approximately 10 kHz of frequency swing when using a 10.1-MHz crystal. L1, the toroidal inductor that connects between Y1 and tuning capacitor C1 of Fig. 1, is made larger. The new inductance value is 12 μH.[2]

The capacitors and inductors of the PA output filter must also be changed in value. Pertinent data are given in Fig. 1. No other changes are required in the transmitter section of the rig. Power output from the MRF476s is the same as for 14-MHz operation — 7 to 8 watts into a 50-ohm load while using a 13.5-volt dc power supply.

Receiver Circuit Changes

T4, L7, C2 and C3 of Fig. 2 (preamplifier module) need to be changed to new values. No additional modifications are necessary for this subassembly.

Two areas require attention in the "Bare Bones CW Superhet" receiver.[3] The VXO crystal (Y4 of Fig. 3) is ground for operation at approximately 13.6 MHz (depending on the band segment desired). The VXO inductor (L1) is increased in value to 10 μH (see note 2).

The remaining receiver-circuit change involves rewinding T1 at the mixer input and adjusting C1 to provide resonance at 10.1 MHz. There are no other changes to the receiver circuit. Simple, eh?

Some Band Aids

Some versions of the original circuit from November 1982 *QST* developed minor "bugs." One condition became manifest as a short-term howl in the earphones as the break-in-delay circuit dropped out of the transmit mode. This was traced to momentary self-oscillation of the receiver mixer during the transfer period of the T-R relay. At that time the receiver remains activated, but there is no load connected to the mixer input. Oscillation occurs at the intermediate frequency (3.579 MHz) because of the high Q reflected to the mixer by the crystal filter. The simple cure is to add a 270-ohm, 1/4-watt resistor (R1) from the high side of the T1 link winding to ground (Fig. 3). This maintains an adequate load at all times, but is high enough in value to prevent degradation of the receiver sensitivity.

Another problem reported by some operators is a missing first dot or dash when the keyer is actuated. This can result

[1]Notes appear on page 21.
*QST Senior Technical Editor

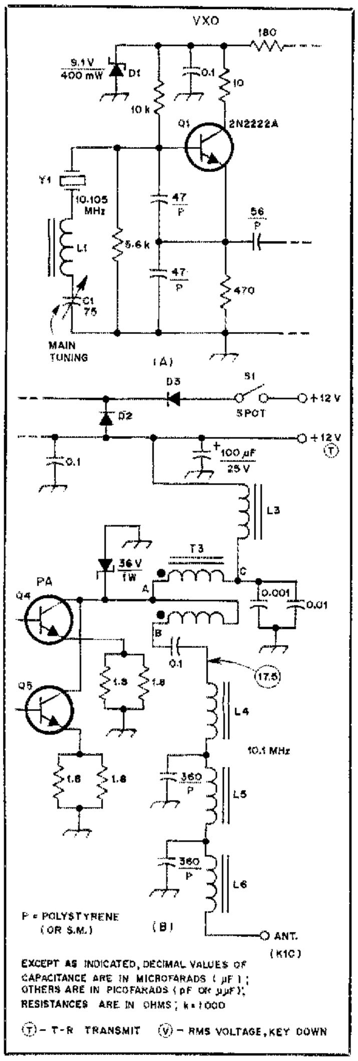


Fig. 1 — The diagram at A shows the transmitter VXO. Y1 is changed to 10.105 MHz for approximate coverage from 10.100 to 10.110 MHz, the lower segment of the 30-meter band. Y1 is an International Crystal Mfg. Co. no. 434110 fundamental crystal. The new L1 value is 12 µH (50 turns of no. 28 enam. wire on an Amidon or Palomar T50-2 toroid core). At B is the PA section of the transmitter. The new cutoff frequency for the harmonic filter is 11.67 MHz. L4 and L6 are rewound to provide an inductance of 0.5 µH (11 turns of no. 24 enam. wire on a T50-6 toroid core). L5 is 1.0 µH (16 turns of no. 24 enam. wire on a T50-6 core).

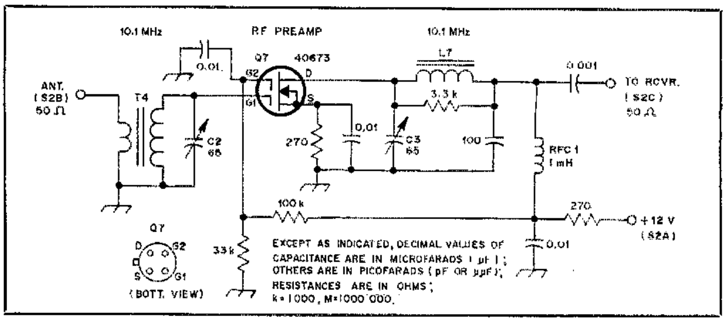


Fig. 2 — Circuit of the modified preamplifier. C2 and C3 are changed to 65 pF, or the existing ones can be shunted by fixed-value capacitors to provide resonance at 10.1 MHz. T4 is rewound to contain 34 turns of no. 28 enam. wire on a T37-2 toroid core. The link has 3 turns. L7 has 36 turns of no. 28 enam. wire on a T37-2 core.

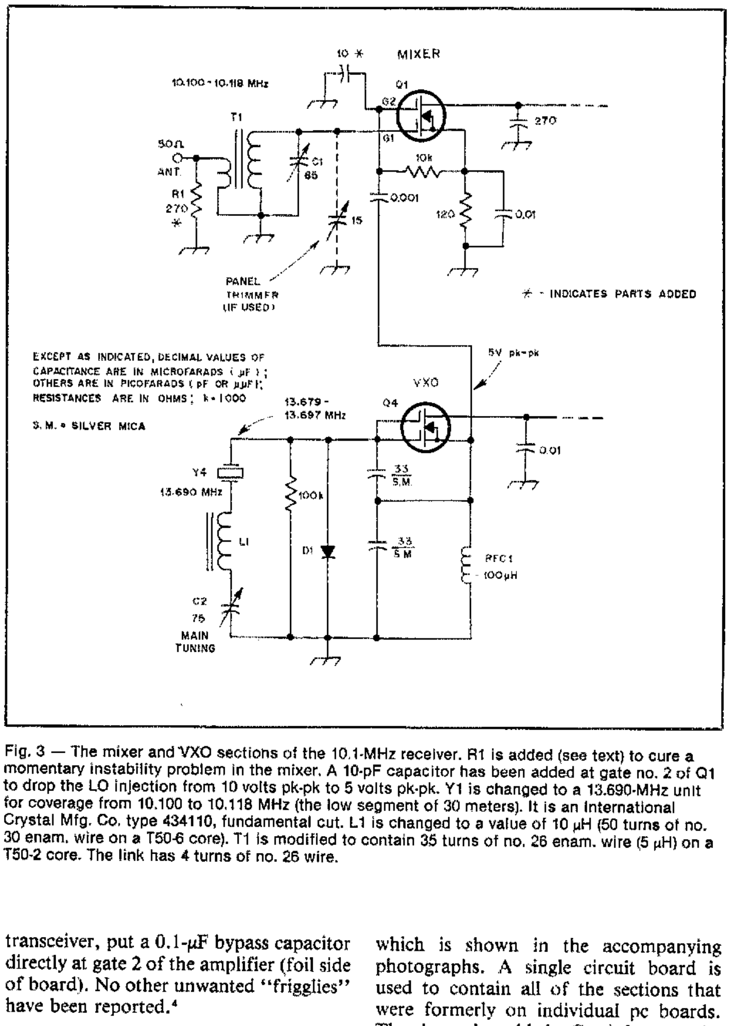


Fig. 3 — The mixer and VXO sections of the 10.1-MHz receiver. R1 is added (see text) to cure a momentary instability problem in the mixer. A 10-pF capacitor has been added at gate no. 2 of Q1 to drop the LO injection from 10 volts pk-pk to 5 volts pk-pk. Y1 is changed to a 13.690-MHz unit for coverage from 10.100 to 10.118 MHz (the low segment of 30 meters). It is an International Crystal Mfg. Co. type 434110, fundamental cut. L1 is changed to a value of 10 µH (50 turns of no. 30 enam. wire on a T50-6 core). T1 is modified to contain 35 turns of no. 26 enam. wire (5 µH) on a T50-2 core. The link has 4 turns of no. 26 wire.

from 2.2-µF blocking capacitor between the gain control and pin 2 of the TL081 op amp holding a charge, which locks out the sidetone signal on the first dit. Substitution of a 0.1-µF capacitor cured the malady.

If instability is noted in the 40673 i-f amplifier of the kit version of this transceiver, put a 0.1-µF bypass capacitor directly at gate 2 of the amplifier (foil side of board). No other unwanted "frigglies" have been reported.[4]

Kit Version

Circuit Board Specialists has developed the kit version of this trans-receiver, which is shown in the accompanying photographs. A single circuit board is used to contain all of the sections that were formerly on individual pc boards. They have also added a Curtis keyer to the new board. The assembled unit (less power supply) measures (HWD) 2-3/4 × 6-3/4 × 6 inches (70 × 171 × 152 mm).

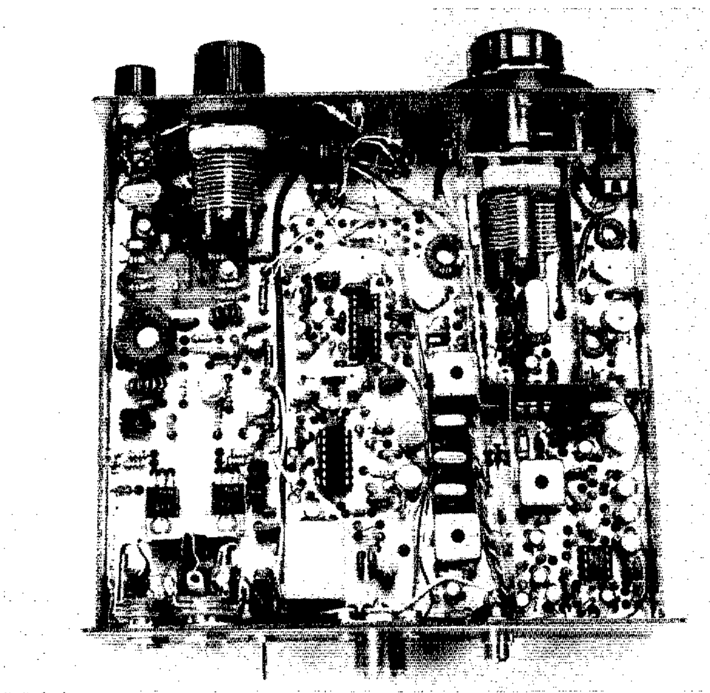


Fig. 4 — Interior view of the CBS kit version of the 30-meter trans-receiver. This lightweight package is ideal for the traveling ham.

This unit is available for 20 or 30 meters (note 1).

Summary Comments

I have enjoyed hopping into the 10.1-MHz activity with low power. With just a sloping dipole and 8 watts I have worked DX easily, and have had a lot of fun chewing the rag with other 30-meter enthusiasts. This new band is ideal for portable operation, since it is open most of the time. Furthermore, a large antenna is not required for the travel pack. I consider 10.1 MHz a great mix of the good qualities of 40 and 20 meters — perfect for the briefcase rig I like to carry afield!

Notes

[1]Circuit Board Specialists (WAØUZO), P.O. Box 969, Pueblo, CO 81002. Negatives, pc boards or full kits available.

[2]The inductance value in both VXOs may have to be increased to more than the amount specified in order to obtain the desired lower limits of the tuning ranges. This will depend upon the crystal used. Use only the amount of inductance that provides the desired range.

[3]D. DeMaw, "Build a Bare Bones CW Superhet," *QST*, June 1982.

[4]Op amps may break into audio self-oscillation when using 8-ohm hi-fi headphones. If this problem occurs, insert a 100-ohm resistor in series with the audio-output at the phone jack. High-impedance phones will give the best results, and are recommended with this circuit.

Strays

QEX: THE EXPERIMENTERS' EXCHANGE

Wonder what you've been missing by not subscribing to *QEX*, the ARRL newsletter for experimenters? Among the features in the March issue were:

- More on the Second ARRL Packet Conference
- "Packet Radio and Radio Communication Requirement," by Karl Meinzer, DJ4ZC
- "A Bibliography on Minimum-Shift Keying," by Den Connors, KD2S
- "ZX81 RTTY Receive Program," by Brian Davis, W9HLQ
- "Data Communications — Digital Phase Lock Loops," by David Borden, K8MMO
- "Components," by Mark Forbes, KC9C

QEX is edited by Paul Rinaldo, W4RI, and is published monthly. The special subscription rate for ARRL members is $6 for 12 issues; for nonmembers, $12. There are additional postage surcharges for mailing outside the U.S.; write Headquarters for details.

No, the FCC is not in the auto parts business, as this sign on Guam seems to indicate to ARRL President Vic Clark, W4KFC, who visited the island last year. *(tnx AH2A)*

CALL FOR PAPERS

☐ The 1983 ARRL National Convention, to be held in Houston, Texas, on October 7-9, will have a new wrinkle for DXers and contesters. The Texas DX Society is inviting leading DXers and contesters worldwide to participate in the First International DX and Contest Symposium. Blue-ribbon panelists will debate pertinent issues of interest to ham radio in the '80s, such as contest ethics and the pros and cons of list operation. A resolution will then be made. The proceedings of each debate will be submitted for publication after the Convention.

The TDX is issuing a call for papers to be submitted from the DX and contesting community on topics of interest to each individual. The TDX will then choose the papers that best represent each side of the most popular issues. These papers will be read and credited to the amateurs who submit them. All contributors whose papers are chosen for presentation will be notified in advance. The papers, one typewritten page on any DX or contest topic, should be submitted by June 1, 1983 to Bob Evans, N5DU, Symposium Coordinator, 13719 E. Cypress Forest, Houston, TX 77070.

I would like to get in touch with . . .

☐ any hams who were members of the 40th Mobile Communications Squadron or the 21st Weather Squadron during WW II. Irvin J. Kirch, W9YFG, 34 Hoss Rd., Indianapolis, IN 46217.

The Search for a Simple, Broadband 80-Meter Dipole

What amateur has not dreamed of finding a broadband antenna for 80 meters? Feed it with coax and work with a 2:1 SWR across the band!

By Jerry Hall,* K1TD

Like many amateurs who operate in the 3.5-MHz band, you may use a simple, single-wire dipole (or inverted V) antenna fed with either 50- or 75-ohm coaxial line. And like every one of those amateurs, you are inconvenienced when you want to operate from band edge to band edge with that antenna. Since most of today's transmitters require an SWR no greater than 2:1, you probably find yourself restricted to operating within a 200- or 250-kHz segment of the 80/75-meter band.

Common practice with a coax-fed antenna is to cut the length to favor either the cw or the phone end of the band. Amateurs who want to operate both modes often cut the antenna for resonance near the center of the band. In this way they can operate in the upper frequency end of the cw portion of the band and the lower frequency end of the phone portion.

But try to operate with such an antenna over the entire 80/75-meter band. No way! Even if the antenna is cut for the center of the band, the SWR will be greater than 4 to 1 at the band edges, as curve A in Fig. 1 shows. Oh, yes, you can always use a Transmatch or other arrangement to transform the line input impedance to something near 50 ohms, but this invariably requires additional controls to tune when you shift from one end of the band to the other. Wouldn't it be fabulous if you could cover the entire band without the inconvenience of adjusting controls outside the transmitter? Curve B of Fig. 1 shows the SWR response of just such an antenna.

The Search for the Ultimate Antenna

Over my years as an amateur I have done considerable experimenting with wire antennas. I tried one kind, then another, then yet another, always looking for that ultimate 80/75-meter radiator. For years I used a center-fed length of wire with an open-wire feeder. At the input end of the line I used a Transmatch. This arrangement gives quite respectable operation over the 80/75-meter band (and our mf band and other hf bands as well!). I've used lengths from 60 ft total to 170 ft total,[1] and each length gave very good results. (Of course, the longer wires excelled somewhat on the lower frequency bands.) I also tried the double bazooka antenna,[2,3] the miscalled Windom antenna,[4] and several others that as far as I know have no names. Uncounted hours upon hours were spent in assembling and installing these antennas, measuring and plotting SWR values across the band, lowering the antennas to make slight changes, and raising the antennas again, only to repeat the whole process.

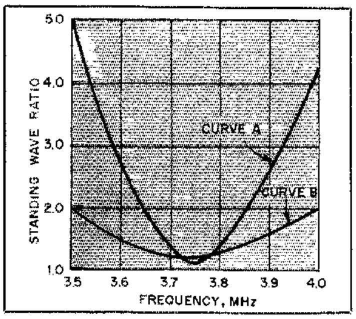


Fig. 1 — Curve A shows the SWR vs. frequency as derived from computer calculations for a no.-12-wire, center-fed, dipole antenna with 75-ohm feeder. The antenna is 125 ft long and is assumed to be in free space with the feed line totally decoupled. This curve correlates very well with that from measured performance of an antenna of this type at a modest height. Curve B shows the SWR response of a broadband dipole discovered in the computerized search described in the text. This antenna has yet to be constructed and tested. See text for details.

Not long ago I realized that rather than devoting so much time to constructing and testing, there was a much better way to continue my search for the ultimate antenna. Why not use a personal computer to perform a preliminary analysis of my ideas? If the computer results looked promising, *then* I could construct and try that type of antenna. In this way I could "test" an antenna in a few minutes, rather than taking a week or more of spare time to check out one or two ideas.

The computerized search proved to be a very worthwhile approach. Over the period of a few weeks I checked out more than 200 different antenna and feed-line combinations. Of all the time spent, a good portion was used in researching the appropriate equations. And of course after that it was necessary to program the computer. But from there things went very quickly. Did I discover anything promising for that ultimate antenna? Well, perhaps. The SWR plot of one of my ideas is that shown in Fig. 1, curve B. But I'm not sure you would classify the antenna as being a "simple" one. More on that later. First, let me tell you how I proceeded with the study.

There are various theories described in the literature for calculating the resistance and reactance of a cylindrical antenna. A wire dipole is a long but, to be sure, a very thin cylindrical antenna. Once the resistance and reactance are known, it is a simple matter to calculate the standing-wave ratio for whatever impedance of transmission line is chosen as the feeder. Thus, with specific physical information to start from, the computer will produce results in terms of resistance (R), reactance (X) and SWR. Details on the math are given in the appendix to this article.

To check the accuracy of the computer program that evolved, I calculated data for several commonly used antennas. Correlation between the computer results and test data I had taken, sometimes years earlier, was excellent. I was more than pleased with the procedure and the results. Changing from 50- to 75-ohm line, for example, was as simple as

*Associate Technical Editor

[1]Notes appear on page 27.

entering a different number at the keyboard. And it was just as simple to change the antenna length, or its diameter. The computer could print out results faster than I could develop new ideas.

Before getting into the results, I should emphasize that the main equation on which these calculations are based applies to an antenna in free space. In the presence of the earth, that impedance will be modified. In general, the impedance will be lower when the antenna is less than 1/4 wavelength above ground. Translated to practical terms, this means that most of my calculations showed a better overall match to 75-ohm line than to 50, and these results are what I'll be presenting primarily. However, a 50-ohm line will likely provide lower SWR values in practice at the usual heights for an 80/75-meter antenna, i.e., below 65 feet. But remember, I wanted only a preliminary evaluation on a comparative basis from the computer. I did not feel it necessary to include earth effects in my calculations.

And here's another important point. The calculations take no feed-line currents into account. It is assumed that the feeder, whether it be balanced or coaxial, does no radiating.

A Simple Dipole

We've already looked into a simple dipole antenna of no. 12 wire. Curve A of Fig. 1 shows the SWR response of this antenna, and Table 1 shows R and X values as well as SWR values. The bandwidth between the frequencies exhibiting a 2-to-1 SWR is 208 kHz. With 50-ohm feed line the computed 2:1-SWR bandwidth is only 152 kHz, but keep in mind the remarks of the paragraphs above.

In Table 1, R represents the radiation resistance. This value goes higher with increased electrical antenna length (or higher with increased frequency for an antenna of fixed physical length). Note that the antenna is being fed with 75-ohm line. Note also that the radiation resistance is 75 ohms at about 3.875 MHz, yet the SWR value is still greater than 2:1. How come? The inductive reactance that is part of the feed-point impedance is to blame. (Positive values in the table indicate inductive reactance; negative values indicate capacitive.) While the reactance does not accept any power, it does prevent some of the power from being transferred to the resistance, creating the mismatch. The higher the reactance, the greater the mismatch, and therefore the higher the SWR.

Antenna resonance exists when the reactance goes through zero. Interpolating from Table 1, we see that this occurs at approximately 3746 kHz. Antenna resonance may not always coincide exactly with the frequency of lowest SWR, although with dipole-type antennas and coaxial feeders the two frequencies would never be more than 1, perhaps 2 kHz apart. It would be impossible to detect the difference with practical measuring equipment.

Table 1
Impedance of a Single-Wire Antenna

Antenna length = 125 ft
Antenna diameter = 0.08 inch
Feed-line impedance = 75 ohms

Freq.	*R, Ohms*	*X, Ohms*	*SWR*
3.500	57.2	−118.6	5.16
3.525	58.2	−106.5	4.43
3.550	59.4	−94.3	3.79
3.575	60.5	−82.2	3.23
3.600	61.6	−70.2	2.74
3.625	62.7	−58.1	2.32
3.650	63.9	−46.1	1.96
3.675	65.1	−34.0	1.65
3.700	66.3	−22.0	1.40
3.725	67.5	−10.0	1.19
3.750	68.7	2.0	1.10
3.775	69.9	14.0	1.23
3.800	71.1	26.0	1.43
3.825	72.4	38.0	1.67
3.850	73.7	50.1	1.94
3.875	74.9	62.1	2.24
3.900	76.2	74.1	2.57
3.925	77.5	86.1	2.94
3.950	78.9	98.2	3.33
3.975	80.2	110.3	3.76
4.000	81.6	122.3	4.22

The 2:1-SWR frequencies are 3.647 and 3.855 MHz and the bandwidth is 208 kHz.
Resonance occurs at 3.746 MHz.

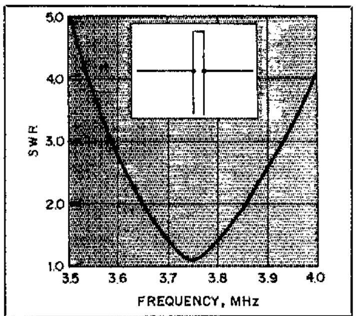


Fig. 2 — SWR curve for antenna with 1/4-wavelength hairpin, as shown in the inset. The antenna is of no. 12 wire, 125 ft long, and fed with 75-ohm line. The hairpin is 52.5 ft of 300-ohm TV ribbon with a 0.8 velocity factor.

Thicker conductors for a single-wire dipole will broaden the SWR response. For example, a dipole made of half-inch-dia wire (such as using the shield of RG-8/U as the radiator) with a 75-ohm feeder exhibits a 2:1-SWR bandwidth of 252 kHz, 44 kHz broader than the antenna of no. 12 wire. In order to maintain resonance at the same frequency as the no.-12-wire antenna, however, its total length must be shortened by about 8 inches.

A Hairpin Match

Now look at the antenna system illustrated in Fig. 2. This antenna is identical to the dipole for curve A, Fig. 1, except that a hairpin of an electrical quarter wavelength at 3.75 MHz has been added. The hairpin will introduce reactance at the antenna feed point as we move away from its quarter-wave-resonant frequency. Below 3.75 MHz this reactance is inductive; above, it is capacitive.

By happy coincidence this is just the opposite of what the antenna itself does, so the hairpin should cancel some of the reactance at the band edges, right? And because the reactance is primarily responsible for the high SWR, then we would think the bandwidth should be significantly broader. But comparison of the SWR plots reveals that there is negligible bandwidth improvement over the dipole alone. The 2:1-SWR bandwidth is 211 kHz, as opposed to 208 kHz without the hairpin. So what went wrong when we added the hairpin?

Most of us amateurs are not accustomed to thinking in terms of paralleling resistances and reactances, and the results often surprise us. For example, assume we have an antenna impedance of 50 ohms resistance and 100 ohms equivalent series reactance at some particular frequency. A shorthand way of writing this impedance is 50 + j100 ohms, where the j indicates that the number behind it represents a reactance, and the values cannot be added directly. The SWR in 50-ohm feed line for this antenna will be 5.8:1. (See what that nasty reactance does to us!)

Now assume we carefully design a hairpin that presents a pure reactance of −j100 ohms at the feed point for this frequency. The actual feed-point impedance will then be the result of the 50 + j100 ohms in parallel with −j100 ohms. We might think the combination will be near 50 ohms with no reactance, but we would be very wrong. The actual feedpoint impedance turns out to be 200 − j100 ohms. The SWR has dropped only from 5.8:1 to 5.1:1.

If we take the extra trouble to completely cancel the feed-point reactance in this example, we will need a shunt reactance of −j125 ohms. The resulting feed-point impedance then will be 250 + j0. With all this care, we've succeeded in bringing the SWR down only to 5:1.

We tend to forget that shunting an impedance with a reactance to obtain a pure resistance gives us an accompanying transformation of the resistance value. As a matter of fact, this is the very principle involved in matching with an L network. So in this example, the very best we can hope to obtain with hairpin matching is a 5:1 SWR.

Different hairpin lengths of 300-ohm line did not yield significant change in the 2:1-SWR bandwidth. The frequency of resonance for the radiator *system* (no reactance at the feed point) changed slightly, however. Making the hairpin length 26 ft instead of 52.5 ft lowered system resonance from 3.746 to 3.712 MHz. With the hairpin lengthened to 75 ft, system resonance occurred at 3.775 MHz. The length of the radiator wire was held at 125 feet for each case here.

The broadest practical system discovered with a hairpin was 254 kHz in bandwidth, as opposed to 208 kHz without the hairpin. This result was obtained with a hairpin of 25 ohms characteristic impedance, 43.5 ft long. The antenna was 125 ft of no. 12 wire, fed with 75-ohm line. A 25-ohm hairpin can be constructed with two lengths of 50-ohm line connected in parallel (i.e., shield connected to shield and center conductor to center conductor). The velocity factor for the hairpin line was taken as 0.66.

Folded Dipoles

It is commonly stated that a folded dipole has improved SWR-bandwidth performance over a single-wire dipole. (Of course, neither has gain over the other.) Computer calculations verified this statement. I "tested" a folded dipole made of 300-ohm TV ribbon. For these calculations I took the antenna as having a 0.5-inch dia, ignoring the fact that it was not truly cylindrical. The antenna was 125 ft long, and a velocity factor of 0.8 was used to determine the shorted-transmission-line effect of the two halves. (This effect is explained in the appendix.) The 2:1-SWR bandwidth in the 300-ohm feeder came out to be 265 kHz, compared with 208 kHz for the single wire with 75-ohm feeder. System resonance occurred at 3.752 MHz.

But wait a minute! We overlooked something that may be important. With a folded dipole the approximate antenna length for resonance can be determined by the old familiar equation, $l = 468/f$. But half of this length is *not* an electrical quarter wavelength as far as the shorted-transmission-line effect goes. To obtain quarter-wave resonance, we must take the velocity factor into account and place the shorting terminations 80% of a free-space half wavelength apart. This is shown in the inset in Fig. 3, and works out to be a distance of 105 ft for 3.75 MHz. The resulting SWR plot is also shown in Fig. 3. The bandwidth is 262 kHz, 3 kHz less than with the shorting terminations at the antenna ends. In addition, system resonance moved from 3.752 to 3.725 MHz (although 3.726 is the natural resonant frequency of a 125-ft single-wire dipole that is 1/2 inch in dia). Thus, it appears to make little difference in bandwidth whether or not the velocity factor is taken into account when placing the shorting terminations.

Some years ago the use of capacitors at the ends of a 300-ohm-ribbon folded dipole was advocated to compensate for the velocity factor. This idea is shown in the inset of Fig. 4. A capacitor value of 5.9 pF per meter of frequency-band operation is used at each end of the antenna. This suggestion seems to have gone unheeded by amateurs, however; I am unaware of any current publications that carry the information. With the computer, I tried this idea with various values of capacitance. The optimum practical value for 3.75 MHz, determined from the Smith Chart, was 470 pF. This was confirmed by the computer, although values as low as 390 pF or as high as 1000 pF provided the same bandwidth results — 266 kHz. Placing the capacitors at the ends essentially does not improve the bandwidth over that with the shorts at the ends, but the results are slightly better than when taking the velocity factor into account for the shorts. Changing capacitor values did change the system resonant frequency slightly, from 3.721 MHz with 390 pF to 3.726 with 470 pF to 3.738 with 1000 pF. The length was held at 125 ft all the while.

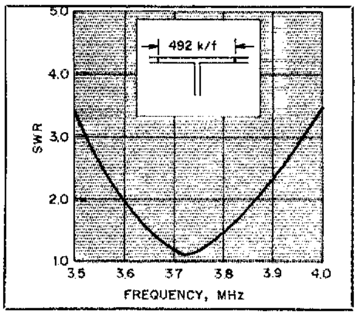


Fig. 3 — SWR curve for a folded dipole as shown in the inset. The antenna and feeder are of 300-ohm TV ribbon. The radiator is 125 ft long, with the shorting connections placed 105 ft apart (k = 0.8). For simplicity in calculations, the radiator was assumed to be cylindrical, with a diameter of 1/2 inch.

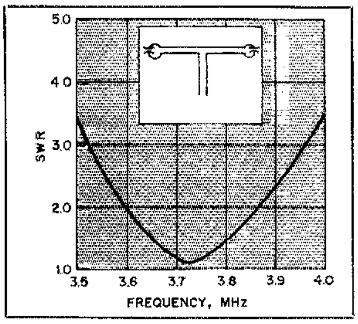


Fig. 4 — A type of folded dipole. The capacitors at the ends compensate for the velocity factor of the 300-ohm TV ribbon for the terminated-transmission-line effect it presents at the feed point. Capacitor values for this plot were 470 pF; the antenna length was 125 feet.

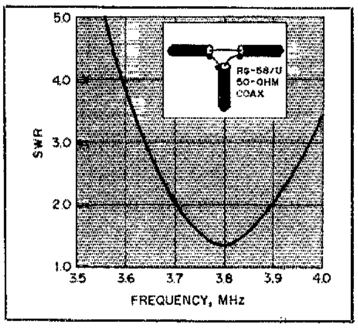


Fig. 5 — The feed-point arrangement of a double bazooka antenna is shown in the inset, and the curve shows the SWR response. The total coax length is 86.67 ft, while the total antenna length is 122.67 ft. The length outside the coax is of 300-ohm open-wire feeder.

The Double Bazooka

The double bazooka antenna is favored by many amateurs because of its "broad bandwidth characteristics." The center portion of this antenna is shown in the inset of Fig. 5. The coaxial line on each side of center is an electrical quarter wavelength of 50-ohm RG-58/U, and is shorted at the outside ends. Because the total coax line length is only 66% of a free-space half wavelength, the antenna is too short for resonance as a radiator. Resonance is obtained by adding a section of 300-ohm open-wire line at each end. Both conductors of each 300-ohm section are tied together, at the connection to the coaxial sections and at the outside ends of the antenna. The antenna is fed with 50-ohm line. The SWR plot of this antenna is shown in Fig. 5. Even though the antenna was "cut" for 3.75 MHz from the double bazooka equations, resonance occurred at 3.803.

Based on computer calculations, the bandwidth of the double bazooka was 190 kHz, compared with 152 kHz for a simple, single-wire antenna fed with 50-ohm line. The improvement is 38 kHz. These computer results agree closely with tests that I conducted in the ARRL lab in 1974, using a General Radio 1606-A rf impedance bridge. From the measured impedances, the 2:1-SWR bandwidth of a double bazooka was 184 kHz, and that measured for a single-wire dipole at the same height and fed with 50-ohm line was 153 kHz. The measured improvement was 31 kHz.

Several variations on the design of the double bazooka yielded some differences in bandwidth. These variations included using 50-ohm line for the radiator and 75-ohm line for the feeder, reversing these arrangements, 75-ohm line for both, and so on. The greatest bandwidth obtained was 268 kHz, using RG-8/U line for the radiator (1/2-in. dia) and a 75-ohm feeder. With a 125-ft antenna and a coax-portion length of 86.5 ft, resonance occurred at 3.725 MHz.

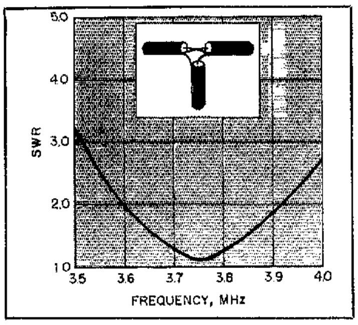


Fig. 6 — SWR plot of an antenna like the double bazooka but with crossed connections at the feed point inset. RG-8/U is used for the coax portion of the antenna (1/2 inch dia), and the feeder is 75-ohm line. The antenna is 124.2 ft long, with 86.5 ft of this length in coax.

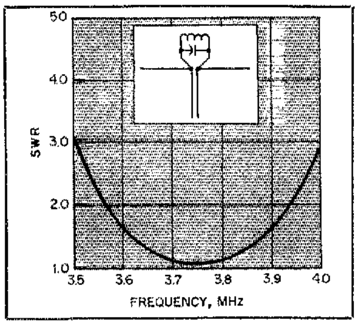


Fig. 7 — A 125-ft 1/2-in.-dia dipole with fixed components at the feed point. The inductor value is 0.5305 μH and the capacitor is 3395 pF. The system is fed with 75-ohm coax. This 124.2-ft antenna has the broadest response of all the practical antennas investigated during the study.

Table 2

Summary of Computer-Study Results for Various Antennas

These calculations assume the antenna is in free space. In actual installations at moderate heights, the bandwidth for 50-ohm feeders will likely be greater than that indicated here.

Antenna Type	*Radiator Dia., In.*	*Feeder Imped.*	*Bandwidth*
Single-wire dipole	0.08	50	152
Double bazooka	0.38	50	190
Single-wire dipole	0.08	75	208
Single-wire dipole with 52.5-ft hairpin of 300 ohms	0.08	75	211
Single-wire dipole	0.5	75	252
Single-wire dipole with 43.5-ft hairpin of 25 ohms	0.08	75	254
Folded dipole of 300-ohm ribbon, shorts 100 ft apart†	0.5	300	261
Folded dipole of 300-ohm ribbon, shorts 105 ft apart	0.5	300	262
Folded dipole of 300-ohm ribbon, shorts at ends	0.5	300	265
Folded dipole of 300-ohm ribbon, 470-pF capacitors at ends	0.5	300	266
Double bazooka, 75-ohm feeder	0.5	75	268
Double bazooka type of antenna, crossed wires at feed point	0.5	75	324
Single wire with coil and capacitor at feed point	0.5	75	375
Cage dipole	36	75	500

†Dimensions from Fig. 61, p. 2-29, *The ARRL Antenna Book*, 14th ed., 1982.

Other Antenna Types

Next I tried a double-bazooka style of antenna with crossed connections at the feed point. This arrangement is shown in the inset of Fig. 6. I found that a greater bandwidth could be obtained than with the double bazooka, as the SWR plot of Fig. 6 indicates. The bandwidth of this antenna is 324 kHz.

W1RN, one of my colleagues at ARRL Hq., suggested using a fixed coil and capacitor in parallel at the feed point (Fig. 7 inset). The effect should be somewhat the same as with a simple hairpin, because the reactance of an L-C parallel combination goes capacitive with increasing frequency. This is just the opposite of what the antenna does by itself.

The results with various L/C ratios were both interesting and surprising. They were interesting because a quite large bandwidth could be obtained — 375 kHz. (The 2:1-SWR frequencies are 3564 and 3939 kHz.) And they were surprising because, with the arrangement plotted in Fig. 7, there really is no frequency that can be defined as system resonance. The reactance swings across zero at three different frequencies in the band — 3563, 3750 and 3904 kHz, being inductive at the low end and capacitive at the high end. For this antenna arrangement, various L/C ratios were tested by assigning different reactance values at resonance. The reactance for optimum bandwidth was found to be 12.5 ohms at L-C resonance.

Although the calculations were for a radiator of 1/2-in. dia, 300-ohm flat TV ribbon or open-wire line with the conductors tied together should work as well. This antenna looks promising. If constructed for transmitting, the capacitor should be a transmitting type, and the coil should be of low loss. There will be high circulating currents in the L-C components. I have not constructed the antenna.

The Search Continues

Table 2 is a summary of the results of all the various antenna types discussed in this article. The antennas are listed in order of increasing bandwidth, and the results are revealing. I hope this information will dispel many rumors that persist about the bandwidths of various antenna types, and perhaps encourage experimentation among you readers.

What about that ideal-looking curve in Fig. 1? Well, there *was* one other idea I tried in my computer search for the ultimate — a cage antenna. A cage is made with several parallel conductors in a circular arrangement. The assembly resembles a round bird cage. Various materials — either conducting or insulating — can be used as spreaders for making a cage. A spreader would have to be placed every several feet along the antenna. If the antenna conductor wires are spaced close enough together, these conductors will have the effect of being a solid wire for the rf current.

It is a cage antenna that yielded curve B in Fig. 1. The 2:1-SWR bandwidth is 500 kHz, while the radiator is 121.5 ft long and is center fed with 75-ohm coax. The cage diameter is, would you believe, 36 inches! The only problem is that I'm not sure what the maximum spacing between adjacent conductors is for a cage to act as a solid conductor at 3.75 MHz. (I've found nothing definitive in the references, although 0.2 λ is a rule-of-thumb value for wire spacing for a "solid" reflector.) Assuming 4 wires would be satisfactory, a little arithmetic indicates that you would need 486 feet of wire and probably two dozen spreaders to build this dipole antenna. I'm still searching for a *simple* way to obtain the same results.

APPENDIX

If you're a computer buff or if you'd like to know about a mathematical method of simulating a near-resonant dipole antenna, you'll be interested in this appendix. Even if the math doesn't interest you, you may pick up a bit of antenna theory.

Of the material researched, a text by Jasik contains information that seemed well suited to use with a personal computer.[3] The technique Jasik presents uses the induced emf method, based upon a sinusoidal distribution. His equation, in simplified form, is

$$Z = R + j\left[Y - 120\left(\ln \frac{24\ell}{d} - 1\right)\cot m\right] \qquad \text{(Eq. 1)}$$

where

Z = feed-point impedance at the center of the antenna, in the form R + jX. R and X are in ohms.

R = radiation resistance, a function of m (see Eq. 2)

j = complex number notation, indicating rectangular coordinates; the entire j term represents the reactance at the feed point

Y = function of length m (see Eq. 3)

ln = natural logarithm

ℓ = total length of antenna, feet

d = diameter of antenna, inches

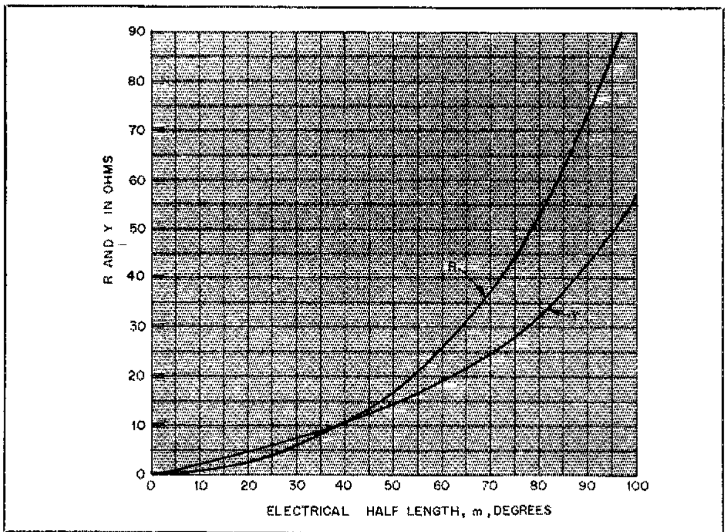


Fig. 8 — Values of R and Y (ohms) vs. m, electrical half length of the antenna, in degrees. (After Jasik; see note 5.)

m = electrical length of half the antenna = 180 (fℓ/984 + 0.013) degrees. (Note that m is not constant across the band.)
f = frequency in megahertz

Jasik states that Eq. 1 is valid only when the half length of the antenna is not much longer than a quarter wavelength, meaning near half-wave antenna resonance. This was no limitation for this study, but be aware that Eq. 1 is invalid for, say, center-fed long wires or harmonic antennas. For short dipoles the equation may be simplified, but Eq. 1 still applies here since we're working near resonance in all cases.

R and Y are complex functions of the electrical half length of the antenna, m. Jasik merely presents a table of values, and also a graph showing the values of R and Y for various values of m in radians. The graph is presented in Fig. 8, with m converted to electrical degrees.

For the computer analysis, it was necessary to arrive at R and Y values from equations, rather than from Jasik's table or graph. Further research indicated that taking sine and cosine integrals was required to compute R and Y.[6] Rather than tackle what I felt would be a monumental task in programming such a procedure, I chose to derive more simple equations that approximate the values Jasik shows. These equations are

$$R = \frac{m^{2.736}}{3048} \quad \text{(m is in degrees)} \qquad \text{(Eq. 2)}$$

$$Y = \frac{m^{2.234}}{549.7} \quad \text{(m is in degrees)} \qquad \text{(Eq. 3)}$$

In these equations, R and Y are in ohms.

Although these equations are only approximations for Jasik's data, they were optimized for the range from 80 to 100 degrees. The maximum variance from Jasik across the 80/75-meter band is less than 0.5 ohm, assuming resonance within the band. The variance increases significantly for short dipoles.

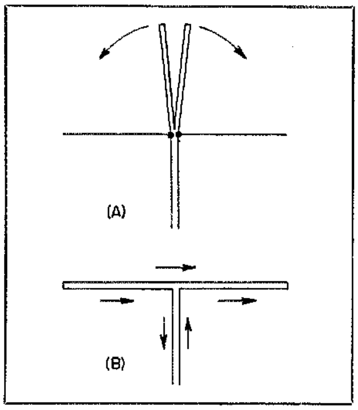


Fig. 9 — At A, a single-wire antenna with two quarter-wave stubs connected in series across the feed point, illustrating the "transmission-line effect" of a folded-dipole antenna. At B, the stubs are brought down to replace the single wire as a radiator, resulting in a folded dipole. Arrows show the direction of current flow in the feeder and the antenna conductors.

Remember that m is designated in electrical degrees in the above equations. If your computer calculates in radians (and most do), substitute the value of pi for the factor 180 in the definition of m in Eq. 1, and use Eqs. 4 and 5 for finding R and Y.

$$R = 21.17\, m^{2.736} \quad \text{(m is in radians)} \qquad \text{(Eq. 4)}$$

$$Y = 15.39\, m^{2.234} \quad \text{(m is in radians)} \qquad \text{(Eq. 5)}$$

Let's say we wish to determine the SWR plot of a single-wire antenna like that resulting in curve A of Fig. 1. The length is 125 feet and the wire diameter may be taken as 0.08 inch for no. 12 wire. For each frequency of interest in the band we may determine the feed-point impedance, Z = R + jX, from the above equations. The resulting SWR value for that impedance is then

$$SWR = \frac{A + B}{A - B} \qquad \text{(Eq. 6)}$$

where

$A = \sqrt{(R + Z_0)^2 + X^2}$

$B = \sqrt{(R - Z_0)^2 + X^2}$

Z_0 = characteristic impedance of feed line

With these equations you now have all the basic information you need to calculate and plot SWR curves for single-wire cylindrical antennas. I'll leave the computer programming details up to you, as the operation involves only entering data, straightforward mathematics and displaying the results.[7] The results of my computer calculations across the 80/75-meter band for this antenna are those appearing in Table 1.

Folded Dipoles, Stubs, Hairpins, and All That

In more complex antennas, or those with some type of matching arrangement incorporated at the antenna feed point, the net impedance result is equivalent in placing an additional impedance in parallel with the R + jX impedance of the antenna alone. Let's add a hairpin in shunt with the 125-ft antenna we've been talking about, as shown in the inset of Fig. 2, in hopes of broadening the SWR response. For simplicity, let's use 300-ohm TV ribbon as the material for the hairpin. The ribbon has a velocity factor of 0.8. So as not to introduce reactance across the entire band, let's make the hairpin length equal to quarter-wave resonance at the center of the band, 3.75 MHz. (At this frequency, the shorted quarter-wave line will represent an "open," or at least a very high impedance at the feed point of the antenna.) The physical length of the hairpin will be 52.5 ft.

As we depart from hairpin resonance and move in frequency toward either band edge, the hairpin will introduce reactance at the antenna feed point. In one frequency direction the reactance will be inductive, and in the other, capacitive, as mentioned earlier. How much and what kind of reactance may be determined from

$$X_H = j\, Z_H \tan h \qquad \text{(Eq. 7)}$$

where

X_H = reactance of hairpin at antenna end

Z_H = characteristic impedance of the transmission line used for the hairpin

h = electrical length of the hairpin

Eq. 7 assumes a lossless hairpin. The resultant impedance at the antenna feed point is R + jX of the antenna in parallel with X_H. Calculations of this nature are simplest when the resistance and reactance terms are converted to either admittance and susceptance or else to polar notation. These procedures have been discussed in detail in earlier *QST* articles.[8,9]

Instead of a hairpin of balanced transmission line, coaxial line might be used. In this case the matching arrangement might be more properly termed a stub. Hairpins or stubs need not be limited to resonant quarter wavelengths.

In the case of a folded dipole, a combination of factors affect the feed-point impedance. One factor is the "shorted-transmission-line" effect of the two halves of the antenna. The result of this factor is the same as placing the reactances of two hairpins in series, and connecting the series combination in parallel with the R + jX of the antenna. This is illustrated in Fig. 9A. With two identical hairpins in series, the total reactance is twice that obtained from Eq. 7.

In Fig. 9B the two hairpins are brought down to become the radiator as well. In so doing, antenna currents are caused to flow in both conductors, as indicated by the arrows. The total antenna current divides equally between the upper and lower conductors, and the direction of current flow will be the same in both, as shown. However, the feed line is connected to only one of the two conductors. With the same amount of power being delivered to the antenna but only half the feeder current flowing when compared with a single-wire antenna, it stands to reason that the feed-point impedance must be four times greater. This comes from the power equation, $P = I^2Z$, where Z is a complex quantity, R + jX. The impedance of the antenna as a pure radiator is therefore four times that obtained from Eq. 1. Thus, the actual feed-point impedance of the folded dipole is

$$Z = 4\,(R + jX) \text{ in parallel with } 2\,(j\,Z_H \tan h) \qquad \text{(Eq. 8)}$$

In a folded dipole type of antenna, the "transmission-line shorts" need not be placed at the extreme ends of the antenna. They may be placed nearer the antenna center, perhaps for better bandwidth. This idea is that shown in Fig. 3. Neither must the line be made of ribbon. Coaxial sections may be used, as is done in the double bazooka antenna.

The preceding paragraphs give the mathematic equations for computing feed-point impedances and then determining the standing-wave ratios for the feeder of your choice. Whatever type of cylindrical antenna might come to your mind can be simulated by applying the equations and reasoning out the configurations, as was done above, to determine the shunt or modifying impedances. Antenna lengths, feed-line impedances, hairpin-line impedances, hairpin lengths and other factors may each be assigned independently. The results may be calculated in terms of R, X and SWR. You can test a whole farm of antennas without ever having to go outside.

Notes

[1]meters = feet × 0.3048; inches = ft × 12.

[2]C. C. Whysall, "The 'Double Bazooka' Antenna," *QST*, July 1968, pp. 38-39. This antenna also appears in each edition of *The Radio Amateur's Handbook* from 1969 through 1977.

[3]W. Maxwell, "A Revealing Analysis of the Coaxial Dipole Antenna," *Ham Radio*, Aug. 1976. This excellent reference provides insight into some of the theory underlying the procedure presented in the appendix of this article.

[4]*The ARRL Antenna Book*, 13th (1974) and earlier editions.

[5]H. Jasik, *Antenna Engineering Handbook* (New York: McGraw-Hill Book Company, 1961), pp. 3-1 through 3-3.

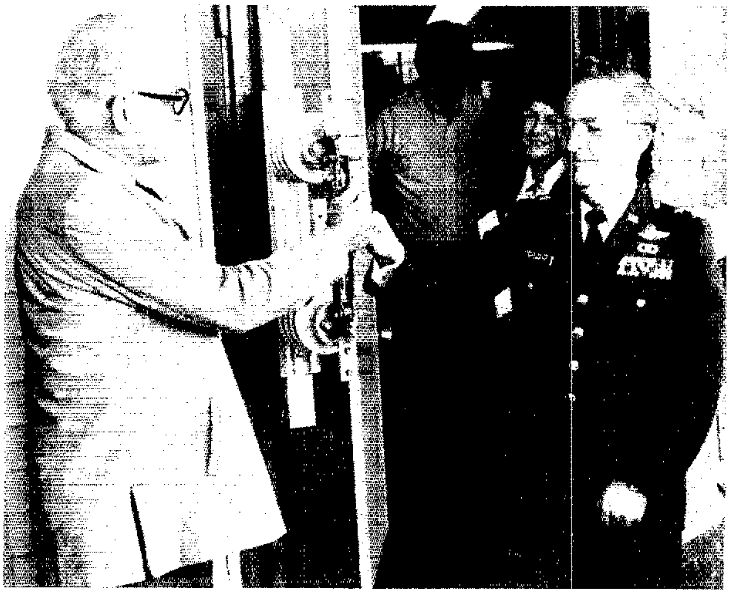
[6]P. S. Carter, "Circuit Relations in Radiating Systems and Applications to Antenna Problems," *Proceedings of the IRE*, Vol. 20, June 1932, pp. 1004-1041.

[7]A photocopy of the program the author used with an IBM personal computer is available upon request, as is the same program written for the TRS-80 model I Level II. Send $2 to cover copy fee, postage and handling to Dept. TD, ARRL Hq., 225 Main St., Newington, CT 06111. Request BASIC program ANTRXSWR/BAS, and be sure to state which computer program listing you desire.

[8]J. Hall, "A Simple Approach to Complex Circuits," *QST*, July 1977, p. 35. Also see Feedback, Aug. 1977 *QST*, p. 41.

[9]J. Bartlett, "Learning to Use Rectangular and Polar Notation," *QST*, November 1979.

Strays



Senator Barry Goldwater, K7UGA (left), and Major General Robert F. McCarthy, commander of Air Force Communications Command, ceremoniously "pull the switch" to terminate operation of MARS station AFC6BG. Since the 1960s, the station has provided a link to home for many armed forces personnel stationed in Southeast Asia and the Pacific.

MARS STATION AFC6BG IS QRT

☐ An amateur station that was a "link with home" for many Vietnam-era military personnel and their families has ceased operations. Military Affiliate Radio System (MARS) station AFC6BG in Paradise Valley, Arizona, went off the air on January 15.

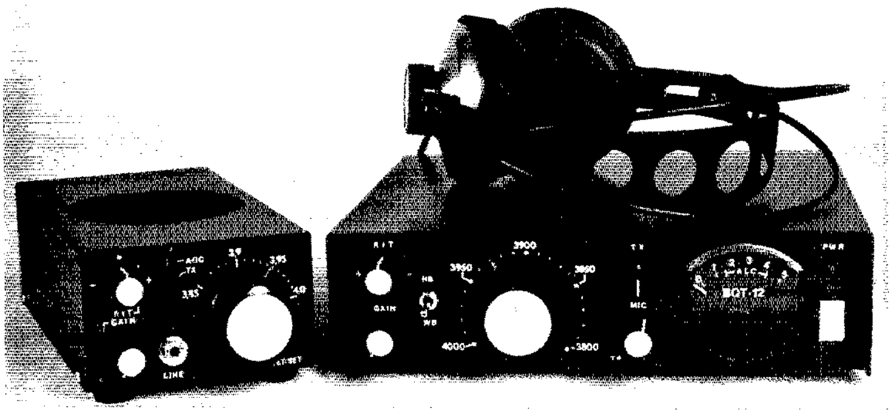
Owned by U.S. Senator Barry Goldwater, K7UGA, AFC6BG served in the 1960s as a phonepatch relay point for those stationed in Southeast Asia to talk to family and friends in the U.S. With the reduction of phonepatch traffic in the 1970s, the station became a radioteletype "gateway" station for the Pacific. The decision to close the station was based on a combination of declining mission activity and the anticipated cost of modernizing the station's aging equipment.

In recognition of the station's years of service, Secretary of the Air Force Verne Orr, WA6IOG, sent the following MARS message to K7UGA: "Operated exclusively by your dedicated volunteers, and at considerable personal expense to you, AFC6BG has provided the Department of Defense and the nation, in times of emergency, a service unparalleled in military communications. We all owe you and your people a tremendous debt of gratitude that can never be adequately repaid." In closing ceremonies at the station, Major General Robert F. McCarthy, commander of Air Force Communications Command, described the volunteers' service as that "which will stand the test of time and will display to others a sense of commitment and dedication." He presented plaques of appreciation to K7UGA, W7FCQ, a long-time station manager, station manager AFA6PU and a dozen volunteers who helped the station in recent years. — *U.S. Air Force News Release*

ATTENTION QRO OPS

☐ The 1983 Dayton (Ohio) Hamvention, scheduled for April 29-May 1, will offer cw operators the opportunity to challenge the world record for copying Morse code. As documented in the *Guiness Book of World Records*, the record of 75.2 wpm was set in 1939 by Ted McElroy. In addition, Cw Proficiency Certificates will again be awarded. For more information, write to Frank J. Schwab, W8OK, Cw Proficiency Chmn., Dayton Hamvention, Box 44, Dayton, OH 45401.

Construct an Audio Amplifier with Agc for Your Simple Receiver



Tired of using headphones with that bare-bones receiver? This module is an inexpensive way to make your little homemade project sound grown up!

By Rick Littlefield,* K1BQT

Many excellent articles describing simple receivers have appeared in amateur magazines. In many respects these receivers perform as well as their commercial counterparts. Most of the designs lack an automatic gain control circuit, however, and this can mean a lot of annoyance and knob twisting while monitoring nets or tuning the band. Some designs do not have the capability of operating with a speaker. In this article I offer one approach to making your next home-built receiver act like a store-bought radio. The module could also be retrofitted to an existing receiver in need of improved audio characteristics.

There are several ICs on the market that satisfy amateur receiver audio requirements. The LM-386, available at Radio Shack stores for about a dollar, is well suited for amateur use. This chip features low distortion, plenty of gain (up to 200), low current drain and high stability. With a 9-V supply it delivers 200 mW of audio power, more than sufficient to drive a speaker.

Automatic Gain Control

Several agc approaches are possible. The one I prefer rectifies a sample of the i-f signal to control the bias of one or more preceding stages. Deriving a control signal from the i-f stage may not be practical in simple designs in which the i-f signal levels are low and BFO leakage is common. Audio-derived agc is usually a safer choice. But deriving audio from a receiver that employs a single high-gain audio chip can also present a problem. The only voltage-sampling point before the volume control is at the output of the product detector, where the voltage level is very low. Thus, a fully functional agc circuit would require the addition of high-gain circuitry to develop a usable control voltage.

The simplest alternative is to derive the control voltage at the audio-amplifier output, where the signal level is high. The agc circuit then keeps the amplifier from being driven past a predetermined output level. Limiting is set somewhere below the maximum undistorted output level of the audio amplifier at a comfortable "maximum" volume level for the radio. A drawback to this approach is that agc action does not occur at very low gain settings. This should not be a problem in quiet listening environments, however, since the human ear can adapt easily to changes of 20 dB or more in normal conversation. In noisy environments the radio would be operated at a higher volume level, and heavy agc action would keep the output more constant. This is a compromise compared to a 90-dB agc system operating independently of the audio gain control; but it meets my design objective without adding much to the complexity of the receiver circuit.

*Box 114, Barrington, NH 03825

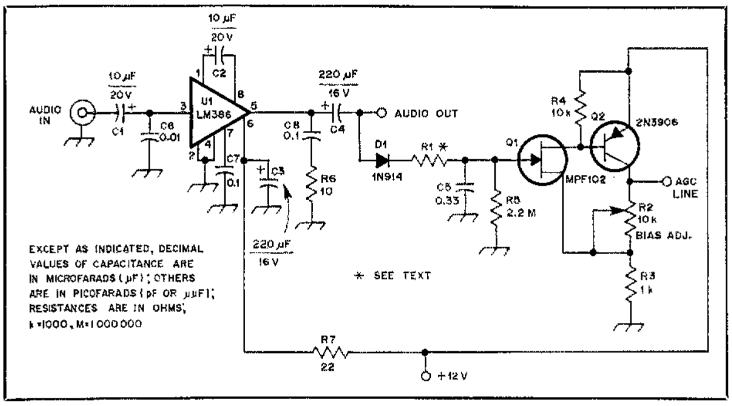


Fig. 1 — Circuit diagram of a simple audio amplifier and agc module for use with homemade receivers.

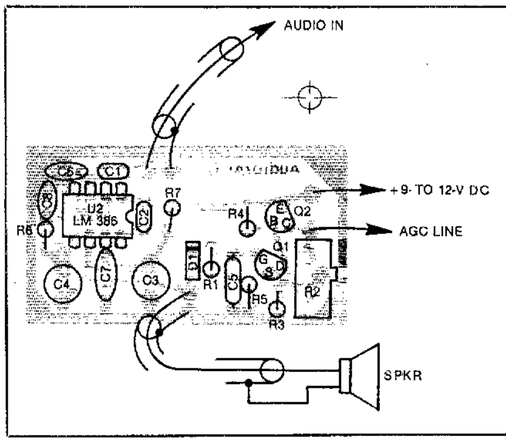
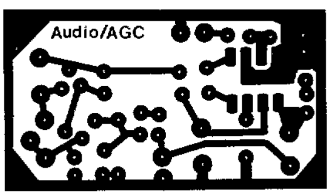


Fig. 2 — Parts-placement guide for the audio module. This view is from the component side of the board. Gray areas represent an X-ray view of the unetched foil. A full-size etching pattern is shown below.



Etching pattern for the audio amplifier and agc module. Black represents unetched copper, viewed from the foil side of the board. Pattern is shown full size. A parts-placement diagram appears in Fig. 2.

Circuit Information

I have used a circuit similar to that shown in Fig. 1 for several small receivers. The LM-386 audio section is compact and stable. Miniature 220-µF, 16-V capacitors (available from Radio Shack) contribute to the small size of the layout (Fig. 2).[1] The circuit is designed to drive an 8-ohm speaker. For the sake of good audio, it should be at least 3 inches in diameter.[2] Operation with headphones is possible by adding a simple attenuator similar to those used in tape recorders and transistor radios. A 10-ohm resistor replaces the speaker, and a 470-ohm series resistor drops the output to a comfortable listening level.

The agc design was borrowed from a more conventional audio-derived configuration and adapted for my application.[3] This circuit has some interesting features that make it universal. First, it samples a wide range of audio levels while remaining virtually "transparent" to the circuit because of the high input impedance. Second, it can be used with either a bipolar or an IC-based i-f section that requires a positive bias-voltage swing, or with MOSFET stages, which employ a negative bias-voltage swing. Since Q1 will respond to either positive or negative voltage, the polarity of D1 determines the direction of the bias-voltage swing. R2 sets the resting-bias level. Adjust this level to the point where agc action begins to lower the receiver background noise on a quiet frequency. R1 is used to adjust the drive. It should be set for smooth response. Overdrive will cause a pronounced overshoot or a cracking sound when strong signals come on. Lack of drive will cause insufficient limiting, and the audio amplifier will be driven into distortion. A substitution box can be used to determine a fixed value for R1 that will work best with your radio. Or, start with a 50-kΩ potentiometer and measure the required resistance.

Receiver Applications

Fig. 3 shows how I used this circuit in conjunction with a receiver that has a 455-kHz i-f and employs a single MOSFET i-f stage. Note that the polarity of D1 is set to provide a negative voltage swing. The agc action is smooth, and the i-f amplifier seems to be resistant to overload when no agc signal is present. Consequently, I did not include an rf or manual i-f gain control.

This receiver was built as a small net monitor for 75 meters. I had doubts about the "audio-derived limiter" approach to agc when I remembered how my old SBE-34 used to snap and crack every time a strong signal appeared. By setting the agc bias at the point of gain reduction and adjusting for the smoothest response, however, I eliminated most of that problem. The scheme worked so well that I decided to try it with a solid-state transceiver, which used an MC-1350P i-f amplifier.

Fig. 4 shows the receiver section of that rig, which uses a 9-MHz i-f stage. The results were good. In practice, I set the gain so that stronger signals activate the agc while weaker ones do not. This provides a reasonable dynamic range while maintaining good audio quality and a low level of background noise.

This receiver does employ a manual i-f gain control, since overload is a problem when signals are strong. Your i-f amplifier is overloading if signals sound distorted at low volume, but seem to clear up at higher levels, when the agc reduces the stage gain. If panel space is not available for the addition of an i-f gain control, a simple attenuator switch using fixed resistors could

[1]Notes appear on page 31.

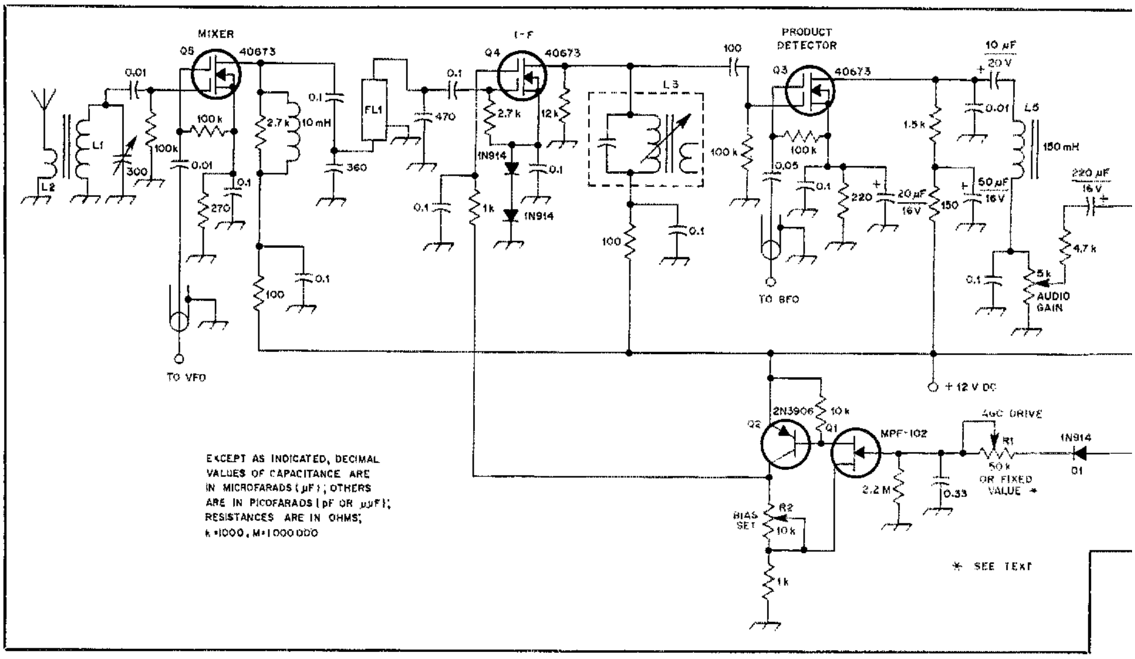


Fig. 3 — Schematic diagram of a 75-meter net-monitor receiver with a MOSFET i-f stage. The audio amplifier/agc module is used as the audio stage in this receiver.

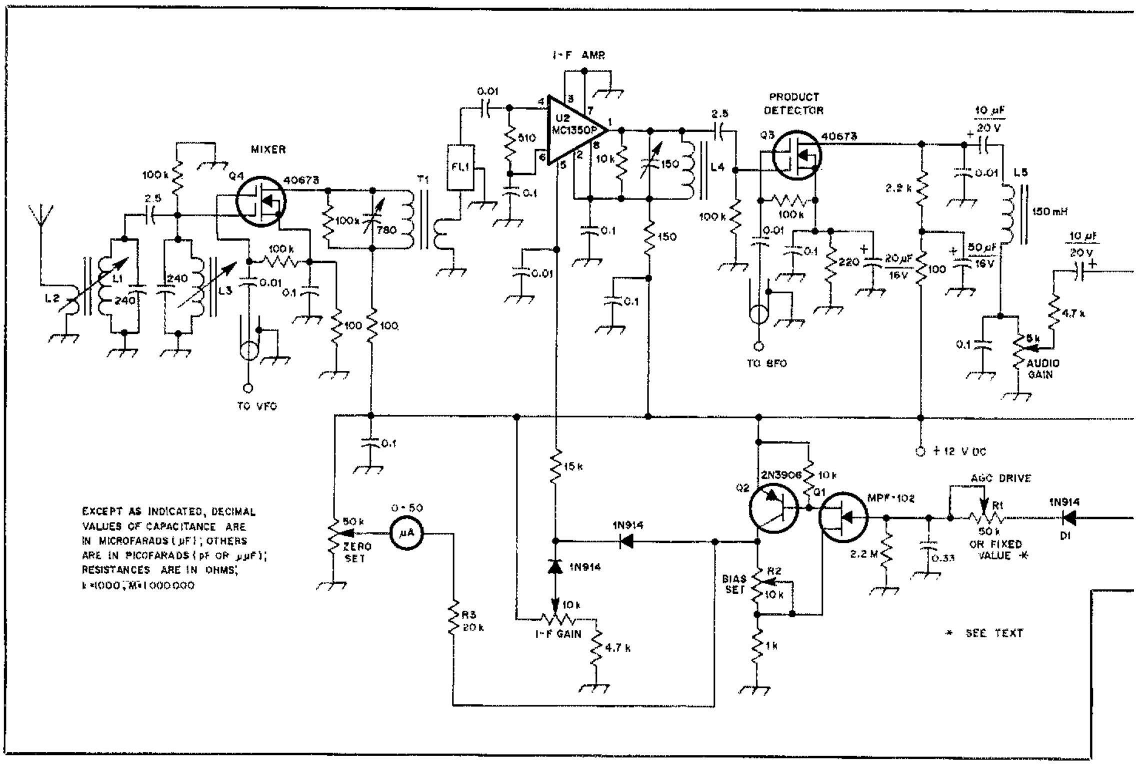
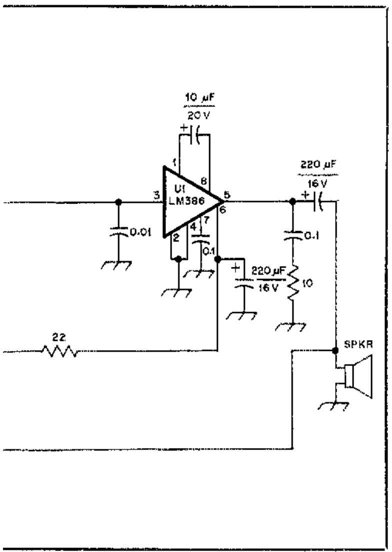
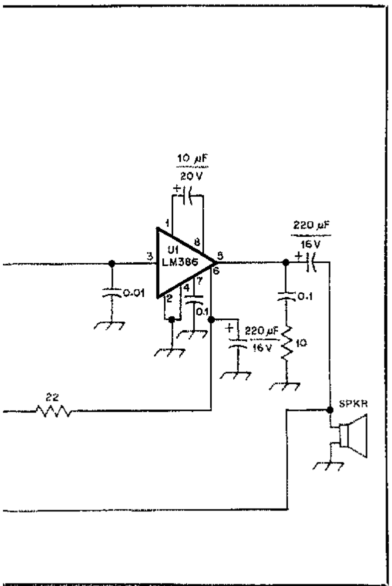


Fig. 4 — Schematic diagram of a simple receiver employing a 9-MHz i-f. The audio amplifier/agc module comprises the audio section of this receiver.



FL1 — 2.1-kHz bandwidth i-f filter. The one used by the author is a Kokusai unit purchased from MHZ Electronics.
L1 — 40 turns no. 24 enam. wire on an Amidon or Palomar T68-6 toroid core.
L2 — Two turns no. 24 enam. wire over ground end of L1.
L3 — Primary of a 455-kHz i-f transformer.



FL1 — Mechanical i-f filter. The author used a McCoy 9-MHz crystal filter.
L1, L3 — Approximately 7 µH, wound on a slug-tuned core.
L2 — Two-turn link wound on ground end of L1.
L4 — 24 turns no. 24 enam. wire on T37-2 toroidal core.
T1 — Primary, 25 turns no. 24 enam. wire on T37-2 core. Secondary, 6 turns no. 24 enam. wire.

be substituted. A 20-dB attenuation level should be about right.

Notice that both receivers employ a single-ended MOSFET product detector. This is a reasonable choice since the circuit is simple and provides some gain. An optional low-pass filter on the detector output reduces high-frequency audio interference. Inductor L5 consists of a 0.6-inch single-bobbin pot core wound to capacity with no. 36 wire. The inductance should be about 150 mH.

My transceiver contained a panel-mounted meter, so I included the meter circuit shown in Fig. 4 for the receiver section. Monitoring the agc voltage does not provide true signal-strength readings, but it does give an indication of agc action. This particular meter has a 50-µA movement, but less sensitive meters can be used with adjustments of R3. The comparative readings obtained with this meter make it a worthwhile addition to the project.

Conclusion

Employing an audio chip that is capable of driving a speaker, and adding a simple agc circuit, can eliminate two drawbacks to building your own "bare bones" receiver. This agc circuit may also adapt to existing projects, including direct-conversion receivers with rf amplifiers. Give it a try. The parts are inexpensive and the results are good.

[Editor's Note: This audio/agc circuit was installed in the W1FB "Bare-Bones CW 'Superhet,' " described in June 1982 *QST*. Few circuit changes were required to replace the original audio circuit. The agc bias voltage was connected directly to Q2 gate 1 through a pair of series-connected 1N914 diodes and a 1-kΩ resistor. The diodes were needed to provide the proper bias voltage range for the i-f amplifier. With no signal present, R2 was adjusted to provide +4 V on gate 1. When the receiver was tuned to a strong signal, agc action reduced this to less than 1 V, decreasing the i-f gain. More or fewer diodes may be needed with other receiver circuits.]

Rick Littlefield received his General class amateur license in 1957, at the age of 13. He holds a BA degree in communications and an MEd in education counseling from the University of New Hampshire. He is employed by the University of New Hampshire in the Department of Media Services. His main duties include writing and producing educational programming and providing consultation for University-affiliated projects, along with some teaching assignments. Rick also owns and operates Lakeshore Media, a small consulting and production company in Barrington, New Hampshire. Rick's other interests include playing several musical instruments and singing. He has stage and recording experience, is a professional narrator, a nationally published cartoonist, illustrator and author, has a Private Pilot rating, and enjoys sailing and cross-country skiing.

Notes

[1] Etched circuit boards, parts and parts kits for the audio amplifier/agc circuit are available from Circuit Board Specialists, P.O. Box 969, Pueblo, CO 81002, and from RADIOKIT, Box 411, Greenville, NH 03048.
[2] mm = in. × 25.4.
[3] W. Hayward and D. DeMaw, *Solid State Design for the Radio Amateur* (Newington: ARRL, 1977), pp. 103-106.

Strays

TA PROFILES

On November 20, 1981, we had the good fortune of adding David A. Zinder, W7PMD, to our team of ARRL Technical Advisors. Because of his more than 24 years experience in engineering and designing electric power control systems, we named him our expert on power supplies and ac power systems. He has been registered as a professional engineer in Wisconsin and Arizona.

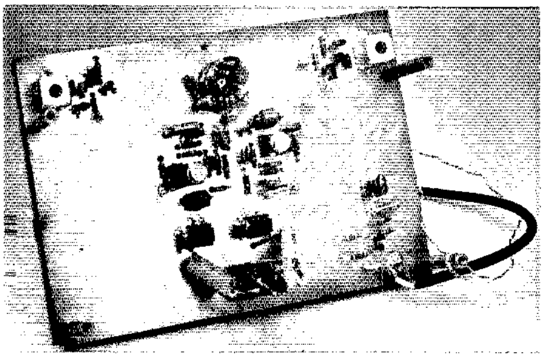
Dave received his first Amateur Radio license in 1951, and currently holds an Advanced class license. He resides in Phoenix, Arizona, and is active on 2 meters. If you are in the area, I'm sure you will hear Dave talking to his son (N7DZF) on 2-meter fm.

Dave received his BSEE degree from the University of Arizona. He is employed as the principal engineer for Sperry Flight Systems. He is also the power-supply designer and test engineer for the F-16 cockpit CRT display system. Dave has authored more than 22 technical papers, many of which were published in engineering technical magazines. One of his articles, "SCR Power Control Fundamentals," was selected as the best paper of the year in 1968 by *Appliance Engineer Magazine*.

After retiring as Scout Master, following eight years of service, Dave is still active in the local scout council and is the committee chairman. Many of his evenings are devoted to teaching "Intro to Electronics" at Phoenix College. When time allows, he enjoys hiking, camping and fishing. — *Marian Anderson, WB1FSB*

TA Dave Zinder, W7PMD

A Dichotic Detector for CW



Have you ever employed "psychoacoustics" to improve your receiver performance? Try this novel circuit and update your station human engineering.

By Douglas A. Kohl,* WØTHM

In spite of marvelous improvements in receiver performance over the years, the audio output circuitry of modern receivers does not take full advantage of our sense of hearing in rejecting unwanted signals. In a typical receiver we attempt to identify a cw station by adjusting the pitch of the "whistle" (beat note) to match a mental "discriminator." Adjacent unwanted signals produce additional tones that we attempt to ignore. Typical narrow-band (250-Hz) receiver operation eliminates most of the unwanted signals, but others within the passband can mask the desired signal. Furthermore, a strong signal on the i-f filter skirt can be louder than a weak signal in the center of the passband.

A Better Way

The human ear can detect a change in pitch of 3 Hz at 100 Hz, or 9 Hz at 1500 Hz, but *not* when two or more tones impinge upon the eardrum simultaneously. For example, if one tone was at 800 Hz, a second one would have to be 800 ± 125 Hz for the ear to be able to clearly distinguish two separate frequencies.[1] (This points out a fundamental limitation of narrow, crystal-filter techniques.)

Pitch recognition is enhanced greatly when the detection process is changed from single-tone to dichotic pitch comparison.[2] If the conventional audio output of a receiver (usb) is fed to the left side of a stereo headphone, and the audio from a frequency-inverted BFO detector (lsb) is fed to the right phone, there will be no audio "beating" or intertone interference caused by the physiologic functioning of the ear.[3]

If a spectrum of various pitched tones from different stations were presented independently to each ear, and if the same frequency were to appear in both the left and right headphones, two effects would occur. First, a sound "source location" would appear directly in front of the listener, as if he or she were listening to a stereo recording in which one instrument in the center of an orchestra is playing. Second, the perceived intensity of this "localized" tone is increased by three automatically.

To implement the dichotic cw detector, a conventional fixed-tuned BFO and product detector must be duplicated, as shown in Fig. 1. The "left" and "right" BFO frequencies are located at the edges of the receiver i-f bandpass (see Fig. 2). A signal appearing at A will produce a high-pitched tone in the right ear and a low-pitched tone in the left ear. (Refer to Table 1 for specific examples of the detector operation.) Because the same code signal produces tones of different pitches in each ear, the mind can easily ignore them because they are not spatially oriented and have no harmonic relationships. The desired signal, C, at the center frequency, however, produces both a stereo "front and center" location and a noticeable increase in sound intensity. One effect of this sensation is that it masks, or hides, adjacent signals, which are the most troublesome in a conventional receiver.[4]

In practice, there are several other effective modes of operating the detector. Because the dichotic system eliminates audio beats among all signals, the musical sensations of consonance (harmony) and dissonance (discord) of the left and right tones of the desired signal may be used effectively. If the receiver tuning is shifted off-center so the dichotic pitch difference is 100 Hz or less, the right and left tones create a harsh discord sound that clearly identifies the desired code signal in the presence of all others that have a "pure" sound.[5]

For those readers who have a "musical ear," tuning the receiver to create a major or minor third, a fourth or a fifth will form a harmonious tone pair from the desired signal. This will make it easy to distinguish from other background tones.

Circuit Construction and Alignment

The schematic diagram of a prototype dichotic cw detector is shown in Fig. 3. This unit was designed to operate with a receiver using a 9-MHz i-f. Two Colpitts oscillators are used to generate the BFO signals. The oscillators are identical except that one has a vernier tuning capacitor. LM3028 differential-amplifier ICs are employed as product detectors and a JFET source follower isolates the receiver i-f from the detector circuit.

Spurious output from the product detectors is attenuated in low-pass filters, and the detected audio is amplified by a pair of LM386 audio-amplifier ICs. The frequency response of the audio ICs is tailored for good low-frequency operation so true discord sensations can be felt by the user.

Perhaps the most important factor in the construction of this detector is isolation of the BFOs from each other and from the receiver i-f. In the prototype, a double-sided pc board was used, with one side of the board serving as a ground plane. Each BFO circuit should be located as far away as possible from the other. The receiver i-f should not be loaded by the detector circuit, although long lengths of coaxial cable from the receiver to the

[1]Notes appear on page 34.

*SPARSA Products, Inc., 417 Sixth Ave., N.E., Osseo, MN 55369

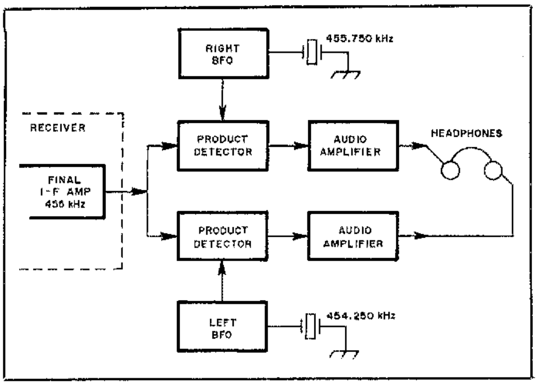


Fig. 1 — Block diagram of a dichotic detector for cw.

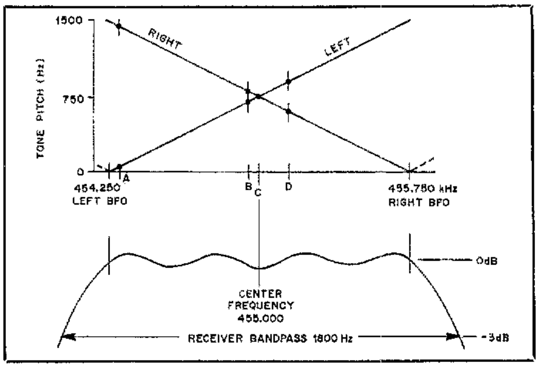


Fig. 2 — Signals present at points A, B and D produce tones of different pitches for the left and right ears.

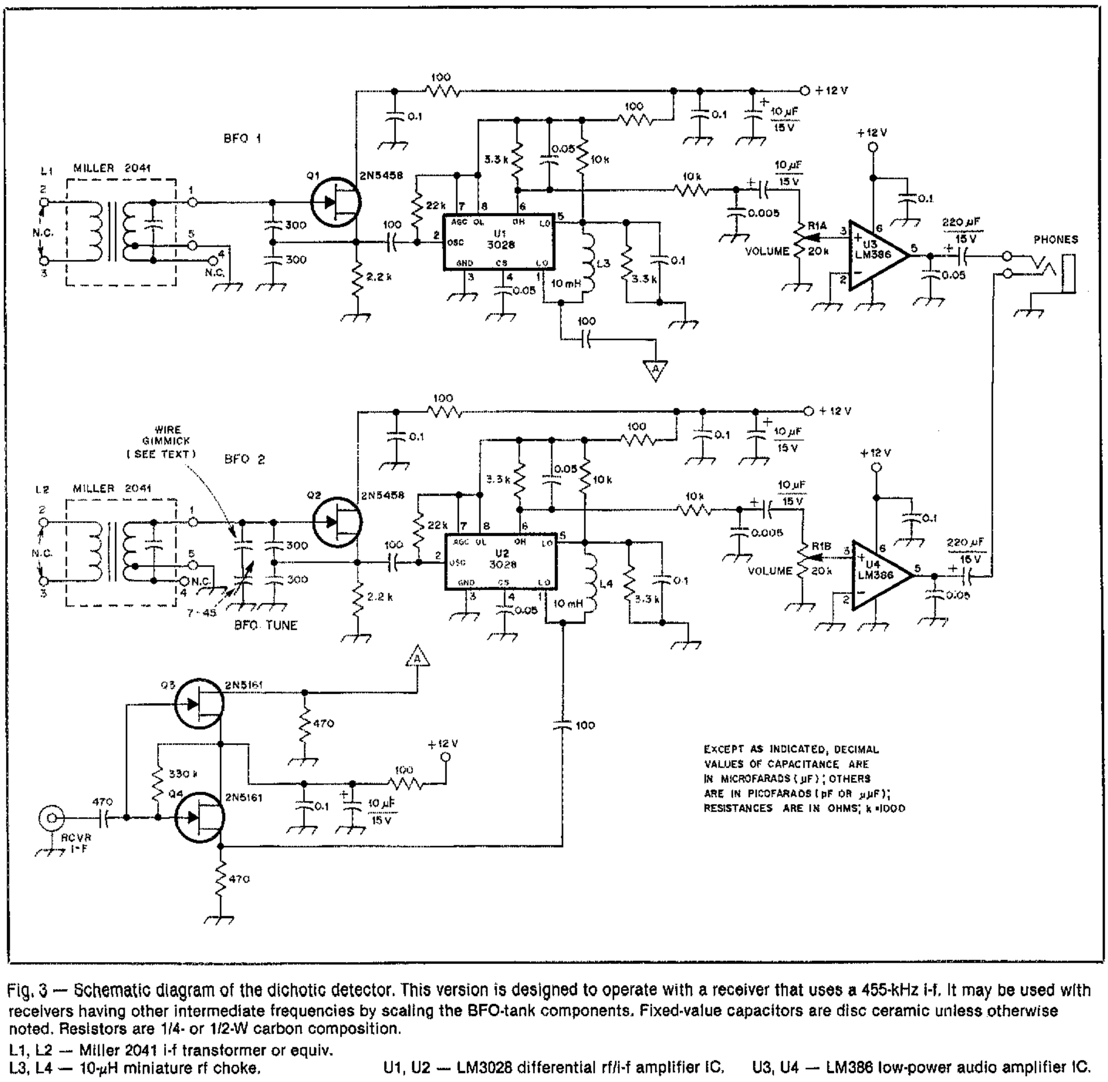


Fig. 3 — Schematic diagram of the dichotic detector. This version is designed to operate with a receiver that uses a 455-kHz i-f. It may be used with receivers having other intermediate frequencies by scaling the BFO-tank components. Fixed-value capacitors are disc ceramic unless otherwise noted. Resistors are 1/4- or 1/2-W carbon composition.

L1, L2 — Miller 2041 i-f transformer or equiv.
L3, L4 — 10-μH miniature rf choke.
U1, U2 — LM3028 differential rf/i-f amplifier IC.
U3, U4 — LM386 low-power audio amplifier IC.

detector may load the i-f output and require slight receiver retuning. As long as good rf construction techniques are employed the circuit should perform flawlessly.

The two BFOs are tuned (using L1 and L2) until one BFO output falls on the high side of the i-f passband, the other on the low side. During this tuning process C5 should be set to half mesh. With the "gimmick" capacitor adjusted to 7 pF the tuning range of C5 should be ± 1 kHz.

Table 1
Typical Detector Frequency Relationships

RF Signal	I-F (kHz)	Audio Tone Pitch (Hz) Left	Audio Tone Pitch (Hz) Right	Dichotic Pitch Difference (Hz)
7.049250 MHz	455.250	0	1500	1500
A 7.049300	455.300	50	1450	1400
B 7.049950	454.950	700	800	100
C 7.050000	455.000	750	750	0
D 7.050150	455.150	900	600	300
7.050750	455.750	1500	0	1500

Operational Technique

With the detector properly aligned and connected, turn on the receiver, select an i-f bandwidth and tune for signals. Remember the headphone volume is now controlled by an external circuit! As the desired signal is located, the two-tone pitches approach each other and merge. When the dichotic pitch difference goes to zero, only one tone centered at 750 Hz, is heard. When centered on the i-f passband, the source of the single tone appears centered directly within the head! All other unwanted signals are pairs of high- and low-pitched tones. Because the high-pitched tone is heard in only one ear and the low-pitched tone in the other, the mind easily rejects them in favor of the localized, intensified tone. Note that the dichotic pitch difference is double the actual difference between the desired signal and the unwanted signal(s) and that exaggerated separation greatly helps in the rejection process.

Atmospheric static comes through the detection system as a localized sound in the center. Under noisy conditions the use of the discord or harmony mode of detection has proven to be an effective way to "tag" the desired signal and copy through the static.

The use of a smaller frequency difference between the two BFOs would shift the center-frequency tone lower in pitch and make the discord or harmony modes easier to use. The effective harmony sensation is a neural processing function within the brain that can create an illusionary but predominate tone sensation that, although derived from two other real tones, has not actually been produced acoustically. Its location also appears to be centered in the head and permits rejection of unwanted left and right ear background signals.[6]

Conclusions

To mentally follow and decode the desired cw signal in the presence of other unwanted signals is a communications challenge. Because of the physical structure of the ear, there are fundamental limitations in recognizing differences in tone pitches that are close together. The dichotic detector takes advantage of the brain's higher-level processing of the auditory neural impulses — psychoacoustics! The result is the same as narrowing the receiver bandwidth to just a few hertz, but without reducing the speed of response, as with a high-Q filter. Furthermore, band searching is accomplished easily because the signals can be tracked readily over ± 1.5 kHz.

Notes

[1]Zwicker, Flottrop and Stevens, "Critical Bandwidth in Loudness Summation," *Journal of the Acoustic Society in America*, Vol. 22, 1958, p. 548.
[2]Dichotic — affecting or relating to the two ears differently in regard to a conscious aspect (as pitch or loudness) or a physical aspect (as frequency or energy) — dichotically.
[3]Roederer, "Introduction to the Physics and Psychophysics of Music" (New York: Springer-Verlag Press, 1975), p. 32.
[4]Eagan and Hake, "On the Masking Pattern of a Simple Auditory Stimulus," *Journal of the Acoustic Society of America*, Vol. 22, 1950, p. 622.
[5]Plomp and Levelt, "Tonal Consonance and Critical Bandwidth," *Journal of the Acoustic Society of America*, Vol. 38, 1965, p. 548.
[6]Houtsma and Goldstein, "Perception of Musical Intervals: Evidence for the Central Origin of the Pitch of Complex Tones," *Journal of the Acoustic Society of America*, Vol. 51, 1972, p. 520.

Strays

ROANOKE DIVISION PLANNING SESSION MAY 7-8

☐ This year's Roanoke Division League Planning Meeting, sponsored by the Grand Strand ARC, will be held on May 7-8 at the Holiday Inn in Myrtle Beach, South Carolina. Among the topics to be discussed are the no-code license, ARRL section-level reorganization and the Special Services Club program. ARRL President Vic Clark, W4KFC, is planning to attend. Pre-registration information is available from Dr. I. F. (Jack) Wood, Jr., WA4KGN, 903 Medical Plaza, 82nd Pkwy., Myrtle Beach, SC 29577.

ARRL PRESENTS IEEE PROGRAM

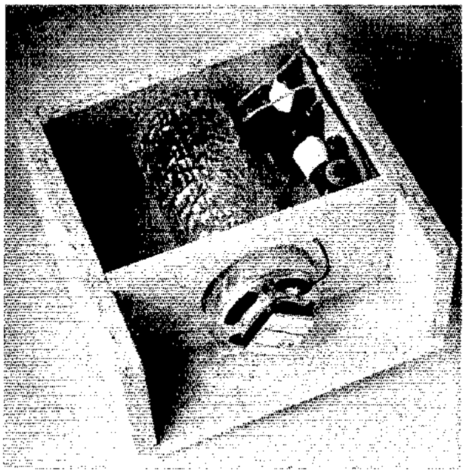
☐ Another ARRL-organized IEEE technical session was presented in December 1982. This time it was MIDCON-82, at Dallas. The theme for ARRL's Session 24 was "Modern Power MOSFETs and Applications — DC through VHF." This was the League's 16th professional IEEE session since 1972, with programs given at Intercon, Electro, MIDCON and SOUTHCON. Jerry Hall, K1TD, of the ARRL Hq. staff, gave a paper in 1982 at another IEEE conference (Globecom-82).

Among the MIDCON-82 IEEE Session 24 attendees were (l-r) ARRL TAs K7QWR, KB6QJ and W5PXH, and WB1FSB, the session chairperson.

Among the attendees at Session 24 were Technical Advisors Roy Hejhall, K7QWR (Motorola), Ed Oxner, KB6QJ (Siliconix, Inc.) and Al Markwardt, W5PXH. Dr. Joseph SooHoo (GE Co.) and ARRL Technical Department Manager Doug DeMaw, W1FB, the primary organizer of Session 24, gave papers. Marian Anderson, WB1FSB, ARRL Technical Department secretary, was the session co-organizer and chairperson.

The ARRL continues to contribute to the IEEE and the professional community in order to help stay abreast of current technology and to provide valuable public relations for Amateur Radio. This regular interface between the amateur and industrial worlds is important in these times of rapid technological advances. — *Doug DeMaw, W1FB*

An Electro-Acoustic CW Filter†



Kind of a fancy name for a simple weekend project. It's a different way to "scrub your ears."

By J. B. Heaton,* G8JFY and R. V. Heaton,* G3JIS

Some experiments we'd performed with hi-fi loudspeaker enclosures of the bass reflex (Helmholtz resonator) type generated a question: Could use be made of a pipe resonator to produce a filter for use with received cw signals? We hoped the resonator would provide good selectivity, greatly attenuate hum and hiss, be low cost and provide easy loudspeaker installation.

Design

The basic resonator principle (see Fig. 1) is similar to that of the organ pipe. An air blast is directed against a sharp edge, creating turbulence that excites the column of air of length *l*. When the column of air is excited sufficiently, an audible tone is produced. The standing wave has a node (high-pressure point) at the sealed end, and an antinode (low-pressure point) at the excited end. When the pressure wave launched at the open end reaches the sealed end, a 180° phase reversal takes place. When this wave returns to the open end, it is in phase with the next wave being launched. This condition is satisfied when the length of the pipe *(l)* is acoustically an odd number of quarters of a wavelength long: $l = n\lambda/4$, where λ = wavelength at the frequency of operation and n = 1, 3, 5, etc.

The relationship between the resonant frequency of the resonator and its physical dimensions is given by

$$f_o = \frac{V_o}{4(l + 0.3d)}\sqrt{1 + \frac{\theta}{273}} \quad \text{(Eq. 1)}$$

where

f_o = resonant frequency in Hz
V_o = velocity of sound in meters/second[1]
l = length of resonator in meters
d = diameter of resonator in meters
θ = temperature in kelvins

For typical values of V_o = 330 m/s, θ = 293 K, this equation becomes

$$f_o = \frac{122.4}{(l + 0.3d)} \quad \text{(Eq. 2)}$$

Thus, for l = 0.13 m and d = 0.06 m,

$$f_o = \frac{122.4}{(0.13 + 0.3 \times 0.06)} = 827 \text{ Hz} \quad \text{(Eq. 3)}$$

The resonator will also resonate at odd harmonics. A filter designed for a resonant frequency of 800 Hz will also resonate at frequencies of 2400 and 4000 Hz. It is necessary, therefore, to prevent harmonic frequencies from reaching the resonator.

A Practical Design

Refer to Fig. 2. A 60-mm-diameter speaker is taped to one end of a glass resonator. A tumbler 130 mm long and 60 mm in diameter was used. The loudspeaker and tumbler are mounted in a wooden enclosure as shown in the photograph. Fig. 3 is a schematic diagram of the input filter circuit. L1 through L3 and C1 through C4 form a low-pass filter. L4, L5 and C5 comprise a 700-Hz series-resonant circuit. These components help extend the filter bandwidth above 650 Hz to counter the roll-off of the low-pass sections; C5 attenuates the passage of low frequencies to the speaker. L6, R1 and R2 damp the free-air resonance of the loudspeaker, but have less effect as the resonant frequency of the pipe is approached.

Audio-frequency components designed for cross-over network use are employed in the filter circuit. Low dc resistance inductors and nonpolarized capacitors are used. The inductors should be well spaced or mounted at right angles to each other to minimize unwanted coupling.

Performance

A filter performance (not *overall* performance) plot is shown in Fig. 4. The 6-dB bandwidth points are at 350 and 900 Hz. Ultimate rejection is greater than 50 dB at 2.1 kHz. Low-frequency output is less than the voltage across the speaker would suggest, as there is a 6 dB per octave acoustic roll-off below the loudspeaker free-air resonance. The roll-off is caused by the inability of small speakers to couple effectively with the air at low frequencies.

†Adapted from a similar article that appeared in *Radio Communication* (RSGB), Oct. 1980.

*20 Tewkesbury Ave., Davyhulme, Urmston, Manchester, England M31 1RJ

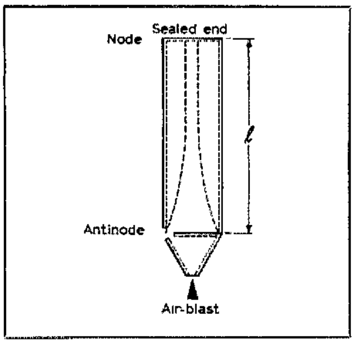


Fig. 1 — Principle of the resonator.

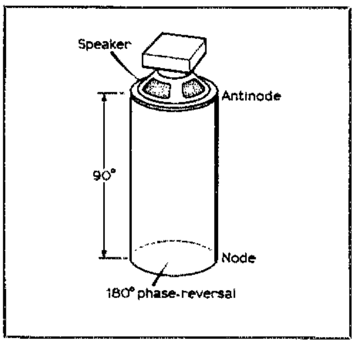


Fig. 2 — Basic form of the acoustic filter.

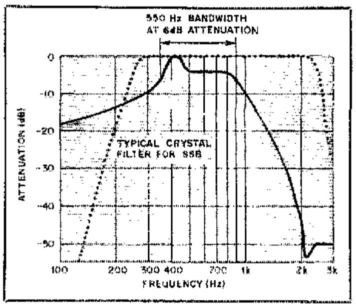


Fig. 4 — Filter performance curve. Note that this is not a curve of overall system performance.

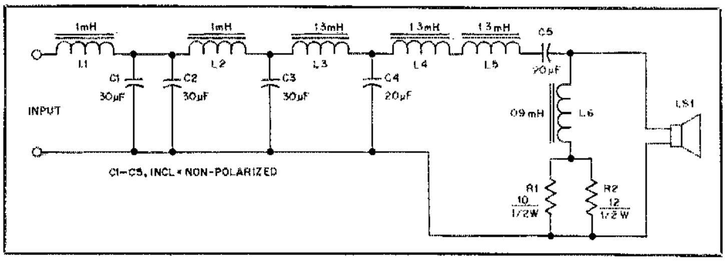


Fig. 3 — Schematic diagram of the harmonic filter.

The unit performs well. When tuning across an S9 + 39-dB marker signal, it is almost uncanny to hear the beat note become virtually inaudible when off system resonance. The steep audio filter roll-off attenuates hiss and hum from all sources. Combined with construction simplicity, don't these features make this acoustic filter an appealing project?

[1]In = mm × 0.039
ft = m × 3.28

New Products

CHANNEL MASTER HIGH-GAIN TV PREAMPLIFIER

☐ A new generation of outdoor antenna amplifiers — Spartan 2 from Channel Master — features new circuitry for higher gain, wide-range stability and lower noise. The 11-model Spartan 2 line allows uhf, vhf and fm reception in high-input as well as deep-fringe areas. Gain as great as 23 dB is provided at the TV broadcast frequencies. High level vhf signals — up to 200,000 μV (0.2 V) — are accepted without harmonic distortion or cross modulation interference effects.

Weather protection is assured by molded ABS unitized housings that can't corrode, and contain no seams or screw holes. The hinged bottom cover gives access for connection or adjustment. It is secured with a captive screw. A weather boot slips over twin lead or coaxial cable to protect connections. Terminal boards mounted at a 45° angle create a drip loop to prevent water from following lead wires into the housing.

Increased lightning protection results from layout and spacing of circuit boards, conductor channels and components. This prevents static charges from arcing through the circuit. The amplifiers are protected from power surges up to 30 kV in both directions.

Spartan 2 antenna amplifiers are designed to operate in temperatures ranging from −40° F to +140° F. They will deliver oscillation- and hum-free performance regardless of humidity and line voltage fluctuations.

To control fm interference, all the Spartan uhf/vhf/fm amplifiers use a combination of switchable and tunable fm traps. A rotatable plug is used to switch the trap in or out. The trap provides a minimum of 25 dB attenuation across the entire fm band; a screwdriver adjustment sets additional attenuation in a selected portion of the fm broadcast band. Other unwanted radio frequencies, such as marine, aircraft, CB and amateur, are blocked by band-rejection filters.

A model 0064C,7361C was tested in the ARRL lab and at my home. Several months of home testing have shown no reason for dissatisfaction. The test amplifier has been used in conjunction with a large LPDA-type TV antenna. TV and fm reception were enhanced considerably. There were no cases of amateur or CB transmissions causing TVI that could be traced to the amplifier.

Testing in the ARRL lab showed good hf band rejection. Attenuation of 10-meter energy was greater than 30 dB; and greater than 40 dB for lower frequencies. Attenuation at 2 meters is 10 dB, 20 to 30 dB with the fm trap in — depending on frequency adjustment of the trap. At 420 MHz, the test amplifier showed a bit over 8 dB gain. This fell off to −18 at 450 MHz.

In the 220- to 225-MHz range there was a slight amount of gain, about a dB or so. The 6-meter operator will not be as happy. There is plenty of gain at 50 MHz — in excess of 10 dB.

The Spartan 2 line is a considerable improvement over previous products of this sort. Spartan 2 amplifiers are manufactured by Channel Master Division of Avnet, Inc. Ellenville, NY 12428. Price class: model 0064C,7361C, $73. — *Charles L. Hutchinson, K8CH*

Hints and Kinks

Conducted By Larry D. Wolfgang,* WA3VIL

GROUNDING YOUR MOBILE INSTALLATION

☐ Immediately after purchasing a new 1982 Plymouth Reliant, I set about the task of turning it into a "ham shack" on wheels. The first job was to install the antenna. The bumper on my new car consists of a rather thin steel frame, coated with rubber. To compound the problem, the energy-absorbing system means that the bumper is insulated from the car chassis. Not much here to provide a ground plane for the antenna system. After connecting four heavy, braided ground straps between the bumper and car body, I fabricated a mounting bracket from 1/4-inch-thick steel (Fig. 1).[1] This is fastened to the bumper by means of three bolts, which go through holes drilled in the bumper (Fig. 2). Cylindrical spaces are used to hold the bracket vertical.

My antenna system consists of a Webster Band Spanner for 80 and 40 meters, and a HyGain 270 2-meter collinear antenna. Only one of these is mounted on the car at a time, with the other one stored in the trunk. I also have a magnetic-mount, 5/8-λ, 2-meter antenna for those times during ARES or RACES drills that I need to operate hf and 2 meters at the same time.

The next step in my project was to improve the ground connection between all parts of the car body and frame. Often the ground wires installed by the manufacturer are inadequate; most of the time the paint is not even removed. I use straps made from heavy copper braid, such as the outer conductor stripped from old RG-8 cable. I installed leads between as many individual pieces of the frame and body as possible. Fig. 3 shows a strap between the hood and fire wall, one from the fire wall to the exhaust pipe, and another to the carburetor. I utilize existing screws to fasten the straps, but I use star washers next to the metal to help ensure good electrical contact. Ground straps are installed between all doors and the car body, as shown in Fig. 4. The trunk lid should also be grounded, along with the axles and exhaust system.

Poor ground connections can be especially troublesome on newer cars, which make increased use of computerized control. When an Amateur Radio station is installed, the problem can be compounded when rf finds its way into these circuits. Proper grounding will help minimize the problem.

A technical-service bulletin from Chrysler suggests that low-mileage wheel bearing failure could be caused by an improper ground connection between the battery and the engine.[2] A current flows through the wheel bearing, causing the bearing and race to become pitted and blackened. This problem may occur with any type of car, not just front-wheel-drive vehicles. If you notice "road static" or "wheel static" on your radio, this could be the cause.

This may sound like a lot of work, but the results are worth it. I have almost no static noise, and mobile operation is a real pleasure. I hope to hear you on the air from your "ham shack on wheels." — *Bob Karl, W8HS, Southfield, Michigan*

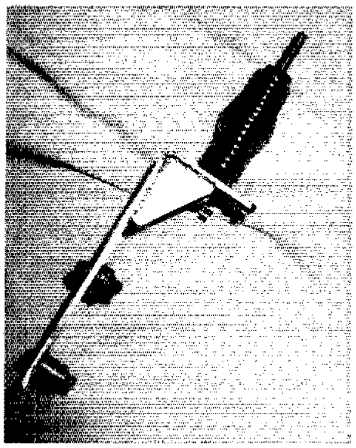


Fig. 1 — Mobile-antenna mounting bracket used by W8HS. Note the spacers used to hold the bracket vertical on the bumper.

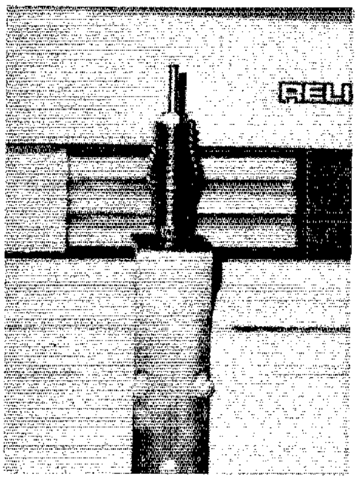


Fig. 2 — Mounting bracket bolted to the bumper of the author's car.

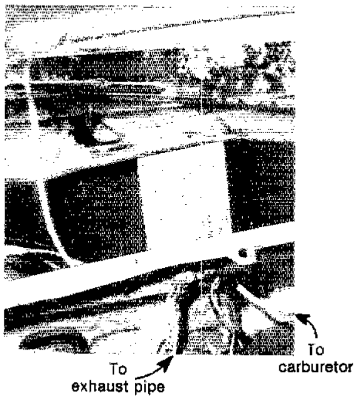


Fig. 3 — Some of the ground straps W8HS installed in the engine compartment of his car.

Fig. 4 — A method of providing a ground connection for car doors. Be sure to scrape off the paint, and use star washers under the screws.

[1]mm = in. × 25.4.
[2]Technical Service Bulletin no. 22-01-80, Chrysler Corporation, July 7, 1980.

*Assistant Technical Editor

FREEZER-BOX SPEAKERS

☐ My 2-meter hand-held transceiver doubles as a mobile rig. I just lay the rig on the seat beside me and use a speaker mike. I found that the speaker had to be held close to my ear when receiving, but that becomes tiring. Besides, I don't like the idea of driving with one hand.

I found a 4-inch speaker in my junk box. This produced good-quality audio as compared to the speaker mike. Next, I purchased a plastic freezer box at a local store. The one I bought is about 5 inches square and 2-3/4 inches deep. The plastic is approximately 1/16 inch thick and is quite flexible. I drilled a grid of 1/4-inch holes in the bottom of the box, then bolted the speaker in place. A 3-foot length of shielded cable was wired to the speaker and brought out through the side of the box. I snapped the cover in place, and the job was complete. The cable plugs into the external speaker jack, and the rig and speaker lay on the seat beside me. The improvement in audio quality is amazing. — *Stan Goddard, W6FBV, Fremont, California*

☐ I use plastic refrigerator boxes to make enclosures for small speakers. These boxes come in all sizes, so it should be easy to find one to fit most any small speaker. I cut a hole in the lid the size of the speaker cone, and drill holes to fasten the speaker. Next, I cut a piece of window screen or perforated metal to fit inside the lid. This serves as a grille. Small bolts through the lid, grille and speaker hold the assembly together. Drill a hole in the side of the box for the speaker wires, then snap the lid in place.

I glue rubber feet on one side of the box. It stands on my operating table. You could also fasten the box to the wall of your shack, or place it in your car for mobile use. — *Charles "Hub" Gilfert, W3PTM, Tamaqua, Pennsylvania*

CURING DRAKE "SCOTCH" S METERS

☐ The Drake TR4C and TR4CW rigs seem to have a reputation for having "scotch" or stingy S meters. I cured this problem with my TR4CW by removing the no. 12 lamp fuse in the antenna circuit and cleaning the contacts. I used a fine abrasive cloth to polish the two contacts on the bulb. When I replaced it, I noticed an increase in S-meter reading. Frank Walsworth, WB8DBP, tried the idea on his TR4C; the results were the same.

In the TR4C, the lamp fuse is located in the final-amplifier cage, on the top side of the chassis. In the TR4CW, it is under the final-amplifier cage, on the bottom side of the chassis. The lamp is about the size of a flashlight bulb.

In my TR4CW, the fuse is easy to reach and can be removed with two fingers. In the TR4C, it is more difficult to get at. Frank solved this problem by using a spring-type clothespin. He cut off the short end to the hollow part and trimmed the sides so it would fit in the space. This made it easy to remove and replace the fuse. — *Ernie Magnuson, KC4EP, Nalcrest, Florida*

MOBILE CONSOLE

☐ My mobile installation includes an hf transceiver, an external VFO, an antenna matching network and a 2-meter rig. I needed a compact way to mount all of this equipment to provide maximum front-panel visibility in the minimum amount of floor space possible. Stacking the units atop each other seemed reasonable. The console mount shown in Fig. 5 is the result of my effort.

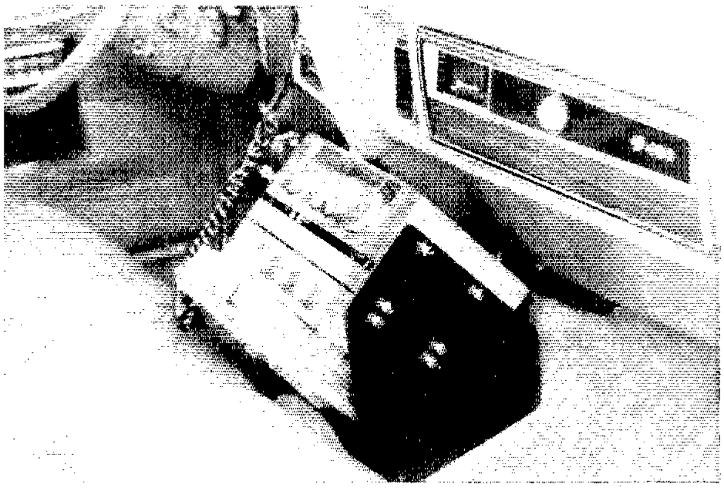


Fig. 5 — Photo of the mobile console installed in Bruce Carson's car. All the equipment is tilted up from the base to provide maximum operating ease.

The main elements are the two side panels made from 1/8-inch Masonite. They are laid out carefully, and a hole pattern is drilled to match the mounting holes in the side of the widest unit. The panels are fastened to the hf rig in my installation. Two wooden cross pieces are cut to fit between the Masonite sides. They are shaped to fit over the transmission hump. The cross pieces are glued in place and fastened with wood screws. I cut a corner out of the back of the left side panel to allow access to the power connectors and the coaxial fittings.

My VFO and antenna-matching network are the same width as my hf rig, so they are fastened directly to the sides of the console. My 2-meter rig is a little narrower; I had to use an alternative method of mounting it. I carefully drilled some holes in the bottom of my Transmatch and bolted the mobile mounting bracket for the 2-meter unit to it. Another option would have been to install a horizontal wooden cross member below the larger unit, then attach the mounting bracket for the smaller unit to it.

When the completed console (with all of the equipment in place) is installed in the vehicle, it is necessary to provide lateral stabilization. I used two small L brackets. One was fastened to the top of my hf rig, and the other was mounted to the dash. When the console is placed in my car, holes in these brackets line up, permitting me to use a no. 8 bolt as a stabilizing pin.

I have made four consoles using this construction technique. I find the materials easy to obtain and to work with. This idea can be adapted to almost any combination of equipment, and the end result is a simple, uncluttered installation. — *E. Bruce Carson, N6AQP, Westminster, California*

SIMPLE STATION CLOCKS

☐ A variety of small stick-on plastic clocks are available at most department stores. These can make a nice addition to your shack. Set the clock to UTC, stick it right on your rig and you will have a ready time reference. These clocks come in a variety of colors, so you should be able to select one that will match the color of your rig. Don't make the mistake that I almost made: The clock may fit nicely on the main tuning knob, but remember: half the time will be upside down! — *Kenneth Trettin, KAØEXG, Rockford, Iowa*

☐ Stick-on LCD clocks are being sold for $3 or less in many department stores. These make a good station clock when set to UTC. In fact, they are so inexpensive that you may want to buy two and set one to local time. Some of these clocks come with a strip of adhesive Velcro® on the back. If you buy some additional Velcro, you can put strips of the mating side at various locations and take the clock with you. The car dashboard is one likely spot. — *Dave Geiser, WA2ANU, New Hartford, New York*

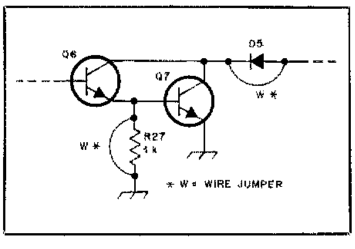


Fig. 6 — Circuit changes for the Heath HD-1410 keyer, to allow it to be used with a solid-state rig.

SOLID-STATE RIGS AND THE HEATH HD-1410 KEYER

☐ When I purchased a new solid-state rig, my Heath HD-1410 electronic keyer would not work with it. The instruction manual for my new rig indicates that the keyer must be able to pull the keyed line down to +0.4 V or less. I measured the key-down voltage with my keyer connected. The potential was +0.6 V. Two simple changes to the keyer circuit cured the problem. D5, at the collector of Q7, and R27, at the base of Q7, must be shorted out of the circuit. I soldered jumper wires around each part (Fig. 6). For those wishing to use the keyer with tube and solid-state types of rigs, a dpst switch can be installed on the back panel. The jumper wires are then connected through the switch. After this modification, the output voltage from my keyer was about +0.2. — *Steve Blomquist, KBØQK, Kansas City, Missouri*

CHOOSING WIRE SIZE FOR TOROIDAL INDUCTORS

□ "What size wire should I use to wind the toroidal inductors with?" is a question frequently asked by people duplicating various circuits and rigs I've designed. The answer is that, in most cases, you can base your choice on availability and convenience, within certain limits. But what are those limits?

Speaking only of toroidal inductors that are found in typical amateur transmitters, receivers and the like, we can describe an inductor by noting its apparent inductance (its inductance modified by stray capacitance) and Q at the frequencies of interest. That is, we could substitute another inductor having the same apparent inductance and Q and get the same performance. There are, of course, some exceptions to this rule, such as VFO tank inductors and others that may be sensitive to very small changes in temperature or physical movement.

To show what effect wire size is likely to have on these two parameters, I wound 22 turns of various sizes of wire on several T-50-6 powdered-iron cores, and measured the inductance and Q of each one at 14 MHz. A home-built test fixture having an approximate accuracy of 5% was used for the experiment. No. 21 wire, which would fit precisely on a single layer, was the largest used, and the turns were evenly spaced around the cores wound with smaller wire. The results of my tests are shown in Table 1.

Within the accuracy of the test fixture, and expected core-to-core variation, the inductors wound with wire nos. 21 through 26 are the same. There is a definite drop in Q for inductors wound with no. 28 and smaller wire, but even this drop in Q won't be noticeable unless the inductor is used in a critical application, such as a VFO tank or a high-Q filter. Single-tuned circuits, transmitter interstage and output matching and filtering networks, etc., are generally of low enough loaded Q that the difference in inductor Q is completely inconsequential. Rf transformers, which are generally wound on ferrite instead of powdered-iron cores, are even less sensitive to wire sizes unless extreme bandwidths (e.g., two or three decades) are required.

A rule of thumb that will lead to the highest Q is to use the largest wire size that will fit on a single layer, but which isn't so large that the wire won't conform well to the core. As you can see from the results of this experiment, using wire as much as several sizes smaller than specified will deliver equally good performance. — *Roy Lewallen, W7EL, Beaverton, Oregon*

Table 1
Wire Size Vs. Inductor Value

Wire Size (AWG)	*Apparent L (µH)*	*Q*
21	2.10	252
22	2.17	237
24	2.17	250
26	2.07	248
28	2.15	220
30	2.13	196
32	2.28	182

*Assistant Technical Editor

OTHER BANDS FOR THE JF ARRAY

□ I've had numerous inquiries concerning my article, "The JF Array" (Nov. 1982 *QST*). Readers wanted additional information about other bands of operation. Because of this response, I've shown applications of my array for other band combinations (Figs. 1, 2 and 3).

I also recommend that expansion of the JF or any other collinear array *not* continue beyond four elements. This is because current diminishes rapidly beyond four elements, and proper phase relationships become difficult to maintain. Also, note that higher gains are possible with the JF concept over conventional perpendicular phasing-stub arrangements because the outer elements are wide-spaced. The "clean" lines of the JF also provide improved multiband performance because little discontinuity is offered at frequencies other than what the antenna was designed for. — *Dick Schellenbach, W1JF, Reading, Massachusetts*

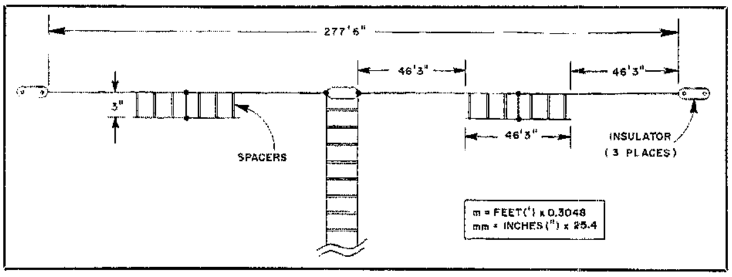


Fig. 1 — JF Array for the following bands and (theoretical gain) configurations: 30 meters — 4-element collinear (5.2 dBd); 80 meters — 2-element collinear (1.9 dBd); 160 meters — 1/2-wave dipole; 40, 20, 15 and 10 meters — extended dipole.

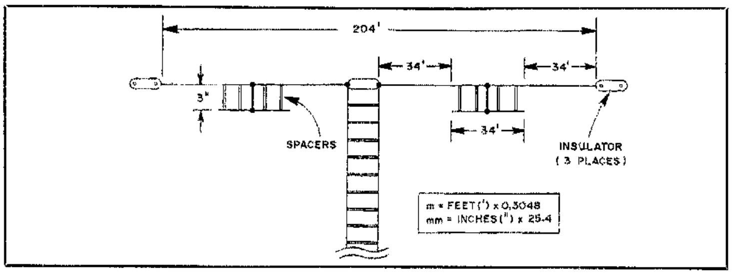


Fig. 2 — JF Array for the following bands and (theoretical gain) configurations: 20 meters — 4-element collinear (5.2 dBd); 80, 40, 30, 15 and 10 meters — extended dipole.

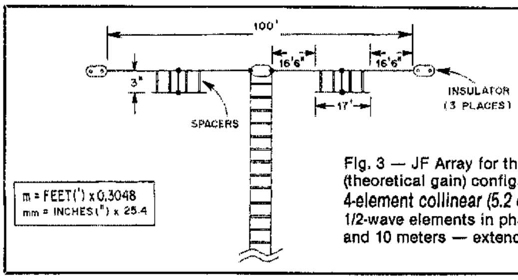


Fig. 3 — JF Array for the following bands and (theoretical gain) configurations: 10 meters — 4-element collinear (5.2 dBd); 30 meters — two 1/2-wave elements in phase (1.9 dBd); 40, 20, 15 and 10 meters — extended dipole.

A HALF TWIN-DELTA LOOP ARRAY

□ By utilizing the apex-driven Delta Loop, you can design a new, 2-element coplanar antenna array, called a Twin-Delta Loop (Fig. 4). This antenna element can be used for practical communication systems from vhf to uhf. In particular, Twin-Delta Loops can be used as

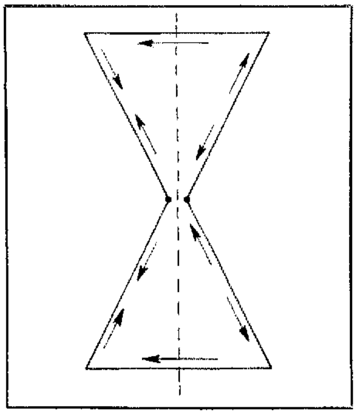


Fig. 4 — Illustration of a Twin-Delta Loop. The total length of the twin loops is approximately 2 wavelengths. Arrows indicate direction of the instantaneous currents on the antenna.

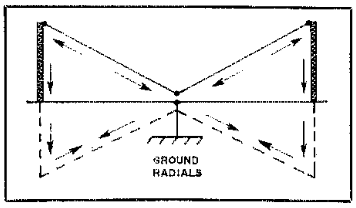


Fig. 5 — The grounded version of the Twin-Delta Loop, with the image in the ground plane.

Table 2

Half Twin-Delta Loop Dimensions

Band	*Tower Height (ft)*	*Length of Sloping Wires (ft)*
160 m†	102.03	204.06
80 m	49.86	99.72
40 m	25.92	51.84

†Optimized for 1.825 MHz.

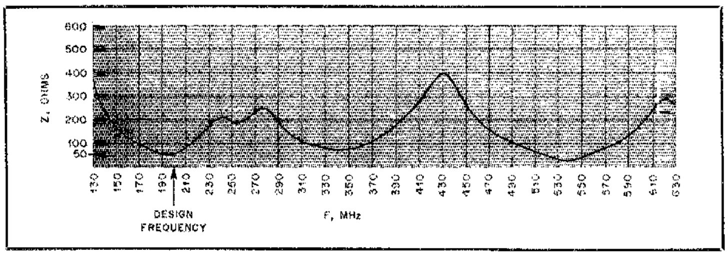


Fig. 6 — Feedpoint impedance of the (grounded) Half Twin-Delta Loop versus operating frequency for the scaled model.

elements of a Yagi-Uda array, providing superior gain and bandwidth characteristics compared to conventional full-wave loop and dipole arrays.[1] If (as with the Half-Delta Loop) this antenna element is rotated 90°, and the lower half is replaced by an image in the ground plane, we now have an M-shaped antenna (Fig. 5) that can be constructed for the hf bands. This antenna offers outstanding performance — even better than the Half-Delta Loop, which in itself is a good hf antenna.[2,3] The new antenna provides a better match to 50-ohm coaxial cable, with greater bandwidth and gain than the Half-Delta Loop.

A Half Twin-Delta Loop was modeled at 200 MHz. That is, *each* Half-Delta Loop and image in the ground plane is 1-wavelength long at 200 MHz. The masts were copper rods of 1/8-in. diameter and 11.12 in. high.[4] The sloping wires (no. 22) were 22.24 in. long. The antenna was mounted on a 98.4-ft-diameter ground screen, which was elevated so that impedance-measuring instruments could be located directly beneath it. A Hewlett-Packard 4191A rf impedance analyzer was used to measure the impedance. Measurements were made through a type-N chassis connector, fed through the ground plane to the measurement instrument located beneath. The electrical length of the feeder cable including the type-N connector was subtracted, so that the measured impedances are referenced to the feed point of the antenna.

The impedance of the Half Twin-Delta Loop was measured over the frequency range of 130-1000 MHz (Fig. 6). Loop resonance (low Z and θ equal to zero) occurred at 201, 346, 540, 755 and 869 MHz, which are close to the resonant frequencies measured for the single Half-Delta Loop over a much smaller ground plane.[5] Notice that at f_0 (201 MHz) the feed impedance is 50 ohms.

The gain in the plane of the ground screen was accurately measured, employing a Hewlett-Packard 8505A rf network analyzer. Gain was measured with respect to a carefully constructed and accurately matched (to 50 ohms) 1/4-wave monopole. The azimuthal gain was measured in two orthogonal directions — in the plane and perpendicular to the loops.

At f_0, a maximum gain of 10.2 dBi (5 dB over a 1/4-wave monopole) occurred in the directions perpendicular to the plane of the loops. The gain was 20 dB less in the plane of the loops. The gain and pattern are similar to that for two monopole antennas spaced $\lambda/2$ apart and fed in phase. At $2f_0$, the maximum gain was nearly the same (9.2 dBi), but it occurred in the plane of the loops. A deep null was not found in the perpendicular directions, for which the gain was 6.7 dBi.

The M-shaped, Half Twin-Delta Loop is bidirectional. Low-band DXers who decide to employ this antenna should therefore give consideration to installing two arrays perpendicular to each other, with a switching arrangement so that either antenna can be selected. The Half Twin-Delta Loop offers superior performance over other wire antennas, provides a good match to a 50-ohm coaxial cable and has a wide bandwidth. In a previous article, DeMaw and I stressed the multiband performance of the Half-Delta Loop, and the dimensions given were a compromise to allow this application.[6] The dedicated low-band DXer may wish to optimize this antenna for his or her favorite band. Table 2 provides dimensions for operation on the 160, 80 and 40-meter bands (based on our model antenna measurements). The bandwidth at f_0 is 12.8% for an SWR < 2:1; therefore, an 80-meter version would operate well over the entire band, without the need for an antenna-matching unit.

I would like to acknowledge the Communications Research Center and the National Research Center for use of their antenna ranges. Also, thanks to Len Bode, who constructed the antennas and made the measurements. — *John Belrose, VE2CV, ARRL TA, Aylmer, Quebec*

[1] T. Tsuliji and S. Tou, "High-Gain, Broad-Band Yagi-Uda Array Composed of Twin-Delta Loops," IEEE Conference Publication No. 195, *Antennas and Propagation*, Part 1: Antennas, 1981, pp. 438-441.

[2] J. Belrose, "The Half-Delta Loop: A Grounded, Vertically Polarized Antenna," *Ham Radio*, May 1982, pp. 37-39.

[3] J. Belrose and D. DeMaw, "The Half-Delta Loop: A Critical Analysis and Practical Deployment," *QST*, Sept. 1982, pp. 28-32. Also see D. DeMaw and J. Belrose, "Design and Practical Deployment of the Multiband Half-Delta Loop," IEEE Southcon-83 *Preprint*, Session 9, Atlanta, Georgia.

[4] mm = in. × 25.4; m = ft × 0.3048.

[5] See note 3.

[6] See note 3.

Feedback

☐ The following additions and corrections to "The Reality of Reflected Power," Technical Correspondence, Feb. 1983 *QST*, should be made. The text should read: ". . . by a distance of 90°.[5] Line voltage measured across the line is the phasor sum of the forward and reflected voltages; line current measured in series with the line is the phasor sum of the forward and reflected currents. The distance between . . ." Additionally: ". . . phase angle between the *line* voltage and current is exactly 0°. At points . . ."

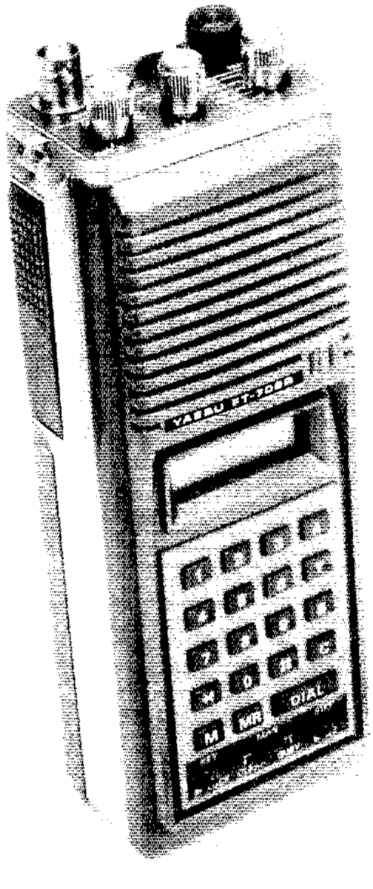
The equation in the last sentence should read $E_{\pm x} \times H_{\mp y} = P_{z}$

Author Maxwell also suggests referring to the article by Kramer, "Reflected Waves and Mismatched Loads," June 1978 *CQ*, for more information on the subject.

☐ Readers who enjoyed the article by Webb, "Electrical Antenna Null Steering," Oct. 1982 *QST*, may be interested in a coincidental British approach to the same problem. The Oct. 1982 issue of *Radio Communication* (RSGB) contains an article by Page-Jones, G3JWI, entitled "An Experimental Adjustable Null Receiving Antenna for 14 and 21 MHz."

Product Review

Conducted By Paul K. Pagel,* N1FB



Yaesu FT-708R 450-MHz FM Transceiver

Yaesu FT-708R 450-MHz FM Transceiver Serial No. 2F100398

Manufacturer's Claimed Specifications	*Measured in ARRL Lab*
Frequency range: 440.000-449.975 MHz with direct programming.	As specified.
Rf output power: 1 W HI; 200 mW LOW.	1.6 W HI; 400 mW LOW.
Spurious emissions: −50 dB or better.	2nd harmonic: −64 dB.
Power requirements: 10.8-V dc battery pack (max. 13 V).	
Current drain: rx 150 mA (20 mA squelched); tx 500 mA (HI) 300 mA (LOW).	Not measured.
Receiver type: Double-conversion superhet 1st i-f — 46.255 MHz. 2nd i-f — 455 kHz.	
Sensitivity: Better than 0.4 μV for 12 dB SINAD. Better than 1 μV for 30 dB S/N.	0.28 μV/20 dB quieting.
Audio power output: (8-ohm load): 500 mW at 10% THD.	600 mW.
Size (HWD): 6.6 × 2.4 × 1.9 in. (168 × 61 × 49 mm).	
Weight: Approx. 25 oz (720 g).	
Color: Brushed aluminum, medium gray.	
Price class: $320	

Available from: Yaesu Electronics Corp., 6851 Walthall Way, Paramount, CA 90723.

Looking for respite from the madding crowds on 2-meter repeaters? Try the 450-MHz band. I did, and found peace, solitude and a few new friends to boot. During the review period for Yaesu's FT-708R 70-cm fm hand-held, I enjoyed relatively long chats with other 450 denizens free from hustle and bustle. I wxs able to converse without the feeling that countless other users were ready to pounce once I said my final "73." It was a nice feeling!

The 450-MHz band repeaters appear to be more popular in the West, but are also found in numbers throughout the rest of the country. The repeater subbands are from 420.0 to 431.0 MHz, 433.0 to 435 MHz, and from 438.0 to 450.0 MHz. In this country, the most popular repeater channels are in the region 440.0 to 450.0 MHz. The national simplex frequency is 446.0. The repeater band plan includes a transmit/receive offset of 5 MHz, with 50-kHz spacing between repeater channels. With Yaesu's new rig in hand, I felt a welcome breath of fresh air on the 70-cm band.

The FT-708R is a durable, easy-to-use, microprocessor-controlled hand-held rig. It's built around an aluminum frame with a heavy-duty plastic case, and is pleasing to the eye. It features 1 watt of rf output, and provides 10-MHz coverage (440.0 to 449.975 MHz) in 25- or 50-kHz steps. Ten memories allow for storage of your favorite repeater frequencies.

Scanning is a nice feature of the radio. Because the 450-MHz band is generally quiet on the repeater subbands, the user can scan for a busy (or clear) channel — over the entire band, certain segments or just the memory channels. Abnormal splits may also be programmed. Frequency choices are made by the front-panel 16-digit matrix keyboard pad, and the last four digits (9.050, for example) are displayed on the large liquid-crystal display. A lamp is provided for nighttime use. Controls include OFF/VOL, SQL and TONE (for silent monitoring of busy channels), SHIFT (for transmit offset), HI/LOW power switch (200 mW, low; 1 watt, high), KEY-LOCK switch, CLEAR-MAN-BUSY scan switch (to select scanning for a clear, or busy channel) and frequency STEP switch for scanning in increments of 25 or 50 kHz. The antenna is a 1/4-wavelength flexible whip equipped with a BNC connector. The 16-digit pad also offers DTMF capability for autopatch and repeater control functions. Jacks provide for a remote speaker/mike (optional), earphone (supplied), and battery charger. The PTT thumb switch is well placed on the left side. Power is supplied by an internal NiCd battery (10.8 V, 450 mAh). A small switch in the battery compartment allows power to be applied to the microprocessor for retention of the memory channels even when the rig is turned off. Accessories included are a wall charger, and sturdy belt clip.

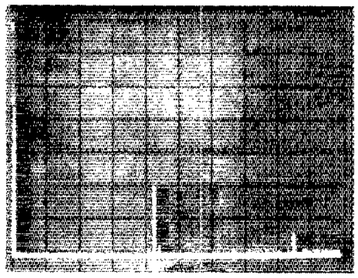


Fig. 1 — Spectral display of the FT-708R. Vertical divisions are each 10 dB; horizontal divisions are each 100 MHz. Output power is approximately 1.6 W at a frequency of 452.65 MHz. The fundamental has been reduced in amplitude approximately 50 dB by means of notch cavities; this prevents analyzer overload.

Operation

Using the FT-708R is a pleasure. Frequencies are entered from the keyboard, and must be divisible by 25 kHz. Once the four last digits of the desired frequency are entered, the DIAL key is depressed to put the decimal point in place, and ready the rig for use at that frequency. To store a frequency in one of the 10 memory channels, simply enter the frequency as described, press the desired memory channel number, and finally press M. To recall a memory channel, the memory channel number is depressed, followed by pressing MR. To return to the dial frequency, the DIAL button should be pressed.

For priority channel operation, first enter all desired memory channels for priority use. Then, enter another frequency onto the dial and recall any of the stored memory channels. Hit #. Now, the display will indicate dial frequency and every few seconds will switch to the memory frequency. If checking for a busy signal (with the CLEAR-MAN-BUSY switch in the BUSY position), the rig will lock onto the memory channel if busy. Similarly, if the switch is in the CLEAR position, the rig will lock onto the memory channel if no signal is present. To use different memory channels, with priority channels, enter the priority frequency and press DIAL, then the number of the desired memory channel and MR. Finally, press # for transceiver search of the memory channel every few seconds from the DIAL frequency.

Repeater operation is simple. Repeater shifts of +5 MHz and −5 MHz are built into the system. Nonstandard splits may be programmed easily.

Scanning is effected with the CLEAR-MAN-BUSY switch and the UP and DOWN buttons on the keyboard. Unconditional scanning is permitted in the MAN (manual) position. To scan for a clear or busy channel, slide the CLEAR-MAN-BUSY switch into the appropriate position. Limited band segments may also be scanned, a nifty feature that saves the operator from having to wade through the entire band. Memory scan is also possible when the UP or DOWN key is pressed with a memory channel on the display. This provides for selectively monitoring one's favorite repeaters.

Performance

I am an avid mountain climber who enjoys lugging varous and sundry fm, ssb and cw vhf rigs up the slopes of mountains in New England. The FT-708R proved itself to be a worthy rig because of its ruggedness and ease-of-operation. It performed well in the cold, and endured bumps and jolts in my backpack. It weighs a little more than some comparable rigs, but the "quality-feel" tips the scale in its favor. The receiver performs well and exhibits no overloading problems. The internal speaker provides full and loud audio even in high ambient noise situations, as in my pick-up truck. The transmitter works well, too. Yaesu's specs are 1 watt for the HI power position, and 200 mW in LOW. In the lab, we observed 1.6 W HI and 400 mW LOW power.

Reports showed the transmitted audio quality to be excellent, although in high-wind situations (on mountain peaks, for example) some wind noise was reported — not altogether unexpectedly! Generally, contacts reported full, deep and crisp audio quality.

The FT-708R flexible whip is small, and a *full* 1/4 wavelength, unlike many 2-meter rubber duckies. Because the band allows the use of inherently small antennas, a 450-MHz fm beam was also easy to pack-in to the mountains.

The battery lasted through many trips, and was easily replaced in the field when it was depleted. A complete battery charge was achieved in about 15 hours by means of the wall charger provided. An optional accessory, the NC-8 ac power supply and quick charger, was a decided advantage, though, allowing the battery to be quick-charged in four hours. The ac power supply allowed constant "home base" use without draining the battery. And with a special adaptor, the spare battery can be charged while the rig uses the primary battery. So whether you're at home or mountain climbing, you can enjoy constant uninterrupted use of the rig.

The manual is complete with information on operating, specifications, accessories and battery operation, and there is a maintenance and alignment section for the more technically minded. (I'm better at mountain climbing than troubleshooting.)

In summary, I would recommend this rig to anyone looking for a state-of-the-art 450-MHz fm transceiver with all of the "bells and whistles." I found no glitches, and will likely purchase one for my own fm arsenal. If you're into mellow fm operation, try 450 MHz. You'll enjoy it. — *Rick Palm, K1CE*

LANCE JOHNSON ENGINEERING D-LAY-5

☐ Before the snow started to fly in the winter of '82-'83 I decided to purchase a new rotator and get it installed. My intent was to get a rugged rotator in place that would easily turn any antenna I'd ever want to install on my tower. After some consideration, I decided on a Telex Hy-Gain T_2X (Tailtwister). Not only is it a rugged "beast," but it's good looking, too. I especially reveled at the appearance of the metered directional indicator — a welcome addition to the shack.

With the T^2X (and some other rotators) the operator must depress the BRAKE switch to retract the brake before energizing the rotator and release the BRAKE switch *a few seconds after* the rotational directional switch is released. It is a precaution that is emphasized in the rotator instruction manual. This allows the rotator/antenna combination to coast to a stop prior to brake insertion and prevents rotator damage. I imagined the awful noise and possibility of rotator damage that might occur should I forget to allow for the delay or if my finger should slip from the BRAKE RELEASE lever. That I could not allow!

Some years ago these circumstances were addressed by other owners of similar rotators. Their solutions to the problem were outlined in past issues of *QST*.[1-4] I recalled seeing a commercially available rotator brake delay unit advertised in *QST*. A letter of request to Lance Johnson Engineering for a D-Lay-5 unit was quickly answered by the delivery of same.

Installation

A glance at the installation instructions promised the job would be quick and easy. This contrasted greatly with the instructions I'd seen for the installation of another commercial rotator delay unit that involved not only rewiring of the rotator control box, but mechanical switch modifications as well!

The D-Lay-5 is neatly constructed on a rugged, single-sided pc board. A 4pdt relay, two electrolytic capacitors and two steering diodes are mounted on the top of the board. Beneath the board are three spacers secured by machine screws. Two of these screws pass through the threaded spacers. They are used to fasten the board to the rotator control box chassis.

No holes need to be drilled in the rotator control box and the existing wiring remains unchanged. Some rotator control boxes may require the physical relocation of an electrolytic capacitor, but this is not required with the T^2X control box. The D-Lay-5 is simply mounted on the underside of the control box chassis using two existing transformer mounting holes, from which the mounting screws have previously been removed. Four color-coded wires are attached to the appropriate switch connections and you're done! Within 15 to 20 minutes after installation has begun, you should have the D-Lay-5 in place and everything up and running in perfect order. You may remove the D-Lay-5 easily should you wish to restore the control box to its original condition at a future date.

[1] R. Myers, "Delayed Action Braking for Antenna Rotors," *QST*, May 1971.
[2] "Ham-M Rotator Brake Modification," Hints and Kinks, *QST*, February 1976, p. 45.
[3] "Upgrade Your Ham-M With Delayed Braking," Hints and Kinks, *QST*, September 1976, p. 40.
[4] A. White, "A Delayed Brake Release for the Ham II," *QST*, August 1977.

It Works!

Control box operation remains unchanged except that now you may release the chosen cw or ccw directional switch and the BRAKE RELEASE simultaneously once the antenna is pointed in the desired direction. You'll note that the BRAKE RELEASE LED remains illuminated for approximately four to five seconds and then extinguishes, accompanied by application of the rotator brake. No more worries about awful noises and rotator damage! Isn't that nice?

The D-Lay-5 is available wired and tested from Lance Johnson Engineering, P.O. Box 7363, Kansas City, MO 64116. Price class: $20. — *Paul K. Pagel, N1FB*

RAK ELECTRONICS "VIC-MORSE" MORSE SOFTWARE

☐ Every month hundreds of pieces of mail, including everything from new-product announcements to reports on the economy, pass through the Hq. offices. One day a new-product release caught my interest. I had just purchased a Commodore VIC-20® personal computer, and had been looking for some Amateur Radio related software.

The new-product release was from a company called RAK Electronics, and the product is a Morse code send/receive program for the VIC-20. I arranged to have a copy sent to Hq. for review.

VIC-Morse is designed to run with the standard 5K VIC-20. The program is written in BASIC, and features a send/receive speed range of 5 to 25 wpm. In the transmit mode, the special characters $\overline{AS}$, $\overline{BK}$ and $\overline{SK}$ are available, along with 4 programmable message memories. Each message may be up to 80 characters long when you're using the 5K version of the VIC-20.

Inside the Program

The program uses BASIC FOR-NEXT loops to set the cw timing sequence. A 1:3:1 dot-dash-space ratio is generated by two loops, one of length X (for dot generation), the other of length 3X (for dash generation), where X is a seed for the speed desired. While transmitting, the program uses the BASIC keyboard-buffer instruction, INKEY$, to get data from the keyboard. This data can be control signals (to set the speed, toggle from transmit to receive, or to run the message memories), or character data. Characters are displayed on the screen *after* the data has been sent. Keyboard data are buffered to 10 characters, after which the program loses data. The receive portion of the program uses an adaptive algorithm to decode the cw. It is also written in BASIC.

Performance of the transmit section of the program can only be described as mediocre at best. The programmer did not include a sidetone function (a simple addition of two POKE statements); therefore, the user must modify the program or hook up an external oscillator or transmitter to test the program. The FOR-NEXT loops used for timing *do not* give the proper 1:3:1 ratio. At 10 wpm, the code sounds somewhat natural, but above that the dashes are much too short. This is because the execution time of a FOR-NEXT loop is not directly proportional to its length (number of loops). Since characters are displayed on the screen *after* they are sent, the user has no way of knowing if he has made a mistake until the mistake is sent. Because BASIC has a slow execution time (compared to machine-code), the maximum speed, buffer space and versatility of the VIC-Morse program is limited. The receive portion of the program was not tested because the transmit section gave overwhelming evidence of the quality of the software.

Conclusions

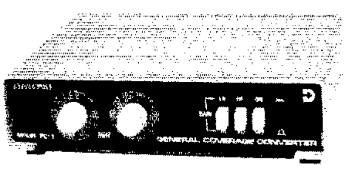
The performance of the RAK program does not justify the price tag. Anyone who has a clear understanding of the BASIC language could write and debug a program as good as VIC-Morse in a few evenings. If you're a ham who is looking for cw software — *caveat emptor!* VIC-Morse is available on cassette tape, and comes with a 9-page instruction manual and an edge connector for TU interface. Receive and transmit signals to and from the VIC-20 are TTL levels. For more information contact RAK Electronics, P.O. Box 1585, Orange Park, FL 32073. Price is $19.95, plus $2 shipping. — *Gerald B. Hull, AK4L*

DATONG PC1

Manufacturer's Claimed Specifications	*Measured in ARRL Lab*
Frequency range: 30 kHz to 30 MHz.	Confirmed.
Overall gain: −3 dB.	−4.5 dB.
Rejection of 144-145 MHz feedthrough: 60 dB.	56 dB (at 144.525 MHz, converter at 15.525 MHz).
Third-order input intercept point: 10 dBm.	18 dBm at 20 and 50-kHz spacings.
Output impedance: 50 ohms nominal.	
Input impedance: 50 ohms, but usable with short, high-impedance antennas.	
Power requirements: 10- to 13-V dc at 140 mA.	Confirmed.
Size (HWD): 1.7 × 7.2 × 6.0 inches.†	Confirmed.
Weight: 35 oz (1 kg).	Confirmed.
Finish: Anodized aluminum wrap-around case. Panels printed white on black.	
Digital dials back-illuminated with green LEDs.	Confirmed.

Optional accessories: Model MPU: Mains power unit (ac-to-dc power supply).
Lead D: Coaxial jumper fitted with a PL-259 coaxial plug at each end.
Model 270: Indoor active antenna.
Model 370: Outdoor active antenna.
Available from AR Technical Products Corp., P.O. Box 62, Birmingham, MI 48012, tel. 313-588-2288.

†mm = in. × 25.4



DATONG PC1 GENERAL COVERAGE RECEIVING ADAPTER

☐ The DATONG PC1 is a receiving converter that accepts the frequencies from 30 kHz to 30 MHz and converts them in 1-MHz bands to a receiver that is tunable from 144 to 145 MHz. Without any modification, your 2-meter receiver tuning range can be modified to include coverage of 30 kHz to 30 MHz! If you have only an hf receiver, a conventional 2-meter converter with output at 28 to 29 MHz may be used.

All that is required for hookup to the PC1 is a 12-V 140-mA dc power supply, a short length of coaxial cable with PL-259s at each end for connection between the PC1 and your receiver, and a shortwave antenna. Good signal reception is possible using an indoor antenna, but use of an outdoor antenna is recommended.

Shortwave listening probably captivates more people than anyone could imagine. After a little time searching for signals from all parts of the world, it was easy to see how interesting it was to explore the "other wavelengths." Amateur Radio operating is great, but to enjoy it fully, you have to become a good listener. This is probably why many SWLs (shortwave listeners) become hams. Having a general-coverage converter for a ham receiver provides the listener the "best of both worlds" — Amateur Radio operating and SWLing.

The PC1 is limited in performance only by the equipment you use with it. Most of the present-day 2-meter multi-mode transceivers will suffice. An ICOM 211 and a Yaesu FT-101ZD with an ARR 144VD vhf converter having a 28-MHz output were used with the PC1. The FT-101ZD has an AM position which gives better reception of a-m broadcasts, than the SSB position of the ICOM 211. If you are going to use a multi-mode 2-meter receiver, having a-m mode selection is important.

Circuit

Construction and quality of the PC1 are top rate. Components are mounted on a single fiberglass, double-sided pc board. The PC1 uses nine ICs, six transistors and 27 diodes. The rf input signal passes through one of the seven band-pass filters and on to the mixer. The mixer output is routed through a 144- to 145-MHz band-pass filter to the 2-meter receiver. If overload is caused by strong local signals, a 12-dB attenuator can be switched in. Two additional filters are used below 1 MHz, one covering 30 to 500 kHz and the other 500 kHz to 1 MHz, both switched in from the front panel.

A single crystal oscillator and LSI frequency synthesizer techniques are used to produce the required LO signals. The circuit has an adjustable trimmer that may be used to set the crystal to the exact frequency. The PC1 as shipped from the factory was found to be within a few hertz of exact received-frequency readout.

Operation

The PC1 frequency coverage is divided into 30 separate 1-MHz bands. Two front panel switches with LED-illuminated windows are all that is needed to select the first two digits of the megahertz segment you wish to tune. If you want to listen at 15.016 MHz, you set the two PC1 switches to read 15 and your 2-meter radio to 144.016 MHz. With the PC1 set to 15 MHz, you can tune to 16 MHz, corresponding to a reading of 145 MHz on your 2-meter radio. No PC1 tuning is required: The diode-switched band-pass filters covering 1-30 MHz are selected automatically when the corresponding frequency is selected. Large frequency excursions can be made with a flip of the two front panel switches. Once the favorite listening spots are found, it is easy to return to them with little effort just by dialing the appropriate frequency.

Once the "new world" of shortwave signals has been sampled, it becomes necessary to find a directory that lists information about the stations being heard. One such directory is the *World Radio TV Handbook,* which contains just about everything you would want to know about shortwave listening.[1]

Many hours of listening were devoted to this review and they were enjoyable as well as informative. Tuning the wavelengths not associated with Amateur Radio could cause a scheduling problem with my spare time. I found that using the PC1 for monitoring Amateur Radio signals on other bands (or on the same band while operating and using it as a backup receiver) made the scheduling less difficult.

Being able to use the PC1 as a ham receiver makes it more attractive. The PC1 is manufactured by DATONG Electronics Ltd., Spence Mills, Mill Lane, Bramely, Leeds LS13 3HE, England. Price class: $260. — *Bernie Glassmeyer, W9KDR*

LOGISTICS CORPORATION FIRE-FIST 1000 CW SYSTEM

☐ If you've grown weary of the relentless "sameness" of cw signals in our amateur bands, brought on by the emergence of keyboards and other computer-generated and -controlled cw transmissions, take heart. No one is required to have a perfect fist: It can be an adventure to break with tradition and send imperfect code when the mood strikes. No doubt you have become obedient to the pressure of peers who claim "everyone should use a buffered keyboard (KB)." Balderdash! How in the world will your friends know who you are if you don't have a distinctive fist? There's no challenge to copying perfect code. It's much more fun to work a station operator who has a bum fist: It keeps you on your toes!

A new manufacturer, Logistics Corporation, Inc., has recognized the need for nonconformity and free style with its Fire-Fist 1000 cw system. The keyer is computerized, which makes it possible to generate a variety of sending styles. Each cw format is selectable by actuating the appropriate push-button switch on the front panel of the keyer box. Some interesting variations are possible by punching up two or more switches simultaneously, but we don't recommend that until you've mastered the basic operating technique. The '1000 also permits sending conventional, error-free cw at speeds up to 100 wpm.

[1]Available from Watson Guptill, 1515 Broadway, New York, NY 10036. Price: $16.50.

Logistics sells a mating paddle. It contains a layered base made from four sections of iron plate. With all four base plates in position the paddle weighs 10 pounds. In this mode the operator with the legendary "kW fist" should have no problems with the paddle wandering about on the operating desk. The less aggressive operator can remove base plates until the paddle-assembly bulk is reduced to his or her liking. By using only one weight-down plate the paddle will slip about at the slightest provocation. This feature was intended for operators who frequently excuse themselves for their bad cw by saying, "this darned key won't stay put on my table." No need to fib with the Feather-Fist 100 paddle.

The "FF-100" paddle has another unique feature of worth to those who tend to become confused during the heat of a cw QSO. There is a digital display (0.5-inch blue flashing letters) located on each side of the paddle arm. One of the readouts says "dashes," and the other says "dots." The designer realized the difficulty that some cw ops have in telling which way to move the paddle for the sending of dots or dashes. (Mensa members will not need this feature.) An extra set of paddle contacts has been included. This can be used to switch the antenna for full QSK (rubber gloves are required during high-power operation to avoid rf burns). Alternatively, the spare contacts can be used by the "personality operator" who has a ham-shack ON THE AIR light above the rig.[6] This can be especially useful for impressing the visitors: The light blinks on and off in unison with the keying! This kind of razzle-dazzle lends itself nicely to some of the high-power amateur stations.

Programming Features

Switch position no. 1 is for the STUTTERING mode. Normal cw can be sent with the paddle, stored briefly in a memory, then processed and sent by a random-error generator within the keyer. The error rate can be increased or decreased by means of a panel-mounted potentiometer. This mode simulates what many operators call "rf getting into the keyer." It can also be used to create a fist characteristic that is frequently blamed on "a long, hard day at the office," or "I think I had one too many before dinner." We hear a lot of these fists on the bands these days, so this should be a popular feature for the nonconformist.

Position no. 2 is labeled BANANA BOAT. It is not too unlike position no. 3, which is called LAKE ERIE.[7] In either mode CQ comes out as NN MA, NN MA or NN GT. We've all heard this one a great many times while tuning the bands. Some refer to this as a "distinctive fist," and take considerable pride in developing this type of swing.

The tag on mode position 4 reads DXPEDITION. In this mode the station call letters are always erased from the memories, but the RST 599 and operator name remain intact. This will enable the user to confound the pileup operators as they try to learn the call letters of the station they just worked. You can keep them waiting for this vital information for hours! I like to think of position no. 4 as the "fun mode."

Next is no. 5 — the DRAG position. This enables one to send dots at several times the baud rate of the dashes. The resultant CQ sounds something like, "drah-di drah-di drah-drah di drah." Some interesting results can be had by slowing the dashes to 10 wpm and running the dots at, say, 30 wpm. We hear a lot of people doing this with bug keys these days. Trimmer potentiometers are available for changing the baud rates of the dots and dashes.

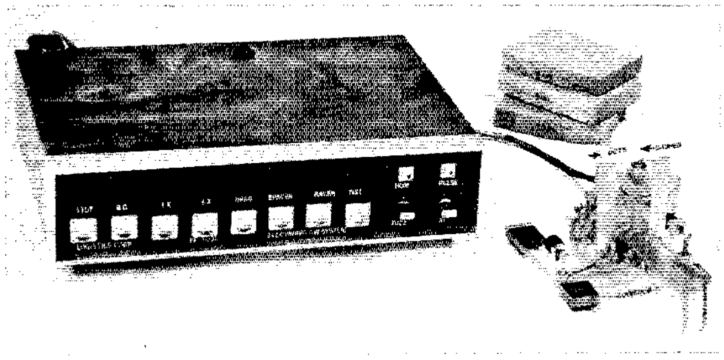
No. 6 position is called SPACER. Here is where you can duplicate much of the cw you find on the bands. Again, two trimmer potentiometers are available. One controls the gaps between letters and the other puts considerable distance between the words. It is possible to send up to 100 wpm in this mode, but to have 10 to 20 wpm spacing between letters and 5 to 10 wpm spacing between words. This technique gives the operator at the other end of the QSO a chance to fill in his log, file his nails or daydream about some new circuit while you're sending. Of course the coherence of your message may never be realized, but why worry!

No. 7 mode is entitled RACER. Irrespective of what you program into your keyer with the paddle, out comes cw that has no spaces between letters and words. By advancing the potentiometer called "race increase," you can actually start piling one character atop the next one to make things even more interesting. There must be a number of Fire-Fist 1000s in service already, for I've been hearing a lot of this type of cw lately.

Mode no. 8 is labeled NST. In this position the computerized keyer enables the operator to test his rig endlessly without identifying his station. Instead of the conventional word "test" going out over the air, the more common NST NST NST is heard. Although this style of operating is illegal (no call letters included), it does seem to be a popular technique in some circles.

Other Features

The foregoing eight program modes represent the main features of the Logistics Corp. system keyer. However, some additional switches are available. (1) Introducing hum on the cw carrier. This provides notes ranging from T3 to T8, depending upon the setting of the "fuzz" pot. (2) There is a switch that controls chirp (requires a coaxial cable to the local oscillator or synthesizer VCO in your transmitter). A varactor diode in the keyer pulls the transmitter oscillator each time the key is closed. The degree of chirp can be regulated by adjustment of still another internal trimmer potentiometer. Finally, there is a PULSE mode switch that can be used to superimpose pulses on the leading and trailing edges of the cw waveform. This will cause key clicks that are difficult to equal by normal means. This is actually a center-off switch. In its opposite function position it creates the effect of excessive shaping, thereby rendering a bell-like tone to the cw note. This type of soft keying imparts a cw note that sounds as if it were coming from the bottom of a deep well. This is great in a pileup, for no one can figure out what your call is!



Logistics Corporation Fire-Fist 1000 CW System

Manufacturer's Specifications

Dimensions (HWD): Keyer unit, 8 × 12 × 16 inches (203 × 304 × 406 mm).
Weight: 8 pounds.
Color: Two-tone gray and black.
Power requirements: 117-V ac at 35 watts.
Output voltage at key-line jack: 5-V dc (negative).
Price class: $290. Dealer inquiries invited.
Manufacturer: Logistics Corporation, Ltd., 70 Braeburn Rd., Bristol, CT 06010.

Electronics Features

Apart from the usual massive collection of OR gates, AND gates, NAND gates, ROMs, PROMs and what have you, the FF-1000 contains 17 go-no-go ICs. These are essential in generating erratic cw. There are also six kant-read ICs in the system — a totally new concept in the LSI technology. These are used to generate cw that can't be deciphered by the most skilled of high-speed operators. Code readers with visual displays or video terminals can't handle this type of cw either.

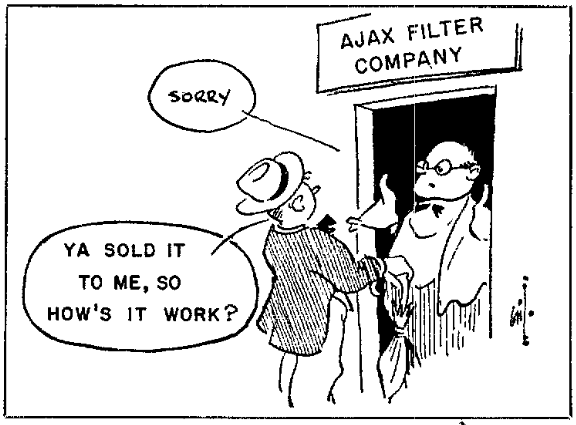
Summary Comments

I highly recommend this system to those who have jaded cw appetites, or wish to do their own thing on the cw bands. The manufacturer is offering a 30% discount for quantities greater than 10 units. This should be especially appealing to the instructors of Amateur Radio classes. The discount deal should be popular also with DXpedition groups. After two months of home use with this system I have finally broken the habit of conforming. No longer do I have that boring fist that puts people to sleep. Now they have to pay attention to what I'm sending, and at last I have a captive audience. — *Zender Bawdrite, YØOP*

[6] See the product review of the Gargler Corp. Profundo 10X Microphone in April 1981 *QST*, p. 49.
[7] a.k.a. "Missouri Valley Swing."

BEGINNER'S BENCH

Understanding and Using Audio Filters



Wondering what an audio filter can do for you? Confused by the plethora of audio filters on the market? This rundown may answer your questions.

By Doug DeMaw,* W1FB

Your friends keep telling you to buy an audio filter to reduce the effects of QRM. But, you've never used one, and you're not sure if such a contraption would really help when the going gets difficult on your favorite cw or phone band. Not only that, you keep hearing about *passive* and *active* filters, and you don't comprehend the difference between them. And, which of the published circuits for home projects would be suitable if you wanted to build your own filter? What about the numerous commercial filters being advertised in the amateur literature? Golly, there sure are a lot of questions that need to be answered, aren't there? An understanding of these filters and how they are best applied should do much toward helping the beginner to make a proper judgment. This article answers some common questions asked of the ARRL staff. I hope it clears up any doubts that may be in your mind.

Types of Audio Filters

I got my introduction to audio filtering back in the 1940s after WW-II surplus radio gear appeared on the market. A popular, low-cost audio filter was available under its military designator, FL-8. It was a passive unit that plugged into the receiver phone jack. A pair of headphones was plugged into the filter box, and away you went with "laundered" audio. The term "passive" simply means there are no devices in the circuit that require an operating voltage. In other words there are no tubes, transistors or ICs that need operating power to make the device work.

An *active* filter, on the other hand, contains electronic devices that require an operating voltage at a specified current. By today's standards the FL-8 aircraft range filter would represent a grave disappointment to the user, but in those days it was definitely better than having no filter at all! It was lossy and it didn't sharpen the receiver output enough to make *anyone* ecstatic. But, it did help to reduce the effects of QRM.

Present-day audio filters offer all manner of high-technology features. The more expensive ones offer various bandwidths and response characteristics. For example we might purchase or build a filter that yielded a low-pass, high-pass or band-pass response, all at the flick of a switch. Other filters are less complex and have only a band-pass response. Such a filter might have up to four bandwidth (selectivity) choices, switch selected. For my cw work I am content with a band-pass audio response.

Perhaps you're wondering how the three filter responses compare. These are depicted in Fig. 1. The curve at A represents how a low-pass audio filter might respond as an audio signal generator was tuned across it. For this example let's assume that the peak (loudest) frequency we want for cw reception is 700 hertz. The filter must then be designed to start rolling off (attenuating) at 700 Hz. As our signal source was moved higher and higher above that frequency, the output from the filter would become lower and lower in level (attenuation). Eventually we would scarcely be able to hear the higher-frequency audio signal. Since our low-pass filter permits all frequencies that lie below 700 Hz to pass without attenuation, the name "low pass" is proper. A low-pass filter will greatly reduce high-pitched QRM energy, but if a signal reaching your receiver creates a tone that is 700 Hz or lower, it will sound just as loud in your earphones as will the desired 700-Hz beat note. Therefore, a low-pass filter will help to reduce some of the QRM, but not all of it. Also, since low-frequency responses will be heard well, static (QRN) will sound as loud as it would with no filter at all. Not much use for QRN reduction, eh? I find a low-pass filter more useful for ssb (single-sideband) reception. If the cutoff frequency (f_{co}) is set for approximately 2000 Hz, the audio from the filter will sound quite acceptable, but the high-pitched "monkey chatter" QRM from stations nearby in frequency will be reduced greatly. Likewise, of course, with high-pitched cw QRM. The sharpness of the rolloff will depend on just how the filter is designed.

Now, let's examine the opposite condition, where the audio filter is built for a high-pass response (Fig. 1B). The same philosophy applies except that the frequency responses below 700 Hz will be attenuated, but those above our chosen frequency will pass unhampered. This type of response characteristic can be useful when a large amount of filtering is not wanted during ssb operation (to preserve

*ARRL Senior Technical Editor

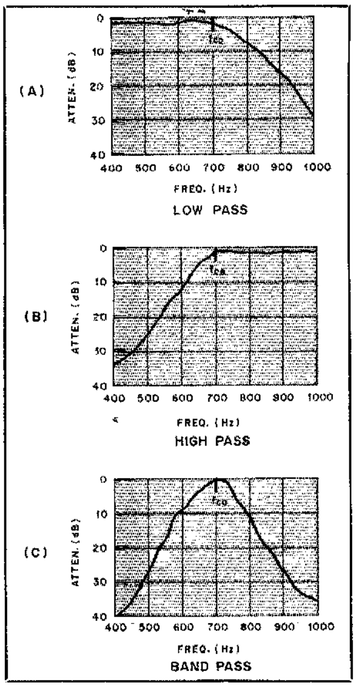


Fig. 1 — Relative response curves for (A) a low-pass filter, (B) a high-pass filter and (C) a band-pass filter.

the fidelity as much as possible), but when we want to reduce any low-frequency (below 700 Hz) rumble (QRM and QRN) that might annoy us or impair readability. I have never found the high-pass audio filter of much use in my shack, but some operators like to use them.

Our remaining filter option is the band-pass type. I find myself "married" exclusively to this kind of filter for cw and ssb work. A typical band-pass response is shown in Fig. 1C. A center frequency (f_c) of 700 Hz would be the choice of many cw operators, since that corresponds to the offset frequency for most of our modern transceivers: An audio filter should have a peak response at or near the same frequency that the receiver BFO (beat-frequency oscillator) provides. This will ensure maximum audio response in the headphones. If you don't know what the offset is for your rig (between the transmit and receive modes) you can tune in a cw signal while using your built-in cw i-f (intermediate-frequency) filter and adjust the tuning dial for peak S-meter response. The pitch you hear will be the approximate frequency to which your audio filter should be set (or designed, if a fixed-frequency filter is used). Most of the better commercial audio filters have a frequency control that permits you to match the audio-filter peak response to that of the receiver offset. Therefore, the correct procedure would be to tune in the cw signal without the filter, then activate the filter and adjust it for maximum response in the speaker or phones.

It is easy to realize the advantage of a band-pass filter over the two other types: QRM and QRN above and below the desired audio frequency are greatly reduced by virtue of the rolloff above and below f_c. This kind of filter generally consists of a low-pass and a high-pass filter connected together electrically. A band-pass response results. It is called a band-pass filter because it will pass only a narrow band of frequencies before attenuation commences. Some electronic filters are designed for an "inverted" band-pass response. They are called "band-reject filters." The approximate response would look like the one in Fig. 1C, except that the curve would be turned upside down. There would be no attenuation above and below the desired frequency range, and that energy within the dip in the curve would be "sucked out." Band-reject filters are used in special applications in which a narrow band of interfering frequencies must be attenuated or removed.

Some commercial and homemade filters have still another function. They are equipped with a "notching" feature. A switch position on the panel is marked "notch." There is also a notch-adjustment control. The function of this circuit is exactly what it is called: It notches out or essentially removes the audio response at a *specific* pitch. Therefore, if you are listening to a cw signal that is peaked at 700 Hz, and there is an annoying QRM beat note at, say, 1000 or 1500 Hz, the notch control can be adjusted to knock down the unwanted pitch without impairing the strength of the tone you are listening to. In my experiences I have applied the notch feature to provide perfect copy, when without it I could not read the station I was in QSO with.

Finally, some filters are equipped with a "Q control." The term "Q" in electronics refers to the relative quality of a circuit or component part. In plain language we can say that a circuit with high Q has low losses. In the filter world we find that the higher the Q the narrower the filter response. Therefore, if we were to increase the Q of the filter that provided the band-pass response of Fig. 1C, the sides (skirts) of the filter response curve would pull in toward the center frequency. This would help to reject unwanted signal energy even closer to the desired signal. An ideal filter response would have skirts that were perpendicular to the base line of the curve, and the top of the curve (nose) would be flat. The resultant response would be rectangular, rather than sloping as it is in Fig. 1.

Filters with Q controls enable the operator to choose the desired amount of bandwidth. Since ssb reception requires a fairly wide bandwidth (300 to 2000 Hz, for example), to ensure reasonable speech reproduction and comprehension the Q control would be set for a low level in order to widen the filter response. Conversely, an ideal cw signal has no appreciable bandwidth, so the Q control would be advanced to narrow the filter response as much as possible.

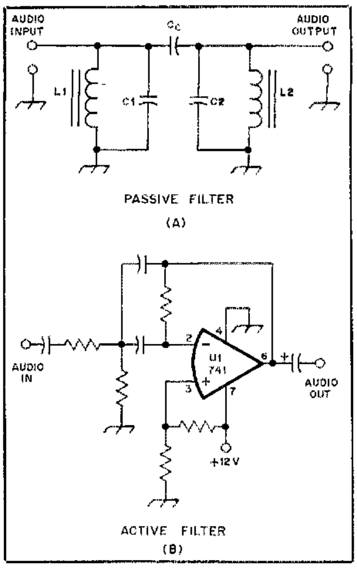


Fig. 2 — Simple examples of (A) an L-C type of passive audio filter and (B) one section of an R-C active filter. The latter requires an operating voltage.

Fig. 2 shows the essentials of an active and a passive type of audio filter. The circuit at A shows a simple two-section passive filter that employs two resonant circuits (C1/L1 and C2/L2). C_c is the coupling capacitor that joins the filter sections. All passive filters have some loss (insertion loss). On the other hand, active filters can be designed to provide gain. However, most designers set them up for a gain of 1 (unity) so that the gain distribution of the receiver remains the same as before the filter was added. The example at B of Fig. 2 shows one section of an active filter that uses an operational-amplifier IC (op amp). Usually, we will find two to four identical filter sections of this type in a band-pass R-C active filter — R-C meaning resistors and capacitors are used to establish the center or cutoff frequencies rather than elements of inductance (L) and capacitance (C). The latter would be called an "L-C filter." The more filter sections used, up to a practical limit, the steeper the sides (skirts) of the filter curve. A single-section filter of the kind shown at B of Fig. 2 would be of little value by itself, since the skirts of the response curve would be rather wide. They could be narrowed considerably by designing the simple filter for a high Q, but this would cause design complications. It is more practical to use two or more low-Q filter sections in series to obtain good skirt response.

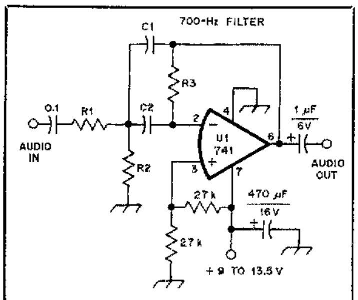
What Will a Filter Do for You?

We already learned that an audio filter will help to banish unwanted signals and QRM. But what else would prompt an amateur to use one of these gadgets? Well, all receivers put out wide-band noise. This noise is generated after the i-f filter, which usually follows the mixer stage. The i-f amplifiers generate this noise, and sometimes the audio stages in a receiver contribute further to the noise. This energy shows up as a hiss in the speaker or phones. It doesn't cause a problem unless the signal we want to copy is weak enough to be on par with the receiver noise level or slightly above it.

Also, background noise on a readable signal can be annoying (I detest it!). An audio filter will greatly reduce the effective noise bandwidth of the receiver and provide "solid copy" in many instances when partial copy would otherwise result. The received signal, irrespective of its strength, will sound much cleaner after the receiver output has been "sanitized" with an audio filter. The filter can effectively "lift" a weak signal out of the noise much of the time, while at the same instant knocking out QRM from other stations.

I can remember the time I was on a DX-pedition to Montserrat (VP2MFW) and had high hopes of working the world on 160 meters. The QRN level near the equator, especially at 160, 80 and 40 meters, is ferocious. It seems to average between S9 and 10 dB over S9 on 1.8 MHz. The antenna was a closed square loop, 1000 feet on a leg and 60 feet high (m = 0.3048 × feet). Having such a magnificent antenna available (courtesy of VP2MF) was like perishing and awakening in Shangri-la. After warming up the station gear I called "CQ" on 160-meter cw. I could hear people answering me, but nary a call came through in readable fashion. After a few more tries I gave up in despair: The static was just too hard to deal with. Suddenly I recalled packing my active audio filter in one of the suitcases, just in case it would be needed. Shortly thereafter I garnered it, hooked it to the receiver and tried another "CQ." Wonderful! I pulled two stations out of the noise and copied them Q5. The rest of the evening was devoted to working as many stations as possible, even though many of them were still too weak to copy, even with the filter actuated. This is but one example of the value of a good audio filter!

Another receiver malady that can be aided or disguised with an audio filter is output hum. Some receivers — especially when hi-fi phones are used (they respond well to low-frequency audio) — cause weak-signal copy to be difficult because of high levels of 60- or 120-Hz hum in the audio output. The low-frequency rolloff of a band-pass or high-pass audio filter will block the passage of low-frequency hum.



Q = choose a value

H_o = desired gain

$\omega_o = 2\pi f_c$

f_c = desired center freq. (kHz)

C1, C2 - choose a value (µF)

Then $R1 = \frac{Q}{H_o \omega_o C}$ = k ohms

$R2 = \frac{Q}{(2Q^2 - H_o)\,\omega_o C}$ = k ohms

$R3 = \frac{2Q}{\omega_o C}$ = k ohms

Fig. 3 — Design information for building your own active audio filter. Two, three or four of these circuits are normally connected in series to make the skirts of the filter response steeper.

Building Your Own Active Filter

An earlier *QST* article described an inexpensive, easy-to-build audio filter for cw work.[1] A parts kit is available for that project.[2] But, if you want to start from scratch and try your skill as a designer you can work with the information in Fig. 3. Let's assume that f_c will be 700 Hz (0.7 kHz), Q will equal 5 and the gain (H_o) will be unity (1). We must pick a value for C1 and C2. Let's use 0.01 µF, since that is a value we have on hand. From this data we can work the equation in Fig. 3 and learn the values for the three resistors.

$$R1 = \frac{5}{1 \times (6.28 \times 0.7) \times 0.01} = 113.7\ \text{k}\Omega \qquad \text{(Eq. 1)}$$

$$R2 = \frac{5}{49 \times (6.28 \times 0.7) \times 0.01} = 2.3\ \text{k}\Omega \qquad \text{(Eq. 2)}$$

[1]Notes appear on page 48.

$$R3 = \frac{10}{4.396 \times 0.01} = 227.479\ \text{k}\Omega \qquad \text{(Eq. 3)}$$

We can see that these are not standard resistor values. Therefore, to ease our design effort we will use resistors of the nearest standard value. R1 will become 110,000 ohms (110 kΩ), R2 = 2200 ohms (2.2 kΩ) and R3 = 220,000 ohms (220 kΩ). These changes in value will not spoil the filter performance appreciably, but will shift the f_c slightly.

One filter section will not be sufficient to help us cope with QRM and noise. A modest filter of amateur value can be formed by placing two sections (from Fig. 3) in series (cascaded). Three sections would be even better, and I would suggest that no more than four filter sections be used. When identical filter circuits are combined it is important that the values of C1, C2, R1, R2 and R3 be closely matched in value throughout the composite filter. Resistors and capacitors of 5% tolerance are usually okay. If the values aren't well matched, the bandwidth of the filter can become too wide, since each filter section will end up with a different f_c. If you have access to a resistance/capacitance bridge, it will help to match your capacitors and resistors as closely as possible before inserting them in the circuit. Composition resistors can be changed in value by filing away some of the carbon inside them. A few passes on the side of a resistor body with a file will increase (raise) the resistance. File and remeasure the resistance until you obtain the desired value. The "wound" on the resistor should then be sealed with glue or epoxy cement. C1 and C2 need to be high-quality (Q) capacitors. Mylar or polystyrene units are recommended.

A Practical Filter Circuit

The filter described in the article of note 1 is shown in Fig. 4. I thought it would be wise to include it in this article in the event your *QST* file doesn't go back far enough. Detailed information and the circuit-board pattern appears in the original article. The circuit contains three active sections and is designed for a center frequency of 750 Hz. You can change the f_c to one of your choice by applying equations 1, 2 and 3.

Radio Shack TL081 op amps can be substituted directly in place of the somewhat generic 741 op amps. The former have FET input circuits and are somewhat less noisy than 741s. But, they cost a bit more!

Hooking the Filter to Your Receiver

Our most convenient installation technique is that of plugging the filter into the receiver phone jack. Headphones are connected to the output of the filter. A 9-volt battery can be used as the filter

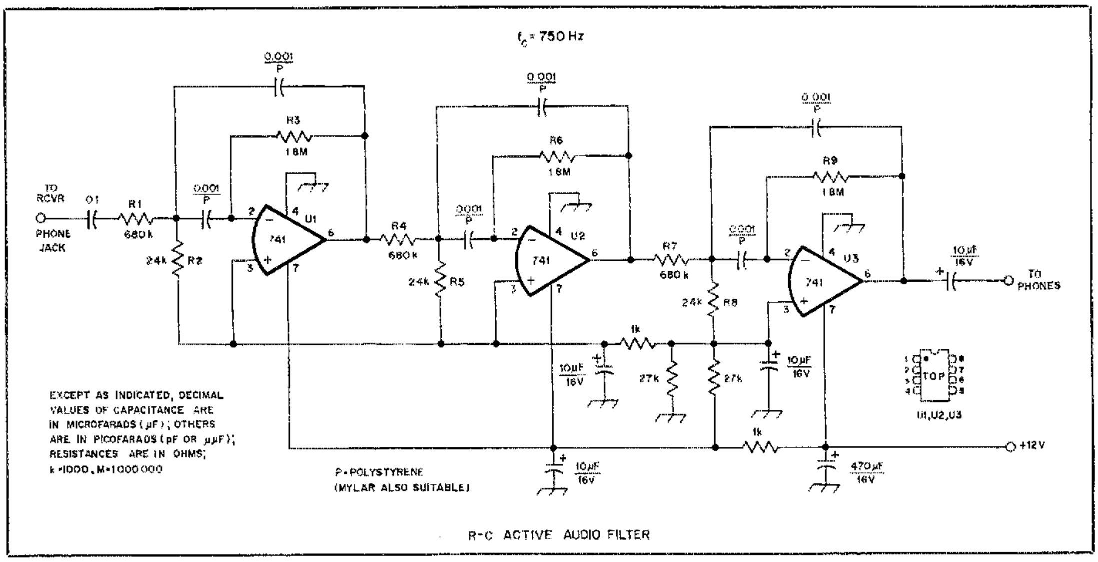


Fig. 4 — A practical three-section R-C active cw filter. Resistors are 1/4- or 1/2-watt carbon types. Capacitors with polarity marks are tantalum or electrolytic.

power supply, and it should last a long time with the circuit of Fig. 4. Alternatively, you can attach a small ac-powered direct-currrent supply. Anything from +9 to +13.5 V dc will work fine.

A switch can be used to turn off the filter and bypass the audio line around the filter when conventional reception is desired. The Circuit Board Specialists kit (note 2) provides that feature.

Care must be taken during use to prevent the filter from being driven too hard by the audio from the receiver. Overdrive can spoil the filter performance and cause objectionable distortion. *Caution:* Use only enough receiver audio output to provide a comfortable listening level in the headphones. You can design your filter for a gain of 2 or 3 if you wish. This will enable you to feed less audio into the filter, while still realizing ample volume in the phones.

A better location for an audio filter would be between the receiver output stage and the low-level audio amplifier that precedes it. But, few of us like to chop into the circuit of our store-bought gear, so we just connect the filter outboard fashion. Many of the commercial active filters contain audio power amplifiers that let us use our loudspeakers. You can learn more about these units by reading *QST* ads and product reviews.

Closing Remarks

I've thumped pretty hard on the subject of filters, so it's time to QRT. I hope you found the answers to some of your questions in this article. Audio filters are easy to build and get operating. If you have not tried your hand at home construction of ham gear, this can be a great first project for you. But more importantly, a good filter can make you king of the weak-signal copiers in your area — assuming that the other ops haven't beat you to the draw in this game!

Notes

[1]D. DeMaw, "Beating Rotten QRM — CW Filtering for the Beginner," *QST*, Oct. 1981.

[2]Circuit boards, negatives and parts kits for the circuit of Fig. 4 are available from Circuit Board Specialists, P.O. Box 969, Pueblo, CO 81001.

Strays

ATV/SSTV FORUM AT DAYTON

☐ Amateurs attending this year's Dayton (Ohio) Hamvention (April 29-May 1) are invited to attend an ATV/SSTV forum on April 30 at the Hara Arena. Stanley Brokl, N2YQ, senior engineer for the Jet Propulsion Laboratory in Pasadena, California, will give a talk titled "The Ultimate SSTV DX." Stan is an ARRL section manager for Los Angeles and a member of the JPL ARC, which retransmitted SSTV pictures from the Viking and Voyager space missions on the amateur bands. Stan's talk will be augmented by slides and pictures never before seen.

Also, Jim Chladek, KA2NSJ, and Larry Horne, N2NY, will give a talk and video demonstration on how ATV is used in putting on the cable television show "Network Two New York," which regularly features new innovations in Amateur Radio. Moderator for the forum is Ron Flynn, KB8LU, SSTV editor for *Worldradio*.

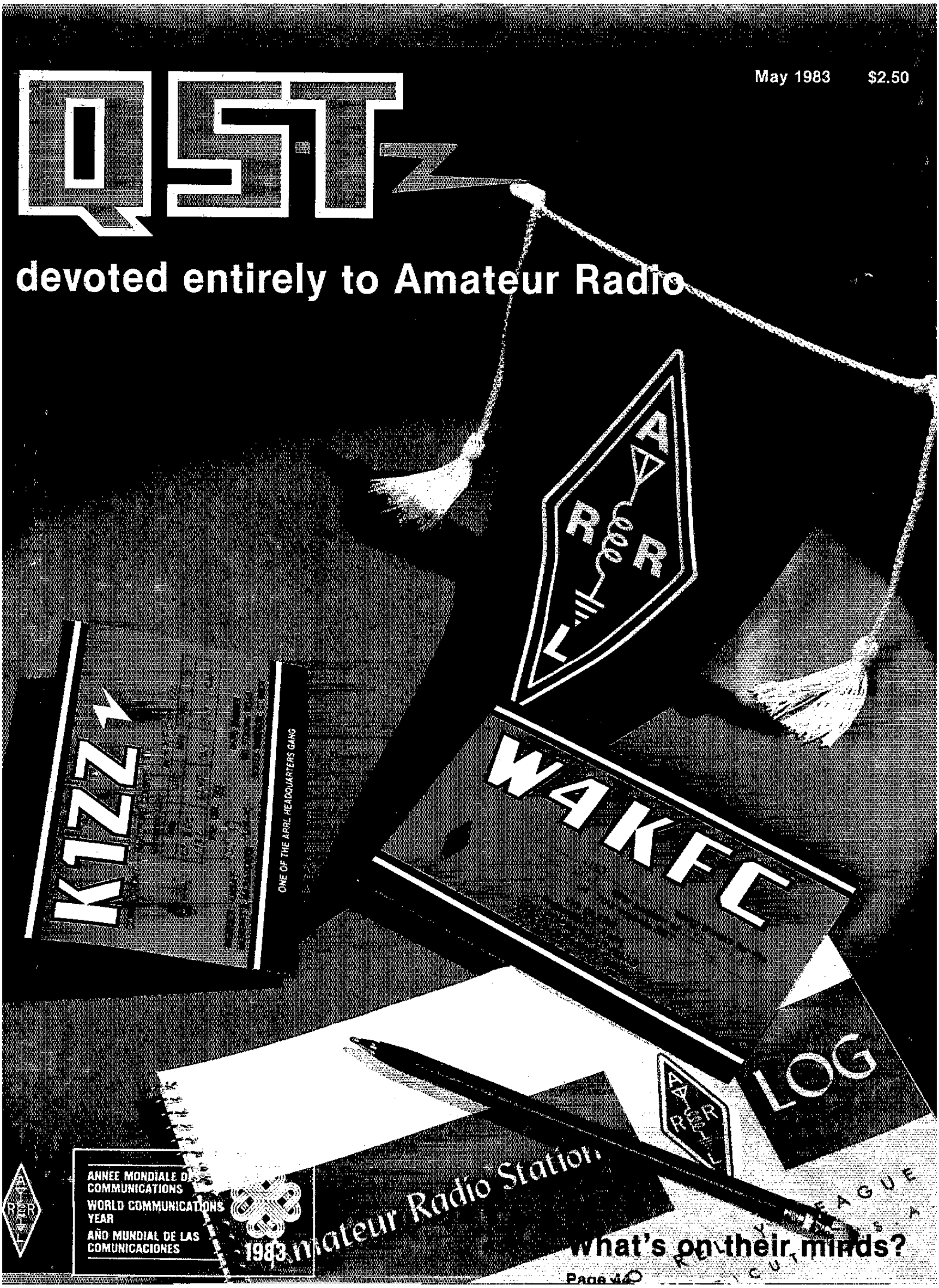
The annual SSTV Get-Together will be held on April 29, beginning at 7:30 P.M., at the Holiday Inn North of Dayton. There will be a social hour, followed by demonstrations of homemade SSTV equipment. — *Ron Flynn, KB8LU, ATV/SSTV Chairman, Dayton Hamvention*

I would like to get in touch with . . .

☐ any cab drivers in the U.S. who are interested in joining a net. Lionel Rose, G4KAB, 164 Regal Way, Harrow, Middlesex, England.

☐ amateurs engaged in air-traffic control who are interested in joining a net. International Federation of Air Traffic Controllers Assn., P.O. Box 196, CH-1215 Geneva 15, Switzerland.

☐ any teenagers in North Carolina, South Carolina or Virginia who are General class or higher licensees and are interested in joining a 75-meter phone net. Jim Lea, KA4UBN, 709 Raleigh St., Roanoke Rapids, NC 27870.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What's on their minds?

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The Weather that Brings Vhf DX

Weather changes can bring spectacular DX openings to the vhf bands. What causes this, and how can you tell when DX conditions are right?

By Emil Pocock,* W3EP

Radio amateurs have known for a long time that weather affects vhf propagation.[1] Under certain atmospheric conditions, contacts of more than 1400 miles over land and 2500 miles over water have been made on the 144, 220 and 432-MHz bands.[2,3] This type of vhf DX is called tropospheric enhancement, or simply "tropo," because it occurs in the weather-producing lower part of the atmosphere called the troposphere. Signals propagated by tropo can be quite strong, and it is not unusual to hear 10-W stations with small antennas making 1000-mile contacts under the right conditions.

Tropo is the most common form of propagation in the bands above 144 MHz, but many vhf amateurs use tropo without realizing its full potential. Although weather changes can dramatically affect propagation, only two weather systems cause most true DX openings: the sluggish high-pressure region that appears most frequently in late summer, and the springtime wave cyclone (see Table 1).[4] Other weather conditions can also improve vhf propagation and sometimes create brief DX openings. Warm fronts and cold fronts bring enhancement to local areas (restricted to a radius of a few hundred miles), but they rarely bring DX conditions. Several unusual types of winds also have DX-producing possibilities.

The Causes of Tropo DX

Tropo DX is quite different from the more familiar ionospheric skip. Ionospheric skip occurs when radio signals are reflected or refracted by a region of ionized air 50 to 250 miles high. In contrast, tropo DX is caused by a layer of warm and often very dry air, known as a *temperature inversion,* that is usually no more than 1 mile high. Normally, the temperature and moisture content of the lower atmosphere decrease with altitude. In an inversion, the temperature suddenly rises and moisture content often decreases dramatically. Inversions may begin right at ground level and rise for a few hundred feet. Most inversions that are useful for propagation, however, are found at altitudes of 1500 to 6000 feet. Fig. 1 shows the temperature and moisture profiles of a normal atmosphere and of two inversions.

Radio signals traveling through an inversion are bent or refracted toward earth. Some inversions do not refract vhf waves enough to bring them all the way back to the earth's surface. Consequently, only stations within the "normal" working range appear stronger. This is often referred to as *local enhancement* because DX contacts do not take place. Tropo DX occurs when an inversion is strong enough to refract vhf signals back to the earth's surface, and when it covers a wide geographical area. Both conditions must be met because tropo signals are not reflected in a single hop (or a series of long hops) at high altitude, as is true for ionospheric propagation. Rather, an inversion continuously refracts radio waves back toward the earth over paths that it covers completely.

The change in the index of refraction of the air within an inversion determines how much the inversion can refract radio waves. Although refractive index change measurements cannot be made by the average ham, they do provide a more precise way of describing inversion conditions. The calculation of the refractive index of air at radio frequencies, designated as N, involves only temperature, moisture and pressure.[5] Since atmospheric pressure always decreases with altitude, variations in temperature and moisture are the only factors that need be considered. Cool, moist air yields high values of N, and warm, dry air produces low values. Refraction is greatest when the index of refraction decreases sharply with altitude.

In a normal atmosphere, temperature and moisture both decrease steadily with altitude for several thousand feet. These changes cause the index of refraction to decrease only by about four N units per 100 meters as altitude increases. (The decrease in N units per 100 meters is noted as dN.) This small change in the index with altitude causes radio waves to be refracted slightly, but not enough to prevent them from going uselessly into space.

Vhf signals are refracted at greater angles as dN increases. When dN equals 16, radio waves are refracted enough to bring them back to the surface of the earth. This condition is known as *superrefraction,* and it occurs in especially strong inversions. Temperature rises and moisture content decreases rapidly in such inversions, causing the index of refraction to decrease sharply. When a superrefracting inversion covers a wide area, DX contacts may be possible.

Even though superrefraction conditions exist, all radio frequencies are not refracted to the same degree. The inversion depth (the vertical distance over which the temperature rises) determines the minimum frequency that is actually superrefracted.[6] Ultra-high frequencies are superrefracted with shallow inversions, but very-high frequencies require [illegible]

[1]Notes appear on page 15.

*810 Henderson St, Bloomington, IN 47401

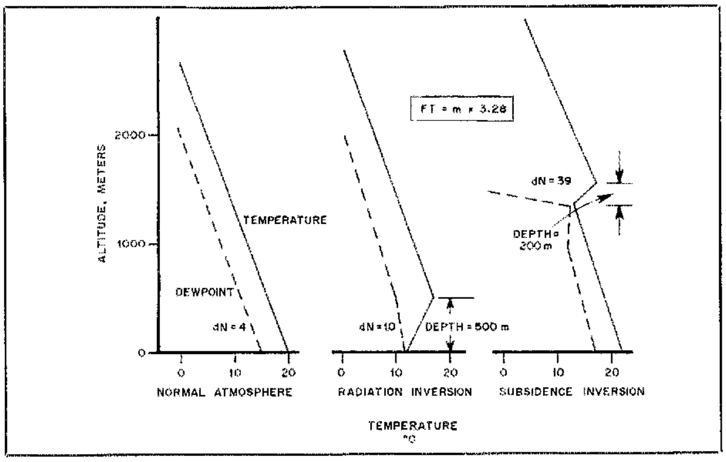

Fig. 1 — Upper atmosphere profiles of temperature and moisture content for normal conditions, a radiation inversion and a subsidence inversion. Moisture is measured by dew point. Note that only the subsidence inversion yields dN large enough for superrefraction. The minimum ducting frequency of this layer is about 100 MHz.

Table 1
Glossary of Terms

Dew point: The temperature to which a given sample of air must be cooled before water condenses; a measure of water vapor content.
DX: Distance over 500 miles.
Index of refraction of air (at radio frequencies): A measure of the potential of air, with given characteristics of temperature, moisture content and pressure, to refract radio waves.
Inversion depth: The vertical distance over which a temperature inversion exists.
Ionosphere: The atmosphere above 40 miles in altitude.
Jet stream: A narrow band of high-velocity winds that flow eastward between 30,000 and 50,000 feet (9000 to 15,000 meters) in altitude.
Local enhancement: An increase in strength of vhf signals within a 500-mile radius of the listener.
Polar front: The worldwide boundary between cold polar air and milder air of the temperate region.
Refraction (of radio waves): The slight change in direction of radio waves caused by their passage between contrasting media, such as layers of air with different temperatures and water content.
Subsidence: The sinking of air in the atmosphere.
Superrefraction (of radio waves): The bending of radio waves in the atmosphere at an angle great enough to bring them back to the surface of the earth.
Temperature inversion: An increase in air temperature with increasing altitude.
Troposphere: The atmosphere from ground level to about 33,000 feet (10,000 meters) in altitude.
Warm sector: The area to the south of the warm front in a wave cyclone.
Wave cyclone: A weather feature characterized by an area of counterclockwise winds that forms along a polar front.

minimum value of 16. When dN equals 16, an inversion 750 feet deep superrefracts 1296-MHz signals, but 144-MHz signals are not superrefracted. With the same inversion depth, 144-MHz signals are refracted back to earth only when dN is greater than 19. Since inversions grow in depth and strength, even over a period of a few hours, a tropo opening may be evident on 432 MHz or higher before it appears on 144 MHz. Monitoring uhf TV channels may provide a good early warning of the formation of inversion conditions on the vhf bands. Signals at 50 MHz are rarely affected by superrefraction because the required large value of dN and depth of inversion are rare.

A large superrefracting inversion is sometimes called a *duct* because it acts like a giant, natural waveguide. Only vhf signals above a certain frequency are trapped below the inversion layer, unable to escape into space because they are refracted back toward earth. Signals within the ducting area are sent long distances with little loss of strength. Unlike ionospheric skip, all stations within the ducting area normally hear each other.

Atmospheric Inversions

Three different weather processes create inversions over land. A *radiation inversion,* caused by the rapid cooling of the earth after sunset, is the most common type, but it is rarely useful for DX propagation. *Subsidence inversions,* which occur primarily in large high-pressure systems, provide strong and long-lasting DX openings. Inversions caused by the movement of air masses are created in a number of weather systems, including wave cyclones, warm fronts, cold fronts and certain wind systems. Wave cyclones often bring DX, but fronts generally do not produce true DX conditions.

Radiation Inversion

This is the most common type of temperature inversion. It often forms over dry land during clear, calm summer evenings and is the result of progressive nighttime cooling.[7] As the sun warms the ground after sunrise, the surface air is rewarmed from ground heating. The inversion is destroyed and any fog or dew present is evaporated.[8]

Radiation inversions occur during all seasons, but they are more common in summer. They are generally weak and rarely create widespread DX. This is because moisture content does not drop off fast enough to add much to the change in the refractive index, and the temperature rise by itself is rarely steep enough to cause a high dN. In addition, radiation inversions are usually not continuous over large areas. They are disrupted by local winds, large bodies of water, mountains and by other factors. Even though radiation rarely causes inversions strong enough for DX communications, it is a common cause of enhancement over paths less than 500 miles long during the evening and morning hours of clear, calm days.

High-Pressure Systems

Late summer and early fall high-pressure systems bring a majority of the strong, long-lasting vhf openings, although not all highs are accompanied by DX. An atmospheric high is a large mass of air whose pressure is relatively higher than the air around it. Steady, clockwise winds force air within the system downward and outward; skies are generally clear and winds are light under high-pressure conditions.

The air within a high is stable because it is sinking. This *subsidence* of air is a most important feature of a high because subsidence creates strong inversions.[9] Fig. 1 shows the temperature and moisture profile of a strong subsidence inversion with a base at 4400 feet.

Subsidence inversions become stronger, more stable and lower in the atmosphere at night. During sunlight hours, subsidence is countered by a strong rising of air caused by the heating of the earth's surface. Low, puffy cumulus clouds are often formed by the daytime rising air movement. As the sun goes down, the ground loses heat and the air stops rising. Subsidence takes over, strengthening the inversion and forcing it farther downward. The clouds are often quickly evaporated by the decending dry air in late afternoon. Since clear skies and calm air prevail after sunset under high-pressure conditions, radiation inversions often form. By morning, radiation and subsidence inversions may exist simultaneously.

A typical high-pressure opening occurred on the evening of September 9, 1979. Many vhf operators remember this as the third day of the spectacular four-day opening that coincided with the 1979 ARRL September VHF QSO Party.[10]

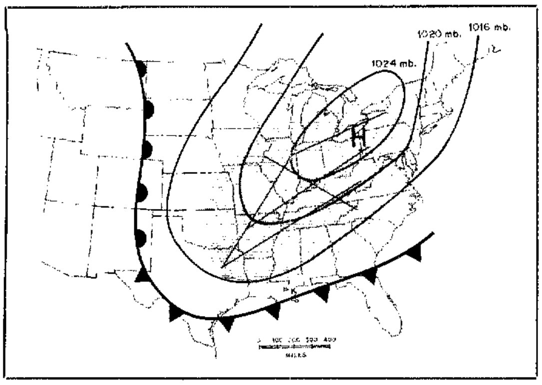

Fig. 2 — Surface map of a stagnating high-pressure system early on the evening of September 9, 1979. The center of the high is over Lake Erie. A few of the known 2-meter contacts are indicated by solid lines. The approximate limits of the area open to DX contacts is shaded. This area expanded to include southern New England by midnight.

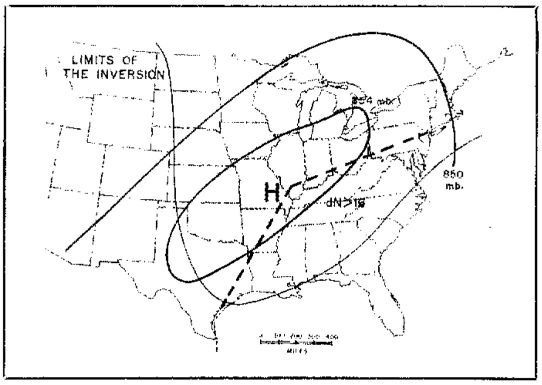

Fig. 3 — Map of the high-pressure system of September 9, 1979, at an altitude of about 4500 feet. The inversion boundary is indicated by a solid line. The useful area of the inversion (where dN is greater than 16) is shaded. Note that the center of the high at this altitude is located over Missouri. The broken line indicates the location of the cross section of upper-air profiles shown in Fig. 4.

Several features of this opening are not obvious from the surface weather map shown in Fig. 2. One might expect, for example, that the longest and strongest radio paths would lie directly across the center of the surface position of the high, where, presumably, subsidence and inversion formation were the greatest. This is almost never the case. The best paths appear consistently to the south or southwest of the center of this high and other highs. It is also unclear why the high has an elongated shape and what clues this might provide to help forecast propagation conditions.

Answers to these questions are found at about 4500 feet above sea level, where the most important inversion lies (see Fig. 3). The superrefracting part of the inversion (where dN is greater than 16) covers the area where 2-meter contacts were made. This area was southwest of the surface position of the high-pressure center. At 5000 feet, the center of the high lies near the midpoint of the inversion layer. The center shifts with altitude because the high leans to the southwest. This leaning is characteristic of "warm-core" high-pressure systems, in which the air is relatively warm. Warm-core highs rise to over 20,000 feet in altitude and lean toward the source of warmest air. Over the U.S., this generally means toward the southwest. The leaning makes the high look elongated on the surface map.

A more detailed analysis of the inversion further reveals that the inversion is bowl-shaped. The center is at about 3000 feet, while the extreme edges rise to nearly 6000 feet (see Fig. 4). The highest values of dN are found toward the center, while the lowest values are at the edges. The depth also decreases at the edges, further making the outer edges of the inversion useless for DX propagation.

Many high-pressure systems originate in the cold, dry air of northern Canada as *cold-core* highs, and slowly evolve into *warm-core* types.[11] Cold-core highs are less likely to bring enhanced propagation conditions.[12]

Winter highs that originate over the Gulf of Mexico and other warm regions sometimes bring very warm air into the Midwest and the East. These warm-air masses may create a strong inversion because cool air is held near the ground by the warm air aloft.[13]

Several high-pressure regions that exist more or less continuously over ocean areas adjacent to the continental U.S. create strong inversions over water. The Pacific high, which sprawls between Hawaii and the West Coast, frequently allows contacts over distances of 2500 miles between the islands and the California coast. The current 432-MHz terrestrial record was made by tropo over this path.[14]

The western side of the Bermuda high, which often lies south of Bermuda off the East Coast of the U.S., has the potential to support long-haul contacts over water as well. This was demonstrated in June 1981, when many Caribbean stations made contacts of up to 1800 miles into Florida and South Carolina.[15] The Caribbean, Bermuda and a large portion of the East Coast of the U.S. may be affected from time to time by a strong inversion associated with the Bermuda high.

Wave Cyclone

The wave cyclone is an entirely different weather system that occasionally brings springtime DX to the central part of the country. Wave cyclones, unlike stagnating warm-core highs, exist for only a few days from birth to collapse.[16] Fig. 5 shows a typical wave cyclone with a well-developed warm sector that covers the entire southeastern part of the country.

The local weather associated with wave cyclones is frequently violent. Severe thunderstorms along the cold front are often accompanied by high winds, hail and tornadoes. Low-level clouds and rain occur along the warm front and around the low. The warm sector often has scattered clouds and showers, especially close to both fronts. Wave cyclones are quite variable, however, and while the cold front almost always brings severe weather, the area within the warm section may remain calm and relatively dry.

On May 7, 1979, a typical springtime wave cyclone brought a strong opening from Iowa, Illinois and Indiana to Louisiana and Texas. Several 1000-mile contacts were made between the three Midwestern states and south Texas. The longest paths aligned generally north-south in the warm sector, but at some distance from the cold front. These are the most typical DX paths in such openings. In some wave cyclones, DX contacts occur along east-west paths in the southern part of the warm sector, such as between east Texas and Georgia. Typically, a wave cyclone is productive only for a single evening and the following morning.[17]

Wave cyclones appear in any season, but they are more common in springtime when the polar front moves northward. Wave cyclones that produce DX propagation appear as early as January along the Gulf Coast, but are most common farther north during May and June. The central part of the U.S., south of the Great Lakes and west of the Appalachians, is most often affected, although the East Coast may experience wave cyclone openings occasionally.[18]

Warm Fronts and Cold Fronts

Warm fronts and cold fronts are common weather features that create inversions. These inversions generally affect vhf propagation only for a short time in limited areas. Fronts rarely bring DX con-

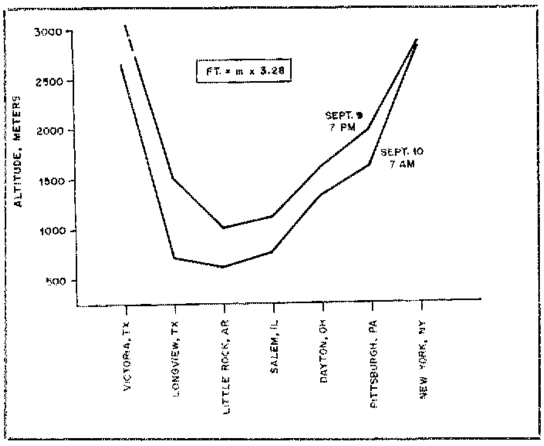

Fig. 4 — Height of the subsidence inversion base of September 9, 1979, at 7 P.M. EST, and at 7 A.M. the next morning. The bowl shape and the lowering of the inversion from early evening to morning are clearly evident.

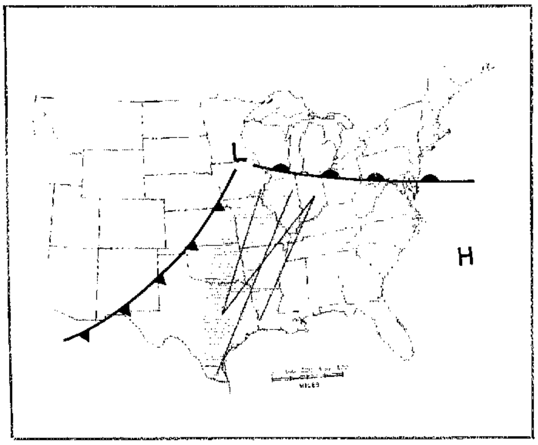

Fig. 5 — Surface map of a wave cyclone on the evening of May 7, 1979. The solid lines are a few known 2-meter paths, and the shaded area shows the limits of the area open to DX.

ditions.[19] Very long or slow-moving warm fronts may provide some exceptional DX conditions, but typically only enhanced local conditions exist for a few hours along and directly ahead of a warm front.

Cold fronts also improve local conditions for a short time, but they, too, rarely produce extended openings.[20] Any enhanced conditions take place immediately after the cold front passes and the atmosphere calms. The winds shift around to the northwest, temperature and humidity drop, and the skies clear. Vhf conditions behind the front may improve dramatically for a few hours. An unusual cold-front opening that produced a few hours of DX contacts between Oklahoma and western New York on the morning of June 29, 1980, is shown in Fig. 7.

Land Breeze

Good vhf conditions reported along the East Coast on clear summer evenings are often caused by land breezes, which develop after sunset. The land cools more quickly than the adjacent ocean. Air cooled over the land flows near the surface, toward the ocean, to replace the air rising over the warmer water. The warm ocean air, in turn, travels inland at altitudes of 600 to 1000 feet, to replace the cool air. The land-breeze circulation of cool air near the surface and warm air aloft may create a low-level inversion up to 50 miles inland and for hundreds of miles along the coast. Land-breeze inversions may last until morning, providing enhanced local conditions and, occasionally, some DX.

Other Weather Conditions

The weather systems described so far do not exhaust all the possibilities for producing useful inversions. Two interesting

Table 2
Symbol Table

Symbol	Meaning
	Warm front
	Cold front
	Stationary front
mb.	Isobars (with atmospheric pressure measured in millibars)
H	Center of high pressure
L	Center of low pressure
	Two-meter radio path
	Limit of temperature inversion
dN>16	Temperature inversion where the change in the index of refraction exceeds 16 "N" units per 100 meters change in altitude (shaded portion)

weather systems with inversion-producing winds have not been reported by vhf amateurs. They are the predictable Chinook winds of the Rocky Mountain foothills and the regular trade winds of the Atlantic Ocean.

Chinook winds occur in late winter and early spring along the eastern slopes of the Rocky Mountains and may produce strong inversions over the Great Plains. The wind originates when a large air mass is forced up the western side of the Rocky Mountains. It is cooled as it rises, and moisture condenses and falls out as rain. When the air mass reaches the crest of the range, it is cold and dry, but as it passes down the eastern slopes it is recompressed and warmed. The air that arrives on the plains below is warmer and drier than when it started. The wind temperature may be as high as 20° C with a relative humidity of less than 10%. This warm, dry Chinook wind overflows a plain that is often cold and still snow-covered, creating conditions ideal for inversion formation: warm, dry air overlaying cool moist air. Vhf contacts may be possible from the prairies of Saskatchewan to as far south as Oklahoma and perhaps east to the Missouri River.

Inversions are also associated with regions of trade winds in the Atlantic Ocean. These warm, dry winds create an inversion at 1500 to 6500 feet above sea level in two broad bands just north and just south of the equator between South America and West Africa. Contacts between Brazil and the coast of southwestern Africa, with Ascension Island at the center of the path, may be possible. In the northern trade-wind belt, useful inversions could potentially stretch from the Caribbean area to the coast of northern Africa. Although vhf activity is not high in these two areas, they may be reliable paths worth investigation.

Conclusion

The vast majority of tropo DX openings over the continental U.S. can be classified as either high-pressure or wave-

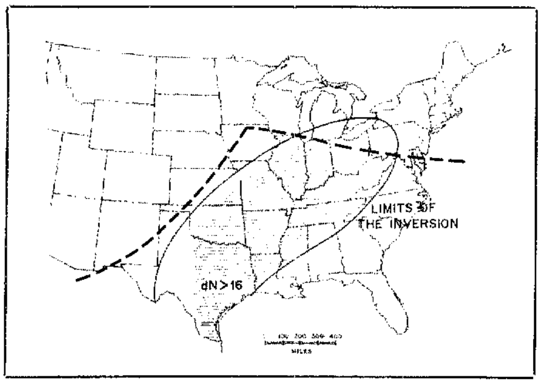

Fig. 6 — The wave cyclone of May 7, 1979, at an altitude of about 3000 feet. A broken line indicates the surface location of the fronts. Inversion limits are indicated by the solid line, and the area of the inversion where dN is greater than 16 is shaded.

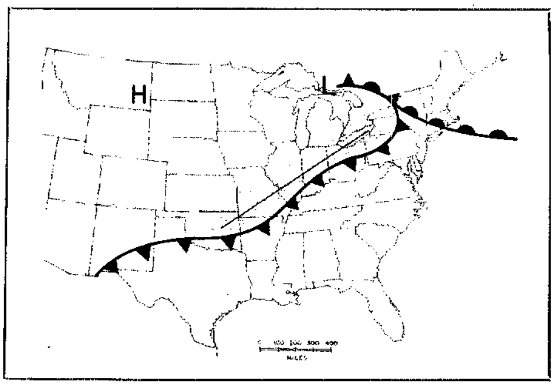

Fig. 7 — Surface map of a cold front of June 29, 1980, that brought a few hours of DX. Known 2-meter paths are indicated by the solid line. The shaded portion shows the presumed limits of DX propagation. Note that the cold front is part of a shallow wave cyclone.

cyclone types. The examples of these two types discussed here are among the best and clearest cases, but they are not unusual. The features of similar weather systems that bring DX openings may be distorted or partially formed. The physical principles are the same, however. Frontal regions and local winds can provide enhanced local conditions, but they are infrequent producers of DX propagation. No doubt there are still other weather configurations that may give rise to inversion formation.

The practical vhfer may wish to make personal forecasts of inversions using TV or newspaper weather maps. "A.M. Weather," which is on public television for 15 minutes each weekday morning, provides us with much useful information. Professional weather forecasts, such as those used by pilots, could be even more helpful, if they are available. Monitoring uhf TV channels is a good way to check on actual conditions. Still, there is nothing to replace getting on the air and making a few calls in likely directions. You never know who may be trying the same thing 500 or 1000 miles away.

Acknowledgments

I am grateful to everyone who provided copies of their logs and informal reports of band conditions over the past few years. These reports of actual radio conditions were an invaluable part of my propagation studies. Analysis of the high-pressure system was aided by an extensive description of the September 7-10, 1979 opening compiled by Curt Roseman, K9AKS, Michael Owen, W9IP and myself. I also appreciated the opportunity to present an early version of this material at the July 1981 meeting of the Central States VHF Society. Curt Roseman and Ann Higginbotham read rough drafts of this article, and their comments were most helpful. Special thanks to Curt Roseman for all his encouragement, and for his valuable comments and suggestions.

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Notes

[1] R. Hull, "Air-mass Conditions and the Bending of Ultra-High Frequency Waves," *QST*, June 1935, pp. 13-18; W. F. Hoisington, "Painless Prediction of Two-Meter Openings," *QST*, Oct. 1949, pp. 22-26; J. Collier, "Upper-Air Conditions for Two-Meter DX," *QST*, Sept. 1955, pp. 16-18. Also, see the *QST* column The World Above 50 MHz.

[2] The World Above 50 MHz, *QST*, July 1982, p. 68.

[3] km = statute mi × 1.609, nautical mi × 1.852; m = ft × 0.3048.

[4] Only paths longer than 500 miles are considered true DX here, since that is the farthest distance the best-equipped vhf stations can cover under ordinary circumstances.

[5] A detailed discussion of the radio refractive index may be found in D. Evans and G. Jessop, *VHF/UHF Manual* (London: Radio Society of Great Britain, 1980), pp. 2.2-2.8. More technical aspects are in B. Bean and E. Dutton, *Radio Meteorology* (New York: Dover, 1968).

[6] The role of inversion depth is presented in B. Bean and E. Dutton, *Radio Meteorology* (New York: Dover, 1968).

[7] As the sun goes down, the land cools by radiating heat. In turn, the air directly above the surface is cooled. Higher in the atmosphere, the air remains relatively warm, creating an inversion. The cooling process continues throughout the evening and pre-dawn hours. As time passes, the inversion deepens, reaching heights of 500 meters in extreme cases. Sometimes the air will cool enough for fog or dew to form, especially in low-lying areas.

[8] Several conditions retard the information of radiation inversions. Wet ground does not radiate heat readily, and high humidity hinders radiation. Cloud cover also prevents heat from escaping, thus insulating the lower levels of the atmosphere and preventing inversion formation. Even under clear skies, wind destroys inversions by mixing the cool and warm layers of air.

[9] As the air is forced downward, it is compressed, heated and dried. The creation of layers of warm, dry air at various altitudes results in inversions. Three or more inversions may form within a well-developed high, but only the lowest inversion is important to vhf propagation. This inversion is most useful when it lies between 500 and 2000 meters because, at this altitude, its depth and dN are generally greatest.

[10] Contesters had made as many as 20 QSOs over 1000-mile paths on 144 MHz between north Texas, Oklahoma and Arkansas to New England and the mid-Atlantic states by the end of that evening. Contacts in the 500- to 1000-mile range were common; many were made by stations using only a few watts. WB4LHD/5 in northwestern Arkansas completed over half a dozen 1000-mile QSOs, the farthest being with N6NB/1 in Vermont — over 1200 miles. The most distant contact that evening was probably 1275 miles, made by WD5CRK in eastern Oklahoma and N6NB/1. Although contacts on 220 MHz and 432 MHz were less numerous, similar paths were covered on these frequencies.

[11] Canadian cold-core highs generally move southeastward, and are heated from below during the day as they travel over the warmer surface. They grow larger and more sluggish, and their speed of travel may decrease to less than 500 miles a day.

Other factors affect the progress of large highs. During the late summer and early fall, the high-altitude wind of the jet stream may not provide its usual strong eastward push because it loops to the north of the high. Then, the high-pressure system slows still further.

Late-summer and early-fall highs frequently bog down over the eastern U.S. because they are blocked by the even larger semipermanent Bermuda high that lies off the Atlantic coast during this season. They may remain nearly stationary for several days or more. The air-mass temperature rises when it stagnates, and subsidence strengthens, resulting in a very strong inversion. The strongest and most stable inversions are produced in these stagnating high-pressure systems. Spectacular DX conditions may last several consecutive evenings as a result.

Although the most productive stagnating warm-core highs occur in late summer, high-pressure systems occur during all seasons. Most of these do not create DX openings.

[12] Unlike warm-core highs, cold-core highs are shallow weather features that do not rise above 3000 meters into the atmosphere. Subsidence is not as pronounced in them, and inversions that do form are weak. Cold-core highs that move too quickly out of Canada are not transformed into large warm-core types. When the surface over which they are moving is cool, as in winter or spring, transformation is even more difficult. Even in summer, highs that move more than 300 or 400 miles a day do not reach DX-producing capabilities before they slide off the continent. In general, fast-moving highs, especially in the cool months, will not produce good DX openings.

[13] Subsidence is not the primary cause of such warm-air-mass inversions, but it helps stabilize the atmosphere. If the surface is snow-covered or wet from recent rains, the additional moisture adds to the change in the refractive index. A superrefracting inversion often results. Such conditions are usually accompanied by dense day-long fog and low clouds. Winter warm-air-mass openings may be quite strong in local areas, and last all day.

[14] The World Above 50 MHz, *QST*, Aug. 1981, p. 72.

[15] The World Above 50 MHz, *QST*, July 1982, p. 68.

[16] Wave cyclones appear along the global boundary *between polar and temperate air regions called the polar front*. This extremely long polar front is unstable, and "waves" appear as low-pressure centers where the jet stream crosses it from the

southwest. The counterclockwise winds created by the low pressure pull cool air around from the northwest, creating a cold front southwest of the low center. Breezes associated with a region of high pressure off to the southeast bring mild Gulf air northward, creating a warm front to the east of the low. The area south of the warm front is known as the warm sector.

As the wave cyclone develops, the cold front is pushed eastward and catches up to the slower moving warm front. In the last stage of the wave, the two fronts meet, the warm sector is closed off and the wave ceases to exist. While the wave cyclone is going through this cycle, it also moves northeasterly. A wave that begins in southern Colorado may collapse over New York state. The whole process takes about three days. Wave cyclones form "family groups," and one wave may form after another along the polar front. The polar front, with its family of wave cyclones, may stretch across North America, the Atlantic and Europe.

In contast to the subsidence inversion of the high-pressure system, wave cyclone inversions are created primarily by strong winds. Mild, moist air flows steadily north near the surface within the warm sector. Above the surface flow, at 4500 feet and higher, very-fast-moving, warm, dry air associated with the jet stream crosses the warm sector from the southwest. The intersection of these two winds, mild and moist below, very warm and dry above, creates a strong and relatively flat inversion at altitudes of about 3000 feet (see Fig. 6). Conditions generally improve after sunset when ground heating and winds decrease. Weak subsidence may also be present if the warm sector stabilizes the inversion layer.

Not all wave cyclones create inversions useful for vhf propagation. The reasons are not clear in every case, but several features seem to be important. Productive wave cyclones are usually well developed with large and distinct warm sectors. They appear to have fewer clouds and less rain in the warm sector, although clouds and even rain do not prevent good vhf conditions from existing. Midwestern stations in the northern part of the warm sector have worked Texas stations while it was raining and while thunderstorms were within a distance of 50 miles.

Wave cyclones must be carefully distinguished from other kinds of cyclonic weather. Wave cyclones form only along the very long polar front; the fronts associated with them are continental in scale, stretching 1000 miles or more. Other low-pressure regions may have fronts associated with them, but these fronts are invariably short. Since wave cyclones are a part of a global weather feature, they are sometimes difficult to recognize on a map as small as that of the U.S. alone. Even so, their formation and development can often be followed in daily weather reports.

A warm front is the advancing edge of a warm-air mass that is moving toward more stationary cooler air. The less dense warm air overrides the cooler air and forms an inversion *ahead* of the surface boundary between the air masses. The long wedge of overriding warm air may extend for 250 miles ahead of the front.

Warm-front inversions are weak. The difference in temperature and moisture between the warm air aloft and the cool air below is usually not enough to create superrefracting conditions. Low-level clouds accompanying the front often produce rain that disrupts and obscures the inversion *boundary*. Warm fronts are also generally too short to create inversion conditions over long paths.

A cold front is the advancing edge of a cold-air mass that is moving toward more stationary warmer air. Since the approaching cool air is denser than the warm air it confronts, it wedges under the warm air, forcing it aloft in a steep grade. An inversion may appear in a narrow band no more than 75 miles wide *behind* the front.

The sharp differences in temperature and moisture between the two air masses in a typical cold front suggest that the inversions might be strong, but usually this is not the case. The frontal boundary is quite unstable. Cold fronts give rise to towering thunderstorms and high winds. These small but violent storms mix the air along the front and destroy the stable conditions needed for an inversion. QST

Strays

QEX: THE EXPERIMENTERS' EXCHANGE

Wonder what you've been missing by not subscribing to *QEX*, the ARRL newsletter for experimenters? Among the features in the April/May issue were:

- "Diode Voltmeters," by Albert E. Weller, Jr., WD8KBW
- "Right on Channel — A Frequency Standard for HF RTTY," by Paul Newland, AD7I
- "Novation D-CAT Modem for Amateur Radio," by Robert Gervenack, W7FEN and Russell Faudree, KA7HVA
- "VHF+ Technology," by Geoff Krauss, WA2GFP
- "Data Communications," by David Borden, K8MMO

Issue 15 of *QEX* is for both April and May. That is to allow the editor, Paul Rinaldo, W4RI, to move from McLean, Virginia to Newington, Connecticut, in early May to take up his new position as manager of the ARRL Hq. Technical Department. Beginning with the June issue, *QEX* will be prepared in Newington.

Please send all subscription requests, manuscripts and correspondence for *QEX* to ARRL Hq. The domestic subscription rate for ARRL members is $6 for 12 issues; $12 for nonmembers. There are additional postage charges for mailing outside the U.S.; write to Hq. for details.

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When writing, please observe the following guidelines so we may provide the best possible service to the greatest number.

1) Before writing for technical assistance, search your files of *QST* and other ARRL publications. The answer you need may be there, available immediately. Consult the annual index of articles in each December issue.

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Do not ask for information on articles published in other magazines. Write to the editor or author of that article.

Do not request custom designs for amateur gear.

Do not ask advice on nonamateur matters. We cannot respond to questions about CB, marine radio, hi-fi, etc. (unless they concern interference caused by amateur gear).

3) Use a typewriter when possible; otherwise, write or print *clearly*. Please be reasonable in the number of questions you ask; try to limit your questions to three per letter.

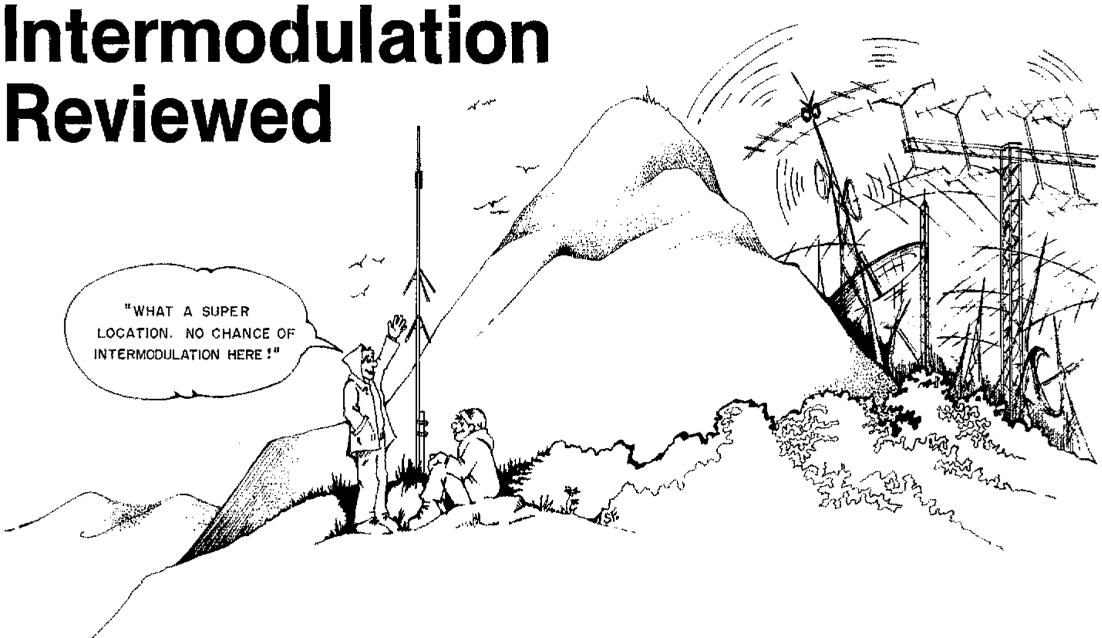
4) When writing, please come right to the point, and be sure to share with us whatever experience you have had with the problem in question. This will avoid our reply covering ground you've already been over.

5) Address all technical questions to Technical Information Service, American Radio Relay League, 225 Main St., Newington, CT 06111. — *Bob Schetgen, KU7G, Technical Information Specialist*

Next Month in QST

If you've ever thought about delving into slow-scan TV, you'll want to read an overview of the subject in June *QST*. Other features in that issue include a report on the April Board of Directors meeting, a discussion of how amateurs can get involved with SETI (the search for extraterrestrial intelligence), and the results of the 1982 Simulated Emergency Test. Whatever your interests, you have something to look forward to in June *QST*.

Intermodulation Reviewed



Hearing unwanted signals on your favorite repeater? Intermodulation may be the culprit!

By David W. Potter,* W2GZD

Intermodulation (IM) is defined as the undesired mixing of two or more frequencies in a nonlinear device, which produces additional sum-and-difference frequencies. Problems with IM seem prevalent in the vhf and uhf bands, because amateur repeaters tend to cluster around hills and mountain-tops — in close proximity to commercial and government vhf and uhf radio services. The intermodulation problem has been understood for years, but a review of the subject can be helpful.

Types of Intermodulation

The simplest kind of intermodulation is the mixing of two frequencies. The equation for *second-order* IM is

$$f_{IM} = f_1 \pm f_2 \qquad \text{(Eq. 1)}$$

where f_{IM} is the frequency of the IM product, and f_x are the mixing frequencies.

Intermodulation products are the sum and difference of the two mixing frequencies. These are similar to the familiar products deliberately generated by the mixing process used in superheterodyne receivers. Note that if f_1 and f_2 are frequencies within an amateur band f_{IM} must be an out-of-band signal.

Third-order IM products can produce both in-band and out-of-band signals when f_1, f_2 and f_3 are all in-band frequencies as given by the equation

$$f_{IM} = f_1 \pm f_2 \pm f_3 \qquad \text{(Eq. 2)}$$

In-band signals are generated by the sum of any two frequencies minus the frequency of the third signal. The out-of-band signal is the sum of all three frequencies. A special case of Eq. 2 is

$$f_{IM} = f_1 + f_1 - f_2 = 2f_1 - f_2 \qquad \text{(Eq. 2A)}$$

Here, the second harmonic of an in-band signal can beat with a fundamental frequency to produce another in-band signal.

Fifth-order IM products are given by

$$f_{IM} = f_1 \pm f_2 \pm f_3 \pm f_4 \pm f_5 \qquad \text{(Eq. 3)}$$

with special cases:

$$f_{IM} = 3f_1 \pm 2f_2 \qquad \text{(Eq. 3A)}$$

$$f_{IM} = 3f_1 \pm f_2 \pm f_3 \qquad \text{(Eq. 3B)}$$

$$f_{IM} = 2f_1 \pm f_2 \pm f_3 \pm f_4 \qquad \text{(Eq. 3C)}$$

$$f_{IM} = 2f_1 \pm 2f_2 \pm f_3 \qquad \text{(Eq. 3D)}$$

Some fifth-order products are in-band and others are out-of-band. Notice that second- and third-harmonic signals may be involved. Odd orders of in-band mixing frequencies produce some in-band products, but when out-of-band mixing frequencies are involved, even-order products may fall in band!

The above equations are valid for steady carriers, and it is easier to understand the concept of intermodulation by using them. Most of the signals we deal with are not steady carriers, however, but modulated ones.

The bandwidth of an IM product may be wider than the bandwidths of the individual signals. This is because the instantaneous frequency of the product is the algebraic sum of the instantaneous frequencies of the mixing signals. For fm, it would be equivalent to adding three voice signals together in a wide-band fm transmitter. Assume that the audio amplitude is limited on *each* signal to produce a deviation no greater than 5 kHz. When the three signals are added together, they could produce an IM product having much greater deviation than any of the individual signals.

Intermodulation may involve any number of frequencies, but let's concen-

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trate on the more common third-order types. I will show some examples to make these abstract concepts more meaningful.

Field Examples of Intermodulation

Consider two repeaters that are physically located close to each other. One repeater transmits on 146.70 MHz, with a 146.10-MHz input, and the second has a 145.31-MHz output and a 144.71-MHz input. Assume that the .31 repeater output causes the .70 repeater first receiver stage to be driven into nonlinear operation (overload). This means that mixing of all frequencies seen by the first stage will occur.

Case 1: The .70 repeater is off, but the .31 machine is operating and a local 145.50-MHz simplex signal is present. The .31 repeater input and output signals and the simplex signal mix to produce an IM product on 146.10 MHz: 145.31 − 144.71 + 145.50 = 146.10 MHz. This signal can key up the .70 machine, which will then repeat both the .31 machine *and* the simplex conversation. We will disregard the other IM products that are generated.

Case 2: The simplex station is off the air. The .31 machine is repeating, and the input to the .70 repeater drops, but the output is still up. An IM signal is produced: 146.70 − 145.31 + 144.71 = 146.10 MHz. The .70 machine will now repeat the signal from the .31 repeater until someone overrides the IM signal — provided that the .70 repeater doesn't time out beforehand. If the system gain is adequate, the repeater could feed back on itself with a resulting characteristic audio howl.

Case 3: The .31 repeater shuts down. The input to the .70 machine drops, but the output is up and *stays* up. You may hear another signal on the output, or the repeater may break into oscillation. Why? You find a strong signal at 147.30 MHz that is overloading the repeater receiver, causing it to be nonlinear. The strong signal mixes with the second harmonic of the repeater: 2(146.70) − 147.30 = 293.40 − 147.30 = 146.10 MHz. This signal falls on the .70 repeater input.

These cases use the popular 2-meter band for illustration. Keep in mind that intermodulation can occur on *any* band, and can entail an endless combination of frequencies.

Receiver as the Culprit

Usually, the first stage of a receiver is the one most likely to overload, causing susceptibility to intermodulation. In some cases, however, later stages may be at fault. Receiver front-end nonlinearity occurs at relatively low input voltage levels, so the signal power involved is small. Therefore, the IM products generated there are also low in amplitude. These signals are fed to the receiving antenna and are radiated. If you track down these weak IM signals, they will lead you back to the receiver site!

There is a big temptation to use a preamplifier to increase the sensitivity of a receiver or the range of a repeater. The use of these devices in repeater service is strongly discouraged. Preamplifiers may not have the dynamic range that early stages of communications receivers have, and they usually lack the front-end selectivity found on well-designed receivers. Therefore, they are red-hot candidates for intermodulation. Preamplifiers may generate signals on the input frequency of the repeater, and users have to override these signals in order to be heard. The use of a preamplifier can sometimes *degrade* system performance!

Reducing Receiver IM Susceptibility

It is important to realize that it takes only *one* signal, located anywhere in the spectrum, to drive a circuit into nonlinearity, which could produce IM products when one or more other signals are present. In order to minimize intermodulation, your receiver circuitry should provide great attenuation to all frequencies except the band of interest. The receiver dynamic range should also be as large as possible. For this idealized case, only a huge signal in the passband could possibly cause overload and nonlinear operation. Any resulting IM signals falling outside of the receiver passband would be severely attenuated.

Transmitter as the Culprit

Serious intermodulation problems can be generated by transmitters when other strong signals present on the antenna (and coupled to the final amplifier) mix with the fundamental and its harmonics. Here the voltage and power levels are much greater than those associated with receiver circuits, so relatively strong IM signals may be coupled to the transmitting antenna and be radiated. The transmitter final amplifier is essentially an rf switch, and unless it is operating Class A (which is uncommon) it will be a nonlinear stage. Class C operation is more likely to cause intermodulation than Class AB1 — because it is more nonlinear. Wide-band, solid-state amplifiers with low-Q circuits tend to be more susceptible to intermodulation than are narrow-band, high-Q configurations.

Reducing Transmitter IM Susceptibility

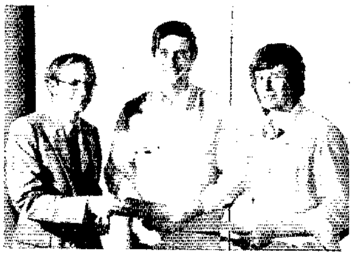
Fm transmitter final amplifiers are quite nonlinear, and intermodulation can occur if other signals mix in this stage. The Q of most final stages is not high, even for tuned final amplifiers, so the resulting bandwidth is wide. Increasing circuit Q helps the IM problem, but it may be undesirable for other reasons. Eliminating intermodulation in transmitters is, therefore, more difficult than in receivers. The use of a *circulator* is effective because it presents a very low impedance to signals going from the transmitter to the antenna, but high attenuation to signals going from the antenna to the transmitter. Since the circulator must carry full transmitter power, it is an expensive cure for intermodulation. When duplexers are used at repeater sites, a degree of attenuation is introduced for out-of-band signals, but this alone may not be sufficient if other transmitters are nearby.

Conclusions

In-band intermodulation products will degrade or destroy your station performance, in addition to interfering with other amateur communications. Out-of-band IM products may play havoc with non-Amateur Radio services. Furthermore, transmitting spurious signals, such as IM products, is illegal. Conscientious operators are knowledgeable about intermodulation and ensure that their stations are free from it.

David W. Potter, W2GZD, is a spokesman for the Long Island GAME Association (Group Against Malicious Emissions). If it sounds like a hunting club, you're right! The group is engaged in tracking down interference for local repeater clubs. David's inspiration to write this article came about as a result of his involvement with the group. Persons interested in working with the GAME Association are invited to contact him.

Strays



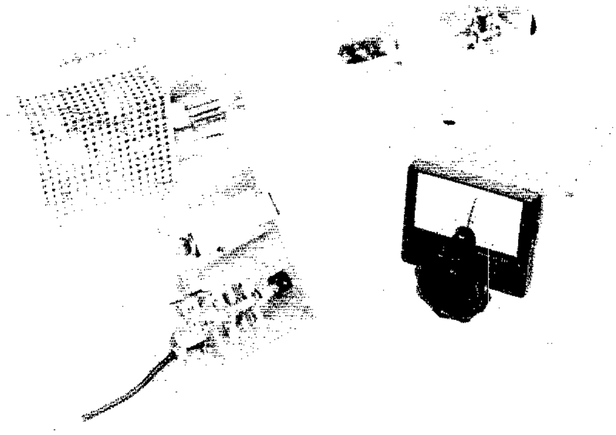
At the 1982 New England Division Convention, ARRL General Manager K1ZZ (left) had the pleasure of presenting awards to the top-scoring W/VE entrants on both modes in the 1982 ARRL DX Contest: K1ZM, operating W1ZM on phone, and K1GQ, operating his own station on cw. Both are members of the Yankee Clipper Contest Club. *(W1VRK photo)*

KERCHUNK!

☐ Advice from a New York photographer to ladies on how to compose their mouth to gain "some advantage to their appearance" when sitting for a picture: ". . . If she wishes to look mournful, she must say 'Kerchunk'. . ." — *Quincy* (Massachusetts) *Patriot*, Dec. 3, 1887 as reprinted in *The New Yorker; tnx W9IWI*

A Minimum 2-Meter Satellite Transmitter

Is OSCAR operation beyond the reach of the budget-minded amateur? Not necessarily! W6IOJ shows you how to do it with a "junkbox special" transmitter.



By John Reed,* W6IOJ

This article describes a simple transmitter for the 2-meter satellite band. It features variable-frequency crystal control (VXO) and up to 40 W of power output. Methods for measuring power, using simple homemade equipment, are also included. The circuit is relatively simple and can be easily duplicated. Components for the projects were chosen on the basis of their availability.

Circuit Details

VXO and Multiplier Chain

A VXO offers crystal stability, high output frequency and circuit simplicity — an ideal signal source for a vhf transmitter. The restricted frequency coverage is not a hindrance in this case, as the 2-meter satellite passband is only 150 kHz wide. My prototype circuit provides a tuning range of up to 300 kHz while retaining excellent stability. Four crystals were tested in the VXO circuit — three overtone units calibrated in the 48-MHz region and one fundamental-type 16-MHz crystal.[1] The internal capacitance of the overtone crystals is somewhat lower than that of the fundamental unit — a characteristic that could influence the tuning range. However, detailed comparative tests indicated no significant differences between the two types. One disadvantage of the overtone crystals is that with 25-pF loading the oscillator frequency will be lower than the marked frequency. The 16-MHz crystal did not exhibit this trait. Regardless of the crystal type, the VXO should operate near the center frequency of 16.230 MHz.

Several versions of the VXO tank circuit were tested to see which gave the maximum tuning range and the best stability. The values shown on the schematic diagram represent the best compromise. Long-term stability will be improved by using silver-mica capacitors rather than disc ceramics in the gate-source capacitive divider. However, Doppler shift during satellite operation far exceeds any instability condition exhibited by the oscillator.

The schematic diagram of the VXO/multiplier circuit is shown in Fig. 1. The VXO uses a 3N211 dual-gate MOSFET in a Colpitts circuit, with the drain tuned to the third harmonic of the crystal. A bipolar-transistor stage triples this 48.6-MHz signal and provides approximately 10 mW of 146-MHz energy to the driver stage. Components are mounted on a 5- × 6-inch double-sided pc board, with the component side unetched and used as a ground plane.[2] The FET drain circuit and the multiplier input circuit are mounted on the foil side of the board to provide maximum isolation.

The crystal frequency is "pulled" by the series circuit, consisting of L1 and C1, with an additional capacitor (C2) used to bring the tuning range into the satellite band. Under certain conditions the crystal would not oscillate at "power up." The combination of D2, D3 and C3 corrects this problem by shorting the reactance network to ground for a few milliseconds. This starts the oscillator as if it were a conventional Colpitts crystal type. Once the circuit is oscillating, the load capacitor charges and offers a high impedance to the reactance network. The completed VXO assembly is shown in Fig. 2.

Driver/Final Amplifier

The schematic diagram of the driver and final amplifier is shown in Fig. 3. The circuit mounts on a 4- × 4-1/2 inch, double-sided pc board, with the component side used as a ground plane. The collector circuit of the predriver stage is mounted on the foil side for isolation. VXO output is coupled directly to the base of the predriver through 20 feet of RG-58/U coaxial cable. A high-Q tuned circuit (which discriminates against spurious signals) follows the predriver stage.

A series-tuned circuit is used to couple

[1]Notes appear on page 22.

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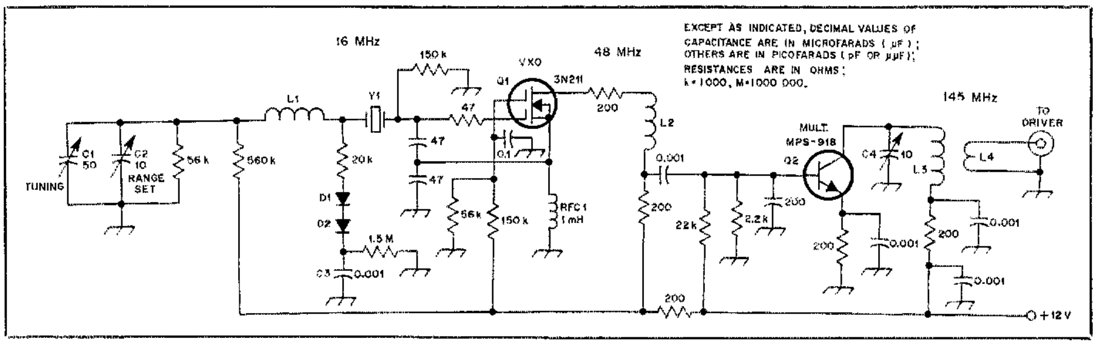

Fig. 1 — Schematic diagram of the VXO/multiplier circuit. Numbered components are identified below. Fixed-value resistors are carbon composition, 1/4 W unless otherwise specified. Capacitors are disc ceramic, except those with polarity marked, which are electrolytic. Part numbers in parentheses are from Radio Shack.

C1 — 7-50 pF variable, Hammarlund MC50-M or equiv.
C2, C4 — 1.5-10 pF piston trimmer.
D1, D2 — 1N914 silicon switching diode (276-1124).
L1 — 35 turns of no. 32 enameled wire on a 3/8-inch-dia. slug-tuned form.
L2 — 7 turns of no. 32 enameled wire on a 3/4-inch-dia. slug-tuned form.
L3 — 9 turns of no. 14 solid wire wound over a 1/4-inch-dia. rod.
L4 — 2 turns of no. 24 plastic-insulated wire wound over the end of L3.
Q1 — 3N211 dual-gate MOSFET (276-2045).
Q2 — MPS-918 silicon npn transistor (276-2011).
Y1 — 16.230-MHz quartz crystal. See note 1.

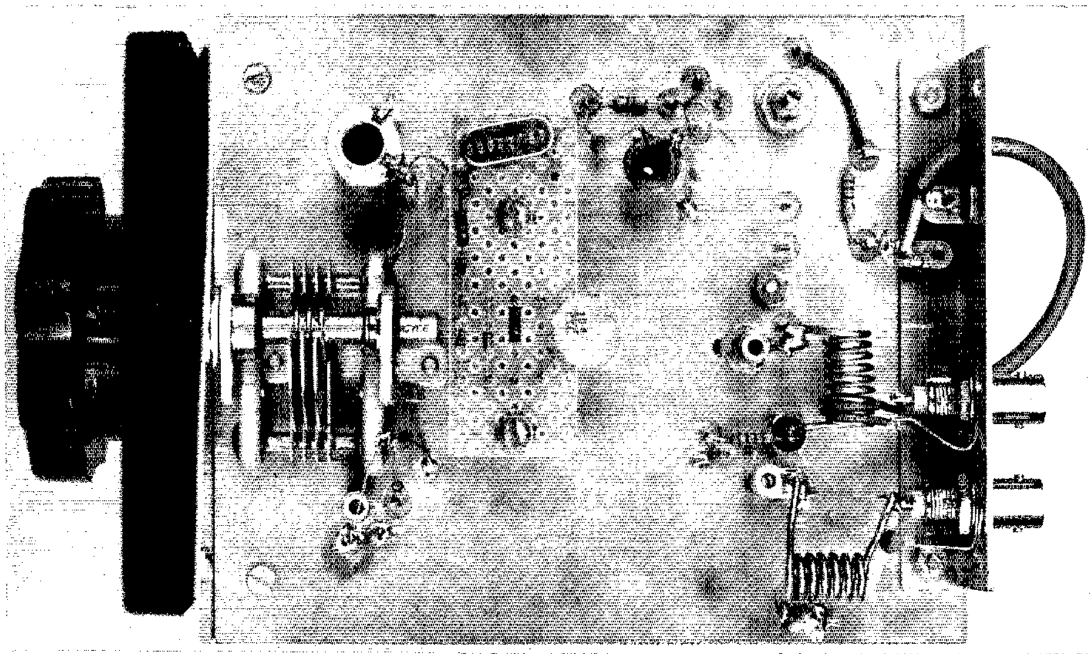

Fig. 2 — Interior view of the completed VXO assembly. Several of the components are mounted on the opposite side of the board.

energy from the driver stage to the 6146 final amplifier. This arrangement provides the best compromise between drive capability and stability. The driver transistor, Q4, must be properly heat-sinked, as it operates at 4-W input. In some of the Radio Shack 276-2038 devices, the case is tied to the collector. In later units, however, the case is insulated and can be clamped to the chassis. In the prototype transmitter, more than adequate heat transfer was achieved with a 1-1/2 × 1-1/2 × 1/16-inch aluminum plate fastened to the printed-circuit board.

Both the chassis and the shields are made of aluminum purchased from a hardware store. Perforated aluminum is used for the shields to permit air circulation around the 6146 envelope. The driver and final amplifier are mounted on a 20 × 5 × 1-3/4 inch "U" chassis. Support for the 6146 is provided by a vertical bracket mounted 5-1/4 inches from one end of the chassis. Next to the vertical bracket is a 4- × 3-1/2 inch cutout for mounting the driver circuit board. Two shields are used in the assembly: One houses the grid circuit/driver; the other covers the plate circuit. Fig. 4 shows the physical layout.

The 1/2-λ plate line is tuned by a 50-pF variable capacitor, which also serves as a mechanical support for one end of the line. Final-amplifier B+ is fed at the rf voltage null on the strip line, approximately 6-3/4 inches from the plate connection. Output power is link-coupled by means of a loop near the rf current node, with coupling varied by a mechanical ar-

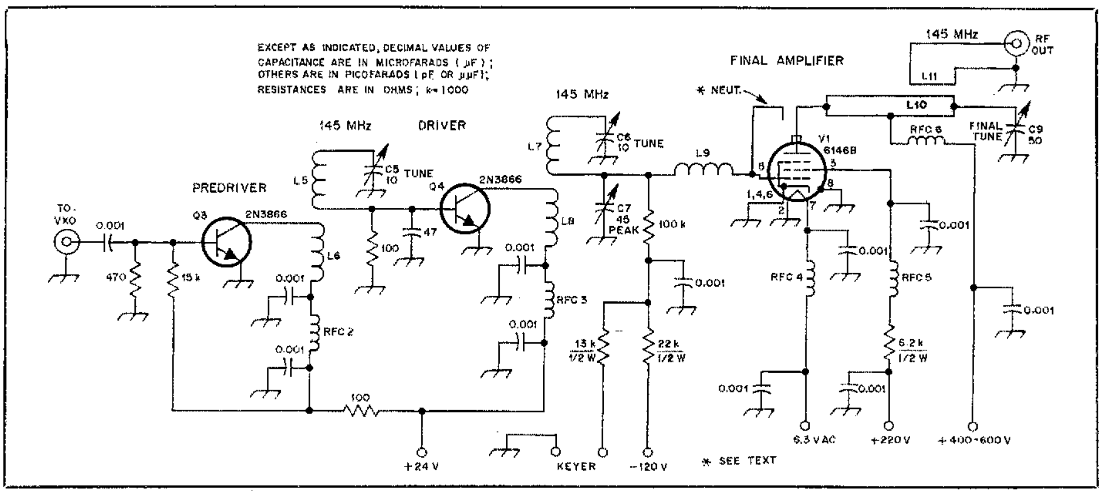

Fig. 3 — Schematic diagram of the driver and final amplifier. Numbered components are identified below. Fixed-value resistors are carbon composition, 1/4 W unless otherwise specified. Capacitors are disc ceramic unless otherwise specified. Part numbers in parentheses are from Radio Shack.

C5, 6 — 1-10 pF piston trimmer.
L5 — 9 turns of no. 14 solid wire wound on a 1/4-inch-dia. rod.
L6 — 2 turns of no. 24 plastic-insulated wire wound over L5.
L7 — 7 turns of no. 14 solid wire wound on a 1/4-inch-dia. rod.
L9 — 3 turns of no. 24 plastic-insulated wire wound over L8.
L10 — Double-sided printed-circuit board, 10 inches × 7/16 inch, spaced 3/4 inch above chassis.
L11 — 2-3/4 inch × 7/16-inch loop made of 5/16-inch copper strip. Mount 5 inches from plate connection on L10 (see Fig. 4).
Q3, 4 — 2N3866 silicon npn rf-power transistor (276-2038).
RFC 2, 3, 4, 5, 6 — 10 turns of no. 32 enamel-coated wire over a 47-kΩ, 1-W resistor.

rangement that permits rotation of the loop approximately 90°. The dimensions provide a 50-ohm output impedance with the loop near the strip-line axis.

One might ask, "Why use a long, 1/2-λ strip line, considering the loaded Q is probably not much greater than a conventional L-C circuit?" The strip-line method was chosen because of the simplicity and predictable performance. A 2-meter L-C tank, loaded by a 6146, turns out to be a hair pin — the parallel tuning and bypass capacitance values become critical. In a strip-line circuit, the tuning capacitor selection is less restrictive: There are no critical bypass capacitors, and there are fewer restrictions on the output coupling arrangement.

Although the 6146B is a rugged and forgiving device (particularly when compared to some solid-state amplifiers), lack of plate voltage will cause an abnormal increase in screen current — exceeding the rated screen dissipation. In this design, a dropping resistor provides screen voltage from the plate supply, which causes an opposite effect: Losing drive causes an abnormal increase in plate current.

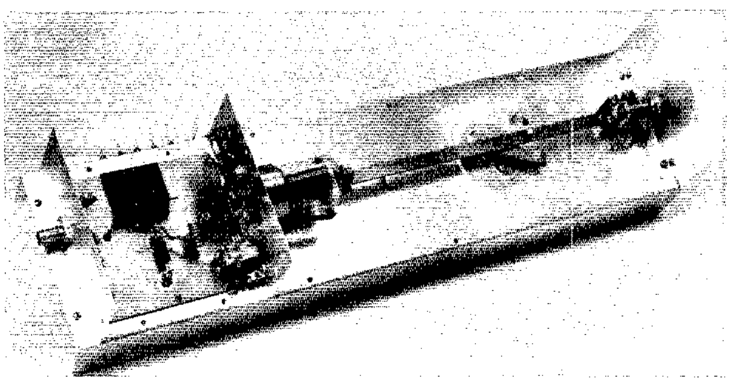

Fig. 4 — View of the driver and power amplifier stages of the transmitter. Power is fed to the unit through the 0.001-μF feedthrough capacitors, which can be seen in the upper left-hand corner of the picture.

Tune-up and Operation

VXO

After checking the circuit wiring, apply power to the VXO and short the junction of Y1 and L1 to ground. This allows the oscillator to be calibrated at the natural crystal frequency. Trim the drain circuit for proper operation at 48.690 MHz — the third harmonic of the crystal. With C1, C2 and L2 adjusted for their minimum respective capacitance and inductance, the VXO fundamental frequency should not change more than 2 kHz when the short is removed. If the frequency drops *lower* than 2 kHz, then L1 has too much inductance. C1 should have a tuning range of approximately 100 kHz (at 48 MHz) when the proper value for L1 is found. C2 can then be adjusted to center the tuning range in the satellite band. Use a receiver, frequency counter or other frequency-measuring device to confirm VXO output on 146-MHz. When the VXO is operating properly, connect it to the driver/final amplifier stage through a length of RG-58/U coaxial cable.

Driver/Final

In the present configuration this amplifier is stable, even without shields. However, a word of caution! Instability during the initial adjustments can destroy the driver transistor. It is recommended that initial adjustments be made with no more than 100/50 V on the 6146 plate and screen, respectively.

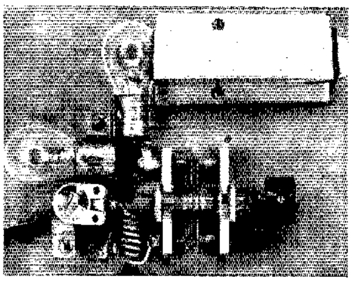

Fig. 5 — View of the homemade power meter. The small piece of "U" aluminum contains the photoresistive cell.

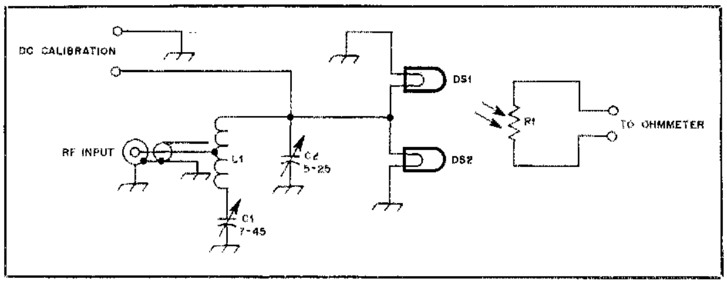

Fig. 6 — Schematic diagram of the "light" rf-power meter. See text for details on construction and operation. R1 is a Radio Shack 276-116 photoresistor, and DS1 and 2 are Sylvania no. 93 bulbs.

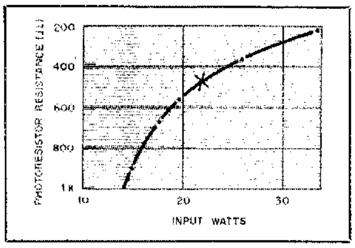

Fig. 7 — Calibration chart for the prototype lamp power meter.

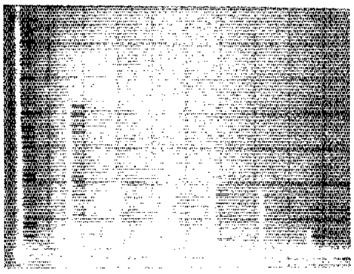

Fig. 8 — Spectral display of the W6IOJ satellite transmitter. Each horizontal division is 100 MHz. Each vertical division is 10 dB. The fundamental signal has been notched 28 dB to prevent analyzer overload. The transmitter does not comply with FCC spectral purity requirements.

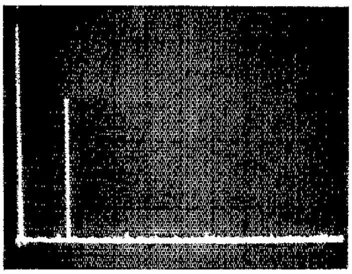

Fig. 9 — Spectral display of the W6IOJ transmitter after the addition of a strip-line output filter. Each horizontal division is 100 MHz, and each vertical division is 10 dB. All spurious and harmonic products are greater than 70 dB down. Several strip-line filter designs are shown on page 15-8 of the 1983 edition of *The Radio Amateur's Handbook* published by ARRL.

Since there is high gain in the final amplifier, the 6146 must be neutralized. Neutralization requires a small capacitance between the grid and plate. This is accomplished with a number 14 wire connected to the grid. The wire is routed along side the tube envelope in close proximity to the plate. A copper strip, 3/8-inch wide, and the length of the tube, is soldered to the wire to increase the capacitance. During adjustments, this is bent in or away from the plate for optimum neutralization. Peak each stage of the driver and final amplifier before applying full plate voltage to the 6146. With 600 V on the plate, power output should be about 40 W.

Power Measurements

Two methods were used to measure the power output. One uses lamps to convert the rf energy to light, and the second employs power resistors for a thermal conversion. Both methods indicated an amplifier efficiency of approximately 55%. Although this is somewhat lower than expected, the measuring devices are inherently lossy and it is probable that several watts were lost in the energy conversions.

The most accurate method was with the lamp power meter (pictured in Fig. 5). Two 12-V high-intensity lamps wired in parallel allow power measurements up to 30 W. Although the dc resistance of each lamp is about 5 ohms, the impedance at 2 meters is high, requiring an impedance-matching network. A good match is obtained by tapping the inductor on a parallel-tuned circuit for the proper input impedance. C1 provides a method for optimizing the tap position, and provides dc isolation. "Eye-balling" the lamps for brilliance is a mistake — they appear very bright even at half the rated input. Guesswork can be avoided by using a cadmium-sulfide photocell for indicating lamp brightness. The schematic diagram of this simple circuit is shown in Fig. 6. The CdS cell resistance varies from approximately 100 Ω to 1 MΩ, depending on the light level. One problem with this method is that the CdS cell is far too sensitive for this application. The detector was modified by placing the cell in a box, with the box facing the lamp and the front blocked except for a 3/32-inch-diameter aperture. The cell is placed in the box in such a manner that it receives only scattered light, avoiding any possibility of hot-spot errors. Calibrate the meter by placing various dc levels on the lamps and recording the photocell resistance. Fig. 7 shows the calibration chart for the prototype meter. Match the unit to a 50-ohm input by inserting an additional 1/4-λ RG-58/U coaxial section in the input circuit and note the difference in loading. Then change the line length by 1/8 λ to maximize the loading difference. Adjust the two capacitors until the loading remains the same regardless of the line length. The line sections involved should use constant-impedance connectors (N, SMA, BNC, etc.).

Conclusions

If the builder uses reasonable care during the construction of this transmitter, he or she will be rewarded by premium performance from a low-cost investment. The transmitter has been in use at W6IOJ for many months, and has provided "solid" satellite contacts.

Notes

[1]Several of the crystals tested in the VXO circuit were purchased from International Crystal Mfg. Co., 10 North Lee, P.O. Box 26330, Oklahoma City, OK 76126. The overtone units are catalog number 031081, and the fundamental 16-MHz crystal is catalog number 031300. Be sure to specify the desired frequency when ordering.

[2]mm = in. × 25.4.

Designing Narrow Band-Pass Filters with a BASIC Program

Computer-aided design will help you add this important circuit to your "designer's toolbox."

William E. Sabin,* WØIYH

In radio-frequency circuit design, whether at 10 kHz or 10 GHz, a very useful circuit is the narrow band-pass L-C filter. This filter type can be designed by first designing a prototype low-pass filter and then converting the prototype into a band-pass filter. This article will discuss this procedure and a BASIC program that can be used to design these filters.[1] This program and an ac circuit analysis program used to analyze filter performance combine to bring this interesting circuit into the repertoire of the experimenter or designer. The analysis program is not presented here, but a suitable program can be found in the literature.[2]

A popular narrow band-pass filter is shown in Fig. 1. There are other ways to configure this filter, but this is the form considered here. The design relies on the idea that for a given center frequency (f_0) and passband width, the end loading and the coupling can be adjusted to obtain predictable passband and stopband responses (Fig. 2). These responses conform closely to those of filters found in catalogs of low-pass prototypes. This applies for passband widths of, say, 1-10% of f_0. At high attenuation values, some departures from the predicted responses do appear. In order to implement the filter design, a means of measuring L, C and Q is required. Some knowledge of coil design is also necessary. Coil measurements should be near f_0 to assure accurate filter designs.

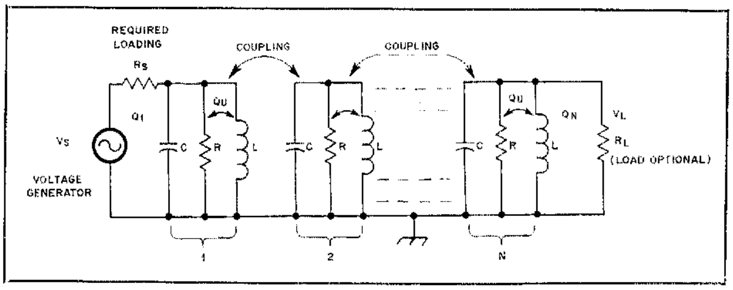

Fig. 1 — Basic narrow band-pass L-C filter shown in nodal form.

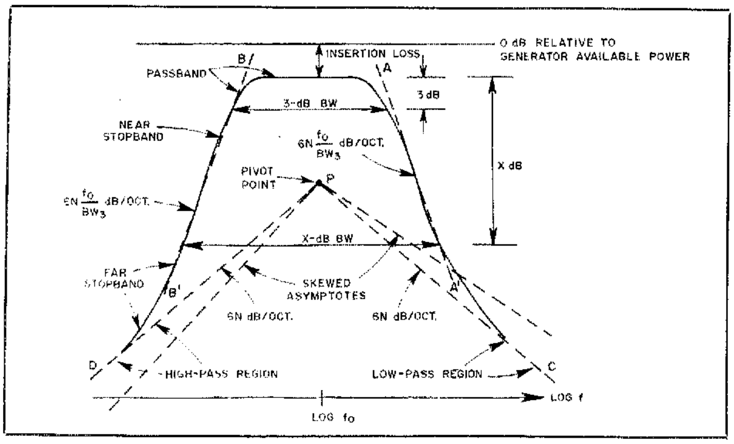

Fig. 2 — Frequency response of a narrow band-pass filter. The passband and *near* stopband responses have almost the same shape as a standard low-pass prototype filter. Lines AA′ and BB′ have slopes of $6Nf_0/BW_3$ in the near stopband region. In the *far* stopband, asymptotes PC and PD, with slopes of 6N, meet at pivot point P. These asymptotes can be skewed, as is shown, by the coupling networks (as in Fig. 13). In the far stopband, the filter reverts to high-pass or low-pass behavior. Point P may be 10 to 120 dB below the maximum response. This behavior is caused by the nonlinear mapping from low pass to band pass, which distorts the shape.

The Low-Pass Prototype

The circuits shown in Fig. 3 (A and B) have approximately the frequency response shown at C. The generator resistance, R_s, is a mandatory part of the filter. To the right of the generator, the first component is a shunt C (if N is odd) or a series L (if N is even). Both networks end with a shunt C. The output load resistance, R_L, is optional. If it is used, it has a definite value that will be determined. At $\omega_c = 1$ ($\omega = 2\pi f$), the response of these networks drops 3 dB. Each L and C has a loss resistance (R_1, R_2, etc.) associated with it, and each component has the same value of Q, called the prototype Q (q_0). Evaluated at $\omega_c = 1$, it is

$$q_0 = \omega_c R_1 C_1 = R_1 C_1 = \frac{\omega_c L_2}{R_2} = \frac{L_2}{R_2} = \cdots \quad \text{(Eq. 1)}$$

*Rockwell International, Collins Telecommunication Products Division, MS 137-138, Cedar Rapids, IA 52498

[1]Notes appear on page 29.

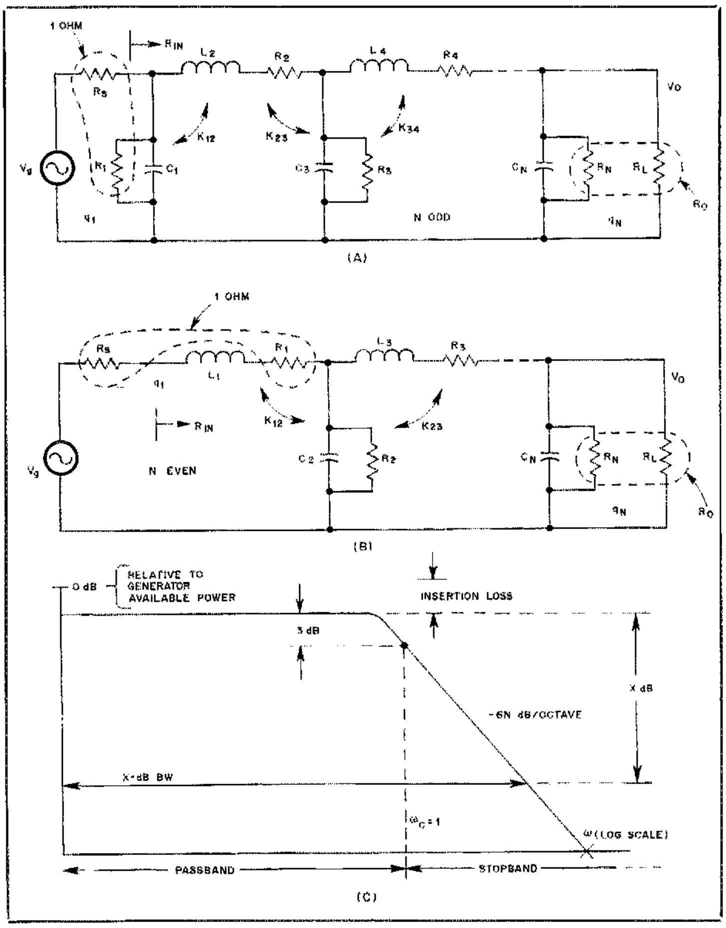

Fig. 3 — Schematic diagrams (A and B) and frequency response (C) of low-pass prototype filters.

Resistors R_s and R_1, in series (Fig. 3B) or in parallel (Fig. 3A), equal 1 ohm (an exception to this will be noted later). The required output loading, R_o, is equal to R_N and R_L in parallel:

$$R_o = \frac{R_N R_L}{R_N + R_L} \qquad \text{(Eq. 2)}$$

Because of loading by R_S and R_1, the first network component has a Q called q_1:

$$q_1 = \omega_c C_1 \frac{R_1 R_S}{R_1 + R_S} = C_1 \text{ (for N odd)} \qquad \text{(Eq. 3)}$$

$$q_1 = \frac{\omega_c L_1}{R_1 + R_2} = L_1 \text{ (for N even)}$$

Likewise, the last filter element (shunt C) has a q_N of

$$q_N = \omega_c C_N \frac{R_L R_N}{R_L + R_N} = C_N R_o$$

(for N odd or even) (Eq. 4)

Note that R_o is not necessarily 1 ohm. The remaining filter elements retain Q values equal to q_0 (from Eq. 1).

Filter elements C_1 and L_2 have a series resonant frequency, ω_{12}, and a "coefficient of coupling," k_{12}, is then defined as

$$k_{12} = \frac{\omega_{12}}{\omega_c} = \omega_{12} = \frac{1}{\sqrt{C_1 L_2}}$$

(for $\omega_c = 1$) (Eq. 5)

Other coefficients of coupling are defined in a similar manner:

$$k_{23} = \frac{1}{\sqrt{L_2 C_3}};\ k_{34} = \frac{1}{\sqrt{C_3 L_4}};\ \text{etc.} \qquad \text{(Eq. 6)}$$

Tables of q and k values are commonly available and are used in this design procedure. The reason for using q and k values, rather than values of L and C, is explained later. Since these q and k values are determined at $\omega_c = 1$, they are the "3 dB down" values, and the tables are labeled in that manner.

In the low-pass prototype, the power available from the generator is

$$P_{GEN} = \frac{V_g^2}{4R_S} \qquad \text{(Eq. 7)}$$

While the power in the load is

$$P_{LOAD} = \frac{V_o^2}{R_L} \qquad \text{(Eq. 8)}$$

From this, an "insertion loss," IL, is defined as

$$IL = 10 \log \frac{P_{GEN}}{P_{LOAD}} \text{ (dB)} \qquad \text{(Eq. 9)}$$

This insertion loss has two parts: A mismatch loss occurs if the input resistance at $\omega = 0$ is not equal to R_S. If q_0 is very large, then $R_{IN} \approx R_L$ and the approximate mismatch loss is[3]

$$L_M \approx 10 \log \frac{(R_S + R_L)^2}{4R_S R_L} \text{ (dB)} \qquad \text{(Eq. 10)}$$

There is also a resistive loss, resulting from dissipation in the L and C elements. For realistic values of q_0, this loss is well approximated by

$$L_R \approx \frac{4.343}{q_0} (C_1 + L_2 + \ldots + C_N) \qquad \text{(Eq. 11)}$$

The insertion loss is, then, with good accuracy, the sum of L_M and L_R.[4] It should be noted that the resistive losses may modify the input resistance at $\omega = 0$ slightly. This results in the actual mismatch loss differing slightly from that given by Eq. 10.

The passband amplitude response of a filter with little or no insertion loss can be flat, rounded or rippled, as shown in Fig. 4. These response shapes correspond to the labels Butterworth, Gaussian (also Bessel or Fano) and Chebyshev. The ripples in the Chebyshev response allow us to define a "ripple bandwidth," as is shown in Fig. 4C. This bandwidth is less than the 3-dB bandwidth and is often specified in filter tables and design specifications. Fig. 5 is a graph that can be used for the conversion from ripple bandwidth (BW_R) to 3-dB bandwidth (BW_3). Fig. 5 will be referred to again.

Note in Fig. 4C that if N is even and q_0 is large the response is reduced at $\omega = 0$. This results from the introduction of a mismatch loss equal to the ripple value. Thus, the output load is not 1 ohm. At the response peaks, however, the transformer action within the filter removes this loss. In the odd-order filters, the mismatch loss occurs at frequencies other than zero and the load is, therefore, 1 ohm.

As the resistive loss increases (q_0

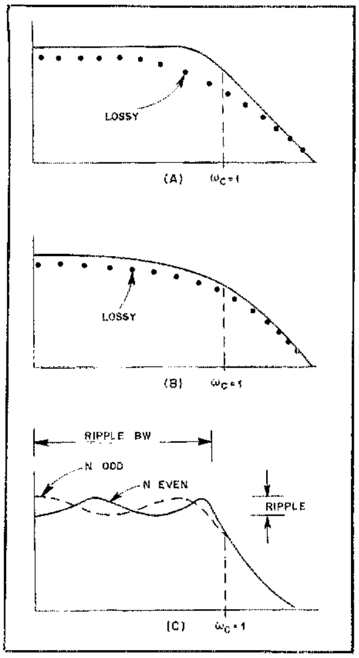

Fig. 4 — Three types of low-pass prototype filter response are possible: Flat (A), rounded (B) and rippled (C).

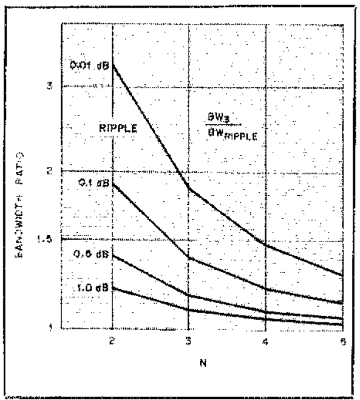

Fig. 5 — This graph can be used to convert from the 3-dB bandwidth to the ripple bandwidth of a Chebyshev filter.

decreases) the passband response changes, as is suggested by the dotted curves in Fig. 4A and B. In the case of the Chebyshev response, the ripples flatten out and the corner becomes rounded as the loss increases. The passband response, particularly the 3-dB bandwidth, also becomes less definite.

The important point is that there are two design approaches that can be used. One is to use the k and q values of the *lossless* prototype, and then to adjust the 3-dB bandwidth until the "lossy" final filter has an acceptable passband response. This is called the "pseudo-exact" approach. It frequently results in very adequate designs having low insertion loss near $\omega = 0$ (this is highly desirable in many applications). The problem is to know how much to change the bandwidth. Reference to lossy filter response curves and trial and error using the computer programs are the methods used in this article.

The second approach is to use a different set of k and q values, and to introduce mismatch loss so that the lossy L-C elements produce exactly the desired response. The additional loss introduced in this approach is often acceptable, especially in filter types other than the Chebyshev, when the exact response is really important. This approach is called the "predistorted" design, and tables of k and q values for various values of q_0 are available.

Now we will consider the stopband response using a geometrical approximation method. At frequencies far above $\omega_c = 1$, the circuits shown in Fig. 3 reduce very nearly to a cascade of voltage dividers, as is shown in Fig. 6. Note that R_1, R_2, R_3, etc. are missing. Using this method, the following "asymptote" formula applies:

$$V_o \approx \frac{1}{\omega^N C_1L_2C_3...C_N} \Big| \text{ N odd}$$

$$\approx \frac{1}{\omega^N L_1C_2L_3...C_N} \Big| \text{ N even} \qquad \text{(Eq. 12)}$$

If we let $\omega = 1$, the value of V_o is

$$V_o \Big| \omega = 1$$

$$= \frac{1}{C_1L_2C_3...C_N} \Big| \text{ N odd}$$

$$= \frac{1}{L_1C_2L_3...C_N} \Big| \text{ N even} \qquad \text{(Eq. 13)}$$

This response, compared to that of the lossless prototype (which also uses a 1-V generator), is shown in Fig. 7. We see that the asymptote at point A does not necessarily "join up" with the prototype response at point B. In the case of the Butterworth filter, they do come together, but for the other types, B is above or below A. For example, a Chebyshev response is shown in Fig. 8. It is apparent that the 20-dB bandwidth estimate is poor, as was the case in Fig. 7. The approximation becomes much better when we apply simple correction factors to the asymptote formulas. Resistive losses less than 2N dB do not invalidate this design procedure. Referring to Fig. 3, the value X is reduced by the value of the resistive loss. For example, if $L_R = 3$ dB, then the 30-dB bandwidth becomes the 27-dB bandwidth.

An interesting prototype filter is formed if we assign a certain uniform value to every L and C element (such as 1H and 1F). The result is an equal element or minimum loss (very nearly true) or Cohn filter. Band-pass versions of this filter are often used in receivers and transmitters. For a given stopband bandwidth, say at 40 dB, the insertion loss at f_0 (for a given q_0) is less than that of any other filter type. Note that for the doubly terminated filter, with a large value of q_0, we have

$$q_1 = q_N;\ k_{12} = k_{23} = ...$$

$$= \frac{1}{q_N} = \frac{1}{q_1} \qquad \text{(Eq. 14)}$$

The two-element pseudo-exact Butterworth is also a minimum-loss filter.

It was mentioned earlier that R_L (Fig. 3) can be an open circuit. The idea of mismatch loss then becomes meaningless, but resistive losses are still present, meaning that the power available from the filter is less than the power available from the generator. Any of the various responses can be achieved over a wide range of R_L values, and catalogs containing the required k and q values are available. The open-circuit case is especially useful when a low-noise common-source FET amplifier is used at frequencies for which the gate impedance is very high, because then no signal power is lost in a terminating resistor.

Frequently, the terminating resistance is the input impedance of a grounded-gate amplifier. In this case, the signal power is not lost, yet the filter is still doubly terminated. Doubly terminated filters are easier to adjust, and singly terminated filters tend to have high input SWRs.

Listed in Table 1 are a few k and q values for various kinds of prototype filters in which q_0 is assumed to be infinite. Values for the lossy, predistorted Butterworth case are also listed. Table 1 is meant only as an illustration and is not complete, but should prove useful.

For the minimum-loss filter, if $N \geq 3$ and $q_0 \geq 5$, an improvement in the shape factor (BW_{60}/BW_3) for a given resistive loss *may* be possible, compared to other types such as the 0.1-dB Chebyshev. For even-order filters, the minimum-loss type avoids the mismatch loss of the Chebyshev filter at $\omega = 0$. The minimum-loss filters are pseudo-exact in that we use the lossless values of k and q for resistive losses up to N dB or so. You should use a circuit analysis program to compare the various

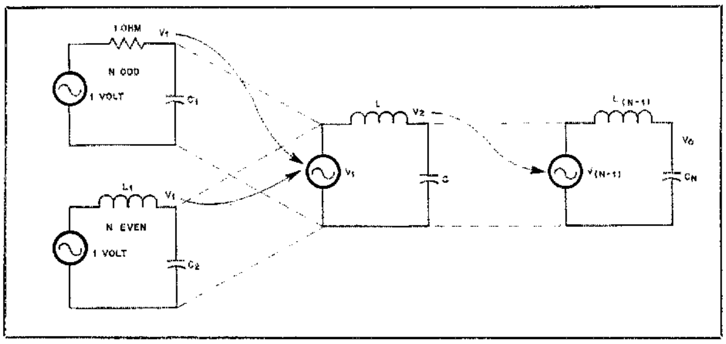

Fig. 6 — Circuit configuration used in calculating the approximate filter stopband response.

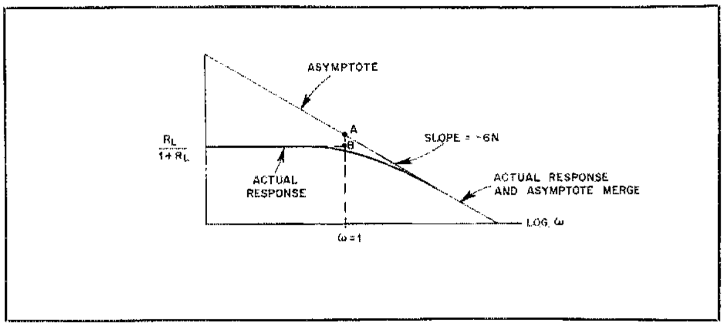

Fig. 7 — The relationship between actual filter response and the stopband asymptote.

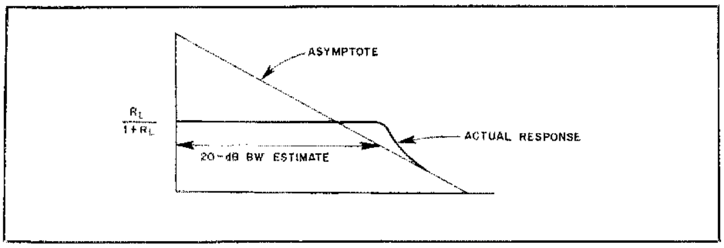

Fig. 8 — The relationship between the Chebyshev response and the stopband asymptote.

prototype filters before making your final selection.

Conversion from Prototype to Narrow Band-Pass Filter

The next step in our filter design is to convert the prototype low-pass circuit into the desired band-pass filter. Fundamental to this conversion is the idea that each prototype L or C element becomes a band-pass filter resonator (as in Fig. 1), with a resonant frequency of f_0.

The loaded Qs of the first and the last resonators are

$$Q_1 = q_1 \frac{f_0}{BW_3},\ Q_N = q_N \frac{f_0}{BW_3} \qquad \text{(Eq. 15)}$$

where BW_3 is the passband 3-dB bandwidth. These values result only from the resonator loss resistance and the adjacent external resistor. Loading caused by the other resonators is not included here. For the internal resonators, the "unloaded" Q_U is used for design purposes:

$$Q_U = q_0 \frac{f_0}{BW_3} \qquad \text{(Eq. 16)}$$

Note that these resonators are not really unloaded.

If q_0 is large and the BW_3 is small, a resonator of very high Q_U is required to satisfy the resistive loss, mismatch loss and the frequency response goals. The resistive loss and the mismatch loss are identical to those given by Eqs. 10 and 11 for the prototype.

Table 1

k, q (3 dB down) Values for Low-pass Prototypes

N	q_0	q_1	q_N	k_{12}	k_{23}	k_{34}	k_{45}
Butterworth (doubly terminated)							
2	any	1.4142	1.4142	0.7071			
3	inf	1.0	1.0	0 7071	0.7071		
3	10	0.8007	1.5359	0.7388	0.6716		
3	5	0.8226	1.7115	0.6567	0.7060		
4	inf	0.7654	0.7654	0.8409	0.5412	0.8409	
4	8.7	0.5407	1.8605	1.0411	0.5373	0.6992	
4	4.36	0.5819	2.2961	0.9051	0.4592	0.7706	
5	inf	0.618	0.618	1.0	0.5559	0.5559	1.00
5	8.1	0.610	0.8157	1.0075	0.4844	0.6887	0.6278
Chebyshev (0.1 dB) (doubly terminated)							
2	any	1.6382	1.6382	0.7106			
3	inf	1.4328	1.4328	0.6618	0 6618		
4	inf	1.3451	1.3451	0.6850	0.5421	0.6850	
5	inf	1.3013	1.3013	0.7028	0.5355	0.5355	0.7028
Minimum Loss (doubly terminated)							
2	inf	1.4142	1.4142	0.7071			
3	inf	1.5214	1.5214	0.6573	0.6573		
4	inf	1.6529	1.6529	0.6050	0.6050	0.6050	
5	inf	1.7458	1.7458	0.5728	0.5728	0.5728	0.5728
Butterworth (singly terminated)							
2	inf	0.707	inf	1.00			
3	inf	0.500	inf	1.22	0.707		
4	inf	0.383	inf	1.56	0.765	0.644	

Generally, Q_u (usually assumed to be the coil Q) is known. Thus, for a desired value of f_0/BW_3, q_0 and, therefore, the resistive loss are established. Consider resonator 1 in Fig. 1: We can freely choose the value of L, C or R_1 (R in parallel with R_S), and the remaining two factors are then defined (since f_0 and Q_U are known) by manipulating the following formulas

$$C = \frac{1}{\omega_o^2 L} = \frac{Q_1}{\omega_o R_1};\ \omega_o = 2\pi f_0 \qquad \text{(Eq. 17)}$$

Often, we choose an inductor that has a high Q at f_0, or a standard capacitance value. In this design procedure and for the program, we choose to make the values of L and C the same for every resonator, although this is not mandatory.

The coupling coefficient between resonators is

$$K_{12} = k_{12} \frac{BW_3}{f_0}, \text{ etc.} \qquad \text{(Eq. 18)}$$

We can now see why k and q values were used for the prototype design. They are related closely to the K and Q values of coupled, loaded resonators. It is apparent, for example, that two slightly "over coupled" resonators will have a Chebyshev response. "Critical coupling" yields a Butterworth response, while "under coupling" produces a Bessel (or Gaussian) response.

It is important to keep in mind that the low-pass bandwidth shown in Fig. 3 (from zero to ω_{LP}) is related to the bandwidth, BW_X, in Fig. 2 by

$$BW_X(BP) = BW_3(BP) \cdot \omega_{LP} \qquad \text{(Eq. 19)}$$

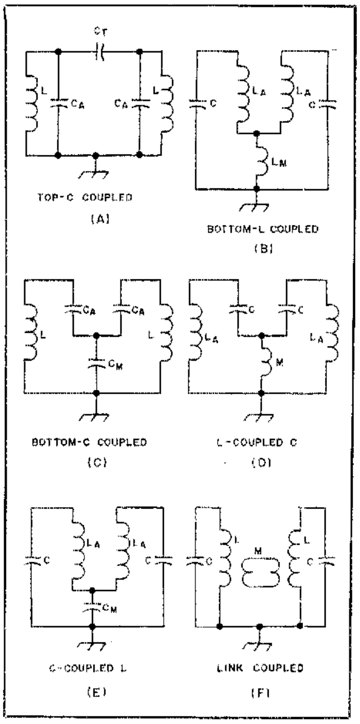

Fig. 9 — Various methods can be used to provide coupling between band-pass filter resonators. The methods shown here are discussed in the text.

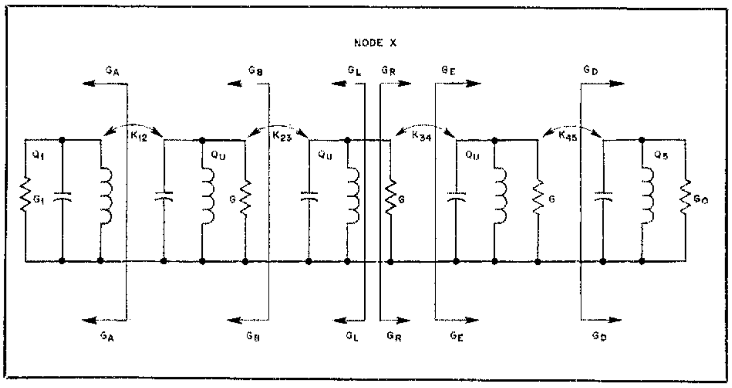

Fig. 10 — Circuit configuration used in calculating the resistance at node X.

The ratio of BW_X to BW_3 is the same in both the low-pass and the band-pass filters. This proportionality holds despite the nonlinear transformation described in Fig. 2. If, for the Chebyshev filter, we want to specify the ripple bandwidth rather than the 3-dB bandwidth (BW_3), it is necessary only to increase the value of BW_3 by the factor obtained from Fig. 5. As previously discussed, further adjustment of the lossy filter may be needed.

Designing the Band-Pass Filter

Shown in Fig. 9 are the various ways to obtain coupling between the band-pass filter resonators. The theory of this coupling can be based on the "J-inverter."[5] In the computer program, the values for coupling method A are computed first. The values for the other methods are then derived by modifications and transformations. Improved attenuation at frequencies far above f_0 is obtained by using the coupling methods shown at B, E and F. Methods A, C and D are better far below f_0. The value of C_T in Fig. 9 is simply

$$C_T = K_{12} C = K_{12}(C_A + C_T) \qquad \text{(Eq. 20)}$$

where C is the value given by Eq. 17. In a gang-tuned filter, we would look for a coupling method that maintains the prototype response over, say, a 3 to 1 frequency range. The correct end loading must also be maintained. As f_0 is changed, the coupling and the end loading must change in such a way that the original prototype k and q values still describe the band-pass filter. This is less difficult to achieve if the ratio BW_3/f_0 remains constant (a constant percentage bandwidth is maintained).

We can now examine how to find the resistance, at f_0, at nodes within the filter. In the example shown in Fig. 10, node X is loaded from the right (G_R) and from the left (G_L). Thus, the node resistance is

$$R_X = \frac{1}{G_L + G_R} \qquad \text{(Eq. 21)}$$

To find G_L, we start at G_I and calculate G_A, G_B and then G_L, looking only to the left:

$$G_A = G_I = \frac{1}{\omega_0 L Q_I} \qquad \text{(Eq. 22)}$$

$$G_B = \frac{G_I^2}{G_A}(Q_I K_{12})^2 + G$$

$$G_L = \frac{G_I^2}{G_B}(Q_I K_{23})^2$$

Then, starting at G_o, we work to the left, looking only to the right:

$$G_D = G_o = \frac{1}{\omega_0 L Q_5} \qquad \text{(Eq. 23)}$$

$$G_E = \frac{G_o^2}{G_D}(Q_5 K_{45})^2 + G$$

$$G_R = \frac{G_o^2}{G_E}(Q_5 K_{34})^2 + G$$

These formulas are theoretically exact, even in low-Q situations.

Of particular interest is the fact the resistance at node X can become very high, making the resonator tune very sharply. In a transmitter, the voltage at node X can become very large; that is, a degree of transformer action occurs in the band-pass filter that is not found in the low-pass prototype. A circuit analysis program can be used to determine these voltages.

When a singly loaded filter is connected to the gate of an FET amplifier, the output resistance looking into the filter is important in determining the amplifier noise figure. A capacitive divider can be used to transform this resistance down to a better value (Fig. 11F). This also helps prevent amplifier overdrive caused by excessive voltage step-up within the filter. More uniform impedance levels can be obtained by using a minimum-loss design. In the minimum-loss design, all the resonators are equally loaded, as verified by Table 1 and Eq. 22. Another technique that can be used is to modify the resonator L-C ratios. This procedure is not provided for in the computer program, but it is covered in the literature.

An "intrinsic" filter impedance is established by R_I, the parallel combination of R and R_S in Fig. 1 and Eq. 17. We must transform the actual source and load resistances to that value so that the filter is properly terminated at both ends. Proper termination is necessary if the desired filter performance is to be obtained. The transformation methods used in the program are shown in Fig. 11.

Often, approximate formulas are not adequate in low-Q cases. Further, if we want to analyze a design with a circuit analysis program, highly accurate component values are useful. Therefore, the program algorithms begin with approximate formulas and then the values are adjusted until the resonance and impedance requirements are satisfied to four significant figures.

In the case of link coupling, the coupling coefficient must exceed a certain value or the transformation cannot be achieved. Also, the link coil Q should be specified. It is also possible to use a tapped inductor method.[6]

There are also certain limitations when top C or top L coupling is used. These

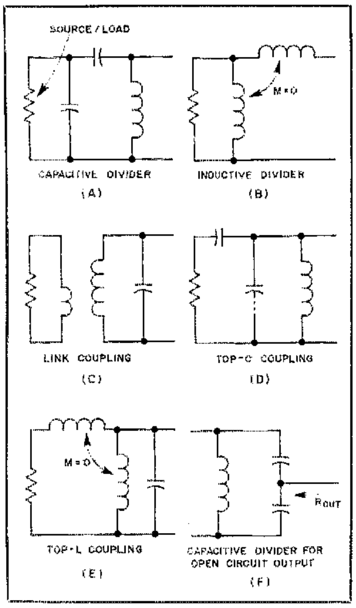

Fig. 11 — Each filter has an instrinsic resistance. Shown here are several methods that can be used to transform the source and load resistances to the intrinsic resistance of the filter.

limitations are indicated by the computer program. Both of these coupling methods have a useful property: The impedance looking from the loading resistor into the filter becomes very high at frequencies outside the passband; that is, the load becomes "decoupled" from the filter.

The C and L dividers shown in Figs. 11A and B can be used to absorb certain amounts of load reactance. If the load is a complex combination of L, C and R, the situation becomes a problem in approximate matching over the passband.[7] Over a narrow passband, a constant load and source impedance can usually be safely assumed. Trouble can occur, however, at frequencies outside the passband (discussion of this complex subject is beyond the scope of this article).

In transmitter and certain other applications, the output and input SWR of the filter are of great interest. Slevin has found that filters using the pseudo-exact procedure for 0.01-dB Chebyshev responses have less than a 1.25 SWR over the ripple bandwidth, especially if the insertion loss is less than 2 dB.[8] Predistorted designs tend to have higher values of SWR. The Butterworth pseudo-exact is quite good in this respect over 60% of the 3-dB bandwidth if N is greater than 2. Minimum-loss designs have good SWR over 40% of the bandwidth.

From Figs. 9 and 11, it can be seen that certain combinations of coupling and termination circuits are somewhat awkward. For example, the coupling circuit of Fig. 9C or D would not combine well with the transformation circuits shown in Fig. 11A or F. The circuits shown in Fig. 11B, C or E would be better choices for use with these coupling circuits. Also, the matching transformers may not be able to accommodate certain values of load or source resistance. The program detects this condition and informs the user. Changing the filter impedance or transformer will solve this problem.

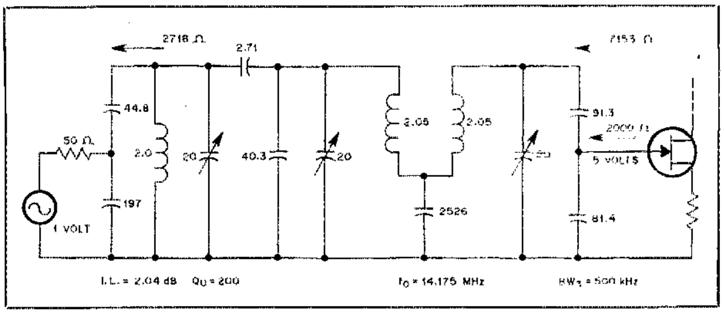

Fig. 12 — This example is a three-resonator, singly loaded Butterworth pseudo-exact design.

The use of link-coupling (mutual inductance) circuits requires that the coupling coefficient be known. This necessitates some supplemental measurements.[9,10] For pseudo-exact designs, the generator and load resistances of the low-pass prototype have the same values as those of the lossless filter. The program makes this adjustment and produces the correct mismatch and insertion-loss values.

The Program

The preceding discussion greatly reduces the need to talk about the computer program at great length. Table 2 is a printout of the design example shown in Figs. 12 and 13. The program requires 8.5K bytes of memory in an Apple II® computer. Another 2K of memory is required for the solution of a 10-resonator problem. Operation of the program is "menu" oriented and is highly interactive. Provisions have been made for the inclusion of a stray or trimmer capacitance value for each resonator in the circuit. The inductance values used are the "effective" values measured on a Q meter at f_0. That is, the inductance value measurement is affected by the coil self capacitance. If the self-resonant frequency of the coil is more than 2.5 times f_0, no significant errors result from this approximation in a narrow-band filter.

In a filter with an open-circuit output, the output resistance at f_0 is determined by the program and a capacitor divider is calculated that transforms the output resistance to the desired value. Also, the values of L, C and R for the low-pass prototype are provided, according to the diagram in Fig. 3. The values of the load resistance and the resistive loss and the

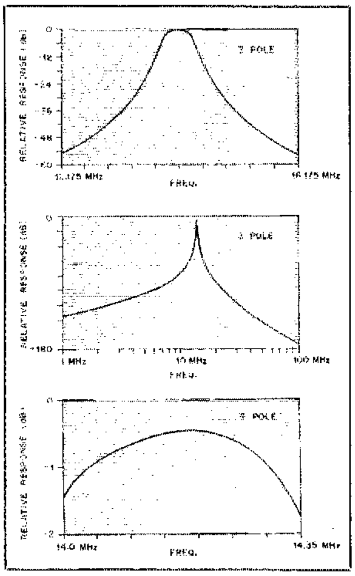

Fig. 13 — Amplitude versus frequency response of the design example discussed in the text and shown in Fig. 12.

mismatch loss (if it is defined) are also determined. The stopband width approximation algorithm has proven to be reasonably accurate in a large number of situations. A circuit-analysis program should be used to verify the filter performance before construction is begun.

Theory and Practice

It is apparent from the discussion that small changes in loading and in the coupling between resonators can make a big difference in the response, insertion loss, and so on. If slug-tuned coils and variable capacitors are used, an infinite and bewildering array of tuning possibilities exist. On the other hand, very accurate measuring equipment is needed to select components of exact values. The following compromise procedure is suggested for practical situations:

1) Use the computer program to design

Table 2

Design Example Printout

```
]RUN
DESIGN DISHAL NARROWBANDPASS FILTER.

2 TO 10 RESONATORS.

NUMBER OF RESONATORS=3
CENTER FREQUENCY=14.175E6
3 DB BANDWIDTH=5E5
ENTER 'INF' FOR INFINITE VALUE
UNLOADED COIL Q=200
NORMALIZED QP OF LOW PASS
 PROTOTYPE=7.05467372

BASED ON QP, DETERMINE THE LOWPASS
 PROTOTYPE 3 DB DOWN K,
  Q VALUES FROM SOME REFERENCE SOURCE

IS QP LARGE ENOUGH (Y/N)?
ENTER 'INF' FOR INFINITE Q
INPUT TERMINAL IS RESISTANCE LOADED
Q(1)=.5
Q(3)=?INF

K(1,2)=?1.22
K(2,3)=?.707

LOADED, DENORMALIZED Q:
Q(1)=14.175      Q(3)= INF

INPUT RESONATOR SPECIFICATION MENU:
SPECIFY 'L', 'C', 'R' :L
L=2E-6

L=2E-06
C=6.30324418E-11
R=2524.9687

ARE L,C,R SATISFACTORY (Y/N)?

ARE THE UNLOADED COIL Q AND THE 3 DB
 BANDWIDTH STILL SATISFACTORY (Y/N)?
COIL RESISTANCE=35625.6607

REQUIRED SOURCE RESISTANCE=2717.57697
REQUIRED LOAD RESISTANCE =INF
```

```
CT(1)=2.7125072E-12
C(1)=6.03199346E-11
CT(2)=1.57192015E-12
C(2)=5.87480144E-11
C(3)=6.14605217E-11

WIRING + AVERAGE TRIMMER C FOR EACH
 RESONATOR=20E-12
C(1)=4.03199346E-11
C(2)=3.87480144E-11
C(3)=4.14605216E-11

TYPE 'SPACE' TO GO ON

COUPLING MODIFICATIONS MENU

TYPE 1 FOR CAPACITOR PI TO TEE
TYPE 2 FOR L COUPLED C
TYPE 3 FOR INDUCTOR TEE
TYPE 4 FOR C COUPLED L
TYPE 5 FOR M COUPLING
TYPE 6 TO CONTINUE
?4

FIRST NODE=2
SECOND NODE=3
C(2)=4.03199346E-11
C(3)=4.30324418E-11
CM=2.52753851E-09
L(2)=2.04987654E-06
L(3)=2.04987654E-06

COUPLING MODIFICATIONS MENU

TYPE 1 FOR CAPACITOR PI TO TEE
TYPE 2 FOR L COUPLED C
TYPE 3 FOR INDUCTOR TEE
TYPE 4 FOR C COUPLED L
TYPE 5 FOR M COUPLING
TYPE 6 TO CONTINUE
?6

DESIGN I/O IMPEDANCE TRANSFORMERS

SELECT TYPE OF TRANSFORMER AT INPUT:
```

```
TYPE 1 FOR C DIVIDER
TYPE 2 FOR L DIVIDER
TYPE 3 FOR LINK COUPLING
TYPE 4 FOR TOP C
TYPE 5 FOR TOP L
TYPE 6 TO CONTINUE
?1

SOURCE RESISTANCE=50

C(1)=4.47660911E-11
BOTTOM CAP=1.97111192E-10

OUTPUT PORT IS OPEN CIRCUIT
TYPE 'SPACE' TO GO ON
FIND INSERTION LOSS
FOR PSEUDO-EXACT, INPUT '1'
FOR OTHERS, INPUT '0' 1
PROTOTYPE Q=7.05467372
PROTOTYPE RB=1
RESISTIVE LOSS (DB)=2.05160031
MISMATCH LOSS NOT DEFINED

LOWPASS PROTOTYPE VALUES
X(1)=.5          R(1)=14.1093474
X(2)=1.34372481 R(2)=.190472991
X(3)=1.48884963 R(3)=4.73833863

APPROXIMATE STOPBAND RESPONSE
ENTER 1 FOR CHEBY, MIN-LOSS,
 0 FOR OTHERS 0

17.9 DB BW=1077098 HZ
27.9 DB BW=1580956 HZ
37.9 DB BW=2320515 HZ
47.9 DB BW=3406034 HZ
57.9 DB BW=4999350 HZ

UNLOADED OUTPUT RESISTANCE AT F0=7153
DESIRED OUTPUT RESISTANCE=2000
C DIVIDER, C1 BOTTOM, C2 TOP
C1=8.13813902E-11
C2=9.13203642E-11

COMPLETE
```

the filter. A good estimate of the coil Q is very helpful.

2) Model the filter with a circuit analysis program, obtaining amplitude response versus frequency.

3) Pick two frequencies in the lower stopband, two in the upper stopband and three in the passband.

4) Build the filter using component values that are as close as possible to the calculated values.

5) Trim the L and C values until the response fits the data points chosen in (3). Use an oscillator, a frequency counter and a simple rf voltmeter or a receiver and attenuator to make the measurements.

6) The top coupling capacitors (C_T) are the most critical. Build a 4-MHz oscillator, and measure the frequency change when C_T is connected across the tank circuit. Use the formula:

$$\left| \Delta f \right| \approx \frac{f_0}{2_C} \cdot C_T \qquad \text{(Eq. 24)}$$

This simple procedure will greatly reduce the alignment effort.[11]

7) Confirm the SWR (if it is important) by using an analysis program or by measurement.[12]

The methods of adjusting narrow bandpass filters described by Dishal should be studied carefully.[13] An interesting, brief discussion has also been presented by Hayward.[14] The job of tuning a filter with several resonators can be quite a challenge, and the formal procedures suggested in these references can be helpful.

Acknowledgments

I wish to express my appreciation to R. C. Edwards and H. L. Landt, of Collins Telecommunications Products Division, for the helpful discussions that enabled me to better understand this subject. Dick Edwards has been especially generous with his involvement in this project.

Notes

[1]The author will place this program on a disk for the Apple II® or the IBM PC® upon receipt of a disk, return postage and an envelope. The program listing can be obtained for $1 from ARRL, Dept. TD-Sabin, 225 Main St., Newington, CT 06111.

[2]R. Steincross, "BASIC Program Performs Circuit Analysis," *EDN*, Sept. 1, 1982.

[3]H. Blinchikoff and A. Zverev, *Filtering in the Time and Frequency Domain* (New York: Wiley and Sons, 1976), Chapters 3, 4 and 6.

[4]S. Cohn, "Dissipation Loss in Multiple Coupled Resonators," *Proc. IRE*, Aug. 1959.

[5]A. Zverev, *Handbook of Filter Synthesis* (New York: Wiley and Sons, 1967, Chapters 3, 6 and 10.

[6]H. Krause, C. Bastaian and F. Raab, *Solid State Radio Engineering* (New York: Wiley and Sons, 1980), Chapter 3.

[7]F. Davis, "Matching Network Design With Computer Solutions," *Motorola RF Design Handbook*, 2nd ed. (Phoenix: Motorola, Inc., 1980), pp. 20-35 to 20-45.

[8]R. Slevin, "Pseudo-Exact Band-Pass Filters," *Microwaves*, Aug. 1968.

[9]*ITT Reference Data for Radio Engineers*, 6th ed. (Indianapolis: Howard Sams, 1975), Chapter 13.

[10]K. Henney, *Radio Engineering Handbook* (New York: McGraw-Hill, 1959).

[11]F. Noble, "The CMO — A Capacitance-Measuring Oscillator," *QST*, Aug. 1979, pp. 38-39.

[12]J. Priedigkeit, "A Reflection-Coefficient Bridge — Impedance Measurements the Easy Way," *QST*, Oct. 1981, pp. 18-20.

[13]M. Dishal, "Alignment and Adjustment of Synchronously Tuned Multiple-Resonator Filters, *Proc. IRE*, Nov. 1951.

[14]W. Hayward, *Introduction to Radio Frequency Design* (Englewood Cliffs, NJ: Prentice-Hall, 1982), Chapters 2 and 3.

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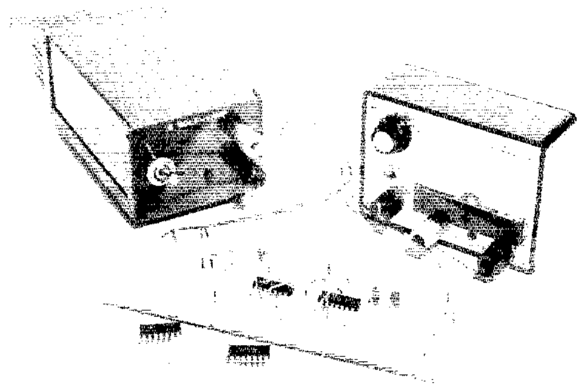
A Beginner's Look at Digital Electronics

ANDs, NANDs and NORs have you mystified? This introduction to digital logic should help clear up the subject for you.



By George Collins,* KC1V

Have you looked at any new rigs lately? Many of them are loaded with digital logic circuits. Keyers, repeater controllers and modern RTTY (radioteletype) gear are all based on digital logic. A basic understanding of digital electronics has become essential to amateurs today. With microcomputers playing an increasing role in our amateur activities, digital techniques will become even more important to the amateur who wants to understand "how it works." Besides, experimenting with logic circuits can be very enjoyable. So let's get started on a beginner's look at logic.

Logic States

The logic devices we will be discussing have only two states, or conditions. They can be "on" or they can be "off," but intermediate conditions are not possible. A toggle switch is a good example of such a device; it can be open (off) or it can be closed (on), but it can't be in between. The on and off states can also refer to different voltage levels, current flows, frequencies, or other physical conditions.

Because we are dealing with only two possible states, the binary number system, which contains only the numerals 0 and 1, is ideal for representing logic-device states. We could, for example, assign the closed-switch position a value of 1, and the open position could be represented by 0. Another way of describing logic states is to define one state as the "true" condition and the other state as the "not true," or "false," condition. In the switch example, we might define the closed position as the true state. The open position would then be the false condition.

Basic Logic Operations

There are three basic logic operations we need to understand. They are AND, OR and NOT. An AND gate is a device that performs the AND operation. It has two or more inputs and a single output. The state of the output is determined by the logic states at the device inputs. If an AND gate has two inputs, A and B, both A *and* B must be true in order for the output to be true. If the AND gate has more than two inputs, *all* of them must be true before the output will be true. We can construct an AND gate by wiring two switches together as shown in Fig. 1A. The switch positions serve as the AND gate inputs, while the output state is displayed by the lamp. Only if both switches are closed will the lamp light (the true state).

Fig. 1B is the diagram of a toggle-switch OR gate. As the name implies, the output of this gate will be true if input A *or* input B is true. The output of an OR gate with more than two inputs will be true if *any* input is true.

The remaining basic logic operation, NOT, is also called the inverse or the complement operation. NOT gates are generally referred to as inverters. Unlike AND and OR gates, inverters have only one input. The output state of an inverter is always the opposite of the input state (Fig. 1C).

Truth Tables and a Few More Gates

By making a list of all possible combinations of input states and the resulting output states, we can describe the operation of any logic gate in an easy-to-understand manner. Such a list is called a truth table. Shown in Fig. 2 are truth tables for the AND, the OR and the NOT gates. In these tables I have chosen to use a 0 to represent the false condition and a 1 to indicate the true state. Next to each truth table is the standard logic symbol for the corresponding gate. These are the symbols we find in schematic diagrams of circuits containing integrated-circuit (IC) logic gates.

Also shown in Fig. 2 are the logic symbols and truth tables for three other gates. Two of these, the NAND and the NOR, are related to gates we've already looked at. The NAND gate is equivalent to an AND gate followed by an inverter. If we compare the truth tables for the NAND and the AND gates, we see that for any set of input conditions the output of the NAND gate is the inverse of the AND gate output. The symbol used for the NAND gate is the same as that used for the AND gate, except that a small circle has been added at the output. This circle indicates that the inversion operation has been applied to the output; thus, we have a NOT-AND gate. This type of relationship also applies to the NOR (NOT-OR) and OR gates. A NOR gate is the same as an OR gate with an inverter placed at the output.

*Assistant Technical Editor

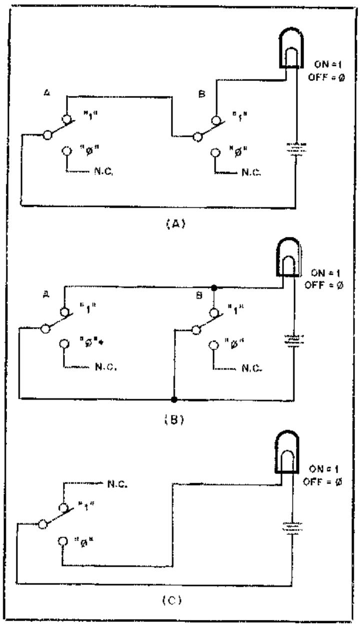


Fig. 1 — A lamp, a battery and toggle switches can be used to demonstrate simple logic operations. The circuit shown at A performs the AND operation, while the OR function is implemented at B. At C is the inverse, or NOT, operation.

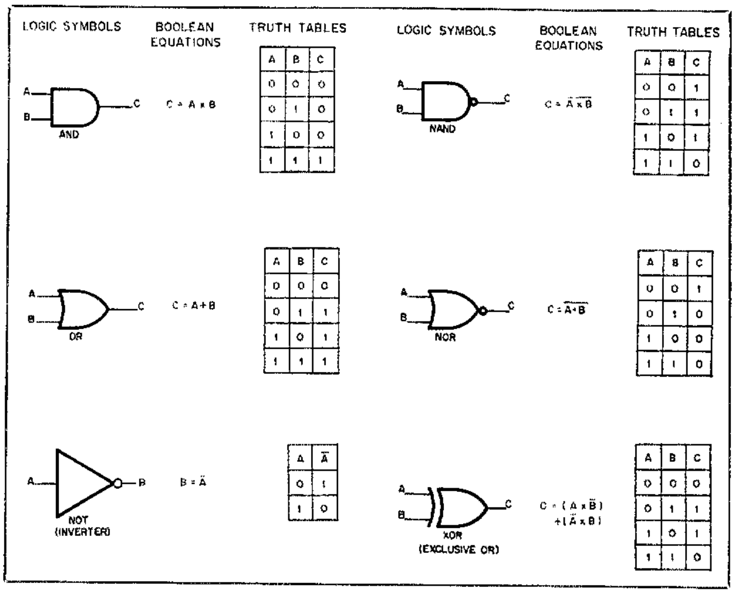


Fig. 2 — Standard logic symbols, Boolean equations and truth tables for the logic gates discussed in the text.

The last gate shown in Fig. 2 is similar to the OR gate, but with an important difference. The XOR, or "exclusive or," gate is like the OR in that the output will be true if input A or input B is true (that's the *or* part). If A and B are *both* true, the XOR output, unlike the OR output, will be false. One input or the other can be true, but not both (that's the *exclusive* part). By placing an inverter after the XOR gate we can form the XNOR function. It's analogous to the NAND and the NOR functions.

Another Way to Represent Logic Functions

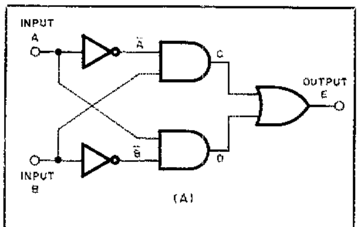
As useful as truth tables are, they do have limitations. One problem is that they take too long to write out for complex circuits. We need a shorter way to represent logic circuits. This brings us to Boolean algebra. Don't worry, we won't be wading through any proofs or theorems. We're just going to "borrow" a few symbols and some basic principles to help us in our logic work.

In normal algebra, there is a symbol for each mathematical operation, such as addition (+), subtraction (−) and multiplication (×). There is also a Boolean algebra symbol for each of the logic operations. The OR operation symbol is "+." For example, we can write the logic expression for a two-input OR gate as A + B = C. The AND operation is represented by x, and a bar over a variable indicates the NOT operation. If the bar is placed over an expression, such as $\overline{A + B}$ = C, it means that you first find the value of the expression (A + B) and then take the inverse. If A + B = 0, then $\overline{A + B}$ = 1. There is also a symbol for the XOR operation, it is ⊕. The Boolean equations for the various gates are given in Fig. 2.

Analyzing Logic Circuits

So far we have been doing "ground work" — learning the basics of logic. Now we can begin to use these basics to help us understand how logic circuits function.

There is usually more than one way to do something, and analyzing a logic circuit is no exception. The circuit shown in Fig. 3A, for example, can be analyzed in a number of ways. One method is to construct the truth table for the circuit. We begin by writing down all the possible input combinations. Since the circuit has two inputs, there are four possible combinations. These are listed in the left-hand columns of the truth table shown in Fig. 3B. To make it easier to keep track of the various logic states, four intermediate circuit points have been labeled (Ā, B̄, C, D). These points are included in the truth table. Circuit point E, the output, is the last column of the truth table. Now, all we have to do is fill in the table and we'll know the output state for any combination of input states!



A	B	Ā	B̄	C	D	E
ø	ø	1	1	ø	ø	ø
ø	1	1	ø	1	ø	1
1	ø	ø	1	ø	1	1
1	1	ø	ø	ø	ø	ø

(B)

Fig. 3 — Analyzing logic circuits is simplified by using a systematic approach. Two methods of analyzing this circuit (A) are explained in the text. The truth table for the circuit is shown at B.

It's really easier than it might sound. First, Ā and B̄ are simply the inverses, or complements, of A and B. If A is a 0, then Ā is a 1. The next step is to determine the states of points C and D. Fig. 3 shows that the state at point C is the AND of Ā and B. Anytime Ā and B are *both* 1, C will be a 1. Using the Ā and the B table columns, we can fill in the C column. We can fill in the D column in exactly the same manner, except that A and B̄ are the inputs. The last step is to form the OR of C and D to produce E, the circuit output. This completes the truth table and the analysis of the circuit. Easy, wasn't it?

In the "truth table approach," we start

at the inputs and work toward the output. We could start at the output and, using the Boolean algebra expressions we talked about earlier, work toward the inputs.

To use this approach to analyze the Fig. 3 circuit, we begin by writing the Boolean expression for point E, the output:

$$E = C + D \qquad \text{(Eq. 1)}$$

Then we move a step closer to the inputs by writing the expressions for C and D:

$$C = \bar{A} \times B \qquad \text{(Eq. 2)}$$

$$D = A \times \bar{B} \qquad \text{(Eq. 3)}$$

By combining these expressions, we obtain the expression for the complete circuit:

$$E = (\bar{A} \times B) + (A \times \bar{B}) \qquad \text{(Eq. 4)}$$

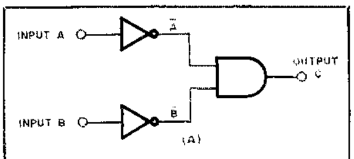
Does this look familiar? It's the same as the expression for the XOR gate shown in Fig. 2. In fact, Fig. 3 is one way the XOR operation can be performed by a combination of AND, OR and NOT gates.

This brings up an important point: We can perform *any* logic operation by using a combination of OR gates and inverters or AND gates and inverters. As an illustration, let's form a NOR gate from an AND gate and some inverters.

Examining the truth tables in Fig. 2, we see that the NOR gate output is true when A *and* B are false. The "and" is the clue. If we invert both inputs and then AND them, the result is exactly the same as a NOR gate (Fig. 4). This result can also be written as a logic expression:

$$\bar{A} \times \bar{B} = \overline{A + B} \qquad \text{(Eq. 5)}$$

By the way, this is an application of what is known as DeMorgan's theorem. It is one of the important results of Boolean algebra. There are many other examples of this idea. For instance, how would you construct an OR gate using AND gates and inverters? The truth tables in Fig. 2 and 4 should give you a hint.



A	B	$\bar{A}$	$\bar{B}$	C
0	0	1	1	1
0	1	1	0	0
1	0	0	1	0
1	1	0	0	0

(B)

Fig. 4 — A particular logic function can be performed by a combination of other types of logic gates.

Some Real Logic Devices

So far, we've been discussing digital logic in general terms, but have not talked specifically about how the logic is implemented in practical circuits. The switch and lamp circuits of Fig. 1, while good illustrations, are not the kind of logic circuits in which amateurs are generally interested. So let's take a look at some IC logic devices — types that are frequently used in Amateur Radio work.

Transistor-Transistor Logic

TTL, short for transistor-transistor logic, is one of the two most popular types, or families, of IC logic devices. Of the various TTL devices, the 7400 series is perhaps the most common. The 7400 series ICs you are most likely to encounter are packaged in dual in-line packages, or DIPs. Shown in Fig. 5 are some 7400 series gates. Like many TTL ICs, these devices contain more than one gate. For example, the 7421 (Fig. 5C) contains two AND gates, each of which has four inputs. This device is referred to as a dual, 4-input AND gate. Because the gates in a single package may be used in *different parts* of a schematic diagram, the representations used in Fig. 5 aren't always convenient. Normally, the individual gates are drawn without the package outline. Pin numbers are then written near the appropriate input and output lines to identify the gates.

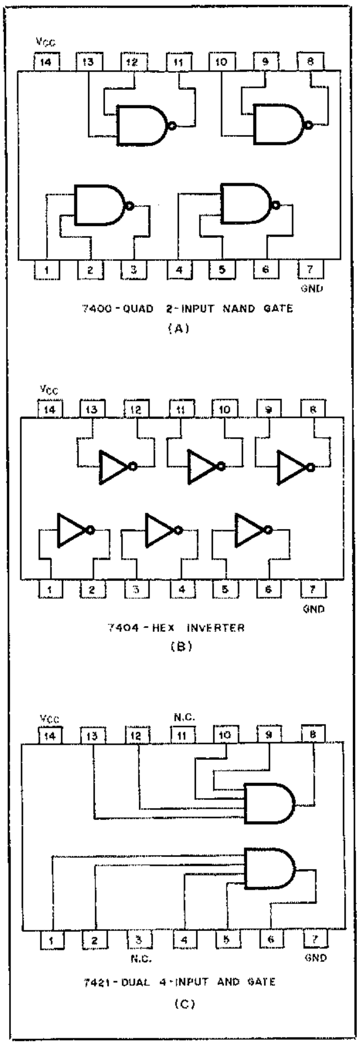


Fig. 5 — Package outlines and pinouts of some common 7400-series devices.

In addition to the simple logic gates, other functions are available in the 7400 series. These include flip-flops, counters, registers and many others (there are over 300 devices in the series!). We'll have to save those devices for another time when we discuss sequential logic.

Electrical Characteristics

Earlier, we related true and false logic states to some physical condition, like the position of a switch or whether or not a lamp is on. When dealing with logic ICs, the physical property most often used to represent logic states is voltage. If an input or output voltage is above a specific value, it represents one of the logic states and we refer to it simply as a HIGH level. Voltages below a specified limit are called LOW levels and represent the other logic state. As long as the voltages are above or below the limits, we don't care what the exact values are. Generally, the true state is assigned to the HIGH level and the LOW level is taken to be the false state. It is also common to assign the binary value 1 to the HIGH level and 0 to the LOW level.

Table 1 contains some of the general electrical specifications for three types of TTL ICs. Keep in mind that these are *general* specifications; specific devices may have slightly different specifications (you'll want to consult a manufacturer's data book when dealing with a critical parameter). First, let's see what Table 1 tells us about the voltage limits we spoke of earlier. Starting with the standard 7400-series input values, we see that the *minimum* HIGH input voltage is 2.0 V. This means that if the voltage at a gate input is 2.0 V or greater (but below a maximum of about 5.0 V), we can be certain that the gate will respond to the input as a HIGH level.

The *maximum* LOW input voltage is listed as 0.8 V, meaning that the input voltage must be at or below this level in order to guarantee a LOW input. Between 0.8 and 2.0 V is a sort of "no man's land." Voltages in this region are undefined and can result in improper circuit operation.

As you may have guessed, the output voltage levels produced by 7400-series devices are compatible with the input level requirements. This allows the output of one device to be directly connected to the input of another, simplifying our logic circuits. The HIGH level output voltage is

Table 1

Electrical Characteristics of Common Logic Families[4]

	Standard TTL (active pull-up) V_{cc} = +5.0 V	*Low-power TTL (74L)* V_{cc} = +5.0 V	*Schottky low-power TTL (74LS)* V_{cc} = +5.0 V	*CMOS (4000A)* V_{cc} +5.0 V	+10.0 V	*CMOS (4000B)* V_{cc} +5.0 V	+10.0 V	+15.0 V	*CMOS (74C00)* V_{cc} +5.0 V	+10.0 V	+15.0 V
minimum HIGH input voltage	2.0 V	2.0 V	2.0 V	3.5 V	7.0 V	4.0 V	8.0 V	12.5 V	3.5 V	8.0 V	—
maximum HIGH input current	40 μA	20 μA	20 μA	—	—	—	—	1.0 μA	—	—	1.0 μA
maximum LOW input voltage	0.8 V	0.8 V	0.8 V	1.5 V	3.0 V	1.0 V	2.0 V	2.5 V	1.5 V	2.0 V	—
maximum LOW input current	1.6 mA	400 μA	400 μA	—	—	—	—	1.0 μA	—	—	1.0 μA
minimum HIGH output voltage	2.4 V	2.4 V	2.5 V	4.95 V (no load)	9.95 V (no load)	4.95 V (no load)	9.95 V (no load)	14.95 V (no load)	2.4 V (360 μA)	9.0 V (10 μA)	—
maximum HIGH output current	800 μA	400 μA	400 μA	300 μA (2.5 V)	250 μA (9.5 V)	510 μA (0.4 V)	1.3 mA (9.5 V)	3.4 mA (13.5 V)	1.75 mA (0 V)	8.0 mA (0 V)	—
maximum LOW output voltage	0.4 V	0.3 V	0.5 V	0.05 V (no load)	0.05 V (no load)	0.05 V (no load)	0.05 V (no load)	0.05 V (no load)	0.4 V (360 μA)	1.0 V (360 μA)	—
maximum LOW output current	16 mA	4.0 mA	8.0 mA	300 μA (0.4 V)	600 μA (0.5 V)	500 μA (0.4 V)	1.3 mA (0.5 V)	3.4 mA (1.5 V)	1.75 mA (5 V)	8.0 mA (10 V)	—

guaranteed to be 2.4 V or greater, and the maximum LOW output voltage is specified as 0.4 V. Both of these values are 0.4 V *inside* the respective input ranges. This "safety factor" is often called the *noise margin*. An output signal can have 0.4 V of unwanted variation, or "noise," superimposed on it and still be within the input limits.

The remaining 7400-series values in Table 1 are the input and output currents. In the LOW state, a 7400-series gate output can pass a maximum current of 16 mA while maintaining an output voltage of 0.4 V or less. An output may pass more than 16 mA, but the voltage could rise above the specified 0.4-V maximum. In the LOW state, current flows *into* the output pin. This is often termed current "sinking." Thus, we can say that the output can sink 16 mA in the LOW state (conventional current, not electron flow). When the output is HIGH, it can *source* (current flows out of the gate) only 800 μA (0.8 mA) of current.

Now turning to the gate input, we see in Table 1 that the maximum HIGH input current is 40 μA. This is the maximum current that can flow when the input voltage is in the HIGH region (2.0 to 5.0 V). This current flows into the gate, or in other words, the input sinks current in the HIGH state. If the input voltage is in the LOW region (0 to 0.8 V), the maximum current that can flow out of (the input sources current in the LOW state) the input is 1.6 mA.

What's the importance of the input and output current specifications? The answer to that question is *fanout*. In all but the simplest circuits, the output of a gate will be connected to several inputs. Fanout is the number of inputs that can be driven from a single output. Since each input can source 1.6 mA in the LOW state and an output can sink a maximum of 16 mA, the fanout is 16 ÷ 1.6, or 10, in the LOW state. Similarly, the HIGH state fanout is 800 μA per output ÷ 40 μA per input, or 20. Because the LOW state is the limiting

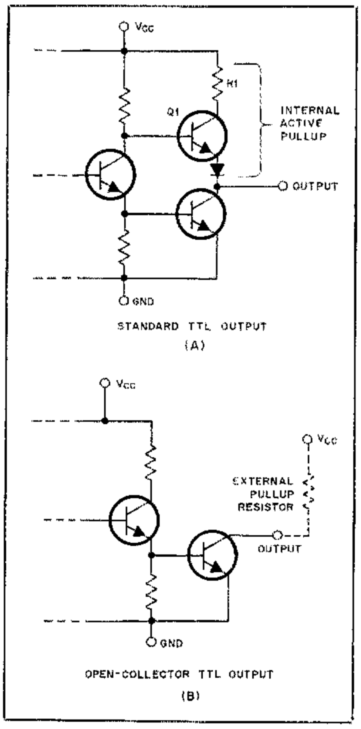


Fig. 6 — Shown at A is the output circuit of a standard 7400-series gate. Q1 and R1 form the internal active pull-up circuit that supplies current to the output pin. Open-collector gates (B) do not have an internal pull-up. In most circuits an external resistor is used to "pull up" the output to V_{CC} when it is in the HIGH state.

condition, the gate fanout is said to be 10. Understanding these input and output specifications is also important when we connect logic ICs of one type to ICs of another type or to other kinds of circuits.

The standard 7400 series is not the only TTL family of devices in use. Other varieties, in which particular characteristics have been optimized, are also available. For example, there is a 74L00 series. The L stands for low power. These devices perform the same function as the 7400 counterparts, but they consume less power. The penalty you pay for this reduced power consumption is lower operational speed. There are several other 7400 variations, but the most popular is the 74LS00 series. These devices offer high-speed operation *and* lower power consumption. Many microcomputer devices are designed to be compatible with the 74LS00 series. The input and output specifications for this TTL family are shown in Table 1.

Open Collectors and Wired ORs

A special type of output circuit is used in certain TTL devices. It is called the "open-collector" output. A "normal" TTL output is shown in Fig. 6A: An internal active "pull-up" circuit, formed by a resistor and a transistor, is used to connect the output pin to the supply when the output is in the HIGH state. An open-collector output stage (B) has no internal pull-up; it is simply an npn transistor with the collector connected directly to the output pin and the emitter internally connected to ground. In the LOW state, the transistor is turned on, providing a low-resistance path from the output pin to ground. The transistor is turned off in the HIGH state. An external pull-up resistor connected between the output and the supply provides a high-voltage level in this state.

The value of an open-collector output lies in the fact that two or more of these outputs can be connected together — something you *cannot* do with other TTL outputs. This type of circuit, often called a "wired-OR," is shown in Fig. 7A. Two truth tables are shown for this circuit: At B, ones and zeros are used to represent

HIGH (1) and LOW (0) levels (in the same manner used in the earlier truth tables). We can see from this truth table that this circuit produces the AND logic function. Why then is this circuit called a wired-OR? To answer that question we must take an "upside-down" look at true and false logic states.

Negative and Active-Low Logic

So far we have assumed that the HIGH level (a high voltage) was the true state, and we have been using a binary one to represent that state. This seems reasonable, but remember that these assignments are arbitrary. We could just as well call the LOW level (0 V) true and the HIGH level (+5 V) false. This is just what is done in many logic circuits. In the second truth table (Fig. 7C), true and false logic states, rather than ones and zeros, are used. We've chosen to let the true state be a LOW level and the HIGH level be the false state. Using this convention, we see that the circuit *does* produce an OR logic function: Input A *or* B must be true to cause the output to be true.

Sometimes, this logic convention is called "negative" or active-low logic. Meaning simply that the low-voltage level is the state that causes something to happen. Fig. 8 is a practical example of a circuit in which active-low logic is used. In order for the relay to close, current must flow from the +5 V supply, through the coil and into the gate output to ground. This is what happens when the output is in the LOW state. If the output is HIGH (+5 V), no current flows through the relay coil (no potential difference across the coil). The truth table shows that inputs A and B must be LOW in order for the relay to be activated. They are active-low inputs. Because the output must be LOW before the relay will close, we call it active-low, also. To help make schematic diagrams more understandable, signal lines are often labeled ($\overline{\text{RELAY}}$ in our example). A bar placed over the label indicates that the signal is active-low. In addition to having many uses in conventional logic circuits, active-low signals are widely used in microcomputer memory and interface ICs.

CMOS

The various TTL families (standard, low-power, etc.) are basically similar, differing only in specific characteristics. The next type of digital IC we're going to examine, CMOS (complimentary-symmetry metal-oxide semiconductor), is much different. To begin with, CMOS devices are constructed from field-effect transistors (FETs) rather than the bipolar transistors used in TTL devices. The result is that CMOS ICs consume far less power than even the 74L00 series ICs. CMOS ICs are also much more tolerant of supply-voltage variations. TTL requires a fairly closely regulated 5-V supply, while CMOS will operate with supplies of about 3 to 18 V. This characteristic, coupled with the low-power requirement, makes CMOS ideal for battery-powered circuits.

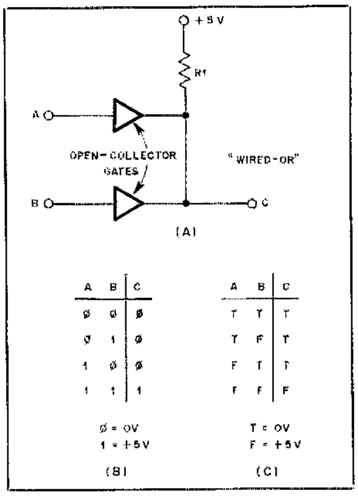


Fig. 7 — An example of a circuit using open-collector gates (A). The gates used here are buffers (the output state is the same as the input state). Other types of gates with open-collector outputs could be used in their place.

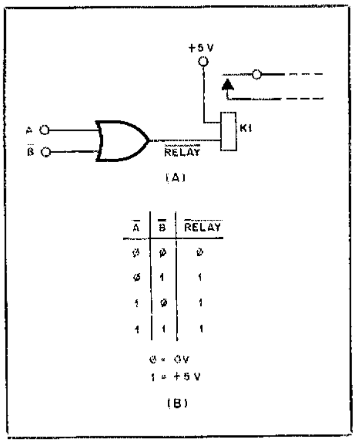


Fig. 8 — A practical circuit example using active-low signals.

As you might expect, CMOS ICs do have certain limitations. Compared to TTL, CMOS is slow. If a particular TTL device will operate at 15 MHz, a similar (in function) CMOS IC may operate to only, say, 4 MHz. In applications where high speed is not required, the advantages of CMOS make it more attractive than TTL.

While we're comparing CMOS and TTL, another characteristic should be mentioned — static sensitivity. TTL is relatively immune to static damage, but CMOS ICs can be damaged easily by static discharges. For this reason, CMOS devices are normally stored in conductive foam or other static preventive material. You should exercise some care when handling CMOS. Simply touching a grounded surface with the conductive foam and your hand before removing the IC from the foam is generally all that is necessary to prevent damage to the device.

There are two other precautions you should take when working with CMOS: Always disconnect the supply voltage before removing CMOS ICs from, or inserting them into, a circuit, and *never* make or break a connection to an IC with the power applied. If you follow these rules, you'll find CMOS to be easy to use.

There are three families of CMOS ICs commonly available. Two of these, the 4000A and the 4000B series, are basically the same, except that the B series has additional output buffer stages. The buffers allow the outputs to operate at higher current levels. There is also a 74C00 series of CMOS ICs. These devices are *functionally* similar to the equivalent TTL device. The 74C02 CMOS IC and the 7402 TTL IC, for example, are both quad two-input NOR gates. While both devices perform the same logic function and have the same pinout, they cannot be interchanged in most circuits.

CMOS Electrical Characteristics

Table 1 contains some of the electrical specifications for the three CMOS families we have mentioned. Because CMOS logic can be operated from a wide range of power-supply voltages, three sets of specifications are listed for each device type. Each set corresponds to a different power-supply voltage (V_{DD}). At a supply voltage of 15 V, the minimum HIGH input voltage (for the B series) is 12.5 V; when a 5-V supply is used, the HIGH input is 4 V. Most of the other specifications also change with supply voltage. In Amateur Radio applications, it is often convenient to power logic circuits from a 12-V supply. The specifications listed for V_{DD} = 10 V in Table 1 can be used for circuits operating at 12 V.

CMOS IC input and output currents are much smaller than those of TTL devices. The maximum input current is less than 1 μA, compared to 1.6 mA for standard TTL! If a 5-V supply is used, a B-series output can pass approximately 500 μA. In the HIGH state, more current can be drawn from the output, but the voltage will drop: At 1.6 mA, the output voltage falls to 2.5 V.

Looking at the input and output current specifications, you might conclude that the fanout of CMOS is very high. In most cases this is correct, but when you are working with circuits that operate at high speeds, another factor enters the picture. Each CMOS input has a capacitance of 7.5 pF. This load capacitance must be charged by the output before the voltage can reach a HIGH level. The larger the load capacitance, the longer it takes to charge it (a series R-C circuit). If you connect 10 inputs to an output, the total load capacitance becomes 75 pF. This increases the length of time required for the output to change from one state of another and limits the operating speed of the circuit. This isn't a problem very often, so we needn't be overly concerned about CMOS fanout capabilities.

Experimenting with Logic ICs

We mentioned earlier that experimenting with logic circuits can be very enjoyable. That's true partly because it's so easy to do. Most of the time you don't even need to heat up a soldering iron! If you use solderless "breadboards," you can wire a circuit simply by plugging in the ICs and making interconnections with no. 26 solid insulated wire.

A typical logic circuit, wired on an AP Products Super-Strip, is shown in the title page photograph.[1] TTL ICs are used in this example, and a homemade 5-V supply powers the circuit. LEDs (light-emitting diodes) and a simple voltmeter serve as output state indicators. Low-cost logic probes, used to determine input and output logic levels, are also available. If you want to experiment with CMOS devices, the same 5-V supply used with TTL circuits can be used to power your CMOS circuitry. A 12-V supply or a 9-V transistor radio battery also make good power sources for CMOS projects.

Finding the necessary components is often difficult for the builder and experimenter, but it shouldn't be a problem when you're working with logic projects. Dozens of mail-order suppliers carry complete stocks of 7400 and 74LS00 series TTL and CMOS devices.[2] Radio Shack stores carry many popular logic ICs and most of the other components the experimenter will need.

Once you have begun experimenting, it's logical that you'll want to increase your knowledge of digital circuits. The *ARRL Handbook* contains an excellent section on digital logic that should help you to enjoy your experimenting even more![3]

Notes

[1]AP Products, Inc., Painesville, OH 44077. The Super-Strip (Model SS-2) is available from Priority-One Electronics (see note 2).

[2]The following suppliers sell a wide variety of electronic components by mail: Jameco Electronics, 1355 Shoreway Rd., Belmont, CA 94002; Priority-One Electronics, 9161 Deering Ave., Chatsworth, CA 91311; Mouser Electronics, 11433 Woodside Ave., Santee, CA 92071.

[3]G. Woodward, ed., *The Radio Amateur's Handbook*, 60th ed. (Newington: ARRL, 1982), pp. 4-48 to 4-61.

[4]This data represents typical IC family characteristics compiled from various sources. Specific device characteristics may differ from those listed.

References

Motorola Semiconductor Data Library/CMOS, Vol. 5, Series B. Phoenix, Arizona: Motorola, Inc., 1976.

TTL Data Book. Mountain View, California: Fairchild Camera and Instrument Corp., 1978.

Strays

FLASH! SHUTTLE FLIGHT WITH HAND-HELD LOOKS PROMISING

☐ As this issue was going to press, we received unofficial word that Astronaut Owen Garriott, W5LFL, will be authorized to carry a specially designed 2-meter hand-held transceiver aboard the STS-9 Space Shuttle flight, scheduled for September. Details will follow.

AFCEA LUNCHEON THIS MONTH

☐ The Washington Chapter of the Armed Forces Communications and Electronics Association (AFCEA) will hold its annual luncheon on May 13 at the Shoreham Hotel, Washington, DC. The guest speaker will be Dr. Robert S. Cooper, director of the Defense Advanced Research Projects Agency. For more information and reservations, call Diana Sibley at 202-457-3060.

MOVING? CHANGING CALL?

☐ When you change your address or call sign, be sure to notify the Circulation Department at ARRL Hq. Enclose a recent address label from a *QST* wrapper if at all possible. Address your letter to Circulation Department, ARRL, 225 Main St., Newington, CT 06111. Please allow six weeks for the change to take effect. Once we have the information, we'll make sure your records are kept up-to-date so you'll be sure to receive *QST* without interruption. If you're writing to Hq. about something else, please use a separate piece of paper for each request.

Ralph Gibbons, W7KV (left), of Seattle, Washington, receives his ARRL 50-year plaque from Director Emeritus Robert B. Thurston, W7PGY. W7KV is a past director of the Northwestern Division, when Hiram Percy Maxim, W1AW, was ARRL President.

WANTED: INFO ON MODIFYING HF EQUIPMENT

☐ *QST* would welcome current information on modifying hf transmitting and receiving equipment for 30-meter operation. We would be most interested in successful modifications to the more popular transceivers. These need not be limited to full-length articles. If manufacturers, their distributors or individuals are offering modification kits or factory modifications, we would appreciate knowing about them. Please direct your related correspondence to the *QST* Senior Technical Editor, ARRL Hq.

When Karl Stephan, KD5DC, of Austin, Texas, checked out his homemade 432-MHz helical antenna last spring, he found an unexpected guest had set up housekeeping on the matching network. The guest has since left, so KD5DC can get ready for another "bird" — the AMSAT-Phase IIIB satellite.

Recognize this equipment? It's a World War II Canadian Army Model 19 set — and it still works! Dick Moores, VE3LRB, of London, Ontario, warms it up once each year, on Remembrance Day, November 11. *(VE3GRO photo)*

Product Review

Conducted By Paul K. Pagel,* N1FB

ICOM IC-290H All-Mode 2-Meter Transceiver

Table 1

ICOM IC-290H Multi-Mode 2-Meter Transceiver, Serial No. 02128

Manufacturer's Specifications	*Measured in ARRL Lab*
Frequency coverage: 143.8000 to 148.1999 MHz.	As specified.
Modes of operation: Fm, usb, lsb, cw.	As specified.
Transmitter power output (HI): 25	30
(LO): 1	2
Frequency readout: Digital, 5-digit green LED display, 100-Hz resolution.	As specified.
Frequency resolution: Ssb: 100-Hz steps; fm: 5-kHz steps (1 kHz selectable).	As specified.
kHz/turn of knob: 5 kHz (250 kHz in fm; 50 kHz selectable).	As specified.
RIT range: ±800 Hz.	As specified.
S-meter sensitivity: Not specified.	S1, 0.95 μV; S9, 3 μV; S9 + 20 dB, 6 μV.
Harmonic suppression: 60 dB below peak power output.	62 dB below peak power output.
Current drain at 13.8-V dc: Approx. 6.0 A (max.).	5.4 A (HI), 2.0 A (LO)
Receiver sensitivity: Fm — more than 30 dB S + N + D/N + D at 1 μV. Less than 0.6 μV for 20 dB noise quieting. Ssb, cw — less than 0.5 μV for 10 dB S + N/N.	Receiver dynamics: Noise floor (MDS): −138 dBm. IMD DR: 85.5 dB. Blocking: Noise limited. Third-order intercept: −9.75 dB. Receiver quieting: 0.26 μV/20 dB
Size (HWD): 2.5 × 6.7 × 8.6 in. (64 × 170 × 218 mm).	As specified.
Weight: 5.5 lb (2.5 kg).	As specified.

The first reaction to this radio has to be, "Boy, they're sure packing a mountain of features into a molehill of gear these days!" At first glance, the '290H, 25-W successor to the 10-W '290A, could be mistaken for an fm-only box; it's the same size and weight as 2-meter fm gear produced only a few years ago. A second glance at the knobs and buttons on the front panel makes it clear that you're looking at a state-of-the-art multimode transceiver.

Well designed for ease of operation and aesthetics, the front panel has no less than 16 knobs, switches and buttons (others are hidden under the top cover). Each control is marked clearly, making mobile operation, even at night, easy to master. The bright, frequency readout LEDs go to four decimal places (100s of hertz for ssb and cw). Other LEDs indicate that a signal is being received or sent, that a priority frequency (more on that later) is in use, and that the radio is in the DUPLEX mode. An S/RF indicator shows S-units and relative output power in decibels using a row of seven LEDs.

Features Galore

It's no longer necessary to ask if a radio like this is synthesized, whether it scans or has memory capability. The appropriate questions are: Is it easy to tune? How many scanning functions and memories does it have? Does it have two VFOs for added versatility? How about RIT? A noise blanker? Automatic gain control? Standard tone pad for repeater autopatch use?

The answers: for the most part, four and five, yes, yes, yes, yes, yes. In fact, you'd be hard-pressed to think of a feature it doesn't have, and to find a control that isn't convenient to find and operate. To provide an idea of what this radio is designed to do, let's take a look at the scanning capabilities.

Pressing the S/S (Scan/Stop) button initiates the SCAN mode, indicated by two flashing decimal points in the frequency display. After programming the five memory frequencies, you can scan them (plus the two VFO frequencies) by pressing the button labeled M-R. Or, if you're anxious to find a QSO, you can scan the entire frequency range — all 4.4 MHz of it. Or, you can select the range to cover, for example, only the ssb subband. Want to have the radio stop at any *empty* frequency? Flick the BUSY/EMPTY switch under the top cover.

One scanning feature in particular caught my eye. In my review of another manufacturer's fm-only transceiver a couple of years ago, I lamented the lack of a scanning delay; whenever a busy frequency was reached, the scanner would pause, but then travel on its merry way the instant the carrier dropped out. I found this annoying, and said so in the review. ICOM must have been listening, as the IC-290H not only has a scan/stop function, but pause time can be adjusted (by means of a control under the top cover) from 5 to 20 seconds. Problem solved! Scanning speed is also adjustable. A convenient 1 kHz button allows scanning (and tuning) in 1-kHz increments rather than the standard 5-kHz (fm) or 100-Hz (ssb/cw) steps.

The IC-290H sports several other features that set it apart from run-of-the-mill 2-meter transceivers. RIT (receiver incremental tuning) allows you to raise or lower the receive frequency by as much as 800 Hz. This is especially handy for satellite work and for keeping drifting stations within earshot. By pressing the AGC button, you can improve the reception of stations that tend to fade quickly. A noise-blanker (NB) switch activates a circuit that reduces pulse-type noise such as that generated by automobile ignitions. Have a favorite frequency? You can monitor it while enjoying a QSO by pushing the PRIORITY button, which sends the receive frequency to your chosen frequency for an instant every five seconds. The rear panel contains jacks for an external speaker and a key, as well as an accessory socket, power socket and antenna connector.

A second VFO was at one time a luxury reserved only for hf transceivers in the $1000-and-up price range. No longer. The two VFOs in the IC-290 make it simple to find a clear frequency. If you're in the midst of heavy

*Assistant Technical Editor

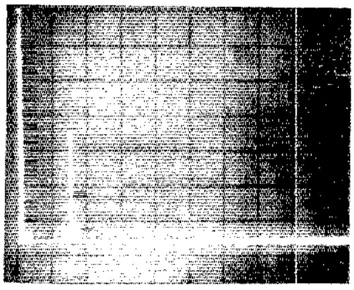


Fig. 1 — Spectral display of the IC-290H. Vertical divisions are each 10 dB; horizontal divisions are each 100 MHz. Output power is approximately 30 watts at a frequency of 146.010 MHz. The fundamental has been reduced in amplitude approximately 34 dB by means of notch cavities; this prevents analyzer overload. All spurious emissions are at least 62 dB below peak fundamental output. The IC-290H complies with current FCC specifications for spectral purity.

QRM and want to find another frequency, just change VFOs, search for a clear frequency, and change back. The frequency you left on one VFO is memorized when the VFO is switched. A second VFO is also useful if you'd like to "keep an ear on" a particular frequency while working stations elsewhere on the band.

Another feature worth mentioning is the nonstandard offset selection. For "oddball" repeater splits, the offset can be changed from ± 600 kHz quickly and easily.

Despite the initial challenge of learning the functions and operations of the various controls, the IC-290H is a breeze to use. Right after unpacking the box, I hooked it up to a power supply, plugged the supply into the wall socket, connected an antenna, set the MODE switch to USB, and proceeded to run a slew of nearby stations. I then realized that I had inadvertently chosen to break in the radio during the height of an ARRL VHF QSO Party!

It was a good omen. The ssb subband hasn't seen that level of activity since, but I've made my share of contacts through various conditions and have had no trouble getting into area repeaters. The receiver sensitivity seems excellent, as one would expect from the manufacturer of the IC-2A.

Technical Description

Frequency control is accomplished with a microprocessor-based, 100-Hz-step digital PLL synthesizer. The circuit serves as the local oscillator for both transmit and receive with output approximately 10.75 MHz below the receive frequency. The operating frequency is controlled by pulse signals generated by the rotary encoder at the tuning knob being added to or subtracted from the preset frequencies in the μP. The μP controls the PLL circuit, which in turn determines the output frequency of the VCO.

Incoming signals are mixed with the LO output from the PLL circuit. In the ssb and cw modes, the circuits function as a single-conversion type with a 10.75-MHz i-f; in fm, they function as a dual-conversion type with an additional 455-kHz i-f.

The transmitter carrier frequency is 10.7485 MHz for usb and 10.7510 MHz for lsb. Carrier

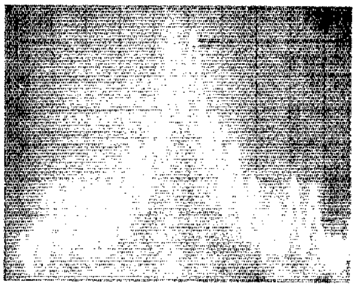


Fig. 2 — Spectral display of the IC-290H output during transmitter two-tone IMD test. Third-order products are 30 dB below PEP and fifth-order products are 40 dB down. Vertical divisions are each 10 dB; horizontal divisions are each 1 kHz. The transceiver was being operated at rated input power on the 2-meter band.

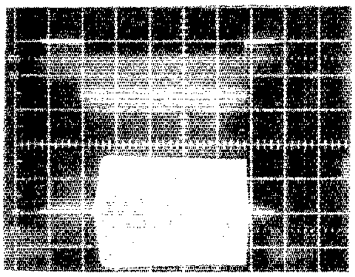


Fig. 3 — Cw keying waveform of the IC-290H. Upper trace is the actual key closure; lower trace is the rf envelope. Each horizontal division is 5 ms. This keyed wave has a rapid fall time.

and voice signals are sent to a balanced modulator, where the dsb suppressed-carrier signal is generated. The unwanted sideband is removed by a crystal filter, and an ssb signal of 10.75 MHz results.

In the cw mode, the usb carrier is shifted up about 800 Hz and is then fed to the transmit mixer. In the fm mode, it uses another crystal oscillator to produce the 10.75-MHz signals that are direct-frequency modulated. The ssb, cw or fm signal is mixed with the LO output from the PLL circuit, and is then amplified, filtered and sent to the antenna.

Manual

Covering 44 pages, the manual is clear, thorough and well-written. There are typographical errors, to be sure, but they are rare. Diagrams liberally sprinkled throughout show how to mount the '290H in a car, an exploded view of the mike plug, the operation of the various scanning functions, AMSAT-OSCAR 8 frequencies and basic operation, the location of key components under the top and bottom covers, and block diagrams of the rf, scan control, main and PLL circuits. In addition, there is a troubleshooting chart, individual descriptions of the various circuits, and photos and identifications of available options, such as ac power supply, scanning mike, mobile speaker and condenser mike.

Separate from the manual are a fold-up, two-color, board-layout diagram and a complete schematic, which should make it easy to identify errant components or to make modifications.

Criticisms

While it has many different features designed to make operating versatile and enjoyable, the IC-290H lacks a couple of circuits that would have made the radio complete. There's no quick and easy way, for example, to decide on the spur of the moment to head for a distant part of the band. If you're listening to a W1AW bulletin on 147.555, for example, and hear that the Perseids are due at any time, you'll want to try to work some meteor scatter at the low end. How do you get there in a hurry? If you haven't programmed an ssb frequency into a memory or a VFO, you can either hit the SCAN button and wait for the frequency to make its way down, or you can rotate the tuning knob about 13 times (in either direction, thanks to the continuous tune feature that prevents out-of-band operation). If you leave the MODE switch in the FM position, the frequency moves in 5-kHz increments (selectable to 1 kHz), so it doesn't take *too* long to find the ssb subband. However you decide to get to the low end of the band, it's a minor inconvenience. A MHz control switch would have been a nice touch. In addition, there is no provision for VOX, and there is no mike-gain control.

On the plus side, the '290H is better suited than its predecessor for working through the amateur satellites, as it tunes in either direction when in the transmit mode. (The '290A tuned only upward in frequency, making it nearly impossible to send a string of dots to tune in your signal through the satellite.)

Two other changes from the '290A are worth mentioning: Bright-green frequency-readout LEDs replace the red ones, making them easier to read in sunlight, and a yellow "caution" sheet warns of a potential readout problem. It explains that the program that initializes the operating condition of the transceiver may malfunction because of an intermittent connection of the power plug or the power switch being turned off and on too quickly. Although the sheet says it's "not an equipment malfunction," I'm not sure how else it can be described. The problem never arose in the review model, however, and it probably doesn't happen too often. In case it does, the sheet gives a procedure that should solve the problem.

The ICOM IC-290H is distributed in the U.S. by ICOM America, Inc., 2112 116 Ave., N.E., Bellevue, WA 98004. Price class: $550. *— Joel P. Kleinman, N1BKE*

HAL COMMUNICATIONS CWR-6850 TELEREADER® RTTY/CW TERMINAL

☐ While searching for RTTY equipment to use on our St. Paul Island DXpedition,[1] I came across the advertisements for the HAL CWR-6850. It looked like it would fit our needs perfectly. We arranged to have one shipped to Hq. for Product Review and use during the DXpedition.

The '6850 Telereader is a portable RTTY/cw terminal that can send and receive the ASCII and Baudot teleprinter codes as well as international Morse code, at the speeds summarized in Table 2. In the unit are tone generators and demodulators for both the high and low afsk tones. Provision for fsk is also made. Receive

[1]"Canadian NewsFronts," *QST*, November 1982.

and transmit data are displayed on a green, 5-inch CRT. Interface circuitry for driving an external video monitor and a Centronics-compatible printer is included. The entire package weighs just 16.5 pounds,[2] and operates from a 13.8-V dc power source.

Features

It would take many pages to describe every feature of the Telereader. I will highlight the basic modes and features.

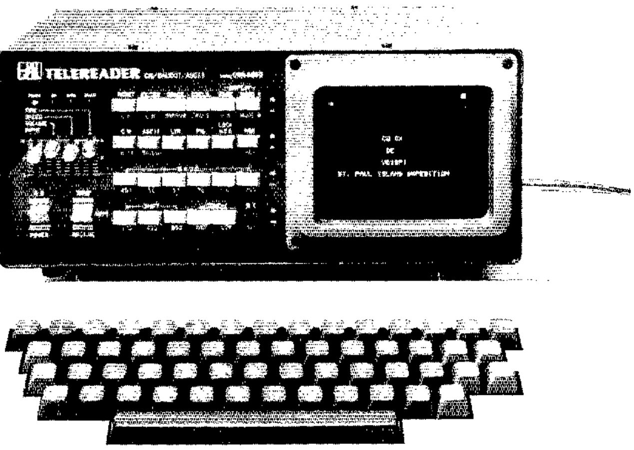
Display: In any of the three modes of operation, four "pages" of data can be displayed on the CRT. A page consists of character data in a 32-character-per-line, 20-line-per-page format. The pages are numbered Ø through 3. Page Ø displays the first 19 lines of the receive buffer (608 characters). Page 1 displays 15 lines of the receive buffer, the status line, tape-storage-buffer line and three lines of the transmit buffer. Page 2 is the reverse of page 1 — it displays the entire transmit buffer (480 characters), three lines of the receive buffer and the status/tape-storage-buffer lines. Page 3 displays the six HERE IS messages, three lines each of the transmit and receive buffers, and the status and tape-buffer lines. The status line indicates the page in use and whatever special functions are active. When using an external (optional) tape recorder to store or send messages, the tape-storage-buffer line displays the data going to (from) the tape. The page selected for viewing can be controlled from the front panel or the keyboard.

Keyboard: The keyboard is a stand-alone, ASCII-encoded unit that connects to the main "box" through a multiconductor cable. The keytops are gray plastic with white lettering, and the "feel" of the keys is quite light. Several useful functions can be activated from the keyboard by use of the CTRL key. A list of a few of these functions is shown in Table 3.

Front Panel: There are *30* front-panel controls on the '6850, four of which are slide potentiometers. The INPUT and VOLUME controls set the audio input level and internal-audio-oscillator level, respectively. Morse code speed is set by the SPEED control. The center frequency of the RTTY demodulators can be altered slightly by the FINE control. Two large lever switches control the main POWER and the transmit/receive condition (RECEIVE/TRANSMIT/AUTO). The remaining 24 switches select the display PAGE, the operating mode (MORSE/RTTY, ASCII/BAUDOT); the RTTY BAUD rate (45, 50, 57, 75, 110 or 300) and the RTTY TONE pairs (HIGH or LOW). Other function switches control the unshift-on-space (UOS) function, NORmal or REVerse RTTY receive sense, the cw receiving program (SPACE/NARROW) and the cw receive-FILTER mode.

Rear Panel: The rear panel is filled with jacks, mostly phono types which are used for input/ouput interfacing with a transceiver. Other types of connectors are used for the keyboard, printer, external video monitor, cw key (1/4-inch phone jack) and the dc power connector (polarized).

Firmware: Much of today's communications equipment uses microprocessors, and the CWR-6850 is no exception. The program controlling the '6850 operation is contained in several ROMs. It is quite versatile — for example, the section used to decode international Morse has a special mode which enables the unit to copy poor fists (QLF?). Another feature



is the ability to copy any mode while sending ASCII data to the printer.

On-the-Air Operation

Initial checkout of the Telereader at my home station indicated no defects. The unit was then repacked in a watertight container for our trip to St. Paul Island.

Our DXpedition RTTY setup consisted of the CWR-6850, a Kenwood TS-520S hf transceiver and a Cushcraft A3 triband Yagi. After our first RTTY CQ, a large pileup ensued. If that wasn't a test for the RTTY filters, nothing was! The unit performed admirably. Although many stations were calling, I had no trouble in "printing" the caller's call signs on the screen. During our entire operation, we had no troubles with the unit. I was amazed of the ability of the '6850 to operate perfectly with widely varying supply voltages; several times the supply voltage dropped to 11 volts and the only ill effect was a slight shrinking of the CRT video.

Back Home

After the trip home, the Telereader was put through the paces at several Hq. employees' ham shacks. Almost all reports were positive. The RTTY filter/demodulator is not as good a performer when copying weak signals, as is true of some other systems. Also, many RTTY terminals provide a "diddle" function that sends the ASCII NUL code or Baudot LTRS code when no character data is present in the transmit buffer. (This function is useful because it keeps the receiving station printer in sync.) On other RTTY systems, the "diddle" function (when enabled) comes on as soon as the operator initiates transmission. The "diddle" then fills in gaps between groups of data. With the '6850 (operating in the CONTINUOUS or "letter" mode) the "diddle" must be turned on and off independently of the status of the transmit line, and it must be turned off before any character data can be sent! In the WORD mode (data transmission being enabled by the detection of a space followed a word) the diddle works as would be expected.

The 32-character-per-line format is somewhat small but necessary when using the built-in 5-inch display. When using an external video monitor, however, 40 or 64 characters is to be preferred.

Table 2

Speeds Available on Telereader Terminal

Code	*Speeds Available*
Morse	4-33 wpm transmit (can be modified); 1-100 wpm receive
Baudot	45.45, 50, 56.88, 74.2, 110 and 300-baud transmit; up to 110-baud receive.
ASCII	Same as Baudot.

Table 3

Keyboard Functions

CTRL:	*Function*
A	On-off control of automatic send-receive (KOS)
D	Extends cw transmit dash length by 25%, increasing the cw weighting.
E	On-off control of the echo from transmit to receive buffers.
I	Changes transmit mode from RTTY to cw, and vice versa.
N	Turns the "diddle" on and off.
P	Turns the printer on and off.
Q	Sets the display page.
U	Slows the rate at which data comes out of the transmit buffer. The effective baud rate is 25 regardless of the speed selected.

Conclusions

What an amazing terminal! It is hard to imagine how they pack all those features into such a small package. The versatility of the Telereader makes it ideal for both portable and fixed-station use. It is unfortunate that a little more design effort was not placed in the display and software. Overall, I would rate the CWR-6850 as an excellent, well-built product worth the asking price. It is one of the first imported products offered by HAL. For more information, contact HAL Communications, Box 365, Urbana, IL 61801. Price class of the CWR-6850 Telereader is $1000. — *Gerald B. Hull, VE1CER/W1*

TOKYO HY-POWER LABS HC-200 TRANSMATCH

☐ If you operate with medium power, go out for portable operation or just need a Transmatch to use your 40-meter dipole on the 30-meter band, consider the HC-200. Size and weight make it a viable candidate for mobile operation as well.

Muted green lettering stands in pleasant

[2]kg = lb × 0.454

Table 4

Tokyo Hy-Power Labs HC-200 Transmatch, Serial no. 829113

Manufacturer's Claimed Specifications	*Measured in ARRL Lab*
Frequencies: 80-10 m (8 bands)	
Power-handling capability: 200-W PEP (100-W cw) at 50 ohms	As specified.
Input impedance: 50 ohms (at matched condition)	As specified.
Output impedance: 10-250 ohms unbalanced	
Meter ranges: 20, 200 watts (forward) and SWR	As specified.
Connectors: Three SO-239 and one single-wire terminal	As specified.
Weight: approx. 4.84 lb†	
Size: 3.3 × 8.26 × 7.36 in. (HWD)	

†kg = lb × 0.454; mm = in. × 25.4

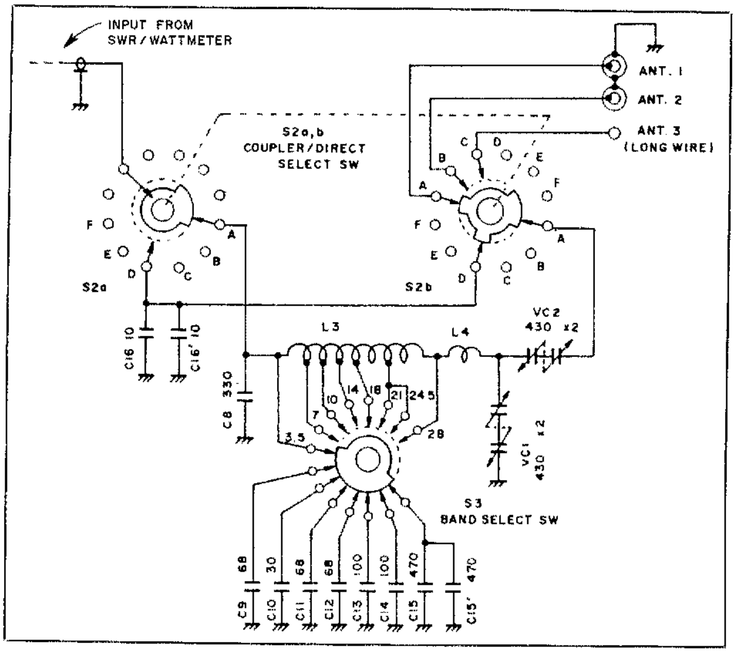


Fig. 5 — HC-200 matching-network schematic diagram copied from the operating manual. Metering section is not shown.

Table 5

Comparison of HC-200 and Bird Model 43 SWR Readings (70 W)

		21 MHz		7 MHz	
R_{load}	*SWR*	*'43*	*'200*	*'43*	*'200*
112.5	2.25	2.2	2.4	2.2	2.1
25	2.0	2.1	1.4	1.9	1.6

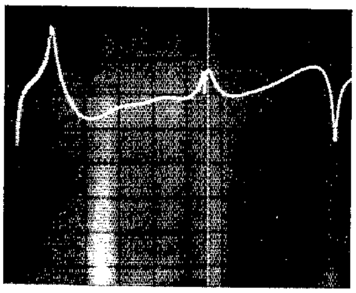


Fig. 6 — Network frequency-response curve. Horizontal divisions are 20 MHz; vertical divisions are 10 dB. The network is adjusted for 50 ohms input and output at 21.050 MHz.

Table 6

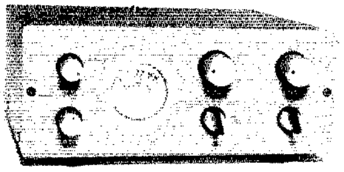
Comparison of HC-200 and Bird Model 43 Power Readings (21 MHz)

'43	*'200*
200-W range	
100	90
60	60
40	42
20	22
20-W range	
20	19
10	9
5	4

relief to the gray of the extra-thick front panel. Bright aluminum knobs, two with black plastic inserts, a round meter and a black cover combine to give the unit eye appeal.

All eight amateur bands from 3.5 to 30 MHz, including WARC bands, are covered by the '200. A power-handling capability of 200-W PEP (100-W cw) is well suited to most modern transceivers. I tried the '200 with a variety of loads on several different bands at home and in the ARRL lab. The results were satisfying — particularly when using my 40-meter dipole in the 30-meter band. During the DX contest, the '200 saved the day by allowing a solid-state transceiver to drive the untuned input of a kilowatt amplifier.

Perhaps the most striking visual feature of the '200 is the round meter. Control of metering functions is by means of METER SELECT and SWR CAL knobs. Those functions include SWR (FWD and REV), 200 W and 20 W (forward power). Separate SWR calibration scales for 200-W and 20-W power levels is an excellent feature. Diodes are notoriously nonlinear, particularly at lower power levels. I compared meter readings from the '200 with readings taken on a Bird model 43 wattmeter. The results are shown in Tables 5 and 6. The metering circuit may not qualify as a lab standard, but it is good enough for the average amateur.



Matching Network

The right-hand side of the front panel has four knobs that are used to control the matching network. The large TUNE and LOAD knobs look good and are easy to use. The two remaining control knobs, BAND and MODE have black plastic inserts that serve the dual function of making it easy to operate the associated switches and providing a positive indication of switch position.

Fig. 5 shows the schematic diagram of the matching circuit. You can see that it looks like a pi network with a series capacitor at the output. That series capacitor increases the matching range of the network.

I was curious as to what the response of the circuit was. The spectrum analyzer and tracking generator in the ARRL lab revealed an interesting situation. The results can be seen in Fig. 6. The matching network exhibits a bandpass response; frequencies above and below the desired band are attenuated. I like that!

The '200 is designed for use with unbalanced loads. There are three antenna connections on the rear panel: two SO-239 coaxial connectors and a single wire terminal. The MODE switch allows any of these three antennas to be selected, routing it either through the matching network or bypassing the matching section altogether. The K8CH antennas are resonant at my favorite portion of each band. Full coverage of some bands does require the use of a Transmatch. This sometimes-yes/sometimes-no situation makes me appreciate the ability to easily switch the matching network out of the line!

The rear panel also contains an SO-239 connector for IN/TX and a bolt and wingnut combination for a GND connection. I found the '200 easy to operate and the documentation complete. There is even a table of typical settings for operation into 50 ohms on each of the bands.

The '200 is available from Encomm, Inc., 2000 Ave. 'G', Suite 800, Plano, TX 75074. Price class is $100. — *Chuck Hutchinson, K8CH*

Hints and Kinks

Conducted By Larry D. Wolfgang,* WA3VIL

"MULTIPEDANCE" BROADBAND AUTOTRANSFORMER

☐ After reading Doug DeMaw's article on a multipedance broadband transformer, I decided this was a piece of equipment I needed.[1] I had plenty of no. 24 wire to wind the transformer, and did not want to buy heavier gauge wire for the project. Some method of providing higher current-carrying capacity was needed, at least for the lower-impedance taps.

Fig. 1 shows my solution. I used six parallel strands of wire to start, then decreased the number of strands as the impedance ratio increased. First, I cut six pieces of no. 24 wire to different lengths, as shown in Fig. 1A. The shortest pieces were long enough to make five turns on the T225A-2 toroidal core, spaced to fill the entire periphery. The fifth-turn tap point was placed alongside the starting point, as shown in Fig. 1B. All six wires connect to the tap, and three of them continue to the next tap. I kept the wires in a flat bundle, with each succeeding five-turn winding adjacent to the last.

To insulate the tap points, I placed the tapped turn in a fold in a piece of 10-mil-thick fish paper (used to insulate transformer windings). Turns 25 through 90 have only one strand of wire, with taps placed every fifth turn. Fig. 2 shows my completed autotransformer. There was just enough room on the core for five turns of fiberglass string between the starting and ending turns of the entire winding.

To finish the project, I sewed the folded ends of fish paper together and coated the entire unit with casting resin, as suggested by DeMaw. You could also place a primary winding over this one and use it as a conventional broadband transformer. — *Frank Thompson, WØOD, Baudette, Minnesota*

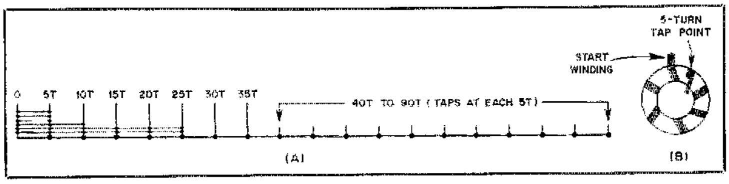


Fig. 1 — The wire layout used by WØOD for his multipedance autotransformer is shown at A. B shows how the first set of five turns was positioned on the core.

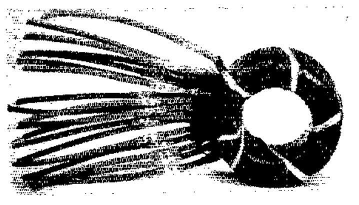


Fig. 2 — Photo of the completed transformer. Note the insulating paper between tap points.

AUXILIARY CRYSTAL SOCKET FOR THE DRAKE T-4X

☐ I recently purchased the crystals needed to operate my Drake T-4X and R-4A on 30 meters. Installing the new crystal in my receiver was simple because there are 10 auxiliary sockets, readily accessible from the rear of the chassis. The transmitter has only four auxiliary sockets, located so that the top cover must be removed each time a crystal is changed. I had already filled the four sockets with crystals covering the 160- and 10-meter bands. (Each crystal provides coverage of a 500-kHz segment.)

My solution to this problem was to install a socket on the right side of the T-4X chassis, so crystals can be changed without removing the cover. Installation of the new socket was facilitated by the fact that my rig had the required holes already drilled in the chassis. The older T-4 had a socket installed on the side, and apparently the same chassis was used for at least some of the T-4Xs. Later versions may not have the holes drilled, but it should not be too difficult to locate a suitable mounting position and drill the holes.

I moved the wire to the crystal-selector switch from socket number four to the front-panel crystal switch. I soldered this wire to one terminal of the new socket and used a short piece of hookup wire to connect the other terminal to the common side of the original sockets. Now the crystal-selector switch selects the new socket when it is in position number four. Changing crystals is now an easy task.

The owner's manuals specify the crystal frequencies for a desired operating range. A 20.8-MHz crystal is required to transmit on the 30-meter band. It is necessary to add 9.7 MHz to the dial reading to determine your operating frequency. This poses no problem because the 21.1-MHz receive crystal provides direct dial readings from 10.0 to 10.5 MHz. I calibrate my receiver to WWV and spot the desired transmit frequency on my receiver. — *Donald Stickle, K2OX, Lake Hopatcong, New Jersey*

JOHNSON VIKING II ON 10 MHz

☐ When the FCC announced the opening of the 10-MHz band for amateur operations, I realized that my four-year-old Kenwood TS-520S would not operate in this new band. I began to think about the possibility of using my old Hammarlund HQ-140X receiver and Johnson Viking II transmitter. I had kept both in operating condition, and even used them for an occasional QSO.

The HQ-140X tunes continuously from 0.54 to 30 MHz, and has a bandspread tuning dial with scales calibrated for the 80, 40, 20, 15 and 10-meter bands. I prepared a frequency chart so that I would know the actual frequency for any given bandspread dial reading. With the main tuning dial set on 10.25 MHz, 10.10 to 10.15 MHz lies between the 7.00- and 7.10-MHz markings on the bandspread dial.

According to the Viking II instruction manual, 10-MHz output can be obtained by using a 5-MHz crystal and setting the band switch to 20 meters. The oscillator, buffer and final stages are all tuned to 10 MHz. Since I wanted to use the Model 122 VFO, I would have to modify it to provide a 5.05- to 5.075-MHz output. This VFO has three ranges: 1.75 to 2.00 MHz, 7.00 to 7.425 MHz and 6.7 to 6.85 MHz. I do not use the last range, so that is the one I modified. A study of the wiring diagram indicated that the only change required was to add some additional capacitance across C56. I found a couple of air-dielectric padder capacitors that would provide 135 pF when wired in parallel. These were installed, and I found the VFO to cover the range 5.05 to 5.075 MHz with dial readings from 7.00 to 7.13 MHz on the original scale.

To ensure good keying characteristics, the Viking and VFO should have the Johnson Time Sequence Keyer modification. Otherwise, you may have a signal with chirps and clicks. — *Walt Bollinger, AF3V, Pittsburgh, Pennsylvania*

DETUNING SLEEVE FOR THE RINGO RANGER

☐ One problem experienced by many users of the original Cushcraft Ringo Ranger 2-meter antenna is the lack of any method to decouple the antenna from the transmission line. This results in the flow of antenna currents on the outside of the coaxial-cable feed line. Owners of the Ringo Ranger II report greatly improved operation because of the decoupling system incorporated with that antenna. Not wanting to replace my antenna, I found a way to retrofit a decoupling system to it.

The operating principle of my system is found in *The ARRL Antenna Book*.[2] A 1/4-λ sleeve, open at the top and shorted to the coaxial-cable shield at the bottom, is placed over the feed line. The impedance at the top of the sleeve is high, and very little current can flow on the outside of the tubing. The detuning sleeve acts as an rf choke, isolating the antenna from the rest of the feed line.

Fig. 3 shows the construction details. A 1/4-λ section of 2-inch aluminum tubing serves as the detuning sleeve.[3] I fabricated a brass bushing for the base of the sleeve. This bushing is made to fit inside the aluminum tubing, and is machined to allow a long double-female coupling (the type with threads over the entire length) to thread through the center. The coaxial cable attaches to this coupling on each side of the bushing, feeding the signal through to the antenna. A Plexiglas® spacer in the top of the sleeve keeps the cable centered. Use a hacksaw to slit the bottom of the tubing, and secure it to the bushing with a hose clamp. — *Rudy Knauck, W7FGQ, Seattle, Washington*

[1] D. DeMaw, "A 'Multipedance' Broadband Transformer," *QST*, Aug. 1982, p. 39.

*Assistant Technical Editor

[2] *The ARRL Antenna Book* (Newington: American Radio Relay League, 1982), p. 5-8.

[3] mm = in. × 25.4.

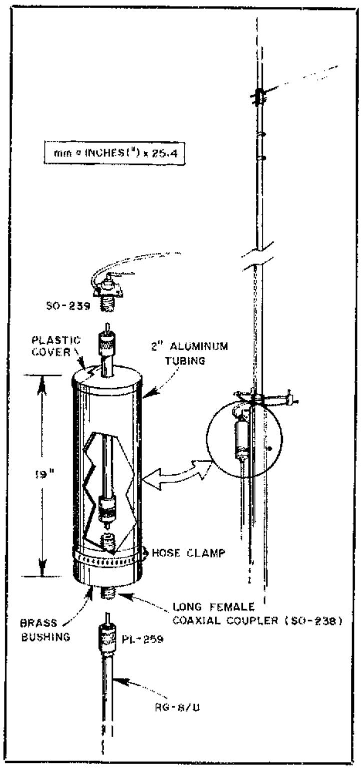


Fig. 3 — Construction details of a detuning sleeve designed by W7FGQ for the Ringo Ranger 2-meter antenna.

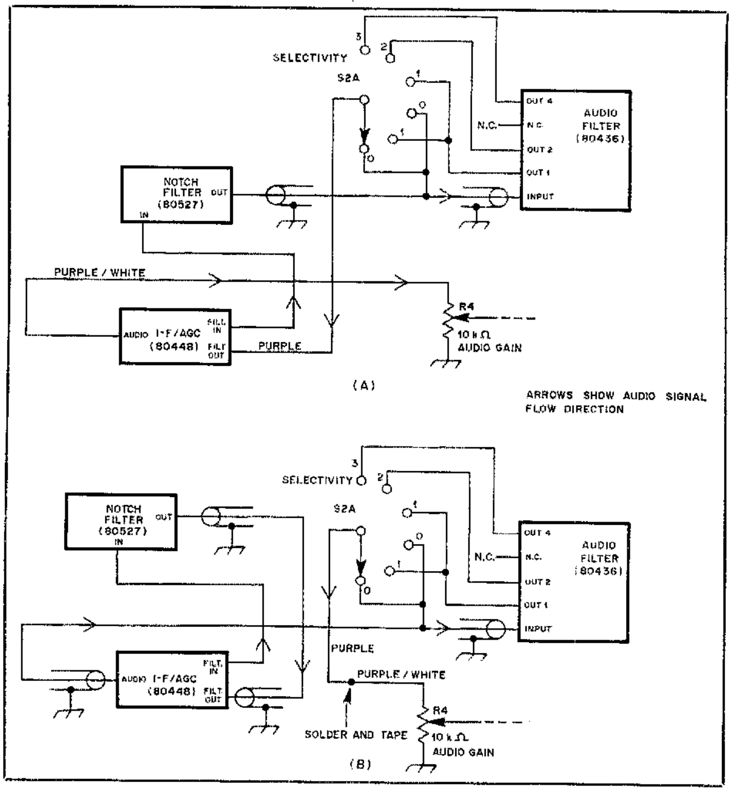


Fig. 4 — Audio- and notch-filter wiring in the Ten-Tec OMNI B before (A) and after (B) K1JD's modification to decrease the agc attack time constant.

REDUCING THE AGC ATTACK TIME IN THE TEN-TEC OMNI

☐ The audio notch and band-pass filters are inside the agc control loop in the Ten-Tec OMNI B transceiver. The advantage of this system is that the agc is not activated by strong signals outside the filter passband or within the notch. You can observe the operation of the filters and agc action by watching the S meter reading decrease as a signal is notched out. Unfortunately, when the audio filters are switched in, a time constant, proportional to the inverse bandwidth of the filter, is introduced into the agc loop. This increases the agc attack time and contributes to the characteristically loud "pop" when the agc responds to a strong cw signal.

After installing the 1.8-kHz and 0.5-kHz crystal i-f filters in my OMNI B, I conducted a short experiment. First, I tuned to a strong cw signal using the 1.8-kHz filter and no audio filtering (0 position). By switching to the 0.5-kHz filter and to audio-filter positions 2 or 3, I noticed an increase in the "pop" as the agc attack time increased. I used this as a measure of the improvement that could be realized by moving the audio filters outside of the agc control loop. With 16 poles of i-f selectivity, performance is not degraded seriously by shifting the additional audio selectivity outside the loop. If your OMNI does not have either a 0.5-kHz or 0.25-kHz i-f filter installed, then I would advise against making this modification.

The only additional material required for the change is a 6-inch piece of shielded wire. The actual circuit changes are shown in Fig. 4. After taking off the top cover, you will have to remove the notch-filter support bracket to gain access to the I-F/AGC board. Remove the I-F/AGC board by pulling gently on the connector pins with pliers. There is an insulating plastic sheet under the board that will also have to be removed. This will provide access to the two filter terminal strips. The board must squeeze past the S meter, but can be removed. Locate the purple wire from the audio filter output in the socket nearest the front and the purple/white audio-output wire from the socket nearest the back. Locate the shielded wire on the notch filter card. This wire is the notch output, even though my manual shows it as the notch input line.

Unsolder the purple/white wire at the I-F/AGC board and the shielded wire at the notch filter output. Connect the center conductor in the shielded wire at the point where you removed the purple/white wire. Solder the braid to the nearby ground lug. This shielded wire *will* reach; remove the audio filter board only if necessary to pull the wire over to the socket. Unsolder the purple wire at the I-F/AGC board and connect it to the loose end of the purple/white wire. Be sure to insulate the joint. Finally, solder the center conductor of one end of a 6-inch piece of shielded cable to the "Filter Out" socket terminal, and the other end to the notch filter output. Ground both ends of the shield at convenient locations. Reassemble the rig, remembering to put the insulating sheets back in place under the boards that you removed.

A check of receiver performance should indicate that the notch filter operates as before. Switching between audio-filter positions 2 and 3 with the 0.5-kHz filter in place will no longer affect the S meter reading. You should also find a noticeable decrease in the audio "pop." — *John DePrimo, K1JD, Honolulu, Hawaii*

COLLINS EQUIPMENT ON 30 METERS

☐ There is an easy way to put the Collins S-Line and KWM-2 on 30 meters. The process is as simple as changing a crystal. I removed the crystal from position 3C (for 14.8 MHz) and replaced it with a 0.001%-tolerance, 13.155-MHz crystal in an HC-6/U holder. This crystal is Collins part no. 290-9042-00. I bought mine from Jan Crystals.

The 14.8-MHz band-switch position now covers 10.0 to 10.2 MHz with the PRESELECTOR or EXCITER tuning control set between 2 and 3 on the logging scale. I did not find it necessary to retune or realign any internal controls after making this change. The 10.100- to 10.150-MHz range is between 100 and 150 on the main tuning dial. There is an added bonus — the unit now tunes WWV at 10 MHz instead of 15 MHz, a frequency that produces more consistent propagation in most of North America. — *Ray Soifer, W2RS, Glen Rock, New Jersey*

Technical Correspondence

Conducted By
Dennis J. Lusis,* W1LJ

TRANSISTOR INPUT AND OUTPUT CAPACITANCES

☐ Some amateurs have indicated during correspondence with ARRL Hq. staff members that they are confused about the C_{in} (input capacitance) and C_{out} (output capacitance) of transistors. They do not understand what effect, if any, these values have on the design of input and output networks. Practically, the effects of C_{in} and C_{out} can be significant if a proper impedance transformation is to be obtained: Maximum power transfer will take place only when the driving source is matched to the input of the transistor amplifier. Likewise with the transistor output and load.

These capacitances need to be absorbed in the matching network when L-C components are used, such as in Fig. 1. When broadband transformers are used in place of L-C networks (Fig. 1B), it is important to consider the effects of the capacitances on the transformers. Knowledge of the C_{in} and C_{out} values is required in order to execute a good design. Unfortunately, not all manufacturers include this data in their specification sheets. Motorola Semiconductor Corp. provides capacitance-versus-frequency curves for most of its power transistors.[1]

We must be mindful, when working with rf bipolar devices, that the input and output capacitances change markedly with operating frequency. The curves in Fig. 2 illustrate this characteristic, which complicates the feedback-network design for wide-band linear power amplifiers. This was not a concern when we worked with vacuum tubes "way back when." The capacitances change significantly with the operating voltage. Each time the collector-supply voltage is doubled, the capacitance increases by an approximate factor of 3. This is true also of power FETs (drains). The capacitance is constantly changing as the sine-wave (signal voltage) goes through its excursion. The changing capacitance results in a varactor type of action, which in turn generates harmonic currents of high magnitude. It is for this reason that we need to use considerably more filtering with a solid-state power amplifier than would be needed in a vacuum-tube model. The usual envelope distortion associated with harmonic generation in vacuum tubes is prevalent also in solid-state amplifiers. It is not uncommon, therefore, to observe second- and third-harmonic levels (before filtering) that are only 10 to 15 dB below the peak output level of the fundamental frequency.

Power FETs have the advantage of being rather similar to vacuum tubes with respect to stable values of C_{in} and C_{out}. That is, the capacitances do not change significantly with changes in drive-power level or frequency. This makes them well suited to use in wide-band amplifiers that employ feedback.[2] Power-FET input capacitances are quite low compared to those found in bipolar transistors. They are on par with most sweep tubes. The output capacitance of big power FETs does, however, compare somewhat in magnitude to those encountered while working with power bipolars.

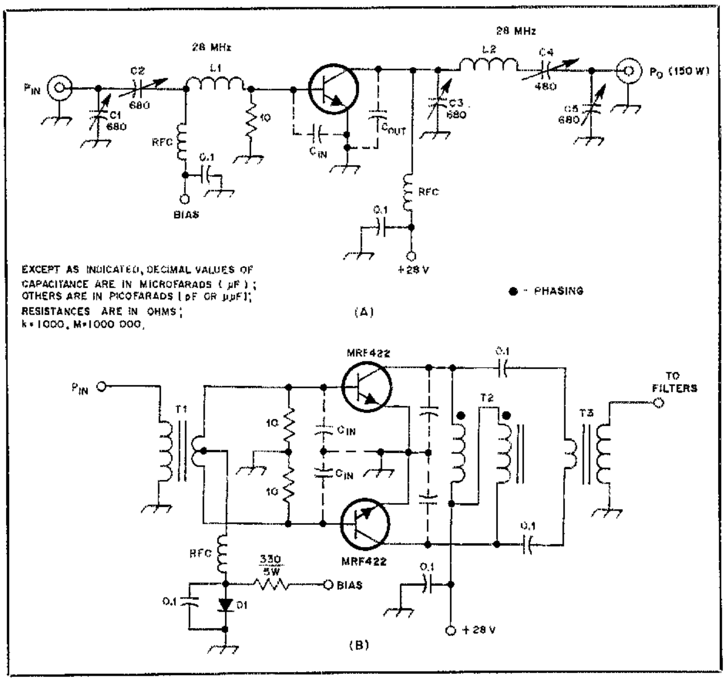


Fig. 1 — C_{in} and C_{out} at A can have a marked effect on the network design, and these values have to be absorbed into the networks. The capacitances need to be accounted for when designing for broadband amplifier service (T1, T2 and T3 of illustration B).

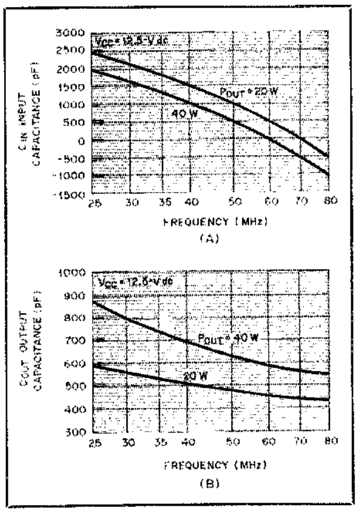


Fig. 2 — Transistor input capacitance versus operating frequency is shown by the curves at A. At B, a similar set of curves shows the change in output capacitance as the frequency is varied. These curves apply to the 2N5849. (*courtesy of Motorola Semiconductor Corp.*)

We should pay attention to the C_{in} and C_{out} characteristics of transistors we plan to substitute in amplifiers, lest the new devices upset the matching networks by virtue of markedly different amounts of capacitance. I was fortunate when replacing a pair of Toshiba S2535s with Motorola MRF421s after I blew out the original transistors in my FT-301D transceiver. Not only did the Motorola parts fit the pc board and match up with the mounting holes, the capacitances were close enough to those of the Toshiba devices to permit instant success upon firing up the repaired transmitter. This might not always be the case when making substitutions, so beware! I hope this discussion has helped to solve some problems you may have had when designing homemade solid-state power amplifiers. — *Doug DeMaw, W1FB, ARRL Senior Technical Editor*

[1] *Motorola RF Data Manual*, 1978 edition, P.O. Box 20912, Phoenix, AZ 85036.
[2] H. Granberg, "MOSFET RF Power — An Update," Parts 1 and 2. Dec. 1982 and Jan. 1983 *QST*.

*Assistant Technical Editor

SOLAR FLUX/SUNSPOT NUMBER CONVERSION FOR MINIMUF

☐ Over the past couple of years, I have spent a fair portion of my hf operating time obtaining a working knowledge of propagation factors, Rose's article, "MINIMUF: A Simplified MUF-Prediction Program for Microcomputers" (Dec. 1982 *QST*), offered me an additional means toward this end. Along with other students of the subject, my primary source of information regarding solar activity and propagation conditions is the hourly broadcast by WWV. Since they use the 10.7-cm solar flux as an indicator, the graph included in the article must be used to convert this figure to the sunspot number, which is utilized in the program.

This presents two problems. First, at the time I began using MINIMUF, solar flux values were well in excess of 250, which is the upper limit of the conversion graph. Second, while the computer would calculate and display the results of the complex algorithm quickly, the solution hinged upon my ability to find the graph among the clutter of my operating position. In order to at least resolve the second problem, I investigated the possibility of performing the solar-flux-to-sunspot-number conversion in the form of an equation, which could then be incorporated into the driver. I could not find any information at our public library dealing with propagation, so I (approximately) characterized the conversion curve with the quadratic equation:

$$SF = (0.0008)(SSN)^2 + (0.73)(SSN) + 65 \qquad \text{(Eq. 1)}$$

where

SF = 10.7-cm solar flux
SSN = sunspot number

Solving, then, for the sunspot number:

$$SSN = \frac{-0.73 + \sqrt{(0.73)^2 - 4(0.0008)(65 - SF)}}{2(0.0008)} \qquad \text{(Eq. 2)}$$

or, in a BASIC-compatible form:

```
S9 = 625 * (SQR((0.73) 2 - 0.00032 * (65 - SF)) - 0.73)
```

(Eq. 3)

These equations do not perfectly represent the conversion curve, but give integer values adequate for setting the S 9 sunspot parameter in the MINIMUF model. The greatest error appears at a solar flux value of 130, where the graph indicates a sunspot number of 83, versus a value of 82 derived from the above equations. — *Clay Bartholow, KA0LEN, Minneapolis, Minnesota*

MORE MINIMUF PROGRAM MODS

☐ I was delighted to find Rose's article on muf prediction in December 1982 *QST*. I promptly entered the program into my computer, and started running muf predictions.

Here are a couple of modifications to the program that some users might want to add. One addition is a conversion scheme for deriving the sunspot number from the solar-flux index. The second addition draws a graph of muf versus UTC and, like the original, is written for the Tektronix 4050 series computers. Once again, thanks for the excellent article. — *Denton Bramwell, K7OWJ, Minden, Nevada*

MINIMUF PROGRAM MODIFICATIONS:

```
102 DIM T0 (24,2)
430 PRINT "SOLAR FLUX NUMBER =";
442 GOSUB 2000
640 FOR T5 = 0 TO 23
650 GOSUB 1000
660 PRINT USING 670:T5,J9
661 T0(T5 + 1,1) = T5
662 T0(T5 + 1,2) = J9
670 IMAGE 5X,2D,7X,2D.D
680 NEXT T5
690 PRINT
700 GOSUB 7000
710 END
720 REM *** LINE 700 "REM" RATHER THAN
"GOSUB" IF NOT USING GRAPHICS
2000 REM CALCULATES SUNSPOT NUMBER
FROM FLUX NUMBER
2010 B1 + 1*(S9<100) + 2*(S9 = >100)*(S9<170) + 3*
(S9 = >170)*(S9<200)
2012 B1 = B1 + 4*(S9 = >200)
2020 ON B1 GOSUB 3000, 4000, 5000, 6000
2030 RETURN
3000 REM COMPUTES SUNSPOT FOR FLUX
BELOW 100
3010 S9 = (S9-64.8)/0.745
3020 RETURN
4000 REM COMPUTES SUNSPOT FOR FLUX
BETWEEN 100 AND 170
4010 S9 = (S9-58.5)/0.875
4020 RETURN
5000 REM COMPUTES SUNSPOT FOR FLUX
BETWEEN 170 AND 200
5010 S9 = (S9-45.3)/0.99
5020 RETURN
6000 REM COMPUTES SUNSPOT FOR FLUX
OVER 200
6010 S9 = (S9-41.3)/1.015
6020 RETURN
7000 REM *** DRAWS GRAPH OF DATA
7001 REM *** DELETE 7000 THRU 7200 IF NOT
USING GRAPHICS
7010 WINDOW 0,23,0,40
7020 VIEWPORT 50,125,20,80
7030 AXIS 1,5,0,0
7040 FOR I = 0 TO 9 STEP 2
7050 MOVE I,0
7060 PRINT "J";I
7070 NEXT I
7080 FOR I = 10 TO 23 STEP 2
7090 MOVE I,0
7100 PRINT "HJ";I
7110 NEXT I
7120 FOR I = 0 TO 40 STEP 5
7130 MOVE 0,I-0.5
7140 PRINT "HHH";I
7150 NEXT I
7160 MOVE T0 (1,1), T0(1,2)
7170 FOR I = 1 TO 24
7180 DRAW T0(I,1),T0(I,2)
7190 NEXT I
7200 RETURN
```

DX AND THE BREWSTER ANGLE

☐ The best DX antenna I ever had on 40 meters was a pair of phased 1/4-λ verticals. I used that array over 10 years ago in rural Michigan. Encouraged by the 40-meter performance, I tried an 80-meter vertical. Again I enjoyed excellent results for my efforts. Curiosity then prompted me to try a 1/4-λ vertical on 20 meters. This time I was disappointed.

According to the books, a vertical should have a radiation pattern lobe aimed at the horizon. By contrast, a horizontal antenna exhibits a null in this direction. Since a low vertical angle of radiation is desirable for working DX, why did a simple dipole outperform my 20-meter vertical? I worked Gus Browning, W4BPD, from his many exotic island locations on 20 meters, and all *he* used was a multiband vertical (Oops — guess that gave my age away!). Why did a vertical always work for Gus and not for me? It was not until several years later that I found out why.

Those beautiful radiation patterns in the books are accurate if you assume that the earth is a perfect conductor. It isn't. For horizontal antennas, this makes very little difference in the low-angle radiation. For verticals, however, the situation is altogether different. Toward the horizon there is an almost complete cancellation of energy because of ground reflection.

The reasons for this are based in optical theory. There is a 180° phase shift in the reflected ray on the horizon. This phase shift causes signal cancellation. At some angle, the phase shift is 90° and the reflected ray is neither a help nor a hindrance. Above this angle, the reflected ray tends to reinforce the direct ray. This angle is called the "Brewster" or "Pseudo-Brewster" angle.

In the RSGB publication *HF Antennas for All Locations* by L. A. Moxon, G6XN, Brewster angles are given for various situations. They are: dry ground, 28°; average ground, 17°; very moist ground, 10°; fresh water, 6° and sea water, 1°.

The final fact in solving the riddle that bothered me is that lower-frequency DX tends to use relatively high vertical angles of radiation. At last it became clear that my antennas mounted over "average ground" exhibited a Brewster angle that didn't affect me on 40 meters, but did on 20 meters. Gus had mounted his vertical right at the edge of the sea, which provided a superior environment — hence a very low Brewster angle.

The answer, you may say, is to lay down 120 radials, like the broadcast stations do, to help cut down ground-return losses in the antenna system. However, those radials will *not* significantly affect ground reflection losses at the lower angles. To lower the Brewster angle of average ground appreciably would require a ground screen that is many wavelengths in radius. Most of us can't afford that kind of installation.

Until I move to an ocean island, I will use horizontal polarization on the hf bands above 14 MHz. — *Chuck Hutchinson, K8CH, ARRL Assistant Technical Editor*

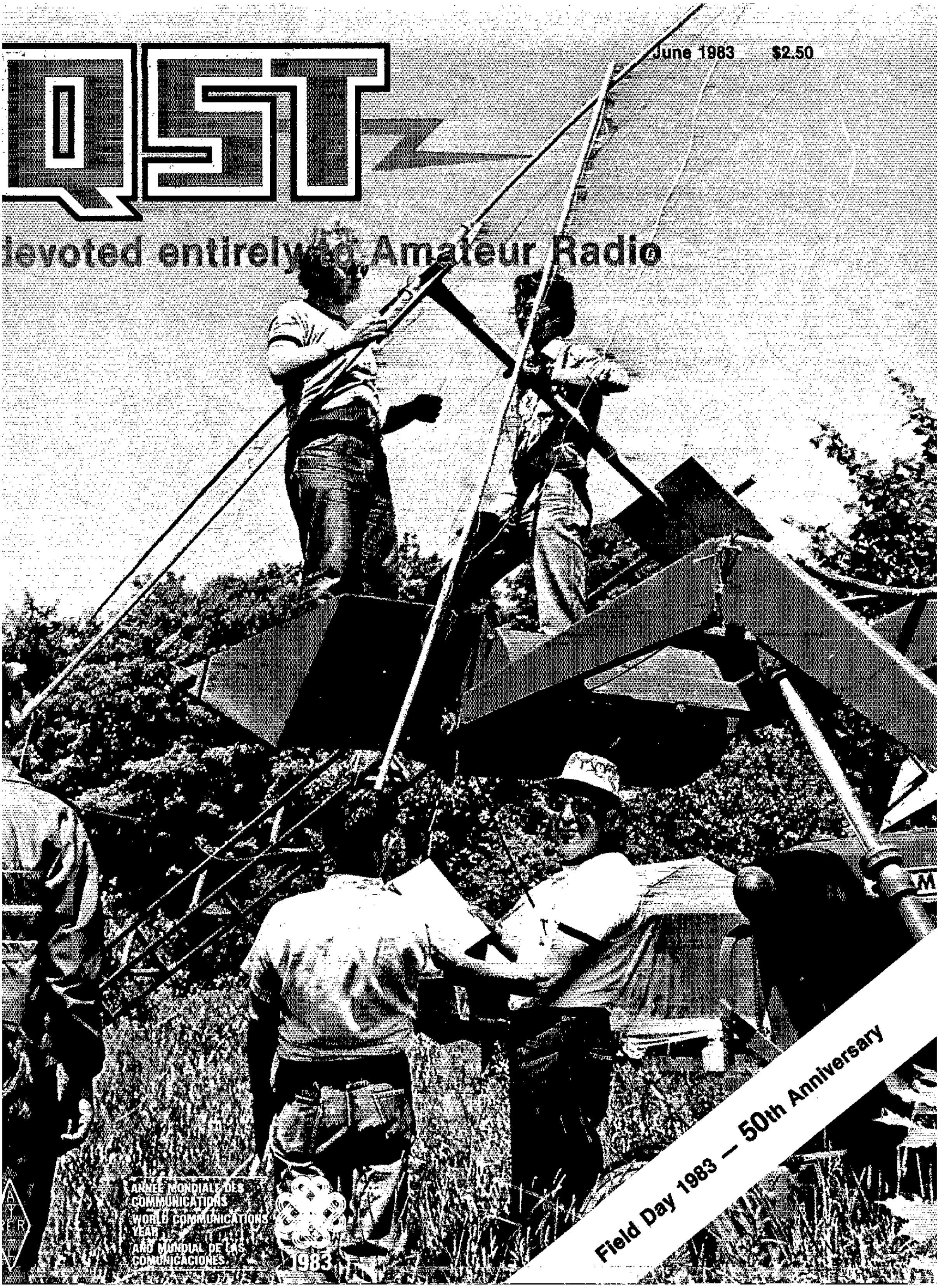
Feedback

☐ The Mirage C22 220-MHz amplifier reviewed in March *QST* and the B23 2-meter version reviewed in May 1981 have been updated since the units were submitted to ARRL for review. The new versions now have slide switches on the front panel (opposite the rf connectors) for ON/OFF and SSB/CW selection. Also, an intermittent voltage regulator problem was discovered in the automobile in which the C22 was mounted. Circumstantial evidence indicates that this contributed to the failure of the one C22 unit.

☐ March 1983 *QST* contains an article entitled "Go Class B or C with Power FETs" by W1FB. The transistors specified are Motorola MRF138s. We have learned since publication that some manufacturing problems have delayed the appearance of these devices on the market. They will be available later this year, according to a Motorola spokesman. Those interested in the amplifier may substitute the Siliconix DV2840S, which has nearly the same characteristics and is in the same package format. The notable difference is that the DV2840S can deliver 40 W of output rather than 30. At 1.5-W input, the power is 30 W at 175 MHz. Less than 1 W of drive should provide 30 W of output from 1.8 to 30 MHz. Further details are available from Ed Oxner, KB6QJ, our ARRL TA at Siliconix.

☐ Please note that the correct street address for Macrotronics, Inc., is 1125 N. Golden State Blvd. An incorrect address appears in the March 1983 Product Review column.

☐ The $2 copy fee quoted in footnote 7, p. 27, of April 1983 *QST*, for the BASIC computer listing for the ANTRXSWR/BAS program is in error. The correct fee for this program, used with Hall's article, "The Search for a Simple, Broadband 80-Meter Dipole," is $1. If you've already submitted $2, you'll receive a credit for the difference.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CONTENTS

OUR COVER

Ah, the purr of the generator and the . . . roar of the tractor? Maybe your club can come up with an even better way to erect an antenna, but you'd be hard-pressed to outdo the Ottawa Valley Mobile RC (VE3RAM). All the FD rules are in May *QST*, page 86.

TECHNICAL

BEGINNER'S BENCH

NEWS AND FEATURES

OPERATING

DEPARTMENTS

SSTV Today

Many new and exciting developments have occurred in SSTV! Read about them now and join in on the fun!

By Ron Flynn,* KB8LU

The appeal of slow-scan television (SSTV) as an Amateur Radio communications mode is visibility. You can *see* a picture of the person you are in contact with, as well as pictures of virtually any other subject. SSTV pictures can be transmitted to any part of the world. The only operational limitations are the subject material (defined by good taste) and the inherent resolution of the equipment being used.

Copthorne MacDonald, now VE1BFL, is credited with inventing amateur SSTV. His articles on the subject were first published in *QST* in August and September of 1958. Over the next six years, MacDonald and other British and American experimenters improved their systems and ultimately settled on the standards still in use today.[1]

Some Basics

SSTV pictures are transmitted on the amateur hf bands as a series of audio tones ranging from 1500 to 2300 Hz. Black corresponds to a frequency of 1500 Hz, white is 2300 Hz, and the frequencies between them define 14 shades of gray. The received tones are demodulated, output as video and fed to a monitor.

Each line of an SSTV picture consists of 128 different pixels (picture elements), or dots, and each line scans from left to right. The 128-line SSTV picture scans from top to bottom in 8.5 seconds. The 1:1 aspect ratio (width to height) produces a square picture. A 1200-Hz sync pulse is used to reset the start of the scan to the top left-hand corner of the monitor.

Equipment

Over the years, many individuals have published articles describing "home-brewed" SSTV equipment. Several manufacturers have produced various types of SSTV scan converters.[2] Robot Research, Inc., introduced the Robot 400 scan converter in 1976; today it remains the most popular and versatile scan converter in use.

In addition to a scan converter, you need a video monitor to display the pictures. A B & W CCTV (black and white closed-circuit TV) type provides excellent results. An inexpensive audio cassette or reel-to-reel audio tape recorder is useful. The SSTV audio tones can be recorded off the air and saved on tape for later viewing or retransmission.

A B & W CCTV camera allows you to create pictures for live or delayed transmission from audio tape. Using a camera, you can create your own SSTV "programs" on tape for transmission during a QSO.

Color SSTV

Several people participated in the early experiments to develop a working color SSTV system. This work brought about the three-Robot system described by Jeremy Royle, G3NOX, in November 1980 *QST*.[3] In this system, three Robot 400s, each capable of storing one full 128-pixel by 128-line SSTV picture, are "slaved." Separate red, green and blue color filters are alternately placed in front of the lens of a B & W CCTV camera focused on a color picture. The filtered B & W representations are alternately loaded into the individual scan-converter memories. Then the memory contents are transmitted in sequential frames, or "frame sequentially," as it is popularly known. The receiving operator loads the picture information into the appropriate converter at his or her station. The combined video outputs produce a composite color SSTV picture on a color monitor. Though this system worked quite well, its main disadvantage was the high initial cost of separate scan converters (three).

At the 1981 Dayton Hamvention, Sam Mormino, WA7WOD, and Howard McAfee, KD6HF, introduced the first commercially available full-color SSTV system.[4] They developed a pc board that mounts inside the '400 cabinet. The 3000C board, with additional circuitry and switching, converts a standard single-memory Robot 400 into a three-memory unit capable of receiving or transmitting one full-color picture or three separate B & W SSTV pictures. In addition, they developed a unique video interface module. It mounts inside the cabinet of specific (but standard) inexpensive uhf/vhf color TV sets that are readily available in the U.S. The complete

[1]Notes appear on page 14.

*Rte. 2, Box 204, 67th St., Bangor, MI 49013

Fig. 1 — An RGB frame-sequential color SSTV picture received from KC5VC on 14.230 MHz. From left to right are the red frame in B & W, the red and green frames combined and the full-color RGB composite picture. *(photos by George Swift, W8AMB)*

Fig. 2 — Hard-copy printout of an SSTV picture using an Epson MX-80 printer and Commsoft's PhotoCaster SSTV system.

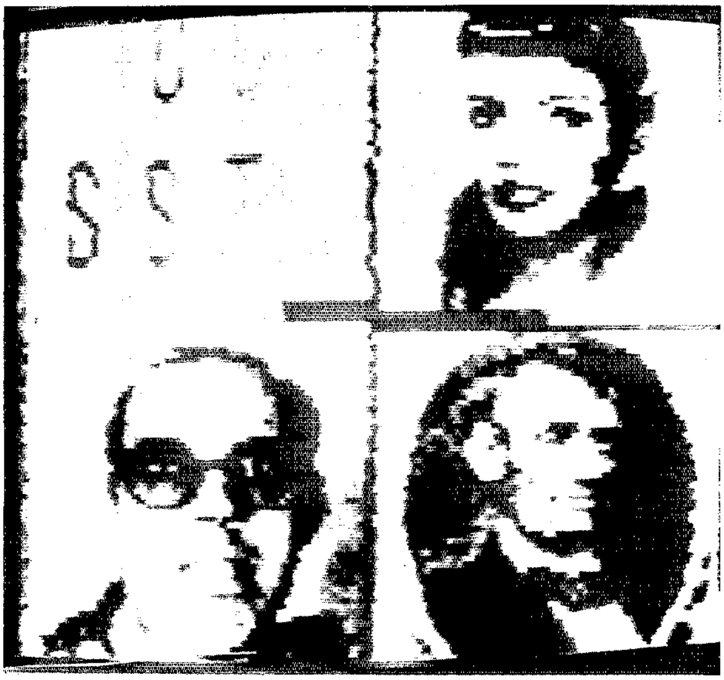


Fig. 3 — Four separate pictures can be loaded into a single-memory Robot 400 using the Four Quad modification of KD6HF.

color SSTV system is available in kit form or wired and tested.

The scan converter RGB (red, green, blue) video outputs feed the color monitor RGB guns, and perfect color reproduction is achieved. A B & W CCTV camera used in conjunction with inexpensive color filters or a color camera can be used to produce excellent quality color SSTV pictures. Color pictures from uhf/vhf TV programs can be snatched (stored) in the converter and transmitted as color SSTV pictures.

This method uses the standard RGB sequential frame transmission mode. The "two, two and two" heard on the SSTV frequencies alerts receiving *stations that* two frames from the red memory will be transmitted first, followed by two from the green and, last, two frames from the blue memory. The transmitting station may select any combination of RGB desired, depending upon band conditions and QRM, to allow all receiving stations to get a good picture.

When receiving a color SSTV picture, the first red frame is usually loaded into all three memories to clear the monitor of noise or a previously stored picture. Therefore, you see the picture first in B & W (see Fig. 1). As the green frame is loaded into the second (green) memory, shades of red and green begin to appear on the monitor. When the final blue frame is received, the full-color SSTV picture is displayed before you. Using this system with proper lighting on the subject, all colors and shades (including flesh tones) are reproduced accurately. With the picture contents of three memories overlying each other, the composite color picture looks smoother and the *visual* resolution is better than a single B & W SSTV picture, which looks somewhat digitized.

Several other color SSTV systems are available to amateurs throughout the world. Syd Horne, VE3EGO, offers a three-memory Robot conversion in kit form called the Colorscan 403.[5] Dave Cowie, ZL1LH, has three-memory color scan converter kits for hams in that part of the world.[6] There is also the German SC422A, a complete, dedicated three-memory color scan converter.[7]

Computers and SSTV

Computers and keyboards have been a part of SSTV for many years, but their only use was to generate large graphics letters and numbers for transmission. This proved to be an aid to many in SSTV QSOs, but no pictures were sent or received using the computer.

The two most widely used computers on SSTV today are the Radio Shack TRS-80® series and the Apple®. An early Apple SSTV program produced very limited results. Only black, white and one shade of gray were displayed. In early 1982, Commsoft introduced their PhotoCaster hardware/software SSTV system for the Apple II.[8] Such a system is capable of resolving the standard 128 × 128 × 16 gray level B & W SSTV picture. It has the ability to display eight colors of sequential-frame RGB color SSTV. The package contains many other sophisticated features, including picture enhancement and hard-copy printouts of SSTV pictures using a

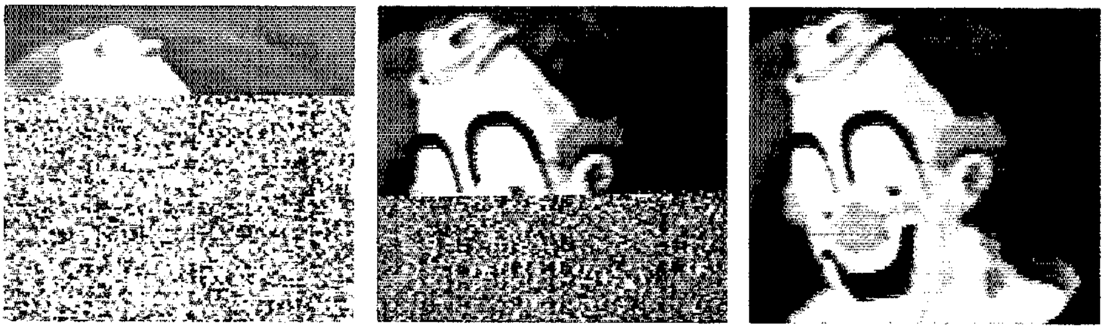


Fig. 4 — A 25.5-second RGB SFC SSTV picture received from KC5VC on 14.230 MHz. From left to right are the picture scan after eight seconds and after 17 seconds, and the full 25.5-second scan.

printer such as an Epson MX-80® equipped with Graftrax® (see Fig. 2).

In search of an economical and simple way for amateurs to participate in SSTV using computers, Clay Abrams, K6AEP, discovered that the TRS-80 Color Computer has great potential.[9] He designed a simple interface board that allows the computer to become a complete SSTV system. Clay has written several SSTV programs with many features, including graphics generation and hard-copy picture printing. The early system would display only four shades of gray and four colors. During the fall of 1982, Clay, along with Bob Blackstock, WB5MRG, developed an improved hardware/software SSTV system for the Color Computer. It provides 256 × 128 × 16 gray-shade resolution and will display 256 different colors.

Robot 400 Modifications

As mentioned earlier, the '400 scan converter is still popular, and thousands are in use today. As hams became more familiar with digital logic, the covers of the converters began coming off and many modifications were developed, especially within the last year or so. Some of these were quite utilitarian, while others simply provided a new dimension of fun and diversity in SSTV QSOs.

QRM can cause a loss of portions of the SSTV picture. Sam Mormino, WA7WOD, and Steve Bruce, WA4OAA, developed a modification that cuts the Robot 400 clock speed in half. In doing so, the SSTV picture is sent and received in 17 seconds instead of 8.5. This reduces the potential for interference to the picture as a whole.

M. H. Emmerson, G3OQD, developed a light-pen addition for the '400. It allows you to display a picture from the scan-converter memory on a monitor, use the light pen to write over the monitor screen, and have the writing superimposed over the picture in memory.

KD6HF has produced many successful Robot circuit alterations. His First Sync modification allows the transmitting station to sync or reset all receiving-station monitors so that everyone receives the SSTV picture scan in unison. Formerly, receiving stations picked up the first picture scan at various points, and it took up to eight seconds to get in sync with the transmitting station.

Howard also developed a Graphic Overlay feature similar in purpose to G3OQD's light pen. With this, graphics from tape, a keyboard, a camera or a received signal can be superimposed on any existing picture. White or black graphics can be used, depending on the picture color content. The graphics to be superimposed may be positioned anywhere on the picture.

The Four Quad modification is another production of Howard's (see Fig. 3). It allows a single-memory '400 to store four different, small B & W SSTV pictures. These pictures can be loaded into any of the four screen quadrants, in any order, from tape or off the air. This greatly enhances the operating capabilities of the '400. In addition, his Zoom modification allows the operator to take any of the smaller pictures from the four quandrants, expand it to a full-size SSTV picture, and transmit it.

Latest Developments

In June 1982, on-the-air experiments began with a new form of color SSTV. They were a departure from the standard 8.5-second SSTV frame scan. By means of a simple modification to the Interface Systems color system, sequential lines (instead of sequential frames) from the red, green and blue memories are transmitted. This is called line-sequential color SSTV. The same amount of video information is being transmitted in a single 25.5-second scan as is transmitted in the RGB frame-sequential overlay method described earlier. Two advantages of this new method are that the true full-color SSTV picture, in all its splendor, "paints" down the monitor screen, and it loads automatically — no switching is required (see Fig. 4). This new approach is called 25.5-second single-frame color (SFC) SSTV.

Sam Mormino, Jim Williams, KC5VC, and I took part in the first on-the-air experiments using 25.5-second RGB SFC. While the immediate effect of seeing full-color SSTV pictures unfold before your eyes is tremendous, there are other considerations. With the slower scan rate, intermittent QRM is less detrimental to the total picture. However, QRM can knock the line-sequential transmissions out of color sync at the receiving end. This results, for example, in the transmitted picture information from the red memory going into the green or blue memory at the receiving end. The color display is then untrue. A fix is available to "kick" the scan back into color sync, but some true color will be lost. Perhaps the insertion of an individual sync pulse for each line or other refinements will perfect this method of transmission. To date, over 50 amateurs are using 25.5-second RGB SFC SSTV. Quite a few owners of standard single-memory '400s have installed this feature and are copying the slower-scan-rate pictures in B & W.

Microcraft offers a high-resolution, two-memory B & W scan converter.[10] This unit employs three scan rates, 8.5, 17 and 34 seconds, to produce different B & W SSTV picture resolutions. I've received very good 8.5-second B & W SSTV pictures from owners of this unit. Instead of the usual 16 shades of gray, this scan converter has 64. The occasional SSTV phenomenon known as "contouring" seems to be eliminated.[11] The three scan rates can produce pictures with resolutions up to 256 pixels with 256 lines and the 64 shades of gray. However, the 17- and 34-second scan rates are not copyable with other SSTV systems.

Normally, SSTV picture lettering sizes of about 1/2 inch or less are difficult to read on a 9-inch monitor. With the higher-resolution picture capabilities of this scan converter, more detailed pictures (such as schematic diagrams) can be exchanged.

In the spring of 1983, Robot Research Inc., announced that it would begin manufacturing color SSTV equipment. For the thousands of Robot 400 owners, they developed a simple do-it-yourself kit to convert the '400 into a 12-second single-frame color SSTV system.

Recent developments in electronic technology, combined with the creativity and knowledge of SSTV innovators, have resulted in the many fine advances that have occurred with this mode over the past two years. SSTV has now become one of the most diverse and interesting modes of Amateur Radio communication. Why not join us?

Appendix

Robot 400 Modification Information Sources

Robot 400 half-speed, 17-second scan
Four quadrant
G3OQD light pen
These modifications were published in the November 1981 issue of *A5 Magazine*. Back issues are not available. Authorized reprints of that issue may be obtained from Ralph Wilson, WBØESF, 4011 Clearview Dr., Cedar Falls, IA 50613. Cost is $2.50 ppd.

Graphics overlay
First Sync
Published in *Worldradio*, June 1982. Reprints may be obtained from *Worldradio*, 2120 28th St., Sacramento, CA 95818.

Graphics overlay
First Sync
Four quadrant
Zoom
Schematic diagrams for these modifications are available from Interface Systems Inc., Rte. 4, Box 634-K, Lindale, TX 75771. Please include a large s.a.s.e. and 37 cents postage.

Ron Flynn was first licensed in 1978 and immediately became involved in SSTV. He currently writes an SSTV column for Worldradio. *Ron is a horticulturalist with a BS degree from Michigan State University. Since 1977, he has been semiretired and does landscape consulting.*

Notes

[1][Editor's Note: The amateur standard, adopted by gentlemen's agreement in 1968 when SSTV was first authorized for general use in the U.S. and Canada, calls for 120 lines per frame, requiring eight seconds to transmit a complete picture. For some years the trend has been toward 128 lines per frame, especially when digital-logic circuitry is used to develop the sync signals, requiring 8.53 seconds to send a picture. With proper adjustment of the vertical sweep circuitry in a receiving monitor, either rate may be received satisfactorily.]

[2][Editor's Note: Scan converters are used to transform picture information from SSTV standards to conventional fast-scan TV standards, and vice-versa. Most (if not all) commercially available scan converters also incorporate a *modulator/demodulator* as an integral part of the unit. As a result, differentiation between the actual scan converter and demodulator sections is at times ignored and the entire unit is referred to as a "scan converter."]

[3]J. Royle, "SSTV in Colour," *QST*, Nov. 1980, p. 11.

[4]Interface Systems Inc., Rte. 4, Box 634-K, Lindale, TX 75771.

[5]Sycel Communications Inc., P.O. Box 893, Belleville, ON K8N 2G6, Canada.

[6]52 Prince Regent Dr., Bucklands Beach, Auckland, New Zealand.

[7]Available in the U.S. from KW Control Systems Inc., South Plank Rd., Middletown, NY 10940.

[8]Commsoft, 665 Maybell Ave., Palo Alto, CA 94306.

[9]Amateur Radio Software, 1758 Comstock La., San Jose, CA 95124.

[10]Microcraft Corp., P.O. Box 5130, Thiensville, WI 53092.

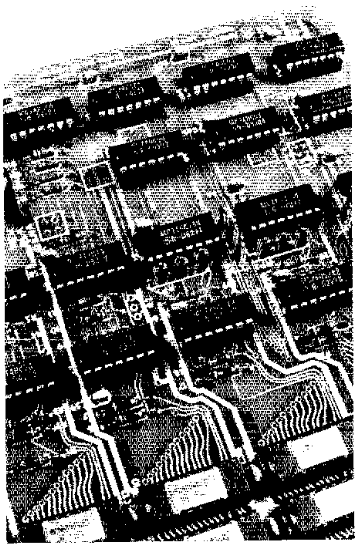
[11]Contouring is a visual imperfection appearing on the SSTV monitor. It occurs when one attempts to display a picture with insufficient contrast or picture detail, or with too few gray levels represented. Parallel wavy lines appear over portions of the picture, similar to the lines on a contour map.

New Products

MOTOROLA HIGH-SPEED CMOS OCTAL INTERFACE DEVICES

☐ The Motorola Integrated Circuit Division announces the availability of a family of high-speed CMOS octal interface devices. For the first time CMOS interface devices are available that will offer a feasible alternative to LSTTL for memory and microprocessor systems. The first of these devices to be released are the MC54/74HC373, MC54/74HC374, MC54/74HC533 and the MC54/74HC534. These devices will be followed by 38 others to be released over the next year, bringing the total to 42 devices in the octal interface subfamily. All devices will be mutually sourced by National Semiconductor. These new devices will satisfy the requirements for high-speed digital interfacing including: data and address buffering to increase fan out for driving additional circuits; address decoding as part of a memory mapping scheme for memory and peripheral chip selection; and data or address storage for synchronizing communications between subsystems operating at different speeds.

For the designer there are four basic groups of devices to choose from: octal decoders, octal buffers/line drivers/line receivers and transceivers, octal flip-flops and latches, and octal level translators.



The bases for these four groups are the MC74HC138, MC74HC240/260 series, the MC74HC370/530 series and the MC74HC4300 series, respectively. Within each basic series a wide variety of options are available. The various members within each series offer noninverted and inverted outputs as well as bus-oriented and conventional pin layouts.

The devices provide the output drive capability and high speed required of interface circuits without sacrificing CMOS high-noise immunity or low power consumption. A 6-mA sink capability exists for the 74HC series of interface circuits across commercial temperature ranges at 0.4 V. In addition to the octal decoders, latches, buffers, transceivers and line drivers, a series of eight-level translators is provided to aid in hybrid systems using a mixture of CMOS, NMOS and TTL devices.

This subfamily within the high-speed CMOS family provides an alternative interfacing solution for microprocessor and memory systems. Now an alternative to LSTTL exists for interfacing with memories and microprocessors through CMOS high-speed logic circuits that provide for systems having much lower power requirements. For further information contact your local Motorola sales office, distributor or Dave Niehaus at Motorola Inc., MOS Integrated Circuits Group, 3501 Ed Bluestein Blvd., Austin, TX 78721. — *Paul K. Pagel, N1FB*

Lightweight Trap Antennas — Some Thoughts

Portable multiband antennas need not be heavy and bulky. Small traps and light-gauge wire can provide a trap dipole that fits in a lunch bag. Try these practical guidelines for your next small antenna.

By Doug DeMaw,* W1FB

Vacationers, campers, sales people and QRPers take note! You need not carry a large multiband trap dipole afield if your transmitter is in the 150-W-output class, or lower. You can construct your own traps inexpensively with ordinary materials, and they can be made quite small without becoming poor performers. This article describes some easy techniques for fabricating homemade antenna traps. Additional hints are offered for keeping the bulk and weight of portable antennas within reason.

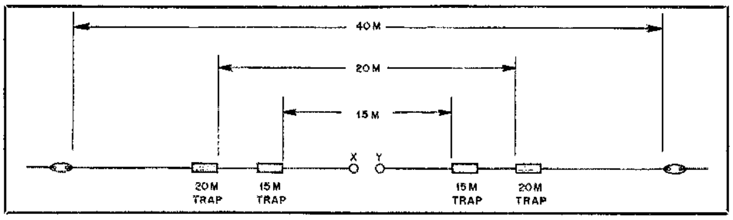


Fig. 1 — Representation of a three-band trap dipole antenna.

A Review of the Trap Concept

A "trap" is exactly what the term implies. It traps an rf signal to prevent it from passing beyond a specific point along an electrical conductor. At some other frequency, however, it no longer acts as a trap, and permits the passage of rf energy.

An antenna trap is designed for a particular operating frequency, and there may be several traps in the overall system — each designed for a specific frequency. Therefore, a 40- through 10-meter trap dipole might contain traps for 10, 15, 20 and 30 meters. On 40 meters, all of the traps are "absorbed" into the system to become part of the overall 40-meter dipole. Owing to the loading effect of the traps, the 40-meter portion of the antenna will be somewhat shorter than a full-size 40-meter dipole with no traps. The antenna bandwidth will be narrower when traps are used. Fig. 1 illustrates the general format for a multiband dipole.

A trap style of antenna is not as efficient as a full-size dipole. This is because there will always be some losses in the traps. But the losses in a well-designed system are usually so low that they are hard to measure by simple means. The losses represent a small tradeoff for the convenience of being able to accommodate many ham bands with one radiator and a single feed line. Yagi antennas contain traps in the parasitic elements (directors and reflectors) as well as in the driven element. Therefore, a multielement antenna of that type may have as many as 12 traps.

Electrical Characteristics

An antenna trap is a parallel-resonant L-C circuit. Therefore, it is similar to the tuned circuit in a transmitter or receiver. A resonator of this kind, if designed correctly, has a moderate Q and a fairly narrow bandwidth. This means that the trap capacitor should have a high Q and the trap coil should contain wire that is reasonably large in cross section. These traits will help to reduce losses.

Fig. 2 shows the equivalent circuit for an antenna trap. Once this network is adjusted to resonance in the desired part of an amateur band, it will not be affected significantly by the attachment of the wires that comprise the antenna. A well-designed and -constructed trap should not change frequency by any great amount when the temperature or humidity around it varies. Therefore, it is important to use a stable capacitor, a rigid coil and some type of sealant.

Mini Trap Using a Toroid Core

In an effort to scale down the size of my antenna traps during a design exercise for a portable antenna, I decided to investigate the worth of small toroid cores upon which to wind the coils. Ferrite cores were ruled out because they aren't as stable as powdered-iron ones. Furthermore, the powdered-iron material has a much greater flux density than an equivalent-size ferrite core, which means that the core will not saturate as easily at moderate rf power levels.

Development work started with Micrometals Corporation T50-6 toroids, which are sold by Amidon Associates, Palomar Engineers and RadioKit (see *QST* advertisements). My first effort resulted in a pair of very small 20-meter traps. A silver-mica capacitor was chosen for the parallel-tuned circuit. Ceramic capacitors were not used because of previous experiences I had with changes in value under temperature extremes; I had better results with dipped silver-mica units.

My rule of thumb for choosing the coil and capacitor values for traps is based on a reactance of approximately 200 ohms, although values up to 300 have also

*ARRL Senior Technical Editor

yielded good results. Using 200 ohms as the basis for the design, I calculated the capacitor to be a value that was very close to a standard one — 56 pF for trap resonance at 14.100 MHz. This was obtained from

$$C(\mu F) = \frac{1}{2\pi f(MHz) X_c} \quad \text{(Eq. 1)}$$

Hence

$$C = \frac{1}{6.28 \times 14.1 \times 200}$$

$$= 0.0000564 \mu F \text{ (56 pF)}$$

Since X_c and X_L are equal at resonance, the coil was calculated by means of Eq. 2:

$$L(\mu H) = \frac{X_L}{2\pi f(MHz)} \quad \text{(Eq. 2)}$$

Hence

$$L = \frac{200}{6.28 \times 14.1} = 2.25 \mu H$$

(approximate)

The value of the coil will have to be adjusted slightly after the trap is assembled to allow for capacitor tolerance and stray capacitance, which accounts for the term "approximate" in Eq. 2.

The Amidon toroid tables were consulted to learn the A_L factor of a T50-6 core (1/2-inch-diameter toroid). The value is 40. From this I calculated the number of turns from

$$\text{Turns} = 100\sqrt{L_{\mu H}/A_L} \quad \text{(Eq. 3)}$$

Hence

$$\text{Turns} = 100\sqrt{2.25/40} = 23.7$$

For practical reasons a 24-turn winding was used: A partial turn is not convenient on a toroid form.

The same procedure was used for the remaining traps in my antenna. This article is not a course in basic math, but the equations can be useful to those who have not previously designed resonant circuits or used toroidal cores.

Toroidal-Trap Adjustment

It's best to use the largest size wire that will fit easily on the toroid core. The stiffness of the heavier magnet wire will help to keep the coil turns in place, thereby minimizing detuning. I used no. 24 enameled wire.

The capacitor leads and coil "pigtails" should be kept as short as possible. Fig. 3 illustrates the layout I used. The leads at each end of the mica capacitor are soldered to the related coil leads before final adjustment is made.

A dip meter can be used to determine the resonant frequency of the trap, as shown in Fig. 4. Although a prominent feature of a toroidal coil is the self-shielding characteristic, which makes it difficult for us to get ample coupling with a dip meter, it is possible to read a dip. I have found that by inserting the dip-meter coil into the area of the winding gap on the tuned circuit (Fig. 4) a dip can be obtained. By approaching the trap from different angles, it should be easy to find a spot where a dip can be read on the meter. Once the dip is found, back off the instrument until the dip is barely discernible (the minimum coupling point). Monitor the dip-meter signal on a calibrated receiver to learn the resonant frequency of the trap.

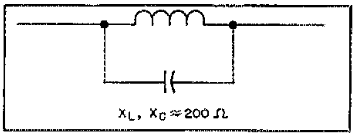


Fig. 2 — Electrical equivalent of an antenna trap. The ac resistance is not shown. A suitable reactance value for the coil and capacitor is 200 ohms.

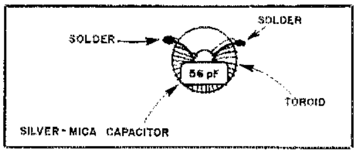


Fig. 3 — Physical arrangement for one of the toroidal L-C traps. Put spaghetti tubing over the capacitor leads to prevent them from shorting to the turns on the toroid.

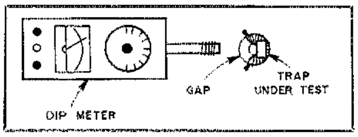


Fig. 4 — Test method for finding the resonant frequency of a trap. Different points around the toroid will yield better dip indications. Experiment with the position of the dipper coil.

Select a part of the related amateur band for trap resonance. I adjust my traps for the center of the frequency spread I am most interested in. For example, I set my 20-meter traps for resonance at 14.025 MHz because I work only cw from 14.000 to 14.050 MHz. For phone-band coverage, I'd pick 14.275 MHz as the trap frequency. A compromise frequency for phone and cw operation would be 14.100 MHz. Owing to the trap Q, coverage of an entire band is not possible without having an SWR of 2:1 or greater at the band-edge extremes. The absolute bandwidth will depend on the trap Q and the Q of the antenna itself.

If the trap is not on the desired frequency, move the turns of the toroid coil farther apart to raise the frequency. Push them closer together to lower the frequency. An alternative method for finding the trap resonance is shown in Fig. 5. The trap being tested is connected to terminals x and y. The coupling is very light in order to prevent the test-circuit capacitance from appearing in parallel with the trap. For this reason the coupling capacitors are only 2 pF. The station transmitter is adjusted for the lowest power output that will provide a reading on M1. The VFO is then swept manually across the band. When the resonant frequency of the trap is located, the meter (M1) will deflect upward sharply, indicating resonance. Adjust the trap for a frequency that is approximately 5% lower than the desired one. This will compensate for the shunt capacitance presented by the 2-pF coupling capacitors.

When the coil turns are set in the correct manner, spread a bead of fast-drying epoxy cement across the turns on the two flat sides of the toroid. This will prevent unwanted position changes that could cause a shift in resonance later on from handling.

Housing the Mini Trap

I learned that a 7/8-inch-OD PVC plumbing coupling, 1-1/4 inches long, would serve nicely as a housing for the toroidal traps.[1] A ridge inside the couplings at the center can be filed out easily to provide clearance for the trap. A rat-tail file does the job quickly. Fig. 6 shows a breakaway view of how the trap is assembled. Slices of dowel rod are used for end plugs. A knot is tied in the antenna wire that enters the trap housing; this prevents strain on the trap coil.

After the antenna wire has been soldered to the trap at each end, add a layer of epoxy glue to the outer perimeter of one of the dowel plugs, then insert it into the PVC coupling until it is flush. Fill the coupling with noncorrosive sealant; I used aquarium cement. Finally, place epoxy glue on the remaining end plug and insert it in the PVC coupling. Allow the trap to set for 48 hours, until the sealant has hardened. Fig. 7 is a photograph of a mini trap, along with a dipole center insulator made from a PVC T-coupling. The coupling is filled with sealant after the wires are soldered to the coaxial feed line. Long plugs are used to close the three open ends of the T connector. A closed nylon loop, made from strong spaghetti tubing, was fed through two small holes at the top of the T-coupling to permit erecting the dipole as an inverted V. A small eye bolt and nut could have been used instead.

There was a minor downward shift in trap resonance after the sealant hardened. Both 20-meter traps shifted roughly 30 kHz lower. No doubt this was caused by increased distributed capacitance across the coil turns with the sealant in place.

[1]Notes appear on page 18.

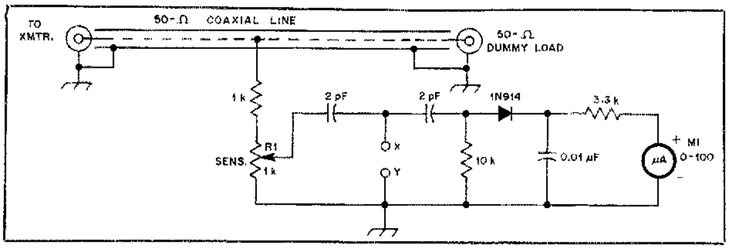


Fig. 5 — Test fixture suitable for checking trap resonance with the station transmitter. Use the least amount of power necessary for meter deflection.

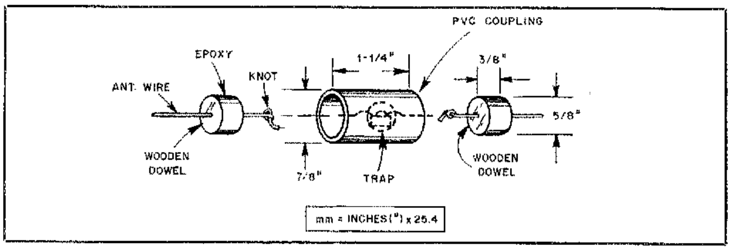


Fig. 6 — Breakaway view of a toroidal mini trap. The knots in the wire prevent stress on the tuned circuit.

This seemed to have no effect on the trap quality; it had a measured parallel resistance of 25 kΩ before and after encapsulation (using the laboratory RX meter for tests). Generally, anything greater than 10 kΩ is suitable for an antenna trap.

Mini Coaxial-Cable Traps

Two very interesting articles concerning antenna traps appeared in the amateur literature during 1981.[2,3] After reading them a second time, I decided to attempt building some traps along the lines discussed in those articles. Some advantages over the usual coil/capacitor style of trap were described by the authors: (1) The traps were not especially frequency sensitive to changes in temperature and climate; (2) the coaxial trap offers greater effective bandwidth; and (3) parallel resistance is quite high — on the order of 50 kΩ.

The articles under discussion contained practical information about the use of RG-58/U and RG-8/U cable for the trap coils. I wanted a small, lightweight trap, so elected to see what could be done with miniature cable — RG-174/U. A completed mini coaxial trap for 20 meters is shown in Fig. 8.

The principle of operation is covered well by O'Neil (note 2). Since this article deals with the practical aspects of traps, we won't delve into the electrical characteristics of the coaxial trap too deeply. However, a diagram showing how it is hooked up is offered in Fig. 9B. A length of coaxial line is wound on a coil form, and the inner conductor at one end is attached to the outer conductor at the opposite end. The distributed capacitance of the two conductors and the inductance of the coil combine to provide a resonant circuit. An acceptable Q results, and the trap can accommodate considerable rf voltage and current without being damaged. A parallel resistance of 50 kΩ was measured for the 20-meter trap of Fig. 8. The bandwidth at the 10 kΩ points was somewhat greater than with the toroidal trap.

Coaxial-Trap Assembly

I found 5/8-inch-OD PVC plumbing pipe to be an acceptable and low-cost material for the coaxial traps. End plugs made from 1/2-inch wooden dowel fit snugly inside the PVC pipe. The completed trap contains a length of bus wire inside it for connecting the braid and center conductor of the cable together, as discussed earlier. The ends of the bus wire and the related cable ends are routed outside the PVC tubing through small holes, then soldered. Aquarium cement was again used, this time to seal the six small holes drilled in the tubing. Epoxy cement was applied to the sides of the wooden plugs before inserting them into the tubing. A layer of vinyl electrical tape can be wound over the coaxial coil if desired, although this should not be necessary. If weather protection is desired, a coating of exterior polyurethane varnish can be applied to the completed close-wound coil. This will keep the turns affixed in the desired position after final adjustment. Tune-up is carried out in the same manner as prescribed for the toroidal traps, using a dip meter or the test fixture described in Fig. 5.

The length of the coaxial cable used will have to be determined experimentally. My 20-meter coaxial trap contains 15 close-wound turns of RG-174 cable (36 inches, 89 pF) to provide resonance at 14.100 MHz. Final adjustment was done by moving the three outer turns at one end until the desired frequency was noted. The coil form for the 20-meter trap is 2-1/2 inches long. The wooden end plugs are 3/8 inch thick. The inside of this trap is not filled with sealant, but it could be if desired. Avoiding the use of filler will make the traps lighter in weight, thereby permitting the use of lighter-gauge wire for the antenna sections.

Trap Performance

Both styles of trap were subjected to rf power tests to determine whether they could handle the output of a typical 150-W class transceiver. A Bird wattmeter was connected between the trap and the transmitter. A 50-ohm dummy load was attached to the opposite end of the trap. Next, 40- and 80-meter rf energy was applied (in separate tests) gradually while observing the reflected power, which of course was not conducive to providing an SWR of 1:1 with the trap in the line. Neither trap showed signs of heating or breakdown at power levels up to 150 W. A key-down period of five minutes was tried during the tests, using a linear amplifier adjusted for 150-W output. Still no sign of power limitation. The SWR did not change under these conditions. I did not advance the power beyond 150 W, but it's safe to conclude that the coaxial-cable trap could sustain substantially more power without damage. This may not be true of the toroidal trap. I lacked the courage to find out!

Toward a Lightweight Dipole

Having solved the problem of lightweight, small traps I set about the task of reducing the bulk of the remainder of my multiband dipole. I am a dedicated miser, so the cost of materials was an important factor in the selection of wire and end insulators. I recalled a type of wire I had used on a number of DXpeditions: It was strong and light in weight, and the price was right! This wire is available from Radio Shack and similar outlets for use as speaker cable. It has a clear plastic outer covering, contains a no. 22 conductor (two each) and costs less than $5 per 100 feet. Hence, for this price we end up with 200 feet of wire (less than 2.5 cents per foot); the parallel conductors can be pulled apart easily without harming the outer insulation. In addition to the insulation aiding the strength of the wire portions of the antenna, it protects the copper

Fig. 7 — View of a toroidal mini trap, an encapsulated toroid and a PVC T-coupling for use as a center insulator. RG-58/U cable is shown in this example (see text).

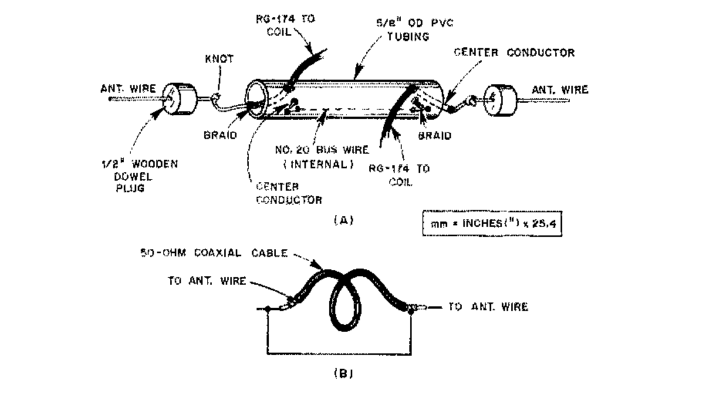


Fig. 9 — At A is a breakaway view of a coaxial trap. The illustration at B shows the electrical connections for a coaxial trap.

Fig. 8 — A completed 20-meter coaxial trap with miniature RG-174/U coaxial cable.

wire from corrosion. This can be especially beneficial in areas where salt water and industrial pollutants affect the atmosphere. The Radio Shack number for this wire is 278-1385. I have observed no apparent deterioration of this type of conductor, even though some of my antennas have been aloft for three years.

Although RG-58/U coaxial cable is less offensive in terms of loss per 100 feet than is true of RG-174/U, we may want to trade losses for portability by using '174. Normally, a 50-foot length of feeder cable is adequate for portable work. In an effort to determine exactly what the hf-band losses per 50 feet might be, I tested this cable from 3.5 through 29 MHz. A Bird wattmeter was connected to each end of the 50-foot test cable. One wattmeter was terminated with a 50-ohm dummy load, and the other wattmeter was connected to a transmitter. The loss in decibels was as follows: 3.5 MHz — 1.19; 7.0 MHz — 1.42; 14.0 MHz — 1.67; 21 MHz — 1.93; 29 MHz — 2.0. Therefore, in a worst-case situation (10 meters), a 100-W power input to the cable would result in an antenna feed-point power of 63 W. RG-58/U, on the other hand, would have a 1-dB loss at 29 MHz, which would mean an antenna feed-point power of 79.4 W. This is not too significant when operating in the 50-150 W range, but it can be important when using a QRP rig with only a few watts or milliwatts of output power. I must say in defense of RG-174/U cable that I operated 20-meter cw with 2 W of output power from 8P6EU while using a dipole with 50 feet of RG-174/U feed line, and I worked the world without difficulty. I received many RST 599 signal reports. The tiny feeder cable and the hookup-wire dipole could be rolled up and stuffed in my hip pocket! The end and center insulators for that antenna were also lightweight. I made them from scraps of pc board from which the copper had been removed. The end insulators for the trap dipole discussed in this article were fashioned from inch-long pieces of 5/8-inch-diameter PVC tubing through which holes were drilled to accommodate the dipole wires and nylon guy lines.

Summary Comments

The overall length of any dipole section in a trap type of antenna will be less than if the dipole were cut for a single band without traps. The exception is the first dipole section after the feed point (out to the first set of traps). The following percentages (approximate) were typical in a coaxial-trap dipole I built for use from 40 through 10 meters, compared to the length of a full dipole (100%) for each band: 10 meters — 100%; 15 meters — 92.4%; 20 meters — 88.8%; 40 meters — 83.6%. The shortening becomes more pronounced as the frequency is lowered, owing to the cumulative loading effects of the traps.

These percentages can be applied during initial structuring of the antenna. Starting with the highest band, the dipole sections for each frequency of interest are trimmed or lengthened for the lowest attainable SWR. After the exact dimensions are known, continuous lengths of wire can be used between the traps. This will add strength to the antenna by avoiding breaks in the speaker-wire insulation, if that type of conductor is used. The percentage reductions listed above are not necessarily applicable to antennas that use toroidal or other coil/capacitor traps. The wire diameter and insulation may also affect the final dimensions of the dipole.

For long-term installations, I would suggest the use of some type of sealer (spar varnish or polyurethane) over the wooden end plugs of the traps. All trap holes need to be sealed securely to prevent moisture from building up inside them.

Miniature antenna traps and lightweight trap dipole antennas are practical and inexpensive to build. Try one during your next vacation or business trip.

Notes

[1] mm = in. × 25.4; m = ft × 0.3048.

[2] G. O'Neil, "Trapping the Mysteries of Trapped Antennas," *Ham Radio*, Oct. 1981, p. 10.

[3] R. Johns, "Coaxial Cable Antenna Traps," *QST*, May 1981, p. 15.

Strays

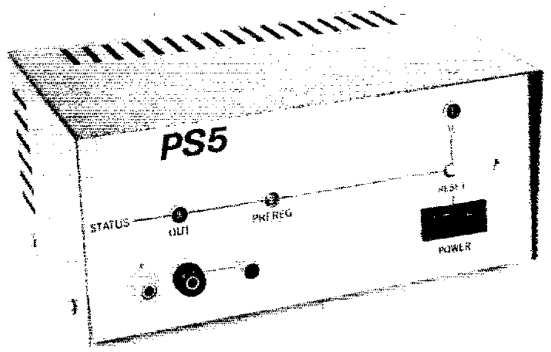
HQ. EMPLOYMENT OPPORTUNITY

☐ There is an immediate full-time opening in the Training Branch at ARRL Hq. — four-day work week, salary and full benefits. An Extra Class license and broad experience in Amateur Radio and education are preferred; solid writing and editing skills are required. If you are interested in this opportunity to help others improve their skills in Amateur Radio instruction, send your resumes and writing samples to "Training Branch" at ARRL Hq.

I would like to get in touch with...

☐ former members of the Communications Division, *USS Volans*, AKS-9, South Pacific area in 1944-45. "Oleo" Boland, K1MZN, 83 Darling St., Central Falls, RI 02863.

Introducing the PS5 — A Dependable, 5-A Portable Power Supply



This compact 5-20 V dc unit is versatile, rugged and practically indestructible. Why not build one to power your portable transceiver?

By Gerald B. Hull,* AK4L

Anytime I set out on a "hamcation" (the combination of a vacation and a portable station operation) I try to gather the proper equipment for the trip. This usually includes a small (25-W or less) transceiver, an antenna system, an ac power source and various station accessories. Perhaps the most difficult item to locate is a suitable 12- to 14-V power supply. Many supplies designed for 100-W transceivers are bulky and heavy; small supplies designed for CB sets or automobile cassette players have poor regulation circuits (very few have over-voltage protection). Still others have sharp cabinet edges or heat-sink fins that can catch on clothing or scratch equipment enclosures.

My requirements for a portable-station supply would include a 13.8-V dc output at up to 5 A of current, short-circuit and over-voltage protection, and small size. Since the majority of commercial supplies did not fit my exact requirements, I decided to design and build my own.

Many commercial and home-built supplies in the 5-A range use a fixed low-current regulator IC or Zener diode and a "wrap-around" pass transistor for regulation (Fig. 1A). This arrangement works well — unless the output is accidentally short-circuited. A short circuit can cause the pass transistor to fail, sometimes with an emitter-to-collector short. This results in the full preregulator voltage appearing at the supply output, which can destroy solid-state components. The better units offer a "sorta" protection circuit for over voltage — an SCR "crow-bar" that shorts the supply output and blows the line fuse when an over-voltage condition is sensed (Fig. 1B). I call them "sorta" protection circuits because they have a response time of a few milliseconds — more than enough time to destroy semiconductor junctions.

*Assistant Technical Editor

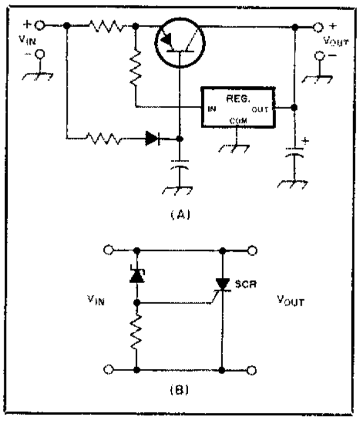


Fig. 1 — A common "wrap-around" regulator circuit is shown at A. An SCR crowbar circuit, which shorts the output when the input exceeds the Zener voltage, is shown at B.

Pass-transistor regulation pitfalls can be avoided by using a high-power hybrid regulator. Fairchild Camera and Instrument Corp. has been a leader in hybrid regulators for years. Fairchild offers the μA78H series of positive regulators in the current range desired for my supply. These units are specified with output currents in excess of 5 A, thermal overload and short-circuit protection, and 50 W of power dissipation. Since I require 13.8-V output, I decided to use the μA78HGASC 5- to 24-V adjustable device as the heart of my design.

Fig. 2 is the schematic diagram for my portable supply. Several components are included to aid the operator. Instead of a line fuse, a small circuit breaker is used to provide the primary circuit protection — this prevents running out of fuses. Three LED indicators monitor different sections of the supply circuit and warn the operator of impending trouble. The PREREG LED indicates output from the rectifier/filter, and the OUT LED indicates output from the regulator IC. If the OUTput LED is off and all others remain on, there is either a direct short at the output or the regulator IC has shut down. The AC LED, of course, indicates that the supply is turned on, but it also indicates the status of the circuit breaker.

Construction

Several special components are used in the PS5. The voltage-setting potentiometer, R1, is a 10-turn precision trimmer. *Do not* substitute a single-turn thumbwheel unit, as they are mechanically unstable and hard to adjust! The power transformer, T1, is a low-profile Hammond unit; others may be substituted, but it may be difficult to locate a replacement with the same current capacity in a small package.

The power supply is built in a Radio Shack aluminum cabinet using point-to-point wiring. *Fig. 3 shows the general* layout of the components. A "universal" heat sink, mounted on the *inside* rear panel,

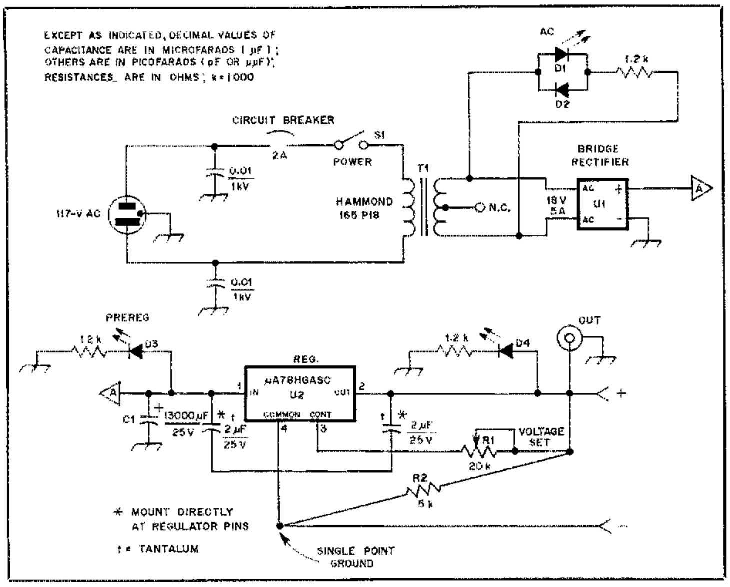


Fig. 2 — Schematic diagram of the PS5 supply. The circuit breaker is a 2-A TV replacement type. Numbers in parentheses are Radio Shack catalog numbers.

D1, D3, D4 — Miniature light-emitting diodes (276-068).
D2 — 1N4001 silicon rectifier diode.
U1 — 25-A, 50 PIV silicon full-wave bridge rectifier (276-1185).
U2 — 5-A, 5-25 V adjustable voltage-regulator IC, Fairchild-type µA78HGASC.
R1 — 20-kΩ 10-turn potentiometer, Bourns type 3006 or equiv.

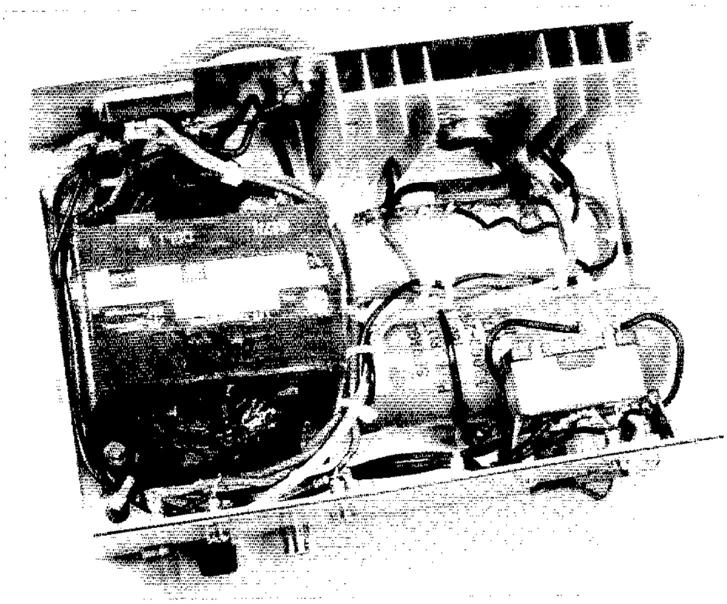


Fig. 3 — Internal layout of the PS5. See text for details.

is used to cool the regulator. Be sure to use heat-sink compound between the regulator/rectifier bridge and the heat sink/cabinet. The most important section of the supply wiring is from the regulator output pins to the output jacks. In the prototype, I used a combination of twisted pairs of no. 14 wire and a solid no. 12 wire.

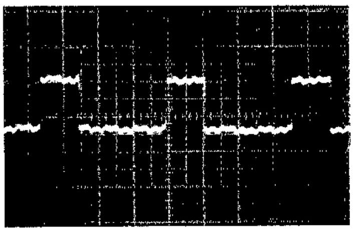


Fig. 4 — Oscilloscope presentation of load regulation in the supply. In this test, a 5.2-A load was pulsed on and off. The vertical divisions on the 'scope are 100 mV.

The dynamic output impedance of my supply is only 20 *milliohms*; therefore, I must stress the need for low-resistance wiring from the regulator to output. As shown in Fig. 2, there are direct connections from the regulator sense resistors, R1 and R2, to the output terminals. By making these connections directly to the output, the voltage drop across the regulator output wires is included in the feedback loop, improving regulation at the load.

Operation

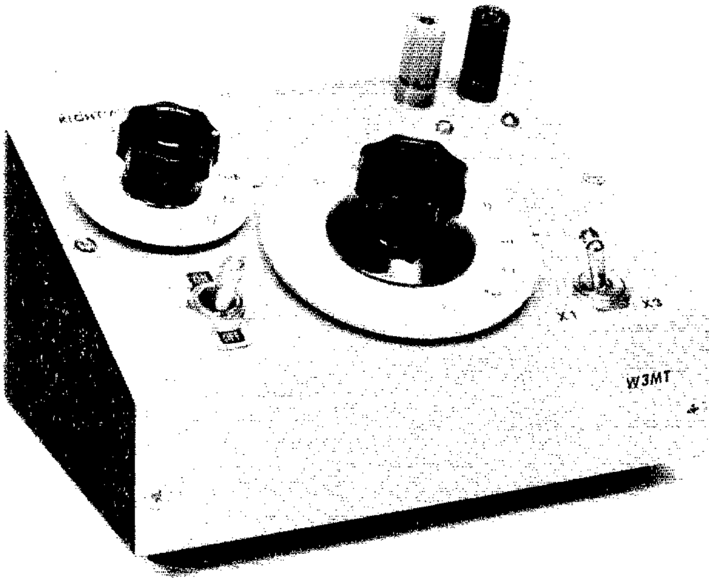
After wiring the supply, *check* and *recheck* your wiring. The short-circuit current available from the power transformer can ignite small-gauge wiring before the breaker has time to open. When you are sure the wiring is correct, locate a multimeter and set it for voltage measurement. Plug the supply in and turn it on. Both the AC and PREREG LEDs should light (the OUT LED will not light unless the regulator output is above 5 V). Measure the voltage at the filter capacitor — it should be very close to 25.0. Move the meter leads to the output terminals and adjust R1 for the desired output voltage. The µA78HGASC will hold regulation, in this circuit, within the range of 5 to 20 V. *Caution*: If you run the regulator at less than 13-V output, be sure to calculate the regulator power dissipation — $P_D = I_{load} \times (V_{in}/V_{out})$. If the dissipation is over 50 W, the regulator could shut down.

Fig. 4 shows the output regulation and ripple in the PS5. A 5.2-A load was pulsed on and off. As you can see, the no-load to full-load regulation is better than 1% and the ripple is less than 20 mV.

Conclusions

With a little design effort a power supply can be simple, yet be a super performer. Commercial products do not fill all needs! The PS5 certainly fits the bill as a dependable power source for my portable station. Now that I have the supply, it's time to plan the next hamcation. Perhaps the Pacific, or the Caribbean, or maybe even Vermont . . .

A Shifty-Eyed Resistance Bridge



When was the last time a test instrument winked at you?

By Frank Noble,* W3MT

The conventional dc Wheatstone Bridge has the virtues of accuracy and simplicity. Its major shortcoming is the requirement of a sensitive meter, which must be protected from overload. Because of this, ohmmeters remain preferred over bridges for measuring resistance — in spite of lower accuracy.

By replacing the balance meter, the Wheatstone Bridge becomes a much more practical circuit. An alternative balance indicator could be a device that only recognizes polarity, since the output polarity of a bridge reverses as the balance point is passed through. A *comparator* is such a device; it switches output state with an input signal change of a few millivolts either side of the threshold voltage. The comparator output state can be indicated through the use of an LED. In fact, if a dual red-green LED is used, the diode color can be made to change at the bridge balance point. To simplify operation, the bridge circuit can be arranged so that a red indication means to turn the knobs clockwise, or "right on red." These are the features I incorporated into my meterless Wheatstone Bridge. Hence, the name "Shifty-Eyed Resistance Bridge."

Circuit Details

As shown in Fig. 1, the bridge is a simple Wheatstone circuit using 5%-tolerance resistors and an Ohmite AB potentiometer as the balance control (1%-tolerance resistors may be used for

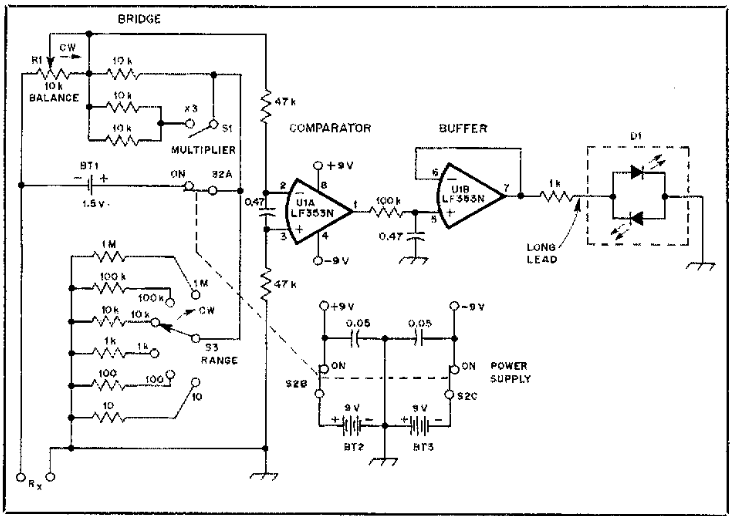


Fig. 1 — Schematic diagram of the Shifty-Eyed Resistance Bridge. Resistors are carbon-composition, 1/2-W, 5%-tolerance or better; capacitors are disc-ceramic or Mylar® 50-V units.

D1 — Tri-color LED, Radio Shack 276-035 or equivalent.
R1 — 10-kΩ, linear-taper potentiometer, Ohmite® series AB or equivalent.
S1 — Spst toggle.
S2 — 3pst toggle.
S3 — Sp6t non-shorting rotary type.
U1 — LF353N biFET dual op amp, Radio Shack 276-1715 or equivalent.

*1004 Belhaven Rd., Bethesda, MD 20817

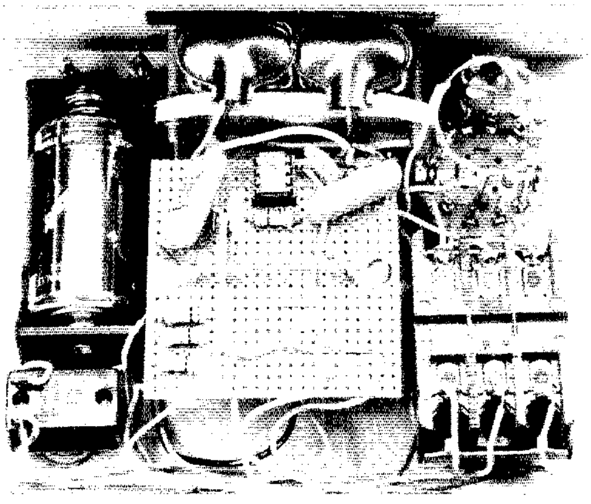


Fig. 2 — Interior of the Shifty-Eyed Resistance Bridge. The unit is built into a Radio Shack 270-264 sloping chassis.

higher accuracy). Switch S1 increases the range by a factor of three, so the bridge has a total of 12 ranges. Resistors from 1 Ω to 3 MΩ may be measured; the accuracy depends on the bridge components used.

The bridge supply voltage should be kept as large as possible, while staying within the dissipation limit of the resistor under test. Increasing the bridge supply voltage is equivalent to increasing comparator gain, which reduces offset and drift. Since dissipation is greatest in the lowest resistance range, the ambitious experimenter might increase the bridge voltage with increasing range by ganging a switch and voltage divider circuit with S3.

Any op amp in an open-loop configuration will serve as the comparator. Under these conditions, the output impedance is relatively high. Although it can directly drive an LED, it is only a little more complicated to buffer the comparator with a follower, which has an extremely low output impedance. The comparator is then thoroughly isolated, with no chance of any loading effects. I selected an LF353N biFET dual op amp for this application because it has very high input impedance, low offset voltage and current, and minimal drift.

In a resistance bridge, high-frequency response is undesirable because it makes the comparator noisy, and will also allow balance-potentiometer wiper noise to interfere with the null indication. Accordingly, I shunted the comparator (U1A) input with a large capacitor and fed it from the bridge through two 47-kΩ resistors, providing a measure of offset balance in the measurement ranges of 10 kΩ and below. The buffer (U1B) is driven through a high-impedance low-pass network, which unloads the comparator and further reduces the bandwidth. The dual LED is wired so that it will change color when the comparator switches output state.

Calibration and Use

The BALANCE control scale is made from a cardboard disc, which is glued to the knob. It may be calibrated by comparing the resistance bridge to a digital ohmmeter of known accuracy. Alternatively, 1%-tolerance resistors can be used as reference standards for calibration.

If the dual LED is wired correctly, a red glow indicates that the BALANCE control must be rotated clockwise to find the unknown resistance value; a green glow indicates that a counterclockwise motion is necessary.

At the balance point, the LED color changes for a very small rotation of the BALANCE control. If this point occurs near either end of the potentiometer travel, S1 may be closed and balance again obtained in a mid-scale position. The resistance is read from the BALANCE-control dial, with regard for the RANGE and MULTIPLIER switch positions.

That's all there is to it! Put a Shifty-Eye together and enjoy the meterless way of measurement. Who says a test instrument can't wink at you?

Frank Noble, W3MT, has done most of his operating on cw since first becoming licensed in 1934. During school days, he obtained both the First Class Telephone and Second Class Telegraph licenses, and gained experience as an engineer for commercial a-m broadcast and police radio stations. He later earned his BSEE and MSEE degrees from Cornell. During WWII, he served as a communications and radar specialist with the Air Corps on Saipan. Frank is now retired, following a 26-year-long career in electronics instrumentation for medical research at the National Institutes of Health. His current hobbies include hunting, fishing and cabin building in West Virginia.

Strays

ARTHUR GODFREY, K4LIB

☐ Former radio and television entertainer Arthur Godfrey, K4LIB, known to millions in the 1950s as the "king of the airwaves," became a Silent Key on March 16. He was 79. K4LIB became a ham through the influence of Curtis LeMay, W6EZV (ex-KØGRL), then the chief of the Strategic Air Command. Godfrey wrote an article in August 1957 *QST* based on his use of Amateur Radio during a big-game hunting expedition to Africa. He also provided his voice for an ARRL film *(The World of Amateur Radio)* in 1979, and was a friend to Amateur Radio.

CONTACT-LENS WEARERS BEWARE!

☐ Now that more people are wearing contact lenses, more hazards peculiar to their use are being discovered. In a recent incident, a shipyard worker wearing plano-type safety glasses over contact lenses was blinded while connecting a welding cable to a 440-V electrical box. When he opened the circuit breaker, the breaker arced and a flash occurred, causing his contact lenses to be "welded" to the cornea of the eye. Doctors later found that the lenses had served to concentrate the heat from the arc flash onto the cornea. Contact lenses must not be worn where a flash can occur or where welding is being done. — Automotive Independent Magazine, *as reprinted in the* APG *(Aberdeen Proving Ground, Maryland)* News; *tnx KA3HMS*

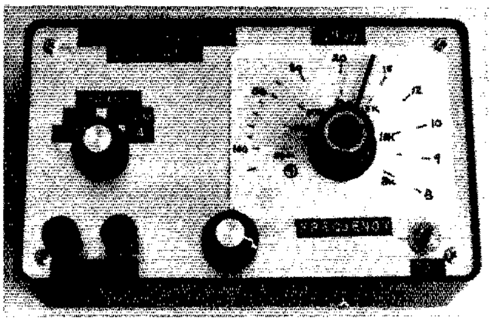
STRAY HINTS

☐ "Strays" are those interesting fillers used when space allows in *QST*. Think you have an item with Stray potential? Here are some hints to help your submission become one. (1) Be sure the information will be of interest to most readers of *QST*. (2) Submit your material before deadline — the 8th of the second month preceding desired publication (i.e. arrive at Hq. before June 8 for August *QST*.) (3) Any photographs you send should be good quality, black-and-white glossy prints. Color prints, slides and instant photos do not usually reproduce well.

Items submitted are normally acknowledged, but that doesn't necessarily mean that your Stray will be appearing in *QST*. We receive far more material than we can find room for. If you want your material returned, please include a statement to that effect and an s.a.s.e.

Follow the above hints and maybe your Stray will find a home in *QST*. — *Andrew Tripp, KA1JGG*

A Wide-Range Variable-Frequency Audio Oscillator



Need a simple and inexpensive audio VFO? This one fills the bill, *and* provides good performance!

By Harry M. Neben,* W9QB

Electronics experimenters (there are still a few of us left!) frequently have the need for an audio-frequency signal generator that has a reasonable output level and relatively low distortion. Such an audio oscillator was described by Shultz,[1] and is included in the Test Equipment and Measurements chapter of the ARRL Handbook.

My variable-frequency audio oscillator is a modification or update of this oscillator, using readily available components. These modifications were the result of difficulty in obtaining a dual, 20-kΩ, linear-taper potentiometer, called for in the original circuit.

Circuit Highlights

The oscillator is a Wien Bridge, or twin-T type (see Fig. 1). If the two R-C arms in this oscillator are of equal electrical value, the frequency is given by

$$f = \frac{1}{6.28 \times R \times C} \quad \text{(Eq. 1)}$$

where
R = resistance in ohms
C = capacitance in farads
F = frequency in hertz

Thus, if the resistance or capacitance is changed by a power of 10, the frequency will be changed by the same factor. This provides a convenient decade-scale relationship.

This concept is great in theory, but when the 741 op amp is operated at fre-

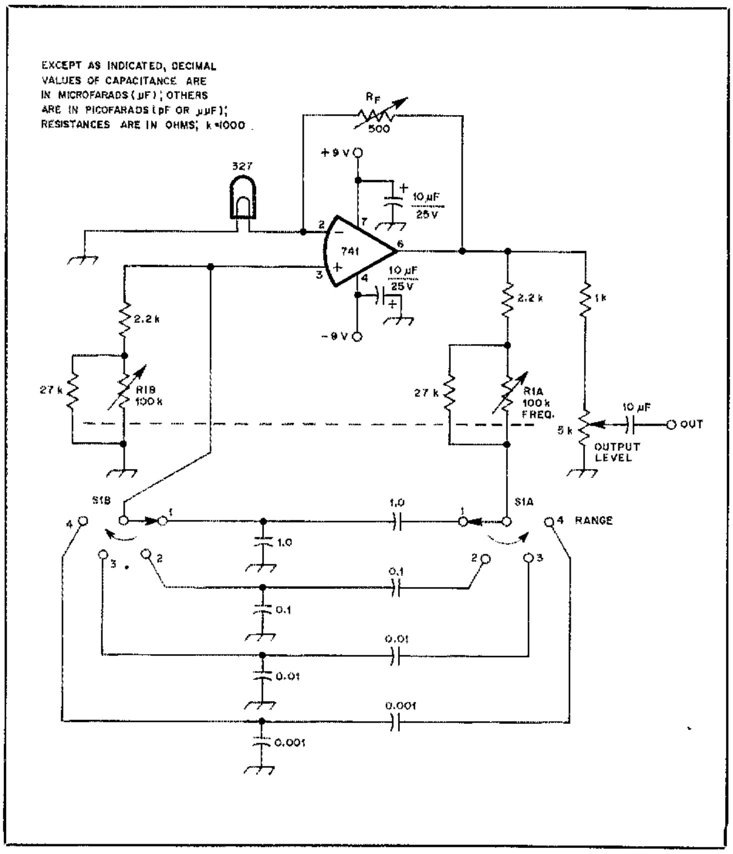


Fig. 1 — Schematic diagram of the Wien-Bridge audio VFO.

[1]Notes appear on page 24.

*1151 Fairway Dr., Dunedin, FL 33528

Table 1
Oscillator Frequency Versus Potentiometer Rotation

Detent	*Rotation (Deg.)*	*R1 Effective Parallel Resistance (ohms)*	*Frequency (Hz)*†		*Detent*	*Rotation (Deg.)*	*R1 Effective Parallel Resistance (ohms)*	*Frequency (Hz)*†
1	7	6	265,317.0		21	154	10,800	147.4
2	15	10	159,214.0		22	161	11,489	138.5
3	22	30	53,110.6		23	168	12,122	131.3
4	29	287	5,547.1		24	176	12,981	122.6
5	37	635	2,507.5		25	183	13,745	115.8
6	44	918	1,734.3		26	190	14,644	108.7
7	51	1,240	1,283.2		27	198	15,242	104.4
8	59	1,599	995.2		28	205	15,955	99.8
9	66	2,034	782.4		29	212	16,732	95.1
10	73	2,455	648.4		30	220	17,532	90.8
11	80	2,861	556.3		31	227	18,217	87.4
12	88	3,635	437.9		32	234	18,809	84.6
13	95	4,638	343.2		33	241	19,326	82.4
14	102	5,559	286.3		34	249	19,782	80.5
15	110	6,580	241.9		35	256	20,187	78.8
16	117	7,297	218.1		36	263	20,491	77.7
17	124	7,816	203.6		37	271	20,796	76.6
18	132	8,308	191.6	Useful	38	278	21,025	75.7
19	139	9,220	172.6	Rotation	39	285	21,073	75.5
20	146	10,047	158.4	Range	40	293	21,121	75.4
				(Approx.)	41	300	21,121	75.4

†Frequency when using 0.1-μF capacitor in the Wien-Bridge circuit.

quencies above 20 kHz, the R-C to frequency relationship is variable, and dependent on the individual 741 used. Regardless, this range is included, but a separate calibration scale is drawn for frequencies above 20 kHz. [This problem can be eliminated by directly substituting a TL081 for the 741 op amp. — Ed.]

The frequency range of the *Handbook* audio oscillator is from 15 Hz to 150 kHz. For my purposes I needed a 10-Hz signal, so the circuit was modified by applying Eq. 1. I used readily available components by doubling the capacitor values to 1.0, 0.1, 0.01 and 0.001 μF. These values provided a low-frequency range starting at about 8 Hz. However, those who want to build this little rascal can choose whatever capacitor values are on hand. Of course, this will change the oscillator frequency range!

My greatest problem encountered with the *Handbook* oscillator was trying to locate a dual 20-kΩ, linear-taper potentiometer. A dual 10-kΩ potentiometer was tried, and all the capacitor values were doubled. This combination was found to be unstable in the 10-Hz region. A dual, 50-kΩ, audio-taper potentiometer was tried. This will work, but the audio taper results in an unsatisfactory logging scale. Another solution was sought.

The local Radio Shack store stocks a dual 100-kΩ potentiometer with an audio taper. This unit has a 41-position detented wheel in it that can be taken off to provide smooth rotation if desired. When a 27-kΩ resistor is placed in parallel with this unit, the equivalent resistance is suitable for use in this audio oscillator (Table 1). However, if the potentiometer is wired into the circuit correctly, the oscillator frequency will *decrease* as the dial is rotated *clockwise*. To the true experimenter this shouldn't matter! Here is a scale that is acceptable, providing about 200° of usable potentiometer rotation. Table 1 can be used to determine the approximate oscillator frequency, according to which detent the potentiometer is turned to.

Other modifications to the *Handbook* circuit are the inclusion of a 10-μF, nonpolarized capacitor in the output to provide isolation from any dc components, and putting a 500-Ω trimmer potentiometer in place of the 220-Ω (R_f) resistor.

Construction and Alignment

Construction of the audio oscillator is simple and straightforward. The unit is built into a box measuring 6 × 4 × 2 inches.[2] Capacitors are wired directly to the rotary switch, and the IC socket and other components are mounted on a small perforated board. Front-panel layout is shown in the photograph. The main tuning knob is made from an ordinary knob with a small piece of clear plastic attached as a pointer.

The only adjustment needed before calibration is in the feedback loop. Connect the oscillator to an oscilloscope and view the wave form. Adjust potentiometer R_f until flat-topping of the sine wave occurs. Slowly back off R_f until a clean sine wave reappears. Check all frequency ranges while sweeping the main frequency control to ensure that no flat-topping occurs at any portion of all ranges. Readjust R_f if necessary.

To calibrate the oscillator, an oscilloscope or frequency counter is needed. All ranges, except the highest frequency one, are related by a factor of 10. If the lowest frequency range is calibrated first, the next two oscillator ranges will also be calibrated. I found it necessary to use a frequency counter to calibrate the highest range (above 20 kHz), since it did not track with the other ranges. The degree of mistracking depends on which op amp is used.

This audio VFO is simple to build and inexpensive, and provides substantial output voltage at a reasonable distortion level. Put one together and you will be delighted!

Notes

[1]T. Schultz, "Audio Oscillator," Hints and Kinks, *QST*, Nov. 1974, p. 43.
[2]mm = in. × 25.4

Next Month in QST

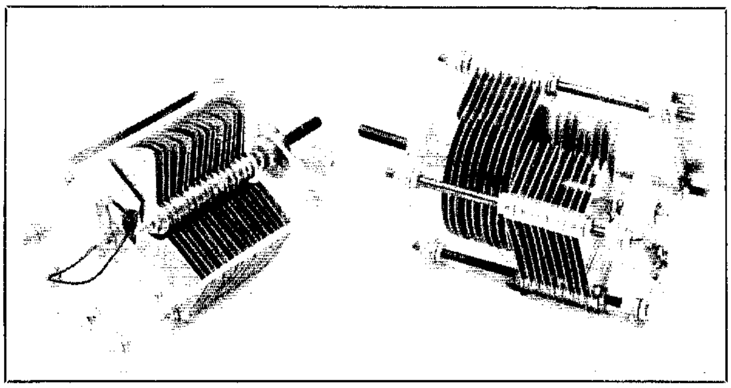
Hams can now use AMTOR. So what's AMTOR, you ask? Why, it's an error-correcting teleprinter protocol. So what, you ask? Find out for yourself, in an article that takes you through the basics of this intriguing new mode.

Also in July *QST*, look for two other articles relevant to those in the forefront of Amateur Radio. One provides a clear, useful explanation of spread spectrum, and the other describes an easy-to-build serial ASCII/Baudot character generator that you can use to test your printer or video terminal. All this good reading, and a great deal more, in July *QST*.

A Homemade High-Power Tuning Capacitor

Transmitting types of variable capacitors are expensive and hard to find! Why not build your own?

By Thomas C. Stephens,* KD6ED



Do you have lots of spare time and good mechanical skills? If so, this project is for you! No longer do you have to pay premium prices for a high-power variable capacitor. One look at the parts list (Table 1) should pique your interest; all the parts should be available from your hardware store and/or junkbox.

First, the mathematics. Before building the capacitor, the number of rotor and stator plates must be calculated. The formula for capacitance between two parallel plates is

$$C = \frac{0.224AK}{d} \quad \text{(Eq. 1)}$$

where
C is the capacitance in picofarads (pF)
A is the plate area in square inches
d is the plate separation in inches
K is the dielectric constant (air = 1)

Table 2 shows typical plate spacing for various peak voltages. I chose a plate spacing of 0.075 inch for my first capacitor — enough to withstand 3000-V peak.[1] Since there is a chance of bending plates, a stack of five no. 8 washers (0.042 inches thick) provide approximately 0.0975-inch spacing between rotor and stator plates, allowing a slight margin for bending. This arrangement is shown in Fig. 1.

To calculate the capacitance per rotor plate, first find the total area of the rotor plate, which includes both sides of the plate minus the center notches (which provide clearance for the rotor shaft) in the stator plates.

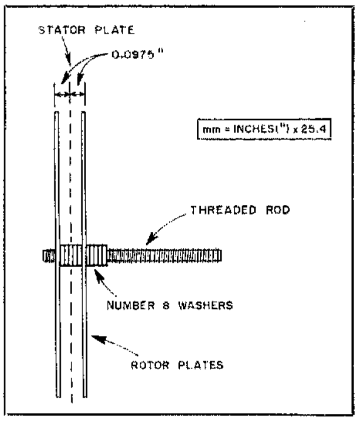


Fig. 1 — No. 8 washers are used to space the rotor and stator plates. See text for details.

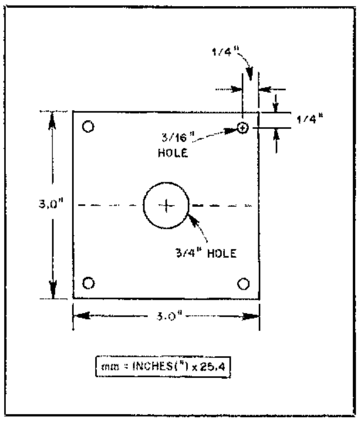


Fig. 2 — Template for the stator, and end-plate jig.

$$A = \pi R^2 \text{ (total plate)} - \pi r^2 \text{ (notch)} \quad \text{(Eq. 2)}$$

where
A is the total area
R is the radius of the total plate
r is the radius of the notch

Substituting the values for my capacitor:

$$A = \pi(1.4375)^2 - \pi(0.375)^2 = 6.05 \text{ square inches} \quad \text{(Eq. 3)}$$

Using Eq. 1, the capacitance per rotor plate can now be calculated:

$$C = \frac{(0.224) \times (6.05) \times 1}{0.0975} = 13.9 \text{ pF} \quad \text{(Eq. 4)}$$

The maximum capacitance of one section of a capacitor (two stator plates plus one rotor plate) is equal to the capacitance of this section at full mesh plus the minimum capacitance, which is about 10% of the maximum. Therefore, each section will have a capacitance of 1.1 × C, or 15.3 pF, in my capacitor.

Construction

Two 3 × 3 × 1/4-inch Plexiglas® plates are used as end supports, and

[1]mm = in. × 25.4.

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Table 1
Parts List for the Homemade Capacitor

Description	*Quantity*
Plexiglas sheet plastic 3 × 3 × 1/4 inch	2
Threaded no. 10 brass rod	5
No. 10 hexnuts	Box of 100 (more than enough; common-size package)
No. 10 brass hexnut	1
Half-hard aluminum sheet, 0.020 inch thick	Depends on number of rotor and stator plates
Copper strip, 1 × 3 × 0.07 inches	1
No. 8 washers, 0.450-inch dia. by 0.042 inch thick	1 lb box

Table 2
Typical Capacitor Plate Spacings

Spacing Inches (mm)	*Peak Voltage*	*Spacing Inches (mm)*	*Peak Voltage*	*Spacing Inches (mm)*	*Peak Voltage*
0.015 (0.4)	1000	0.07 (1.8)	3000	0.175 (4.4)	7000
0.02 (0.5)	1200	0.08 (2.8)	3500	0.25 (6.3)	9000
0.03 (0.8)	1500	0.125 (3.0)	4500	0.35 (8.9)	11,000
0.05 (1.3)	2000	0.15 (3.8)	6000	0.5 (12.7)	13,000

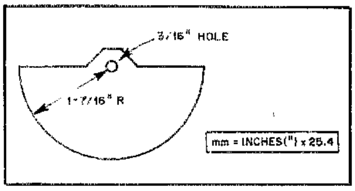

Fig. 3 — Rotor-plate template.

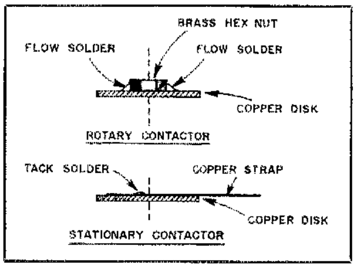

Fig. 4 — Construction details of the stationary and rotary contactors.

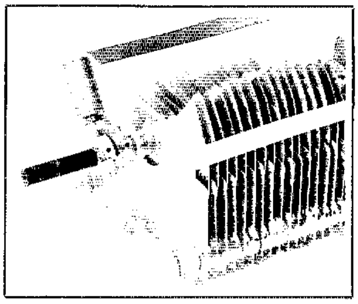

Fig. 5 — Rotor balancing modification. Use enough washers to just balance the capacitor.

no. 10 rods form the capacitor frame. The length of these rods will vary with the size of the capacitor; it is best to use long rods and cut them to size once the unit is completed.

A drilling jig aids in constructing the end and stator plates. The jig outline is shown in Fig. 2. Use a fairly stiff material for the jig, such as phenolic or 0.04-inch aluminum sheet stock. Drill the end plates while using the jig and place a screw in each drilled hole to ensure proper alignment. Cut out and drill the 3-inch aluminum sheets, again while using the jig. These plates will be cut in two to form the stator plates. Once the sheets have been cut and drilled, enlarge the center hole to 3/4 inch (the best method is to use a chassis punch) and cut each sheet in half. The resultant half-moon shape in each stator plate provides clearance for the rotor shaft.

Each rotor plate is made from the same aluminum sheet, using Fig. 3 as a guide. Note that the rotor radius is 1-7/16 inches instead of 1-1/2 inches. This provides clearance for the frame rods. Once the rotor plates have been cut, you can use a file to smooth and round them. One filing method I tried was to stack all the plates on a rod and fan them out to form a circle. The assembly is then tightened with a pair of nuts and placed in an electric drill. While spinning the plates a file is used to smooth the edges. Be *very* careful if you try this method; be sure the drill is secured firmly.

Two contactors are needed to provide an electrical connection for the rotor assembly. These contactors are made from 3/4-inch-diameter copper, brass or silver (best) disks, copper strap and a brass nut. Construction details of the stationary and rotary contactors are shown in Fig. 4.

The rotor shaft is made from a brass no. 10 threaded rod (length depending on the design capacitance) and two 1-1/2 inch lengths of 1/4-inch aluminum shaft material. Drill one end of each aluminum shaft to a depth of 1/2 inch and tap them with a 10-32 *bottoming* tap.

Assembly

Assemble the rotor first. Select one of the no. 10 threaded rods as a rotor shaft, screw one of the threaded aluminum shafts over the end of this rod, then secure it with a drop of glue. Next, slide two washers over the rod and follow with a nut. Tighten the nut slightly; the washers are used as end-play bearings. Slide on another flat washer, followed by a rotor plate, more washers (number depending on chosen spacing) and another rotor plate. Repeat this process until all the plates are in place. Slide one more washer next to the last plate and secure with a nut. Screw on another nut, leaving a 1/16-inch space from the last nut; this nut will serve as a locking device for the contactor assembly. Screw on the rotary contactor so that the flat surface is facing away from the rotor plates. Now, screw on the stationary contactor so that the flat surfaces mate with each other. Finally, screw on the other threaded aluminum shaft. This completes the rotor assembly.

Next, mount the potentiometer bearings to the plastic end plates and install two frame rods in the lower corners of one end plate. Use nuts and washers on both sides of the plate. Install the stator plates on these two parallel rods, using the same procedure as for the rotor plates.

After all the stator plates are mounted, slip the coil spring on the rotor shaft and attach the rear end plate. The end-to-end spacing should be enough to allow some play in the rotor, but the spring should provide a fair amount of tension. Finally, attach the two top frame rods. This completes the assembly of the capacitor.

Alignment

It will take a few minutes of adjusting to ensure free rotation of the capacitor. Some of the plates may have to be bent; it is hard not to bend them when assembling the unit. After building several capacitors, I found they needed a little counterbalancing. This was accomplished with a collar on the rotor shaft that has a 1-1/4 inch set screw. This modification is shown in Fig. 5.

Although this project is time consuming, it surely is cost effective. If you use reasonable care when drilling, cutting and bending, you will have a capacitor of which you can be proud.

Worldwide Beacon Net: The Possibilities Abound

By John G. Troster,* W6ISQ and Cameron G. Pierce,** K6RU

Monitor 14,100 kHz any time, day or night, for 10 minutes and see how many of the eight worldwide power-attenuating beacon net stations you can hear as they transmit automatically one after the other. In that short time, you can get a quick, general appraisal of the various 14-MHz paths that may (or may not) be open around the world, and the direction and the quality of the opening.

One of the most interesting observations is the consistent strength and readability of the 1-W (and even the 0.1-W) dash from the distant beacons. It's enough to make the QRO fellas turn off their finals and the QRPers cheer!

The beacon net was organized and financed by the Northern California DX Foundation (NCDXF). The overall beacon transmitter concept and rf power-level switching was designed by Dave Leeson, W6QHS. Jack Curtis, K6KU, of Curtis Electro Devices, designed the clock, the microprocessor and the programming components. The engineering, production and packaging was done by Cam Pierce, K6RU.

Do-It-Yourself

Besides monitoring, the beacons can also be used for several do-it-yourself practical observations or projects. The beacons

1) provide an "in band" time/frequency standard

2) are a means of comparing antennas or receivers by switching back and forth during the same nine-second dash. Check DX versus short-range characteristics of different antennas. When a beacon signal is steady, rotate the beam antenna (from center to side) to check beamwidth.

3) enhance propagation investigation. Record the lowest signal level copied from each beacon (or note "no copy") hour-to-hour and day-to-day. How do your observations and interpretations of the beacon's reception compare with published propagation forecasts, or correlate with various A K indices, WWV broadcasts, sunspot activity or just generally observed band activity? Is there any correlation at all? Can beacon signals be used to improve forecasts of conditions?

This propagation-observation possibility is the principal purpose of the power-attenuating system. Indeed, it is hoped that amateurs will be stimulated by monitoring the beacon net to learn and investigate more about the vagaries and conditions of propagation and propagation forecasting, and to share that knowledge with others.

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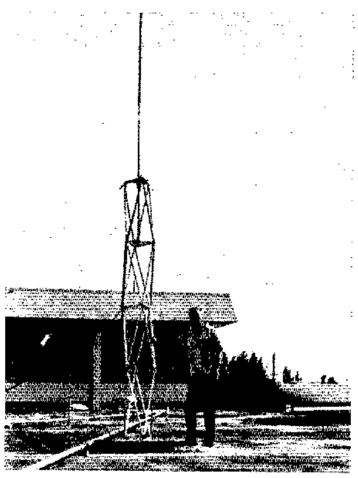


Bob Jones, KH6O, displays the beacon antenna designed and built by his science-class students at Honolulu Community College, one of eight NCDXF beacon sites worldwide.

Table 1
NCDXF Beacons Transmitting Sequence

Time	*Station*	*Location*
0000	4U1UN/B	United Nations, New York
0001	W6WX/B	Stanford University, California
0002	KH6O/B	Honolulu Community College, Hawaii
0003	JA2IGY/B	JARL, Mt. Asama, Japan
0004	4X6TU/B	Tel Aviv University, Israel
0005	OH2B	Helsinki Technical University, Finland
0006	CT3B	ARRM, Madeira Island
0007	ZS6DN/B	Transvaal, South Africa

Same sequence repeats every 10 minutes. Beacons are crystal-controlled and are on 14,100 kHz, First "Q" of each beacon's "QST" begins within a fraction of a second of the assigned time, plus or minus human starting error. The sequence may vary as more beacons are added to the net.

Table 2
Text Transmitted by Each Beacon

Power Level	*CW Message*
100 W	QST de (W6WX/B) beacon
100 W	· ——— 9-second dash
10 W	·· ——— 9-second dash
1 W	··· ——— 9-second dash
0.1 W	···· ——— 9-second dash
100 W	SK (W6WX/B)

Transmission time: ±58 seconds
Speed: 20+ wpm
Power attenuated in 10-dB steps

Total Solar Eclipse Reports Requested

There will be a total solar eclipse in the area of Indonesia on June 11 from 0209.5 to 0715.8Z. San Francisco State University Professor Lou Berman, K6BW, requests that observers around the world continuously monitor all stations in the NCDXF beacon system (14,100 kHz) an hour or so before, during and after the eclipse time. Please send log of signal reports and other radio or solar observations to K6BW, 1020 Laguna Ave., Burlingame, CA 94010. Even a "no change" report would be much appreciated. — *W6ISQ*

Volunteer Observers Needed

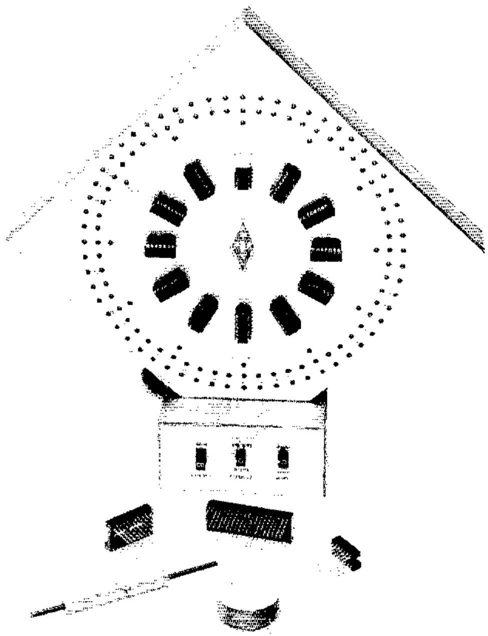
The beacon net is looking for volunteers worldwide who can make regular observations and submit them about once a month. (It would make a good science class project.) The observers' interpretations and conclusions would also be of great interest. Anyone interested should write to Observation Coordinator Al Lotze, W6RQ, 46 Cragmont Ave., San Francisco, CA 94116.

Strays

SAFETY BELTS

☐ All the folderol and a considerable amount of balderdash about using safety belts when climbing the tower is a complete waste of time. Anyone with an ounce of brain will tell you that by the time you lace up the harness and strap the belt on your body you can be up the tower and back down! If anyone would like further info on the folly of safety belts you can drop me a line at this address: R. F. Inshack, Memorial Hospital, 7388 Traction Rd., Brokentutu, Pennsyltucky. — *Tamiami ARC, Venice, Florida*

The Pizza Clock: An Exercise in Wire Wrapping



Wire wrapping is a useful circuit wiring technique for the hobbyist. Add a new "twist" to your home-built gear collection by building this unique conversation-piece.

By Robert Shriner,* WAØUZO and Gerald B. Hull,** AK4L

This article combines education and fun; you'll learn the technique of wire wrapping and build a unique electronic clock in the process. With the advent of ICs that implement almost any function, the process of wire-wrapping circuits has become popular among hobbyists and the industry alike.

What is "Wire Wrapping"?

Wire wrapping is a circuit-wiring process in which connections are made by wrapping wire around a square terminal post. A close-up view of a wire-wrapped connection is shown in Fig. 1. An obvious advantage of this technique is the solderless connection — circuit changes and repairs are made easily. The major disadvantage of wire wrapping is the need for a square post — passive components with round leads (resistors, capacitors, and so on) must be soldered to a square wire-wrap post to ensure a good connection. Wire-wrapped connections are as good as, if not better than, a soldered connection; in fact, aerospace electronics equipment is usually wire-wrapped for improved reliability.

Generally, hobbyists construct their wire-wrapped projects on a punched phenolic board (often called "perf board") or a prototype card, which is a perf board with copper traces for dual in-line package (DIP) IC power supply pins etched on one side. Special square posts are available for use as tie points for wire-wrapping passive components, but the majority of wire wrapping is done with special wire-wrap IC sockets. These sockets come in all the standard DIP sizes, as shown in Fig. 2.

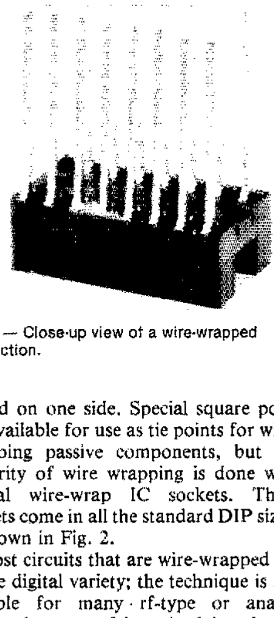


Fig. 1 — Close-up view of a wire-wrapped connection.

Most circuits that are wire-wrapped are of the digital variety; the technique is not suitable for many rf-type or analog designs because of long lead lengths required and large numbers of passive (round lead) components. Wire-wrapped digital circuits can be made to operate into the lower vhf spectrum if care is used when wiring the circuit.

Tools

Wire-wrapping tools come in many shapes and sizes. The most common (and least expensive) is the hand-operated tool, shown in Fig. 3. One end of the tool has a sharp-edged V-shaped groove which strips insulation from wire; the other end has a hollow tube with two holes. The wire is inserted in one hole and the wire-wrap post fits in the other.

Once the wire has been inserted into the tube, the tool is placed over the square post and twisted. These steps are illustrated in Fig. 4A, B and C. Other types of wire-wrapping tools are modifications of the simple hand tool. Some tools strip insulation while wrapping. This feature will allow the user to "daisy-chain" connections; in other words, to connect many terminal posts without stopping to strip wire at each connection. Daisy-chain wrapping is illustrated in Fig. 5. Choosing a tool is basically a matter of cost. Usually, the easier a tool is to use the more it will cost.

Wrapping requires wire with a special size of insulation (to fit the stripping tools). The industry standard wire size is no. 30 with Kynar® (plastic) insulation. Most of the tools are designed to accept this size, although a few models use larger wire. The wire is available in many colors and forms; project construction time can be cut considerably by using precut and stripped wire (Fig. 6).

Wiring a Circuit

A schematic diagram tells a lot about how a circuit can be wired. For example, Fig. 7 is the schematic diagram of our sample wire-wrapping project, the "Pizza

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**Assistant Technical Editor

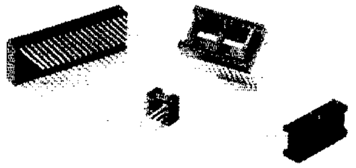


Fig. 2 — Wire-wrap IC sockets.

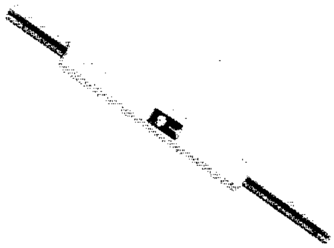


Fig. 3 — Hand-operated wire-wrapping tool.

Clock." Most of this circuit uses the wire-wrap technique, but several portions are soldered together because they are impractical to wire-wrap — these include the power supply section, the indicator LEDs and the time-base oscillator.

When building a project as complex as the Pizza Clock, it is a good idea to make up a "connection table" or wrap chart. Once a project has been completely wired, it is very difficult to check individual connections; a wrap chart provides an invaluable reference during assembly and debugging. A sample of the wrap chart for the Pizza Clock is shown in Table 1.

The Pizza Clock

Building a discrete-IC 24-hour clock seemed to be a good exercise for wire wrapping, but since the cost of 24-hour digital clocks has dropped to the $10-15 range, a clock project using digital readouts would be like "reinventing the wheel." The idea for the Pizza Clock came from a unit of similar design that one of the authors had seen in a store. That unit used TTL ICs and lamps to indicate hours, minutes and seconds. The ac power mains provided the power and frequency reference.

Low-power CMOS logic is used in conjunction with LED indicators to produce a nice-looking clock that is independent of the ac line. Another unique feature of the clock circuit is that it uses exactly 12 ICs — one for each hour.

The schematic diagram of the Pizza Clock is shown in Fig. 7. An MM5369 oscillator/divide-by-59659 chip provides a 60-Hz output from a 3.579-MHz color-burst crystal. U2 and U3 divide this 60-Hz signal by 10 and 6, respectively, which yields a 1-Hz signal for the counter/display chain. U4, U6, U7, U8, U10 and U12 are 4017 decimal counter

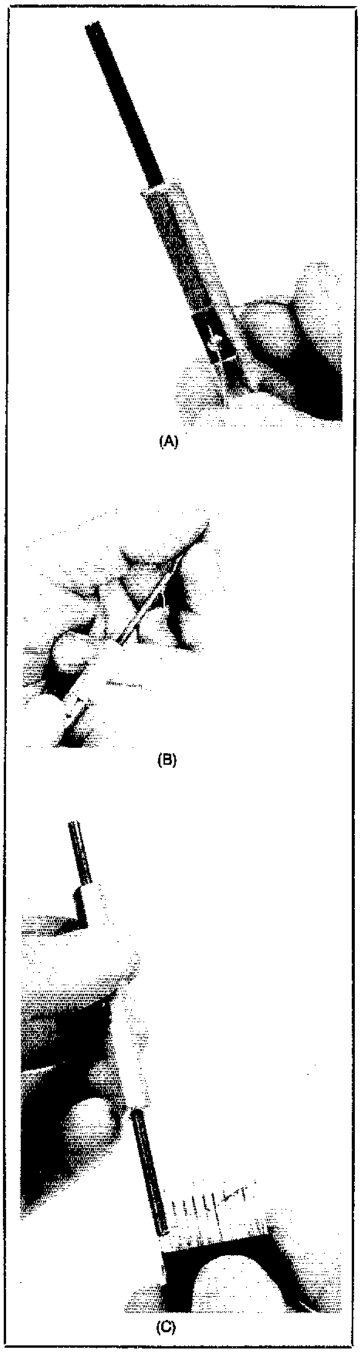


Fig. 4 — The wire-wrapping process — (A) strip insulation from wire (1/2-1 inch); (B) insert wire in tool; (C) place tool over post and twist.

ICs, whose outputs drive the LED indicators. D4 through D33 provide isolation for the counter ICs when the in-

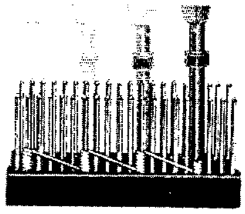


Fig. 5 — Daisy-chain wrapping.

Table 1
Portion of a Wrap Chart for the Pizza Clock

From	*To*		
U1 pin 1	U2 pin 14	U1 pin 4	N.C.
U1 pin 1	S1A	U1 pin 5	R2
U1 pin 2	GND	U1 pin 5	C1
U1 pin 3	N.C.	U1 pin 5	Y1

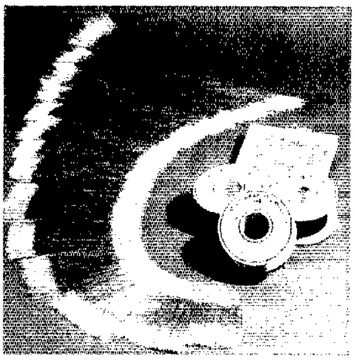


Fig. 6 — No. 30 wire for wrapping comes in many forms. Pre-cut and stripped wire speeds construction time. Photo courtesy of OK Machine and Tool Co.

dicator LEDs are reverse biased. U5 and U9 provide signal inversion to properly drive the LEDs. The minutes and seconds counter circuits are exactly the same; only the hours counter is different — it uses a quad NAND gate (4011) to sequence 12 LEDs instead of 10 and resets the hours-counter circuit.

Construction

The Pizza Clock case is constructed from double-sided printed circuit board; this material provides a pleasing finish *if* care is used during assembly. This is not a one-evening project: Take your time and you will be rewarded by a good-looking station accessory.

Fig. 8 and Table 2 show details of the pc-board cabinet and base. The faceplate of the clock is etched with the IC pinout and power supply bus on one side and the

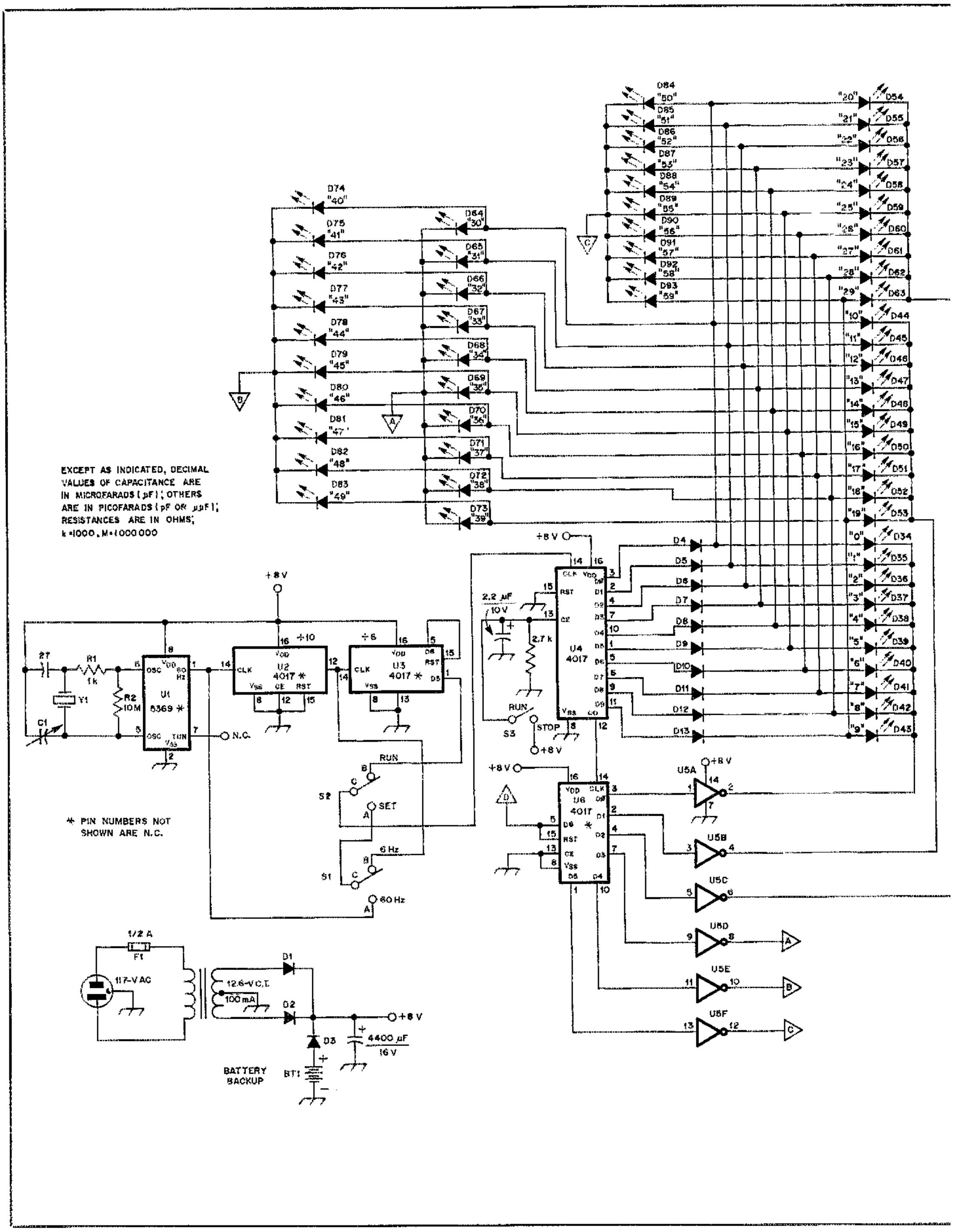
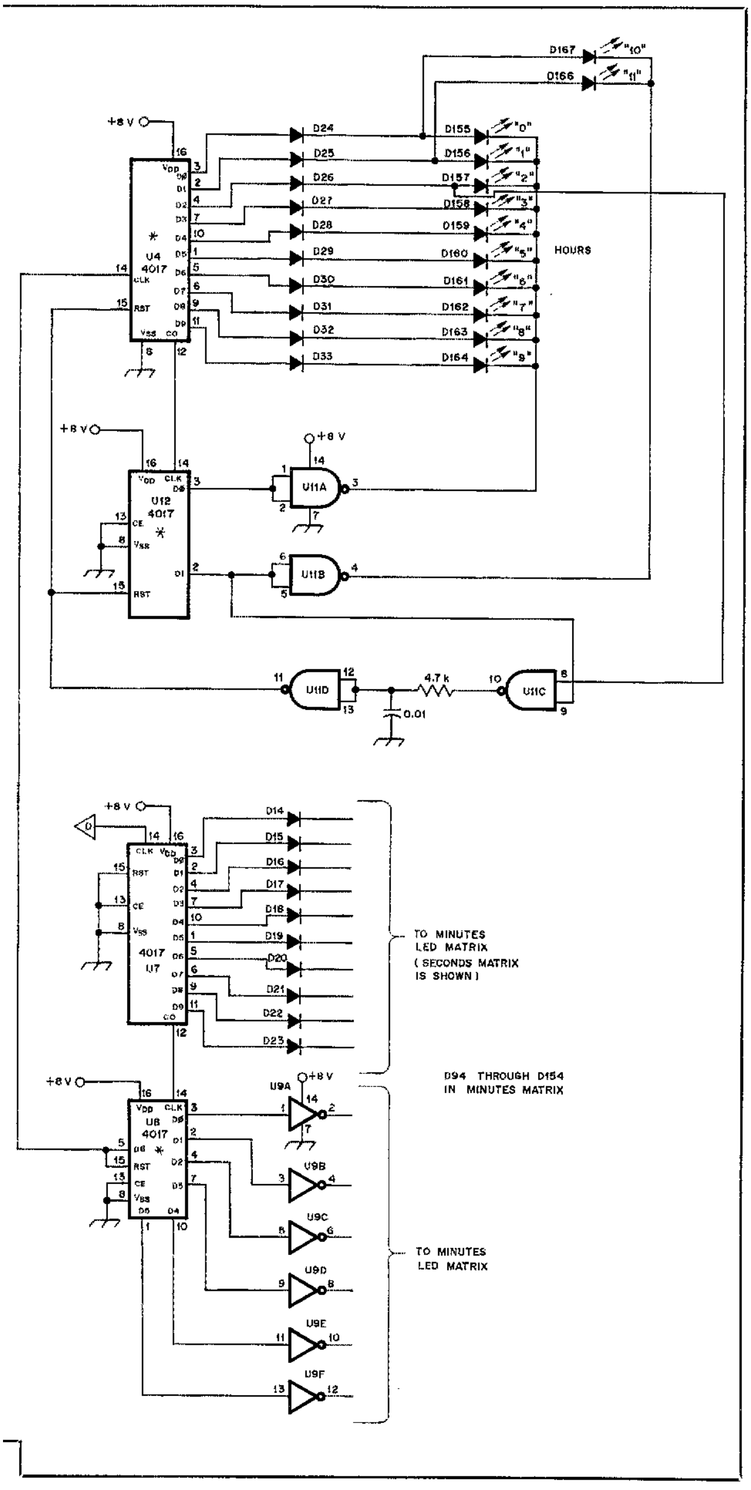


Fig. 7 — Schematic diagram of the Pizza Clock. Fixed-value capacitors are disc ceramic unless they are polarized; these are electrolytic. Resistors are 1/4-watt unless otherwise specified.

C1 — 5-30 pF plastic trimmer capacitor.
D1-D3, incl. — 1N4001 silicon power diode.
D4-D33, incl. — 1N914 silicon switching diode.
D34-D166, incl. — red light-emitting diode.
S1-S2 — SPDT slide switch.
S3 — spst slide switch.
U1 — MM5369 oscillator/divider IC.



U2, U3, U4, U6, U7, U8, U10, U12 — 4017 CMOS decimal counter IC.

U5, U9 — 4049 CMOS hex inverting buffer IC.
U11 — 4011 CMOS quad 2-input NAND gate IC.

Table 2

Circuit-Board Material Required for the Pizza Clock Cabinet†

Quantity	*Size*
Base	
2	3-1/2 × 1-3/4 inches (sides)
2	4 × 1-7/8 inches (front and back)
1	4 × 3-1/2 inches (top)
Face	
1	8 × 5-3/4 × 3-1/2 × 5-3/4 × 8 inches (face)
2	8 × 1 inches (side panel)
2	5-3/4 × 1 inches (side panel)

† See note 1. (mm = in. × 25.4)

time numbers on the other.[1] The best way to build the clock is to break the assembly down into four stages. Each stage can be completed in an evening or two. First, build the base. Sand the edges of the pc board material using dry, 150-grit sandpaper and check for a tight, smooth fit between edges.

Tack solder the box together and again check for a proper fit. Once you're sure of the fit, solder along the inside edges of the pc material. Then file the outside edges of the box at a 45° angle to bevel the edges smooth. Now mount and wire the power-supply components and install the setting switches. These slide switches can be soldered to the box wall. Check the power supply for proper operation — output should be around 8 volts.

Next, construct the clock face. Follow the methods illustrated. A construction diagram for the clock face is shown in Fig. 9. Once the cabinet has been soldered together, buff both assemblies with steel wool until the copper shines brightly (steel wool leaves an oily finish that can be cleaned away with a damp cloth). After buffing the parts, spray a clear protective finish on the board to prevent tarnishing (Polyurethane varnish in a spray can works best).

The third step is to wire the IC portion of the circuit. If the circuit-board pattern is used in making the clock, the GND and +V pins of the wire-wrap sockets can be soldered to hold them in place. Try to wire-wrap the connections so there are no more than two wraps on each post. When the wiring is completed, place the RUN/STOP switch in the STOP position and apply power. A multimeter (set to the 10-V dc or greater range) attached to pin 1 of U3 should rise and fall at one-second intervals; this indicates proper operation of the timing chain. For precise timing, a frequency counter should be attached to pin 7 of U1. Adjust C1 for a reading of exactly 3.579540 MHz.

Finally, the LEDs are installed and

[1]Complete parts kits, including components and cabinet parts, are available from Circuit Board Specialists, P.O. Box 969, Pueblo, CO 81002. The pc-board patterns for the clock are available from ARRL Hq. for a business-size s.a.s.e.

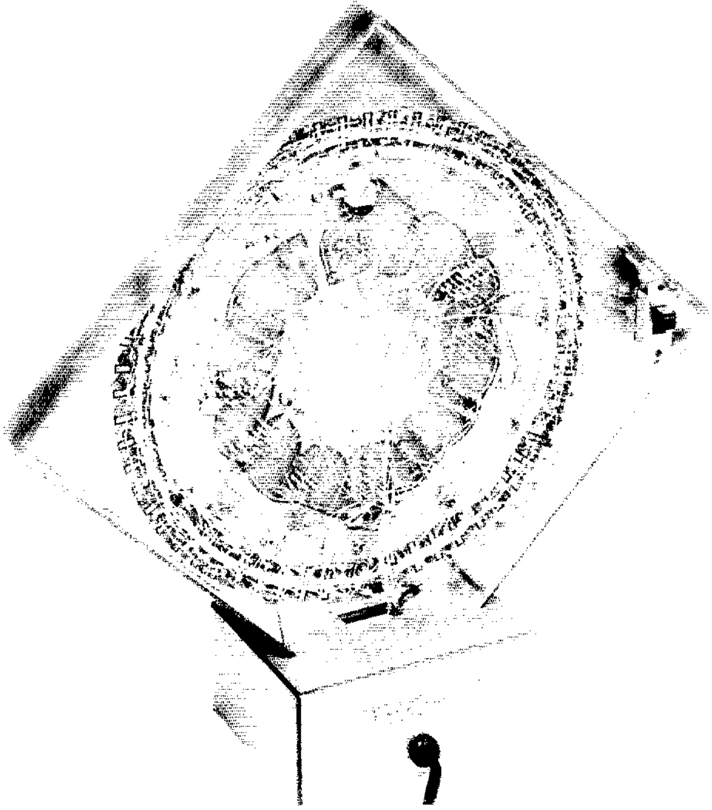


Fig. 8 — Rear view of the Pizza Clock face.

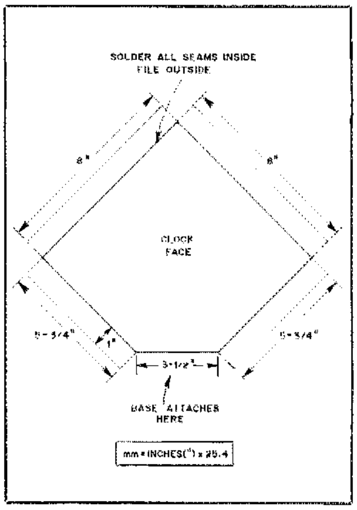


Fig. 9 — Method of assembly for the clock face.

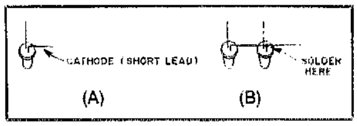


Fig. 10 — (A) The minutes and seconds LED cathodes are bent in this shape prior to mounting. (B) Each cathode of a 10-LED group is soldered as illustrated.

wired in. This is the most time-consuming portion of the assembly. Count out 120 LEDs from the 132 required and bend the cathode lead (usually the short one) into the shape of an L as shown in Fig. 10A. Mount the LEDs in groups of 10 in the minutes and seconds holes with the cathode leads outside (Fig. 10B). Drop a little epoxy on each LED to secure it and solder the cathode leads of each 10-LED group together. Now mount each hour LED (securing it with a drop of epoxy) and solder a 1N914 diode to each anode. Solder diodes to the anode of each minute and second group also. Wire-wrap wire is used to connect the LEDs to the counter ICs; it is best to solder to the LED terminals.

The large number of connections in the Pizza Clock allows lots of room for error. Check each connection against the schematic diagram *before* applying power. If all is okay, the clock should start running immediately. Each setting switch is self-explanatory. What's your next wire-wrapping project going to be? QST

Strays

ATTENTION HIGH SCHOOL SENIORS

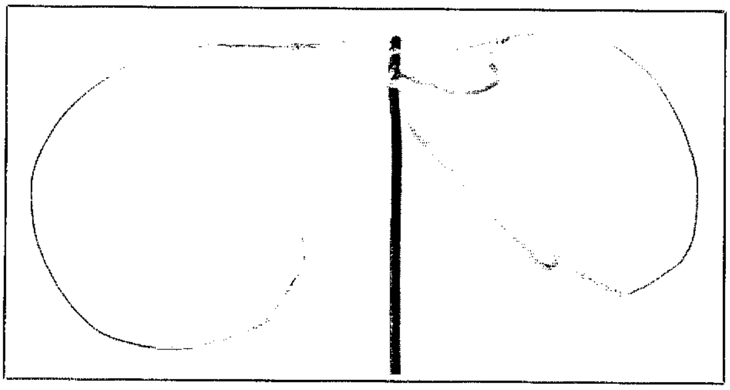
☐ The Atlanta Radio Club will award three $500 scholarships to graduating high school seniors entering an accredited college or university in the fall of 1983. Recipients must be duly licensed Amateur Radio operators at the time of application. Completed applications and high school transcripts must be postmarked no later than July 31, 1983. For more information and application forms, write to Phil Latta, W4GTS, Secretary, Atlanta RC Scholarship Committee, 259 Weatherstone Pkwy., Marietta, GA 30067.

IS YOUR EVENT ACCESSIBLE?

☐ I would like to urge all radio clubs who sponsor flea markets, auctions, club meetings, hamfests or any other event to include in their announcements whether they are wheelchair accessible. It is very discouraging to travel great distances in anticipation of fun and be greeted by stairs or other stumbling blocks to us hams who are real mobile stations on wheels. — *Steve Rich, WA1DFL, Revere, Massachusetts*

FRED BERRY, WØMNN

☐ Electronics inventor and designer Fred M. Berry, WØMNN, of Leawood, Kansas, became a Silent Key on January 6. He was 69. A very active operator in the early days of single sideband, Fred had a hand in the first post-WW II ssb rig in the Midwest. He also wrote articles on that subject for *QST* in the late 1940s and the early 1950s.



Wire Antennas for the Beginner

Every ham knows how to make and install wire antennas. But if you've never done it, you probably have a few questions. Here are some answers.

By George H. Woodward,* W1RN

Ah, spring — when girls' thoughts turn to boys and boys' thoughts turn to baseball. But if you're a ham of either sex the pleasant weather probably inspires thoughts of improving your antenna farm. Although many experienced amateurs insist that an antenna will work best if it's installed during a howling blizzard, such an approach is a bit intimidating for a newcomer. There's no need to fight the natural elements until you've mastered the electrical and mechanical elements.

Do it With Wire

Almost every ham likes to experiment with antennas. Your first antenna very likely won't be your last, but you have to start somewhere. Towers and beams are great, but save all that money and effort for later and start with a wire antenna. Why wire? "Skyhooks" made with wire are relatively inexpensive compared to aluminum-tubing types. And they're easy: easy to build, easy to install, easy to disguise and easy to modify.[1] Further, wire antennas are fun to experiment with. Your practical experience will bring your antenna textbooks and tutorial articles to life. Finally, making wire antennas is a rite of passage — you haven't really paid your dues until you've successfully deployed one or two.

Convinced? Good. But there are so many different antennas — which one should you start with? Here are several considerations that will influence your decision:

1) Safety
2) Available real estate
3) Available supports
4) Single-band or multiband operation
5) Possession of or need for a Transmatch
6) Type of operation (stateside or DX)

There may be some compromises inherent in these conditions. For example, if your goal is to work the long path between New England and Japan consistently on 80 meters, you won't make it with a 40-foot wire that's only 20 feet above ground.[2] Above all, follow this safety rule: *Never install an antenna where it can contact a public utility wire.* You may have a pair of 70-foot trees perfectly positioned to support a wire beam aimed at your favorite DX area. But if a power line runs between the trees, forget it.

If you have plenty of room for a real antenna "farm," then a separate antenna and feed line for each band is the way to go. Space is a real constraint for most of us, though, and we usually must ask a single antenna system to work on several bands. A modest multiband radiator can provide satisfactory results if your expectations aren't too high. A reasonable and realizable goal for a Novice station using one antenna on four bands is to work all states and Canadian provinces, along with some casual DX (Europe and South America for East Coast stations, and Japan and Australia for West Coast stations).

The most common multiband system in use today is a random-length wire coupled to the transmitter or transceiver through a Transmatch, as sketched in Fig. 1. The beauty of the method is its simplicity — just get the flat top portion of the radiator as high as possible, using whatever safe supports are available. Here's the major drawback: You'll need a Transmatch (also called an antenna tuner or antenna coupler) to transform the antenna feed-point impedance to a value the transmitter can deliver power into. A Transmatch is an extremely useful device, but it would be nice if you didn't have to use one, especially if you've just spent your last penny on a transceiver.

The other "rub" is the ground connection. If the radiator is near an odd multiple of a quarter wavelength, a poor ground connection will degrade the radiation efficiency. If the radiator is near a multiple of a *half* wavelength, a poor ground can introduce "rf-in-the-shack" problems. Symptoms of this malady include rf burns from equipment chassis, erratic keyer operation, automatic shutdown of solid-state transmitters and poor VFO notes. If you can locate your station

[1]Notes appear on page 38.

*143 Carroll St., New Britain, CT 06053

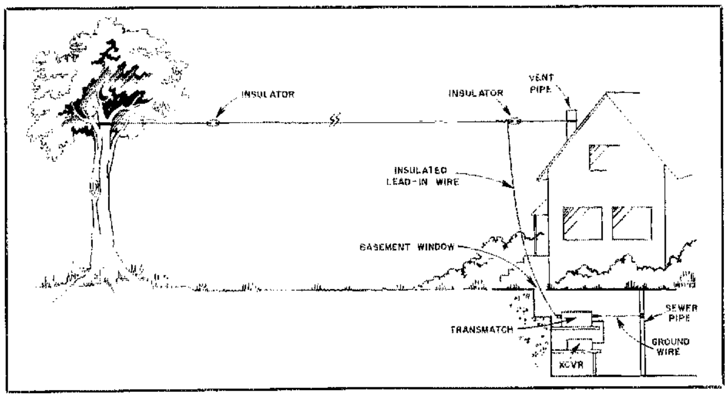


Fig. 1 — A typical end-fed random-wire antenna usable on all hf bands. The assets and liabilities of this system are discussed in the text.

in a basement utility room, the cold water inlet and/or sewer pipe will provide a good ground almost free. The shield braid from RG-8/U coaxial cable makes a good grounding strap. Sand, scrape or brush the pipe and attach the strap with an automotive hose clamp. Test your grounding pipe with a VOM — you should read full ac line potential between the "hot" side of a power outlet and the pipe. If you don't get this reading, your pipe probably has plastic couplings, which must be shunted with braid.

If you can't operate from the utility room, the next best spot is a ground-floor room with a window. A few 6- or 8-foot copper-clad stakes driven into the soil just below the window may provide an acceptable ground.

It's still possible to use an end-fed antenna from an upper floor, but it can get a little tricky. Water pipes aren't much good as rf grounds up there, so you'll have to use a *counterpoise* to create an artificial ground. A counterpoise is a quarter-wavelength wire (radial) that is connected to the Transmatch chassis at one end and open at the other end. One such wire is required for each band. They needn't be straight, but can run around the baseboards. The open ends will have high rf voltage, so make sure they're well insulated. A counterpoise supplies the "missing half" of the antenna system and will radiate a fair amount of energy. Your shack will be in the rf field when you use a counterpoise, but the equipment chassis should be cold. It should be possible to tame the rf in the shack by bonding all the chassis together and keeping all cables short. For safety you should still have a water-pipe connection or ground rod.

Whys and Hows of the Hardware

There's a lot of information in Fig. 1, but you might still have some nagging questions. For example, what size wire should you use? Electrically, it doesn't matter. Here's why: The current in an antenna varies along its length, being zero at the ends and maximum at the center of a half-wavelength section. (A current maximum is called a current *loop*, and a minimum is called a *node*.) The ratio of voltage to current at any point along an antenna is called the *radiation resistance*. This has nothing to do with the wire resistance, but accounts for the work the antenna performs as a *transducer* in converting the electrical energy produced by the transmitter into an electromagnetic field. The radiation resistance is the equivalent value of resistance that would dissipate the transmitter energy as heat instead of radiating it as an electromagnetic wave.

A half-wavelength antenna installed at a typical height over typical earth has a radiation resistance of about 50 ohms at the current loop. The transmitting equipment used by most Novices can produce about 100 watts of rf output. These two pieces of information and Ohm's Law will allow us to calculate the maximum current in the wire.

Recall that voltage equals current multiplied by resistance:

$$E = I \times R \qquad \text{(Eq. 1)}$$

and that power equals voltage multiplied by current:

$$P = E \times I \qquad \text{(Eq. 2)}$$

Substituting I × R for E in Eq. 2 we get:

$$P = I^2 \times R \qquad \text{(Eq. 3)}$$

Solving for I yields:

$$I = \sqrt{\frac{P}{R}} \qquad \text{(Eq. 4)}$$

When we plug in 100 watts for the power and 50 ohms for the resistance, the current comes out to be $\sqrt{2}$, or about 1.4 amperes. A bare no. 20 copper wire is good for 11 amperes dc in free air, so we can get away with a pretty small conductor for only 100 watts.

But the direct current rating can't be strictly applied to rf service because of the *skin effect*, which causes rf current to flow only on the surface of the conductor. This means that a no. 20 wire couldn't be counted on to withstand 6 kW in an antenna. On the other hand, the current has its maximum value in only one place in every half-wavelength section — the current gradually decreases toward the ends, so the low-current portions can act as heat sinks for the current loops. Perhaps that's more theory than you were expecting, but now you understand why you can use skinny antenna wire with your 100-watt transceiver even though the power-supply cable is fat.

With the electrical matter disposed of, the question of wire gauge becomes one of strength, ease of handling, cost, availability and visibility. The strongest wire that's suitable for antenna service is copper-clad steel, also known as Copperweld®. The copper coating is necessary for rf service because steel is a relatively poor conductor. Practically all of the rf current is confined to the copper, because of the skin effect. Copper-clad steel wire is outstanding for permanent installations, but can be difficult to work with. Kinking, which severely weakens the wire, is a constant threat when handling any solid conductor.

Enamel-coated "magnet wire" is a good choice for experimental antennas because it's easy to manage and the coating protects the wire from the weather. Although it stretches under tension, this will have no serious electrical consequences in a random-length antenna such as the one sketched in Fig. 1. If periodic adjustment of the tension is a nuisance after the antenna has been installed, the wire can be prestretched. A local electric motor rebuilder might be a good source for magnet wire.

It doesn't really matter whether the wire is insulated or bare (in the hf range, at least). Portions that could be touched by persons or animals, however, should be insulated to prevent shocks and burns. The lead-in in Fig. 1 is an example of a wire that should be insulated. This wire may be subject to some movement, so it's best to use stranded conductors. Hook-up wire, speaker wire and even ac zip cord are suitable for this service. Solid or stranded wire can be used for the main radiator, provided the ends are properly prepared. This subject will be taken up shortly.

The next question might be "how high should the antenna be?" Almost invariably the answer is "as high as possible." The height affects the radiation angle, azimuthal directivity, radiation resistance and ground losses. For Novice work, there's no need to be concerned much about directivity, and matching the radiation resistance isn't a problem. But ground losses can eat up a large chunk of your signal if the antenna is too low. The "high as possible" rule has its limitations, too — DXers often argue that 200 feet is too high because of unfavorable radiation angle and other considerations such as feed-line loss. These arguments are academic for most of us — if you can get most of the radiator up 25 feet or higher, that should be satisfactory for many good contacts on all bands. *The ARRL Antenna Book*[3] has a thorough treatment of the effects of antenna height.

What about length? So far, all we've said is "random," along with some references to fractional wavelengths. A *wavelength* is the distance between two points in the radiation field having equal intensity and phase. Wavelength is related to the speed of light and the signal frequency:

$$\lambda = \frac{c}{f} \qquad \text{(Eq. 5)}$$

where

λ = wavelength in feet,

c = velocity of light in space ($\approx 9.8 \times 10^8$ ft/s)

f = frequency in cycles/s (Hz)

A wavelength at 7.1 MHz would be

$$\frac{9.8 \times 10^8}{7.1 \times 10^6} = 138 \text{ feet.}$$

Using the metric conversion factor given in note 2, $138 \times 0.3048 \approx 42$ meters, which explains the name for the "40-meter" band.

To exhibit reasonable radiation efficiency an antenna should be at least a quarter wavelength long. Shorter antennas can be made to work well, too, but only if you pay careful attention to conductor and ground losses. So if you want the antenna of Fig. 1 to work on 80 meters, it should have a total length (flat top plus lead-in) of at least 60 feet. An end-fed antenna that is near a quarter-wavelength long is called a *Marconi,* and one that is near a half-wavelength long is called a *Hertz.* A 60-foot end-fed wire is a Marconi on 80 meters and a Hertz on 40 meters. If a 30-foot length is all you can manage, then 40 meters might be your lowest band, but you should give 80 a try anyway — you might be suprised.

You'll need insulators at the ends of the wire. High voltage exists at the free end of an antenna, so you can't just tie the wire to a tree. The tree is neither a good insulator nor a good conductor — it's a resistor. And a resistor is what's inside your dummy load. A perfect insulator would be transparent to your signal and a perfect conductor would reflect or reradiate it. But a tree will absorb and dissipate your signal and may even catch

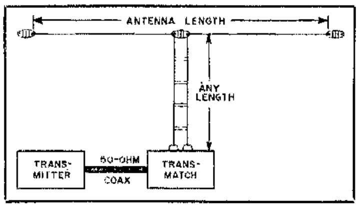


Fig. 2 — This classic antenna is back in vogue because of the widespread use of Transmatches with modern transceivers.

fire. For that reason, the radiator should terminate in an insulator outside the branches. Antenna insulators can be purchased from several *QST* advertisers, but it's easy to make your own. Quarter-inch-thick Plexiglas® is excellent insulating material for the hf bands and can be cut easily with a hacksaw. Drill holes near the ends for the wire loops to pass through, and chamfer these holes to reduce wear on both the wire and insulator. Both commercial and homemade insulators are shown in the title photo. Don't make the loops too tight — they should bear freely on the insulator holes. Twist the wire end back on itself and solder the splice thoroughly, filling in the crevices. Do this indoors if possible, because a cool breeze blowing on a long copper wire will suck the heat out of a soldering iron in no time. You may have to use a propane torch to solder outdoors. A liberal coating of clear acrylic lacquer will retard corrosion of the splice.

Putting It Up

Buildings often have built-in antenna supports, or at least the inspirations for supports. Examples of these are chimneys, vent pipes and balcony railings. A couple of sections of TV mast strapped to a chimney can really put an antenna up in the clear. Watch those power lines, though!

Use nylon cord for the halyard between the antenna end insulators and the supports. Nylon is the material to use because it can withstand high shock loads. A diameter of 1/8-in. is sufficient for all but the heaviest antennas. Depending on the weave, this cord is called either no. 4 mason line or parachute cord. The halyard can be passed through an eye hook screwed into the top of a mast for hoisting.

The support for the antenna end away from the house is likely to be a tree. Getting a halyard into the highest part of a tree can be a challenge. The most direct approach is to climb the tree, but this is often impractical. The alternative is to propel the line by some mechanical means. Many ingenious methods have been devised for this task, including using a bow and arrow, fishing rod or slingshot. Stay away from harpoon guns and nautical shot lines unless the tree is really high and the area is free of persons, animals or property that could be damaged.

If you're striving for only modest height, a small power transformer will carry a line over a branch and to the ground. Hold the line about 2 feet above the weight and sling it underhand for a healthy centifugal boost. Streamline your projectile as much as possible to keep it from snagging branches and to make it easier to retrieve — several tosses will probably be necessary to get the halyard where you want it. Use very small line on the projectile and then tie on the heavier halyard and hoist it with the light line. Tape your knots to prevent snagging.

What About Center Feed?

A random-length, end-fed wire can perform well, but there are situations that call for a different approach. Perhaps your house is *between* two good antenna supports. In such a case a center-fed antenna may be a good idea. A significant advantage of the center-fed configuration is that you don't need a ground to establish an rf current return. You still need one for electrical safety, but its length isn't important. A flat top of 65 feet or longer that is center fed with open-wire line is usable over the entire hf spectrum. Fig. 2 illustrates this versatile antenna. The lead-in doesn't radiate because the current in the two conductors is flowing in opposite directions at any instant. Because the conductors occupy almost the same space, the two fields cancel each other almost completely. The feeder should be dropped at right angles to the flat top to preserve the current balance.

A half-wavelength flat-top center fed with open-wire line is sometimes called a center-fed *Zepp,* after its end-fed cousin that was trailed from airships called zeppelins. Operated at the second harmonic, it becomes a *double Zepp.*

The Transmatch used to couple the open-wire feeder to a 50-ohm transmitter must be capable of working into a balanced load. Many commercial Transmatches have ferromagnetic baluns that are somewhat particular about the impedance they look into. A link-coupled Transmatch[4] is especially well suited for feeding open-wire line.

Resonant Antennas

Most modern transceivers are designed to work into a 50-ohm resistive load. If the load deviates too far from this value, a solid-state transmitter will reduce power or trip off completely. Most hams who have solid-state rigs have resigned themselves to the seeming necessity for a Transmatch. Is there a way out? Of course — use an antenna whose feed line presents a proper load to the transmitter.

The simplest antenna that can be made

to present the proper impedance to a solid-state transmitter without a Transmatch is a half-wavelength dipole fed in the center with coaxial cable. A miniature model is featured in the title photograph.

Coaxial cable is a convenient transmission line because the rf current flows (in opposite directions) on the outside of the center conductor and the inside of the outer conductor. The field is contained within the outer conductor, so the cable is *shielded*. Ideally, no current flows on the *outside* of the outer conductor. This property makes coaxial cable easy to handle because it can be installed anywhere (and the outer conductor can contact anything) without upsetting the internal rf currents. (Practical considerations sometimes deface this beautiful picture, but we'll deal with them later.)

How long should a dipole be? Earlier, we derived a crude formula for wavelength, but the effects of insulators and loops at the wire ends weren't considered. The formula most often used for cutting an hf dipole is

$$\ell = \frac{468}{f} \qquad \text{(Eq. 6)}$$

where
ℓ = length in feet
f = frequency in megahertz

This is the total length between the far ends of the insulator loops. The pigtail leads from the coax cable to the flat top are part of the radiator, too. Their length is significant at 28 MHz but practically unnoticeable at 3.5 MHz.

A popular variation of the horizontal dipole is the inverted V. This antenna is supported in the center and the ends droop to form an interior angle of 90 to 120 degrees. Don't make the angle any more acute than this because the fields from the two radiator halves will cancel just like a transmission line.

As you might expect, drooping the ends modifies the resonant frequency and feed-point impedance. Antenna impedance has been neatly explained by Hall in a recent *QST* article.[5] Bill Orr recently reported some length formulas developed for inverted Vs by Ohi, JA5COY.[6] The total length for a 120° inverted V is

$$\ell = \frac{465.6}{f} \qquad \text{(Eq. 7)}$$

For the 90-degree model the formula is

$$\ell = \frac{463.3}{f} \qquad \text{(Eq. 8)}$$

where ℓ and f have the same meanings as in Eq. 6. Lengths for all three dipole configurations are given in Table 1. These are *starting* lengths. Proximity to metal structures (such as other antennas) may alter the resonant frequency and impedance.

The inverted V is attractive not only because it requires just one support, but also because it needs less horizontal space than a standard dipole. Some high-school geometry shows that the 80-meter Novice dipole from Table 1 needs only 109 linear feet when configured as a 120° inverted V and can be shoehorned into 89 feet if the apex angle is 90 degrees. The same geometry shows that the minimum pole height is 32 feet for the 120° model and 45 feet for the right-angle case. We'd like to keep the ends at least 10 feet off the ground to prevent detuning or accidents (the ends can have high voltage, remember?). That calls for an apex height of 55 feet, which may be difficult to achieve with an existing or inexpensive support.

Table 1
Half-Wave Dipole Lengths for the Novice Bands

	Configuration		
Freq. (MHz)	*Horizontal (Z ≈ 73 Ω)*	*120° inv. V (Z ≈ 50 Ω)*	*90° inv. V (Z ≈ 30 Ω)*
3.725	125',7-1/2"	125'	124',4-1/2"
7.125	65',8-1/4"	65',4"	65',1/4"
21.15	22',1-1/2"	22'	21',10-3/4"
28.15	16',7-1/2"	16',6-1/2"	16',5-1/2"

But we can soften the height requirement by taking some liberties with the antenna ends. For example, if you must use a 90-degree inverted V and have only a 45-foot apex support, the antenna ends would just touch the ground. If we can bend a dipole in the vertical plane, however, we can bend it in the horizontal plane as well. Simply tie off the last 14 feet of each leg horizontally. The horizontal portions will be the desired 10 feet above ground. The horizontal sections can extend outward in the plane of the V or can be perpendicular to the plane. If you run them perpendicular to the V, placing them on opposite sides of the V plane is preferable to placing them on the same side. This is because the fields from the extensions will partially cancel each other if the conductors are parallel. The small loss of radiation isn't important but the resulting change in impedance and resonant frequency may be tricky to deal with. Similarly, don't run the extensions in toward the support if you don't have to — this can also complicate the adjustments.

Dipole Details

Techniques for putting up a dipole are similar to those for putting up an end-fed random-length wire. The dimensions need to be measured carefully, of course, to permit feeding the antenna without a Transmatch. Other than that, the only new ground to cover is the center insulator and the feed line.

The center insulator serves as an attachment point for the feed line. In addition, this component is the suspension point in an inverted V. Center insulators are available commercially, but here again it's cheaper to make your own. A Plexiglas® center insulator is shown in the title photo. It's made in much the same way as an end insulator except for some additional holes to anchor the feed line. A hole or notch at the top of the insulator locates the coaxial cable and two smaller holes on either side allow a locking cable tie to be passed through the insulator and around the cable. Notice how the cable is routed up the back of the insulator, over the top and down the front.

This arrangement serves the dual purposes of strain relief and moisture resistance. Having the pigtail leads pointing down forces water to overcome gravity to enter the cable. Water can still migrate up (and down) the braid by capillary action, so a dab of RTV sealant at the cable end is in order. Solder the pigtail leads to the wire loops and spray some clear lacquer on the connections.

A dipole center insulator needn't be as long as an end insulator. That's because the center is a current loop, where the voltage is minimum (voltage node). Voltage as a function of power and resistance is given by

$$E = \sqrt{P \times R} \qquad \text{(Eq. 9)}$$

where
E = rms potential in volts
P = power in watts
R = resistance in ohms.

Since a typical Novice transmitter puts out about 100 watts and we're shooting for a resistive impedance of 50 ohms, the potential will be $\sqrt{100 \times 50}$ or 70.7 volts, which doesn't require much of an insulator.

Now we come to the feed line. I recommend RG-58/U cable because it's light, easy to handle and relatively inexpensive. True, it *is* lossier than the larger RG-8/U, but unless you're using 100 feet or more at 28 MHz, the difference isn't significant. For long runs, you can compromise by using RG-8/U for the major section and RG-58/U for the portion that hangs from the center insulator. The radiator doesn't support the feed line on an inverted V, so you can run RG-8/U all the way up if desired.

The title photo shows the center conductor and dielectric neatly separated from the shield braid, but it doesn't show how to separate them. This process is explained in Fig. 3. You'll need a connector at the station end of your feed line. Most likely, your equipment will have an SO-239 (UHF female) receptacle, for which you'll need a plug from the PL-259 (UHF male) family. Fig. 4 illustrates the assembly procedure for common combinations of cables and connectors.

If you use the solder-on style of connector, soldering the braid can be tricky. In general, coaxial cable braid coverage isn't what it used to be, so don't be discouraged if you're not successful on the

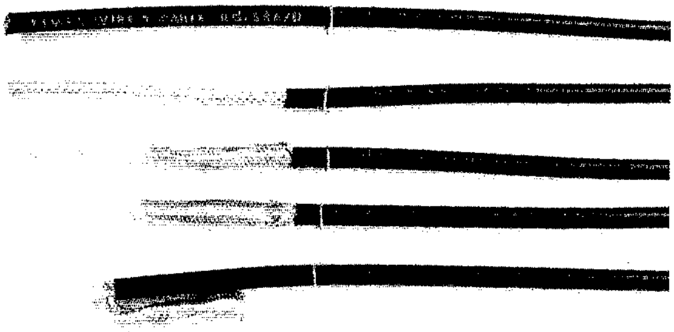


Fig. 3 — The conductors of a coaxial cable can be separated in four easy steps. A — Remove the outer sheath with a knife, being careful not to nick the braid. B — Push the braid accordian fashion against the sheath. C — Make a hole in the braid by separating the strands. D — Bend the cable at the hole and pull the center conductor and dielectric through.

first attempt. When the braid is sparse, pretinning it will ease the soldering job. Use a big iron or gun to solder the braid through the connector body holes — the pencil iron you use for pc boards just won't cut it. The idea is to heat the connector body *quickly* to soldering temperature and remove the iron as soon as possible. If you have foam-dielectric cable, the foam will bubble and ooze through the holes if you heat the body too long. When the connector assembly is complete, check for short circuits and continuity with a VOM.

Making it Work

Once the dipole is up, it's time to test it. The first thing to do is listen for signals. If the band isn't active, listen for natural or man-made noise — alternately connecting and disconnecting the antenna should have a marked effect on the background noise if the receiver sensitivity is adequate. Failure of this test means something has gone wrong — perhaps there's a short circuit or a connection has fallen apart.

The transmitting test depends on the equipment you have. If you have a transmitter with tube finals (and tune and load controls), just fire it up into the feed line. If the controls behave normally, that is, produce the proper plate dip and loaded current without running out of range, that's good enough — test no more.

You can try something similar with a solid-state rig, too. The test here is to see if the transmitter will draw the rated collector current with the same drive level used with your dummy load. These transmitting tests are best conducted when the band isn't too active, meaning daylight hours for 80 meters and late night for 10.

If the transmitter won't deliver power to the antenna, some different tests are called for. Ask a friend with a higher-class license to try to find the resonant frequency by trying several spots across the band. If the antenna resonates at a frequency that's 5% too low, shorten the wire by the same percentage and try again. If the antenna doesn't resonate inside the band, you can find the resonant frequency with a noise bridge and a general-coverage receiver.[7]

Suppose the antenna resonates in the band, but the system just won't take power. In that case, the SWR is probably too high for your transmitter. The radiation resistance of a horizontal dipole can be altered without significantly affecting the resonance by changing the height. Inverted V antennas are a little easier to tune. Lengthening the radiator will lower the resonant frequency, of course, but sharpening the apex angle will lower both the resonant frequency *and* the radiation resistance. The Novice subbands are only 50- and 100-kHz wide, so it should be possible to achieve an SWR of 1.5:1 or less across this segment. Any transmitter should be happy with that.

What if everything is cut right and it still won't load up? It sometimes happens, and to explain why we must expose the imperfection of coaxial feed. Recall that the transmission-line current in coaxial cable flows on the outside of the center conductor and on the *inside* of the outer condutor. The current is confined to the inner surfaces by induction and the skin effect. The skin effect completely isolates the inside of the shield from any currents that may be flowing on the outside. The outside of the shield becomes a *third conductor* connected to the feed point. Depending on the length and routing of the cable, this conductor can severely detune the dipole. It is this effect that sometimes causes the SWR reading to change radically with the length of the feed line, in apparent defiance of transmission-line theory.

The solution is to choke off or decouple the outside of the cable. One way is to attach an earth ground to the shield an odd number of quarter wavelengths from the feed point.[8] This ground will be transformed into a very high impedance at the feed point, keeping rf current off the shield. The limitation of this approach is that it only works on one band or bands that are odd multiples, 40 and 15 meters for example. If you need a broadband decoupler, some excellent designs are described by Cooper and Reisert.[9,10]

Antenna current can appear on your cable from *induction* as well as conduction. Induction can be minimized by orienting the feed line as symmetrically as possible with respect to the radiator, running the cable at right angles to the antenna. Chances are, you won't need any extra devices at the feed point. Direct coaxial feed usually works fine.

Multiband Dipoles

Separate dipoles for each band is an excellent antenna system if you have room. That's a pretty big "if"! A practical multibanding scheme that avoids a Transmatch is to excite several dipoles from a common feed point. Electrically, this arrangement places the antennas in parallel. The interaction isn't too severe, however, because a dipole exhibits a high impedance when operated at a frequency that is far removed from resonance. An 80-meter dipole, for example, shows a very high radiation resistance on 40 meters. Taking the opposite case, a 40-meter dipole fed with 80-meter energy has a low radiation resistance but very high reactance. In either case, the antenna current (and therefore the radiation) is very small. The current applied to the feed point of a multiband dipole takes the path of least impedance, automatically selecting the proper-length radiator.

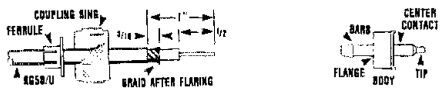
A Four-band Novice Dipole

Here's one you can build! It covers the Novice segments of 80, 40, 15 and 10 meters, and you should be able to use it without a Transmatch, even if it takes some cut and try. Fig. 5 shows the configuration in a single plane. The radiating elements need not lie in a single plane, however. In fact, my model had the 40/80-meter inverted V skewed about 45° with respect to the 10/15-meter dipole.

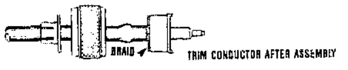
Open-wire line is used for the elements for ease of fabrication. It's a simple matter to measure and assemble a 22-foot piece of the line and then cut back one conductor on each side to the proper length for 10 meters. You can do the same thing for the 40/80-meter sections, but you can reduce the amount of open-wire line needed by splicing on a single wire for the 80-meter extension. Open-wire line is sold by several *QST* advertisers, but you can probably obtain it locally from a TV parts jobber. It comes with a characteristic impedance of 300 or 450 ohms. Use either one.

Some of the element lengths given in Fig. 5 deviate considerably from those listed in Table 1. This is because the

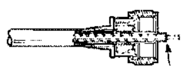
83-58FCP



1. Strip cable — *don't nick braid, dielectric or conductor.* Slide ferrule, then coupling ring on cable. Flare braid slightly by rotating conductor and dielectric in circular motion.



2. Slide body on dielectric, barb going under braid until flange is against outer jacket. Braid will fan out against body flange.



3. Slide nut over body. Grasp cable with hand and push ferrule over barb until braid is captured between ferrule and body flange. Squeeze crimp tip only of center contact with pliers; alternate-solder tip.

83-1SP PLUG (PL-259)

1. Strip cable, *don't nick braid, dielectric or conductor.* Tin exposed braid and conductor. Slide coupling ring on cable.

2. Screw body on cable. Solder braid through solder holes. Solder conductor to center contact.

3. Screw coupling ring on body.

83-1SP PLUG WITH ADAPTERS

1. Strip jacket. *Don't nick braid.* Slide coupling ring and adapter on cable. Note — use 83-168 (UG-175) adapter for RG-58/U and 83-185 (UG-176) for RG-59/U, RG-8M or RG-8X.

2. Fan braid slightly, fold back over adapter and trim to 3/8". Strip dielectric and tin exposed conductor. *Don't nick conductor.*

3. Screw body on adapter. Follow 2 and 3 under 83-1SP plug.

Fig. 4 — Cable stripping dimensions and assembly instructions for several popular coaxial cable connectors. This material courtesy of Amphenol Electronic Components, RF Division, Bunker Ramo Corp. (Dimensions on this drawing are in inches.)

Novice bands don't have an exact harmonic relationship and the nonresonant sections couple reactance into the working section. The only way to tune out this reactance is to adjust the element lengths. While the interaction necessitated a lot of adjustment, the shortening of the 80- and 40-meter sections is a blessing, in that less height and horizontal space are required. The 10-meter dipole (actually somewhat a "noninverted V") had to be made a little longer than formula, but the 15-meter antenna was about right. Start with the lengths in Fig. 5, but be prepared to make some adjustments for your installation.

The supports for this antenna need be only 22 feet apart. Perhaps your house has a chimney and vent having this much separation. Then the inverted V can straddle the peak of the roof. The test model had an apex angle of about 120°.

Here's a final word of caution about multiband antennas — they don't discriminate against harmonics. That means you must be careful to tune up your transmitter properly, especially if it's one of the older oscillator-multiplier types. If the plate current dips at more than one position of the tuning control, use the dip corresponding to maximum plate tank capacitance.

I installed this antenna on the roof of the headquarters building on Ground Hog's Day, with the help of Mike Kaczynski, W1OD. We put it up in a temporary fashion in balmy (for winter) weather. We didn't have the time-tested performance advantage gained from erecting it during a howling blizzard, but the thing (surprisingly) has survived three major snowstorms. As soon as it was up and trimmed, I answered the first CQ I heard on each of the four Novice bands. Starting at about 8:30 in the morning I worked Vermont on 80, Kentucky on 40, Luxembourg on 15 and Germany on 10, all with good signal reports. Not a world-beater by any means, but entirely respectable. Have fun with *your* antenna. Despite what others may tell you, start now, while the weather is nice!

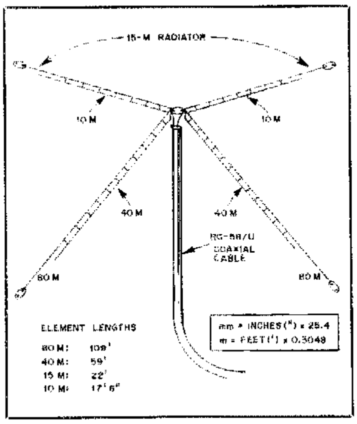


Fig. 5 — A multiband dipole that will work over all four Novice bands without a Transmatch. Construction details are given in the text.

Notes

[1] D. DeMaw, "Antennas for Those Who Can't Have Antennas," *QST*, Feb. 1983, pp. 15-17.
[2] meters = feet × 0.3048, mm = in. × 25.4.
[3] G. Hall, ed., *The ARRL Antenna Book*, 14th ed. (Newington, CT: The ARRL, Inc., 1982), pp. 2-16 to 2-23.
[4] *The ARRL Antenna Book*, pp. 4-5, 4-6.
[5] Hall, "A Simple Approach to Antenna Impedances," *QST*, March 1983, pp. 16-19.
[6] W. Orr, "Ham Radio Techniques," *Ham Radio*, Dec. 1982, p. 58.
[7] G. Woodward, ed., *The Radio Amateur's Handbook*, 60th ed. (Newington: ARRL, 1983), pp. 16-34 to 16-36.
[8] R. Dome, "Balanced Dipole Antenna Fed By Coaxial Cable," (Technical Correspondence), *QST*, May 1979, pp. 43-44.
[9] W. Cooper, "The Coaxial Cable Balun," *Hints and Kinks for the Radio Amateur*, 11th ed. (Newington: ARRL, 1982), pp. 5-10, 5-11.
[10] J. Reisert, "A Simple and Efficient Broadband Balun," *Ham Radio*, Sept. 1978, p. 12.

Conducted By Larry D. Wolfgang,* WA3VIL

SIMPLER RTTY DIVERSITY COMBINER

☐ I was intrigued by George Woodward's RTTY Diversity Combiner in the September 1982 Hints and Kinks column. After giving some thought to the circuit, I realized that it included more parts than necessary. There are two logic inputs, A and B. If they agree (both 1 or both 0) then the output is the same. If they disagree, however, the desired output must be the opposite of what it was the last time the inputs agreed. This requires another variable, C, representing the value of A and B the last time they areed. The Schmitt trigger hysteresis in W1RN's circuit provides this function.

I did some work with truth tables and Karnaugh maps to show that the logical function of Eq. 1 provides the desired result:

$$AB + (A + B)\overline{C} \qquad \text{(Eq. 1)}$$

C can be generated by a simple R-S flip-flop, which can be implemented using NAND gates.

Fig. 1 shows a RTTY Diversity Combiner that requires 8 NAND gates. My unit uses two SN7400N quad NAND gates, with no other parts. The total cost was about $1.20. — *Carl Hayes, KD5BH, Dallas, Texas*

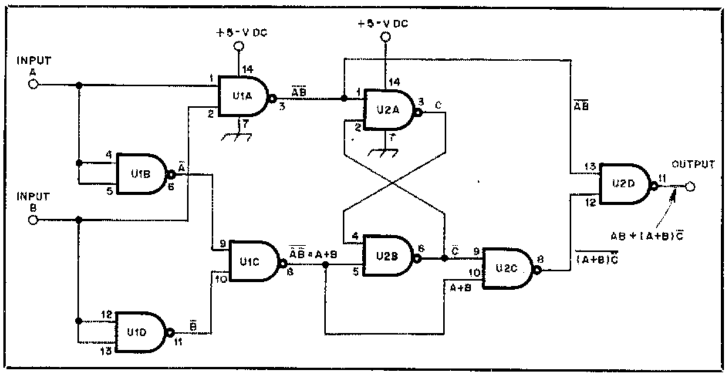


Fig. 1 — Schematic diagram of a simplified RTTY Diversity Combiner requiring only two ICs.

INEXPENSIVE 30-METER BEAM ANTENNA

☐ In about two months of operating on the new 30-meter band I have worked all 50 states and over 50 countries. My antenna is simple but effective. It is a rotatable inverted V beam. Fig. 2 shows the construction details. The antenna boom is suspended from a tree branch about 50 feet in the air.[1] The antenna can be rotated 360° simply by moving the two ground stakes. All of the materials to build this antenna cost me less than $25.

Eq. 2 gives the driven element length, Eq. 3 gives the director length, and Eq. 4 gives the element spacing that I used.

$$\text{D.E. length} = 476/f_{MHz} \qquad \text{(Eq. 2)}$$

$$\text{Dir. length} = 450/f_{MHz} \qquad \text{(Eq. 3)}$$

$$\text{Spacing} = 120/f_{MHz} \qquad \text{(Eq. 4)}$$

The feed-point impedance is around 30 ohms. I used a matching transformer made by connecting two 1/4-λ sections of RG-59/U coaxial cable in parallel. One end of the transformer connects to the antenna, and the other end goes to 50-ohm cable to the shack. Fig. 2B shows how this is wired. Perhaps the easiest method to join the two pieces of 75-ohm cable is to use coaxial T connectors. You should use a balun at the antenna feed point to prevent rf from flowing on the outside of the shield braid.

I also built an antenna of this type for 40 meters, and it works great. I guess the key word is rotatable! — *Jon Ferrara, N9DWR, Chattanooga, Tennessee*

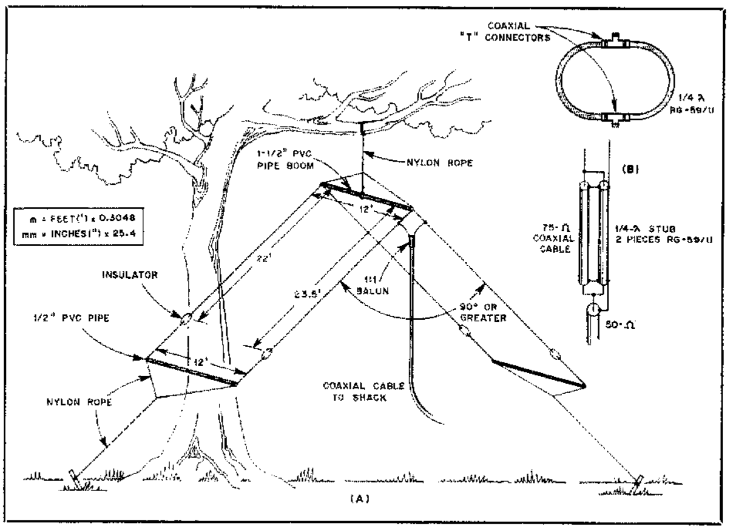


Fig. 2 — Construction details for a 30-meter inverted V beam are given at A. A coaxial-cable impedance-matching transformer is shown at B.

LOW-COST ANTENNA WIRE

☐ Radio amateurs frequently need low-cost copper wire for use in antenna systems. The wire is needed for radiators as well as for on-ground or buried radial systems. Having faced that need a number of times myself, and being a chap who objects to paying premium prices for such ordinary material, I sought some low-cost sources of suitable wire. You may find this information useful when you begin to accumulate material for that next antenna project.

Multiwire rotator cable for TV antennas is an excellent source of wire for many amateur antenna systems. The five-conductor cable from Radio Shack (15-1201) contains no. 20 wire. A 100-foot length sells for $10. By separating the five lines from one another you end up with 500 feet of insulated wire, which calculates out to 2 cents per foot. I have used this cable many times to put down a radial system. The conductors pull apart easily, and each retains a coating of plastic insulation.

Another type of wire that I have used extensively for the radiating elements of antennas is two-conductor speaker or "pot" cable. It has a heavy plastic insulating material that lends strength to the wire. This is especially useful for dipoles or similar antennas. As is the case with rotator cable, the individual conductors can be pulled apart to form separate insulated wires. Radio Shack light-duty cable (278-1387) costs $5.99 per 100 feet. The roll provides two lengths of no. 20 wire, which equates to 3 cents per foot. Not bad! Heavy-duty speaker wire (16 gauge) is also available from Radio Shack in 100-foot rolls at $11.95 (278-1384). I have used this vinyl-insulated, two-conductor wire for balanced feeders with dipole antennas and 3-element wire Yagis during DXpeditions. I never measured the loss, but it seemed to perform okay at power levels up to 100 W.

Those wishing to put up long spans of wire (Beverage antennas and such) may want to con-

[1]mm = in. × 25.4; m = ft × 0.3048.

*Assistant Technical Editor

sider using aluminum electric-fence wire. It is available from Sears, Roebuck and Co. in 1/4-mile-long rolls for a very low price. Another fine low-cost wire from Sears is no. 8 solid-aluminum wire with vinyl jacketing. It is listed as overhead power wiring for farms. The price per foot is very low. I have used this wire for buried radials. The insulation helps prevent corrosion of the aluminum conductor caused by soil acids or alkalies. Electric-fence wire, on the other hand, can dissolve in a few months when buried in some soils.

Finally, when seeking low-cost wire, don't overlook those deflection yokes from junked TV sets. They contain plenty of formvar-insulated magnet wire that is useful for winding coils and making lightweight antennas. — *Doug DeMaw, W1FB, ARRL Hq.*

TAGS FOR SHACK-WIRING IDENTIFICATION

☐ Recently, I was redoing my station wiring. I was making small tags to identify the interconnecting coaxial cables and affixing them with masking tape. My XYL Florence, WA1IKR, watched for a while, then left. She was soon back with a jar full of the plastic clips used to close bread wrappers. These clips have a small opening that is just right to fit over RG-58/U or RG-59/U coaxial cable. They can also be attached to power cords or to external speaker leads. Florence used a felt-tipped marking pen to write on the clips. The ink from most pens of this type will rub off, but we used a piece of clear cellophane tape over the tag to protect the label. [The ink from a Pilot permanent-marking pen will not rub off. — Ed.] Now it is easy to connect the individual cables that run from the receiver to antenna switch, between the amplifier and exciter, and the many other interconnecting wires that occupy the cable trough behind my operating table. This idea should prove especially useful for those who change equipment often, or for setting up a temporary station such as for Field Day operation. — *Stirling Olberg, W1SNN, Waltham, Massachusetts*

QRP TRANSMITTER REMOTE CONTROL

☐ My transmitter is a version of the W7ZOI "Universal QRP Transmitter."[2,3] I use this rig, along with a direct-conversion receiver for 40, 20 and 15 meters, at my home station and for portable operation. Trying to optimize the setup, I became concerned about the signal loss in the more than 50 feet of coaxial cable to my antenna.

I decided to mount my transmitter in the tree that supports the antenna and to key it from my operating position. A length of RG-59/U and a single + 12-V dc lead go to the transmitter. A 12-V spdt relay in the transmitter connects the RG-59/U to the antenna for receive, or to the key lead on transmit. A dpdt switch on the receiver cabinet completes the control circuitry. On receive, the coaxial cable is connected to the receiver; on transmit, it is connected to the key. The second pole of the switch provides + 12 V to the receiver or transmitter. Fig. 3 shows the wiring details. By mounting the transmitter close to the antenna more of the power is radiated, and that is important for successful QRP operation. — *Bert Halfpap, WØKOA, Denver, Colorado*

[2]D. DeMaw, "Experimenting for the Beginner," QST, Sept. 1981, pp. 11-15.
[3]Feedback, Nov. 1981 *QST*, p. 50.

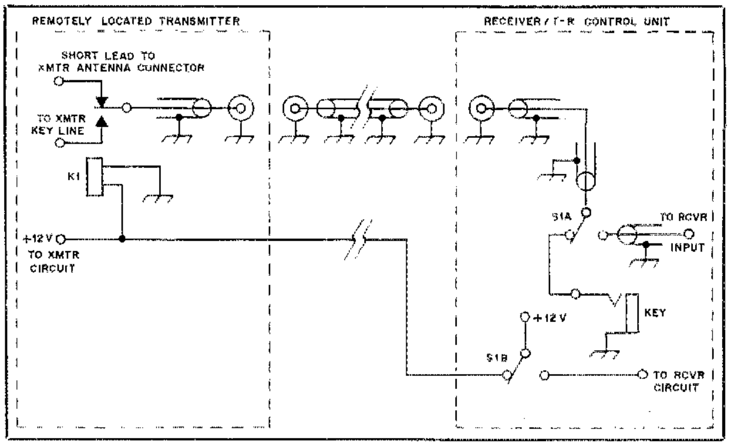


Fig. 3 — T-R control wiring used by WØKOA with his remotely mounted QRP transmitter.

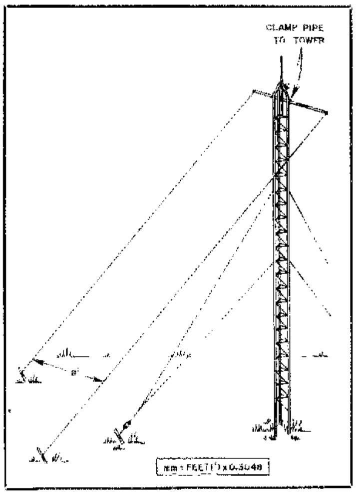


Fig. 4 — The method used by WA1OEZ to raise a beam antenna without the guy wires getting in the way.

RAISING BEAM ANTENNAS

☐ I decided to replace a Mini-Quad antenna with a 4-element triband Yagi on top of my 50-foot tower. A set of guy wires at the 35-foot level prevented pulling the new antenna straight up the side of the tower. My plans also included installing a 2-meter beam above the Yagi. I clamped a 6-foot pipe to the top of the tower and stretched ropes from each end of it to the ground (Fig. 4). I secured the ropes with stakes about 8 feet apart on the ground.

By removing the rotator, I was able to lower the mast until about 1 foot extended above the top of my tower and secured it at the thrust bearing. The new beam was laid on the ropes and pulled to the top of the tower. I placed it over the mast, but did not tighten it in place. I raised the mast, secured the 2-meter beam, then slid the mast up to the final position. I installed the rotator and positioned the triband antenna. One or two ground assistants are helpful to steady the slide ropes. — *Robert Mims, WA1OEZ, Taunton, Massachusetts*

TS-820S GOES TO MARS

☐ Those interested in MARS activity must find a way to tune their rigs to frequencies just outside the amateur bands. For those using a TS-820S, with or without the VFO-820, a simple adjustment will allow almost full-band coverage, plus coverage of several MARS frequencies.

Either the VFO-820 or the TS-820S VFO can be adjusted in the same manner. The lower frequency limit will be 10 to 15 kHz above the lower band edge, but this would only affect Extra Class licensees. The upper frequency limit will be approximately 40 kHz above the amateur-band edge.

The modification procedure should take no more than 15 minutes, and requires no instrumentation. Remove the top of the VFO or '820S cover. Looking at the VFO with the dial facing you, remove the silver tape covering the two access holes on the top-left side of the enclosure. Set the band switch to 3.5 MHz and turn the power on. Rotate the main tuning knob to the upper limit (reading 4.000 on the display). Adjust the analog dial on the front of the VFO to read "blank 40" using the dial-calibrate knob behind the main tuning knob. Turn the rear adjustment screw until the digital display reads 4.040. Now, turn the main tuning knob to the lowest-frequency stop, and adjust the front screw so the digital display agrees with the analog dial (about 3.500 to 3.550). These two adjustments are interactive, so you will have to repeat the process several times. With some persistence, the digital readout will be synchronized with the analog dial from 3.510 to 4.040 MHz, or whatever limits are set. Do *not* adjust the exposed adjustment screw on the right side of the VFO top.

This adjustment will affect the other bands in a similar manner. If you ever want to reverse the modification, the band limits are returned easily with a similar procedure. — *E. Benson Scott II, M.D., AE5V, West Monroe, Louisiana*

Conducted By Dennis J. Lusis,* W1LJ

COLLINS 32S AND KWM TRANSMITTERS IN THE WARC BANDS

☐ When this classic series of transmitters is first placed on the 30-meter band, it is often found that the transmitter has practically no grid drive to the final amplifier. It is tempting to simply trim the inductor nearest the rear of the tuning gang — the most critical one of the group. Unfortunately, this can create a parasitic oscillation near 17 MHz, detectable only with a second receiver.

A better solution is to obtain whatever grid current is available at a dial setting of 3 on the preselector logging scale (Fig. 1). Then peak the grid current by rotating each inductor *no more than one-half turn.* Next, peak the capacitors on 20 meters, as prescribed in the Collins manual, and then return to 30 meters for a second try with the inductors. By making no more than one-half turn at a time, you will avoid trouble. Be sure the receiver preselector is peaked correspondingly if you are in transceive operation.

If a general-coverage receiver is available, tune in the parasitic oscillation near 17 MHz and vary the driver plate inductance and the mixer capacitance, noting that the "parasite" occurs only when both these adjustments are not peaked at the same position of the tuning gang.

The 10-MHz power output may be less than that obtainable on 20, 15 and 10 meters, as the tank circuit Q is somewhat high for good efficiency (high circulating current). Reasonable output usually can be obtained with 225 mA of plate current. Do not tune for more current than this, or discoloration of the final tank coil may occur, indicating excessive values of circulating current. — *Cliff Buttschardt, W6HDO, Los Osos, California*

*Assistant Technical Editor

ABOUT "RUBBERING" CRYSTALS

☐ Inductance-loaded VXO design is facilitated by the use of the graph in Fig. 2. The operating frequency varies from series resonance, ω_s, downward. C_0 is the equivalent parallel capacitance of the crystal; r is the ratio of C_0 to C_1 (equivalent series-branch capacitance of the crystal) and is the electrical equivalent of the mechanical elastance of the vibrating plate. L is the equivalent load inductance; it is made continuously variable by various configurations of adjustable capacitors.[1] HC-6/U fundamental crystals have a C_0 in the range of 5 to 9 pF and an r value in the vicinity of 250. Almost all fundamental high-frequency crystals have r values between the limiting values of 125 and 500, as shown by the dashed lines in Fig. 2. Stability varies roughly with the slope of the curve. For r = 250, the best compromise of range versus stability occurs at about 1.5 kHz/MHz. Stability at unregulated room temperature is about one part per million per hour. — *Frank Noble, W3MT, Bethesda, Maryland*

[1]F. Noble, "Phantom-Coil VXO," *Ham Radio*, Jan. 1982, p. 66.

NOTES ON COAXIAL BALUNS

☐ [Editor's Note: The following is taken from a letter by John Belrose, VE2CV, to D. A. Christie, WB5KFP, who asked Belrose if the antennas he modeled in his Dec. 1982 *QST* article, "The Effect of Supporting Structures on Simple Wire Antennas," were fed using a balun or not, and what the effects of this were.]

The model antennas we built were all connected to the feed line through a coaxial balun of either a 4:1 or a 1:1 type. For the inverted-V dipoles, a 1:1 coaxial balun was constructed by sliding a decoupling sleeve ($\lambda/4 \times 0.95$ long) over the RG-174 coaxial feed line. This arrangement proved to be unacceptable at 200 MHz.

We noticed a problem while making impedance measurements. When the rf bridge was "balanced," bringing one's hand near the feed line unbalanced it. The decoupling sleeve, which was formed from the braid of a coaxial cable, was then made "solid" by soldering; after this the bridge no longer became unbalanced when a hand was brought near. We did, however, notice asymmetry in the vertical-plane pattern of the inverted V, for vertical polarization measured in the plane of the antenna. This effect was more noticeable when a conducting support tower was employed (Figs. 3 and 4). The effect was attributed to an imperfect transformation in the balun.

We subsequently learned how to make better coaxial baluns for use at vhf and uhf, but we did not repeat our measurements while using them. The outer sleeve should be made of solid material and the coaxial cable should have a solid outer sheath, or be constructed from tubing of a similar size, with an insulated wire (of proper diameter for the desired impedance) running down the center. There should be sufficient air space between the coaxial cable and the inner wall of the sleeve so that dielectric material can be inserted to electrically "lengthen" the inner sleeve.

I mention *inner* sleeve because it has different electrical lengths on the inside and outside. The outer surface is excited by the radiated field, so from a radiation point of view the physical length is shorter than the electrical length by a factor that depends on the electrical diameter of the

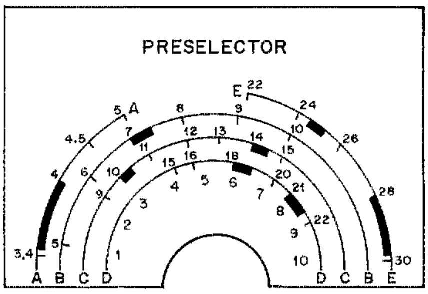


Fig. 1 — New preselector template (including WARC bands) for Collins S-Line and KWM transmitters.

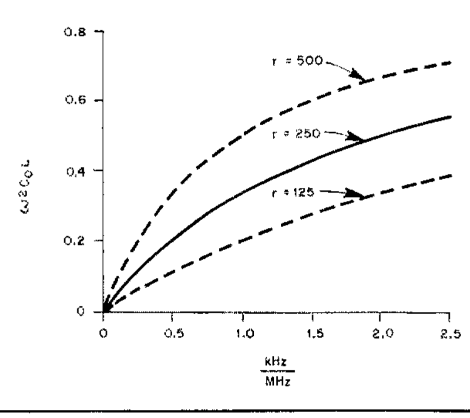


Fig. 2 — Graph used to aid in the design of VXOs.

sleeve. At vhf, this factor is 0.7-0.85. On the inside, the sleeve behaves like a transmission line. So for air dielectric the physical length is shorter than the electrical length by a factor of about 0.975. Hence, if the sleeve is "cut" to the right outside length, it is too short to be an electrical 1/4 λ on the inside. By dielectric loading (dielectric material should be low loss and have a constant of about 1.8), the electrical length on the inside can be made equal to that of the outside.

Let us now look at the effect of supporting towers on the patterns for an inverted-V dipole. In my *QST* article I showed only "symmetrical" smoothed patterns. In Figs. 3 and 4 are the unsmoothed, measured vertical-plane patterns for vertical polarization, measured in the plane of the antenna for △ = 90° and 127°. You will notice that there is indeed some asymmetry, which is more marked for a conducting tower and at low elevation angles.

This effect is attributed to a slight unbalance in the feed to the dipole, and therefore I decided (for my *QST* article) to plot a smoothed symmetrical pattern. The field strength plotted was the geometric mean value at each elevation angle. Since care was taken to construct the balun, and yet the pattern unbalance at low elevation angles (say 10°) can be as much as 5 dB, I conclude that without a balun the bidirectional nature of the antenna would be completely spoiled . . . but I do not have any measurements to verify this.

Incidentally, I have been reexamining our notes on these antenna measurements, and I have noticed that the inverted-V having a △ = 127° was almost a perfect match for the 50-Ω feeder. The impedance of the half-wave dipole was 67 Ω (should have been 72 Ω), and the impedance of the inverted-V for a △ = 90° was 36 Ω. My statement in the *QST* article that a configuration of △ = 90° provides a better match to 50-Ω feed line than a horizontal dipole does was therefore not correct. However, if you wish to work DX, the configuration of △ = 90° provides better low-angle radiation in the plane containing the antenna.

I should emphasize, so that there will be no misunderstanding, that the extreme sensitivity to any unbalance is only for the vertically polarized field. The balun was "plenty good enough" for the horizontally polarized field in the orthogonal directions . . . the pattern was symmetrical. — *John S. Belrose, VE2CV, Aylmer, Quebec*

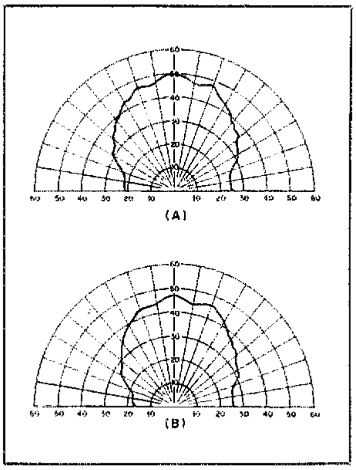


Fig. 3 — Vertical-plane polar diagrams of 1/2-λ inverted Vs, for vertical polarization measured in the plane of the antennas. The angle between dipole arms (△) is 90° in both cases. A nonconducting support tower is used in A, and a conducting support in B. Note increased asymmetry in B. [The patterns are drawn on a *linear* scale and are not in *QST* style. They are included for relative comparison only. — Ed.]

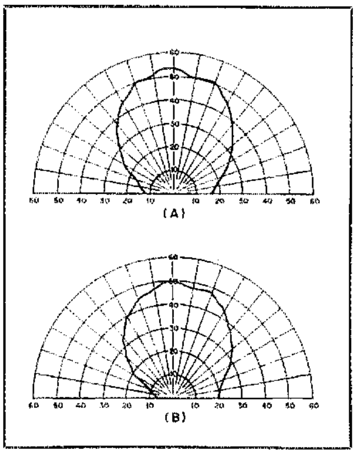


Fig. 4 — Same as in Fig. 3, except △ = 127°.

LEGAL WIRE SIZE

☐ Over the years, I have read numerous articles concerning so-called "invisible" antennas made of small-diameter wire. The appearance of a recent article in *QST* prompts me to observe that many hams are apparently unaware of the National Electrical Code Table 810-52 — which specifies that the minimum size of antenna conductors be *no. 14.*

Amateurs should consider the possible implications of presenting a claim to an insurance carrier as the result of a liability caused by an antenna erected in violation of this code. — *William H. Tilton, K7OKC, Centralia, Washington*

POWER-TUBE FILAMENT CONSIDERATIONS

☐ When the filament opened up on my 8877 tube, I found that I couldn't pick up a $1.98 replacement at the corner drug store. I figured it was time to do a little investigating to see what could be done to help prolong the life of this tube. The Eimac specifications for the 8877 note the following:

1) The filament voltage should be 5.0 ± 0.25 V.

2) The cathode warm-up time should be a minimum of *180 seconds.*

The warm-up period in my Alpha 77 was *only 60 seconds,* and the instruction book stated that the warm-up period should be between 80 and 120 seconds (still too short!). I changed the R-C network in the timing circuit to over 180 seconds — assuring I didn't apply drive too soon.

Page 192 of the 1977 ARRL *Handbook* specifies 6-V ac for the 8877 filament. [Later editions have been corrected. — Ed.] I also did not notice any mention of tube warm-up time.

The bottom line is a warning that anyone designing or using an amplifier using the 8877 tube should be aware of the proper filament voltage. Also, the required warm-up period is at least 180 seconds before drive power is applied. — *Howard A. Miller, W4KXE, Luray, Virginia*

Feedback

☐ Please note these corrections to "Amateur Use of Solar Electric Power — Part 2," November 1982 *QST*. Fig. 3 should have a connection dot joining the negative output line of module 17 to the BANK 2 negative bus. In Fig. 10, the 20- and 50-A breaker labels should be swapped.

☐ The book review "Three New Directories For Amateurs," in March *QST*, page 35, should have made clear that the *Amateur Radio Call Directory* is the type with which we are all familiar. That is, call signs, names and addresses are listed by districts. The other directories are a *Geographical Index* and a *Name Index*, respectively.

☐ Two items in the April 1983 Hints and Kinks column have apparently confused a number of readers. We know of no way to produce a 24-hour-format display on the small stick-on clocks. Dave Geiser, WA2ANU, suggested setting one to 12-hour-format UTC. If anyone does know of a simple modification to provide a 24-hour display on these clocks, please send details to the Hints and Kinks Editor.

☐ Sharp-eyed readers have been quick to point out references that are appropriate to append to the list on page 27 of April 1983 *QST* for the article "The Search for a Simple, Broadband 80-Meter Dipole" by J. Hall, K1TD. They are:

Lawson, J. L. "160/80/75-Meter Broad-Band Inverted-V Antenna." *QST*, Nov. 1970, p. 17 (considers parallel-fed dipoles of different lengths).

Vissers, W. "Build a Double Bazooka." *73*, Aug. 1977, p. 36 (indicates the broad-band characteristics of the parallel-connected coaxial dipole, a more apt name for what Hall calls a "double bazooka with crossed connections at the feed point").

Harbach, A. B. "Broad-Band 80-Meter Antenna." *QST*, Dec. 1980, p. 36. (treats number of conductors versus equivalent diameter in a cage antenna, and mechanical and other considerations).

Strays

SSCs — WE HEAR YOU!

☐ When applying for the Special Service Club program through your section's Affiliated Club Coordinator (see December 1982 *QST* or contact Hq. for further information), you may now list your club's ARRL members by *call sign* instead of by membership number. We hope this will ease the process of applying so that your club can more quickly and easily get down to the business of working more effectively within your community for the betterment of Amateur Radio.

Product Review

Conducted By Paul K. Pagel,* N1FB

Yaesu FT-230R 2-Meter FM Transceiver

The Yaesu FT-230R is a microprocessor (μP)-controlled, synthesized fm transceiver that packs 25 W of output power (at least) and is designed for use with a 13.8-V dc negative-ground supply. Front-panel layout is superb; the most often used controls are in easy reach. Because of the small size of the radio, the controls are close together yet far enough apart to avoid accidental operation of another control. Five '230 models are produced, one for the American market and the others for the rest of the world.

Features

The '230 is equipped with a microphone, a fused power cord, a mounting bracket, a removable wire stand and an SO-239 connector on the rear panel. A large heat sink takes up most of the back panel. The rest of the space is occupied by the power connector, a 1/8-inch jack for an external speaker, a tone burst ON/OFF control and a control for the frequency scanner.

A frequency coverage of 143.500 to 148.495 MHz surpasses the rated specifications. With the five-digit display, a resolution of 0.1 kHz is achieved. (When the display shows 5.545.0, the actual frequency is *14*5.545.0 MHz; the first two digits are assumed.) The front-panel-mounted VOLUME control doubles as the ON/OFF switch, and an outer ring on the same knob sets the SQUELCH level. A rotary switch selects the repeater offsets of ±600 kHz, and simplex operation. TONE CALL and TONE SQUELCH controls are located on the front panel. (In the European models, a HIGH/LOW [25/3W] power switch replaces the TONE SQUELCH control.) Red and green LEDs indicate a transmit condition or an opened squelch, respectively. A horizontal meter indicates power output on a scale of 0 to 10 and received signal strength. The MIC jack accepts microphone audio input and control lines for the PTT switch and frequency scanner. The μP and frequency controls occupy the rest of the front panel.

Yaesu FT-230R VHF FM Transceiver Serial No. 2F 050470

Manufacturer's Claimed Specifications	*Measured in ARRL Lab*
Frequency coverage: 144.00 to 147.995 MHz in 5- and 10-kHz steps.	143.500 to 148.495 MHz in 5- and 10-kHz steps.
Mode of operation: Fm.	
Readout: Five-digit LCD array.	
S-meter: Horizontal bar meter.	
Receiver sensitivity: 1 μV for 30 dB S/N.	0.27 μV for 20 dB.
Audio output power (8-Ω load): 1 W.	1.32 W.
Transmitter rf power output: 25 W.	30 W.
Spurious suppression: Better than 60 dB.	See Fig. 1.
Current drain: Receive, 0.3 A; transmit, 5.0 A.	Receive, 0.21 A; transmit, 5.85 A.
Size (HWD): 2 × 5.9 × 6.8 in. (51 × 150 × 173 mm).	
Weight: 2.9 lb (1.3 kg).	

The Microprocessor

In addition to the small size, the most attractive aspect of this transceiver is the microprocessor. While operating frequencies *can* be selected using the 1-inch-diameter click-stop tuning dial, why not take advantage of all the conveniences a four-bit μP can provide? With switches on the microphone and on the front and back panels, the operator can make full use of the frequency-selection flexibility of this rig. Ten memory channels are available; each will recall a repeater offset of ±600 kHz or will store transmit and receive frequencies for simplex operation. To load the memories, the desired frequency is selected using the main dial, one button is pressed and the necessary information is entered, erasing whatever frequency was stored previously in that channel. A scanner, controlled by UP and DOWN buttons on the microphone, will zip through the 4-MHz frequency range in 45 seconds when using the 10-kHz steps provided by the PLL frequency synthesizer. A 5-kHz step rate may be used, doubling the scan time. The scanner can also go through the memory channels; a front-panel button selects the scanner range. The scan will stop for five seconds for any of three commands: a clear frequency, a busy frequency — when the squelch is opened — or manually, by pushing the UP, DOWN or PTT buttons on the microphone. The scanner is activated by holding the UP or DOWN buttons for more than 1/2 second. If pressed for less than 1/2 second, these same buttons will advance the synthesizer one step — 5 or 10 kHz. The microphone has a LOCK switch that disables the frequency-selector buttons.

Other Specifics

Two VFOs are selected by an A/B button on the front panel. The VFOs may be used with the memory channels to cover unusual repeater pairs. In this mode, the receive frequency is fixed and controlled by the memory channels, and the main VFO dial selects the transmit frequency. When this function is activated (by pressing two buttons on the front panel), a small bar appears on the LCD. A priority channel is available, the desired frequency being checked every five seconds during operation. The scanner will stop on the priority frequency if it is busy or clear, depending on how the scanner is set. A lithium battery is provided to retain the last function and frequency selected prior to disconnecting the transceiver from a power source. Whenever a major function is varied, or if the scan locks for five seconds, a beep sounds, alerting the operator to the change in status.

The manual is easy to read; approximately half of its 52 pages are devoted to control operations, installation and operating instructions. A thorough maintenance and alignment section is included.

Impressions

The FT-230R feels right at home in a car. I took the rig on several trips around the northeastern U.S. With a 1/4-wavelength whip on the roof of my car, I was easily able to work repeaters 50 miles away. The high-power output

*Assistant Technical Editor

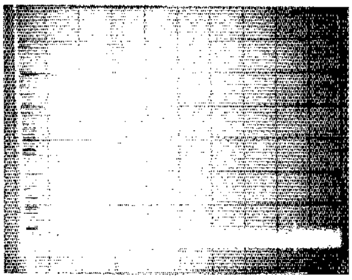


Fig. 1 — Worst-case spectral display of the FT-230R. Vertical divisions are each 10 dB; horizontal divisions are each 50 MHz. Output power is approximately 30 W at a frequency of 148 MHz. All spurious signals are at least 62 dB below peak fundamental output. The fundamental has been reduced in amplitude approximately 32 dB by means of notch cavities; this prevents analyzer overload. The FT-230R complies with current FCC specifications for spectral purity.

is a big plus. Transmit audio quality was never criticized. All front-panel controls have a decisive feel to them, and the status-change beep was a help while I was concentrating on my driving. Before a trip, I cut out portions of an *ARRL Repeater Director* and taped them in my atlas next to the various cities I would be passing through. It was a simple matter, when coming close to a metropolitan area, to load up the 10 memories with the appropriate frequencies and offsets. The µP memory feature was a blessing here. All further frequency control was performed through manipulations of the microphone buttons, making my driving quite a bit safer.

I noticed several minor drawbacks when using the rig. The sound from the built-in, bottom-mounted speaker tends to be mushy. While this is no trouble when the rig is used as a base station, careful listening is required when driving at highway speeds. The large LCD is easy to see, although in bright daylight the digits are unreadable when viewed at an angle. I never found the priority channel feature to be of use. If I had a favorite net or repeater frequency, though, it would be indispensable. The rig draws a hefty amount of current commensurate with its high power. It is unfortunate that the American version has no low-power setting. With such high power, the heat sink gets *hot*. Air should be permitted to flow past the rear of the rig. I found that out-of-band transmission can occur, but only when the simplex mode is selected.

The FT-230R is an excellent 2-meter rig. If you're looking for small size and high power, this rig may be for you. The Yaesu FT-230R is available from Yaesu Electronics Corp., P.O. Box 49, Paramount, CA 90723. Price class is $300. — *Leo D. Kluger, WB2TRN*

FLESHER CORPORATION TU-300 AND -470 RTTY TUs

□ One of the more popular RTTY terminal units (TU) to have entered the Amateur Radio ranks is the Flesher TU-170.[1,2] RTTY buffs found this unit to be an economical and highly effective modem, and many are still in use today. But there were many '170 users who craved a little more in the way of operating features. The TU-300 and -470 should fill this need.

My TU-300 review was on its way to be typeset when I received news of the birth of the TU-470. After I spoke with Joe Elliott of the Flesher Corporation, we agreed it would be a good idea to combine the reviews of the two units because of their similarity.

To a great extent, the circuitry of the two units is the same. Some circuit changes, a different board layout, a built-in loop supply, and a front-panel face lift distinguish the '470 from the '300. It comes with all the plug-in boards that are options with the TU-300. Perhaps the most distinguishing characteristic of the '470 is that it is not available in kit form. Only wired and tested units are available from Flesher.

Description

The TU-300 and -470 are each housed in a smooth-lined, heavy-duty, blue-gray steel cabinet. The units are powered from the ac line by a built-in, triple-voltage-regulated power supply. A series of five function-indicator LEDs and a 10-segment LED bar graph display occupy the upper-right portion of the front panel. The lower front-panel section supports a neat row of flag-type push-button switches.

The bar-graph display is used as a signal strength indicator and provides a relative filter-output level indication. Most of the five LED indicator functions are self-explanatory: POWER, ac power on; SEND shows the TU-300 is in the transmit mode; RDA (receive data available) indicates the presence of a signal and closure of the autostart relay; MARK/SPACE, illuminated when a signal is present at the respective filter output when in the receive mode and during transmit, indicates the presence of the appropriate tone at the afsk input.

Up Front

TU-300 front-panel push-button switch functions include: POWER ON/OFF, OPERATE/STANDBY, RECeive/SEND, REVERSE SHIFT SEND/REC, and AUDIO FREQUENCY-SHIFT 170 HZ, 425 HZ, 850 HZ. The '470 has added RTTY/CW and FILTER NARROW switches. In the STANDBY position, the demodulator output is locked in the MARK state and the autostart relay is energized. It takes *approximately* one second for the relay to drop out after restoring the switch to the OPERATE position.

With the REC/SEND switch in the REC position, the filters selected by the FREQUENCY SHIFT switch are enabled and the demodulator output placed on the RS-232C and TTL output lines. In the SEND position, the demodulator output is locked in the MARK condition, the afsk audio output is enabled and a pair of auxiliary switch contacts is closed. (Access to these contacts is made via a 25-pin connector on the rear panel.)

Locking the REVERSE SHIFT REC switch reverses the MARK and SPACE assignments of the selected audio frequencies. The SEND switch section reverses the output frequency assignments to the MARK and SPACE afsk inputs. The two switches are used when receiving or transmitting "upside down." (Convention calls for the use of lsb when using afsk on the hf bands.)

Mutually exclusive switches are used to select the FREQUENCY SHIFT used. In each case, the switch selects a separate filter board for the SPACE frequency, MARK remaining at 2125 Hz. For 170-Hz shift, a 2295-Hz filter is chosen, a 2550-Hz filter for 425-Hz shift, and a 2975-Hz filter for 850-Hz shift. For the '300, remember that the individual boards should be installed for the shifts desired, but MARK-only copy is possible for any of the shifts for which no SPACE frequency filter is present. All boards are installed in the '470.

The RTTY/CW switch of the '470 selects either mode of operation. The cw demodulator has a center frequency of 750 Hz. That corresponds closely to what most transceivers use nowadays for the audio pitch delivered when a cw signal is centered correctly in the filter passband. As the cw demodulator board's center frequency is adjustable, some deviation from the nominal 750-Hz figure is possible.

A FILTER NARROW switch on the '470 inserts a 170-Hz shift preselector filter that has a bandwidth of approximately 350 Hz. This filter is

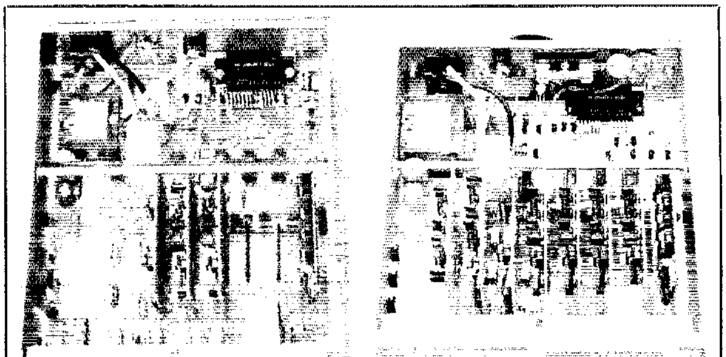


Fig. 2 — Views of the attractive interiors of the TU-300 and TU-470. The TU power transformer is at the left rear of the cabinet; the autostart relay is to the right of the transformer. All ICs are socketed.

[1] "Flesher Corporation TU-170 RTTY Terminal," *Product Review*, March 1979 *QST*, p. 42.

[2] The successor to the TU-170 is the TU-170A, announced in December 1982.

disabled when any frequency shift other than 170 Hz is used and should be manually disabled when using the cw demodulator feature. The manufacturer does not recommend the filter be used for 300-baud operation.

Bringing Up the Rear

The TU-300 and -470 rear panels are similar. On the '300 rear panel there is a DB-25 socket for I/O connections (a mating plug is supplied), a four-pin plug for use with the optional external loop supply, a grounding post, entry point for the ac line cord, and a three-conductor auxiliary ac socket that will provide autostart-relay-controlled 117-V ac line power to an external device such as a printer. The four-pin plug is absent from the '470 rear panel. A 1/4-inch key jack is located in the upper-left corner instead. It is used as the 20- or 60-mA loop output connector. Loop current is determined by the use of a jumper.

Other Differences

Some control pin function differences exist between the two TUs. Pin 23 of P1 on the '300 (now, more appropriately, S1 on the '470) is unused. In the '470, a SEND-P line has been added here. This is essentially a dupe of SEND-N, but where SEND-N is a TTL level, SEND-P is a bipolar ±3-V "RS-232C" level. XMIT-P and XMIT-N provide transistor keying for positive or negative PTT or cw key lines. These functions take the place of the mechanical SW1/SW2 function of the TU-300. XMIT-P/N are keyed by the front panel SEND/RECeive switch, the SEND-N or SEND-P inputs.

TU-300 Assembly

There are bound to be a number of readers who have constructed the TU-170 and are familiar with the (shall I say "concise"?) assembly instructions. In contrast, the '300 assembly manuals are quite detailed.

It took me a little over 6-1/2 hours to build the TU-300. Before construction began, information from some addenda sheets had to be transferred to a couple of the manuals (of which there are four, not counting the optional afsk unit: an operator's manual, and assembly manuals for the main/display, filter and demodulator boards.) These changes have been incorporated into the later editions of the manuals.

The main circuit board has provisions for a plug-in demodulator board, four filter boards, the optional afsk board and two expansion slots. Such construction makes assembly/disassembly and troubleshooting easy. You'll find that the plug-in boards can be installed backward, so be careful and pay attention to the illustrations. If you're installing the optional afsk unit, leave one end of D1 (at the gate of the output FET) lifted from each of the filter boards. Otherwise, you'll have to unsolder it later.

The unit I received called for some below-the-board changes to be made for better autostart/antispace action. I added these after first determining that the circuit was working normally otherwise. Later production units have these changes included.

During the review period, I received notice that a cw-demodulator option had been developed. This board is placed into one of the expansion slots on the main board. This option also requires the mounting of a front-panel switch for which a panel hole and space have been provided.

Low-tone kits for the 1275- and 1445-Hz frequencies are available, too. This option is not meant to be a modification for existing filter boards, however. Flesher recommends the filter boards be constructed initially with your choice of high or low tones.

Component quality is excellent. The pc boards are all glass-epoxy types, double-sided with plated-through holes, solder masked and with screened parts locations. A metal bracket fitted between the chassis side rails is slotted to accept the edges of the plug-in boards and secure them, as well as provide some additional structural strength to the overall assembly. The front panel/chassis assembly slides into the wrap-around steel cabinet and is secured by two screws at the rear. The complete unit is as rugged as it is attractive.

Alignment

This procedure consists simply of adjusting three potentiometers on each of the filter boards for maximum response. A number of alignment methods are offered, the first requiring no test equipment whatsoever! With this method, the output of the optional afsk unit is connected to the demodulator input by placing a wire jumper between two pins on the DB-25 connector. Since the tones produced by the afsk unit are crystal-controlled, they are smack on the nose — 2125 and 2295 Hz (measured with an Optoelectronics 7010A frequency counter) — with the 170-Hz filter in use. The afsk output adjustment potentiometer is set to produce a usable indication on the front-panel LED bar graph display, and the three potentiometers on the individual filter boards are adjusted for a maximum display. The SEND/REC switch is used to select the proper filter and tone. This is possible because, during assembly, certain selection diodes have not been soldered in place.

If you have a VTVM or an oscilloscope, I'd recommend you use it during alignment because you can obtain a finer degree of filter tuning, but I doubt you'd notice the results in on-the-air use. Should you not have ordered the afsk unit, you'll have to supply the proper tone frequencies to the demodulator. You can use an audio generator or you may try using the marker signal from your receiver or transceiver. Just make sure you have the proper input frequencies.

Once the filters are aligned, solder the remaining diodes in place and assemble the case. You're now ready to put the '300 to work. Since the '470 is up and ready to run when you get it, no alignment should be necessary. There are alignment instructions in the manual for each of the boards should you ever need to perform those tasks.

On Line

The rear panel TU connectors provide for interconnection among the station radio equipment, a 20- or 60-mA current loop teleprinter (with the optional '300 loop supply), and a computer or other device requiring RS-232C or TTL levels. A TTL-compatible SEND control pin permits placing the TU in the transmit mode by external control. The cw key input (for required station identification) is TTL compatible, too. The latter provides a shift of 100 Hz.

If you're not planning on using the loop supply, the manufacturer recommends the supply be disabled. Instructions for accomplishing this are given in the manual. The purpose of this is to eliminate excessive heat build-up in the chassis from power dissipated by the loop-keying transistor. The manufacturer also recommends the key-down time for the loop supply not exceed 15 minutes.

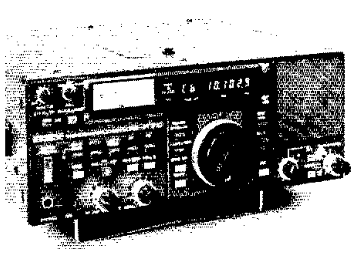
You'll find the TUs easy to use. Receiver tuning is done while observing the MARK and SPACE LEDs, and the LED bar graph indicator. Tuning is adjusted for a maximum display on the bar graph, with alternating action on the MARK/SPACE LEDs. (I would have liked to have had a bit more resolution on the bar-graph display at the upper end of its range.)

If you wish, you can connect an oscilloscope to observe the RTTY cross pattern. While not an absolute necessity, it provides you with a fine-tuning indicator and permits observation of selective fading of the incoming signal. (If you've got visitors in the shack, they're sure to be impressed!) The display consists of two ellipses rather than a perfect cross, but it is still quite usable.

Unlike the TU-170 and TU-470, there is no threshold adjustment in the '300; it is fixed. Under most conditions, the fixed level is such that good copy is obtained, but under some signal conditions, one might consider altering the threshold level to provide better copy. The TU-470 has an internal threshold-level adjustment potentiometer. It is R64 and is accessible through a hole in the right-hand side of the cabinet.

The TUs proved themselves to be good performers. I'm sure either will run rings around some others you've used, both commercial and "homebrewed."

The TU-300 and TU-470 are available from the Flesher Corp., P.O. Box 976, Topeka, KS 66601, tel. 913-234-0198. Kit price classes: TU-300, $300; optional afsk unit, $40; loop supply, $48; filters, $30 each; TU-470, $500. — *Paul K. Pagel, N1FB*



ICOM IC-R70 COMMUNICATIONS RECEIVER

☐ I was especially interested in reviewing the ICOM IC-R70 receiver because of my long-term interest in high-performance receiver design. Therefore, I did not hesitate to say "yes" when the review editor, N1FB, asked if I'd volunteer my time for the project! Over the years I have evaluated more inferior commercial receivers than good or excellent ones. It is difficult to rule out subjectivity when a person with an rf-engineering background lays hands on a piece of gear he or she did not design. But for the purpose of reviewing a product it is essential that we stick to the performance facts and ignore what we might have done differently in designing the unit. I like to compare the performance

ICOM IC-R70 Receiver, Serial No. 01234

Manufacturer's Claimed Specifications	*Measured in ARRL Lab*	*80 m*	*20 m*
Frequency coverage: 0.1 to 30 MHz, general coverage or ham-band only, via switch.	As specified.		
Modes of operation: cw, ssb and a-m; fm when optional fm module used.	As specified.		
Frequency readout: six-digit blue luminescent display.	3/8-inch digits. Has 2-digit mode indicator also. Skirt of tuning knob has analog increments to match the programmed frequency tuning rate of the receiver.		
kHz per turn of knob: 1 kHz, 10 kHz or 100 kHz (tuning rate switch selected).	As specified.		
Backlash: Not specified.	None discernible.		
RIT range: Greater than ± 800 Hz.	± 1000 Hz.		
S meter response (μV/S9): Not specified.	All amateur hf bands and 1.8 MHz; −61 dBm (200 μV); preamp off.		
AF/i-f notch filter depth: Not specified.	Approximately 70 dB.		
Receiver sensitivity (preamplifier on): cw, ssb, RTTY — less than 0.15 μV for 10-dB S + N/N. For a-m — less than 0.5 μV for 10-dB S + N/N above 1.6 MHz.	Noise floor (MDS) dBm	−130	−130
	Blocking DR (dB)	90.5	94.5
	Two-tone 3rd-order IMD (dB)	94	87.5
	3rd-order intercept (dBm)	+ 11	+ 12.5
Frequency stability: Less than 250 Hz from one minute after on to 60 minutes.	Less than 50 Hz.		
Audio output (8-Ω load): Greater than 2 W.	2.25 W.		
Noise blanker: Dual timing mode (selectable) for normal pulses and OTHR QRM.	Satisfactory.		
Synthesizer noise: Not specified.	Not measured, but very low noise in terms of reciprocal mixing, even when strong signals are nearby in frequency.		
Color: Black.			
Size (HWD): 4-3/8 × 11-1/4 × 10-7/8 in. (110 × 285 × 275 mm).			
Weight: 10.3 lb (4.85 kg).			
Power requirements: 117- or 234-V ac at 50-60 Hz, 30 VA.	As specified.		

against the cost and circuit complexity. That is, if a large number of components are used in the product, it stands to reason that performance in keeping with the price tag should be easy to realize. Basically, it is how the parts are used that determines the relative quality of the product. Two ingredients are essential: (1) knowledge of the amateur's operating and equipment-performance needs, and (2) competent design engineering.[3] It appears to me that the criteria were met with respect to the IC-R70 receiver. I would definitely rate this unit as a clean, high-performance box.

Significant Operating Features

1) General coverage (0.1-30.0 MHz) or ham-band only (1.8-28 MHz, including the WARC bands), at the push of a switch.

2) Quadruple superheterodyne circuit with continuously variable bandwidth control.

3) Filters for a-m, ssb, RTTY and cw are standard equipment.

4) Selectable preamplifier (panel switch).

5) Selectable tuning rate of 1 kHz, 100 Hz or 10 Hz per step.

6) Two internal VFOs. Can be used independently in any chosen band.

7) Band-pass tuning.

8) No preselector to peak (contains broadband front-end filters), thereby permitting rapid UP-DOWN band changing by means of panel buttons.

9) A panel switch permits turning on a monitor for the transmitter signal, thus enabling the operator to check the actual signal quality.

10) Variable-frequency notch filter.

11) Front-panel RTTY mode switch automatically selects the cw (500 Hz) and ssb (2.3 kHz) i-f filters for RTTY reception. Both filters are used for RTTY.

12) VFO memory switch permits retaining the programmed VFO frequencies even when the main power switch is turned off. There are numerous other features, but those mentioned should be of special interest to prospective buyers.

Circuit Highlights

The following features are probably of interest to the performance-conscious amateur or engineer:

1) Drift: 250-Hz maximum first hour; less than 50 Hz after one-hour warmup.

2) Passband tuning is ±500 Hz for ssb, cw and RTTY; ±2.7 kHz for a-m.

3) A 250-Hz cw filter (FL-63) is available as an option. A 500-Hz cw filter is supplied with the receiver.

4) An internal frequency-adjust control permits the operator to align the PLL oscillator to WWV at 10 MHz.

5) A panadaptor/scope monitor jack is located on the rear of the chassis. It samples the 70-MHz i-f at the first mixer output.

6) The selectable preamp is broadband and has a gain of 10 dB. The attenuator (same switch) reduces the signal by 20 dB.

7) A DBM (diode ring) is used as the first mixer. The i-f of this mixer is 70 MHz. A monolithic filter follows the mixer.

The PLL contains three phase-locked loops. Output from the first is 13.15 to 23.14 MHz, with 10-kHz steps. The loop output is divided by 100 and used for the reference frequency of the second loop. The LO frequency for the first loop is 20.48 MHz.

The second-loop output is from 30.8515 to 30.9514 MHz, with 10-Hz steps. The local oscillator is a VXO from which the output is tripled to 30.72 MHz.

The third (main) loop has an output from 70.4515 to 100.4514 MHz. This output is used as the first LO for the receiver. Four VCOs are employed to divide the LO frequency range into four segments.

Performance Impressions

I put on my demon's mask, smiled sardonically and subjected the IC-R70 to the customary W1FB receiver-torture test — just two blocks from the 1-kW, multiband onslaught of W1AW. I fully expected to see the excellent receiver dynamic range negated by reciprocal mixing problems resulting from synthesizer output noise. I was amazed to note that no discernible evidence of the malady existed, even a few kilohertz away from the W1AW 80-meter frequency!

Why check this on 80 meters? Well, I once terminated my 80-meter vertical antenna in 50 ohms and measured the W1AW energy across the resistor with a Tektronix 453 scope: The rf level was 5 V peak to peak! It is not nearly so high on the other hf bands. So, 80 meters has always been my most difficult challenge with respect to keeping receivers from "crunching" at my QTH. I did not observe blocking, cross-modulation or excessive IMD responses while testing the receiver on 80 meters (or any other band), and I was able to copy weak cw signals as close to the W1AW frequency as 5 kHz. There was no apparent performance degradation when I switched in the 10-dB preamp, and that *was* a surprise!

If I were to "pick nits" about the performance, I would mention that the agc tends to lock up more readily from strong signals in the passband than is characteristic of some other receivers I have used. Also, the audio-output signal becomes distorted at room-listening level and higher when using the speaker (see note 3), even though the af channel is rated for a minimum of 2-W output. This malady seems to afflict most of the solid-state receivers we have tested, but it is not a problem when headphones are used or when the audio-speaker level is moderate. If you weren't born with "golden ears" you may not observe this condition. I'm just a grouch about audio quality, even in communications receivers.

I was delighted to note an absence of agc-caused clicky signals. The attack time is entirely acceptable, and the decay time can be controlled by means of a panel switch to provide fast or slow agc. There is also an agc-disable switch position, which comes in handy when a big signal does lock up the agc circuit and desenses the receiver.

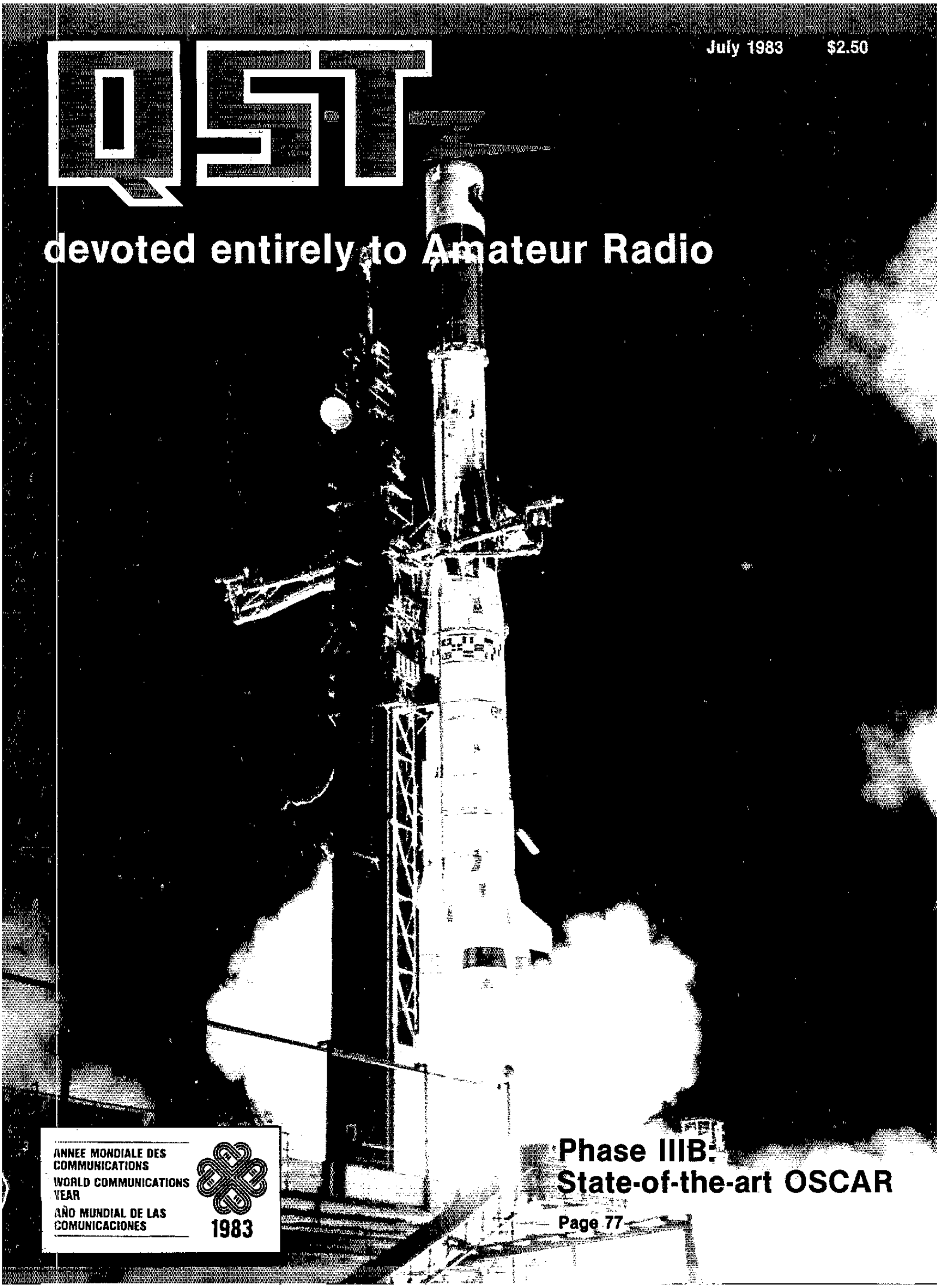
Noise-blanker operation is good, and there is no appreciable deterioration of the receiver dynamic range observed when it is actuated. I am favorably impressed, because many receivers completely "fold up" at my QTH when the blankers are turned on.

This receiver should be fine for amateur use. It would also be a good choice as a laboratory-grade instrument. I'd like to own it for that purpose myself!

Price class: IC-R70, $750. Available from ICOM America, Inc., 2112 116th Ave. N.E., Bellevue, WA 98004.

— Doug DeMaw, W1FB

[3] D. DeMaw and W. Hayward, "Modern Receivers and Transceivers — What Ails Them?" *QST*, Jan. 1983.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OUR COVER
Far more than just a pretty launch, AMSAT-OSCAR Phase IIIB represents a prodigious accomplishment. Hams around the world should now have a spanking-new spacecraft to use, thanks to the volunteers and contributors who have made it possible. *(photo courtesy European Space Agency)*

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An Introduction to AMTOR

This mode adds space age excitement to RTTY!

By Paul Newland,* AD7I

On January 27, 1983, the FCC approved the error-correcting teleprinter protocol known as AMTOR for use by American radio amateurs. What is AMTOR, anyway?

AMTOR (*Am*ateur *T*eletype *O*ver *R*adio) is an exciting new mode of radioteletype (RTTY). It offers all the utility and enjoyment of conventional RTTY plus one great advantage: AMTOR removes virtually all errors caused by interference, fading or other disruptions. It also has built-in capability for selective calling. If your station is configured for AMTOR, it has the option of not responding to ASCII, Baudot or cw; only AMTOR signals will print on your teleprinter. This new system should prove useful for avid RTTY operators and traffic handlers as well.

I will provide an overview of how AMTOR operates and how amateurs can make use of it. Those interested in more of the technical details of AMTOR are encouraged to read the excellent *QST* article by Peter Martinez, G3PLX.[1]

AMTOR is a derivation of the commercial teleprinter system, SITOR, which stands for *Si*mplex *T*eletype *O*ver *R*adio.[2] SITOR is a commercial hf RTTY error-correction system used by many ships and coastal marine radio stations. SITOR allows a shipboard teleprinter to be connected to the international Telex network via hf radio so ship owners can send messages directly to their ships in real time. In addition to having error-correction capability, SITOR also includes selective calling. With this feature, coastal stations can call any ship that is monitoring, and forward a message to it without error or operator intervention.

AMTOR, in most respects, is the same as SITOR. With an AMTOR code converter, however, it is possible to monitor other stations using either AMTOR or SITOR operating in Mode A. This feature is important for radio amateurs so they can continue to self-police their bands. Most SITOR code converters do not have this capability.

Modes of Operation

AMTOR features four modes of operation.[3] They are: A — Automatic Request for Retransmission (or fills), B — Broadcast, L — Listen, and S — Selective Broadcast. Before discussing the modes in detail, it is useful to define some terms:

Master Station (MS) is the station that starts the communications. It does not show who is sending data at any one time, but only who began the contact. All stations synchronize their transmissions to those of the master station.

Slave Station (SS) is the station that is called by the MS.

Information Sending Station (ISS) is the station that is generating or sending data. I will sometimes use the term *sending station* when I mean, more correctly, ISS.

Information Receiving Station (IRS) is the station that is receiving and printing data. I will sometimes use the term *receiving station* when I mean, more correctly, IRS.

Of all the modes, Mode A is the most immune to errors, but is also the most complex. In Mode A, both the ISS and IRS send bursts of data to each other. This requires both stations to operate in a VOX-like manner. The ISS sends three AMTOR characters in a burst, and the IRS sends one AMTOR character in a burst. After each station sends a burst, it switches to receive so it can detect and decode bursts from the other station. This requires radios capable of switching from transmit to receive or receive to transmit in 10 to 20 ms. Most modern ssb radios meet this requirement.

To begin a QSO in Mode A, the operator tells the AMTOR code converter the selective calling characters of the station to be called.[4] The MS code converter begins to send the selective call characters, using bursts of three characters and switching to receive between the bursts. The gaps between the bursts provide time to listen for the receiving station. The receiving station, when correctly decoding its own selective call characters, will respond with a special AMTOR signal that says, "I hear you and I am ready." When the MS code converter hears that signal, the MS becomes the ISS, and it will alert the operator (usually through the use of LEDs on the code converter) that the IRS is ready to copy. The ISS may now send data to the IRS. During communications, the ISS (either the MS or SS) transmits characters in groups of three and the IRS replies with a one-character AMTOR signal that says either, "I received the last group of three correctly; send the next three," or "I missed that last group; send it again." Because AMTOR enables the ISS to repeat characters until they are received without error, the number of errors actually printed on the teleprinter is reduced drastically in comparison to conventional RTTY.

When the ISS wants to let the other stations send data, it sends the three characters, FIGS Z B.[5] This sequence of three characters is the AMTOR "over" signal for Mode A. Now, the IRS becomes the ISS, and vice-versa. Similarly, when the

[1]Notes appear on page 13.
*P.O. Box 205, Holmdel, NJ 07733

new ISS wants to convert back to an IRS, the operator types in the characters FIGS Z B. In addition, the IRS station is able to break into the ISS transmission at any time. To do this, the IRS operator presses the OVER button on the code converter. Instead of the IRS sending the usual "send it again" or "give me the next block of data" bursts, it responds with another special one-character message that says, "I want to send you some data; let's exchange places." When the ISS receives this message, the exchange is made and the ISS may begin sending data.

When the ISS wants to close the QSO, the operator presses the QRT button on the code converter, causing both converters to "handshake" and go off the air. If the IRS wants to close the contact before the ISS is finished, the IRS can "break" into the ISS and exchange places. The operator may then press the QRT button to close the contact.

Listen for AMTOR over W1AW

If you can receive AMTOR, try listening for W1AW AMTOR transmissions, Mode B, following the 11 A.M. EST/EDST (presently 1500 UTC) RTTY/ASCII transmissions Monday through Thursday on 14.095 MHz. W1AW has been transmitting AMTOR bulletins since February 8, first under an STA (Special Temporary Authority granted by the FCC) and continuing after the Commission authorized AMTOR for general use in the Amateur Radio Service. The W1AW AMTOR schedule will soon expand; watch the W1AW schedule, published in alternate issues of *QST*, for full details.

W1AW has also made several two-way contacts using Mode A, mostly with stations in Europe, where AMTOR has been popular for some time. It's still a thrill to have a DX station say, "Please QSL — you're my first U.S. QSO on AMTOR!" — *Chuck Bender, W1WPR, Chief Operator, W1AW*

Mode B

Mode B is useful for broadcasts when the sending station wants to reach more than one station simultaneously, or when the IRS cannot send acknowledgements to the ISS. Examples of this type operation are W1AW bulletins or net operations.

In Mode B, the transmitter is on continuously, unlike Mode A, in which the data is sent in bursts. The sending station AMTOR converter sends each teleprinter character twice, with the second transmission of a character delayed from the first by 280 ms. These transmissions are made synchronously, as opposed to conventional RTTY, which uses start and stop signals before and after each character. In synchronous communications, all data characters are sent one after the other, without start and stop signals, and at a rate that is controlled by an accurate time base, such as a crystal oscillator. When there are no characters from the teleprinter to send, the code converter fills this "dead time" with AMTOR idle characters. These characters are used to fill the periods of time when the sending teleprinter is not generating data, but they do not print on the receiving teleprinter; they are like a conventional RTTY "diddle."

The advantage of Mode B, compared with conventional RTTY, is that the teleprinter only receives data from the code converter when there are valid AMTOR signals present. Because the characters are transmitted in pairs, AMTOR B stores the first character in memory and waits for the second. When it receives the second character, it checks both in the pair for errors. If one of the two characters has been received correctly, it is printed. If both are in error, a space will be printed to mark the error. This space is easy for the operator to spot and provides a place on the hard copy for a manual fill to be made.

Mode S

Mode S is just like Mode B, except that it includes selective calling. It is rarely used today, although it probably will be used often in the future. Mode S is useful if you are unable to operate in Mode A and only want to receive messages that are specifically intended for you. Teleprinter paper can be used up at a fast pace if you are on a channel that is active with Mode B traffic!

Mode L

Mode L is useful for monitoring two stations that are using Mode A to communicate. Of course, since the sending station is not getting acknowledgments from the Mode L station, but instead from the other Mode A station, the Mode L station will not obtain the benefits of error correction. This is hardly a problem, because the only purpose of this mode is to allow for monitoring, not error-free communications.

Which Mode?

It is important to understand that for all modes except L an AMTOR controller that is monitoring a channel will automatically switch to the proper mode when it is called by an MS. When someone calls your station, you don't have to be concerned with which mode is being used. The AMTOR controller will determine whether the call is in Mode A, B or S, and will adapt to that mode immediately. Only when a station *places* a call does the operator choose what mode to use.

Error Detection

AMTOR is able to process data without error because it rejects data that it believes is incorrect. How does AMTOR know when the data is wrong? It converts each five-unit Baudot character to a seven-unit character that always has four 1s and three 0s before transmission. At the receiving station, AMTOR checks each received character to see if it meets the 4/3 criterion. If it does, the code converter assumes that the data is correct.[6] If it does not, the character is assumed to be in error.

Even though this method of error detection may appear to be overly simplistic, it provides adequate performance for "outwitting" the error-causing phenomena found on hf radio circuits.

How to Hook It All Up

AMTOR equipment is available for moderate cost and is *simple* to hook up. Those totally unfamiliar with RTTY might want to review the information on this mode found in the ARRL *Handbook*. The basic equipment used for AMTOR is the same as that used for normal RTTY, with the addition of an AMTOR code converter between the terminal unit (modem) and the teleprinter. As an added plus, with AMTOR you don't need a terminal unit with auto-start, mark hold, and so on. The AMTOR code converter provides all those features for you.

Where and When

Most AMTOR operation is centered around 14,075 and 3637.5 kHz. These two frequencies are typically used for calling. After contact is established, and if both operators are present(!), they will usually move the QSO to another frequency.

Some stations have their radios configured for scanner operation. They will tell other AMTOR operators the frequencies they scan and what their hours of operation are. When someone wants to contact them, they locate a frequency that the other station is known to scan, make sure it is inactive, and then begin a Mode A call. Usually, if the "receiving" station is not in a QSO on another channel, it will respond to the Mode A call within 40 seconds and the QSO can proceed. If stations are using scanners, there is no need to move off channel because the QSO can begin on a *working* channel, instead of a calling channel.

Some stations, such as G3PLX and HB9AK, monitor 14,075 kHz continuously and have computers attached to their AMTOR systems. The computers can be used to store or retrieve messages intended for their station or others. Some AMTOR computer systems can also act as store-and-forward repeaters. When you call these systems, you can request that your message be relayed to another station on completion of your QSO.

As an example of this capability, G3PLX and I were trying to pass some information on 20 meters, and the path between us was just not there; we found that by relaying messages from G3PLX to A4XFW in Oman, to HB9AK, to AD7I near New York City and back again, we could communicate. It is noteworthy that there were *no* operators at HB9AK or A4XFW during these relays. [This type of unattended transmitting is not presently allowed under FCC rules without special authorization. — Ed.] Everything at these two stations was under computer control. Now *that* has to be a new form of long path!

Simple, But Robust

I have been involved with AMTOR for some time now and I continue to be impressed with its robust capabilities, while maintaining simplicity. It has proved to be a simple and inexpensive addition to any amateur RTTY station. Finally, the most compelling reason I can offer for becoming active with AMTOR is that RTTY is more fun without errors!

Readers' comments and suggestions are always welcome. I would like to hear from anyone interested in getting involved with AMTOR. Please include an s.a.s.e. with your correspondence.

Notes

[1] J. Martinez, "AMTOR, an Improved Error-Free RTTY System," *QST*, June 1981, pp. 25-27.

[2] The exact specifications of SITOR are outlined in CCIR (International Radio Consultative Committee) Recommendation 476-2.

[3] Not all controllers include all modes. If you consider building or purchasing a code converter, you should know which capabilities you want. In addition, if you purchase a commercial AMTOR code converter, make certain the unit meets or exceeds all specifications outlined in CCIR Recommendation 476-2 for the modes you want to use.

[4] AMTOR uses four-letter combinations for selective calling. With a 26-letter alphabet, this provides for 456,976 unique calls. Although it might be desirable to have more calls available, this is all the internationally agreed upon standard allows. Most amateur stations use the first letter and last three letters of their call sign for the selective calling characters. Using this algorithm, the calling characters for W1AW would be WWAW.

[5] On CCITT no. 2 coded terminals, FIGS Z B will print as ±?. On terminals commonly used by American radio amateurs FIGS Z B will print as "?.

[6] This leads us to why I must say that AMTOR provides *virtually*, rather than *absolutely*, error-free communication. There is a chance, although small under normal conditions, that some "hits" during transmission will result in a data error that will pass the 4/3 test at the receiving station.

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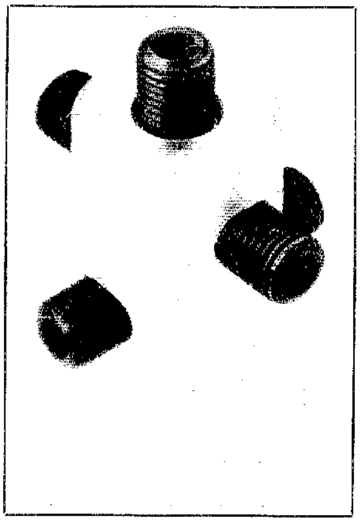
Paul Newland, AD7I, was first licensed in 1971 and is currently active on hf and vhf with new modes such as AMTOR and packet radio. He is presently focusing his Amateur Radio activities on development of hardware and firmware for AMTOR code converters and modems. Professionally, Paul is employed by a major research and development company where he designs microprocessor-based communications interfaces for computers. He received a BSEL degree from California Polytechnic State University in 1979 and a MSE-ECE from the University of Michigan in 1980. Paul is a member of the ARRL and IEEE.

New Products



X-PANDA-FIVE

☐ A new accessory for the hf mobile enthusiast is available from JL Industries. The X-PANDA-FIVE is an adaptor that permits the installation of up to five resonators on one mobile mast.

The manufacturer claims ease of installation and convenience of operation. The mobile operator no longer needs to stop to change resonators — a distinct advantage, especially during bad weather. In addition, hams who own motor homes or RVs, or who live in modular homes, apartments or condominiums, might find the X-PANDA-FIVE a useful accessory for erecting a multiband hf antenna where limited space is available.

Complete installation and resonator tuning instructions are furnished with the X-PANDA-FIVE. The adaptor is machined from solid aluminum stock and incorporates resonator mounting studs of high-tensile carbide steel. All threads are 3/8" × 24 to accept either Hustler or HyGain resonators and masts. The suggested retail price is $14.95, and the unit is available direct from the manufacturer, JL Industries, P.O. Box 030413, Fort Lauderdale, FL 33303 — *Sandy Gerli, AC1Y*

UNIMEC SWITCHES

☐ A/S MEC-75 of Copenhagen, Denmark, offers a unique modular switch available in two forms: momentary contact and push-push action. Both types provide five different contact configurations. These are determined by selecting the appropriate switch terminals with pc-board traces or hard wiring. Essentially, each switch is a programmable 4pdt type.

In addition to the two basic switches, MEC offers a 10-color variety of keycaps and bezels. Bezels illuminated by LEDs are available, too.

MEC also provides a matrix mounting system (UNIMEC/VARIO-SUPPORT) for their switches. This enables a designer to realize a custom keyboard or switch panel almost instantly. The desired layout and switch functions are determined simply by inserting the appropriate switches into the matrix. According to the manufacturer, this makes panel mounting easy and stops over-pressure as well as flexing of the pc board.

Switches and pricing information are available from Forsbert Sales, Inc., 646 Summer St., Brockton, MA 02402, tel. 617-522-6300, and Switches, Inc., 949 Stierlin Rd., P.O. Box 4248, Mountain View, CA 94040, tel 415-962-8649. — *Paul K. Pagel, N1FB*

Strays

QEX: THE ARRL EXPERIMENTERS' EXCHANGE

☐ The June issue of *QEX* featured:

- "Many Modifications to the Ten-Tec Omni," by Robert E. Helms, AF5Z.
- "Lowpass Speech Filter Using Surplus Inductors," by Ed Wetherhold, W3NQN.

It also reviewed the Ninth Annual Eastern VHF/UHF Conference and carried a request from Hank Magnuski, KA6M, for information from potential users of the AMSAT OSCAR Phase IIIB AMICON Special Service Channel, which has been designated for packet-radio operations.

This issue, the 16th, was the first one produced at ARRL Hq. Please send all subscription requests, manuscripts and correspondence for *QEX* to ARRL Hq. The domestic subscription rate for ARRL members is $6 for 12 issues; $12 for nonmembers. There are additional postage charges for mailing outside the U.S.; write to Hq for details.

DX SERVICE AVAILABLE TO BLIND OPS

☐ The Braille DX Service promotes DXing among blind hams by providing a monthly cassette recording of current DX activity, including QSL information, propagation reports and DXpeditions. For more information, contact Phil Scovell, AFØH, 8347 West 6th Ave., Lakewood, CO 80215, tel. 303-233-4335.

Spread-Spectrum Applications in Amateur Radio

Through the properties of their coded modulation, spread-spectrum systems can provide multiple-access, low-interference communications to radio amateurs.

By William E. Sabin,* WØIYH

Traditionally, the emphasis in Amateur Radio has been to make a transmitted signal as narrow in bandwidth as possible. Also, receivers are made as narrow and as interference-immune as possible. In this way, many signals can occupy a ham band successfully. This approach has been successful, to a point. But if a group of stations is on one frequency (the pileup!), the system does tend to break down, with disastrous results.

A new approach is being advanced that amateurs should take a look at. Military and commercial organizations are developing spread-sprectrum systems. Such systems deliberately occupy a wide band of frequencies, as part of a strategy to make communications more reliable and more secure, or private. (The word "privacy," as used here, has a special meaning in Amateur Radio, which will be considered later.)

To be more exact, a transmitter sends its message in such a way that a wide spectrum is used, according to a very carefully designed plan. The receiver has the ability to use this same plan in reverse, to convert the signal back to narrow-band form. By performing these actions in the right way, privacy and interference immunity are improved. Fig. 1A shows a conventional transmitter output of, say, 1 kW. With spread-spectrum operation, this same power is spread out as shown in Fig. 1B. There is no strong carrier at any one frequency. Within a 3-kHz band, the amount of signal is greatly reduced. In fact, it may be less than the noise level. But after "despreading," the signal once again looks like the signal at A.

This scheme is different from wide-band fm in that the message itself does not produce the spread spectrum. Instead, another agent is employed to spread the signal. Also, there is no carrier, as in an fm system.

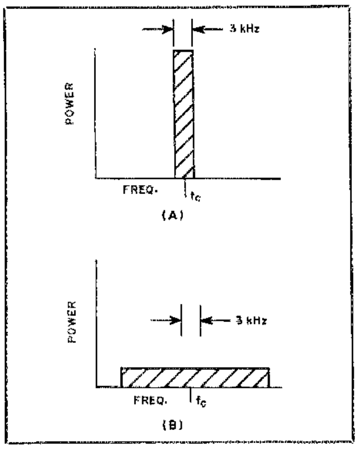


Fig. 1 — The power distribution of a conventional communications signal (A) versus a spread-spectrum signal (B). The same total power is contained in both signals.

Why Amateur Spread Spectrum?

Consider the network shown in Fig. 2. Using spread spectrum, stations A and B can communicate privately, while C and D do the same. Or, A can address all stations. Also, a member of this net can address an entirely different net or a single member of the other net. Or, B could address all nets simultaneously. In a particularly large region, these nets all use the same frequency band, and no equipment retuning is needed for any of the above operations.

In any of these operations, a degree of "privacy" is achieved, in the sense that communications are programmed according to the requirements of the moment. This "selective calling" is achieved by using the microcomputer-based "protocol," or message-routing procedures. The use of spread spectrum is an enhancement of the "packet radio" techniques that advanced amateurs are now experimenting with. By adding to this packet system a carefully managed spread-spectrum protocol, it should be possible to greatly reduce "collisions," avoid interference and add significantly to the repertoire of the packet system.

This extra element of spread-spectrum management, in addition to time management, makes it possible to reduce the guesswork with respect to frequency selection, which is a major problem in Amateur Radio. The difficulty is that a clear frequency at my station may not be clear at

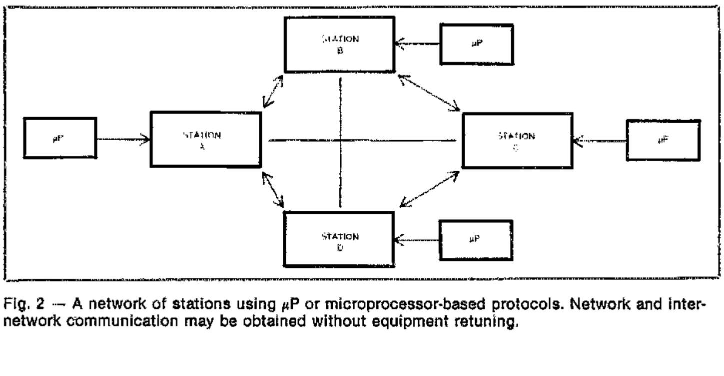


Fig. 2 — A network of stations using μP or microprocessor-based protocols. Network and inter-network communication may be obtained without equipment retuning.

*Rockwell-Collins, Telecommunication Products Div., Cedar Rapids, IA 52498

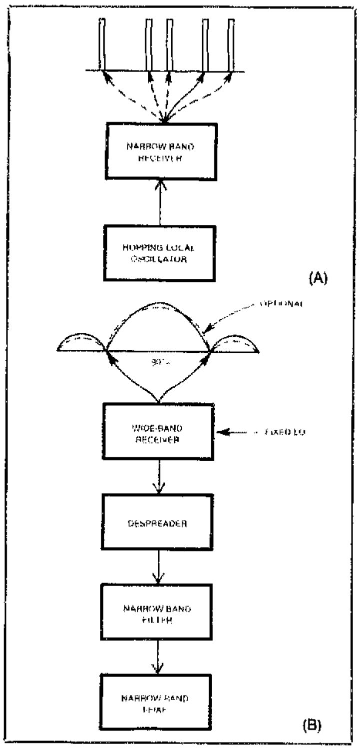


Fig. 3 — The two major types of spread-spectrum operation. At A, the frequency-hopping mode is illustrated; at B is the direct-sequence mode. The spreading is caused by applying a coded modulating signal at the transmitter that is independent of the intelligence modulation. The same predetermined code is applied in reverse at the receiver to despread the signal, and the intelligence is then demodulated in conventional fashion. If different spreading codes are appropriately chosen, interference-free operation may be obtained with different stations sharing the same frequency spectrum.

your station. We need a way to improve this situation. The communication is not restricted to data. Digitized voice messages can be sent, stored at the receiving station and converted back to speech at the completion of the message.

The information below summarizes the possible advantages of combining spread spectrum with packet-network protocols. There is no particular argument in favor of using spread spectrum by itself.

- Voice or data
- Simultaneous net combinations
- Network privacy
- Reduced collisions between
  - A) spread and nonspread systems
  - B) other spread systems
- Combines well with packet protocols
- Enhances packet repertoire

The important extra element is the voluntary "discipline" to which the various players can subscribe. The discipline is handled by the various personal computers involved, leaving the operator with a great deal of freedom.

Fig. 3 illustrates the two major types of spread-spectrum reception that amateurs might consider: frequency hop and direct sequence. At A, a narrow-band receiver is rapidly tuned through a predetermined set of frequencies. The desired signal is available at each of these frequencies when the receiver tunes there, according to a predetermined plan. Other signals are programmed so that they are seldom on the frequency to which the receiver happens to be tuned at one particular moment, although they may share this frequency at different times.

At B, a wide-band receiver, tuned to a fixed frequency, listens to a signal that has been carefully spread out in the manner shown, according to a predetermined plan. The receiver possesses the key by which to despread this signal and put it through a narrow filter. All other signals using the same frequency band are essentially ignored. A different spread-spectrum signal fails to despread. A conventional signal is likewise unable to penetrate the narrow filter, because it is converted to a wide-band signal by the receiver.

On the hf bands, frequency hopping would be better because of its narrow-band nature. At uhf, direct sequence offers advantages. By carefully selecting the kind of modulation in direct sequence, the spectrum is improved as shown in the dashed line in Fig. 3B. This could be msk (minimum-shift keying) modulation.

Direct-Sequence Spreading

Fig. 4A shows a conventional phase-shift-keyed data signal and the transmitted spectrum it produces. The width of the main lobe is twice the data rate, and in an amateur RTTY system would be less than 1.5 kHz wide. In Fig. 4B, each data bit is modulated by a PN code, where PN means pseudo-noise. There may be thousands of PN code bits for each data bit. The data bit has the effect of inverting, or not inverting, a group of PN bits, depending on a data 1 or 0. After combining these code streams, a mixer with a fixed-frequency local oscillator produces the spread-spectrum rf signal. The width of the main lobe is twice the PN code rate.

The design of the PN code is critical to the performance of the spread-spectrum system, and will be covered in the next section. The spectrum in Fig. 4B shows that, when listened to by a conventional receiver, a weak, hissing, noise-like signal is heard. It is not truly random noise, because of the nonflat nature of the spectrum, and because the PN code does repeat itself after some long time interval. The expression "pseudo" noise is therefore appropriate.

Fig. 5A shows a PN code generator. A shift register with N stages, initially loaded with all 1s, generates an output that is

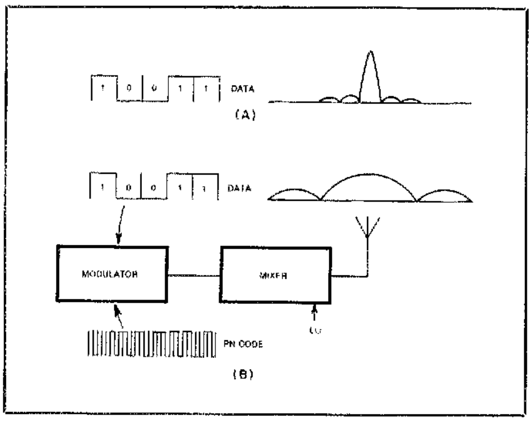


Fig. 4 — At A, a conventional phase-shift-keyed (psk) data signal and the transmitter spectrum that it produces. At B, the same data signal and a pseudo-noise (PN) code are combined to produce a direct-sequence psk signal with its much broader spectrum.

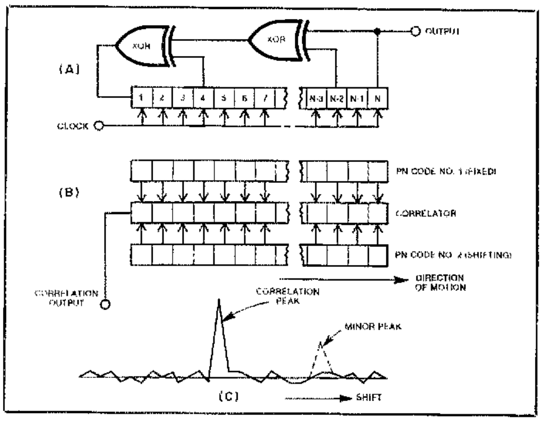


Fig. 5 — A PN code generator, A, and a correlator, B. The strings of blocks represent shift registers. At C is shown the correlation output with shifts of data in the lower shift register. This output peaks when synchronization is obtained.

influenced by the XOR feedback gates. After as many as $2^N - 1$ pulses, the code repeats itself. Code length can be from, say, 10 bits, to 10 million or more. If my transmitter and your receiver use the same PN generator circuit, we can communicate.

Fig. 5B shows a correlator. One input is a fixed (or static) PN code that has been stored. The other input is the received signal, which ripples through the bottom section. The output of the correlator, Fig. 5C, is very small (± 1 or 0) except when the two codes exactly coincide, at which time the output equals N. When this peak occurs, the two codes are "synchronized," or correlated.

A well-designed code has only one of these sharp peaks. Other codes will have other minor peaks that can produce "false" synchronization. Also, a short segment of a good code may not be so good in this respect. Therefore, the register length should be as long as the code, or at least as long as possible. A good possibility for amateur use is to combine two short Gold codes.[1] The resultant codes have good correlation properties and are easy to generate.

Fig. 6 shows two types of direct-sequence receivers. In Fig. 6A, the despreading is done at i-f. A narrow-band crystal filter lets only the despread signal get into the data or voice detector. In B, a double conversion takes place, and the correlator, operating at baseband, performs the despreading. A low-pass filter then passes the desired signal only, which can be derived from the correlator output. Some receivers use both methods, that at A for data detection, and that at B for synchronization.

Direct-Sequence Synchronization

Fig. 7 shows, in a very general way, how synchronization is achieved and maintained. One way is to slow down or speed up the PN code generator in such a way that it can search backward and forward in time to "acquire" the incoming code. Once the code has been acquired, a tracking operation takes place, so that the PN code stays closely aligned with the desired signal. Acquisition is greatly aided by the use of very stable clocks and by prealignment of the codes, so that only a small amount of searching is needed. For example, using WWV, all code generators could be initialized each hour.

Another method uses a special, short preamble that the receiver quickly recognizes. This recognition starts the tracking operation. In an amateur application, this would be easier and cheaper to implement. In this case, the prealignment

[1][Editor's Note: Gold codes are a family of codes named after the developer, R. Gold. The significance of the family is that a large number of codes may be obtained with relatively short shift registers in the code generator.]

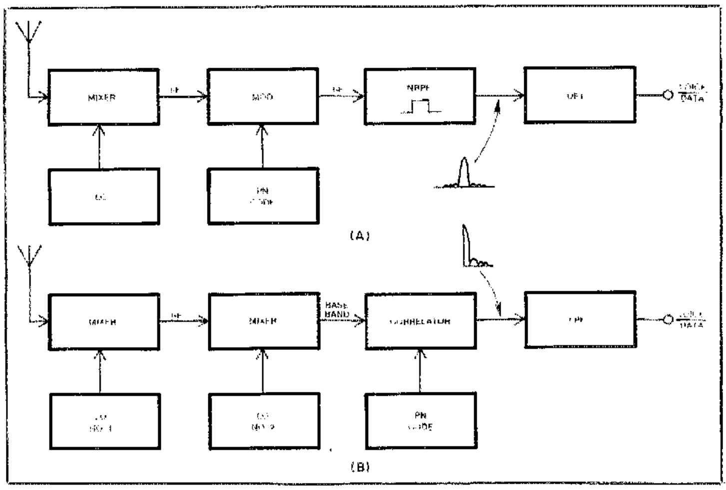


Fig. 6 — Block diagrams of two types of direct-sequence receivers. See text.

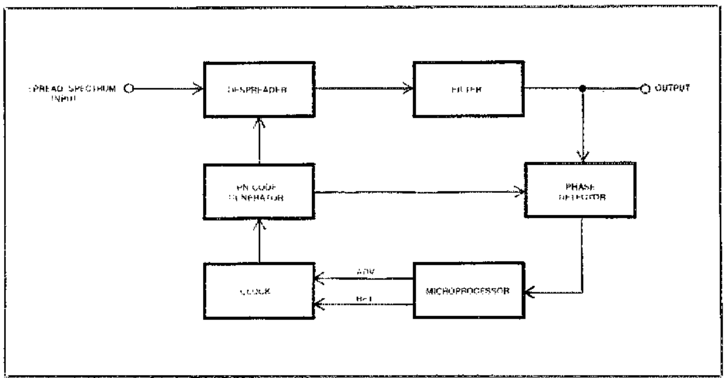


Fig. 7 — A system for synchronizing and tracking with direct sequence.

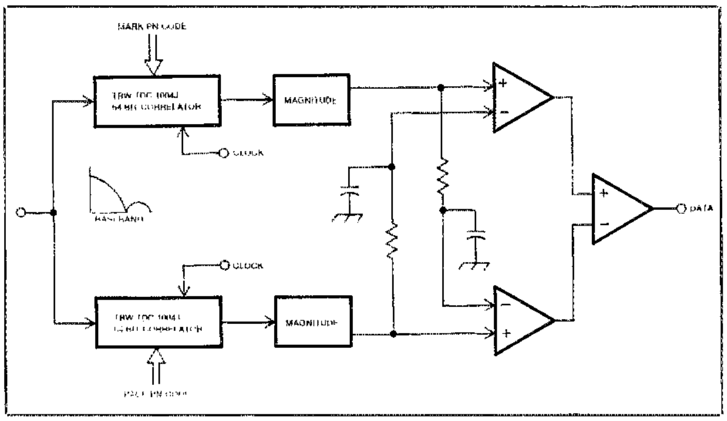


Fig. 8 — A correlator that operates at baseband with direct sequence.

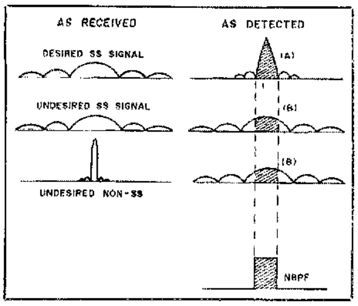


Fig. 9 — Illustrating interference rejection in direct sequence. Processing gain is obtained with despreading, and in decibels is equal to 20 log (A)/(B), where (A) and (B) are the amplitudes depicted in the shaded areas at the right.

of codes and time-of-day clocks would not be needed. When the preamble is sent, it is immediately followed by the start position of the long PN code. The receiver performs the identical operation. Sometimes, the preamble is sent as a frequency-hopping signal rather than direct sequence.

The circuit of Fig. 8 uses two TRW 64-bit correlators. One recognizes a PN sequence that identifies a mark, and the other recognizes an unrelated sequence that signifies a space, using bi-phase modulation. The correlator outputs can be positive or negative, so a full-wave rectifier (using op amps and diodes) is needed.

Each comparator looks at the instantaneous output of its correlator and the average output of the opposite correlator. This circuit, with very little modification, can detect marks and spaces in a frequency-hop radio, in which up to 64 bits of information would signify the message bit.

Fig. 9 shows how a desired spread spectrum and an undesired signal, either spread or narrow-band, are interpreted by a spread-spectrum receiver. The narrow-band signal is "smeared" by the receiver so that little energy is passed by the narrow band-pass filters. If the undesired local signals are strong enough, however, they can still override the desired distant signal, even after it has been despread, as shown. This "near-far" problem in direct sequence is an important limitation. Frequency hop seems to be more immune.

Frequency-Hop Spreading

Fig. 10 is a block diagram of a frequency-hop transmitter. A fast-frequency-change local oscillator is needed. The time that it takes to settle on the next frequency should be less than 10% of the dwell time on that frequency. Hop rates of 10 to several thousand per second are feasible with today's technology, and amateurs should experiment with speeds over the entire range. For example, a real opportunity exists for innovation in a low-cost, fast-hopping synthesizer design.

Shortly before changing frequency, the signal is smoothly attenuated, as shown. After the hop, the signal is brought up again, smoothly. This is necessary to reduce the transmitted spectrum (key clicks) as much as possible and to allow the LO time to make its frequency change. The hopping pattern is all under microprocessor control and determines which station or network will be addressed.

The analog voice or fsk data is filtered by the narrow band-pass filter. In other words, frequency hop is basically a narrow-band mode at any one frequency. Following the mixer, wide-band amplifiers and a wide-band antenna are needed so that equal power output occurs at each frequency.

For voice, a-m (with carrier) is preferred to single sideband because ssb causes phase jumps between hops, which produce excessive noise and distortion. Analog voice is not a preferred mode, in general, in frequency hop. At very low hop speeds, however, ssb should be considered as a possibility.

Fig. 11A shows the transmitted signal, smoothly attenuated between hops. The dashed lines show possible amplitude variations with a-m operation. Fig. 11B shows the spectrum. The tapered Christmas-tree shape at each frequency is caused by the turning on and off (Fig. 11A), and also by

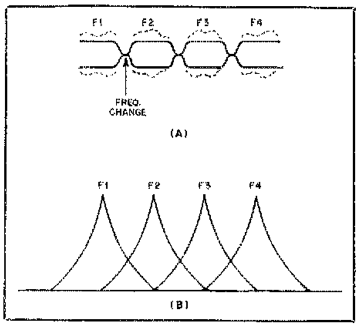


Fig. 11 — At A, an amplitude-versus-time representation of a frequency-hopping signal. The broken lines indicate possible variations arising from amplitude modulation of the signal. At B, an amplitude-versus-frequency representation of the same signal; any predetermined arrangement of frequency hopping may be used.

the mark/space information or a-m sidebands. To minimize interference, the drop-off in the spectrum should be as fast as possible, consistent with good communication. We can also visualize that the receiver bandwidth should be only wide enough to receive, say, 90% of the total signal energy at each frequency.

A frequency-hop receiver is shown in Fig. 12. An antenna input switch controls the turn-on at each frequency in a way that reduces intermodulation with strong, undesired signals on nearby frequencies. After mixing, a narrow filter leads to the signal detectors. The outputs of these detectors provide signal information to the microprocessor to control the synchronization algorithm and to determine that sync has occurred. When synchronization is achieved, a tracking operation is started in which the hop clock rate is adjusted momentarily. Later paragraphs cover this topic more thoroughly.

Instead of a single filter plus discriminator, consider two narrow filters, one for mark and one for space. The outputs are rectified and compared to determine the

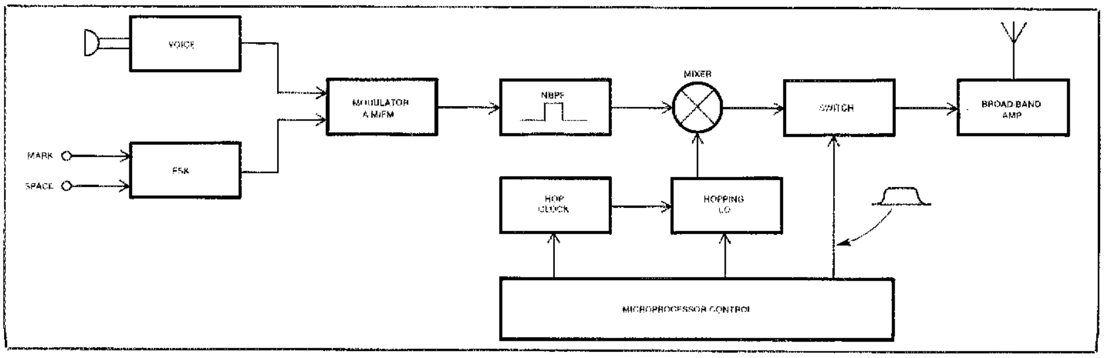


Fig. 10 — A frequency-hop transmitter. The hopping LO must be capable of making fast frequency changes with short settling times.

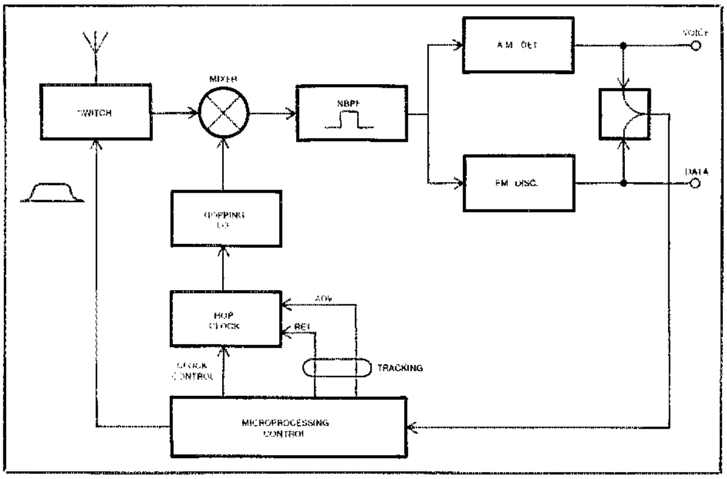


Fig. 12 — A frequency-hop receiver. The antenna switching reduces intermodulation from strong adjacent-frequency signals.

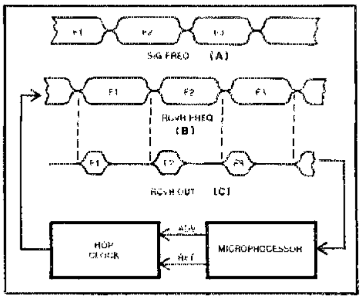


Fig. 13 — Illustrating frequency-hop sync searching or tracking.

mark or space condition. For data reception, each filter has a 3-dB bandwidth that in hertz is about 1.25 times the number of hops per second. Linear-phase filters, called "matched" filters, are needed.

Frequency-Hop Synchronization

One way to synchronize is to use very stable hop-clock oscillators. The hopping code patterns are then all initialized at some time, say each hour, using WWV as a time reference. Then, only a slight amount of searching back and forth in time is needed to align the receiver with the signal.

At the start of each reception time, a small sync adjustment is made. A block of data would be preceded by a special segment that sets up the receiver to copy data. A latecomer in a net would also need opportunities to get fully synchronized.

Sync searching is illustrated in Fig. 13. We see that the incoming signal, at A, does not completely coincide in time with the receiver tuning, at B. The signal switches to frequency f2 before the receiver is ready to switch to f2. The result is that the receiver output has a signal only during the interval shown at Fig. 13C. The receiver uses this information to advance or retard the hop clock slightly until the overlap has improved sufficiently. In the example shown, the receiver hop clock would be speeded up.

In another synchronization method that is somewhat more complicated, the receiver slow hops until the computer recognizes that sync has occurred. Then, the receiver fast hops in sync with the signal. The advantages of this method are that very stable hop-clock oscillators are not needed, and no prealignment of hopping codes is needed.

An additional method is to reserve certain frequencies as "start" or "stop" frequencies. When the start frequency is detected, hopping automatically begins according to the hop pattern plan. A stop frequency advises that the net control is available.

Interference Rejection

In Fig. 14, two kinds of interference are shown. In one case, a steady signal appears on one of the frequencies. In the second case, another hopping signal occasionally occupies the same frequency at the same time. Also shown in Fig. 14 is a common occurrence, signal fading from time to time on various channels.

Ways to combat these conditions are (1) make the receiver bandwidth narrow, to reduce interference; (2) design the hop codes to minimize "collisions" between nets and between net members; (3) use a fast-responding agc; (4) use a lot of message redundancy, i.e., repeat everything on several frequencies; (5) use error-correction codes; and (6) delete occupied frequencies and insert clear frequencies. Possibly certain frequencies would be reserved for backup use only.

Network Protocol

How could frequency hop be used to enhance an amateur packet network? Consider the format in Fig. 15.

Block 1: The net control has a hop code that addresses the net. The stations in the net are using this hop code to monitor the net control station and synchronize to it.

Blocks 2 and 3: The net control also sends data that contains the information in blocks 2 and 3.

Block 4: Having acquired frequency-hop sync, the net members acknowledge.

Block 5: The net control listens for my message using my frequency-hop code. Other stations in the net communicate with each other at the same time, using predetermined hop patterns that do not interfere with other members.

Block 6: The stations use two kinds of code simultaneously: frequency-hop pattern and mark/space code. This makes it possible, for example, to address a particular station and to identify the caller at the same time, or to inform the recipient how to respond to the call. Once the net members become synchronized, it should be possible to maintain sync for a long period of time, with tracking adjustments as required. If a member requires sync, he can send a sync-request message to which the recipient responds.

These ideas are offered for illustrative purposes and do not represent any known system in use. The important thing is that

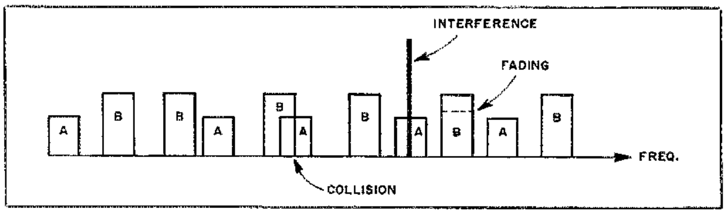


Fig. 14 — This drawing depicts interfering signals and fading in a frequency-hop system. The letters A and B identify emissions from two different transmitters sharing the spectrum, but with different hopping codes. The two frequency sets are almost orthogonal. The coding system is designed to reject interference, fading and collision. The same hop frequencies may be used by both stations, A and B, but at different times.

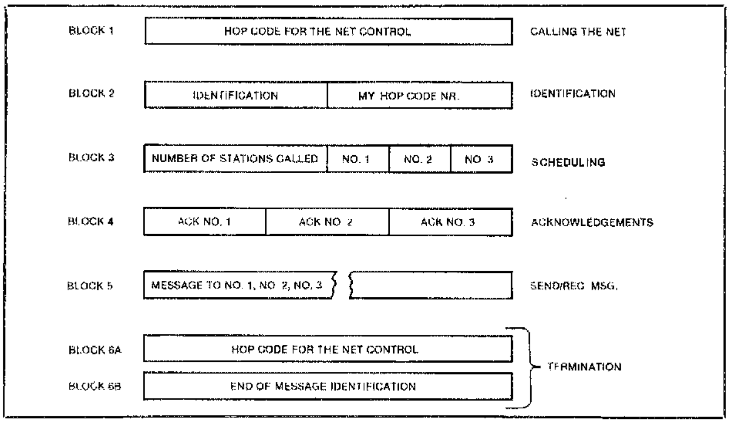


Fig. 15 — Possibilities for network protocol.

amateurs are free to devise schemes that are right for them.

The key to success in frequency hop is to repeat the message often and to provide error correction. This means the message rate must be reduced to improve reliability. In frequency hop, the possibilities of interference are strong. This means that any part of the message must be repeated on several different frequencies. In a slow-hop system, a block of data is sent on each frequency and then the frequency is changed. A typical dwell time on each frequency might be 0.1 second.

It is important to have some way of detecting "bad" frequencies and moving away from them. The network protocol should include an avoidance strategy. The key is to find out if excessive errors occur on certain frequencies. This frequency management is under computer control. My computer tells your computer what action is needed to improve the message reliability.

In a fast-hop system, the dwell time on each frequency might be 1 ms. The message rate is the same as with slow hop, but now each data bit is repeated on several different frequencies. After each bit is received, a vote is taken and the majority decides whether the bit is a 1 or 0. The addition of error correction adds to the reliability. The fast-hop method should be a better scheme for amateur use because, with good design of hopping code patterns, collisions and loss of data should be less.

The Amateur Radio Research and Development Corporation (AMRAD) is a group of amateurs who are dedicated to advanced technology in Amateur Radio. Within this group are subgroups interested in spread spectrum and packet networks. The *AMRAD Newletter* is helpful to anyone wanting to learn more, or to get in touch with others having similar interests. (Another interesting newsletter, *QEX*, an experimenter's exchange, is published by the ARRL.) AMRAD is also in touch with the FCC and has obtained special permits to do various kinds of experimental work. The address is AMRAD, P.O. Drawer 6148, McLean, VA 22106. Terry Fox, WB4JFI, is president of AMRAD. Hal Feinstein, WB3KDU, heads the spread-spectrum subgroup, and Dave Borden, K8MMO, heads the packet-protocol group.

New Books

MICROPROCESSORS AND MICROCOMPUTERS

by Ronald J. Tocci and Lester P. Laskowski. Published by Prentice-Hall, Inc., Englewood Cliffs, NJ 07632. Second edition, 1982. Hardbound, 7-1/4 × 9-1/2 inches, 404 pages including index. $20.95.

This comprehensive book does exactly what the authors intend it to do. It addresses a broad spectrum of readers, providing a practical introduction to the world of microprocessors and microcomputers. Concentrating on the fundamentals of microprocessor-based systems, it leans specifically toward the 6502.

The authors' writing style is about the nearest you can come to having a personal tutor without actually employing one. If you can't find the time (or money) to attend classes on this subject, you can bring the class into your home for a small cash outlay. And should you attend a formal class, you'll still find this book to be a valuable reference.

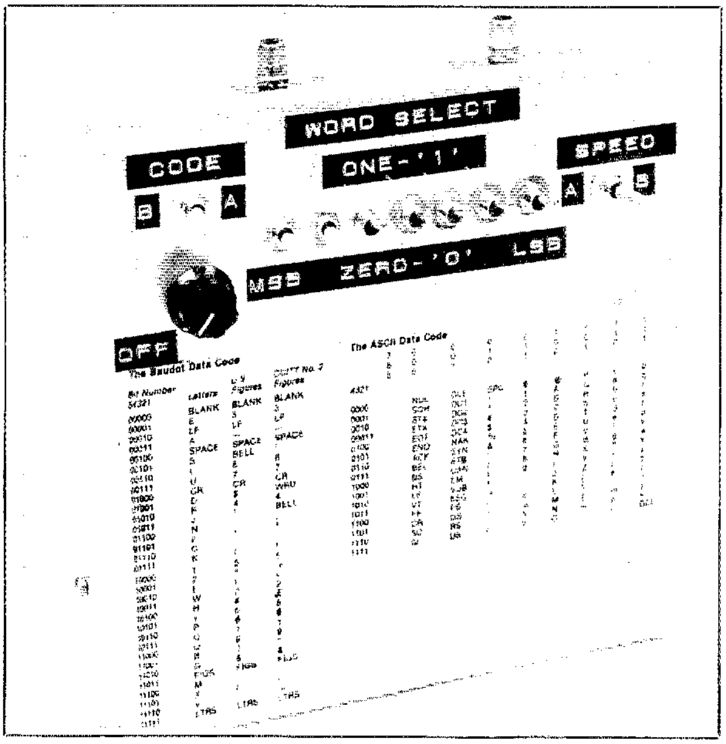
From number systems and codes used in the world of computers, you're stepped through digital circuits, an introduction to computers and then into microcomputer hardware. After voyaging through the inner workings of the microprocessor itself, you're led to the ports of I and O — input/output interfacing. These are important ports of call, as this interfacing allows you to communicate with the computer, and the computer to respond to you. The many facets of microcomputer software are covered in the last section, and the appendix contains the 6502 instruction set and op codes. Budding assembly-language programmers will find some helpful examples.

Each chapter has a number of questions, plus problems for the reader to solve. This helps you to discover gaps in your understanding of the material. More than likely you're not going to remember *everything* you've read the first time around, but it's a simple matter to turn back a few pages and refresh your memory — sort of a dynamic RAM operation!

Weak points? My only gripe is that there is no answer section. It would be nice to have absolute verification of the answers you've provided for the example questions and problems.

If by now you haven't gathered that I recommend the book, let me state emphatically that I do! I wish all my schoolbook texts had been written this clearly. This is one volume that shouldn't collect dust on your bookshelf. — *Paul K. Pagel, N1FB*

A Serial ASCII/Baudot Character Generator You Can Build



Looking for a simple way to test your printer or video terminal? This "weekend" project may be just what you need.

By J. Robert Witmer,* W3RW

Wouldn't it be nice to be able to *easily* generate serial Baudot or ASCII characters for test purposes? I constructed a simple, low-cost device to do just that. It provides Baudot or ASCII characters at the most frequently used speeds, has a variable character rate and can be battery powered. Easy-to-add modifications include 8-bit ASCII operation, selectable parity generation and additional speeds. The unit is based on a UART (Universal Asynchronous Receiver Transmitter), an LSI (large-scale integration) IC widely used in data communications applications.

Brief UART Theory of Operation

At first glance the UART appears complicated, but it's really very easy to use. Several articles have described its operation in detail, so we'll cover just enough of the theory to understand how to put the UART to work.[1,2] A UART is a circuit subsystem (many functions in one package) that can receive and transmit binary information of varying word lengths. In doing this, it incorporates the start, stop, control and error-detecting information necessary for asynchronous communications with many standard data terminals and computers.

This may be easier to understand by first looking at the transmission time sequence of the Baudot (Murray) and the ASCII character codes (Figs. 1 and 2). Circuitry within the UART allows you to concentrate on the binary code for the Baudot or ASCII character and to forget about the generation of the start, stop and optional parity bits. You don't have to worry about the timing, the spacing or the sequencing, either, as the UART handles all of that, too! Most UARTs have control pins that are used to select the format and size of the Baudot or the ASCII word (or any 4- to 7-bit code) you wish to use. A description of the function of each UART pin is given in Table 1.

How the Character Generator Works

The schematic diagram of the character generator is shown in Fig. 3. The heart of the circuit is the UART, U2. S2 through S8 control the UART data inputs and are used to enter the binary code for the Baudot (see Table 2) or ASCII (see Table 3) character you wish to send. S9 enables you to send 5-bit (Baudot) or 7-bit (ASCII) characters. If you wish to be able to send 8-bit ASCII characters, a switch can be added to connect UART pin 38 to the positive supply (8-bit) or to ground (5- and 7-bit). I hard

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[1]Notes appear on page 23.

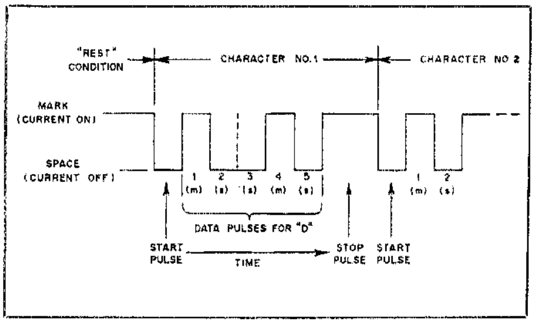


Fig. 1 — Time sequence of a typical Baudot character, the letter D.

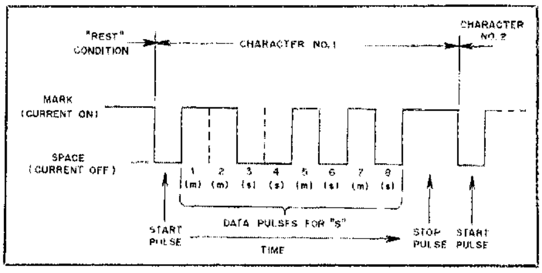
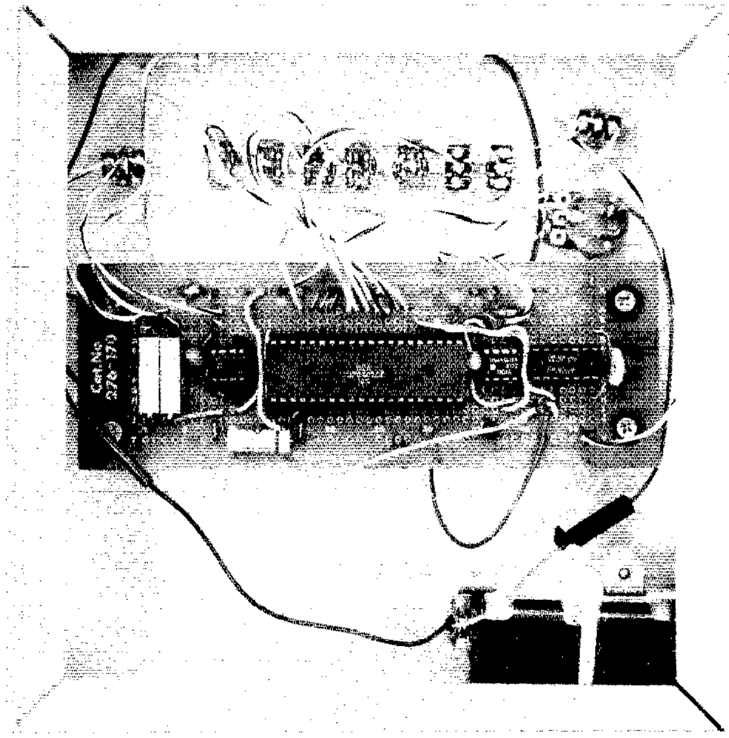


Fig. 2 — Time sequence of a typical ASCII character, the letter S. The eighth, or parity, bit (when used) may be set to one of four possible conditions: (1) always mark, (2) always space, (3) odd parity or (4) even parity.

Table 1
UART Pin Descriptions

Pin no.	*Function*
1	Positive supply: Usually +5 V (4 to 6 V used in this circuit).
2	Some UARTs require −12 V on this pin (not used).
3	Ground.
4	Receive data output enable: Parallel data outputs are enabled when this pin is low (not used).
5-12	Receive data outputs (not used).
13-16, 18, 19	Receive status outputs (not used).
17	Receive clock input (not used).
20	Receive serial input (not used).
21	Reset: Operational status of the UART is reset when this pin is pulsed high.
22, 24	Transmit status outputs (not used).
23	Transmit strobe: Data transmission is initiated when this pin is pulsed high.
25	Transmit serial output.
26-33	Transmit parallel inputs.
34	Control strobe: Control inputs are entered into the UART when this pin is high.
35	Parity enable: Parity generation is enabled when this pin is low.
36	Number of stop bits: Two stop bits are generated when this pin is high. One stop bit is generated when this pin is low and the select-data word length is greater than 5 bits. One and one-half stop bits are generated when this pin is high and the data word length is 5 bits.
37, 38	Data word length: This pin selects the number of data bits per word: *pin 37 / pin 38 / bits/word* 0 / 0 / 5 0 / 1 / 6 1 / 0 / 7 1 / 1 / 8
39	Parity select: If parity is enabled (pin 35), this pin is used to select even or odd parity.
40	Transmit clock input.



A look at the interior of the ASCII/Baudot character generator. The CHARACTER RATE control is above and to the right-hand side of the circuit board. At the bottom right-hand side of the chassis is a NiCd battery pack used to power the unit.

wired pin 38 to ground for 5- and 7-bit-only operation.

A similar situation exists for pin 35: If you wish to be able to send parity-encoded ASCII characters, a switch can be connected to this pin to allow parity to be selected. I chose to disable parity by hard wiring pin 35 to the positive supply. If you do elect to utilize parity, another switch should be connected to UART pin 39 to enable you to select even or odd parity.

An R-C circuit (R2 and C2) connected to the UART RESET input (pin 21), ground and the positive supply provides a reset pulse to the UART when power is first applied. U3 provides the data-strobe signal to the UART. Each strobe pulse causes the selected binary code to be loaded into the UART and transmitted. By varying the data-strobe pulse rate with R1, you can adjust the rate at which characters are sent. R1 does *not* affect the words-per-minute speed or baud rate of the individual characters, but simply controls the time interval between characters. The circuit shown in Fig. 3 provides a rate of approximately 1/2 to 60 characters per second. You could eliminate U3 and use a momentary-contact push-button switch to

Table 2
The Baudot Data Code

Bit Number 54321	*Letters*	*U.S. Figures*	*CCITT No. 2 Figures*
00000	BLANK	BLANK	BLANK
00001	E	3	3
00010	LF	LF	LF
00011	A	—	—
00100	SPACE	SPACE	SPACE
00101	S	BELL	'
00110	I	8	8
00111	U	7	7
01000	CR	CR	CR
01001	D	$	WRU
01010	R	4	4
01011	J	'	BELL
01100	N	,	,
01101	F	!	!
01110	C	:	:
01111	K	(	(
10000	T	5	5
10001	Z	"	+
10010	L	)	)
10011	W	2	2
10100	H	#	£
10101	Y	6	6
10110	P	Ø	Ø
10111	Q	1	1
11000	O	9	9
11001	B	?	?
11010	G	&	&
11011	FIGS	FIGS	FIGS
11100	M	.	,
11101	X	/	/
11110	V	;	=
11111	LTRS	LTRS	LTRS

Note: FIGS-H (10100) may also be used for MOTOR STOP function. "1" = mark = hole in punched tape

Table 3
The ASCII Data Code

Bit Number 7	0	0	0	0	1	1	1	1
6	0	0	1	1	0	0	1	1
5	0	1	0	1	0	1	0	1
4321								
0000	NUL	DLE	SPC	Ø	@	P	\	p
0001	SOH	DC1	!	1	A	Q	a	q
0010	STX	DC2	"	2	B	R	b	r
0011	ETX	DC3	#	3	C	S	c	s
0100	EOT	DC4	$	4	D	T	d	t
0101	ENQ	NAK	%	5	E	U	e	u
0110	ACK	SYN	&	6	F	V	f	v
0111	BEL	ETB	'	7	G	W	g	w
1000	BS	CAN	(	8	H	X	h	x
1001	HT	EM	)	9	I	Y	i	y
1010	LF	SUB	*	:	J	Z	j	z
1011	VT	ESC	+	;	K	[	k	{
1100	FF	FS	,	<	L	\	l	/
1101	CR	GS	-	=	M	]	m	}
1110	SO	RS	.	>	N	^	n	~
1111	SI	US	/	?	O	—	o	DEL

ACK = acknowledge
BEL = signal bell
BS = backspace (←)
CAN = cancel
CR = carriage return
DC1 = device control 1
DC2 = device control 2
DC3 = device control 3
DC4 = device control 4
DEL = (delete)
DLE = data link escape
ENQ = enquiry (WRU)
EM = end of medium
EOT = end of trans.
ESC = escape
ETB = end of block
ETX = end of text
FF = form feed (home)
FS = file separator
GS = group separator
HT = horizontal tab (→)
LF = line feed (↓)
NAK = not acknowledge
NUL = null
RS = record separator
SI = shift in
SO = shift out
SOH = start of heading
SPC = space
STX = start of text
SUB = substitute
SYN = synchronous idle
US = unit separator
VT = vertical tab (↑)

Note: "1" = mark = hole in punched tape

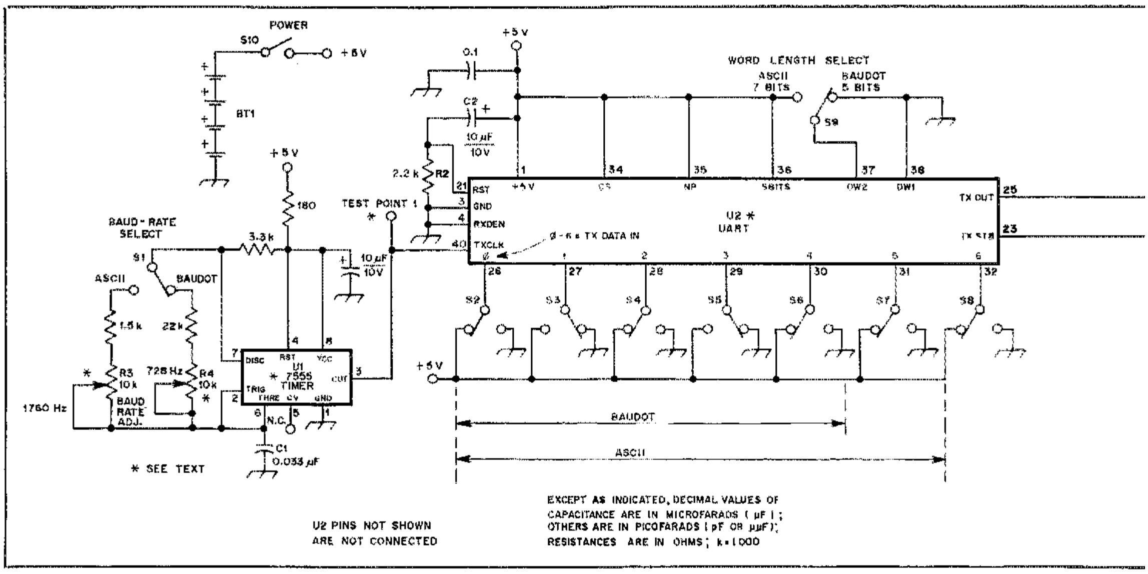


Fig. 3 — Character generator schematic diagram. All resistors are 1/4-W, 5%-tolerance carbon types. Except as indicated, capacitors are disc ceramics. Polarized capacitors are electrolytic or tantalum types.

BT1 — Four AA-size NiCd or alkaline cells in series.
C1 — 0.033-µF Mylar or polystyrene capacitor.
R1 — 100-kΩ panel-mount control.
R3, R4 — 10-kΩ pc-mount multi-turn control.
S1-S9, incl. — Spdt toggle switch.

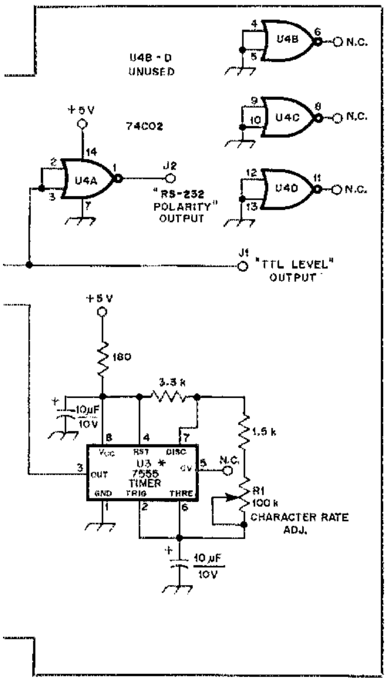
Table 4
Part Comparison

	CMOS	Standard
U2	CDP6402 IM6402	AY-3-1015 AMI S 1602
U1, U3	7555	555
U4	74C02	74LS02
Current drain	3.5 mA at 5 V	60 mA at 5 V
Approximate total IC cost	$13	$9

pulse pin 23 high when you want a character to be transmitted. Unused sections of U4 could be used to debounce the switch.

U1 provides the transmit-clock signal to the UART (pin 40). It is the frequency of this signal that determines the words-per-minute speed or baud rate at which each character is sent. S1 allows you to select between two preset operating speeds. I have one set for 60-wpm Baudot operation and the other set for 110-baud ASCII.

Four "AA" NiCd or alkaline batteries provide the power to operate my unit. The total current drain with the parts I use is less than 4 mA. The current requirements and voltage ratings of various component alternatives are given in Table 4. Once you've selected the parts, choose the battery type and capacity necessary for the length of service you desire. I've operated my unit at supply voltages from 4 to 6 V without trouble. Be sure that the circuit being driven can handle the maximum voltage of the battery pack you plan to use. Be careful not to exceed the maximum voltage ratings of the parts you use.



S10 — Spst toggle switch.
U1-U4, incl. — See text and Table 4.

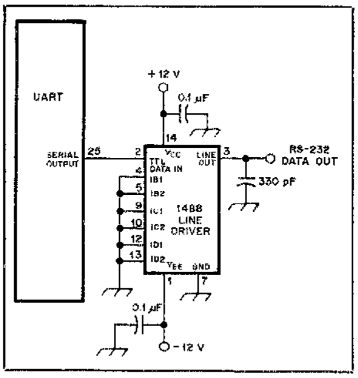


Fig. 4 — RS-232 interface circuit.

Construction and Alignment

The construction technique you use when building the character generator is not critical. I used a Radio Shack experimenter's board (cat. no. 276-170) and provided sockets for all ICs. I mounted the board in a 7 × 7 × 2-inch aluminum chassis with all the switches and controls mounted along the top edge.[3] This arrangement permitted sufficient room to attach copies of the ASCII and Baudot code tables below the switches for easy reference.

I am presently using the low-power CMOS parts listed in Table 4. The UART is an RCA CDP6402. The IM6402, available from JDR Microdevices, should be an exact replacement for the RCA part.[4] The 74CO2 is also available from JDR Microdevices, and the ICM7555 CMOS timers are available from Radio Shack. I also tried non-CMOS parts in the circuit, with no change in performance except for a large increase in battery drain and a higher supply-line noise level. The non-CMOS UART I tested was the AMI S1602, which is an exact replacement for the GI AY-3-1015D (available from Radio Shack). A Mylar or polystyrene capacitor should be used for C1 to ensure frequency stability under varying temperature conditions.

Alignment is simple: Connect a frequency counter to the output of U1 (TEST POINT 1, Fig. 3) and adjust the trimmers (R3 and R4) for a clock frequency equal to 16 times the desired baud rate. For 60-wpm Baudot (45.45 baud) and 110-baud ASCII, the frequencies are 728 Hz and 1760 Hz, respectively. If you wish to operate 100-wpm Baudot (74 baud) or 300-baud ASCII, the frequencies are 1187 Hz and 4800 Hz.

Using the Character Generator

I use the output of U2 (J1) to drive (through an appropriate bias resistor) the keying transistor in my Model 15 printer loop system. The output from U4A (J2) will drive the RS-232 input of my ADM3A CRT terminal. Although this output does not meet the RS-232 standard specifications, I have used it successfully with several RS-232 printers and terminals.[5] The circuit shown in Fig. 4 can be used to ensure full compliance with the RS-232 standard if you desire. Notice, however, the need for additional power-supply voltages. You could probably use 9-V batteries to supply these voltages, but I would try single-supply operation first.

To operate the character generator, you simply connect your equipment to J1 for transistor keying or to J2 for "RS-232 polarity" (mark = 0 V). Set the switches to the desired mode and character code, turn on the power and adjust the character rate by means of R1. That's all there is to it! You now have a simple way to test your Baudot or ASCII terminal or printer.

Notes

[1] J. A. Titus, "The UAR/T and How it Works," *Ham Radio*, Feb. 1976, p. 58.
[2] S. Leibson, "Input/Output Primer, Part 4: The BCD and Serial Interfaces," *BYTE*, May 1982, p. 202.
[3] mm = in. × 25.4.
[4] JDR Microdevices, Inc., 1224 S. Bascom Ave., San Jose, CA 95128.
[5] G. Woodward, ed., *The Radio Amateur's Handbook*, 60th ed. (Newington: ARRL, 1982), p. 14-46.

References

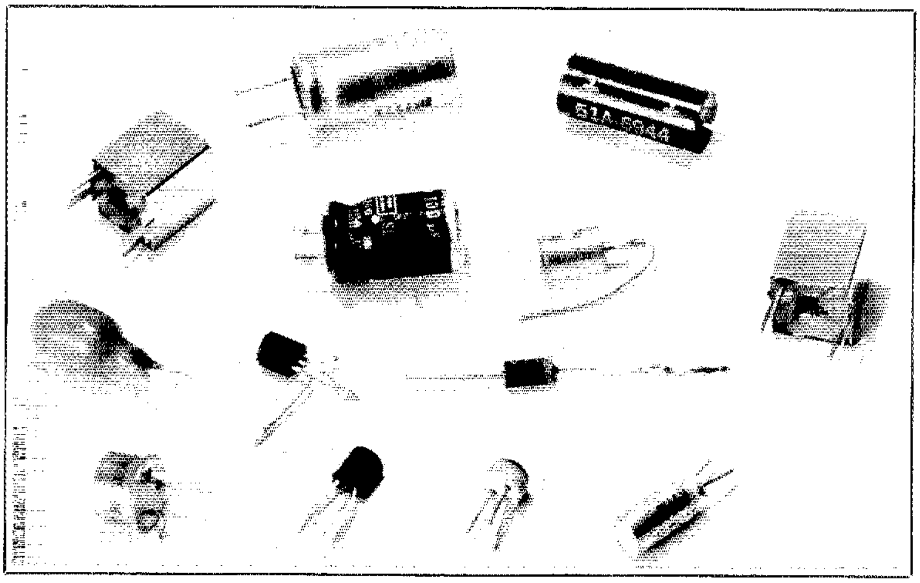
Barden, W. Jr. "Using the Model I/III RS-232 Port." *BYTE*, July 1982.
Derynck, R. R. "Bit Rate Clocks for Your Serial Interface." *Microcomputing*, Oct. 1979.
Kalin, E. B. "A Programmable Serial-Communication Interface." *QST*, Sept. 1982.
Kirchner, E. "Serial Converter for 8-Level Teleprinters." *Ham Radio*, Aug. 1977.

Next Month in QST

It's being called the "Amateur Radio Event of the Decade" — Astronaut Owen Garriott's upcoming operation aboard the Space Shuttle *Columbia*. How can you take part? Check out August *QST*, which will provide details on when and where to listen, and how to have the best chance to complete a QSO with W5LFL while he conducts the first Amateur Radio activity from space.

On the more mundane side, the August issue will feature several top-notch technical articles. One will describe a means of approaching fast-scan TV quality using slow-scan methods. Another, aimed at beginners, will delve into rf power measurements — including how to use an rf ammeter, an rf probe and a scope. Finally, "A Boom-Excited Antenna" will tell you how to put an existing Yagi on 10 MHz and below by gamma-matching the boom.

Salvaged Parts: A Gold Mine for the Radio Amateur



Parts procurement a problem? If you're a builder, you're a "scrounger." Here's a map to another treasure trove of parts and a sample of what can be done with salvaged parts.

By Clint Bowman,* W9GLW

After my retirement in 1976, I engaged in part-time work that involved me with a-m/fm digital clock-radio circuits. This work had two results. It triggered a desire to put together some gear and get back into Amateur Radio activities using current technology. (From 1931 to 1945 I had done a considerable amount of experimenting, building and constructing on the 160- through the old 2½-meter bands.) I also realized that understanding the intricacies and eccentricities of the clock-radio circuits could be extremely useful to amateurs whether they build anything or not. Also, there is a gold mine of salvageable parts in defunct clock-radios. These parts may be retrieved easily using a solder sucker or wicking material and a 25- to 45-W soldering iron. Bypass and electrolytic capacitors, transistors, diodes, fixed resistors and some useful miniature a-m/fm i-f transformers are there for the taking.

Discovering the I-F Transformers

The i-f transformers are housed in bright metal solderable containers coded with paint dots on the container or on the end of the adjustable core. Typically, those with yellow or black dots are used for a-m i-f circuits at a nominal frequency of 455 kHz. The red-dot units provide inductance for the a-m local oscillator/mixer tuning circuit at 455 kHz, and have an output coupling loop. The ferrite-core material employed is suitable for use at frequencies from somewhat less than 455 kHz to at least 2500 kHz.

Transformers with orange, green, blue or lavender dots are used in 10.7-MHz fm circuits. The green and blue transformers have collector-matching taps and are therefore preferred for general amplifier service. These coils will generally be satisfactory in applications requiring tuned circuits at frequencies from about 14 MHz to less than 7 MHz, depending on the amount of lumped circuit capacitance present. These coils may also have an integral fixed capacitor across the input winding that should be removed to increase the upper resonance range. This capacitor may be located within the shield can or externally just below the plastic base.

The external capacitor can be removed easily by breaking it with long-nosed pliers. The shield will have to be removed to accomplish the job if the capacitor is mounted internally. For frequencies below 10.7 MHz, this capacitor may be left as is and additional capacitance can be added to resonate the inductor at a frequency as low as 7 MHz without an excessively dominant C to L ratio. For lower frequencies, it is

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desirable to rewind the transformer. This can be done with a little patience.

Rewinding the Transformers

A supply of no. 40 enameled copper wire should be on hand. Be certain it is of the self-stripping variety that strips the insulation from the wire during soldering. A small vise, good illumination, a low-wattage soldering iron with a clean tip, a penknife and long-nosed pliers should also be available.

Usually, the shield can is crimped to the plastic base. This crimping can be undone by passing a knife blade around the base several times between the shield and the base. Then, one at a time, grasp each corner solder pin with your pliers and, while holding the shield can with your other hand, gently work the can loose. Inside the can you will find a threaded plastic insert with a threaded ferrite cup core. Up and down movement of the cup changes the flux density of the coil and the coil inductance. The unique design of the cup core completely shields the coil from outside influences.

Secure the coil base in a small vise by catching opposite edges in the vise jaws. The coil is wound on a dumbbell-shaped ferrite core and will probably be surrounded by a thin plastic protective membrane that should be removed with the aid of a knife blade. The core is usually held in place with wax in a seat provided in the plastic base. The coil also will be wax-impregnated and will have to be heated briefly with the soldering iron to remove the existing wire turns. During this procedure, the core probably will loosen from the base. It can be reseated by reheating the assembly with the tip of the iron. Or, the bottom end of the core and the mounting space on the base can be cleaned of wax and a drop of epoxy cement used to make a solid assembly.

Coil winding is best accomplished with the core in place. Leave enough slack in the wire terminations to the solder pins so these leads may be dressed closely to the winding area, because the cup core must clear these fragile leads. A 3:1 winding ratio between input and output windings will work for usual common-emitter amplifiers. The same ratio applies for determining the input tap for the driving-stage collector. Ensure that certain windings are polarized properly. Make a drawing of the pin arrangement showing the winding connections, because after the shield can is replaced this information may prove invaluable!

The ferrite material used in the 10.7-MHz transformers is probably useful up to at least 150 MHz. The cup core arrangement limits use of practical coils to something less than 40 MHz, however, because of the high flux density and correspondingly few turns of wire. At least one of these dumbbell-core coils is found in the

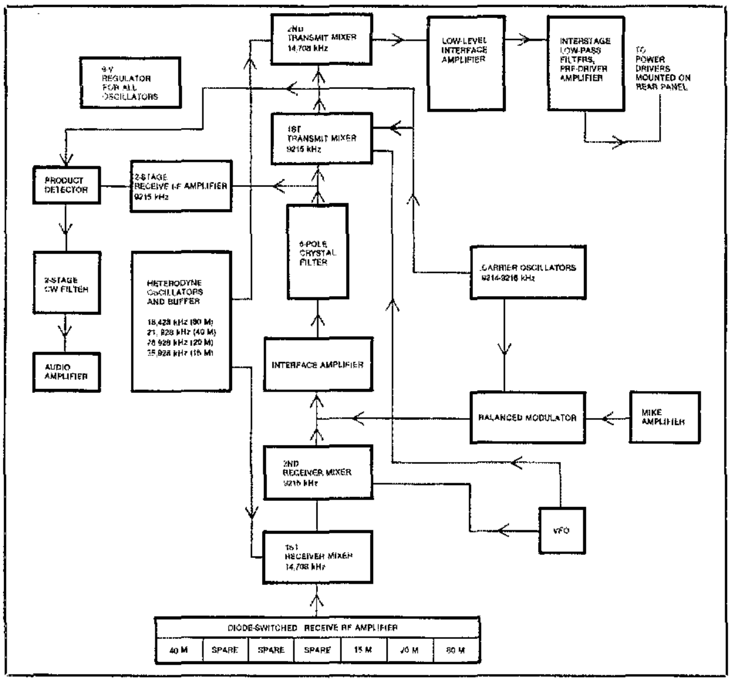
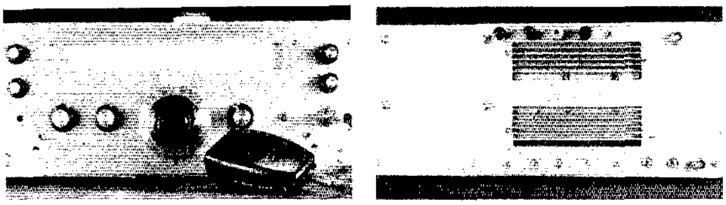


Fig. 1 — A block diagram of the author's homemade transceiver.



Front and rear views of the homemade transceiver. *(photos by Russ Planck, W9RGH)*

emitter circuit of the high-frequency, common-base fm mixer/oscillator of these clock-radios as a tuned choke for a dc return.

These ferrite-core transformers exhibit a lower Q than air-wound coils. For use as rf/i-f stage transformers for the bands from 160 through 20 meters, however, they exhibit flatter loaded-Q characteristics and far better shielding.

Other Components

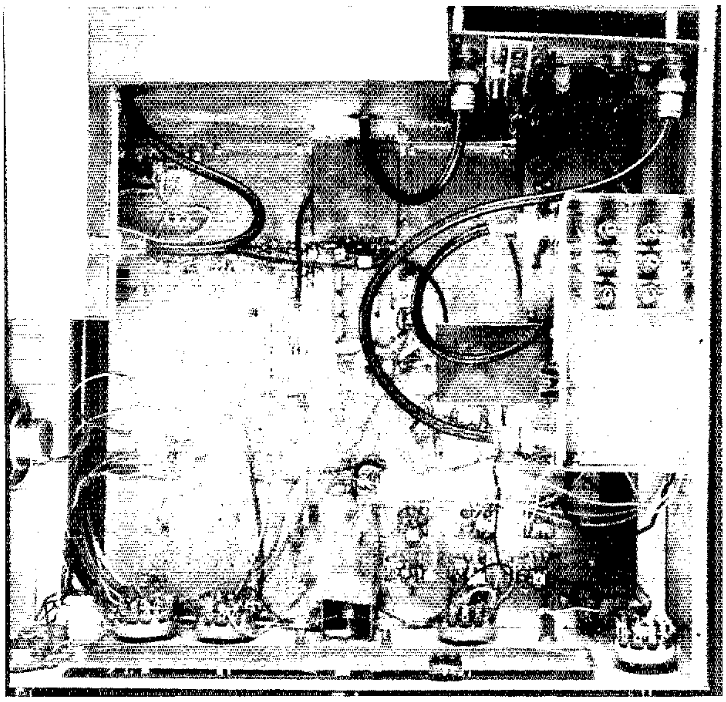
The front-end rf transistors in these sets will oscillate and amplify beyond 150 MHz. They are suitable for noncritical circuits at 2 meters, but leave something to be desired as low-noise, high-gain amplifiers at the higher frequencies. These devices are quite suitable for 9-MHz i-f service, as are transistors found in the radio i-f circuits.

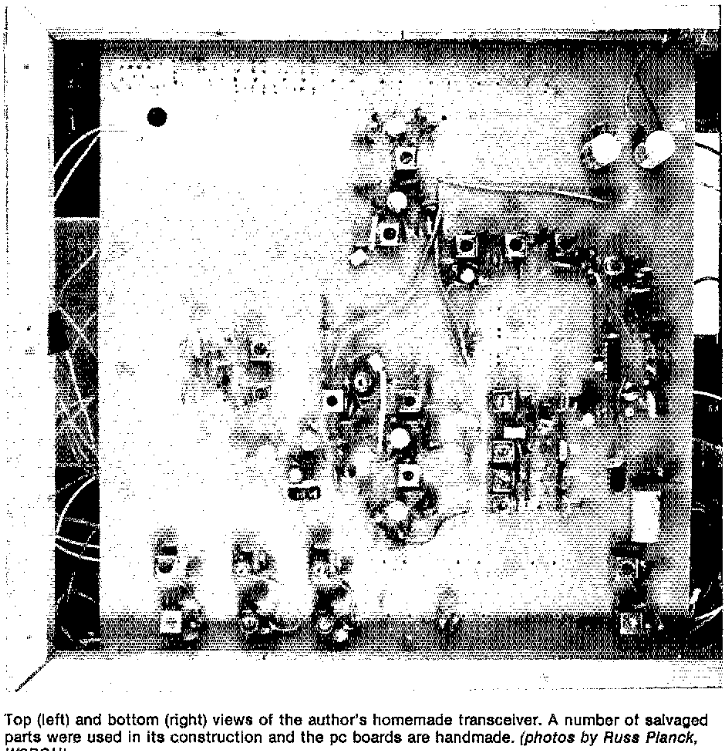
Generally, high-frequency germanium diodes of the 1N60 variety can be retrieved from these sets. One such diode is located in the fm mixer collector circuit, and others in the ratio detector. These diodes exhibit low-threshold, low-capacitance characteristics and are suitable for use in rf probes at frequencies up to at least 1000 MHz.

Other devices include npn and pnp transistors suitable for use in low-frequency service, rectifiers of the 1N4000 type and electrolytic capacitors ranging in value from 1 to 1000 μF usually with 16-V dc ratings. Fixed resistors with values ranging from a fraction of an ohm to perhaps 1 megohm are abundant.

A Transceiver is Born

As mentioned earlier, the part-time work triggered a desire to put together some gear





Top (left) and bottom (right) views of the author's homemade transceiver. A number of salvaged parts were used in its construction and the pc boards are handmade. *(photos by Russ Planck, W9RGH)*

and get back into Amateur Radio activities using current-technology equipment. A transceiver with multiband switching capability and enough power to permit operating during prime times was an absolute requirement. It had to be all solid-state (with a minimum of direct, mechanical switching), be self-contained (with the exception of the power supply), incorporate an antenna-matching network and SWR indicator, and be definitely more than a novelty since it would have to serve as the only hf equipment at W9GLW.

Several goals were established: (1) tuning must be consistent from band to band and progress from left to right (using a slide-rule dial), with the lower-frequency end on the left; (2) all tuned circuits must present negligible levels of unwanted signals, especially in the mixer; (3) remote-diode or transistor switching would be employed for all low-level, multiple-frequency or multiple-function circuits. A block diagram of the transceiver is shown in Fig. 1.

A regulated power supply provides a 12-V dc source for all low-level transmitter and receiver circuits, a 20-A high-current source adjustable from 6- to 13-V dc for the power stage drivers and final amplifier, and a source of 117-V ac. The latter is switched from a hand-held microphone through a low-current dc circuit and is used to operate an external coaxial antenna T-R switch.

I selected a pair of Motorola MRF454 transistors for the final amplifier. These transistors have a 250-W dissipation rating and a maximum continuous collector current rating of 20 A. At 13-V dc, this amplifier functions satisfactorily at better than 200-W cw input without forced-air cooling. Bias is set for about 800 mA of collector current for satisfactory linear operation.

The power amplifier driver is a two-stage affair consisting of an input stage using push-pull MRF237 transistors and an output pair of MRF475s. This amplifier operates at a lower power level, supplying from 1 to 5 W of driving power. Both amplifiers are broadband types, and no tune-up is necessary.

The high-current power-supply output is fed directly to the driver and final amplifiers continuously, even during receiving periods. A 30-A meter provides for current monitoring.

Driver and final-amplifier modules are mounted vertically on the rear panel of the transceiver. This panel may be removed for servicing without unsoldering a connection. I found it imperative that the final amplifier be completely shielded; otherwise, massive regeneration resulted from coupling between the output and the transmitter second mixer circuits. Little shielding is required for the driver module. Cables equipped with BNC connectors are used for interconnections. Separate antenna connections for transmit and receive are used, and an external coaxial switch allows for antenna switching.

The consistent tuning goal dictated a double-conversion scheme for transmit and receive. Cascaded balanced mixers satisfy this requirement. The interstage transformers were salvaged and modified as outlined previously, and no interstage amplification is required.

To a great extent, the antenna interface circuit treatment determines the overall effectiveness of any receiving system. This transceiver uses a single rf amplifier stage with diode-switched, individually tuned input and output transformers for all bands.

The two-stage i-f amplifier is patterned after circuits found in a-m/fm radio sets, and the cw filter and audio amplifier are similar to circuits found in *QST* and semiconductor manufacturers' applications notes. A three-position panel switch selects cw sharp, cw medium and ssb bandwidths. The transistors and transformers used are salvaged items. Occasionally, an LM386 will be found in imported radios. I used one for the audio amplifier. A small salvaged speaker provides adequate audio for communications-quality sound.

The inclusion of a receiver agc circuit is, in the final analysis, one of personal preference. An agc loop is included in this circuit. It is without dc amplification, excludes the rf amplifier, and is inoperative except on extremely strong signals. Separate rf and af gain controls are provided. This arrangement is entirely satisfactory for digging out weak DX signals and working through full-power neighboring-station signals. An S meter wasn't feasible with this circuit, so that device was omitted...and hasn't been missed!

The VFO uses inductive tuning and has a nominal range of 5700 to 5200 kHz, providing for 500-kHz-wide bands. The rather unusual oscillator end-points were dictated as a convenience to the manufacturer of the carrier oscillator and filter crystals; they enabled the use of on-hand blanks. John Holmbeck, W9KZO, was of help in this area.

The heterodyning oscillator circuits are diode-switched. A set of steering diodes, energized by the microphone PTT switch or cw setup switch, diverts the buffered oscillator signal to the transmit or receive mixer. All oscillators are supplied with 9-V dc from an inboard regulator and filter system. This approach isolates these critical circuits for emergency operation from an unknown power source. Rear-panel jacks are included to permit operation from a single power source.

Seven-band operation has been provided. At the present, only the 80, 40, and 20-meter bands have been activated for transceive, and 15 meters for receive only. Seven low-pass transmitter output filters are mounted in a shield box on the left side panel of the transceiver. Filter selection is made by means of a two-section rotary switch. A relay, activated by the microphone PTT switch or the cw setup switch, controls distribution of low-current supplies to both transmitter and receiver circuits.

The cabinet is made of individual pieces of aluminum for the sides, front and back panels, and top and bottom covers. A length of U-shaped 1-inch-wide aluminum countertop stock was formed into a rectangle.[1] To this is attached the main circuit board, the four cabinet sides and the bottom cover. The top cover attaches to a lip formed along the top edge of each side panel. Top and bottom covers are removable for transceiver alignment or examination purposes without affecting the mechanical integrity of the vertical members of the assembly.

PC-Board Construction

All pc boards are handmade, including the 12-inch-square main circuit board. I use a high-speed rotary hobby tool and dental burr.[2] Component-mounting holes are drilled using a drill press. I have used this method of circuit-board construction for many years with excellent results.

Should you wish to try this method of pc-board construction, here's an outline of the procedure from circuit conception to finished board: (1) a schematic diagram is drawn; (2) rough artwork is sketched disregarding dimensions, but placing components and electrical runs to eliminate as much as possible the need for jumpers; (3) final artwork is drawn to scale on 0.1-inch-ruled graph paper; (4) the artwork is taped to the copper-clad board and all component and mounting holes are drilled; (5) the artwork is removed and all connecting runs and other features are drawn free hand on the board using the drilled holes as guides; (6) the board is then screwed to 1-inch lumber stock for ease of handling, and only the lines encompassing the electrical runs and pads are routed. Most of the copper remains on the board for ground plane use or the later addition of other circuits; (7) the board is then sanded to a smooth finish, and all circuits are checked with an ohmmeter for shorts and continuity.

Complicated single- and double-sided boards filled with 6- to 40-pin ICs have been prepared in this manner. To the uninitiated, they look like etched boards with more copper remaining than usual. After you've had a little practice, an audio amplifier or oscillator project can progress from an idea to a finished board in a couple of hours.

Try your hand at a salvaged-parts project. You'll save money and gain some valuable experience!

Notes

[1]mm = in. × 25.4

[2]Router burrs supplied by the tool manufacturer have never been found to be suitable for this service; the shanks are too long, and control of the tool is lost. In addition, the burrs lose sharpness quickly and must be replaced often. Your friendly dentist might give you a package of used burrs or tell you where to purchase them. Dental burrs have short shanks and retain their sharpness for a long time.

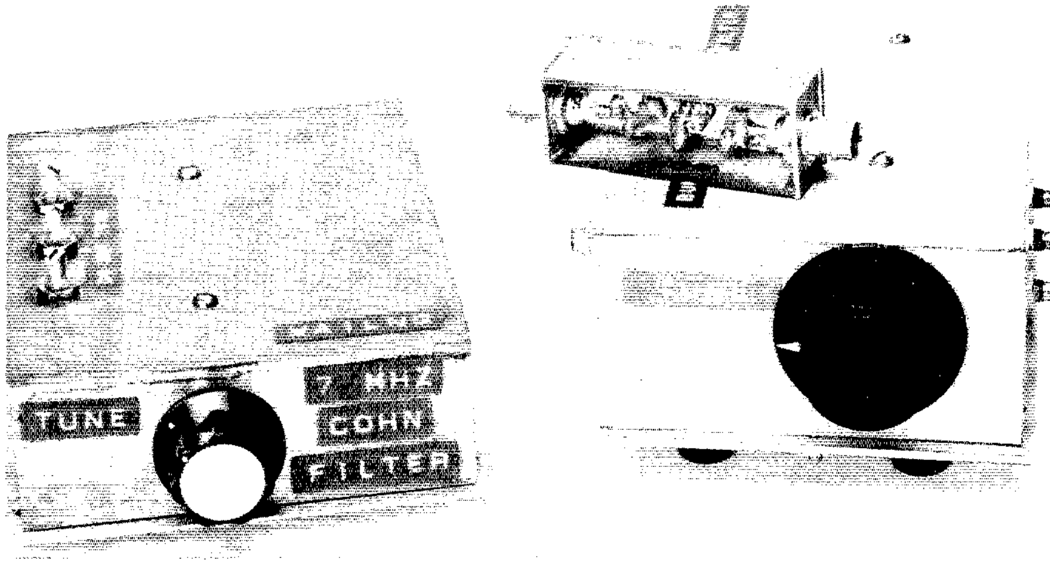
Strays

NARL, ARRL SIGN PUBLIC-SERVICE PACT

□ In April, officials of the Newington Amateur Radio League (NARL) and the American Radio Relay League put their signatures on an agreement that formalized the club's assistance to local police, particularly during emergencies. In a cermony at W1AW attended by the local news media, NARL President Tom Namnoum, KM1O, and W1AW trustee John Lindholm, W1XX, signed a Memorandum of Understanding between the club and W1AW/R. Under the terms of the agreement, NARL will provide operators to the Newington Police Department for public service events and during emergencies. NARL members will also give information over W1AW to visitors requesting directions to ARRL Hq. ARRL is responsible for supplying and maintaining the equipment.

With the signing of the pact, the issuing of photo identification and the completion of mock disaster drills, NARL is now the official backup communications group to the Newington Police Department. Police Chief Thomas Ganley said that NARL members and their equipment will be put to use during major disasters, such as floods and blizzards, and if the Police Department's communication system should fail. The amateurs will also help police during parades and on holidays. The NARL volunteers have been incorporated in the town's civil preparedness division to further formalize the assistance program.

Filter Systems for Multi-transmitter Amateur Stations



Have you ever operated from a station that has had two or more transmitters on simultaneously? The intrastation interference can drive you crazy! Here's how to cure the "multi-multi" syndrome.

By Gerald B. Hull,* AK4L/VE1CER

Modern hf receivers are designed to tolerate very large input signals; after all, the bands *are* crowded with QRO signals. But there are situations in which receivers will "fall apart" when they are most needed. Multi-multi contest stations, emergency communications centers and Field Day sites are prime examples. In each case, fundamental and harmonic energy from close-by transmitters can overload a receiver front end, causing intermodulation distortion (IMD) and other spurious products to appear at the output.

Just how bad can this problem be? To illustrate by way of a mathematical example: Imagine a transmitter delivering 600 W (+58 dBm) at 3505 kHz to a dipole antenna.[1] The second harmonic is 40 dB below the fundamental, or +18 dBm. If we neglect antenna system loss and/or gain, the free-space attenuation between two antennas placed 200 feet apart would be 25 dB on 7 MHz.[2] This means a 7-MHz receiver would be subjected to a signal of −7 dBm (+58 − 40 dB − 25 dB = −7 dBm), or 100,000 μV, at its input! Signal levels of this magnitude can raise havoc within a receiver.

Another cause for intrastation interference, although not as prevalent, is rf leakage into the ac power lines. This can be caused by improper bypassing at the equipment supply, or leakage from cabinet openings.

The Solution

A sure-fire cure to the above problems is simple — external filter networks. Stand-alone low/high-pass filters can reduce harmonic energy to a near-zero level. Fixed-frequency or tunable band-pass filters can be used as narrow preselectors to reduce "crud" in a receiver front end. "Brute-force" ac-line filters can reduce rf energy on power lines drastically.

Which filter should you use? Each type has its particular advantages and disadvantages, and each application must be chosen

*Assistant Technical Editor

[1]Notes appear on page 31.

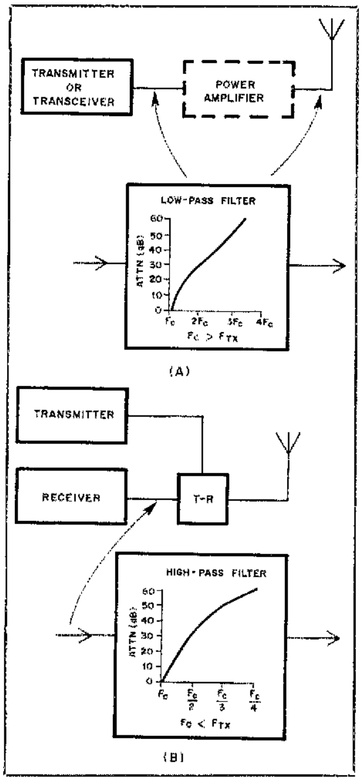


Fig. 1 — Block diagram of a high/low-pass filter system. When an external power amplifier is used, low-pass units should be located at the *output* of the amplifier.

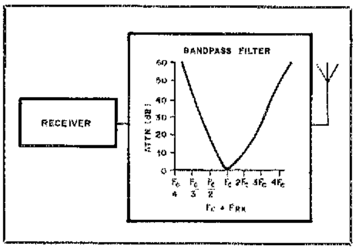


Fig. 2 — Band-pass filter system diagram. When these filters are used with a transceiver, the equipment must be modified to prevent transmitter energy from entering the filter.

Table 1

Chebyshev High-Pass and Low-Pass Filter-Attenuation (dB)

No. Poles, Ripple	*VSWR*	*$2f_c$*	*$3f_c$*	*$4f_c$*	*$5f_c$*	*$6f_c$*	*$7f_c$*
3 pole, 0.1 dB	1.36	12.24	23.60	31.42	37.39	42.22	46.29
5 pole, 0.1 dB	1.36	34.85	54.21	67.27	77.21	85.26	92.04
7 pole, 0.1 dB	1.36	57.72	84.83	103.11	117.03	128.31	137.80
9 pole, 0.1 dB	1.36	80.60	115.45	138.96	156.86	171.36	183.55

Note: For high-pass filter configuration $2f_c$ becomes $f_c/2$, etc.

Table 2

Chebyshev Low-Pass Filter — Pi Configuration

No. Poles, Ripple	*C1*	*C2*	*C3*	*C4*	*C5*	*L1*	*L2*	*L3*	*L4*
3 pole, 0.1 dB	3283.6	3283.6				9.131			
5 pole, 0.1 dB	3650.4	6286.6	3650.4			10.91	10.91		
7 pole, 0.1 dB	3759.8	6673.9	6673.9	3759.8		11.32	12.52	11.32	
9 pole, 0.1 dB	3805.9	6794.5	7019.9	6794.5	3805.9	11.48	12.87	12.87	11.48

Component values normalized to 1 MHz and 50 ohms. L in μH; and C in pF.

Chebyshev Low-Pass Filter — T Configuration

No. Poles, Ripple	*L1*	*L2*	*L3*	*L4*	*L5*	*C1*	*C2*	*C3*	*C4*
3 pole, 0.1 dB	8.209	8.209				3652.3			
5 pole, 0.1 dB	9.126	15.72	9.126			4364.7	4364.7		
7 pole, 0.1 dB	9.400	16.68	16.68	9.400		4528.9	5008.3	4528.9	
9 pole, 0.1 dB	9.515	16.99	17.55	16.99	9.515	4591.9	5146.2	5146.2	4591.9

Component values normalized to 1 MHz and 50 ohms. L in μH; C in pF.

carefully. An improper choice can result in minimal interference reduction or an unnecessary expense.

Low-Pass/High-Pass

External low-pass filters can be used when transmitter harmonic signals are causing receiver overload. Simple 3- to 7-pole filters can reduce harmonics to a noninterference level. The principal disadvantage of discrete low-pass filters is the requirement for filter components that will withstand high power levels. High-voltage capacitors for a 1-kW filter are expensive and almost impossible to locate. Low-pass filters are installed at the transmitter or final-amplifier output (see Fig. 1).

An alternative to the reduction of transmitter harmonics, although not as effective, is the addition of a quarter-wave coaxial stub at the transmitter output. The coaxial line should be 1/4-λ at the *harmonic* frequency. One end of the line is open-circuited and the other end connects to the transmitter output through a coaxial "T" connector.

A high-pass filter mounted at the receiver input will prevent overload from strong transmitter fundamental signals. Since relative signal levels at the receiver input are low, standard-size components may be used.

Fixed/Tuned Band-pass

Band-pass filters are useful for eliminating interfering signals that may be present in a receiver front-end passband but outside the i-f system passband (see Fig. 2). Typical fixed-tuned band-pass filters for 1.8 to 30 MHz have bandwidths of 50 to 200 kHz. This type of filter can be used to attenuate energy that falls outside the filter passband, e.g., to lower the effective level of broadcast carriers present in the upper portion of the 7-MHz band. The disadvantage of fixed-tuned band-pass filters is the need for several units to cover a single band. Tunable bottom-coupled, minimum-loss band-pass filters have been described in amateur literature.[3,4,5] Commonly called "Cohn filters," these units have the advantage of tuning over an entire band and, depending on the design center frequency, *very* narrow bandwidths. Cohn filters work well up to 14 MHz, after which the bandwidth and loss increase dramatically. The alignment of Cohn filters is tricky, but despite this fact they remain one of the most popular external interference-rejection filters.

AC Line

When interference exists between stations that have proper input/output filtering, there is a good chance that rf energy is being coupled into the receiver via the ac power line. To confirm this, transmit into a dummy load and monitor the second-station receiver. If the interference persists, the path is probably the ac line. A "brute-force" ac line filter should clear up this type of interference.

Construction Data

Tables 1, 2, 3 and 4 are used in conjunction with Figs. 3 and 4 to design each of the above mentioned filters for a specific frequency or band. The high-pass and low-pass filter component values in Tables 1 and 2 are normalized to 1 MHz.[6,7] To find a component value for the frequency of interest, divide the normalized value by the frequency in megahertz. For example, to calculate values for a low-pass filter that has a cut-off frequency of 8 MHz, you would divide the table values by 8. A practical 7-pole low-pass filter is shown in Fig.

Table 3

Band-pass Filter Component Values

BW_L(3 dB)	L1,L2 (μH)	C_o (pF)	C1,C5 (pF)	C3 (pF)	BW_L(3 dB)	L1,L2 (μH)	C_o (pF)	C1,C5 (pF)	C3 (pF)
1.8-1.9 MHz	5.12	1446	291	55.3	7.0-7.2 MHz	2.05	245	42.4	4.88
1.8-1.85 MHz	5.12	1485	192	28.8	7.0-7.3 MHz	2.05	242	53.2	7.17
1.8-1.9 MHz	8.74	847	221	32.4	7.0-7.2 MHz	8.74	57.5	19.1	1.14
1.8-1.85 MHz	8.74	870	148	16.8	14.0-14.2 MHz	0.8	158.7	14.4	1.59
3.5-3.6 MHz	5.12	393	75.1	7.8	14.0-14.2 MHz	2.08	61	9.2	0.61
3.5-3.6 MHz	8.74	230	56.6	4.58	14.0-14.4 MHz	2.08	61	14.7	1.2
3.5-3.7 MHz	5.12	352	110	15	21.0-21.3 MHz	0.49	115	9.2	1.15
3.5-3.7 MHz	8.74	224	83.4	8.8	21.0-21.5 MHz	0.49	114	14.1	1.9
3.8-4.0 MHz	5.12	325	93.2	11.8	21.0-21.5 MHz	0.82	68.4	11.4	1.14
3.8-4.0 MHz	8.74	191	70.7	6.9	28.0-28.5 MHz	0.48	65.3	7.2	0.82
7.0-7.1 MHz	2.05	248	26.6	2.5	28.0-29.0 MHz	0.48	64	11.8	1.6

$C2 = C_o - C1 - C3$
$C4 = C_o - C3 - C5$

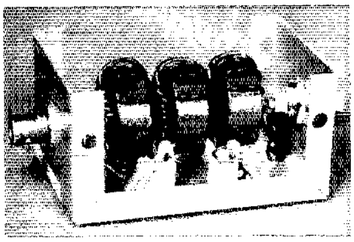


Fig. 5 — A practical 7-pole low-pass filter built around the information given in Table 2. The unit is housed in a small aluminum box.

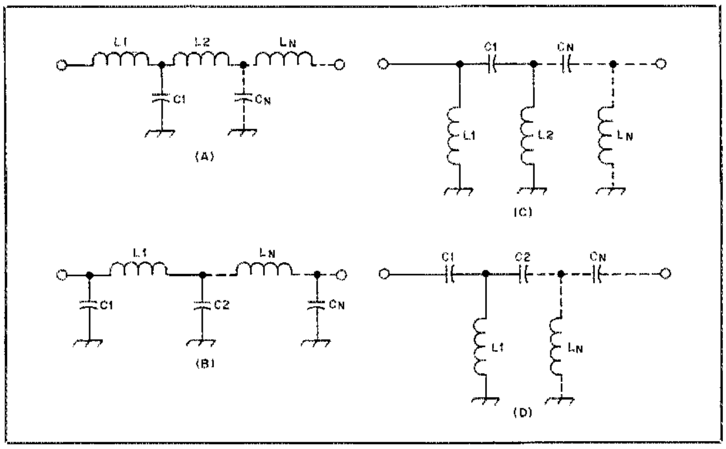


Fig. 3 — Schematic diagrams for low- and high-pass networks. The choice of a "T" or "pi" configuration will depend on the components on hand.

5. Toroid cores are used for the inductors. They are self-shielding, which will decrease the internal coupling in the filter, improving the ultimate attenuation. The filter is mounted in a small aluminum box to prevent leakage.

Table 3 shows the component values for a fixed-pole band-pass filter for 1.8 through 30 MHz. The construction techniques described in the previous section can be used for the band-pass design.

An important factor to consider when constructing band-pass filters is insertion loss. This loss is directly proportional to the Q of the inductors used. For optimum performance, always choose the highest-Q core material for the frequency range desired, keeping in mind the trade-off between highest Q (maximum selectivity) and insertion loss.

Variable Passband

Tunable Cohn filters are perhaps the most complex to construct and align. The schematic diagram of a three-resonator bottom-coupled Cohn filter is shown in

Table 4

Cohn Filter Parts Values

Band	L1	L2*	L3	L4	L5*	L6	C2	C3	C4**	C5	C6	C7	C8	CX†	CY†	CZ†
80	29	0.8	60	60	0.8	29	150	33	2-14	2-14	2-14	33	150	5	33	5
40	14.5	0.38	29	29	0.38	14.5	68	18	1-20	1-20	1-20	18	68	—	15	—
20	2.37	50 nH	3.0	3.0	50 nH	2.37	240	20	1-20	1-20	1-20	20	240	—	—	—

Inductance values are in μH; capacitance is in pF unless otherwise noted.
On 80 and 40 meters, L1, L3, L4 and L5 are wound on 2-mix ferrite cores (μ = 10, $Q_u \gtrsim 250$).
On 20, 6-mix cores are used (μ = 8, $Q_u \gtrsim 250$).
*L2 and L5 are air core inductors, except on 40 meters, where a T37-2 core is used.
**C4, C5 and C6 2-14 pF units are Johnson 193-6-1; others are Cardwell 160-107. See note 4 for availability.
†May not be required.

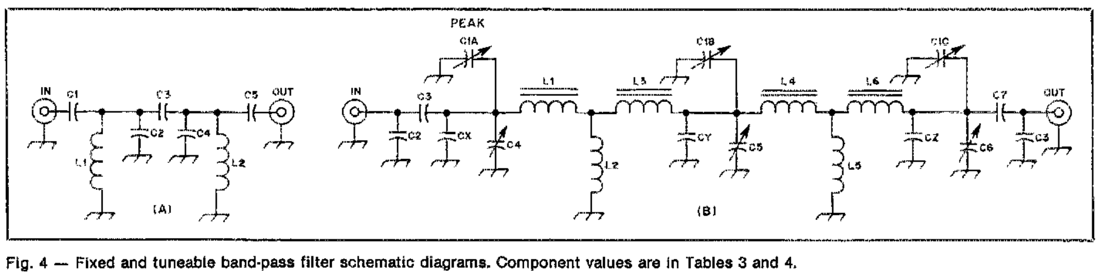


Fig. 4 — Fixed and tuneable band-pass filter schematic diagrams. Component values are in Tables 3 and 4.

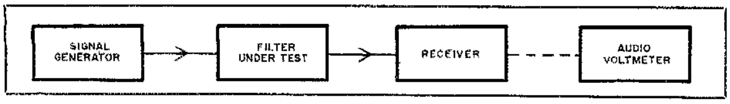


Fig. 6 — Cohn filter alignment setup. See text for details.

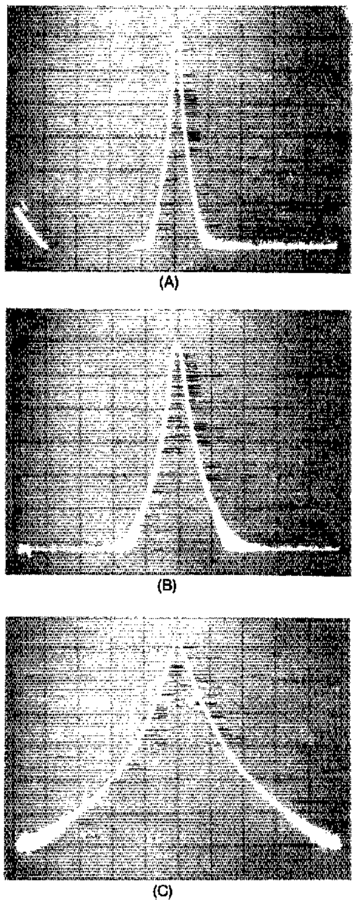


Fig. 7 — Response curves of three Cohn filters for 3.5 (A), 7 (B) and 14 (C) MHz. Top of the picture represents 0-dB loss. Vertical divisions are each 10 dB; horizontal divisions are each 500 kHz.

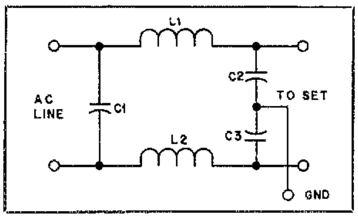


Fig. 8 — "Brute-force" low-pass ac line filter schematic diagram. All capacitors are 0.005-μF disc-ceramic units. L1 and L2 are 15 μH, with wire size depending on current demand.

Fig. 4B. The parts values for 3.5- through 14-MHz filters are given in Table 4. Inductance values in the table must be adhered to *strictly*, or the filter will be impossible to align! The three-gang tuning capacitor, C1, is a Miller 1460 (23 pF per section), which is available from Radio Kit.[8] Substitutes for this capacitor can be used, but be sure the three gangs track properly and that the capacitance value for each section is the same.

Two pieces of test equipment will be required to align the filter — an hf receiver and a signal generator, both of which cover the filter frequency. Hook up the test gear and filter as shown in Fig. 6. The receiver S meter or an external audio voltmeter can be used to measure the receiver output. Set the generator and receiver to the filter center frequency. Apply a signal to the filter and adjust C1 for maximum output. Next, adjust C4, C5 and C6 for maximum output level. These three capacitors will interact, so they will have to be adjusted and readjusted to ensure minimum insertion loss. L4 and L5 determine the mutual coupling between filter sections. The coil turns may have to be compressed or expanded to find the coupling that will give a single-hump response. Over-coupling will cause a three-hump response, and under-coupling will severely restrict the bandwidth and increase insertion loss.

Fig. 7 shows the response curves of three Cohn filters built by Jay Rusgrove, W1VD. The loss in these filters varies from 7 to 10 dB. By carefully adjusting the coupling inductors and trimmer capacitors, these filters can be adjusted for minimum loss or minimum bandwidth, whichever is more important to the builder.

AC Line

A "brute-force" ac line filter is shown in Fig. 8. The inductors should be air core, and wound with enamel-coated wire of sufficient cross-sectional area to handle the current of the equipment involved. The unit should be enclosed in a metal box with a ground lug connected directly to an earth ground.

Conclusions

These filters should enable the multioperator-oriented amateur to design a station in which intrastation interference does not exist. These filter methods are not meant to take the place of proper station engineering, i.e., proper placement of antennas, use of high-dynamic-range receivers, and so on.

I have operated both multi-single and multi-multi contest stations that incorporate these filters, and have met with great success. Thanks to ARRL Hq. Lab Supervisor Phil Accardi, AJ1N, for his assistance in testing and confirming the Cohn filter designs.

Notes

[1]The unit "dBm" is used to express power in reference to 1 mW, or 0 dBm. A level of +58 dBm is 58 dB above this reference.
[2]m = ft × 0.3048.
[3]D. DeMaw, "His Eminence — the Receiver, Part 1" *QST*, June 1976.
[4]W. Sabin, "Solid-State Receivers," *QST*, July 1970.
[5]Cohn, "Dissipation Loss in Coupled Resonator Filters," *Proc. IRE*, Aug. 1959, pp. 1342-1348.
[6]G. Woodward, ed., *The Radio Amateur's Handbook* (Newington: ARRL, 1982), pp. 6-11 and 6-12.
[7]D. DeMaw, ed., *ARRL Electronics Data Book* (Newington: ARRL, 1976), pp. 55-56.
[8]Trimmers and three-gang tuning capacitors are available from Radio Kit, Box 411, Greenville, NH 03048.

Strays

QRP MOVEMENT IN EUROPE

☐ Enthusiasm for QRP (low power) in Great Britain and other parts of the world has heightened considerably in recent years. According to a U.S. spokesman for the American QRP movement, some 40,000 amateurs around the world are now actively engaged in QRP activities in the high-frequency spectrum. This seems due in part to the appeal of building inexpensive, simple homemade projects. Another reason for the keen interest in low-power operation is the challenge offered by using less than 10 W of transmitter output power to span the globe with cw and ssb signals.

The Rev. George Dobbs, G3RJV, who founded the G QRP Club in the UK some time ago, publishes *SPRAT*, the club's quarterly QRP bulletin. He reported on April 1, 1983 that the club has gained more than 450 new members since Christmas of 1982. This remarkable surge in affiliation is indicative of the interest being shown in QRP activities. Membership in the G QRP Club is available to any amateur, irrespective of nationality. The club also offers an interesting QRP type of booklet (100 pages) called *Circuit Handbook*. RSGB will soon reprint and distribute this publication for the G QRP Club. Those interested in the club may contact G3RJV, 17 Aspen Dr., Chelmsley Wood, Birmingham B37 7QX, England. — *Doug DeMaw, W1FB*

I would like to get in touch with...

☐ anyone with information or a schematic diagram for an amplifier using an 833A or 8122 tube. Nick Ferro, KU2A, Box 167, Lake Placid, NY 12946.

A Simple Computer Model for VHF/UHF Propagation

Understanding propagation basics will help you predict vhf/uhf coverage. A BASIC program makes the calculations easy.

By Jack Priedigkeit,* W6ZGN

Wouldn't you like to be able to predict how far beyond the horizon your vhf or uhf signals will travel? By using the simple propagation model presented in this article, you can estimate terrain clearance and line-of-sight (los) coverage from, for example, a repeater site. You'll also learn how to account for the effect of varying meteorological conditions. To ease the job of making the necessary calculations, a simple BASIC computer program has been included.

Basic Concepts

The idea that the velocity of propagation along an rf transmission line depends on the dielectric constant of the material separating the conductors is familiar to most amateurs. However, it is not generally appreciated that this concept also applies to the bending of a radio wave as it propagates through the atmosphere. The velocity of propagation depends on the refractive index — the square root of the dielectric constant. The dielectric constant of air is a function of pressure, temperature and humidity and it generally decreases with altitude. It is this decrease with altitude that is important.

A radio signal at a slightly higher altitude, $h + \Delta h$, travels just a little faster than the same signal traveling at the reference altitude, h. The effect is to tilt the radio wave front ever so slightly, so as to bend the ray path toward the earth. This is illustrated in Fig. 1. If a curve with a radius greater than the radius of the earth is used to draw the ray path, it can be seen that the refracted, or bent, ray will intercept the earth beyond the geometric tangent point.

The distance to the geometric horizon (the tangent point for the straight ray path) is calculated easily because the tangent to a circle is perpendicular to the radius of the circle. This distance is

$$d(ft) = \sqrt{(a + h)^2 - a^2} \qquad \text{(Eq. 1)}$$

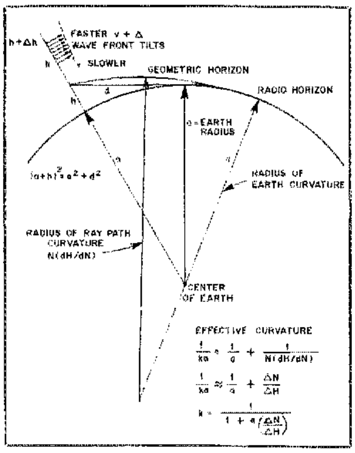


Fig. 1 — The geometry used to calculate the geometric and the radio horizons.

where
a = radius of the earth in feet (3690 mi × 5280 ft/mi)
h = antenna height in feet[1]

Neglecting the h^2 term, which is small compared with 2ah, the distance in miles to the geometric horizon is very closely approximated by the expression

$$d(mi) = \sqrt{1.5h} \qquad \text{(Eq. 2)}$$

where h = antenna height in feet.

Meteorological Considerations

The typical long-term average refractive index of the air at the surface of the earth is 1.000300. This normally decreases with altitude at an average rate of 12 parts per million per 1000 feet of elevation (39 parts per million per km). It is the custom to describe the refractive index in terms of N-units. The N-unit being defined as (1 − the refractive index) × 10^6. With this definition, the surface refractive index is 300 N-units, and the gradient, $\Delta N/\Delta H$, is −12 N-units per 1000 feet, or −39 N-units per km.[2,3,4]

The effective earth-radius factor, k, which determines the distance to the radio horizon, is related to the refractive-index gradient by the expression

$$k = \frac{1}{1 + a(\Delta N/\Delta H)} \qquad \text{(Eq. 3)}$$

where a = the true earth radius

$\Delta N/\Delta H$ = the refractive index gradient in the same units as the earth radius (i.e., both in feet or in meters).

The distance to the radio horizon in miles is

$$d(mi) = \sqrt{1.5\,kh} \qquad \text{(Eq. 4)}$$

where h = antenna height in feet.

The refractive-index gradient is determined by the changes in air pressure, temperature and humidity with altitude along the propagation path. Temperature inversions and sudden changes in humidity, as occur at the interface between fog and clear air, can cause rapid changes, or discontinuities, in the rate of change of N with altitude. Since it is difficult, if not impossible, to measure the detailed micrometeorology of the propagation path, it is necessary to use an average k value for the path. Over a period of many years, the average value of k is 4/3. This results in the familiar expression

$$d(mi) = \sqrt{2h} \qquad \text{(Eq. 5)}$$

where h = antenna height in feet.

The propagation model to be described allows the user to specify a value for k. The computer uses this value to calculate the

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[1]Notes appear on page 33.

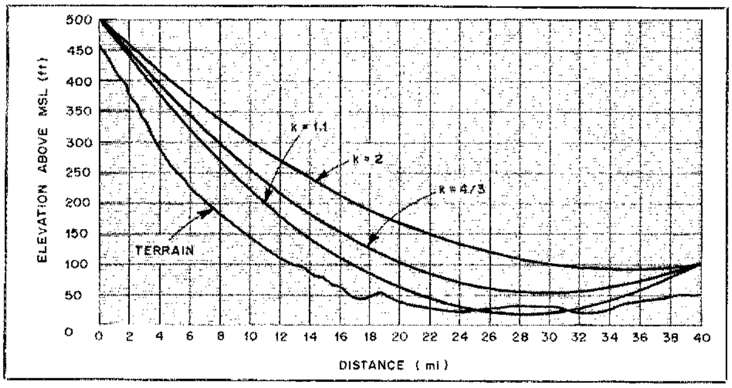


Fig. 2 — Shown here are the ray-path-versus-distance plots for the example described in the text. Three values of k are used to illustrate the effect of varying atmospheric conditions.

height of the ray path above a smooth earth as a function of the distance between the two antennas. Different meteorological conditions can be simulated by using different values of k.

Computer Results

Fig. 2 is a plot of the computer-calculated ray-path height between two antennas, 40 miles apart, located at 500 and 100 feet above mean sea level. This might be a propagation path between a repeater site and a fixed station. A terrain profile was sketched in using elevation data taken from a topographic map. From Fig. 2 it can be seen that under average propagation conditions (k = 4/3) the ray path clears the terrain by 50 feet. For meteorological conditions where k = 1.1, however, the ray path is blocked by the terrain, and a reduction in signal level would be expected. It can also be seen that when conditions are such that k = 2 the ray path clears the terrain by 100 feet. These curves illustrate nicely the effect of short-term meteorological conditions on this propagation path.

For other combinations of antenna height, separation distance and terrain, the ray path may be blocked for the average (k = 4/3) conditions. In this case, the path will open up when meteorological conditions increase the effective earth radius. This partly explains the extended repeater coverage and the interference between repeaters sometimes observed.

The Computer Program

The vhf/uhf ray-path computer program (Table 1) is written in BASIC and should run on most home computers with little or no modification. For simplicity, extensive operator directions and input prompts have been omitted from the program. These features can be added, if desired, by anyone familiar with BASIC programming.

The operator enters, by means of the keyboard, the altitude of the antennas, H1 and H2, in feet above mean sea level, the separation distance, D0, in miles, and the earth radius factor, K0. The computer output is a listing of the distance in miles from the antenna at H1 and the height of the ray path in feet above mean sea level.

Table 1
Program Listing

```
10 INPUT H1
20 INPUT H2
30 INPUT D0
40 INPUT K0
50 DH=H1-H2
60 K1=1.5*K0
70 K2=(D0^2)/K1
80 IF ABS(DH)>K2 THEN 170
90 D1=0.75*K0*DH/D0+D0/2
100 H0=H1-D1^2/K1
110 IF H0<0 THEN PRINT "LOS NOT POSSIBLE"
120 FOR N=0 TO 20:  D3=N*D0/20
130 H3=H0+(D3-D1)^2/K1
140 PRINT D3, H3
150 NEXT N
160 END
170 K3=DH/D0+D0/K1
180 FOR N=0 TO 20:  D3=N*D0/20
190 H3=H1-K3*D3+D3^2/K1
200 PRINT D3, H3
210 NEXT N
220 END
```

If the antenna heights are too low for the separation distance, the ray path will be blocked by the curvature of the earth and the computer will print "LOS NOT POSSIBLE." The computer will then print negative altitudes where the ray path falls below mean sea level. This is useful information, as it can be used to estimate the increase in antenna height required for a line-of-sight path.

The ray-path height is calculated at 20 equally spaced points along the propagation path. Twenty points were chosen because it is a convenient number of lines to display on the computer terminal. By changing the number 20 in program lines 120 and 180, the number of points and lines printed can be changed.

The ray-path height calculations use either of two programs: lines 120 through 140, or lines 180 through 200. If the absolute difference in the antenna height, ABS(DH), is less than the difference in height needed for line of sight at the separation distance, D0, the path height is calculated by lines 120 and 130. In this case, it is necessary to first calculate the distance, D1, to the point of minimum path height, H0. This is done in lines 90 and 100. If the absolute difference in antenna height is greater than that needed for line of sight, the ray path is calculated by lines 180 and 190. In this case, the distance D1 and the height H0 are not used, as the minimum height will be either H1 or H2. However, it is now necessary to include the height change caused by the slope of a straight line connecting H1 and H2. Line 170 calculates the slope, and line 190 calculates the ray-path height with distance, including the height changes resulting from the curvature of the earth.

Final Comments

No mention has been made of Fresnel zone clearance criteria used in selecting sites for microwave relay stations where highly reliable paths are required to minimize circuit outage.[5] These have been omitted because the midpath first Fresnel zone clearance for a 40-mile path at 146 MHz requires a 600-foot terrain clearance. Since there are few, if any, amateur repeater sites that can meet this criterion for the majority of the users, first Fresnel zone clearance is, in this case, academic. We amateurs are conditioned to accept varying signal strengths and an occasional path outage. Since Fresnel zone clearance decreases as the square root of the signal frequency, in the gigahertz frequency range it is considerably less than the 600 feet required for 2-meter operation.

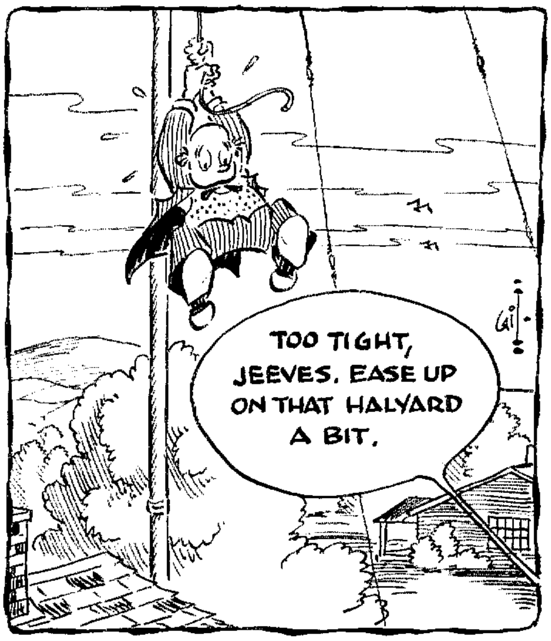
It has been assumed, for this simple model, that the refractive-index gradient is constant with altitude and that it extends over the entire propagation path. Discontinuities caused by a temperature inversion or a sudden change in humidity, as occurs at the interface between fog and clear air, have not been considered. Since it is these discontinuities that produce "ducting," or "trapping," of the ray paths, extended propagation coverage resulting from these effects is not predicted by this simple model. Nevertheless, this simple computer model is a useful tool for estimating terrain clearance and potential coverage of a repeater site. It should also help you to better understand vhf/uhf propagation.

Notes

[1] m = ft × 0.3048; km = mi × 1.609.
[2] B. R. Bean and E. J. Dutton, "Radio Meteorology," *National Bureau of Standards, Monograph 92* (Washington, DC: NBS, 1966).
[3] B. R. Bean, J. D. Horn and A. M. Ozanich, Jr., "Climatic Charts and Data of Refractive Index for the United States and the World," *National Bureau of Standards, Monograph 22* (Washington: NBS, 1960).
[4] D. L. Haarsager, "Microwave Path Evaluation," *Ham Radio*, January 1978, pp. 40-42.
[5] *Reference Data for Radio Engineers*, 5th ed. (New York: Howard W. Sams and Co., Inc., 1969) p. 26-14.

Getting the Most out of Your Antenna

BEGINNER'S BENCH



All dipoles are created equal, but some work better than others. Is your antenna doing the best it can for you?

By C. L. "Chuck" Hutchinson,* K8CH

Asking which is the best antenna is about like asking which is the best automobile. Almost everyone has a different opinion. Personal operating habits have a lot to do with one's choice. But more important, consider *where* and *how* your antenna can be installed!

For the amateur who wants a simple, effective and low-cost antenna for hf, a 40-meter dipole is a great way to get started. The cost is minimal, it will fit on most housing lots, and you get the bonus of 15-meter coverage. That's right, two bands for the price of one and no Transmatch (tuner) is needed. If casual contacts are what you seek, this antenna may be all you'll ever need. Should the "DX bug" bite, you will not be "uncompetitive" on 40 meters; on 15 meters, it is unlikely that yours will be "the loudest signal on the band."

How well your dipole performs depends on a number of factors, not the least of which is where you install it. It is easier to contact China using a dipole installed in Japan than with one in Vermont, but that is not what I mean! For a dipole to work best, it should be erected high and in the clear. To understand why requires a knowledge of how horizontally polarized antennas act when installed over the earth. Furthermore, one should grasp the concept of how radio waves get from one place to another.

Dipole Radiation Patterns

Consider a dipole in "free space." That means completely removed from the vicinity of all other objects, including the earth! A dipole in those circumstances will exhibit the familiar radiation pattern shown in Fig. 1. You may have seen the figure-8 pattern before, but be aware that this is only a partial representation. A more complete picture is shown in Fig. 1B. The doughnut shape represents the three-dimensional dipole radiation pattern.

All antenna patterns are three dimensional. They are seldom represented that way in books and magazines because there is no easy way to represent three-dimensional-pattern intensities accurately on a flat surface. Most of the time you can get a good idea of pattern shape by looking at pattern plots taken in horizontal and vertical planes. Fig. 1A illustrates the pattern of a dipole in a plane that contains the antenna. Let's call it horizontal, even though that means nothing in space. The pattern in a perpendicular or "vertical" plane would be a circle. Combining the two patterns results in the one shown in Fig. 1B.

Getting Down to Earth

What happens when the dipole is moved from free space to your backyard? As you have probably guessed, the presence of earth will modify the dipole radiation pattern. Energy striking the earth will be reflected from it. This reflected energy will combine with the directly radiated energy causing reinforcements and cancellations (lobes and nulls) at various angles.

There is a major difference in the way horizontally and vertically polarized radio waves reflect off the earth. In Fig. 2 you can see this difference represented graphically.

A radio wave has two components — magnetic and electric; by definition the two exist at right angles to each other and at right angles to the direction of travel. All

*Assistant Technical Editor

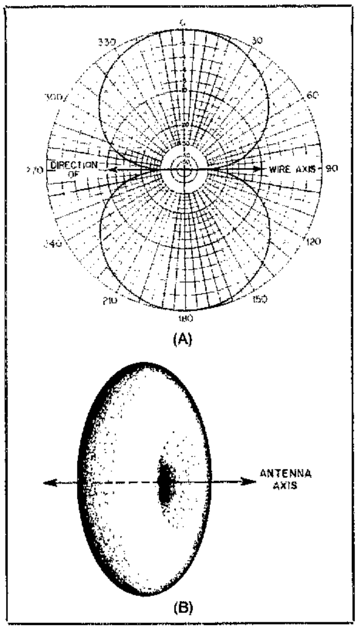


Fig. 1 — At A, radiation pattern of a half-wave antenna in the plane containing the wire axis. At B, the solid pattern of the same antenna.

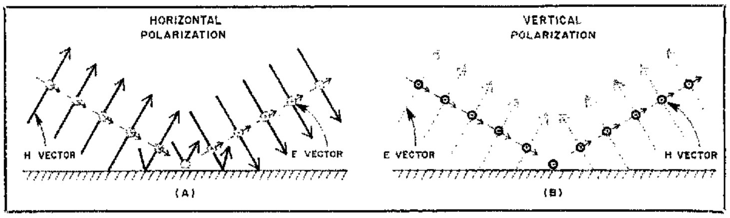


Fig. 2 — Horizontally polarized waves are shifted 180° in phase upon reflection, as shown at A. Vertically polarized waves, as shown at B, are not shifted. Circles with an X in the center are vectors (arrows) pointing into the page; those with dots in the center are pointing out of the page.

three can be thought of as vectors, or arrows, all mutually perpendicular. To visualize how this works, think of the seams in the corner of a room — where the two walls meet and where each wall meets the floor. These three seams are all at right angles to each other.

The electric field of a horizontally polarized antenna is horizontal, and the magnetic field is vertical. With a vertically polarized antenna the relationship is reversed; the electric field is vertical, and the magnetic is horizontal.

Alignment of the magnetic component (H vector) of the radio wave determines the nature of the reflection. In the case of horizontally polarized radio waves, the H vector is coming to the ground at an angle. You can visualize what happens during reflection by thinking of that H vector as an arrow that is made up of thousands of short pieces. Each piece reflects from the earth as it hits. The result, as seen in Fig. 2A, is that the vector is turned over. The electric field (E vector) must retain the "right sense" to the magnetic field. Therefore, the E vector (shown going into the page before the reflection) is shown coming out of the page afterward. This change in direction represents a 180° phase shift.

In the case of vertical polarization, the magnetic vector comes down parallel to the earth. This time there is a simple reflection with no phase shift or reversal. This assumes that the earth is a perfect conductor.

The effect of the ground reflection is the same as if a second "image" antenna were placed as far below the earth as the "real" antenna is above the earth. The image antenna acts as if it were fed the same amount of energy as the real antenna. For horizontally polarized antennas, the two are 180° out of phase. That means that when a point on the real antenna is at a positive peak the corresponding point on the image is at a negative peak (Fig. 3). For vertical polarization, the antenna and image are in phase.

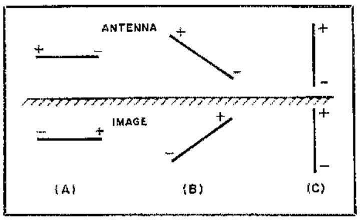


Fig. 3 — Antenna and image phase relationships are shown at A, B and C.

Let's assume that a dipole is installed 1/4 wavelength above ground. We could also say the dipole is 90 (electrical) degrees above ground. (One wavelength represents a full cycle or 360°.) Can you imagine the pattern that exists in a plane perpendicular to the antenna? In free space it would be a circle. What will it be here?

Fig. 4 shows two antennas spaced $\lambda/2$ apart and fed 180° out of phase. The drawing "freezes" an instant in time. The heavy solid lines can be thought of as positive peaks and the lighter, dashed lines as negative peaks. Where solid lines (or dashed lines) touch, the energy from the two antennas is in phase and they combine by adding. Where solid and dashed lines touch each other, there is a cancellation or nulling of energy.

The situation depicted in Fig. 4 is like placing a horizontal dipole $\lambda/4$ above the ground. There is one difference, however. We can ignore the bottom half of the figure — radiation exists only above the ground. You can see that there is maximum radiation straight up. Also notice that there is a complete cancellation, or null, along the horizontal center line. That line corresponds to the earth's surface.

Let's review the situation as we now understand it in the case where we have a horizontal dipole $\lambda/4$ above the ground. First, the doughnut-shaped pattern of free space is altered by the presence of the earth. Second, the resulting pattern is the same as if another dipole antenna were placed $\lambda/4$ below the ground and fed an equal amount of power, 180° out of phase with the real antenna. The radiation pattern will have a maximum straight up and a minimum along the ground.

How can we determine the pattern intensity of our dipole at angles between the horizontal and the vertical? The method is illustrated graphically in Fig. 5.

At the angle of interest two rays contribute to the radiation, one direct and one reflected. The reflected ray will travel farther than the direct ray (refer to Fig. 5). We see that the difference would be zero (wavelengths or degrees) if the antennas were located at A and C. Line AD (the distance from A to D) is equal to BD. The reflected ray travels a distance equal to BC farther than the direct ray. This effect creates a delay or phase difference at a receiving antenna. Even if the rays were to travel 100 or even 1000 miles, the reflected wave would still have to travel that relatively short distance farther.

It is not enough to know how long in feet or meters the distance BC is. That distance must be converted to electrical degrees.[1] The result is the phase difference, or phase angle, between the two rays. With that information, the two can be combined mathematically. This sounds involved and difficult, and it can be. However, through the use of trigonometry the calculations turn out to be straightforward.

Vertical radiation patterns for horizontal dipoles at various heights can be seen

[1]Notes appear on page 37.

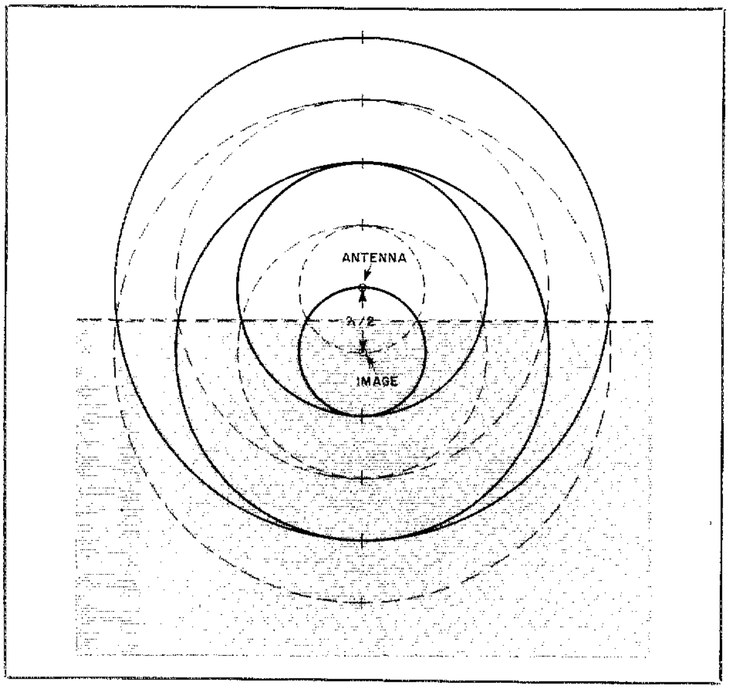


Fig. 4 — Interference between waves causes directional effects. See text.

in Fig. 6. When the antenna is placed at odd multiples of 1/4 wavelength above ground, there is maximum radiation vertically; at even 1/4-wavelength spacings (multiples of 1/2 wavelength) there is a vertical null. The higher the antenna is raised, the lower the angle of the first lobe.

Why the Vertical Radiation Pattern is Important

Radio communication is possible at hf because, depending on conditions, the ionosphere refracts radio signals, returning them to earth. Fig. 7 shows an ionized layer above the earth. When radio signals enter that layer at an angle, they are bent or refracted; when conditions are right they return to earth. Lower-angle radiation returns to earth at a greater distance; higher angles yield shorter distances. Actual conditions are much more complicated than this. A future article in the Beginner's Bench series will cover radio propagation in more detail. We will be concerned only with a few general principles for now.

Our first rule of thumb is that lower radiation angles yield greater distances. The second is that lower frequencies tend to use higher angles of radiation. It is useful for you to capture those two concepts. Let's look at some numbers.

Table 1 is taken from *The ARRL Antenna Book*. Observe that the angle of arriving signals varies with frequency. A close examination reveals three numerical values and relationships for each band on the England-to-New Jersey path. First, useful radiation angles will be below the "upper angle." The upper angle is determined by merely changing the meter-band designation to degrees. For example, on the 40-meter (7-MHz) band, the upper angle is 40°. Second, the most useful angle will be about half the upper angle. Third, at times you may use angles at or less than one quarter the upper angle.

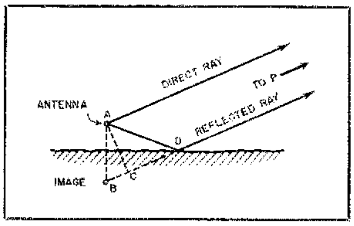


Fig. 5 — At any distant point, P, the field strength will be the resultant of two rays, one direct from the antenna, the other reflected from the ground. The reflected ray travels farther than the direct ray by the distance BC, where the reflected ray is considered to originate at the "image" antenna.

The numbers in Table 1 apply to the path across the North Atlantic; they don't apply to all situations. As you can see, there is no single angle that is best for a given path or band. There is, however, a preferred range of angles. Let's look at a practical example.

Earlier, I stated my preference for a 40-meter dipole as a first antenna. Let's suppose you have two supports 69 ft high (λ/2) and properly situated to support a dipole.[2] Fig. 6C shows the vertical radiation pattern off the sides of the antenna. (There will be little useful radiation off the ends.) Comparing with Table 1, we see that the radiation from our dipole is concentrated in angles a bit higher than optimum for the North Atlantic path. To center dipole radiation on the angles indicated in the table would require that the dipole be raised to 100 ft or more! Few of us have trees, or other supports, that are anywhere near that high. (Perhaps I can find some 30-ft-high planters for the 70-foot oak trees in my front yard.) Most of us will be content if we get our dipoles close to 69 ft high.

Our friendly 40-meter dipole also works on 15 meters, but the pattern is a bit more complicated. The horizontal "free space" pattern is shown in Fig. 8. A 15-meter dipole 69 ft above the earth would have the pattern of Fig. 6F off the sides. Our (3/2-wavelength) dipole will have a pattern that is the sum of these two patterns. As with the 40-meter band, radiated energy is concentrated in angles a bit higher than optimum. Nevertheless, this also represents a reasonable compromise.

What happens if we erect an 80-meter dipole at 69 ft (λ/4)? The broadside, vertical pattern will be that of Fig. 6B. This is not a very good DX antenna; most of the radiated energy goes straight up. This combination should work well for local contacts and for communication out to several hundred miles, however.

Most experienced hams agree that it is best to get your antenna up at least 1/2-wavelength if you want to work DX. But there is nothing magical about 69 ft for the 40-meter (or any other ham) band. Think of it as a desirable height, not as an absolute minimum height.

Getting your antenna up high will not only give you a lower radiation angle, it will help minimize losses. Trees, shrubs, buildings and the like will tend to scatter and absorb radio waves. This phenomenon increases with frequency, becoming critical in the uhf and microwave ranges. Whenever possible, you should try to get your antenna above such objects.

But I Can't Get It up That High!

Not all, perhaps not many, of you have a pair of well-spaced, tall supports for a dipole. In fact, you don't need two tall supports. Maximum antenna radiation takes place at the high current points. In a half-wavelength dipole, the high current point is at the center. For that reason, you may choose to put up your dipole as an "inverted V."

For best radiation, the angle between the two sides of an inverted V should be as

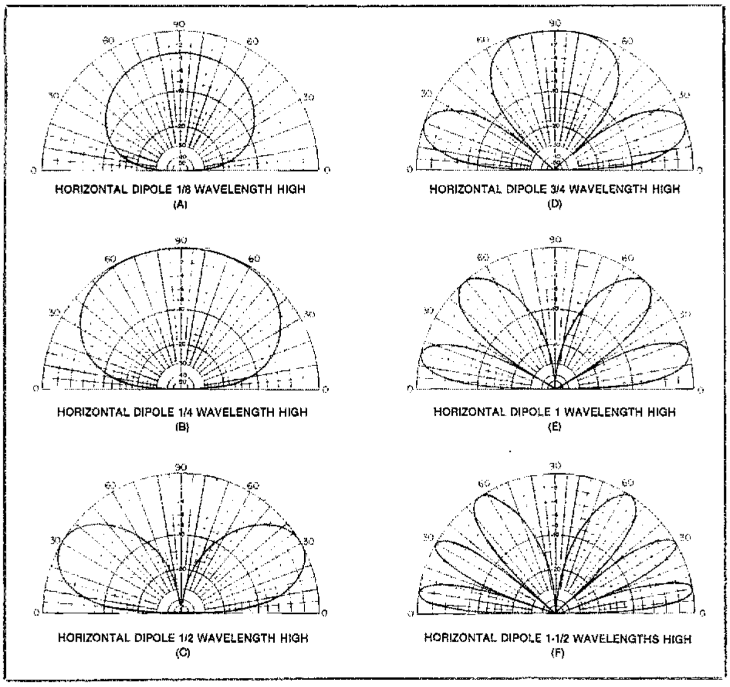


Fig. 6 — Vertical radiation patterns for horizontal dipoles at various heights above ground. The free-space pattern would be a circle at −6 dB on the scale. For directional antennas, these are factors to which the free-space radiation pattern should be added to include the effect of reflection from perfectly conducting ground. These patterns are only for the vertical angle of radiation of horizontal antennas.

Table 1

Measured Vertical Angles of Arrival of Signals from England at Receiving Location in New Jersey

Freq. (MHz)	*Angle below which signals arrived 99% of the time*	*Angle above which signals arrived 50% of the time*	*Angle above which signals arrived 99% of the time*
7	35°	22°	10°
14	17°	11°	6°
21	12°	7°	4°
28	9°	5°	3°

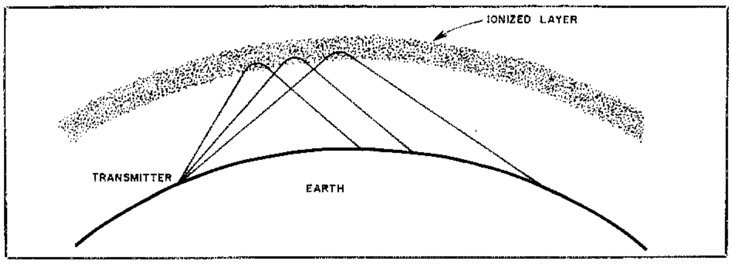


Fig. 7 — Simplified model of the ionosphere showing how hf radio signals are propagated over long distances.

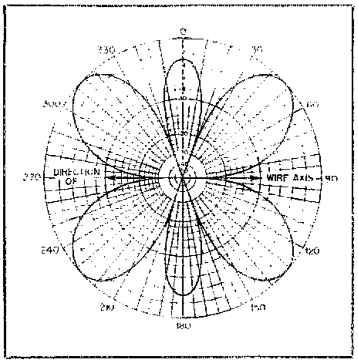


Fig. 8 — Radiation pattern of a 3/2-λ dipole in the plane containing the wire axis.

great as possible. If the sides were parallel (zero-degree angle), the results would be the same as from a feed line: There would be no appreciable radiation. With a 120° angle between sides, losses in effectiveness should not be noticeable; at 90°, they will be. Try to avoid angles of less than 90° between sides of an inverted V. My 40-meter inverted V has its center at about 60 ft; the angle between sides is between 120° and 90°. I have had good DX and domestic performance from that antenna.

Vertical antennas can be good DX performers, particularly on frequencies below 14 MHz. But they depend on ground conductivity and certain installation factors for good low-angle radiation. While vertical antenna DX performance can be very good, the same cannot be said for performance at a range of more than a few miles to a few hundred miles. Much more could and should be said about vertical antennas, but that must await another time.

For a low-cost, versatile antenna it is hard to beat the horizontal, center-fed dipole. For best results, get your dipole up as high and in the clear as you can. Good luck.

Notes

[1]The distance, in feet, divided into 984 gives the number of wavelengths. Wavelengths times 360 equals degrees. To change meters to wavelengths, divide into 300. To change wavelengths to radians, multiply by 2 times π (3.1415927...). Most computers use radians in making calculations.

[2]m = ft × 0.3048.

Strays

I would like to get in touch with...

☐ former members of the Lawrence Institute of Technology ARC, W8QOA, of Southfield, Michigan. Eric Janle, KA8OBP, 3032 Golfhill Dr., Pontiac, MI 48055.

☐ anyone having a parts list or schematic drawing for a Side Band Engineering Del Mar Model 225 12-channel vhf marine radio transceiver. Francis E. Erdle, N3AJ, 21930 New Hampshire Ave., Brookeville, MD 20833.

Hints and Kinks

Conducted By Larry D. Wolfgang,* WA3VIL

A 400-Hz POWER SUPPLY FOR A RADIO COMPASS

☐ I needed a 400-Hz supply that could power two surplus radio compass units, one for azimuth and one for elevation with my EME antenna system. After many unsuccessful attempts to wind a 400-Hz transformer for a previous design, I decided to try a 555 timer and a flip-flop to switch a pair of transistors.[1] Fig. 1 shows the schematic diagram for my circuit. Most of the parts were ones I found in my junk box, but the parts list gives Radio Shack part numbers for anyone who wishes to buy all new components. I tried a variety of transformers at T1, and they all seemed to work well.

As an additional use of this circuit, I discovered that it is ideal for finding the operating frequency of transformers. The oscillating frequency of the timer can be adjusted over a wide frequency range, and the transformer under test connected as T1. Find the frequency of maximum output voltage and current, with minimum current drawn from the dc supply. This will be the optimum operating frequency of the transformer. — *Gene Wasson, K0UDZ, Rapid City, South Dakota*

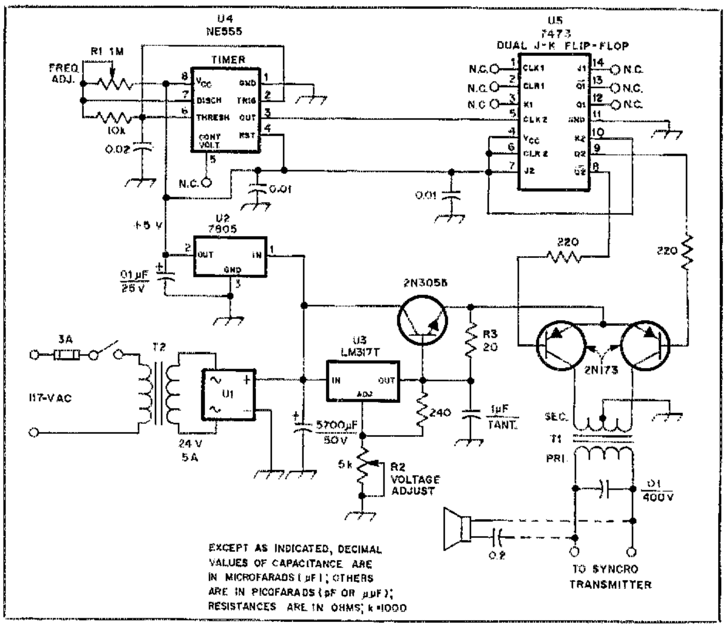


Fig. 1 — Schematic diagram of a 400-Hz power supply that can be used with a radio compass for antenna-direction readout (part numbers in parentheses are Radio Shack).

T1 — 117-V primary, 25-V secondary, used as a step-up transformer (273-1512).
T2 — 117-V primary, 12- to 18-V secondary, at least 4-A rating.
U1 — 6-A, 50-PIV full-wave bridge rectifier (276-1180).
U2 — LM 7805 5-V regulator (276-1770).
U3 — LM 317T adjustable voltage regulator (276-1778).
Q1 — 2N3055 or equiv. (276-2041).
Q2, Q3 — 2N173 or equiv. (276-2006).

RETUNING MOBILE ANTENNA, FAST AND EASY

☐ I use a Hustler mobile antenna mounted on the rear bumper of my car. During extended trips from home, I keep cw and ssb schedules on several bands. Peak performance of this antenna requires resetting the adjustable whip to resonance when changing from the cw to the ssb portions of a band. Hustler provides a collar on the adjustable tip. It acts as a preset guide for the correct length at the higher frequency, but resetting for a preferred lower frequency requires some other method of measuring tip length. The method I devised involves placing strips of electrical tape on the rear bumper. These are spaced a distance equal to the resonant length of the tip at my preferred frequency. I simply lay the entire resonator along the bumper, adjust the tip to match the spacing of the tapes and tighten the jamb nut. Of course, a pair of tape strips is required for the resonator of each band. — *Victor Woodling, W9JNH, East Peoria, Illinois*

MULTI-BAND HF MOBILE ANTENNA

☐ My mobile station consists of a Kenwood TS-120S and four Anixter-Mark Heliwhip antennas. These antennas are helically loaded. A separate fiberglass rod is used for each band. Stopping along the highway to change the antenna for each band change became a nuisance.

I fabricated a three-band adapter. Later, I made a four-band adapter, using 1/8-inch hard aluminum plate.[2] For the four-band version, I used a T-shaped piece of aluminum. The three end portions of the bracket were bent down at an approximate 30° angle. I drilled four holes in the bracket to clear the standard 3/8-inch × 24 antenna-mounting stud. To assemble the antenna, I mounted the longest whip through the top of the T and into the base. The remaining three coils were then secured to the adapter by means of lock washers and nuts.

I have experienced no problems with this system. Band hopping is now effortless, and my mobile operating is more enjoyable. This principle should be useful with other types of mobile antennas, such as those that employ interchangeable loading coils. — *Merritt Scarff, W4FAH, Lakeland, Florida*

COLLINS EQUIPMENT 40-METER COVERAGE CHANGE

☐ While changing the 8.9775-MHz crystal in position 3C of my Collins equipment to provide coverage of the 30-meter band, I decided to make another crystal change. By removing the 10.355-MHz crystal in position 2B and replacing it with a 10.255-MHz unit, I shifted the coverage on this 40-meter segment from 7.2-7.4 MHz to 7.1-7.3 MHz. This enables me to tune the entire 40-meter phone band in one segment. I believe this was a worthwhile change. — *John Wasmuth, W8BP, Marquette, Michigan*

TOWER THRUST BEARING PROTECTION

☐ I have a 65-foot tower with a 5-element triband beam antenna. To make the rotator easier to service, I have it mounted at the 20-foot level. I use a 50-foot mast pipe (1-11/16 inch OD) from the rotator through a heavy-duty thrust bearing mounted on the top plate of the tower. The thrust bearing supports the entire weight of the mast and antenna.

I was concerned about what would happen to my investment if water were allowed to find its way into the bearings over a period of time. A protective cover will prolong the useful life of the bearing and minimize the need to replace this expensive piece of hardware at frequent intervals.

My solution was to use a "plumber's helper" to form a cover. Carefully enlarge the hole that the handle threaded into until it is about 1/4-inch smaller in diameter than your mast. Slide it down the mast until the cover is against the bearing

[1] E. L. Campbell, "Antenna Rotators and Indicators," *QST*, April and May 1967.
[2] mm = in. × 25.4; m = feet × 0.3048.
*Assistant Technical Editor

plate. A small bead of silicone sealer around the top of the helper completes the job. Fig. 2 illustrates the final arrangement. I inspected the bearing after six months of exposure to our inclement Michigan weather, but found no traces of rust. — *Dick Carter, N8DMO, Jenison, Michigan*

(NOT) STORING IC CHIPS IN STYROFOAM®

☐ When I opened my copy of the new *Hints and Kinks* book I was shocked to see the suggestion about using foam-plastic meat trays to store ICs.[3] The trays I have seen are completely nonconductive and are capable of holding large static charges. Static-sensitive ICs would be destroyed instantly! That's why commercial foam for storing IC chips is conductive, and will not hold a static charge. You can easily verify this with an ohmmeter. — *Kjeld Hvatum, N1BSP, Cambridge, Massachusetts*

☐ A word of caution is in order about the use of foam-plastic trays to store IC chips. As integrated circuits become more compact, and operate on less and less power, they become even more sensitive to the ravages of static electricity. How sensitive? This was the subject of a videotape shown recently to the employees of NCR Corporation, a computer-systems manufacturer.

You don't have to walk across a carpet in a dry room to build up enough charge to damage or destroy an integrated circuit. With the right combination of synthetic fibers in your clothes, you can ruin an IC through mishandling it while sitting at a workbench. But the most dramatic demonstration on the videotape was a perfectly good commercial-grade IC destroyed by the static electricity in a foam-plastic cup waved once over the chip! No spark, no sound, just one defective IC.

Most amateurs don't deal with such high-performance, large-scale integrated circuits — yet. But we can all save ourselves some trouble and expense by keeping our IC chips away from nonconductors capable of holding even a small electrostatic charge. This would include cellophane tape, foam-plastic and ordinary plastic bags. Computer manufacturers use specially treated plastic bags with conducting surfaces. Only then can the components be safely shipped in a box of foam-plastic "peanuts." — *Ken Nollet, KØEN, St. Paul, Minnesota*

TOWER THRUST BUSHING PROTECTION

☐ For several years, I have been using a rubber boot to keep the thrust bushing on my tower clean and dry. The boot was fashioned from a rubber toilet-tank ball. [This is the part that fits in the drain hole at the bottom of the tank. — Ed.] Cut the top of the ball to fit snugly around the mast, and cut the bottom so it will stretch fit over the thrust bushing. Slide the boot down the mast until the bottom can be stretched over the bushing. A stainless-steel hose clamp placed over the boot just above the thrust bushing holds it it place (Fig. 3). Be sure to pack the bushing with grease before installing the protective cover. — *Thomas Kruszon, WB2PXL, Riverhead, New York*

[3] *Hints and Kinks for the Radio Amateur*, 11th ed. (Newington: ARRL, 1982), p. 1-2.

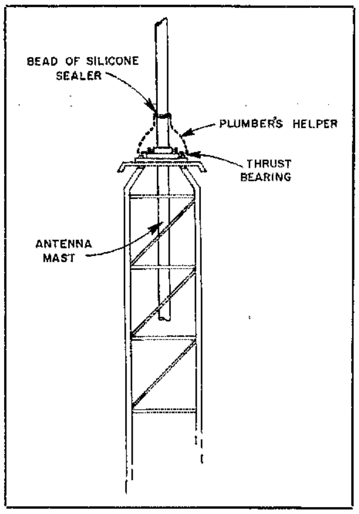


Fig. 2 — A "plumber's helper" can be used to protect a thrust bearing from the weather.

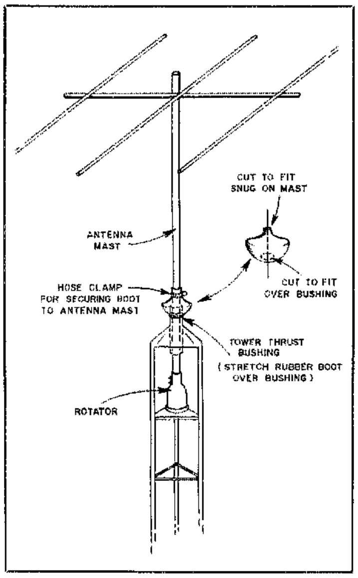


Fig. 3 — WB2PXL uses a rubber toilet-tank ball to form a protective boot for the thrust bushing on his tower.

☐ I use a piece of rubber-impregnated material, such as the cuff from an old rubber glove, to make a weatherproof cover for the thrust bushing on my tower. Fig. 4 shows this technique. The cuff is secured with a stainless-steel hose clamp, just above the top of the bushing. Be sure to apply a good coat of bearing grease between the mast pipe and the bushing. — *Ed Lynch, WB2YKX, Latham, New York*

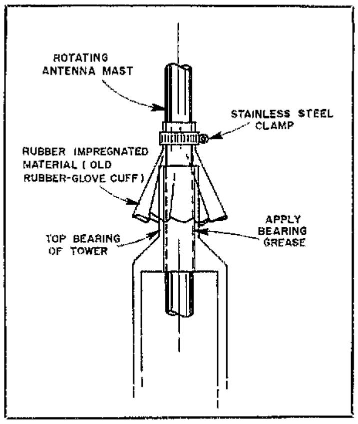


Fig. 4 — Illustration of the protective cover used by WB2YKX to weatherproof the thrust bushing on his tower.

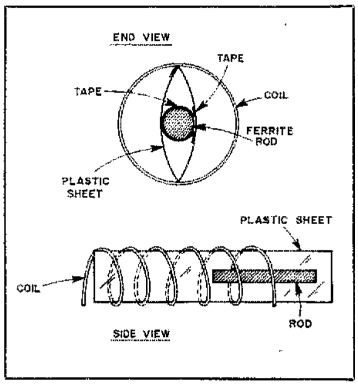


Fig. 5 — A piece of stiff plastic and a ferrite rod can be used to make a homemade variable inductor.

FERRITE-ROD SLIDER FOR VARIABLE INDUCTOR

☐ Here is an idea that I have employed to make a homemade variable inductor for use in a Transmatch or other piece of equipment. I mount a ferrite rod inside the coil, and slide the rod through the coil to change the inductance. To mount the slider inside the coil, I use a piece of stiff plastic (the type used for covers on greeting-card boxes works fine) that is twice as long as the coil. Next, I cut the plastic to a width of 2.75 times the inside diameter of the coil. Tape the ferrite rod to the sheet as shown in Fig. 5 and fold the sheet in half. The assembly will fit into the coil with just a slight compression. This arrangement will minimize any dielectric loss and provide an easy way to slide the rod back and forth through the coil. You could fasten a wood or plastic handle to the end of the assembly and add a scale for setting the inductance. — *Glenn Morrison, KO9L, Evanston, Illinois*

Conducted By
Dennis J. Lusis,* W1LJ

OPERATING WITH AMTOR

☐ [Editor's Note: The following is taken from a report outlining on-the-air experiments with AMTOR. The report was submitted to the FCC on November 12, 1982 in accordance with the terms of STA 7130. This STA allowed K4PA to experiment with AMTOR prior to legalization of this mode on hf for all duly licensed American amateurs.]

A Report on Experiments With the CCIR 476 ARQ Teleprinter Code:

1.0 INTRODUCTION

AMTOR (MODE A) is a teleprinter system that is described in CCIR Recommendation 476 and uses a fast-break simplex radio channel to automatically request repeats (ARQ) of garbled data.[1,2] This "handshaking" process is a significant improvement over traditional RTTY operation and provides nearly error-free copy over marginal hf circuits.

Beginning in January 1982, I (K4PA) commenced two-way communication with other AMTOR stations around the world, in an effort to field test microprocessor equipment used to generate and detect CCIR 476 signals, and to determine whether the AMTOR system was suitable for use by amateur RTTY stations. From January through October, 22 AMTOR stations in 11 countries were contacted, with a total operating time of 80 hours. All contacts were made on 14,075 kHz.

2.0 AMTOR PERFORMANCE

Is It Really Error-Free? Yes, almost perfectly so! AMTOR performance is so near to its promise that a meaningful error-rate estimate is difficult to make. Typing errors are the most significant source of garbled copy.

AMTOR is much like wideband fm communication in that, as propagation conditions vary, a sharp threshold marks the boundary between perfect copy and an unusable circuit. That boundary, however, is substantially below the signal/noise ratio usually necessary for hf RTTY work. When signals fade, AMTOR automatically reduces the "throughput rate," maintaining perfect copy until finally giving up at what seems to be the last possible moment. If signals fade and return repeatedly, AMTOR operators communicate with ease while conventional RTTY operators try to fill in the garbled portions of their messages.

What is the Data Rate? The block timing specifed in CCIR 476 permits a maximum rate of 66 wpm. This "throughput" is realized if no repeats are requested. Band conditions determine how often repeats are needed. On the average, circuits seem to run at 30 to 60 wpm, probably a result of the operators using AMTOR in conditions that are otherwise unacceptable. Speeds of less than 10 wpm may be accompanied by significant errors and are not generally used.

Unfortunately, CCIR 476 ARQ timing is somewhat complicated and adapting it to faster rates is not a simple clock-frequency change. In most stations, changes to the microprocessor code in ROMs would be required. To my knowledge, no amateur experiments along these lines have been conducted.

3.0 THE AMTOR OPERATING ENVIRONMENT

AMTOR has several unique characteristics related to the basic ARQ process:

Message Receipt Confirmation. By virtue of the ARQ handshaking technique, operators using AMTOR are certain that all transmitted data blocks have been received correctly. They can monitor the transmission process at any time to learn exactly how much of a message has been sent.

Full Break-In Operation. CCIR 476 allows the changeover between sending and receiving data to be initiated from either end of a circuit. Information-receiving stations may interrupt incoming messages by pushing a control key.

Transmitter Look-Thru. With AMTOR, the radio equipment at both ends of a link is continually cycling between transmit and receive at a 0.45-second rate. Activity on the channel is noticed easily during the listening periods.

Effective Power Adjustment. When propagation is good, the AMTOR operator can reduce power and immediately observe the effect on channel speed. He or she can do this while either sending or receiving data, and can experiment without affecting reception quality. It is quite dramatic to observe how superfluous a "full gallon" is on good circuits — and how helpful a 10-dB power increase is with marginal conditions.

Reduced Interference Potential. An interesting feature of AMTOR is that both ends of a link are "always" transmitting. This precludes the common situation in which a frequency is erroneously thought to be clear due to inability to hear one side of a contact. No QRL? transmissions are needed.

4.0 EQUIPMENT REQUIREMENTS

Almost all AMTOR equipment takes the form of a hardware interface between conventional data terminals (ASCII or Baudot) and the RTTY station modem. The signaling rate is 100 baud, within the 45 to 110 range commonly employed. The transmitter shift may be anything the operators agree upon, and has been exclusively 170 Hz in my experience.

AMTOR operation requires radio equipment capable of relatively fast break-in operation. A nominal requirement is that the transmitter achieve nearly full output within 20 ms after changeover is signaled. AMTOR equipment generally provides a delay adjustment to withhold transmission until receive/transmit changeover is completed. Tardiness at this point limits the maximum range at which other stations may be contacted. CCIR 476 block timing allows 170 ms for the round-trip delay, but receive/transmit changeover time must be subtracted from this value for range calculations. I have learned that a 30-ms delay prohibits ARQ communication between Virginia and Australia!

Most amateur equipment uses mechanical relays as part of the transmit-receive changeover process, most often in antenna-switching circuits. A subtle difficulty arises when other transmitter circuits can switch faster. In this case, we find that an antenna relay will "hot-switch" the transmitter output and generate a click over considerable portions of the band. The problem, of course, has been solved with most modern rigs operating in the cw mode. Yet, this does not guarantee a particular transmitter will be clean when running RTTY or when speeded up for AMTOR. One solution, employed by KB6BT and me, uses a gated afsk generator that is delayed and shaped to provide a clickless turn-on after all relays have settled.

A second timing requirement with AMTOR is that the transmit-to-receive changeover be made as quickly as possible. That is, when the transmitter is turned off, the receiver agc must allow full sensitivity in 20 ms or less. Interestingly, this delay affects the *minimum* range over which AMTOR stations may communicate. If a nearby slave station begins control signal transmission 20 ms after changeover, the master station must be ready to receive at that time. When the two stations are distant, signal propagation delay mitigates this constraint.

5.0 TYPICAL AMTOR EQUIPMENT

Most stations are using the Mark 2 AMTOR kit from G.P.W. Electronics, Ltd.[3] This is a single circuit board designed by G3PLX that translates CCIR 476 to and from Baudot code at 45 or 50 baud. A 1023-character typing buffer is also provided. Input/Output are at TTL levels.

A few stations are reportedly using software-only modifications to popular hobby computers and single-board microprocessor systems. I use a machine-language program for AMTOR Mode A in a Kim-1 microcomputer dedicated to RTTY operations.

6.0 AMTOR SIGNAL CHARACTERISTICS

Compatibility with Other Signals. The frequency shift and signaling rate used with AMTOR are commensurate with present amateur RTTY equipment. Only the characteristic "chirp-chirp" of the simplex ARQ process distinguishes these transmissions from other amateur RTTY. I believe that most RTTY operators are now familiar with AMTOR signals as a result of experiments conducted by U.S. stations and the considerable activity in other parts of the world. It is estimated that at least 80 stations in 20 countries are now active with AMTOR.

An interesting aspect of AMTOR operation is that it is fully compatible with commercial TOR equipment now gaining acceptance in maritime use. AMTOR stations have adopted the practice of using the alphabetic portions of their call signs to determine a four-letter selcode unique to each station. Commercial apparatus uses a five-digit sequence assigned to each sta-

[1] J. Martinez, "AMTOR, an Improved Error-Free RTTY System," *QST*, June 1981, p. 25.
[2] J. Martinez, "AMTOR, an Improved Radioteleprinter System Using a Microprocessor," *Radio Communication*, Aug. 1979, p. 714.
*Assistant Technical Editor
[3] 55 Cobham Rd., Ferndown Industrial Estate, Ferndown, Wimborne, Dorset BH21 7RA, England.

tion, which is converted internally to four characters, in accordance with CCIR Recommendation 491. Nevertheless, amateur and commercial stations may communicate easily with one another once the appropriate selcode is determined.

Is the AMTOR Block Design Appropriate? Recognizing that AMTOR achieves nearly perfect elimination of transmission error, one may ask what can be done to improve transmission speed. CCIR 476 yields a communications efficiency of 50%, in the sense that a 100-baud radio channel (with ARQ handshaking) produces a 50-baud teleprinter output. To increase the AMTOR data rate one may adjust either the signaling speed or the ARQ efficiency.

The question of efficiency recalls a most striking feature of AMTOR — the relatively few characters (three per block) transmitted at a time. One may ask if increasing the block length will substantially reduce handshaking overhead. In my opinion, the block length cannot be increased usefully beyond a factor of 5 or 10. The problem is that with marginal hf circuits the probability of a "hit" in an interval of several seconds is very high. Using a block length of 80 characters (one line of text) seems foolhardy if communication is to be attempted during poor conditions. Perhaps an adaptive system that matches block length to channel statistics, or a method of correcting individual errors within a long block, would yield useful improvement in ARQ efficiency.

7.0 PREDICTIONS FOR AMTOR USE

AMTOR performance is such a significant improvement over conventional RTTY that enthusiasts are given to forecasting its exclusive use within a few short years. Comparison with the history of ssb telephony is unavoidable. Yet AMTOR, as is, suffers from the 50-baud limitation on transmission rate. This is adequte for any link involving a typist, but how soon will the amateur's need for computer-to-computer communication grow beyond this modest speed? Stations may already be heard lamenting that more than 40 minutes is required to send an 8-K program over the air.

Thus, the future of AMTOR involves an estimate of amateur data transmission needs and our progress with the proposed wideband vhf data networks. If the need does not materialize quickly, or if the networks still require years of development, AMTOR may well establish itself as the dominant mode of amateur data communications.

Another factor affecting my prediction of exclusive AMTOR use on hf is my preoccupation with long-haul circuits that are reliable 24 hours per day. These considerations are not typical of the great bulk of amateur RTTY communication. 100 wpm (or, possibly, 300 wpm) Baudot and ASCII codes work well on continental U.S. paths during good band conditions. Nevertheless, I find that, when a station has the option, AMTOR is invariably selected. It extends our communication range, increases the time "window" during which a circuit is open, and provides that wonderful certainty of perfect print on the other end. — *William C. Meyn, K4PA, Reston, Virginia*

AID FOR PROGRAMMERS

☐ First, congratulations on publishing the fine article by Rose, K6GKU, on the MINIMUF Program (Dec. 1982 *QST*). I was, however, somewhat disappointed when I read the editor's note on page 38. There is, and has been for some time, a book addressing the problem of using a program on a computer for which it has not been written. The book is David Lien's (W6OVP) *BASIC Handbook*, available from CompuSoft Publishing, 535 Broadway, El Cajon, CA 92021, for $19.95 + $1.65 postage and handling.

With the *BASIC Handbook*, anyone can translate the many different BASIC language "dialects" to their own computer syntax. The book is really a "must have" for any serious microcomputer owner. — *Leslie S. Smith, KA6AXX, San Diego, California*

LINE LOSS AND SWR

☐ While it is true that moderately high SWRs on transmission lines of average length at the lower frequency bands are of minor importance, things *can* get out of hand. When the 10-MHz band legally opened, my friend W7IR was anxious to give it a try, since his new rig includes that band. He has an inverted V cut for 3.6 MHz, and reasoned that since the third harmonic of that band is not too far from 10 MHz, it might work on the new band. The SWR at 10 MHz was about 5:1; this was overcome easily with a matching network. Signal reports were disappointing, however, even allowing for the 100-W output of the transceiver.

He then realized that his transmission line (mil-spec RG-8A/U) is 350 feet long, and perhaps the attenuation was causing the SWR to look much better than it really was. (At 10 MHz, RG-8A/U has an attenuation of 0.55 dB per 100 feet, or about 2.0 dB for the whole line.)

This attenuation figure means that, even if there is no antenna connected to the end of the line, the return loss measured at the shack end can be no greater than 4.0 dB; *the SWR can never look higher than 4.4:1!*

Actually, the antenna (there really *was* an antenna at the end of the line!) was radiating a small amount of power, but W7IR had not intended to operate QRP! W7IR is now looking for a place to hang a 10-MHz antenna, preferably closer to the shack. — *Harry R. Hyder, W7IV, Tempe, Arizona*

Feedback

☐ W2HXF correctly points out that the pictorial drawing of T1 in Fig. 4 of "Go Class B or C with Power MOSFETs," March 1983 *QST*, p. 27, has an error. Leads D and E should be joined, rather than leads C and D. The schematic diagram for T1 is correct, however. Also, T2 should have been described as a bifilar choke.

☐ There is a typographical error in "More MINIMUF Program Mods" (Technical Correspondence, May 1983). The following line should read:

2010 B1 = (remainder per May issue)

Author Bramwell is credited with the correction.

☐ VE2CV has pointed out that in Fig. 7 of "The Effect of Supporting Structures on Simple Wire Antennas," Dec. 1982 *QST*, p. 34, the angles in the small insert have been reversed. The angle θ is the angle of elevation for launch of skywaves; ϕ is the azimuthal angle ($\phi = 90°$ is the plane orthogonal to the plane containing the antenna).

☐ In "SOLAR FLUX/SUNSPOT NUMBER CONVERSION FOR MINIMUF" (Technical Correspondence, May 1983), there are errors in Eq. 3. It should read

S9=625*(SQR((0.73)↑2-0.0032*(65-SF))-0.73) (Eq. 3)

Thanks to K4JW for picking this up.

☐ Author Belrose points out that there are several errors in the Appendix of his January 1983 *QST* article entitled, "Beverage Antennas for Amateur Communications." Eq. 1 should read: Z_o equals 60 times the natural logarithm of 2h/a. While the symbol n is defined correctly, this definition should have appeared in the text after the list of definitions following Eq. 6. Eq. 5 should read

$$G = \frac{377\pi \sin^2\delta}{Z_o \lambda^2} \left| \frac{1-e^{-\ell\Gamma}}{\Gamma} \right|^2 \qquad \text{(Eq. 5)}$$

Eq. 4 was converted incorrectly from metric to English units. It should have read

$$Z_o = 60 \ln \frac{2(3.7)(12)}{0.0404} = 462 \text{ ohms (Eq. 4)}$$

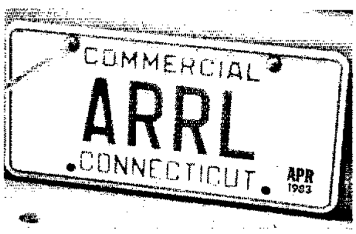
where

h = 3.7 feet (1.13 m)

a = 0.0404 inches (0.1026 cm)

☐ *QST* has received a virtual landslide of information from various sources telling us that the June Stray entitled "Contact-Lens Wearers Beware!" (page 22) is erroneous. Apparently, we (along with several other publications) were the victims of a hoax: The incident described in the Stray, according to the information we've received since the Stray appeared in print, did not occur. The underlying message is this, according to the National Society to Prevent Blindness: "Contact lenses, of themselves, do not provide eye protection in the industrial sense. For occupational use, contact lenses should be worn only in conjunction with appropriate industrial eye protection."

Strays

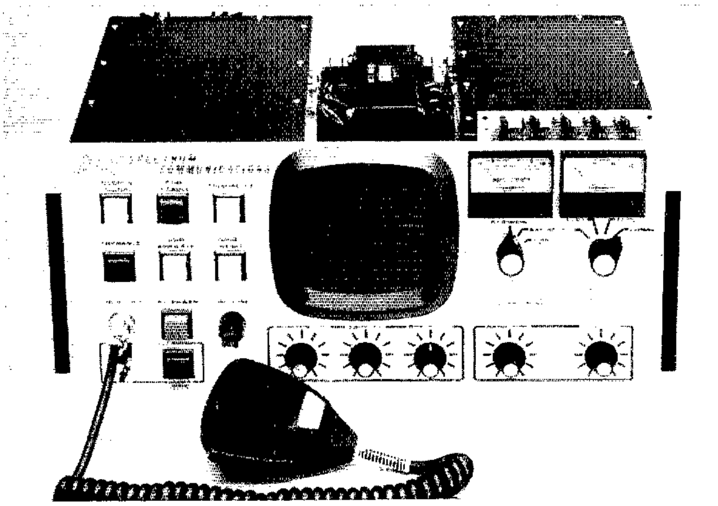


The League's 1979 GMC pickup truck, used mainly for hauling mail to and from the Post Office, is sporting a new license plate, compliments of the family of the late Francis E. Handy, W1BDI. The family graciously transferred ownership of the license plate to the League in memory of Handy's 42 years of service to the ARRL and Amateur Radio. Known affectionately as "Mr. Amateur Radio," Handy is best known for creating *The Radio Amateur's Handbook* ("Handy's Handy Handbook") and for founding the ARRL section-level organization and Field Day, which, by the way, celebrates its 50th anniversary this year.

Product Review

Conducted By Paul K. Pagel,* N1FB

Spectrum Communications SCR 1000 2-Meter FM Repeater



Reliable, trouble-free, solid-as-a-rock — those are the words that first come to mind when I think of the Spectrum Communications SCR 1000 2-meter repeater. Spectrum has relied heavily on time-proven circuits and components to put together a package that should be a good performer for any group that doesn't want to "roll their own."

The review unit was in use daily at W1AW/R during a period that stretched from mid-July through the following January. After overcoming some initial difficulties, we experienced no significant problems with the unit. W1AW/R is located in an elevator penthouse in a large public building in Newington. The penthouse is neither heated nor air conditioned, which means the temperature extremes are *not quite* as severe as would be experienced if the unit were housed in a metal box hanging on the side of a tower.

During the first couple of weeks of the review, the unit twice failed to transmit. On both occasions, we pulled it from service and returned it to the ARRL lab for testing. The first time we found nothing wrong with it back at the lab, so we returned it to service. After a few days it again failed to transmit. This time we sent it back to the factory. Although there was some minor damage in shipping, the factory could find nothing basically wrong with it. They returned it to us after a few days, and we put it back into service. Craig Baker, one of Spectrum's engineers, suggested that the problem might have been falsing in a remote inhibit circuit.

The inhibit circuit can be activated or deactivated by remote means. On a few occasions after that, the transmitter failed to operate. The first time, one of the technical crew made the trip to the site and found the INHIBIT indicator light on, indicating that the inhibit circuit had been set. (Because we were unfamiliar with the unit and were not testing it at the site, we neglected to notice if the light had been on during the first two "failures.") The transmitter also failed to transmit on three other occasions. Each time, we were able to deactivate the inhibit function remotely and return the repeater to service without "making a house call." Because the repeater could be inhibited remotely, we were never really certain whether the incidents had been true "falses" or were the acts of some whimsical jester. Also, we could not make it "false" in the lab. Those were the only difficulties experienced with the unit during the entire review period.

Spectrum Communication SCR 1000 2-Meter Repeater, Serial No. 1844

Manufacturer's Claimed Specifications	*Measured in ARRL Lab*
Frequency range: 136 to 174 MHz.	
Sensitivity: 0.3 μV for 12-dB SINAD.	0.42 μV for 20-dB quieting.
Squelch/COR threshold: 0.1 to 0.2 μV typical, 0.25-μV maximum.	0.13 μV.
Hang time: 0.1 to 6 seconds.	Less than 0.5 second (minimum) to approx. 10 seconds (max.).
Time out range: 0.5 to 4 minutes typical.	Approximately 12 seconds (min.) to approx. 5 minutes (max.).
Power output: 30 W at 13.8-V dc, 25-W minimum.	28 W.
Modulation: True fm, 7-kHz maximum.	Not measured.
Preemphasis: 6 dB per octave.	Not measured.
Spurious emissions: −75 dB typical, −70 dB minimum.	See Fig. 1.
Size (HWD): 7 × 19 × 13 in.†	
Weight: 21 lb.	

†mm = in. × 25.4; kg = lb × 0.454.

On-the-Air Reports

The repeated audio fidelity is excellent. Frequently, I tried listening to the input and output of the repeater. Except for the obvious variations in signal strength, there was little, if any, difference between the tonal qualities of the two. The audio is handled the "right way." Audio from the fm detector in the receiver is deemphasized at 6 dB per octave rolloff according to EIA specifications. In the transmitter, the audio is preemphasized, again according EIA specifications, before being applied to the modulator, a modified Clapp oscillator configured to produce true fm. The acid test for such a unit is the application of a human voice having a deep "gravelly" quality. We are blessed with two such test cases on W1AW/R, and I am happy to report that the SCR 1000 passed both voices with flying colors.

The SCR 1000 replaced a real "clunker." Users were ecstatic with the increased range of the new machine. The output power seemed well-balanced with the sensitivity of the receiver. Transmit and receive ranges appeared to be nearly equal for a typically equipped 10-W mobile station with a 5/8-λ antenna. Our unit came outfitted with a CTCSS decoder built in. Being located near the coast, we are subject to periods of tropospheric ducting. For hours at a time, the repeater may key up on weak distant signals. Experiments with the CTCSS decoder indicated it is an effective means of keeping the distant weak signals from activating the repeater without decreasing the range. If your repeater suffers from this sort of "interference," it may be worthwhile considering adding a CTCSS decoder for use at such a time, if not all the time. Modern encoders are inexpensive and easily installed in most transceivers, including hand-held portables.

The receiver front end consists of eight high-Q resonators with an amplifier following the first two and another amplifier following the next three resonators. A doubly balanced mixer converts the 2-meter output of the front end to the first i-f at 21.4 MHz. Two monolithic crystal filters follow the mixer and deliver the i-f signal to a single-stage amplifier. The output from the amplifier is then filtered by an additional set of monolithic crystal filters. From there, the signal is processed by a MP5071, which provides the functions of second mixer, amplifier, limiter, fm detector and squelch noise amplifier. The squelch

*Assistant Technical Editor

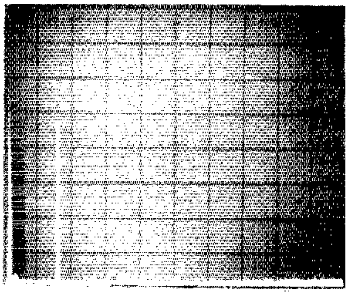


Fig. 1 — Spectral display of the SCR 1000. Vertical divisions are each 10 dB; horizontal divisions are each 100 MHz. Output power is approximately 28 W at a frequency of 145.45 MHz. The fundamental has been reduced in amplitude approximately 33 dB by means of notch cavities; this prevents analyzer overload. All spurious emissions are about 78 dB below peak fundamental output. The SCR 1000 complies with current FCC specifications for spectral purity.

circuit detects noise above 35 kHz, which makes it immune to many of the shortcomings found in some simpler squelch circuits.

Audio from the receiver, local microphone and i-d generator is applied to a varactor diode supplying load capacitance to the 18-MHz crystal. From the modulator, the signal is multiplied, filtered and amplified to provide approximately 2-3 W of 144-MHz energy at the exciter output. From there, the signal is fed to the power amplifier that boosts the level to the 30-W range. Spectrum currently uses a 2N6082 in this stage. A 75-W PA is also available as an option.

Control circuitry consists of TTL-compatible CMOS logic. Relays are not used; current consumption for the circuitry is quite low, a critical parameter when operating from emergency backup power sources. Squelch range, hang time and time-out functions are all adjustable, as would be expected.

One observation about the squelch is in order. Aside from operating on 35-kHz noise, the squelch circuit has hysteresis. That is, it takes a stronger signal to open the circuit than it does to keep it open. Without this provision, a weak, fading signal is apt to be "chopped" up by the squelch opening and closing as the signal crosses back and forth over the critical point that opens the squelch. We did not notice such chopping with this circuit, but we have with our "clunker." The hysteresis is a true blessing for the operator attempting to talk with another station having a weak, fading signal.

Besides the circuits necessary for the functioning of the repeater, the SCR 1000 comes with a number of goodies that make life a lot easier for the repeater technical crew. Two meters on the front panel provide means for making a quick check of the entire system without the need to remove a single screw. The first meter can be switched between EXCITER RF, FINAL RF and RECEIVED SIGNAL STRENGTH. The second permits reading of the 5-V SUPPLY VOLTAGE, 12-V SUPPLY VOLTAGE, EXCITER CURRENT and FINAL AMPLIFIER CURRENT. MONITOR VOLUME, SQUELCH, REPEATED AUDIO, HANG TIME and TIME OUT can all be adjusted with controls located on the front panel. An ac line fuse holder is also found on the front panel. A push-button switch turns the ac power on and off (as with the other switches on the panel, this one is illuminated when activated). The control operator can either inhibit the COR or simulate it with another pair of push-button switches. Another switch provides for manually keying the i-d circuitry. A final push-button switch will inhibit the transmitter when activated (illuminated). The switch will also light when the transmitter has been inhibited remotely. The switch (local control) can override the remote inhibit, but a remote reset will not override a local inhibit. Another indicator illuminates with an incoming signal.

The owner's manual for the SCR 1000 is a pleasure to read. Aside from explaining the functioning of the repeater circuitry in detail, the manual provides general information on repeater operation and general troubleshooting of the system — not just the repeater itself. Solutions are suggested for several commonly encountered problems. Complete technical specifications, schematic diagrams and alignment data are provided.

Although the unit came with a built-in (optional) autopatch circuit, we were unable to give it a test at the repeater site. (Because of conditions beyond our control, we were unable to get a telephone line installed at the repeater site.)

Users have found the SCR 1000 delightful. It was a sad day when we boxed the SCR 1000 up to ship back to the factory and brought the "clunker" out of storage. Price class for the SCR 1000 is $1600; for the CTCSS option, $120; for the autopatch, $600. Additional information can be obtained from Spectrum Communications Corp., 1055 W. Germantown Pike, Norristown, PA 19401. — *Peter O'Dell, KB1N*

CUSHCRAFT CORPORATION 40-2CD 40-METER SKYWALKER YAGI

□ Most amateurs who take their contesting seriously eventually ask themselves the following question: "What can I do about 40 meters?" Why 40 in particular? Well, it's on this band that the Sweepstakes is won and lost, and it's on this band that the outcome of most cw DX contests is decided. Forty is the one band where nine times out of 10 the average dipole or inverted V just doesn't "cut the mustard." Most successful contest stations have gain antennas on 40 for two reasons: The added signal strength puts them a cut above the rest in the pileups, and the front-to-back and front-to-side ratios help get rid of some of the trash that makes hearing the weak ones so difficult.

What kind of gain antenna to use depends on several factors. A vertical array is nice, but an extensive ground system is essential if it's going to be effective. And often the radiation angle is so low that a vertical system isn't optimum. Fixed wire beams and quads are popular too, but they're only good in one or two directions. Eventually, the choice becomes clear: Some sort of rotatable horizontal antenna is the solution.

Again, the would-be 40-meter "big gun" is faced with a bewildering number of options. Currently on the market are full-sized 2- and 3-element Yagis, 2-element quads and shortened 2, 3, 4 and 5-element Yagis. The full-sized antennas offer good bandwidth and gain. They are also large and heavy, often requiring use of spacial towers, masts and rotators. The shortened beams are lighter and somewhat smaller, but many are a compromise in terms of performance and SWR characteristics.

Enter Cushcraft Corporation's most recent addition to its line of hf "Skywalker" antennas. Designated the 40-2CD, this 2-element beam

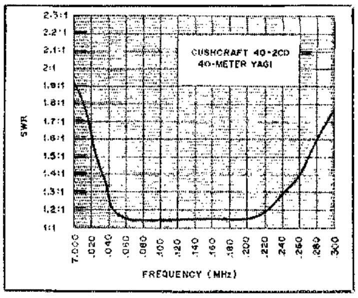


Fig. 2 — SWR curve of the Cushcraft 40-2CD 40-meter Yagi.

offers good performance in a practical package. The antenna consists of a driven dipole element and a reflector element mounted on a 22-foot boom.[1] Each element is shortened to approximately 43 feet through the use of loading coils and small capacitance hats. A 3-element 40-meter beam of similar design has been described in the ARRL *Handbook* for the past several years.

The antenna is made from high quality, polished-aluminum stock. The boom consists of four separate pieces. The center section is made from two 2-1/8 inch OD pieces joined by a coupler at the boom-to-mast plate. A 56-inch-long piece of 2-inch-OD stock is inserted into each end of the center section to complete the boom. The larger-diameter tubing is slotted at each junction, and a steel worm gear-type hose clamp holds the sections together securely.

Each element is made from telescoping sections of tubing. Reflector halves are joined at the boom by a 3-foot length of 1-3/8 inch OD tubing. Driven element halves are joined at the boom by a black plastic insulator tube that is 8 inches long. Element tubing decreases from 1-1/4 inches OD at the center to 1/2-inch OD at the tip.

The elements are secured to the boom by 1/4-inch-thick aluminum plates and U bolts. These plates have grooves machined in them to help prevent element twisting. The 1/4-inch-thick boom-to-mast plate and associated U bolts and saddles provided by Cushcraft will accommodate masts up to 2-1/8 inches OD. A boom brace made from aluminum tubing runs from the mast to the ends of the boom to provide support.

The hardware provided with the beam is all first-rate. What isn't stainless steel or aluminum is well-plated and should hold up without rusting for a long time. In addition to the construction steps outlined in the owner's manual, I coated the telescoping tubing sections with conductive grease at each junction to prevent aluminum-oxide buildup, which can cause antenna performance to deteriorate over the years. I also coated the threads of every screw with noncorrosive silicon sealant to prevent any vibration from loosening the nuts. Plastic end caps for the boom and elements completed the installation.

The antenna went together in about three hours. The instruction manual is fine. Cushcraft seems to subscribe to the theory that a picture is worth a thousand words; the clearly labeled detailed drawings left no questions in my mind as to which bolt went where. My only complaint is that four of the hose clamps were missing from

[1] m = ft × 0.3048; mm = in. × 25.4; km = miles × 1.609.

Cushcraft Corporation 40-2CD 40-Meter Yagi

Manufacturer's Claimed Specifications	*Measured by ARRL*
Boom length: 22.3 feet.	As specified.
Longest element: 43 feet.	As specified.
Turning radius: 24 feet.	As specified.
Assembled weight: 44 pounds (20 kg).	As specified.
Wind surface area: 6.38 ft.^{2}	Not measured.
Frequency coverage: 7.0-7.3 MHz.	As specified.
Bandwidth (2:1 VSWR): 200 kHz.	See Fig. 2.
Material: 6063-T832 seamless tubing.	

the parts package. Although Cushcraft is very good about sending replacements for missing or damaged pieces, I decided to purchase replacements at a local hardware store to save time.

It took Gary Hitchner, WA2OMY, and me about half an hour to install the completed antenna atop 100 feet of Rohn 25. We installed the antenna in one piece. Gary pulled a rope rigged through a pulley on the mast while I followed the antenna up and fished it around guy wires. The relatively light weight and small size of the antenna made it an easy job. Caution should be taken, however, to watch the capacitance hats at the element tips. They do tend to get caught on guy wires, and too much bending and twisting could detach them.

The completed antenna is mounted on a no. 4130 seamless steel tubing mast about 5 feet below a 3-element 20-meter monobander. It is fed through a 12-turn decoupling choke made from RG-8/U cable, as suggested in the manual. The feed line is 165 feet of RG-8/U. Physically, the antenna is about the same size as the 20-meter beam. The elements do droop some, caused in part by the weight of the loading coil assemblies, but this hasn't caused any problem.

In the construction manual, Cushcraft gives three options for the resonant antenna frequency: 7.025 (cw), 7.140 (middle) and 7.220 (phone). I assembled mine for the middle of the band. The resultant SWR curve (Fig. 2) was obtained using a Bird Thruline wattmeter connected at the antenna end of the feed line. As the curve shows, my transmitter is extremely happy on both cw and phone! This type of SWR curve is not characteristic of some of the other shortened 40-meter beams on the market, but other 40-2CD owners have had similar results.

On-the-air performance of the antenna is excellent. Transmitted signal reports confirm that the antenna works significantly better than a dipole at 70 feet. Observations on receive indicate significant nulls off the side and back of the antenna. Perhaps the biggest thrill is being able to hear the many weak signals that are inaudible on the dipole.

During the three-month evaluation period (October through December 1982), more than 110 different countries were worked with the antenna. It acquitted itself well during the CQ Worldwide DX Contests and the ARRL CW Sweepstakes. Although we have had no ice yet at the time this is written, the antenna has weathered several storms with winds in excess of 50 mph with no apparent problems. The antenna still works as well as it did when first installed.

I would highly recommend this antenna to anyone looking for that added "something" on 40, but who doesn't have the desire or resources to install a full-sized antenna. The 40-2CD lists for $380. Manufacturer: Cushcraft Corporation, P.O. Box 4680, Manchester, NH 03108. — *Mark Wilson, AA2Z*

VIBROPLEX "BRASS RACER" AND EK-1 PADDLES

☐ Many electronic-keyer-paddle designs have appeared in the past 10 years or so. Each manufacturer tries a new approach to certain aspects of paddle construction: base material, paddle shape and material, type (single lever or iambic), tensioning method and overall weight. The Vibroplex Co.'s latest entries, the Brass Racer and the EK-1, are fashioned after the former HAMCO "Scotia" design.[2] The two models discussed here are constructed of hardwood, polished brass, steel and molded plastic.

Triangular paddles are used in the 'Racer — similar to the Bencher and HAL "FYO" designs. Fig. 3 is a close-up of the pivoting system; it is similar to that of the Bencher and incorporates locking screws to prevent the paddles from flying apart (remember that malady, "FYO" owners?). Paddle tension is adjusted by two permanent magnets mounted just behind the paddle arms. (Ten-Tec uses a similar tensioning scheme in one of their designs, although it uses electromagnets.) One word of caution when adjusting the tension: *Do not* pull the magnets out of their holes. There is a tiny spring and a small plastic ball that keep tension on the magnet. These parts will fly out of the paddle, and may be impossible to locate. The paddles come with a 2-foot, three-conductor cord and an unattached 1/4-inch phone plug.[3]

The EK-1 has a Curtis 8044 IC keyer built into the base. A pc-mount potentiometer is used as the SPEED control, and it protrudes through the right-hand side of the keyer base. The keyer is powered by an internal 7.5-V battery (EP-175) that the manufacturer claims should have a lifetime of up to 12 months under average use. To gain access to the battery or keyer, two recessed, bottom-mounted screws must be removed.

The keyer is constructed on a single-sided, glass-epoxy pc board. No battery polarity markings were on the board of the review unit, but the manufacturer has stated that later production models do have such identification. The instruction sheet does caution that the positive terminal be placed toward the SPEED control potentiometer.

Adjusting sending speed is a bit awkward, especially when sending with your right hand. Also, the potentiometer movement is stiff, and precise speed control settings are made difficult. Employing a centrally mounted miniature rotary control might be preferred.

The EK-1 is wired as a "right-handed" keyer. The dot and dash contact posts are connected to foils on the pc board, and no provisions have been made to swap the dot/dash keyer inputs to the opposite posts.

Initially, the review EK-1 wouldn't send over 20 wpm. A 150-kΩ board-mounted fixed-value resistor in series with the SPEED control had to be changed to 47 kΩ to reach the higher sending speeds. This, too, is a production-line change incorporated in later models.

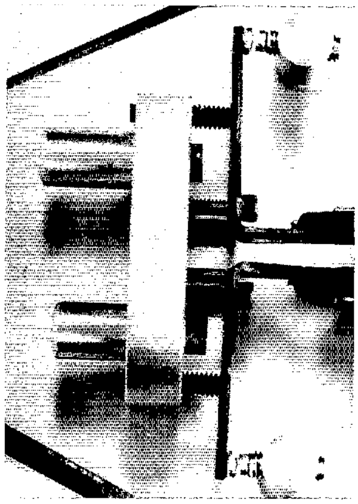


Fig. 3 — A close-up of the Vibroplex paddle.

Output line keying is accomplished by means of an MJE-350 high-voltage pnp transistor. The collector and emitter are "floating," enabling the single transistor to be used to key both positive and negative key-line voltages by correctly connecting the shield and center conductor of the two-conductor output line to the transmitter. A diode placed across the output line serves to protect the transistor in case the wrong polarity is inadvertently used. The transistor has a V_{CEO} rating of 300 V and should handle key-line voltages found in most amateur equipment. The presence of a 50-V disc-ceramic capacitor across the output line surprised us, but no problems occurred when keying a TS-820S with a key-line voltage of −65.

Although the Curtis IC has provisions for weighting control and sidetone output, these functions are not used with the EK-1. Weighting could be changed, however, by altering the value of a fixed-value resistor across pins 15 and 16 of the IC.

The paddles have a different feel, but we found them easy to adjust, and soon became confortable with the light touch required to operate them. The heavy weight of the paddle prevents it from "walking" across the operating desk. Wood, brass and black metal are combined to make an attractive addition to any ham shack. For those with a crowded operating position, the EK-1 will free up a little extra desk space.

The Brass Racer and EK-1 are available from The Vibroplex Co., Inc., P.O. Box 7230, 476 Fore St., Portland, ME 04112. Price class: Brass Racer, $75; EK-1, $110. — *Paul K. Pagel, N1FB and Gerry Hull, AK4L*

MICROCRAFT CODE*STAR READER KIT

☐ In the last few years, it has become common practice for many hams to abandon the quiet murmur of clanging gears and the sweet

[2]Product Review, Dec. 1978 *QST*.
[3]mm = in. × 25.4

fragrance of stale machine oil associated with mechanical RTTY equipment in favor of the ubiquitous home computer with an RTTY or cw send/receive program. As the price of microprocessor chips dropped, it became conceivable to build a dedicated "reader" around these chips. Such is the Microcraft Code*Star.

The amazingly compact unit (7-3/4 × 5-3/4 × 3-3/8 inches) houses a dedicated microprocessor chip, eight multisegment LED readout chips and all the circuitry needed for filtering audio signals and translating them into logic levels the microprocessor can understand.[4] Only about half of the circuit board is used; the rest is reserved for adding an optional ASCII output port — and there is plenty of room inside for building in an ac-operated supply or installing the otherwise necessary 12-V battery.

The obvious question, then, is: "If it is so small and has so few parts, does it work?" The answer is yes, it does work — and I think it works well, considering the cost. There are a few drawbacks to using the unit, though. Several other staffers used the Code*Star briefly during the review period. The most often heard complaint was that at the higher speed settings (110-baud ASCII and 100-wpm Baudot), the user found it difficult to read the display as the letters shifted left from one display unit to the next. I did not find this to be a problem, but I am a speed reader and found it easy to follow the characters on the display as long as I used a soft focus. That is simply a matter of training and conditioning.

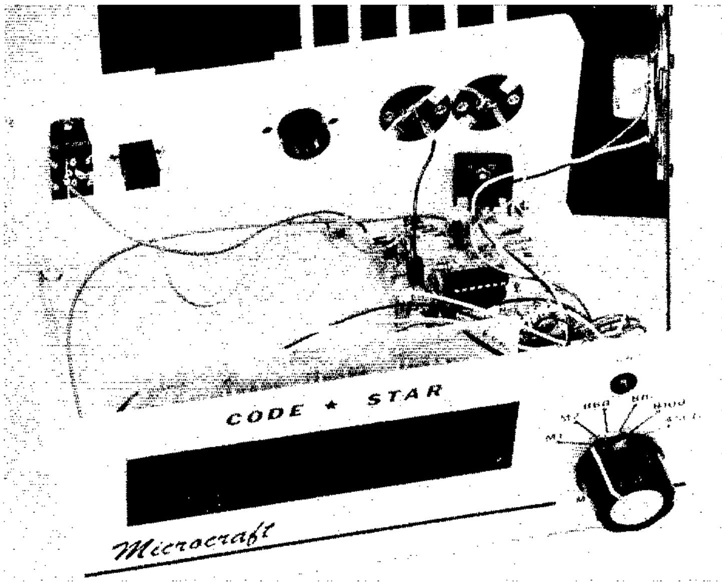
A somewhat more serious complaint is that the unit tends to "false" on sloppy code and on good code if another signal is in the passbands of the internal filters or the receiver filters. Careful tuning, additional filtering, lowering the receiver volume-control setting and similar measures can be used to minimize the effects of other signals in the passband, but nothing short of the vengeance of T.O.M. can abolish sloppy code. Fading band conditions also tend to disrupt the functioning of the Code*Star.

Provisions are made for connecting a key (or keyer) to a special input of the Code*Star for sending practice. It is a truly humbling experience to watch the machine display what you are *actually* sending!

Six modes of operation are available to the Code*Star user. M1 and M2 copy cw; M1 utilizes "heavy filtering" and has a speed range of 3 to 33 wpm, while M2 has light filtering and a speed range of from 3 to 70 wpm. Using my (memory) keyer directly into the Code*Star, I found that it copied accurately up to about 85 wpm in the M2 range. Modes M1 and M2 both auto-track, which means they automatically adjust to the speed being received within a few characters. Speed adjustments (but not mode adjustments) are handled by the microprocessor, and there are no manual means for adjusting the speed, save the MODE switch on the front panel.

Baudot (five-level RTTY code) can be copied at three speeds — no need to change those oily gears, just flip the switch. Speeds of 60, 67 and 100 wpm are available. Since amateurs tend to use either 60 or 100 wpm, I spent very little time using the 67 wpm code. It is on hand for those who like to listen to RTTY stations found outside the amateur bands.

[4]mm = in. × 25.4.



As with the Baudot mode, I found it rather tricky to tune in an ASCII station (despite the tuning indicator LED on the front panel). I understand from friends who are RTTY addicts that this is one of the curses of the mode. Once I found the proper receiver dial setting, the Code*Star functioned beautifully. After some practice, it did seem to become considerably easier to tune stations in. With these modes, the only glitches occurred during deep signal fades or when interfering signals were in the passbands. (The perfection of mere machines will never conquer the human tendency to err. Some RTTY operators compensate for their mechanically flawless sending and receiving equipment with terrible spelling and atrocious grammar.)

The Code*Star kit is intended for intermediate to advanced builders. Microcraft offers to refund the purchase price minus a handling charge for any unassembled kit within 10 days of the original shipping date if the purchaser feels he or she is not able to construct the kit. The Code*Star is available in assembled form at a slightly higher price.

The kit instructions are straightforward, and there are only 3-1/2 pages of them. That is a little misleading, though. The second instruction is, "Install all fixed resistors and solder. Save scrap leads." There are nearly 60 resistors. An average kit builder could expect to complete the kit over a weekend.

The microprocessor, three support ICs and six transistors are installed on the main board. A second board, mounted perpendicular to the main board, holds the displays and drivers. The two boards are attached with several jumpers made from the scrap leads saved in step 2. Both boards have parts-placement guides silk-screened onto the front sides to aid construction.

All parts except the custom microprocessor are covered by a 90-day warranty. According to the manual, the custom microprocessor is "100% tested and burned-in. However, it can be damaged by mishandling and incorrect voltages and therefore is not included in the warranty."

Initially, the Code*Star did not function correctly. I returned it to the factory, and they found I had installed a transistor backward. Afterward, it *usually* functioned okay, but there were times I couldn't get it to copy anything. My initial assumption was that rf was affecting it (I was using it in the ARRL lab while W1AW was on the air). That assumption proved incorrect. Finally, after a few months of sporadic failure, it "died." Back to the factory, where the technicians found a faulty capacitor in the input circuit. Apparently, it had intermittently failed. They replaced the capacitor, and I have experienced no more problems with the unit. Microcraft reports that this was a highly unusual failure.

In my opinion, the unit has a lot of value per dollar. It does a reasonably good job of copying cw, Baudot and ASCII. It is small, lightweight and convenient to use, having no external controls other than the MODE switch and the ON/OFF switch. To connect it to the receiver, one merely plugs a jumper into the headphone output of the receiver. A tuning LED indicates when you are "in the ballpark" while tuning in a station.

When the unit is in use, receiver audio is passed to a 2-inch speaker in the Code*Star. At times, I would have preferred having the ability to use headphones and the Code*Star. The volume control is a pc-mounted trimmer potentiometer that is accessible only by removing the cover. Both of these shortcomings could be remedied with a couple of minor circuit modifications. (Gee, I've got 20 minutes, a 25-k Ω volume control and a phone jack. Think I'll go turn the soldering iron on!)

Price class of the Code*Star kit is $160; wired, $220. The optional ASCII output port kit has a price class of $60; wired, $80. Additional information can be obtained from Microcraft Corporation, P.O. Box 513Q, Thiensville, WI 53092, tel. 414-241-8144. — *Peter O'Dell, KB1N*

W5LFL: First Ham in Space

Ever work an Astronaut aboard an orbiting spacecraft? Not likely! Here's your chance.

By Roy Neal,* K6DUE

"W5LFL, from the flight deck of the Space Shuttle *Columbia.*" Those exciting words came a step closer to reality when ARRL President Vic Clark, W4KFC, and other ARRL officials, accompanied by Vern Riportella, WA2LQQ, President-Elect of AMSAT, went to Houston on May 17 for a meeting with Dr. Owen Garriott, W5LFL, and a team of NASA officials and engineers.

NASA has granted permission for Garriott, an Advanced class amateur and electrical engineer, to carry a 2-meter transceiver on the flight of STS-9. The Space Shuttle is now scheduled for a nine-day mission to be launched on September 30. He will operate up to an hour a day with the understanding that his hamming will not interfere with his work as an astronaut.

The *Columbia* is now at the Kennedy Space Center, being modified to carry the European Spacelab 1. The flight path, at an inclination of 59°, will carry it over most of the heavily populated areas on earth.

It was agreed that W5LFL will operate on the low end of the 2-meter band, between 144 and 146 MHz. Many foreign amateurs are restricted to those frequencies.

Garriott will transmit on several preannounced frequencies and listen for replies on perhaps a dozen others. He will probably transmit for one minute, identifying himself and the stations he has been able to copy. Then he will listen for a minute, logging calls received. Exact transmit and receive frequencies and format will be announced once they are determined.

He will fly at an altitude of around 155 miles, with a speed of some 17,000 miles an hour. This will permit line-of-sight communication up to 8 minutes over most locations on the flight path. Orbital tracks and times also will be made available and disseminated widely.

NASA engineers estimate that small gain antennas will be needed to ensure good reception. Designs will be suggested for simple, inexpensive construction. Garriott, by the way, will probably use a special helical type of antenna on a printed-circuit board, built by NASA and constructed to mount in one of the windows at the rear of the flight deck. He will probably use 5 W on fm and may even try to pound brass for a few cw contacts.

He will use his astronaut's mini-cassette recorder to tape all activity, including the signals received. This will constitute the log of amateur space station W5LFL. The League will act as his QSL manager, and cards will acknowledge listener reports as well as QSOs.

There are many details still to be worked out, but the teams are now at work to coordinate the effort to write some pioneering history into the annals of Amateur Radio with the first ham in space, Dr. Owen Garriott, W5LFL, this fall. Stay tuned for the "how to" details. They'll be released soon by NASA, AMSAT and the ARRL.

The STS-9 NASA Space Shuttle-Amateur Radio project is one step closer to reality. Meeting recently at Johnson Space Center in Houston to work out the details were (l to r): AMSAT Executive Vice President WA2LQQ, NBC Science Editor K6DUE, ARRL Satellite Program Manager W9KDR, ARRL President W4KFC, ARRL Public Information Coordinator KB1N, and NASA Astronaut W5LFL. *(NASA photo)*

During a tour of the Shuttle trainer, Astronaut W5LFL shows ARRL Public Information Coordinator KB1N the window where the antenna will be mounted in the *Columbia. (NASA photo)*

*c/o NBC News, 3000 West Alameda Ave., Burbank, CA 91523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August 1983
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OUR COVER

Have (or have access to) a 2-meter fm transceiver? Able to put together a simple turnstile antenna? If so, you've got as good a chance as anyone of working W5LFL, the First Ham in Space. All you need to know is in the article beginning on page 50. *(official NASA photo)*

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High-Resolution SSTV

High-resolution pictures approaching the quality of fast-scan TV can be sent at audio bandwidth. It just takes a little time.

By Dr. George R. Steber,* WB9LVI

Amateur-standard 8.5-second slow-scan television (SSTV) has the capability of conveying much higher picture definition than first-generation scan converters can process. A new generation scan converter, the VIDEOSCAN 1000 from Microcraft Corporation, is completely compatible with amateur-standard SSTV yet conveys all of the resolution inherent in the signal. Perhaps of greater interest is that it can also operate at scans of 17 and 34 seconds for truly spectacular high-resolution SSTV.

This article discusses certain technical aspects of high-resolution scan conversion, including the impact on picture quality and the reasoning for the added scan periods. To begin, a short review is given of some of the significant historical developments and their effect on SSTV signal standards and picture resolution.

The age of slow-scan television began in 1958 when a young engineering student, Copthorne Macdonald, realized his dream of reducing a wide-band television signal to a narrow bandwidth signal that could be sent over a voice-grade communication system.[1] Macdonald spent a number of years in perfecting the system.[2-4] His persistence and that of other dedicated amateurs finally paid off, and after considerable experimentation, field tests and a number of petitions to the FCC, slow scan was permitted on the ham bands as part of the incentive licensing plan in 1968.[5]

The SSTV system that evolved was essentially a method of conveying low-resolution images over a 3-kHz-bandwidth voice channel. Each picture was composed of about 128 scan lines transmitted at a horizontal rate of 15 lines per second (15 Hz). Thus, 8.5 seconds were required to complete a single picture transmission.

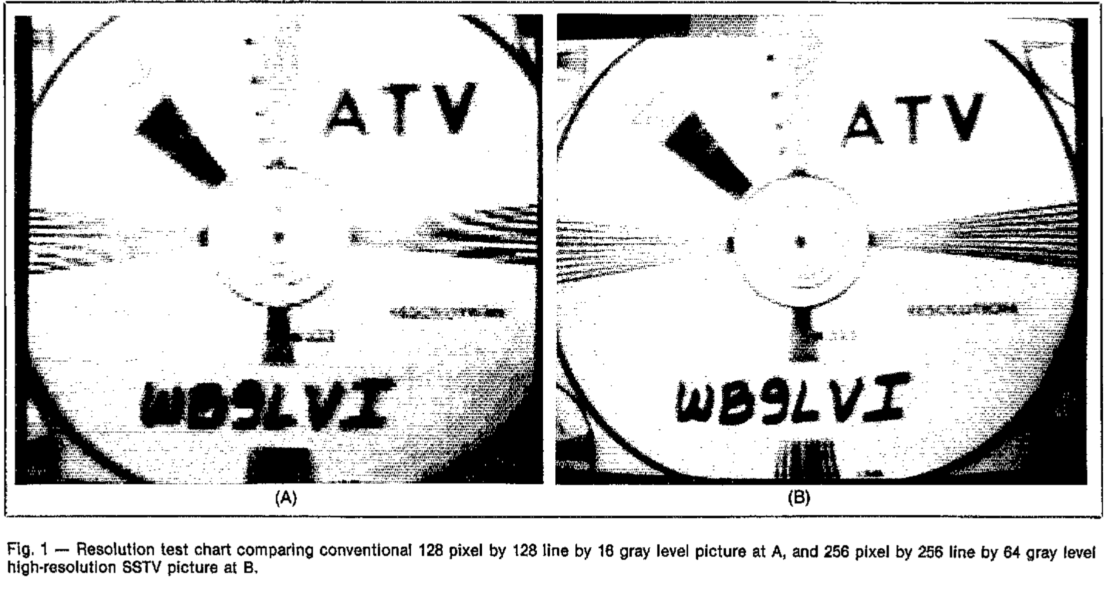
Some method of storing the 8.5-second picture was needed so that a complete frame could be viewed at one time. The only inexpensive method available at the time was the long-persistence P7-phosphor radar tube. The yellow afterglow of P7 was used as a short-term memory that enabled one to view a whole picture. Unfortunately the afterglow was very weak and faded in a few seconds, and the pictures had to be viewed in a darkened room.

In the March 1975 issue of *QST*, I described an all-digital method for storing SSTV pictures that could be viewed on an ordinary TV monitor.[6] The "LVI" scan converter revolutionized SSTV since it allowed ordinary TV equipment to be used. Its main attraction was that it produced a bright stationary image that could be frozen on the screen of a TV monitor.

The basic idea was to divide each SSTV line into 128 picture elements (pixels), quantitize the brightness to 16 levels for each pixel and store the digitized image in a memory that could be read out at fast-scan TV rates. This procedure required a

*Electrical and Computer Science Dept., Univ. of Wisconsin-Milwaukee, Milwaukee, WI 53201

[1]Notes appear on p. 13.



(A) (B)

Fig. 1 — Resolution test chart comparing conventional 128 pixel by 128 line by 16 gray level picture at A, and 256 pixel by 256 line by 64 gray level high-resolution SSTV picture at B.

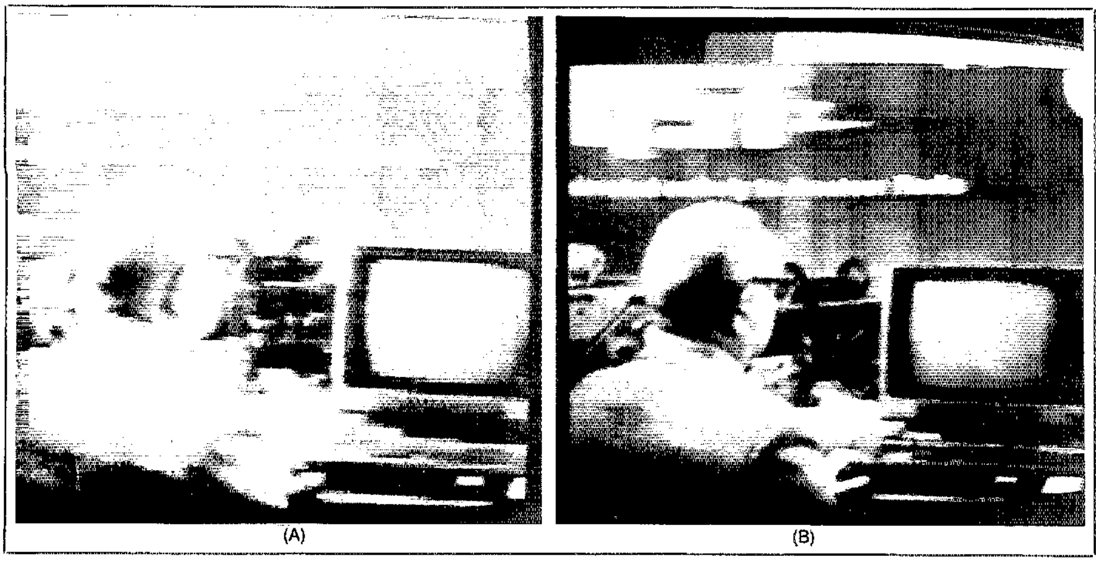


Fig. 2 — Example of pictures comparing 128-pixel by 128-line-by-16-gray-level image at A and 256 pixel by 256 line by 64-gray-level-high-resolution image at B. (These pictures were taken of the cover of May 1981 *QST.)*

large (for 1975) digital memory of 128 × 128 × 4 = 65,536 bits.

Because only 128 pixels were used on each line and only 16 levels of gray (4 bits) were used, it was recognized immediately that digital SSTV was inferior to P7 type SSTV in terms of resolution and gray scale. Nevertheless, the thrill of viewing SSTV on a regular TV set outweighed these factors and over 150 "LVI" scan converters were built world-wide based on the *QST* article. Some time later Robot Research introduced their popular model 400 scan converter, which used the same allocation of digital memory. The model 400 has since become the de facto standard for SSTV because of its wide acceptance by amateurs.

High-Resolution SSTV

With the introduction of VIDEOSCAN, pictures with greatly improved resolution may be conveyed via SSTV. This increase in picture fidelity is possible mainly because of the availability of low-cost, high-density digital memory ICs and other LSI ICs which allow more pixels of the SSTV picture to be processed and stored.

VIDEOSCAN uses 256 pixels per TV line with each pixel quantitized to 64 levels of gray (6 bits). In most cases this degree of quantitization produces a digital video signal that is almost indistinguishable from an analog signal. Objectionable contouring effects (false edges) associated with 16 gray levels are therefore eliminated.

It is important to realize that SSTV signals are bandwidth limited by the equipment used for transmission. In order for SSTV to be transmitted, the SSTV signal is converted to an fm subcarrier so that sync pulses and low frequencies in the SSTV signal can be transmitted. This fm subcarrier corresponds to 2300 Hz for white, 1500 Hz for black and 1200 Hz for sync. Since these frequencies fall within the voice spectrum, SSTV can be transmitted and received using conventional ssb equipment.

Analysis of the fm SSTV spectrum is complex. Assuming an ssb audio filter of 3 kHz, it can be shown that the maximum recoverable video frequency will be near 1.1 kHz. This corresponds to about 134 lines of horizontal resolution. Theoretically speaking this resolution could be achieved with 134 samples (pixels) per SSTV line. However, in practice, sampling a signal at the minimum rate generally leads to problems in accurately reconstructing the signal.

It has been shown experimentally that a standard 8.5-second analog SSTV signal can be closely reproduced with 170 pixels per line.[1] In other words, at least 170 samples per SSTV line should be employed to equal the resolution of the analog system. Thus the early "LVI" scan converter was somewhat deficient in taking only 128 samples. VIDEOSCAN, on the other hand, over-samples each SSTV line since 256 pixels per line are used. This oversampling enhances the detail on each line and more accurately reproduces the video waveform.

The designers of VIDEOSCAN realized that higher resolution SSTV would benefit from having more TV lines than 128. A logical choice would be to use 256 TV lines. Experiments confirmed that 256-TV-line pictures were vastly superior to 128-TV-line pictures, but they need a longer transmission time. Thus, if 256 TV lines are transmitted at 15 lines per second, this yields a vertical frame period of 17 seconds. Hence, 17 seconds was chosen as an alternate frame period for VIDEOSCAN. This period produces a very excellent picture in a reasonably short time. The overall resolution is considerably better than the early 128-pixel by 128-line scan converters.

To fully utilize the digital memory of 256 pixels per line, a slower horizontal rate is needed. Some amateur and commercial SSTV gear has used a half-rate 7.5-Hz horizontal rate. This doubles the transmission time from 17 to 34 seconds. After extensive experimentation it was found that this would be an excellent choice for a second frame period for high-resolution SSTV. There are several reasons for this choice. Foremost is the fact that the half-speed line rate of 7.5 Hz and the resulting 34-second transmission time produces exceptionally sharp, clear TV images that are comparable to commercial-quality TV images. This is true because band limiting problems in the fm modulation and demodulation process associated with the 15-Hz line rate are no longer a problem. In other words, more of the picture detail can pass through the ssb and SSTV filters when the lower line rate is used.

Of equal importance, this 34-second mode has the ability to cut through QRM, QSB (multipath) and QRN much more effectively than 8.5- or 17-second pictures. This is probably true because the interference usually represents a small percentage of each video line. Another advantage is that 34-second pictures recorded on cassette tape are affected less by wow and

Table 1

Comparison of Conventional SSTV Standards to High-Resolution SSTV Standards

	8.5-Second SSTV	*17-Second SSTV*	*34-Second SSTV*
Horizontal line rate	15 Hz	15 Hz	7.5 Hz
Horizontal samples	128/256 pixels	256 pixels	256 pixels
Vertical period	8.5 seconds	17 seconds	34 seconds
Number of TV lines	128	256	256
Horizontal sync pulse	5 ms	5 ms	5 ms
Vertical sync pulse	30 ms	30 ms	30 ms
Sync subcarrier freq.	1200 Hz	1200 Hz	1200 Hz
Black subcarrier freq.	1500 Hz	1500 Hz	1500 Hz
White subcarrier freq.	2300 Hz	2300 Hz	2300 Hz
Overall transmission bandwidth	300-3000 Hz	300-3000 Hz	300-3000 Hz

flutter of the recorder than 8.5- or 17-second SSTV pictures. Hence, an ordinary cassette recorder can be used to store high-quality SSTV pictures with little degradation.

Table 1 compares regular SSTV standards to the high-resolution standards employed by VIDEOSCAN. As can be seen, the 17-second mode is very similar to regular 8.5-second SSTV with the exception that 256 TV lines are used. The transmission of the additional lines serves only to double the transmission period. A 17-second picture, if copied on existing SSTV gear, will show only one-half of the picture and it will appear elongated. The 34-second picture mode is achieved by reducing the line rate to half, which effectively doubles the transmission time.

Comparison of Pictures

To illustrate the difference that can be expected when using these SSTV modes, a series of SSTV pictures were made as shown in Figs. 1 and 2. The original pictures were captured using a TV camera and a scan converter. It should be noted that resolution loss will be caused in the pictures by band limiting in the fm modulation and demodulation process. Additional resolution loss will be caused by the limited audio bandwidth (300 to 2700 Hz) of most modern ssb transceivers (the fm sidebands extend beyond 3000 Hz for 8.5- and 17-second SSTV). Also the photographic and printing processes have degraded the images.

Figs. 1A and 2A were generated on a WB9LVI scan converter using 128 pixels × 128 lines × 16 gray levels. The remainder of the pictures were generated and displayed using the VIDEOSCAN high-resolution scan converter.

As can be seen, the 8.5-second pictures on the original WB9LVI scan converter lack resolution and exhibit quantitization effects. The pictures made on VIDEOSCAN offer a much higher degree of clarity in all modes and have the capability of presenting very detailed images.

On-the-air experience has shown that 34-second pictures are much clearer and sharper than corresponding 17-second pictures. This advantage is offset somewhat by the two-to-one ratio of transmission times. The operator therefore has a choice: Send a 17-second picture with good resolution or a 34-second picture of very good resolution. In all likelihood, other factors such as signal strength, QRM and other band conditions will influence the decision of which high resolution mode to use at any given time.

Final Remarks

This article has presented some of the philosophy behind the VIDEOSCAN standards for high-resolution SSTV. It is hoped that these standards will prevail in the amateur community to promote growth and experimentation in the fascinating field of SSTV.

High-resolution SSTV now offers us the capability of conveying very high quality images in a few seconds almost anywhere in the world that we can talk to via ssb. This should open new horizons to Amateur Radio operators and help promote friendliness and goodwill in the international amateur fraternity.

Notes

[1]Macdonald, C., "A New Narrow-Band Image Transmission System," *QST*, August and September 1958.

[2]Macdonald, C., "S.C.F.M. — An Improved System for Slow-Scan Image Transmission," *QST*, January and February 1961.

[3]Macdonald, C., "A Compact Slow-Scan TV Monitor," *QST*, March 1964.

[4]Macdonald, C., "A Slow-Scan Vidicon Camera," *QST*, June, July and August 1965.

[5]"Slow Scan TV in the U.S. . . ." and ". . . And Slow Scan in Canada," Happenings of the Month, *QST*, September 1968, p. 80.

[6]Steber, G., "Slow-Scan to Fast-Scan TV Converter," *QST*, March and May 1975.

[7]DeWitt, B., "In Focus," *CQ*, September 1976, p. 30.

George R. Steber, Ph.D., P.E., is a professor of electrical engineering and computer science at the University of Wisconsin-Milwaukee, where he teaches courses in integrated circuits, microprocessors and control systems theory. He is a registered professional engineer in the state of Wisconsin. Dr. Steber earned his Ph.D. in electrical engineering from Marquette University in 1969. His research interests are in the areas of image processing, control systems and computer simulation, where he has published several IEEE papers.

A Life Member of the ARRL, Professor Steber has been licensed as WB9LVI since 1972. He currently holds an Advanced class license. Along with W9NTP, W3EFG, WØLMD and W6MXV, WB9LVI has been authorized by the FCC to conduct A5 and F5 (experimental medium-scan television, 36-kHz bandwidth) transmissions on the 10-meter band. Dr. Steber resides in Mequon, Wisconsin, with his wife, Gloria, and two children. In his spare time he enjoys slow-scanning on 15 meters, racquetball, photography, astronomy and visiting hamfests.

Strays

HAM HISTORY ON DISPLAY IN SOUTH AFRICA

☐ One-and-a-half years ago, there was no museum of Amateur Radio equipment in South Africa, so Julius Lieberman, ZS6AF, decided to start one in his ham radio store in downtown Johannesburg. Since then, he has managed to accrue an interesting variety of transmitters, receivers, tubes, QSL cards and other items of days past that are on display for the public. Of particular interest is the collection of telegraph keys, which now numbers about 100 types.

All items in the museum have been donated or lent by hams. Each is labeled with the ham's name and call sign, and the history of the item is recorded and filed. Donations or loans of telegraph keys are welcomed and appreciated. For more information, write to ZS6AF at J. Lieberman Electronics, P.O. Box 8628, Johannesburg 2000, South Africa.

Next Month in QST

September: the month we've all been waiting for. While aboard the STS-9 Space Shuttle, Astronaut W5LFL will conduct the first Amateur Radio operation from space. Beginning shortly after the launch, now scheduled for September 30, the operation will put Amateur Radio in the public eye as it's never been before. Elsewhere in this, the August, issue you learned how to have the best chance of garnering a W5LFL QSL card. Next month's *QST* will tell you about Owen Garriott himself: how he got into Amateur Radio, and how he became a NASA Astronaut. Fittingly, NBC Science Editor Roy Neal, K6DUE, who has covered hundreds of space shots over the years, conducted the interview.

In a related story, learn what went on behind the scenes to obtain approval for the historic Amateur Radio operation aboard STS-9.

Also in September *QST:*

• for RTTY enthusiasts, construction details for an all-electronic Amateur Radio modem (terminal unit), and an account of an experimental graphics system.

• an adapter that allows a 2-meter transmitter to control a transceive-type Mode A satellite station.

• an update on AMSAT-OSCAR 10, the long-awaited "Phase III" spacecraft that will soon be humming with QSOs.

The Boom-Excited Beam Antenna

Here's a great way to put your beam on 10 MHz or below — and cheaply, too!

By Edward C. Pienkowski,* W8BEB

With decreasing sunspot activity and the recent addition of the 30-meter band, many of us would like to operate in the lower frequencies. And we'd prefer to do this without any extra strain on our antenna budget. Why can't we use an existing Yagi?

Yagi as Dipole

Boom-exciting is a way of operating a Yagi on an additional band. It can be done without impairing correct operation on the original band. This technique is accomplished by treating the boom and the elements at the ends of the boom (usually the reflector and "last" director) as a half-wave dipole. The half wavelength is measured from the tip of the director to the boom, along the full length of the boom to the center of the reflector and then out to the end of the reflector.

My 20-meter beam is a HyGain 204BA (Fig. 1). The length from one end of the director to the boom is 15 ft 7 in. The boom is 26 feet long, and it is 18 feet from the boom to one end of the reflector.[1] This gives a total length of 59 ft 7 in., slightly less than the ideal 64 to 68 ft for a 40-meter dipole. But there is another consideration: The configuration is not a straight piece of aluminum. Electrically, it is an end-loaded dipole, in which the boom acts as the center of the dipole and the reflector and director serve as end-loading elements. It may be resonant on 40 meters, even though it is physically a little shorter than it "should" be. End loading gives it the extra electrical length that is needed.

Although end loading may not sound familiar, perhaps top loading does.[2] Top loading a vertical antenna requires placing a horizontal wire or grid work at the top of the antenna so the vertical section can be made shorter. It loads the antenna and shortens the required height by increasing the capacitance between the top of the vertical section and the ground. In our case, end loading is essentially the same thing, and results from bringing the dipole ends closer together — increasing the mutual

*40 W. 938 Whitney Rd., St. Charles, IL 60174

[1]Notes appear on page 15.

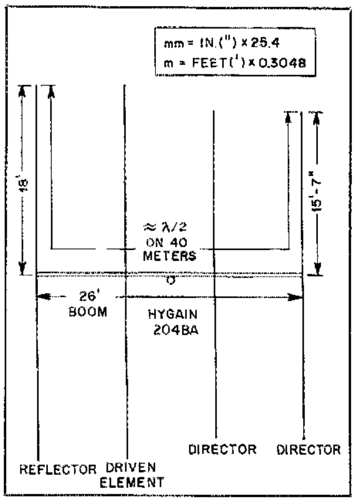


Fig. 1 — Example of how another amateur-band resonant length may be found on a Yagi.

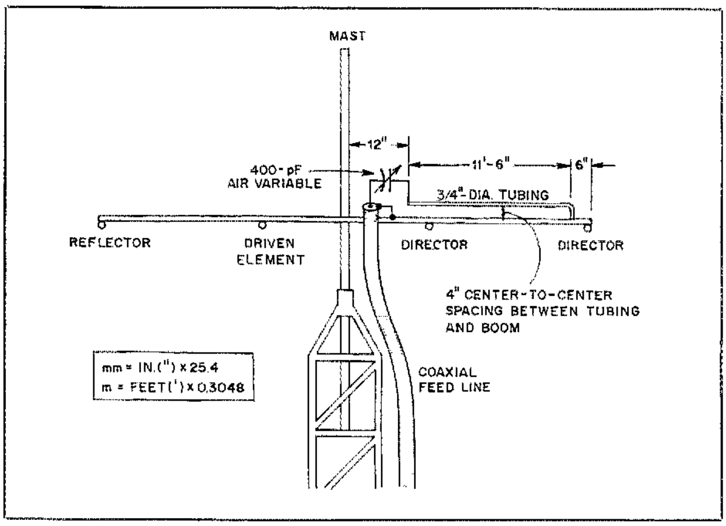


Fig. 2 — Gamma-match feed system used on the author's 20-meter Yagi.

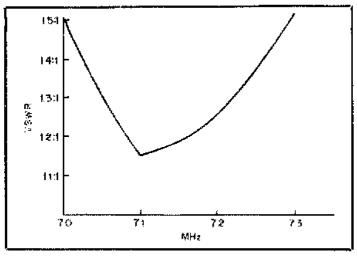


Fig. 3 — SWR curve of the boom-excited 20-meter Yagi when used on 40 meters.

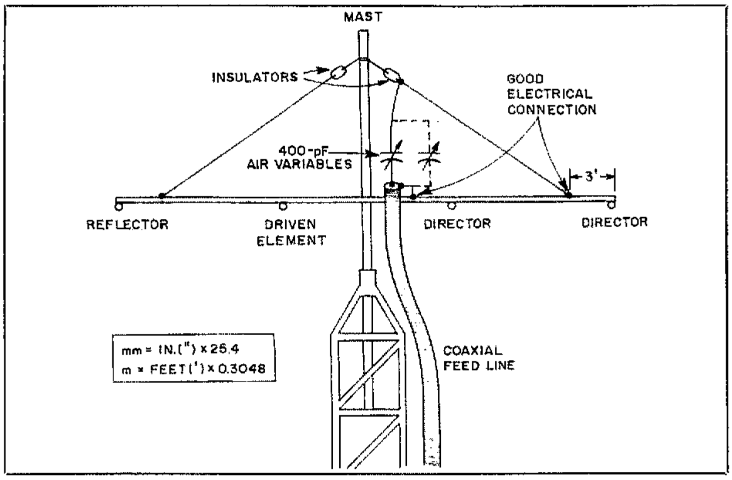


Fig. 4 — Alternative method of boom-exciting a Yagi, using the boom truss wire as part of the matching system. Capacitor in dashed lines is optional, but may be necessary to obtain a match in some systems.

capacitance. Additional capacitance or loading is provided by the other elements in the array.

Feed Systems

The first question is, "What's the best way to feed power to this system?" It's impractical to place an insulator in the center of the boom. Instead, I mounted a gamma match along the boom (Fig. 2). A delta match, a T match or an omega match should also work well.[3] I chose a gamma match because it was a simple technique for my antenna. The resulting SWR curve is shown in Fig. 3.

Another feed method that works well on a 1/12 scale model of my Yagi is to insulate the boom truss wires at the point where they attach to the mast. One of these is connected to the center conductor of the coaxial feed line through a series capacitor, as shown in Fig. 4. To obtain a good match, it is necessary to adjust the length of the truss wires and the series capacitor value. Also, care must be used in making a good electrical connection between the boom and the truss wire. On some antennas, a second capacitor connected in an omega match configuration might help. Experimentation will help determine your individual needs.

If your antenna is not the correct physical length for the desired band, the electrical length can be adjusted in various ways. One way is to insulate the outermost elements from the boom. Inductors (to lengthen) or capacitors (to shorten) the antenna can then be connected between the boom and the center of the elements. (I believe it is possible to develop a combination of coils and capacitors allowing operation on more than one additional frequency.)

Antenna length can also be adjusted through the addition of boom extensions. Based on a 1/12 scale model, a 10-foot extension on each end of the boom makes my Yagi resonant on 30 meters. This provides a trapless tribander on 20, 30 and 40 meters!

Another possibility involves adding boom extensions for 80-meter operation. Because of size limitations, this would unfortunately involve the use of lossy loading coils.

Performance

As with all antennas, there are some disadvantages to the boom-excited beam. Maximum radiation from this antenna is at right angles to that from the Yagi. There is no front-to-back ratio because the antenna is essentially a dipole, which has little gain at best. On the other side of the ledger, the boom-excited beam *does* have nulls off the ends (as all dipoles do), and they can be pointed at an interfering station. On my antenna, the nulls are about 15 to 20 dB in depth. Because there are no lossy loading coils and the antenna has heavy conductors (the boom) in the area of maximum antenna current, efficiency seems to be very good. The full length of the antenna is at one height, which helps to lower the overall radiation angle. On-the-air tests have yielded good signal reports.

All these techniques are bound to raise questions about boom-exciting other types of antennas. I am sure this is possible, but it will take some experimenting to obtain the answers. If you decide to experiment, I'd be anxious to learn of your results. Perhaps additional findings will be reported in a future issue of *QST*.

If nothing else, after boom-exciting your antenna you can impress your friends by telling them you now have a 4-element, trapless 40-meter beam. Well, that's sort of true. . . .

Notes

[1]mm = in. × 25.4; m = ft × 0.3048.

[2]G. Hall, ed., *The ARRL Antenna Book*, 14th ed. (Newington: ARRL, 1982), Chapter 2, p. 25.

[3]*The ARRL Antenna Book*, Chapter 5.

Strays

INTERCONTINENTAL PACKET RADIO A REALITY

☐ A successful two-way 10-meter packet-radio QSO took place between the U.S. East Coast and New Zealand on May 27, 1983 at 2300 UTC. Tom Clark, W3IWI, in Maryland, and Ian Ashley, ZL1AOX, near Auckland, made this record 13,850-km contact. The Tucson Amateur Packet Radio (TAPR) terminal node controller (TNC) was used at both ends, running at a speed of 1200 baud and using amateur AX.25 link-level protocol. On May 30, Vern Riportella, WA2LQQ, joined W3IWI and carried on a one-hour contact at 600 baud.

THANKS, ICOM

☐ Icom America, Inc., recently donated an IC-251A 2-meter multimode transceiver to the ARRL Technical Department. This unit will be used as an i-f source for developing and evaluating transverters for a book on uhf and microwaves. The ARRL expresses its gratitude for this gift.

AMTOR PHOTOS NEEDED

☐ The ARRL Technical Department is looking for good-quality B & W photographs of AMTOR stations, especially homebuilt ones. Those accepted would be used in *QST* or the *Handbook*, with appropriate photo credit given. Please send the photos to the attention of Paul Rinaldo, W4RI, at ARRL Hq.

A Low-Cost, Modular Approach to RTTY

Here's a helpful review of several approaches to setting up for RTTY operation. Included is a description of an expandable, low-cost system.

By J. Robert Witmer,* W3RW

After operating receive-only RTTY for approximately 18 months with a surplus model 15 teleprinter, the '15 developed a problem: It just "hammered away" at the same spot on the paper. The family wasn't upset at all. More than once they had complained about the noisemaker in the basement!

In trying to determine the cause of the problem, I became convinced of two things: Whoever designed the model 15 must have been an excellent mechanical engineer, and it was time to upgrade my equipment! To simplify matters, I made a list to help me evaluate the various new equipment options available, and decided to use the TU (terminal unit) I had on hand. The following areas were taken into consideration during the evaluation: cost (should be low), noise (had to be family compatible), capability of using multiple speeds, complexity, ASCII operation as well as Baudot, information storage and display-unit features. I forged ahead with these items in mind.

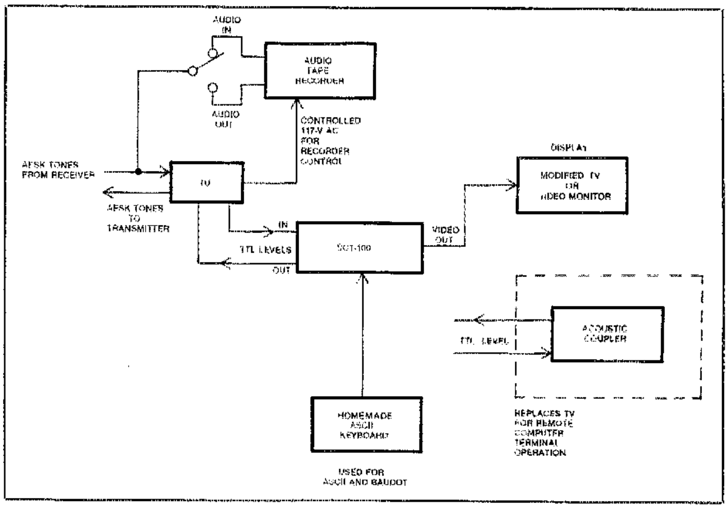


Fig. 1 — Block diagram of the XITEX RTTY station setup.

Surplus TTY Gear

The teleprinter models 15, 19, 28, 33 and some newer units are available in most areas. It's hard to beat this type of gear for cost. In fact, I got my original model 15 (completely reconditioned) for $20! However, these teleprinters, especially the models 15 and 19, are quite noisy. Multiple-speed capability is mechanically possible only if spare gears are on hand and you want to be changing them each time you shift speeds.

From an electrical standpoint, it is hard to get much simpler equipment. Mechanically, it's tough to get much more complex! This may be a trade-off that depends on your personal talents. ASCII/Baudot capabilities are singular. You can have one or the other — not both with the same machine.

By nature, this type of equipment provides a permanent record ("hard" copy on paper and/or punched tape) of transmitted and received information. The ability to see many lines of copy is an advantage. This is almost impossible to duplicate except with another type of printer or with a multiple-page CRT-type terminal (very expensive, usually!).

Dedicated Terminals

New multifunction terminals are appearing all the time. Since the majority of this equipment is new, it usually carries a pretty high price tag. But, the noise is gone! Most operation takes place using a video monitor, a TV set or an LED/LCD readout. One of the biggest advantages of this type of system is the flexibility. ASCII, Baudot and Morse code capabilities are usually inherent or available as options. Permanent message storage requires the use of an external printer (extra cost, noisy) or an audio tape recorder. Transmitted and received messages can usually be recorded on audio tape using the afsk tones.

The display device characteristics vary greatly from unit to unit. Some display only one line or several characters of text at once. Those that use an external TV set are usually limited to a maximum line display of 32 to 40 characters because of TV-set bandwidth limitations. This can be a problem since many operators send 60 to 70 characters per line. If you plan on copying pictures, the shorter line length is a definite disadvantage. If a video monitor is used, the wide bandwidth permits displaying up to 24 or more lines of 80 or more characters each.

Home Computers and Interfaces

These combinations are popular. If you already have a personal computer, this may not be a bad approach, but the interfaces

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and software can be as expensive as the computers! The machine noise is gone, but where the family is concerned, you may still have a problem — you'll have to fight for the use of the computer!

Most interface/computer combinations allow the use of multiple speeds as well as ASCII, Baudot and Morse transmission and reception. ASCII, after all, is the "natural language" of the computer (the ZX81/Timex 1000 being a notable exception)! The message storage and display features are similar to those offered by the dedicated terminals.

Another Approach

After reviewing the options, I wasn't happy with the overall results. The two-part article by Greg McIntire, AA5C, in January and February 1982 *QST*, reminded me of an ad I had seen in *QST*.[1] After digging through back issues, I found it — the advertisement was for a single-card video terminal, manufactured by XITEX Corporation, the SCT-100.[2]

The SCT-100 is designed to operate as an ASCII/Baudot terminal with two speeds for each mode: 110- and 300-baud ASCII, and 60- and 100-wpm Baudot. The display provides for sixteen 64-character lines.

There are three SCT-100 versions: an assembled unit, a kit and a semi-kit. With the latter, you supply some commonly available parts. The manual, while not Heathkit® style, is excellent and provides detailed interfacing and operational information. Since this is a type of dedicated terminal, the noise is nonexistent. I've found having a separate (from the home computer) unit to be a big plus. With the SCT-100, I use an audio tape recorder to store the information. The 64-character line is compatible with most RTTY operation. A summary of the evaluation is shown in Table 1.

Table 1

Evaluations Summary

Surplus Teleprinters

Advantages: Low cost, permanent storage capability, good readout.
Disadvantages: Mechanically complex, single speed, single mode, noisy.
Comments: May be the best choice for single-mode operation.

Dedicated Terminals

Advantages: Speed/mode flexibility.
Disadvantages: High cost, variable readout parameters, lack of permanent storage capabilities in some units.

Personal Computer Adapter

Advantages: Speed/mode flexibility.
Disadvantages: Relative high cost, requires use of personal computer, variable readout parameters, lack of permanent storage capabilities with some units.

XITEX System

Advantages: Relatively low cost, speed/mode flexibility, reasonable display.
Disadvantages: Lack of permanent storage (can be overcome by use of a tape recorder).

Station Description

How I integrated the SCT-100 into my station can be seen in Fig. 1. I use the TU described in the December 1980 and November 1981 issues of *QST*.[3,4] It is TTL-compatible and interfaces nicely with the '100. The autostart relay contacts are used to switch the 117-V ac supply to the audio tape recorder. This allows me to keep a record of TTY activity on a local 2-meter fm repeater when I'm out of the shack.

The SCT-100 board has provisions for 20-mA, four-wire full duplex, 60-mA two-wire simplex, modified RS-232C and TTL level I/O. I use the TTL level I/O for interfacing with the TU. For computer/telephone accoustic coupler use, the modified RS-232 output is employed. The '100 has an onboard 5-V power supply that requires only 8-V ac for operation. A 5-V transformer in series with one-half of a 6.3-V transformer secondary provides this potential. For powering an ASCII keyboard, 5 V at up to 250 mA is available at the I/O connector. Selection of ASCII/Baudot operation and speed is made by opening or grounding two points on the I/O connector by means of a switch.

Both composite video and discrete vertical and horizontal sync outputs are available to simplify the interfacing. The composite video output of the SCT-100 complies with EIA standard RS-170. It consists of a 1.5-V peak video signal and a 0.5-V peak sync pulse. This output drives a surplus video monitor. A modified TV set could also be used.[5,6]

It is possible to interface the SCT-100 with an unmodified TV set by using an rf modulator and feeding the rf to the TV antenna input. This is not recommended since the bandwidth required for the 64 × 16 display (64 characters per line, 16 lines displayed) is somewhat beyond that of most TVs. While this will not damage the TV set, the resulting picture may be distorted and unusable.

I use a "homebrewed" keyboard that provides the required 7-bit ASCII, TTL-compatible, positive-true logic signal with a strobe signal. One feature of the SCT-100 is that the LTRS and FIGS shift characters are generated automatically when a respective lower- or upper-case entry is detected.

Acquiring the Equipment

One advantage of this versatile system is that it can be expanded as your budget permits. A suggested first step would be to acquire the semi-kit. After all the other parts are gathered, the SCT-100 can be assembled and used for receive-only operation with a modified TV set. Next, add a keyboard for full transceive operation. A printer and a tape recorder could subsequently be added if desired.

These suggestions should help you decide on how to assemble your Baudot/ASCII RTTY station. Come on and join in on the RTTY fun!

Notes

[1]G. McIntire, "Designing a Microprocessor-Based RTTY Speed and Code Converter," *QST*, Jan. and Feb. 1982.
[2]XITEX Corporation, P.O. Box 2952, Garland, TX 75041.
[3]M. DiJulio, "A State-of-the-Art Terminal Unit for RTTY," *QST*, Dec. 1980, p. 20.
[4]J. Witmer, "Auto-start and Anti-space for the State-of-the-Art TU," *QST*, Nov. 1981, p. 28.
[5]S. Bach, "Converting a Bargain TV to a Video Monitor," *Kilobaud Microcomputing*, Jan. 1980.
[6]T. Loos, "Use Your Television Set as a Video Monitor," *BYTE*, Feb. 1979, p. 46.

Strays

QEX: THE ARRL EXPERIMENTERS' EXCHANGE

☐ The July issue of *QEX* featured:
• "Continuous RTTY Reception on the ZX80," by Kenneth Heitner, WB4AKK. This full-length feature article includes full program listings and interface circuitry.
• "Data Communications," by Dave Borden, K8MMO, a discussion of packet-radio network design issues. Also included is a report on U.S.-to-New Zealand packet-radio contacts between ZL1AOX and W3IWI, and subsequently with WA2LQQ.
• In an editorial, Steve Place, WB1EYE, describes the successful launch of AMSAT-OSCAR 10 and gives initial orbital elements.

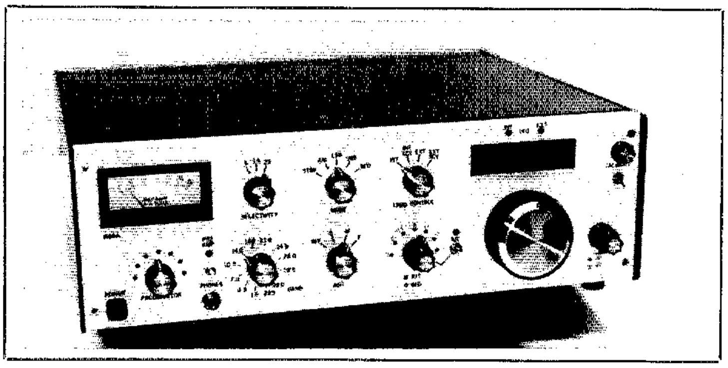
WANTED: ELMER-OF-THE-YEAR NOMINEES

☐ Nominations for the 1983 Elmer of the Year award are being sought by the Northern New Jersey Chapter of the QCWA. Nominations may be made by any licensed radio amateur in northern New Jersey, and eligible candidates must be licensed amateurs who reside in the area. Send your nomination, which should be accompanied by a statement (500 words or less) detailing why the person is worthy of the award, to Gordon S. Gregory, N2IN, 8 Winding Way, Denville, NJ 07834, tel. 201-627-4426. September 1, 1983 is the deadline for receipt of nominations.

I would like to get in touch with...

☐ anyone who has any information on the "Morris Coil Winder." Robert Morris, WB6MUM, 116 South F St., Lompoc, CA 93436.

A Structured Engineering Approach to the Design and Construction of Electronic Equipment



Want to increase the chances of success of your next construction project? Then follow these guidelines!

By Jerry L. Pittenger,* K8RA

Some hams believe that "homebrew" construction and experimentation are activities of the past. You hear about the ever-increasing price (and decreasing availability) of electronic components, and about the proliferation of "appliance operators." Also, many amateurs seem overwhelmed by the fast-moving pace of electronic technology. All this has convinced many would-be builders that constructing a quality station component such as a receiver or a transmitter is beyond their abilities.

Yet, the amateur journals are packed with technical information and construction articles. Hamfests sport flea markets with literally millions of electronic components at reasonable prices. This is proof that more than just a few amateurs are actively building and experimenting with electronic equipment. "Homebrewing," these hams have found, gives a feeling of pride that builds confidence and even a sense of self-respect.

You don't have to be an electronics engineer to build high-grade equipment. The real ingredients are the courage to get started, use of a well-structured approach to accomplish the task and the perseverance to complete the project.

Most hams don't start with a box of parts and quickly produce a quality product. A first-class job involves research, planning, more planning, design and plenty of testing. No miracle design and construction techniques exist that eliminate errors and problems. However, the approach presented here can help circumvent many of the problems that the inexperienced — or seasoned — builder is likely to confront.

The proposed approach is adapted from state-of-the-art Computer Application Engineering (CAE) principles to which I have been introduced in my profession. This technique provides a structured, well-organized approach to bringing a complex construction project to a level that can be understood easily. I have used these principles to construct many projects ranging from keyers, linear amplifiers and antenna-matching networks to a respectable solid-state receiver. The results of some of these projects are shown in the accompanying photographs.

The steps involved in using this technique include

1) Define the system.
2) Design the system.
3) Plan and document testing.
4) Implement system components.
5) Test the system.
6) Test for acceptance.
7) Provide user education.

The system is interpreted as the electronic unit being constructed.

To illustrate the functions of each of the steps, the solid-state receiver shown in the accompanying photos is used as an example. It is an amateur-band (including the WARC frequencies), solid-state design that uses diode switching throughout. Emphasis is placed on sensitivity, selectivity and high dynamic range.

Define the System

First, define the requirements of the unit. In CAE terminology, this is referred to as "developing a functional specification." Many potentially good projects have gone astray because the builder didn't think clearly about the required features and document them, to make the unit acceptable. The result is usually a device that almost meets requirements, but with just enough missing to justify pitching the unit aside later on.

I spent more than three months researching the literature before designing this receiver. A more experienced person may be able to shorten the effort. Because of my inexperience in receiver design, however, I read all I could about receivers to make sure the requirements I defined were technically current and achievable. Reviewing the alternative operational features for any electronic project provides a shopping list of features from which to choose. This results in a unit that best meets an individual's needs.

The options must then be evaluated, based on added capabilities, cost, complex-

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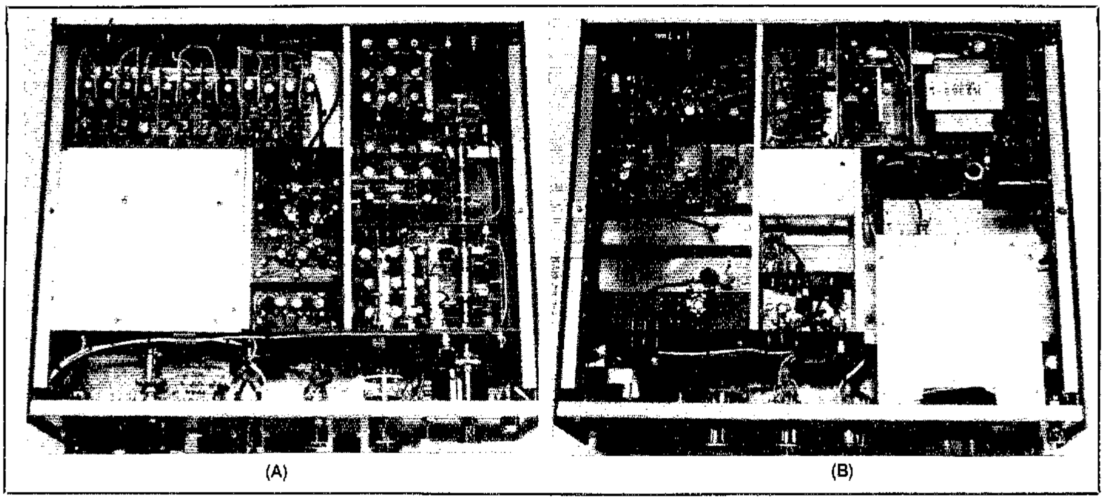


Bottom (A) and top (B) views of the homemade receiver.

ity, size, parts availability and any other constraints unique to a given option. Table 1 summarizes the specifications of the receiver. They emphasize "what" the unit should do, not "how" it will do it.

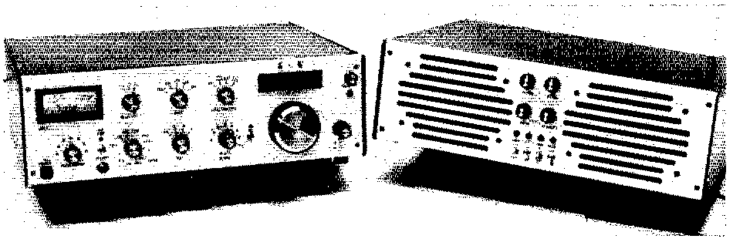
Design the System

Functional Design

This step translates the "what" into "how" it will be done. First, divide the unit into pieces. As in CAE, the division is carried out in steps: Divide the unit into major components, divide each major component into sections, and then subdivide these sections until they are small enough to be designed easily in detail. Each section should be small enough to be independent from other sections while serving a unique, identifiable function. I like to use the block-diagram approach for this process.

The method is best illustrated by an example. Fig. 1 shows the top-level block diagram for the receiver. This diagram is basic, but it provides a starting point. Then, each major component is broken down to the next level of detail. Fig. 2 shows the next level of refinement for the frequency-control module. Note that this module has been separated into several submodules. Fig. 3 shows the breakdown of the VFO module of Fig. 2. Similar diagrams were created for each module of Fig. 2. The refinement should go to the point where a set of simple submodules is defined, each performing an independent function. My recommendation is to separate a module into simple sections that you understand completely, and not go to extremes.

The bottom level of detail produces a set of lowest-level submodules that are referred to as the system "primitives." In Fig. 3, the VFO, buffer and amplifier are all primitives of the VFO submodule. Now, design the circuit for each primitive. This step may be nothing more than selecting a circuit from the available literature. On the other hand, you may develop your own circuit — something better than that which currently exists.



The author's homemade receiver and speaker/filter unit. Notch, width and filter frequency controls are included on the latter.

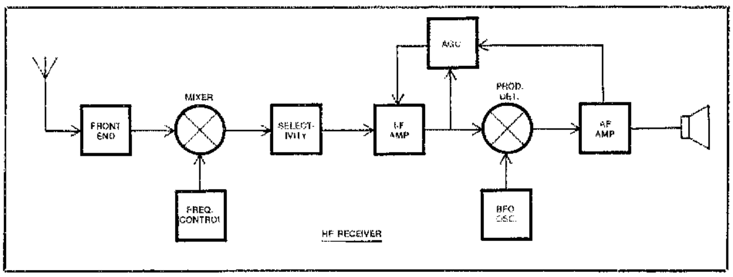


Fig. 1 — Block diagram of the receiver.

The simple circuits of the primitives combine to make the complete circuit of the project. Believe me, there are lots of parts in the example receiver! The complete schematic diagram is complex and would be nearly impossible to use in its entirety during construction of the receiver. When you look at the project as a simple set of much smaller subunits, however, complex projects become understandable and achievable.

Physical Design

You must also consider the physical design of the unit. Thought should be given to the physical appearance of the front and rear panels, as well as to how the modules fit together inside the unit. Few people are satisfied with an end result that works okay but looks like a piece of junk.

The original receiver front-panel design differed slightly from that of the final product. I mention this only to illustrate that

Table 1
Receiver Specifications

Frequency coverage:	28-30, 21-21.5, 18-18.5, 14-14.5, 10-10.5, 7-7.5, 3.5-4, 1.5-2 MHz.
Sensitivity:	0.5 μV for 10 dB S/S+N.
Selectivity:	Three optional filters, 2:1 shape factor at 6:60 dB points.
Dynamic range:	95 dB.
Noise floor:	−135 dBm.
Third-order input intercept point:	+5 dBm.
Frequency stability:	500 Hz from cold start to one hour later.
BFO:	Crystal-controlled on ssb; variable on cw.
Agc:	I-f derived, hang type.
Audio output power:	5-W minimum.
S meter:	Accurate within 20% with 6-dB/S unit calibration.
Transmitter interface:	Allow for mating with transmitter VFO control and split-frequency operation.
Power supply:	Internal.
Physical Specifications	
Appearance:	Compatible with existing station equipment. Cabinet dimensions (HWD): 5 × 17 × 14 in.[†]
Construction:	Modular, to allow for future experimentation.

[†]mm = in. × 25.4.

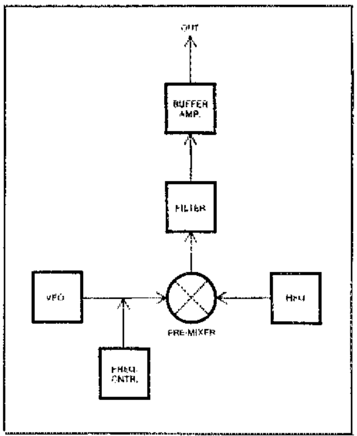


Fig. 2 — Breakdown of the frequency control module — the first step in simplifying the design approach.

the initial design is not sacred. The design is an iterative process throughout the life of the project. When it is advantageous or necessary to modify the design, do so. But always consider all the impacts of changes, and update the design documentation whenever changes are made.

Implementation Plan

In what order will the primitive submodules be built and tested? Try to capitalize on any advantageous sequence that allows easy module integration and testing. The plan used for the receiver is given in Table 2. Note that the sequence proceeds in reverse order from the speaker. After each submodule is constructed and tested independently, it is integrated with the other completed modules.

Design Documentation

All of the design data should be assembled to form a detailed design document. I usually assemble this information in a three-ring notebook. Take time to develop clear, precise diagrams, schematics and textual documentation, outlining why the final primitive designs were selected. Take careful notes and document literature references for future use. The design document provides guidance in the building phase, but it is even more valuable when something fails later on. One final recommendation: Keep the documentation current. When you get into the build/test phases of the project, it can be tempting to go forward and not take the time to change the documentation neatly as design changes are made. Avoid the pitfall!

This probably looks like a lot of extra work. I think this approach saves time. I am convinced that doing the detailed design on paper (where changes are easy to make) is worth the time invested. It avoids confronting such problems as not enough room inside the cabinet halfway into construction, trying to find room on the front panel for that extra switch you forgot, or mistakenly drilling holes in the front panel.

Table 2
Receiver Implementation Plan

1) Power Supply
2) Audio amplifier
3) BFO crystal oscillator
4) Variable BFO
5) Product-detector module
6) Post i-f crystal filter module
7) I-f amplifier
8) Agc module
9) Crystal filter
10) VFO
11) Frequency counter
12) HFO
13) VFO/HFO premixer
14) Premixer filter
15) Premixer buffer amplifier
16) Front-end mixer
17) Front-end tuned filters
18) Front-end preamplifier
19) Crystal calibrator

Plan and Document Testing

Careful consideration should be given to how the modules and completed unit will be tested. Writing a good test plan also provides the best possible design review. In designing and constructing electronic equipment, three levels of testing are recommended: module, integration and system.

The test plans are organized and documented prior to the building step. Two major pieces of documentation are required in any good test plan: The test-execution procedure and a list of the expected results. Writing good test plans is a difficult task, since it requires a designer to foresee problems.

Module Test Plan

A module is the lowest level at which testing is performed. The size and the number of modules are determined by the level of division performed in the system design step. Examples of modules in the receiver include the audio amplifier, the product detector, the i-f strip and the agc circuit. Consider the test plan (Fig. 4) developed for the product-detector module. Note that it includes the test setup, procedure and expected results. A plan similar to the example should be developed for each module identified in the Implementation Plan (Table 2).

Integration Test Plan

The second level of testing occurs during module integration. The tests should be defined in accordance with the Implementation Plan (Table 2) developed in the System Design phase. The objective of integration testing is to assemble one or more modules and test the assembly to ensure the modules work correctly together. (Remember, each module has been tested individually.)

The Integration Test Plan for the receiver is given in Table 3. RX MODULES lists all the receiver modules. PLAN defines where testing is to occur. For example, step 2 plans an integration test of the speaker, the audio power amplifier, the BFO crystal oscillator, the product detector and power supply modules. Note that these modules are completed in sequence for this test as defined in the Implementation Plan. For each integration test, a test procedure is developed. Fig. 5 outlines the test plan for step 2 of Table 3. Note that this includes a test-setup description, instructions on

Table 3

Receiver Integration Test Plan

RX Modules		*Plan*	
A)	Speaker	1)	A+B+O
B)	Audio power amplifier	2)	A+B+C+E+O
C)	BFO crystal oscillator	3)	A+B+D+E+O
D)	BFO variable oscillator	4)	A+B+D+C+E+O
E)	Product detector	5)	F+G+O
F)	I-f amplifier	6)	A+B+C+D+E+F+G+O
G)	Agc	7)	A+B+C+D+E+F+G+H+O
H)	Crystal filter	8)	I+J+K+O
I)	HFO crystal oscillator	9	I+J+K+L+O
J)	Premixer	10	A+B+C+D+E+F+G+H+I+J+K+L+O
K)	VFO	11)	A+B+C+D+E+F+G+H+I+J+K+L+M+O
L)	Front-end mixer	12)	All — proceed to System Test
M)	Front-end filters		
N)	Crystal calibrator		
O)	Power supply		

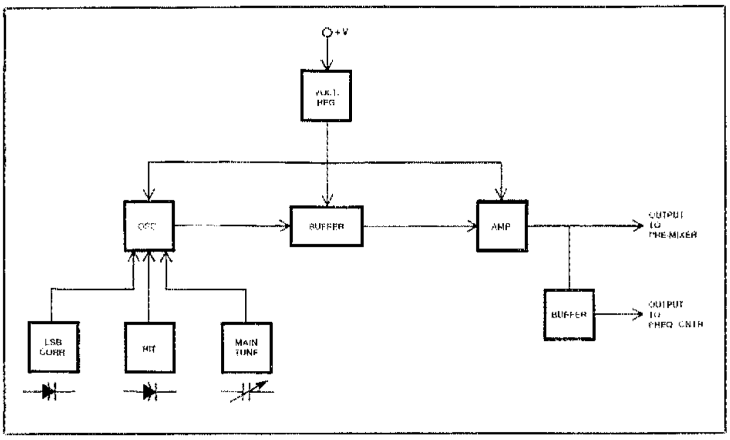


Fig. 3 — Taking the VFO module of Fig. 2 down to the "primitive" level.

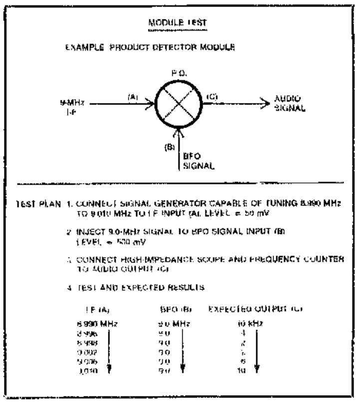


Fig. 4 — Module test plan for the product detector.

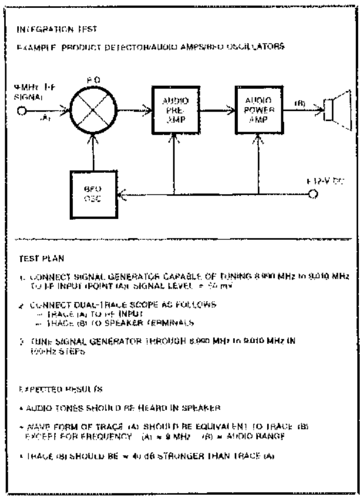


Fig. 5 — The Integration Test plan. At this stage, a number of modules are interconnected and tested as a unit.

how to perform the test and the expected results.

System Test Plan

Once the modules have been integrated, the last test recommended is a System Test. This is performed to verify that the complete unit meets all peformance requirements defined in the System Specification (Table 1). A procedure should be formulated for each specification. Fig. 6 defines a test for the receiver noise floor.[1]

Remember, test plans don't have to be complex. A frequency-stability test for the receiver was easy to do because the included 10-MHz band provided WWV as a frequency reference. Do what is necessary with the minimum effort to evaluate total system performance.

Test Plan Comment

Development of module, integration and system-test plans may seem tedious. Experience has shown that the time is well spent and provides a way to avoid problems early in the project when they are easiest to find and fix. Since all test results are documented, if problems occur later or if the unit is modified, the tests can be repeated. The results can be compared to those of earlier tests to see if the problem is cured, or if the modifications have caused any detrimental effects.

System Components Implementation

This step is often referred to as the building step. We are finally at the point of assembling and testing the modules. By now, approximately half the project time should have been expended. Unfortunately, it is usually the starting point for those who do not use a structured engineering approach. This lack leads to false starts, construction errors and total project disasters.

Implementation should be a noncreative step, all major decisions having already been made. The modules are constructed in the defined sequence (i.e., Table 2). Each module should be tested as it is completed.

I offer no construction advice except to observe neatness. I used glass-epoxy board with silver termination pins in a point-to-point wiring technique. Other construction methods might include perf-board, terminal strips or etched board. Whatever technique you feel most comfortable with can be used.

The following advice is based on many hours of experience with some successes and some failures:

- Take your time — don't rush!
- Follow your planned design. You spent a lot of time planning the project, so use the product of your efforts.
- Use only quality components. Don't take shortcuts.
- Don't settle for "almost working." Several small problems add up to a big problem during system integration.
- Expect problems. Redesign when necessary.

[1]*The Radio Amateur's Handbook*, 60th ed. (Newington: ARRL, 1983), p. 16-41.

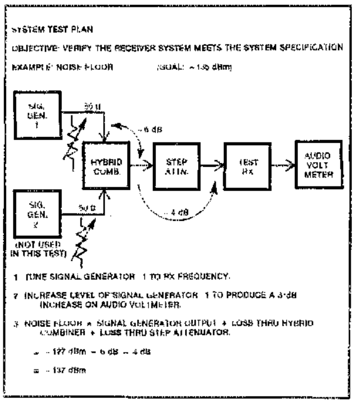


Fig. 6 — The System Test plan for the receiver. In this case, the receiver noise floor is being measured.

- Document any design changes.
- Keep organized notes.

As modules are completed, integrate them accordingly. The test results should be documented and included with the test plans for future reference. Temptations to move fast, skip testing and slight documentation are strong at this point. Don't succumb! Do the job methodically with all the diligence and attention to detail that you are capable of, and that the project deserves. The results will be rewarding.

System Test

The purpose of this test is to verify that the system (in this case, the electronic device) meets the operational and physical specifications defined in the System Specification (Table 1). With a proper design approach and the extensive module and integration testing performed, I've discovered few major problems during system testing. When changes are needed, carefully design the changes, implement them and update all documentation. Changes in the system design may require iteration of some integration tests to ensure that the overall system performance has not been disturbed. The module and integration test results provide a reference point to evaluate the performance gains or degradations resulting from changes.

Acceptance Testing/User Training

Acceptance testing is usually performed upon delivery of a unit to a customer. The test is executed according to a plan developed by the customer. Because in our case the builder is also the user, the system test also serves as the acceptance test.

User Training is not required. The idea of a person not knowing how to use the device after defining, designing, building and testing the unit is not a realistic consideration.

Some Final Comments

The CAE techniques appear to add more work to a given project. But I feel strongly that using such an approach will save time by avoiding the problems and pitfalls so often encountered. In fact, the use of this or some similar method could make the difference between achieving complete success or total failure with a construction project. I make this claim based on personal experiences, and from seeing the problems and results others have had.

I encourage you to try the approach, in its entirety or partially. I will be interested in hearing about possible improvements to the approach and any experiences you may have in constructing your electronic equipment. QST

Strays

INTERESTED IN LEAGUE-SPONSORED INSURANCE?

☐ The Membership Affairs Committee is studying the feasibility of offering League members additional insurance programs beyond the present ARRL Ham Radio Equipment and Club Liability programs. Participation in the plans would be voluntary. Please take a few moments to complete the following questionnaire and send it to the Membership Services Department, ARRL, 225 Main St., Newington, CT 06111. Make a photocopy if you'd like, or simply answer the questions on a separate piece of paper. Thanks.

I would be interested in the following types of voluntary insurance plans, should the ARRL offer them:

	I am interested	I am not interested
1) Life Insurance	☐	☐
2) In-Hospital Insurance (to pay a fixed amount each day during hospitalization)	☐	☐
3) High-Limit Accident Insurance (to pay specified amounts for specific levels of injury due to accident)	☐	☐
4) Disability Income Protection	☐	☐
5) Excess Major Medical (to pay medical charges above a deductible of $15,000 to $25,000)	☐	☐
6) Major Medical (to pay medical expenses above a deductible of $500 to $1000 but with a maximum benefit of $20,000)	☐	☐

I do not believe the ARRL should sponsor insurance programs that have no direct relation to Amateur Radio. ☐

Please note: An expression of interest in no way binds you to a particular course of action. This survey is designed only to gauge membership interest in these insurance programs.

MAKING PATTERNS AND TEMPLATES TO SCALE

☐ Starting with this issue, all pc-board patterns and other templates requiring full-size reproduction that appear in *QST* are accompanied by a 1-inch (25.4-mm) scale. This will enable everyone concerned with the pattern to ensure it has been reproduced full size; simply place a ruler alongside the scale and compare the 1-inch markings. — *Paul K. Pagel, N1FB*

THANKS, KLM

☐ KLM Electronics, Inc., recently donated 143-150-14C and 420-450-18C 2-meter and 70-cm circularly polarized antennas to the ARRL Technical Department. These antennas will be used for satellite communications. The ARRL expresses its gratitude for this gift.

Last year, the Conde brothers, of Granger, Wyoming, came up with a unique Father's Day gift: As a surprise for their Dad, WA7USI, they prepared for and passed their Novice exams. Left to right are Wally, KA7OMP, Andy, KA7OGQ, and "CQ," KA7OGR. *(W1YL photo)*

New and Improved Formulas for the Design of Pi and Pi-L Networks

Published equations relating to the design of pi networks are often inaccurate where circuit operating Q is concerned. If Q is of primary consideration, use the equations in this article.

By Elmer A. Wingfield,* W5FD

This article introduces new formulas for the design of pi networks and pi-L networks. These new formulas permit the network to be designed based on the resistances to be matched and on the actual circuit operating Q value, Qo. Design formulas now in common use have only a partial circuit Q value as the circuit Qo. In many practical design cases the error involved in the usual formulas can be quite large and the resulting network will not perform as intended.

The pi networks, the pi-L networks and the resonant-L networks considered in this article are only those that are configured in the low-pass arrangement with the inductance as the series element and capacitors as the shunt elements, as shown in Fig. 1. This is the arrangement in which these circuits are invariably used in amateur rf applications. These networks, consisting of two, three or four reactance elements, have only one absolute requirement: At least one of the reactances must be of opposite sign to the others for a "match" or transformation of a load-end R2 to a desired R1 value at the source end. This is because resonance is impossible otherwise. A low-pass configuration is not required for a match, but it is the only arrangement of any importance in rf uses of such networks and therefore the only arrangement considered in this article.

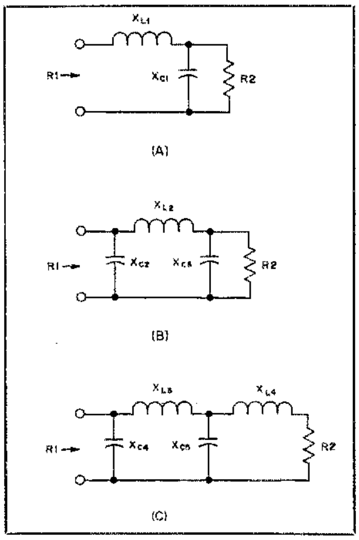


Fig. 1 — An L network is shown at A, a pi network is at B, and a pi-L network is at C. These networks are shown in the low-pass configuration. Each network matches the load, R2, to the input resistance, R1.

R1 > R2

$$N > \sqrt{\frac{R1}{R2} - 1}$$

$$X_{C2} = \frac{R1}{N}$$

$$X_{C3} = \frac{R2}{\sqrt{\frac{R2}{R1}(1 + N^2) - 1}}$$

$$X_{L2} = R1\,\frac{N + \frac{R2}{X_{C3}}}{N^2 + 1}$$

Fig. 2 — The "old standard" equations for pi-network design. These equations are those appearing on page 54 of *The Radio Amateur's Handbook*, 55th (1978) edition, changed here for agreement with the reactance designations shown in Fig. 1. In these equations, N is the same as Q1 in the article.

The "Old Standard" Equations

The old standard pi-network design formulas are those familiar equations that have been published in the ARRL *Handbook* for many years, in ARRL *Solid State Design*, the *ARRL Electronics Data Book*, the *Radio Handbook*, and Motorola *Application Note AN-267*. They appear in Fig. 2, and apparently were first published in the amateur literature by Pappenfus, WØSYF, and Klippel, WØSQO, in 1950.[1] The tabulated pi and pi-L network data given in current editions of the ARRL *Handbook* are based on these formulas, as is similar data in the *Radio Handbook*, 21st edition.

These old standard formulas are derived by treating the pi network as two back-to-back L networks. (The derivation is given in the appendix.) These formulas give correct results for the network reactance values for matching an R1 source end to the R2 load-end resistance. The only error involved is in assuming that the Q1 source-end L-network Q value used in the formulas is the resultant total pi-network Qo value. It is not, and understates the Qo value by the amount of the Q value of the load-end L network section, Q2.

In terms of the assumed formula Q1 value and the R1 and R2 to be matched,

$$Q2 = \sqrt{\frac{R2}{R1}(Q1^2 + 1) - 1} \qquad \text{(Eq. 1)}$$

This is the error involved in taking Q1 as the Qo value, since Qo = Q1 + Q2. The error becomes progressively greater as the ratio R1 to R2 becomes lower. For R1/R2 = 1, the error is 100%, as Qo equals

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[1]Notes appear on page 29.

2 × Q1. The error is greater for R1 × R2, a common arrangement in matching the required R1 load resistance of rf transistor amplifiers to the usual 50-ohm R2 load.

If the R1-to-R2 ratio is moderately high, say 20 to 1 or higher, then the error in taking Q1 as the circuit Qo will be relatively small. Therefore the old standard formulas may be used for the design of pi-network output tuning circuits for the usual tube-type rf amplifiers. For example, for a pi-network design of R1 = 2500 ohms, R2 = 50 ohms and Q1 = 12, the actual Qo will be 13.38 instead of 12, and this error is unimportant. This design would be representative for a pi network for rf linear amplifiers operated at the 2-kW peak level at a plate voltage of 3000 V or so, as is common for current amateur-band linear amplifiers. About the worst case for an amateur tube-type design would be four sweep tubes in parallel requiring a plate-load resistance on the order of 400 ohms. Then, with a formula Q1 entry of 12 and a 50-ohm load, the actual network Qo would be 16.14 — too high, perhaps, but still workable.

The conclusion is that the old standard pi-network formulas provide satisfactory pi-network output tank coupling circuit designs for amateur tube-type rf amplifiers and have done so for more than 30 years. The Q1 value used for Q differs only moderately from the correct circuit Qo.

Unsatisfactory Q Results

The old standard formulas do not give satisfactory designs for Q-based pi-network circuits intended for the fixed-tuned input circuits of cathode driven (grounded-grid) rf amplifiers. The rf exciter normally requires a 50-ohm load, and the fixed-tuned pi-network circuit must match this R1 source end to the rf-amplifier cathode-driving impedance at the R2 load end. At a minimum this load is perhaps 50 ohms, and it may be as high as 100 to 150 ohms or so for some tube types and operating conditions. For the tubes currently popular for the amateur services — zero-bias triodes operated as grounded-grid Class B linear amplifiers — the cathode driving impedance will be in the range of 50 to 75 ohms when operated at the 2-kW-PEP level.

Since the input tuned circuit is fixed tuned to obtain a broadband response, the Q of the circuit is kept at the lowest level that will just provide an adequate "flywheel" effect. In practice, a Q of 2 to 3 is used. If the network is designed using the old standard formulas for the "best" case (where R1 = 50 ohms, R2 = 50 ohms and a Q of 3 is chosen), an actual circuit Qo of 6 will result. The error is 100%. The error will increase as R2 is larger than 50 ohms, as it usually is. Because of the higher than intended (or needed) Qo, the input circuit will not be sufficiently broadbanded, and solid-state broadband exciters may not be able to drive the amplifier to the full intended power level because of SWR turndown in the exciter. This might be the case especially on the relatively wide 75- and 80-meter band. These fixed-tuned-input pi networks should be designed using the formulas given later in this article.

As an example of the very large errors that can occur in the Qo value when pi networks having low R1/R2 ratios are computed by the old standard formulas, the first data entry in Motorola *AN-267* indicates a Q value of 1 for the network that matches an R1 of 1 ohm to the R2 load of 50 ohms.[2] The actual Qo for this network is 10.95, almost 11 times the stated Q value.

Pi-L Networks

The standard formulas are not suited for the design of pi-L networks in which the circuit Qo is an important factor, as it is in tube-type and other rf amplifiers. The reason for this is that in the pi-L the output load resistance, R2, is stepped up to an intermediate value, Rm. Rm becomes the load-end resistance value for the input-end pi-network section. Then, the ratio of R1 (source end) to Rm (load end) for the pi-network section is quite a bit lower than the R1-to-R2 ratio in a straight pi-network tank. The Qo of the pi- network input section of a pi-L will therefore have an added increment of Q caused by the lower ratio R1 to Rm. In addition, there is an added increment of Q from the output-end L network. The resulting error in assuming Q1 as the network Qo is large.

As an example of the large error involved, the pi-L tabulated data in Table 10 on page 6-31 of the ARRL *Handbook* (1979 and later editions) indicates a circuit Qo of 12 for the network data given. For that data, the true circuit Qo for the high end of the band varies from 19.47 for Z_{in} = 1500 ohms to 16.29 at Z_{in} = 8000 ohms. The errors are too large, and the networks given will have unnecessary additional losses because of the higher-than-optimum Qo of 12. In practice, the pi-L output tuning network would be expected to accommodate the line input variations for a line having an SWR of 2, at least. Tuning the network to match actual impedances that may be encountered could result in the pi-L network Qo value going well above 20. The old standard formulas should not be used for the design of the pi-network section of a pi-L tank network circuit.

Two sets of formulas and procedures are given later for the design of pi networks and pi-L networks. These procedures are based on the R1 and R2 to be matched and on the desired circuit operating Qo value. Both of the formula sets are simple algebraic formulas similar to the old standard formulas. They are equally simple and straightforward in use, but with the decided advantage of giving precisely correct results for all quantities.

It was noted earlier that the old standard Q-based pi-network formulas would provide satisfactory design for a pi-network output coupling circuit for most amateur tube-type rf amplifiers. Only a small-to-moderate error in the actual circuit Qo value will result. Since the new formulas given later are equally simple and straightforward, and since they give exact results for all quantities for any possible Q-based pi network, it is recommended that they be used for the design of all pi networks in which Q is a factor.

Gibson, G2BUP, published an article pointing out the Qo error in the old standard formulas and giving a graphical procedure for a correct pi-network solution.[3] This material was later reprinted in another publication.[4] In private correspondence in 1978, Gibson clearly demonstrated that the Q1 or input L-net-section Q value used in the pi-network formulas did not account for all of the stored energy or reactive power associated with the output capacitor, C2, or its equivalent capacitive reactance, X_{C2}. The actual circuit Q is given by

$$Qo = Q1 + Q2 \qquad \text{(Eq. 2)}$$

or

$$Qo = R1/X_{C1} + R2/X_{C2} = X_L/Rs \qquad \text{(Eq. 3)}$$

where

Rs is the series equivalent resistance value of the parallel-to-series conversion of the R2 and X_{C2} components.

The standard pi-network formula Q error was also noted by Whyman, W2HB, and by Kajii, JA1FG.[5,6] Kajii gave a graphical design procedure to obtain a pi-network solution based on the correct value of the circuit Qo. The graphical design procedures have the limited accuracy and the limited range of variables inherent in charts and graphs. The algebraic formula methods do not have these limitations.

The algebraic formula procedures given below as Procedure 1 and Procedure 2 are intended so the user can simply plug in the network R1, R2, and Qo values and crank out the network reactance values. Use of Procedure 1 or Procedure 2 is at the choice of the user. The formulas and procedures are simple to use, but close attention must be paid to circuit-element designation including subscripts and superscripts. This is particularly so in the pi-L-network solution, which has the added L network to account for.

To avoid the clutter involved in adding too much detail in the procedures, some preliminary background basic ac circuit material is shown in the appendix. This includes the series-to-parallel and parallel-to-series conversion formulas. An addition to these results provides a resonant circuit, including a resonant L network, which is the basis for deriving the old standard back-to-back L-net-derived pi-network formulas that are given. The fundamental energy-related Q is shown for the circuits for circuit components involved.

Of the two new procedures and formula sets given here, either will yield precisely correct design of pi networks and pi-L networks as used in amateur practice. This first set is based on an adaptation of the standard Q-based back-to-back L-network derived formulas. The second set is based on an adaptation of the non-Q-based general matching pi-network equations given by Everitt.[7]

Procedure 1, Pi Network

1) Refer to Fig. 3 and select the desired Qo value. Qo must be selected to satisfy the following:

$$Qo^2 > \frac{R1}{R2} - 1 \text{ and } Qo^2 > \frac{R2}{R1} - 1$$

2) Compute Q1.

$$Q1 = \frac{R1 \cdot Qo - \sqrt{R1 \cdot R2 \cdot Qo^2 - (R1 - R2)^2}}{R1 - R2} \quad \text{(Eq. 4)}$$

3) Then,

$$Q2 = Qo - Q1 \quad \text{(Eq. 5)}$$
$$X_{C1} = R1/Q1 \quad \text{(Eq. 6)}$$
$$X_{C2} = R2/Q2 \quad \text{(Eq. 7)}$$

$$X_L = \frac{R1 \cdot Qo}{Q1^2 + 1} \quad \text{(Eq. 8)}$$

4) For the special case where R1 = R2, select a value of Qo greater than zero. Then

$$X_{C1} = 2 \cdot R1/Qo \quad \text{(Eq. 9)}$$
$$X_{C2} = 2 \cdot R2/Qo \quad \text{(Eq. 10)}$$

$$X_L = \frac{R1 \cdot Qo}{\frac{Q_o^2}{4} + 1} \quad \text{(Eq. 11)}$$

Note: Eqs. 9 through 11 are valid only for the special case where R1 = R2.

Procedure 2, Pi Network

1) Refer to Fig. 3 and select the desired Qo value to satisfy the following:

$$Qo^2 > \frac{R1}{R2} - 1 \text{ and } Qo^2 > \frac{R2}{R1} - 1$$

2) Compute X_L.

$$X_L = \frac{Qo(R1 + R2) + 2\sqrt{R1 \cdot R2(Qo^2+4) - (R1+R2)^2}}{Qo^2 + 4} \quad \text{(Eq. 12)}$$

3) Compute Q1 and Q2.

$$Q1 = \sqrt{\frac{Qo \cdot R2}{X_L}} - 1 \quad \text{(Eq. 13)}$$

Q2 = Qo − Q1, or

$$Q2 = \sqrt{\frac{Qo \cdot R2}{X_L}} - 1 \quad \text{(Eq. 14)}$$

4) Then,

$$X_{C1} = R1/Q1 \quad \text{(Eq. 15)}$$

$$X_{C2} = R2/Q2 \quad \text{(Eq. 16)}$$

Note: Eqs. 13 and 14 for determining Q1 and Q2 above are valid only for Procedure 2, using X_L from Eq. 12 and the variable values Qo, R1 and R2 used in Eq. 12.

The Pi-L Network

For the following discussion, refer to Fig. 4. Select the intermediate Rm value that is supplied by the output L net as the load for the input pi network. For example use $Rm = \sqrt{R1 \cdot R2}$ or other as appropriate within the requirement that R1 > Rm > R2. Then compute the L-net Q value.

$$Q_L = \sqrt{\frac{R_m}{R2} - 1} \quad \text{(Eq. 17)}$$

Then

$$X_{L2} = Q_L \cdot R2 \quad \text{(Eq. 18)}$$

$$X_p = Rm/Q_L \quad \text{(Eq. 19)}$$

Select the desired pi-L network Qo value. Next, compute the required pi-network-section Q value.

$$Qo_\pi = Qo - Q_L \quad \text{(Eq. 20)}$$

Use either the Procedure 1 or Procedure 2 pi-network formulas to compute the values for X_{C1}, X_{C2} and X_L for the pi network. Use the value specified for Rm as the value for R2 in the selected procedure. Also, use the Qo_π value for Qo. Note that the X_{C2} value obtained will be the X'_{C2} value in the pi-L network arrangement. Then,

$$X_{C2} = \frac{X'_{C2} \cdot X_p}{X'_{C2} + X_p} \quad \text{(Eq. 21)}$$

This equation yields the pi-L X_{C2} value. This completes the pi-L solution.

A program in BASIC for the Procedure 1 pi-network formulas is given in Table 1. Table 2 gives reactance values for a Qo value of 12 for pi-network tube-type tank circuits. The range of plate load resistances in Table 2 should cover most amateur requirements. Tables 3 and 4 show reactance values for cathode-driven tuned-input circuits for Qo values of 2 and 3 for a range of cathode drive impedances that should cover most requirements. The computer program and printouts are by Don Reaves, KC5JH.

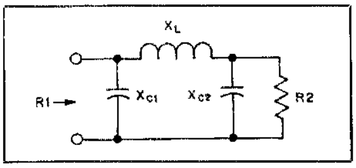

Fig. 3 — Pi-network component designations for use with Procedures 1 and 2. In these procedures, a single set of equations and conditions satisfies both step-up and step-down impedance transformations. Thus, it is unnecessary to specify R1 > R2 or R2 > R1. R1 is the source resistance, and R2 is the load resistance.

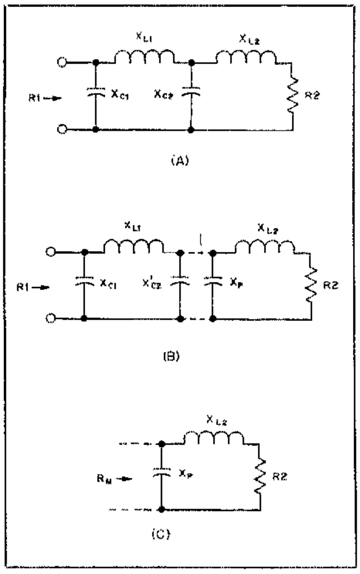

Fig. 4 — The pi-L network, A, is designed as a resonant L network, C, combined with a pi network, as shown at B. X_{C2} is determined through the use of an intermediate value, X'_{C2}, in parallel with Xp.

APPENDIX

For the circuit of Fig. 5A to be equivalent to that of 5B, it is assumed that

$$Z_{in}(A) = Z_{in}(B) \quad \text{(Eq. 22)}$$

$$P_A(A) = P_A(B) = \text{active (real) power} \quad \text{(Eq. 23)}$$

$$P_X(A) = P_X(B) = \text{reactive power} \quad \text{(Eq. 24)}$$

and

$$\frac{Rp}{Xp} = \frac{Xs}{Rs} = Q = \frac{\text{reactive power}}{\text{active power}} \quad \text{(Eq. 25)}$$

$$P_A(A) = I^2Rs = \frac{E^2Rs}{Rs^2 + Xs^2} = P_A(B) = \frac{E^2}{Rp} \quad \text{(Eq. 26)}$$

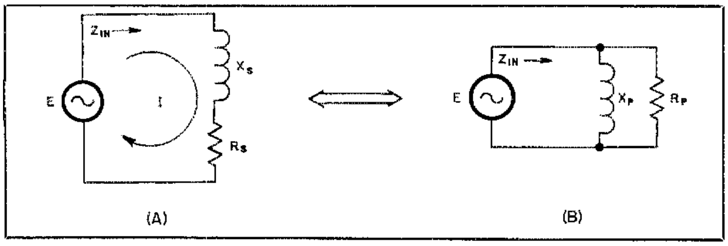

Fig. 5 — A series circuit, A, and its parallel equivalent, B. For the same value of Z_{in} in both drawings, Xs is not equal to Xp, and Rs is not equal to Rp.

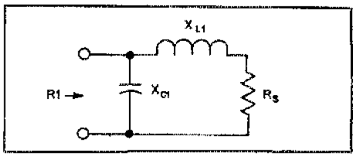

Fig. 7 — The resonant L network, redrawn from Fig. 6A.

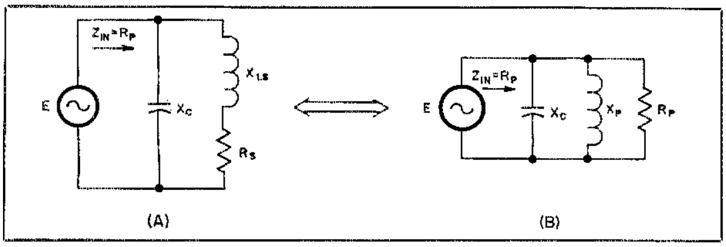

Fig. 6 — The resonant L network is shown at A. At B is the parallel-equivalent circuit.

Therefore,

$$Rp = \frac{Rs^2 + Xs^2}{Rs}$$

$$= Rs\left[1 + \left(\frac{Xs}{Rs}\right)^2\right] = Rs(Q^2 + 1) \qquad \text{(Eq. 27)}$$

$$P_X(A) = \frac{E^2Xs}{Rs^2 + Xs} = P_X(B) = \frac{E^2}{Xp} \qquad \text{(Eq. 28)}$$

Therefore,

$$Xp = \frac{Rs^2 + Xs^2}{Xs} = \frac{Rs}{Q} + Q \cdot Rs$$

$$= \frac{Rs(Q^2 + 1)}{Q} = \frac{Rp}{Q} \qquad \text{(Eq. 29)}$$

From the first terms of Eqs. 27 and 29,

$$Rp \cdot Rs = Xp \cdot Xs \qquad \text{(Eq. 30)}$$

From Eqs. 25 and 27,

$$Rs = \frac{Rp}{1 + \left(\frac{Rp}{Xp}\right)^2} = \frac{Rp \cdot Xp^2}{Rp^2 + Xp^2} \qquad \text{(Eq. 31)}$$

And from Eq. 25 and 29,

$$Xs = \frac{Rp^2Xp}{Rp^2 + Xp^2} \qquad \text{(Eq. 32)}$$

These formulas are for series-to-parallel and parallel-to-series equivalent conversions.

Now see Fig. 6. The series circuit, A, and its parallel equivalent, B, are made parallel resonant by adding a capacitive reactance X_C of a magnitude equal to Xp in shunt, as shown. At the resonant frequency, fo, the circuits appear to the generator as a purely resistive impedance, Z_{in} = Rp. The Q remains the same as before the addition of X_C.

$$Q = Rp/Xp = X_Ls/Rs \qquad \text{(Eq. 33)}$$

where X_Ls = Xs in Fig. 5.

The circuit of Fig. 6A is the resonant L net. This circuit is redrawn in Fig. 7 with the circuit elements redesignated for convenience to use in developing the "standard" Q-based pi-network formulas that follow.

$$Q = R1/X_{C1} = X_{L1}/Rs \qquad \text{(Eq. 34)}$$

$$R1 = Rs(Q1^2 + 1) \qquad \text{(Eq. 35)}$$

where R1 > Rs. Other alternate forms are

$$Q1 = \sqrt{R1/Rs - 1} \qquad \text{(Eq. 36)}$$

$$X_{L1} = \sqrt{R1 \cdot Rs - Rs^2} = \frac{Q1 \cdot R1}{Q1^2 + 1} \qquad \text{(Eq. 37)}$$

$$X_{C1} = R1\sqrt{Rs/(R1 - Rs)} = Rs\,(Q1^2 + 1)/Q1 \qquad \text{(Eq. 38)}$$

In the L net, R1 must be greater than Rs. That is, the shunt reactance element is placed in shunt with the larger of the resistances to be matched.

The pi network is divided into two back-to-back L nets, as shown in Fig. 8. Rs is not a physical resistor, but is rather a virtual resistance. It is the transformed value of R2, the load-end resistance value, and is the value of resistance required to afford R1 at the input or source-end L network for the Q1 value selected.

The standard pi-network Q-based formulas follow. Select the Q1 value. Then,

$$X_{C1} = R1/Q1 \qquad \text{(Eq. 39)}$$

At the input end in Fig. 8A,

$$Rs = \frac{R1}{Q1^2 + 1} \qquad \text{(Eq. 40)}$$

And at the output end,

$$Rs = \frac{R2}{Q2^2 + 1} \qquad \text{(Eq. 41)}$$

Table 1

BASIC Program for Designing a Pi Network Using Procedure 1

```
100 REM  PINET4
110 REM Q BASED FORMULAS PER W5FD FOR CALCULATING PI-NETWORK REACTANCE
    VALUES
120 X1=0:X2=0:XL=0:R1=0:R2=0:QO=0:Q1=0:Q2=0
130 INPUT "R1";R1
140 INPUT "R2";R2
150 INPUT"QO";QO
160 IF (R1*R2*QO)<=0 THEN PRINT "Not a pi network": GOTO 340
170 REM  SPECIAL CASE WHERE R1 = R2
180 IF R1=R2 THEN X1=(2*R1)/QO:X2=(2*R2)/QO:Q1=QO/2:XL=(R1*QO)/Q1*Q1+1):GOTO 280
190 IF ABS((QO*QO+1)-(R1/R2))<.01 THEN PRINT "L network": GOTO 340
200 IF ABS((QO*QO+1)-(R2/R1))<.01 THEN PRINT "L network":GOTO 340
210 IF R1/R2>QO*QO+1 THEN PRINT "No solution":GOTO 340
220 R2/R1>QO*QO+1 THEN PRINT "No solution":GOTO 340
230 Q1=(R1*QO - SQR(R1*R2*(QO*QO) - (R1-R2)*(R1-R2)))/(R1-R2)
240 Q2=QO-Q1
250 X1=R1/Q1
260 X2=R2/Q2
270 XL=(R1*QO)/(Q1*Q1+1)
280 PRINT "R1 = ";R1
290 PRINT "R2 = ";R2
300 PRINT "QO = ";QO
310 PRINT "X1 = ";X1
320 PRINT "X2 = ";X2
330 PRINT "XL = ";XL
340 PRINT: RUN
```

Table 2

Pi-Network Calculations for Tube-Type RF Amplifiers

R2 = 50 ohms, Qo = 12

R1	X_{C1}	X_{C2}	X_L
1500	145	31	166
1550	149	31	170
1600	153	32	175
1650	158	33	179
1700	162	33	184
1750	166	34	188
1800	171	34	193
1850	175	35	197
1900	179	36	201
1950	184	36	206
2000	188	37	210
2025	190	37	212
2050	192	37	215
2075	194	38	217
2100	197	38	219
2125	199	38	221
2150	201	39	223
2175	203	39	226
2200	205	39	228
2225	207	39	230
2250	209	40	232
2275	212	40	234
2300	214	40	236
2325	216	41	239
2350	218	41	241
2375	220	41	243
2400	222	42	245
2425	224	42	247
2450	227	42	249
2475	229	42	251
2500	231	43	254
2525	233	43	256
2550	235	43	258
2575	237	44	260
2600	239	44	262
2625	241	44	264
2650	244	45	266
2675	246	45	268
2700	248	45	271
2725	250	46	273
2750	252	46	275
2775	254	46	277
2800	256	47	279
2825	258	47	281
2850	260	47	283
2875	263	48	285
2900	265	48	287
2925	267	48	290
2950	269	49	292
2975	271	49	294
3000	273	49	296
3050	277	50	300
3100	281	51	304
3150	286	51	308
3200	290	52	312
3250	294	53	317
3300	298	54	321
3350	302	54	325
3400	307	55	329
3450	311	56	333
3500	315	57	337
3550	319	57	341
3600	323	58	345
3650	327	59	349
3700	331	60	353
3750	336	61	357
3800	340	62	361
3850	344	62	365
3900	348	63	369
3950	352	64	373
4000	356	65	377

From this,

$$\frac{R1}{Q1^2 + 1} = \frac{R2}{Q2^2 + 1} = \frac{R2}{1 + \frac{R2^2}{X_{C2}^2}} \quad \text{(Eq. 42)}$$

Table 3

Pi-Network Calculations for Cathode-Tuned Input Circuits

R1 = 50 ohms, Qo = 2

R2	X_{C1}	X_{C2}	X_L
50	50	50	50
55	53	52	52
60	55	55	55
65	58	57	57
70	60	60	59
75	63	62	61
80	66	65	63
85	68	67	65
90	71	69	67
95	74	72	69
100	77	74	71
105	81	76	72
110	84	78	74
115	88	80	76
120	92	83	77
125	96	85	79
130	100	87	80
135	105	89	81
140	110	91	83
145	115	93	84
150	121	95	85

Table 4

Pi-Network Calculations for Cathode-Tuned Input Circuits

R1 = 50 ohms, Qo = 3

R2	X_{C1}	X_{C2}	X_L
50	33	33	46
55	35	35	48
60	36	38	51
65	37	40	53
70	38	42	55
75	39	44	57
80	40	46	59
85	41	48	61
90	42	49	63
95	43	51	65
100	45	53	66
105	46	55	68
110	47	57	70
115	48	59	72
120	49	61	73
125	50	63	75
130	51	64	77
135	52	66	78
140	53	68	80
145	55	70	81
150	56	71	83

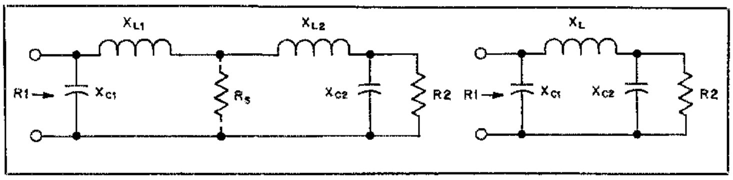

Fig. 8 — The pi network, B, may be considered as two back-to-back L networks, as shown at A. X_L equals the sum of X_{L1} and X_{L2}. Rs is a virtual resistance, not a real resistor.

Solving for X_{C2},

$$X_{C2} = R2 \sqrt{\frac{R1/R2}{Q1^2 + 1 - R1/R2}} \quad \text{(Eq. 43)}$$

Further,

$$X_{L1} = Q1 \cdot Rs = \frac{R1 \cdot Q1}{1 + Q1^2} \quad \text{(Eq. 44)}$$

and

$$X_{L2} = \frac{Rs \cdot R2}{X_{C2}} = \frac{R1 \cdot R2}{(Q1^2 + 1)X_{C2}} \quad \text{(Eq. 45)}$$

$$X_L = X_{L1} + X_{L2} = \frac{R1 \cdot Q1 + R1 \cdot R2/X_{C2}}{Q1^2 + 1} \quad \text{(Eq. 46)}$$

Equations 39, 43 and 46 are the old standard Q-based pi-network formulas. They are in slightly different form from those of Fig. 2, but in fact are the same formulas. The formula for X_L may be put in a form involving only Q1, R1 and R2, by substituting the formula for X_{C2} (Eq. 43) into Eq. 46. There is no advantage in doing this, however, because the value of X_{C2} must be found from Eq. 43 and is already available.

These Q-based pi-network formulas give correct results for the network reactance values required for matching R1 to R2 and the selected input-end L-network Q1 value. It is to be noted, however, that this Q1 value is not the pi-network operating Qo value, which is

$$Qo = Q1 + Q2 = R1/X_{C1} + R2/X_{C2} \quad \text{(Eq. 47)}$$

A pi-network solution to match R1 to R2 at the Q value selected for Q1 is possible when $(Q1^2 + 1) > R1/R2$. When $(Q1^2 + 1) = R1/R2$, X_{C2} goes to infinity and an L-net solution results. As Eq. 27 indicates, $Q1^2 + 1 = R1/R2$ is the L-network equation. If $(Q1^2 + 1) < R1/R2$, no solution can be afforded. Note that both R1 and R2 must exceed Rs. (The $Q1^2 + 1 - R1/R2$ term of Eq. 43 takes care of this.)

Now we examine the Qo of pi networks and pi-L networks. In Fig. 9,

$$P_A = \text{active power} = E1^2/R1 = E2^2/R2 \quad \text{(Eq. 48)}$$

$$P_{X1} = \text{reactive power} = E1^2/X_{C1} \quad \text{(Eq. 49)}$$

$$P_{X2} = \text{reactive power} = E2^2/X_{C2} \quad \text{(Eq. 50)}$$

$$Q = \frac{2\pi \text{ max. energy stored}}{\text{energy dissipated per cycle}} = \frac{\text{reactive power}}{\text{active power}} \quad \text{(Eq. 51)}$$

$$Q_\pi = \frac{\frac{E1^2}{X_{C1}} + \frac{E2^2}{X_{C2}}}{\frac{E1^2}{R1}} \quad \text{(Eq. 52)}$$

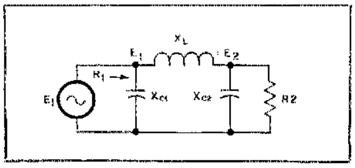

Fig. 9 — The pi network identified for a discussion of the circuit operating Q, Qo.

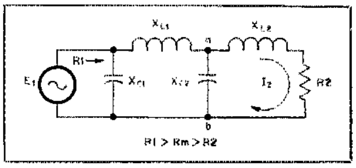

Fig. 10 — The pi-L network identified for a discussion of Qo.

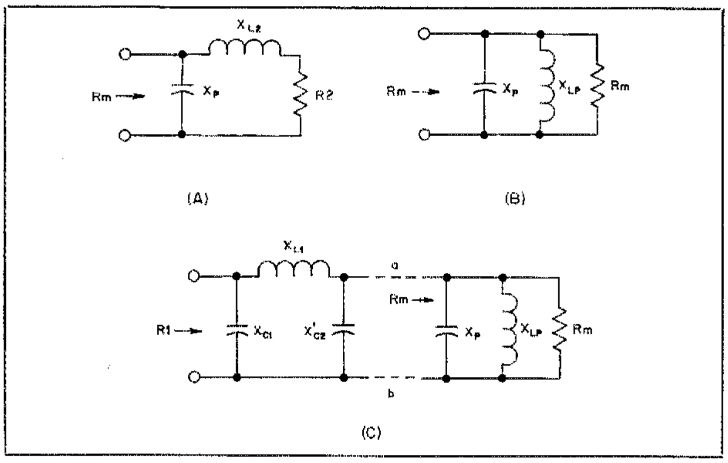

Fig. 11 — Derivation of the pi-L network from an L section, A, and its parallel equivalent, B. The parallel equivalent is added to a pi section, C.

where, from Eq. 48,

$$E2^2 = \frac{R2 \cdot E1^2}{R1} \qquad \text{(Eq. 53)}$$

Therefore,

$$Q_\pi = \frac{R1}{X_{C1}} + \frac{R2}{X_{C2}} = Qo_\pi \text{ (pi-net Qo)} \qquad \text{(Eq. 54)}$$

And for the Qo of the pi-L network, refer to Fig. 10. In that drawing,

$$Z_{AB} = R_{mid} = Rm \qquad \text{(Eq. 55)}$$

where R1 > Rm > R2.

$$Q_{\pi L} = \frac{R1}{X_{C1}} + \frac{Rm}{X_{C2}} = Qo_{\pi L} \text{ (pi-L net Qo)} \qquad \text{(Eq. 56)}$$

The reactive power associated with $I2^2 \cdot X_{L2}$ is included in $E_{AB}{}^2/X_{C2}$.

The Pi-L Network

Refer again to Fig. 4. The circuit Qo value for the pi-L network may be expressed as follows:

$$Qo_{\pi L} = \frac{R1}{X_{C1}} + \frac{Rm}{X_{C2}} \qquad \text{(Eq. 57)}$$

Only the arrangement where R1 > Rm > R2 is considered here, the output-tank-coupling-circuit operation. The Qo of the pi-L is shown as follows. To the series combination of X_{L2} and R2 there is added a capacitive reactance, Xp, to provide a resonant L net with R_{in} = Rm. This is shown in Fig. 4C, as well as in Fig. 11A. In that circuit,

$$Q_L = Rm/Xp = X_{L2}/R2 \qquad \text{(Eq. 58)}$$

The parallel equivalent for this circuit is shown in Fig. 11B. In this circuit,

$$Q = Rm/Xp \qquad \text{(Eq. 59)}$$

Adding the input-end pi network results in the circuit of Fig. 11C. From that diagram,

$$Qo_{\pi L} = \frac{R1}{X_{C1}} + \frac{Rm}{X'_{C2}} + \frac{Rm}{Xp} \qquad \text{(Eq. 60)}$$

$$Qo_{\pi L} = \frac{R1}{X_{C1}} + \frac{Rm}{X_{C2}} \qquad \text{(Eq. 61)}$$

where

$$X_{C2} = \frac{X'_{C2} \cdot Xp}{X'_{C2} + Xp} \qquad \text{(Eq. 62)}$$

$$\frac{Rm}{X'_{C2}} + \frac{Rm}{Xp} = Rm\left(\frac{X'_{C2} + Xp}{X'_{C2} \cdot Xp}\right) = \frac{Rm}{X_{C2}} \qquad \text{(Eq. 63)}$$

Derivation of the Q1 Formula Used in Procedure 1

Refer again to Fig. 8. From the back-to-back L-net pi-network equations,

$$Qo = Q1 + Q2 \qquad \text{(Eq. 64)}$$

$$\frac{R1}{Q1^2 + 1} = \frac{R2}{Q2^2 + 1} \qquad \text{(Eq. 65)}$$

$$Q2^2 = \frac{R2(Q1^2 + 1) - R1}{R1} \qquad \text{(Eq. 66)}$$

$$Q2 = Qo - 1 \qquad \text{(Eq. 67)}$$

$$Q2^2 = (Qo - Q1)^2 \qquad \text{(Eq. 68)}$$

$$\frac{R2(Q1^2 + 1) - R1}{R1} = Qo^2 - 2 \cdot Qo \cdot Q1 + Q1^2 \qquad \text{(Eq. 69)}$$

Solving for Q1 yields

$$Q1 = \frac{Qo - \sqrt{\frac{R2}{R1}\left(Qo^2 + 2 - \frac{R2}{R1}\right) - 1}}{1 - \frac{R2}{R1}} \qquad \text{(Eq. 70)}$$

This formula is rearranged in simpler form and is presented as Eq. 4 in Procedure 1, given earlier. This gives Q1 in terms of the known Qo, R1 and R2, and enables solution equations for the pi-network reactance element values in terms of Qo, R1 and R2.

Derivation of the X_L Formula Used in Procedure 2

Refer now to Fig. 3. Everitt's equations for the low-pass pi network follow.[6]

$$X_{C1} = \frac{R1 \cdot X_L}{R1 + \sqrt{R1 \cdot R2 - X_L{}^2}} \qquad \text{(Eq. 71)}$$

$$X_{C2} = \frac{R2 \cdot X_L}{R2 + \sqrt{R1 \cdot R2 - X_L{}^2}} \qquad \text{(Eq. 72)}$$

$$Q1 = \frac{R1 + \sqrt{R1 \cdot R2 - X_L{}^2}}{X_L} \qquad \text{(Eq. 73)}$$

$$Q2 = \frac{R2 + \sqrt{R1 \cdot R2 - X_L{}^2}}{X_L} \qquad \text{(Eq. 74)}$$

Qo is the circuit Q value, and may be expressed as

$$Qo = Q1 + Q2 \qquad \text{(Eq. 75)}$$

Substituting the Q1 and Q2 expressions in Eq. 75 and solving for X_L yields

$$X_L = \frac{Qo(R1 + R2) + 2\sqrt{R1 \cdot R2(Qo^2 + 4) - (R1 + R2)^2}}{Qo^2 + 4} \qquad \text{(Eq. 76)}$$

This gives X_L in terms of the known Qo, R1 and R2, and enables solutions for X_{C1} and X_{C2} from Everitt's equations or from the equations given in Procedure 2.

The networks discussed in this article are termed linear networks. In addition to other things, this means that they have a simple linear Ohm's law response to an input. It means that these networks work in both directions. That is, a pi, a pi-L or an L network may be designed to transform a load-end R2 value into a value of R1 ohms at the source or input end. If the physical resistor R2 is removed and a physical resistor R1 is placed at the R1 end terminals, then this value of R1 will have been transformed into a value of R2 ohms at the R2 end terminals.

It is to be noted that there is no restriction placed on the ratio of R1 to R2 in the new formulas given in this article. This is unlike the way in which the old standard Q-based pi-network formulas were normally stated, in which R1 had to exceed R2. The new formulas work equally well, whether R1 is greater than or less than R2. The only restriction is in the requirement that $Qo^2 + 1$ must exceed the higher value of R1/R2 or R2/R1. The requirement that R1 had to exceed R2, as the old standard formulas were normally stated, was never correct and, in fact, the old standard formulas work equally well for $R1 > R2$ or $R1 < R2$, as an application to an example pi network solution will indicate. This is in consequence of the fact that the old standard Q-based formulas did and do provide an exact "mathematical model" of the physical pi network insofar as computing the reactive network element values for the R1-to-R2 transformation. The only shortcoming of the old formulas is that the computation was based on a partial circuit Q value, and this partial value was taken as the circuit operating Q value. Even so, once the network was solved, the correct operating Qo value was readily available by adding the partial circuit Q value that was not accounted for. The new formulas allow the network solution based on the correct circuit Qo value selected at the outset.

Notes

[1]E. W. Pappenfus and K. L. Klippel, "Pi Network Tank Circuits, *CQ*, Sept. 1950, p. 27.

[2]F. Davis, "Matching Network Designs with Computer Solutions," *Application Note 267*, Motorola, Inc., p. 6.

[3]H. L. Gibson, "An Improved Design Method for Pi and L Pi Network Couplers," *Radio Communication*, June 1969, p. 390.

[4]*Radio Data Reference Book* (3rd edition). London: RSGB, 1972.

[5]E. W. Whyman, "Pi-Network Design and Analysis," *Ham Radio*, Sept. 1977, p. 30.

[6]K. Kajii, "Design of Pi-Type Circuits," *CQ Ham Radio* (Japan), June 1974, p. 264.

[7]L. Everitt, *Communication Engineering*, 1st edition (New York and London: McGraw-Hill Book Company, Inc., 1932).

[8]See note 7.

New Books

☐ *70 Years of Radio Tubes and Valves* by John W. Stokes. Published by the Vestal Press Ltd., Vestal, NY 13850. First edition, 1982. Hardbound, 8-3/4 × 11 inches, 247 pp., $21.95.

☐ Amidst the furor of home computers, large-scale integration and satellite TV, there remains a small, but highly devoted contingent of vacuum-tube enthusiasts. And why not? Admittedly, present-day use of tubes is largely restricted to high-power (usually transmitting) applications, but not a single modern-day radio circuit has developed without some footing in tube-based principles. For this reason alone, we owe homage to the vacuum tube with books that preserve the history of this noble device. *70 Years of Radio Tubes and Valves* is such a book.

Even today's "silicon generation" should find the story of vacuum tubes' golden age extremely interesting — or perhaps fascinating is a better word. John Stokes does much to promote this fascination in *70 Years,* which gives the book broad-based appeal.

The title may seem a bit misleading, for what, if *any,* difference is there between a valve and a tube? Technically, there is none. But Stokes, who is a New Zealander, writes as an impartial historian of the parallel but distinctly separate developments in the American and European tube industries. With American developments, he refers to the vacuum tube, while with European counterparts he refers to the valve, after Fleming's early name for the device.

Stokes has ordered his book according to the chronological development of the tube. He begins by documenting the early experiments of Edison, Fleming and DeForest, while carefully noting the ambiguity of this developmental period — primarily surrounding each man's claim to be *the* original inventor. As the book progresses, entire chapters are devoted to specific aspects of tube development. For example, there are chapters entitled, "The Grid," "Metal Envelopes," "Penta-Hodas" and "Miniaturization." Each chapter is extremely well-researched and so packed with information that I occasionally found it necessary to stop reading and digest a bit. My only criticism (and a minor one at that) stems from Stokes' tendency to cram perhaps a wee bit too much information into each chapter.

It should be noted that readers specifically interested in the developments of companies such as Telefunken and Phillips will not be satisfied by this book. Stokes has intentionally limited his research to the USA and England, citing language difficulties as the reason. These companies are mentioned (one minor chapter is on Phillips), but only when their achievements affected the course of tube development on a worldwide basis.

Some mention should be made of the excellent selection of photos and graphics in *70 Years.* There is an abundance of original tube advertisements, which alone often tell the state of the art at that time. It's interesting to see ads that directly brand competitors' products unauthorized under patent law. Also, there was a great deal of emphasis placed on "magic breakthroughs" in early tube technology that could easily cure *all* receiver problems — much the same as "snake oil" for your health! Photos are equally interesting, and practically every tube type that is mentioned finds its way into a photograph or excellent illustration.

As the subtitle, "...A Guide for Electronic Engineers Historians and Collectors," claims this book should prove to be a valuable reference work both now and in the future. You don't fit into any of the categories? Don't despair, for *70 Years* will prove to be interesting to *all* who have at least a passing fancy on the subject. — *Dennis J. Lusis, W1LJ*

Strays

WRITING TO HQ.?

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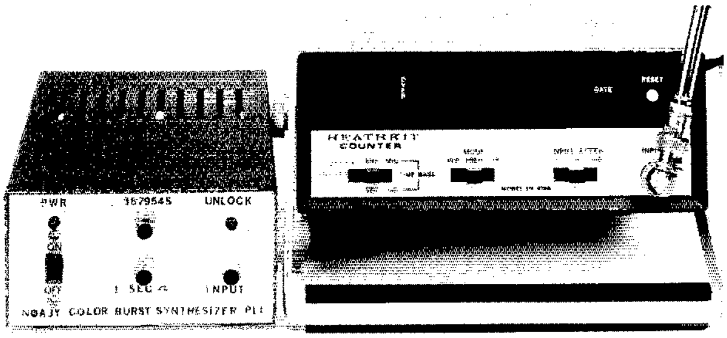
4) Enclose a business-sized self-addressed, stamped envelope if a reply is required.

5) Address your request to a particular individual or department, if possible, especially when responding to correspondence received from Hq.

6) Send a check or money order (IRCs for foreign requests) when applicable. *Do not* send cash.

The NØAJY cb Standard

Another Citizens Band radio conversion project? *No sir!* The term cb means *c*olor *b*urst — the ultimate in frequency-counter accuracy!



By David Bissen,* NØAJY

Have you tried to measure the frequency of your vhf transceiver with an inexpensive frequency counter? The results can be disheartening. This paper describes several methods of upgrading your counter to lab-quality status. Simple CMOS circuits are used to extract the 3.579545-MHz color-burst (cb) frequency from an ordinary TV set, and this signal is divided down and used to gate your counter. Why the cb frequency? During broadcasts by the four major TV networks, the cb signal is locked to a standard with stability and accuracy measured in parts per *billion!*

I own a Heathkit IM-4100 30-MHz counter. The accuracy was adequate for my requirements as a Novice, but after moving to vhf, I felt the need to make frequency checks that were above the range of the counter. After adding a divide-by-ten prescaler, I noticed heat inside the unit caused drift in the 10-MHz time-base oscillator. Changing the crystal and oscillator components did not cure the problem. A crystal oven, consisting of a Plexiglas® enclosure around the oscillator and a copy of a commercial temperature-control circuit, was tried as a solution. This improved the frequency stability to a point where a very accurate reference signal would be required to test the unit. Then, I remembered the color-burst frequency, 3.579545 MHz, which is available from a color-TV set.

Initial Tests

I found a spot in my TV, near the color-burst crystal, where a stable frequency reading was available. Measuring the signal frequency with my counter, I found the counter internal stability to be one part per million (ppm) after a two-hour warmup. This was not acceptable. The oscillator also made violent frequency excursions during warmup.

Color-TV Gating Circuit

The block diagram of my initial standard is shown in Fig. 1. It has selectable outputs of 1, 10 and 60 Hz, along with 3.58 MHz. The circuit uses the National MM5369 chip to divide the cb frequency by 59,569. This produces an output of 60.000084 Hz, which is a 1.4-ppm error. Counter accuracy remains good because the error is fixed and can be computed out mentally. The disadvantage of this approach is the requirement of a color TV receiver.

A "Universal" Circuit

There is a direct relationship between the horizontal-scan rate and the color-burst frequency in TV transmissions — the cb frequency in hertz is equal to the horizontal rate divided by 2 and multiplied by 455. The circuit shown in Fig. 2 was chosen for its low cost and locally available parts, and because it works! It consists of two sections. One section locks onto the horizontal sync of *any* television set (black-and-white or color) and develops the cb frequency, 3.579545 MHz. The other section divides this signal to a 1- or 1/10-second gating interval.

Input signals from a pickup loop near the TV set are cleaned and brought up to CMOS levels by means of transistor amplifiers Q1, 2 and 3. The horizontal-sync signal (15734.264 Hz) is then used to lock the PLL to 3.579545 MHz via dividers U1 and U3. The cb signal from the PLL is then corrected to 3.579540 MHz by a "one and only one" circuit consisting of flip-flops U8, U9 and U10. This circuit inhibits the input signal once for every 715,908 clock cycles. The 3.579540-MHz signal is then applied to the MM5369 divide-by-59,659 chip, which yields an output of exactly 60.000 Hz. A 1- or 1/10-second gating pulse is then delivered via U9 and U11.

Components and Construction

Transistors used in the standard circuit are general-purpose npn types with an h_{fe} of 40 or better. The 4046 PLL is rated at a maximum frequency of 3 MHz. I have found conflicting specifications, but conclude that they will work to at least 4 MHz with 15 volts V_{DD}. If Murphy's Law

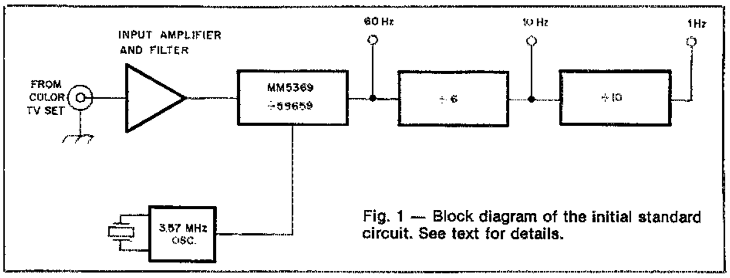


Fig. 1 — Block diagram of the initial standard circuit. See text for details.

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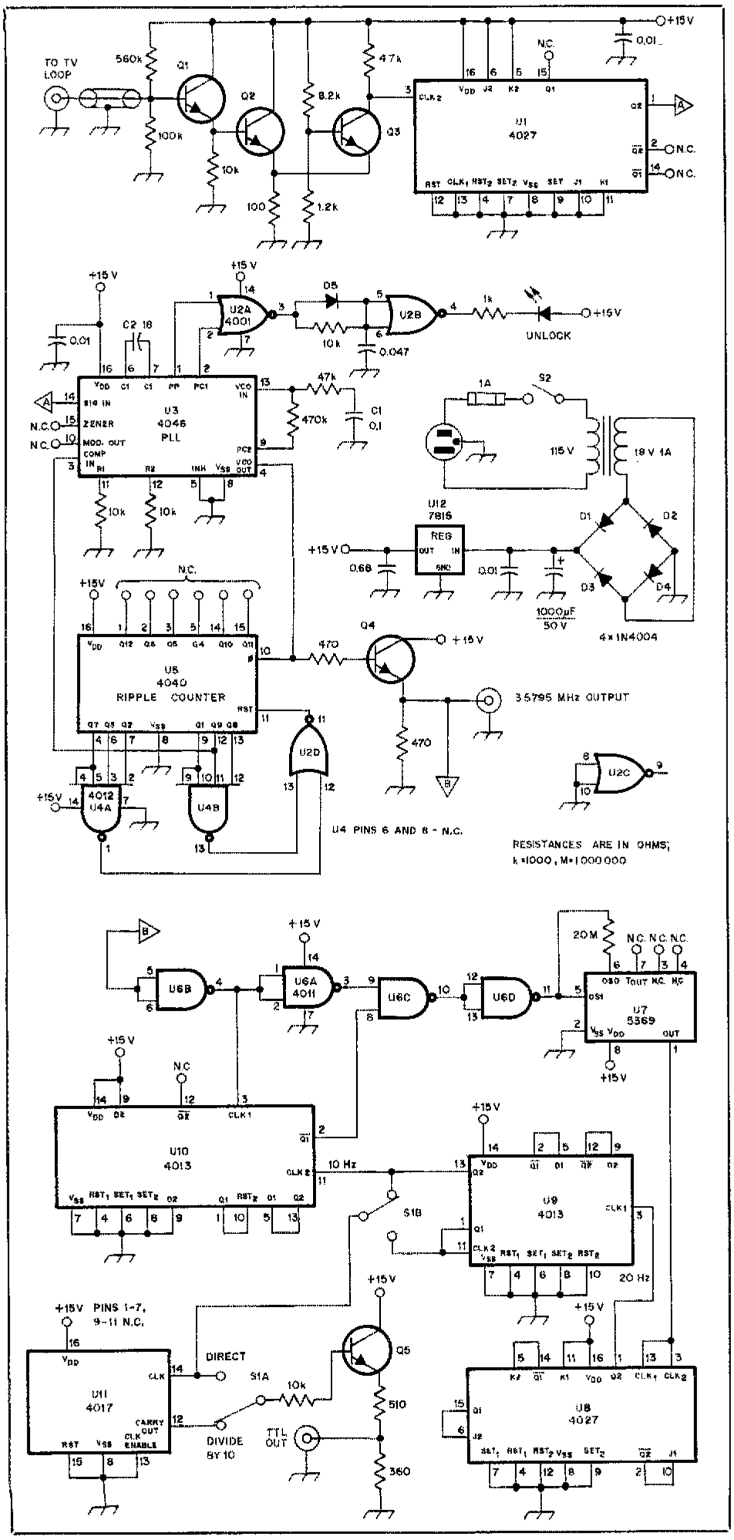


Fig. 2 — Schematic diagram of the NØAJY standard circuit. All resistors are 1/4-W carbon-composition or film types.

D5 — 1N4154 silicon signal diode.
Q1-Q4 — Small-signal npn silicon transistor, 2N2222A or equiv.
U1, U8 — CMOS Dual J-K Master-Slave flip-flop, CD4027 or equiv.
U2 — CMOS Quad 2-input NOR gate, CD4001 or equiv.
U3 — CMOS Micropower phase-locked loop, CD4046 or equiv.
U4 — CMOS Dual 4-input NAND gate, CD4012 or equiv.
U5 — CMOS 12-stage binary/ripple counter, CD4040 or equiv.
U6 — CMOS Quad 2-input NAND gate, CD4011 or equiv.
U7 — CMOS 17-stage programmable oscillator/divider, MM5369 or equiv.
U9, 10 — CMOS Dual "D" flip-flop, CD4013 or equiv.
U11 — CMOS decade counter/divider, CD4017 or equiv.
U12 — 15-V, 1-A positive-voltage regulator, type 7815 or equiv.

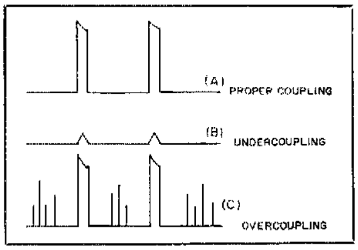
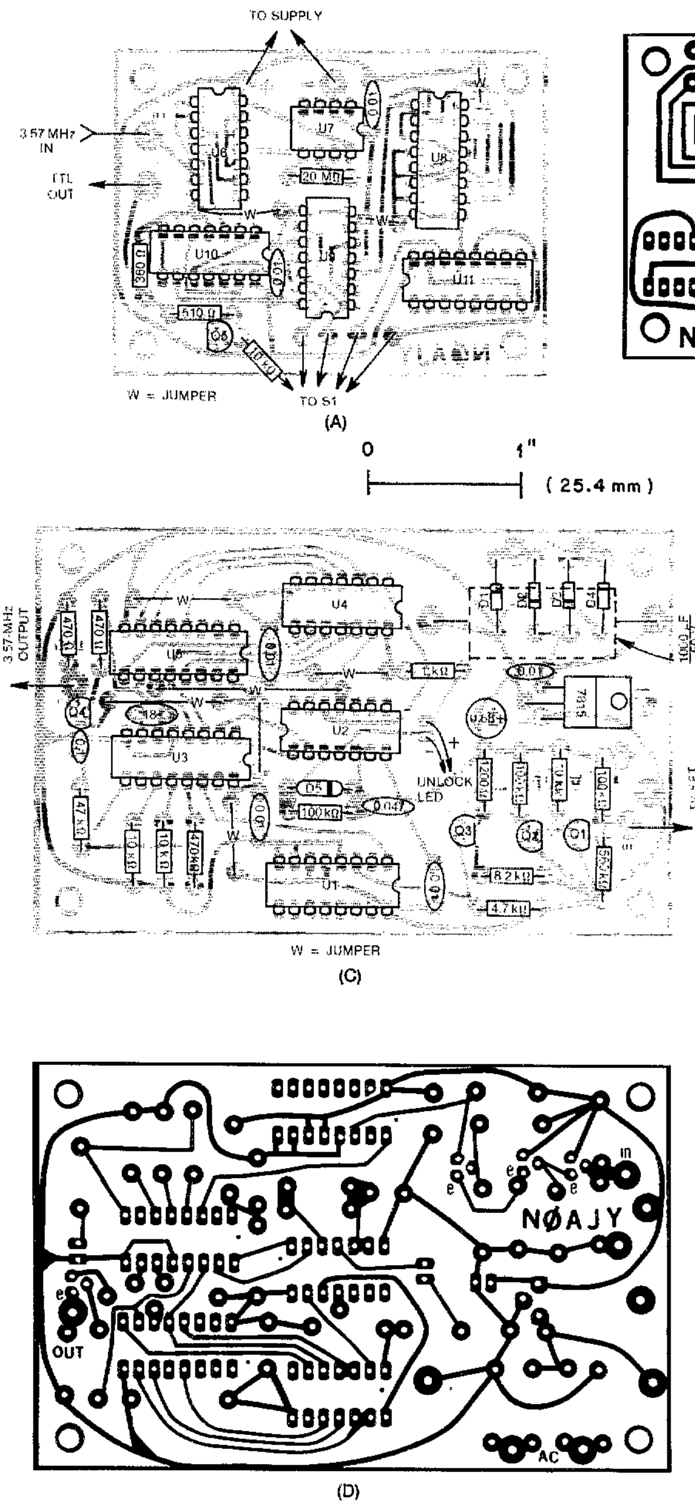


Fig. 3 — Under- or over-coupling the horizontal sync from the TV set will cause improper operation of the PLL.

prevails and you purchase a slow one, try substituting another. The impedance at C1, the loop-filter capacitor, is very high; the capacitor should be a good low-leakage unit, such as a tantalum type. The prototype unit was built using two printed-circuit boards, and mounted in an aluminum case similar to a Radio Shack model 270-253. S1 is optional; the connections can be "hard wired" to suit your counter.

Check Out

Check the power-supply section of the project before installing any of the solid-state devices. After the ICs are in place and the power is applied, the UNLOCK LED should light. Check for proper VCO operation in one of the following ways: Connect a short lead to the base of Q1 and hold it near the screen of your TV set. The LED should flicker and go out; if it does not, readjust the position of the pickup cable until the LED stops flickering and stays off. Another method of confirming PLL operation is to monitor the frequency of the VCO while alternately connecting pin 9 of the 4046 from ground to V_{DD}. The VCO should swing above and below 3.58 MHz during this test, with 3.58 MHz being



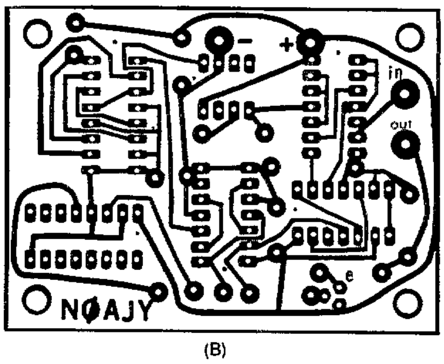


Fig. 4 — Parts-placement diagram for the reference boards. Components are mounted on the nonfoil side. Gray areas represent unetched copper. (A) and (C): Parts-placement diagrams for the reference components mount on the nonfoil side. Gray areas represent unetched copper. (B) and (D): Etching patterns for the reference boards. Patterns are shown actual size from the foil side of the board. Black areas represent unetched copper. Good engineering practice requires all unused inputs to be grounded. This was not done on these boards. Wire jumpers can be added to correct this fault.

the approximate center frequency. The free-running frequency of the VCO is set by capacitor C2, the value of which is quite critical. Values between 15 and 20 pF worked satisfactorily in the prototype. An 8- to 25-pF trimmer should be suitable for those desiring an exact adjustment.

Look Out for Troubles

Over-coupling to the TV set will cause ringing, which will unlock the loop. Examine the VCO output with a scope; it should look like Fig. 3A. Some counters may count noise or load the standard output circuit. It is important for the builder to be familiar with the ins and outs of his own counter. Occasionally, a glitch in the TV signal will wipe out enough sync signal to change the counter a few cycles; however, this is the exception, as the loop filter averages out minor noise pulses.

Summary

It is not often that the typical amateur has the chance to get his or her hands on test equipment of commercial quality. Less likely would be the ability to build equipment that rivals commercial standards. Those of you who build this unit will be amazed at the accuracy and stability. With gating locked to network standards, you will have a stable counter that can allow you to check for drift without wondering if the counter or equipment under test is at fault.

Thanks to Jim Summers, WB9QFC, for his technical help with this article. Please send an s.a.s.e. with any correspondence.

A Battery Low-Voltage Indicator

Don't get caught with dead batteries. Here are some simple circuits that provide advance warning.

By Harry M. Neben,* W9QB

Mobile and hand-held equipment usually does not contain an indicator to alert the operator to battery condition. One solution: Simple low-voltage indicators can be made from components found in many shacks.

The most common batteries found in amateur service are the nickle-cadmium and lead-acid types. NiCds are used in most hand-held transceivers, and lead-acid batteries are found in motor vehicles. Both are susceptible to being overlooked until it's too late to keep your rig going!

Nickel-cadmium cells have a no-load potential of 1.4 V and nominal output of 1.25 V; they will power your transceiver until they drop to a critical value of 1.1 V. This is shown in Fig. 1A.

Lead-acid batteries have a no-load voltage of 2.1, with a nominal output of 2.0 V, and are useful until they drop to a 1.8-V critical value. This is shown in Fig. 1B.

I constructed a simple device that pro-

*1151 Fairway Dr., Dunedin, FL 33528

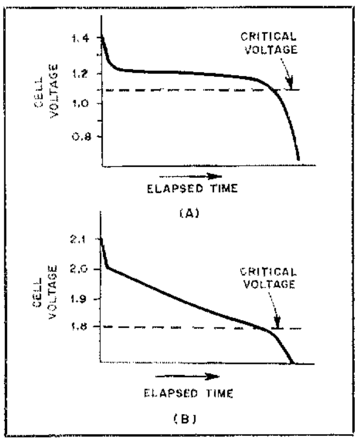


Fig. 1 — Voltage/time discharge curves (at constant current drain) for NiCd batteries (A) and lead-acid cells (B).

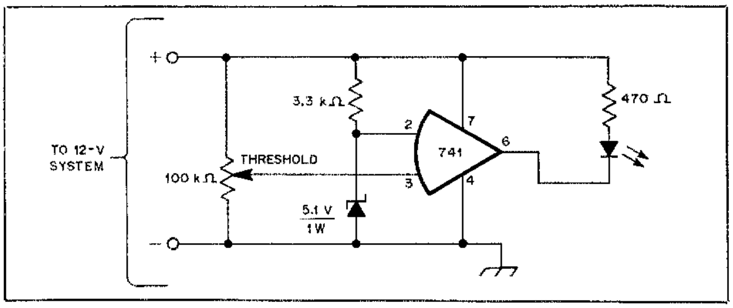


Fig. 2 — Basic comparator circuit used to indicate low battery voltage.

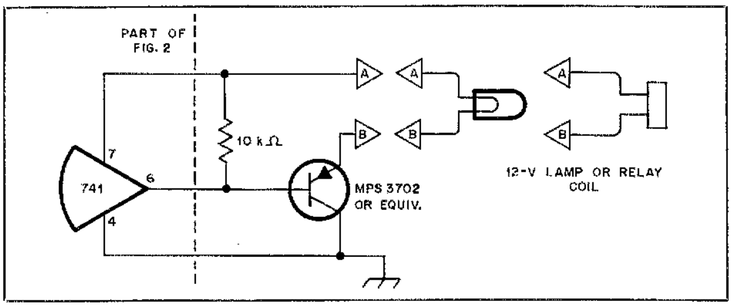


Fig. 3 — The simple addition of a transistor amplifier allows the use of a pilot lamp or relay.

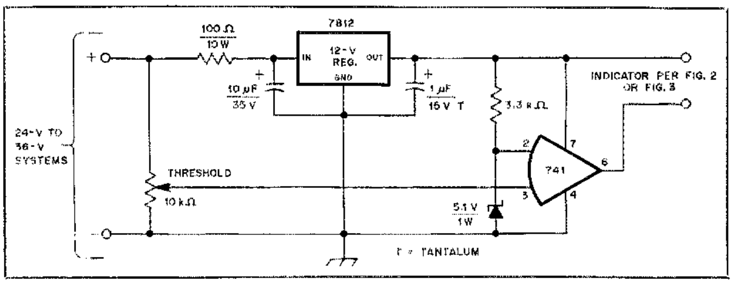


Fig. 4 — Higher-voltage systems such as found in airplanes and boats, necessitate voltage regulation before the comparator to avoid damaging it.

vides an indication of when battery voltage approaches the critical point. My circuit uses a 741 op amp in a comparator circuit (Fig. 2). This functions by comparing the voltage at pin three of the op amp with that found at pin two. As long as pin three is higher in voltage than pin two, the comparator output will be near the full supply voltage, keeping the LED off. However, when the voltage at pin three drops below that at pin two, the comparator output drops to near ground potential, and the LED lights. Simple, isn't it?

Reference voltage for the comparator is provided by a 5.1-V Zener diode, although any other value below the critical voltage will work as well. (Be sure to change the series resistor value, though!) The voltage at pin three is set by adjusting the 100-kΩ potentiometer. To calibrate, apply a dc voltage equal to the cut off value you wish to indicate. Simply adjust the potentiometer until the LED comes on. Check your calibration by *slightly* increasing the input voltage to the circuit. The LED should turn off. That's all there is to it!

To operate a relay or pilot light from this circuit, the LED and resistor are deleted, and a dc amplifier is added. Any general-purpose PNP transistor can be used as an amplifier. Fig. 3 shows these additions.

For those who need an indicator to use on a 24 to 36-V circuit, such as found in boats, airplanes or golf carts, the circuit in Fig. 4 should suffice. At these relatively high input voltages, the rating of the 741 op amp can be exceeded by a transient condition. A voltage-dropping resistor and regulator are added to bring the operating voltage within the rating of the 741. This little "rascal" is the hit of the golf-cart set!

All of these circuits may be left connected to an automobile battery with little effect on the capacity. In the standby mode (LED off), the current drain is only 2.5 mA. This will increase to about 30 mA with the LED on, and more if a relay or pilot light is used. If you are using one of these circuits with a NiCd battery pack, a switch to disconnect the unit is a useful addition.

Strays

It must be difficult for K2BLA to decide which use is best for these elements of his pipe organ: "Shall I build a 6-meter cavity resonator from this thing, or shall I make music with it?"

TA PROFILES

☐ This month, we introduce ARRL Technical Advisor Albert D. Helfrick, K2BLA, our expert on specialized modes, uhf and microwaves. We extend our hearty thanks to Al for the many services he has provided as a TA since 1979. He is a Life Member of the ARRL, and has written numerous *QST* articles (the first one for the December 1969 issue). Al has also contributed excellent material for the ARRL *Radio Amateur's Handbook*. In addition, he has written three non-ARRL books: one on Amateur Radio, one on communications equipment repair and a third, on aviation electronics, to be published early in 1984.

An Extra Class licensee, Al has kept the same call he was issued in 1957. He has operated in all bands, from 160 meters to 70 cm, using all modes. Al has been awarded WAS, WAC, WPC, WPX and DXCC certificates, all on cw. If you're looking for Al, you'll find him operating RTTY and cw in most of the hf bands.

Residing in Boonton, New Jersey, Al is employed by Cessna Aircraft. His principal responsibility at work is the design of distance-measuring equipment (DME) and navigation computers. He is an adjunct faculty member of Kean College of New Jersey, where he teaches electronics. He has received degrees in physics and mathematics, and has taught both subjects at the college level.

Aside from Amateur Radio, Al spends much of his free time playing a 7-rank pipe organ at home. He also plays low brass instruments. He is the organizer of and principal tuba artist in a brass chamber-music group that plays an ambitious concert season each year. If you search the rivers and ponds of North Jersey, you may spot Al paddling his kayak and enjoying the serenity after a busy day's work. — *Marian Anderson, WB1FSB*

HELLO, TEST...

☐ "Break, break...W6ISQ here... ahhhhh, anybody out there on the CATS net can give me a check on a new mike?

"Yeah, bad hum."

"Too much lows."

"Lotta background noise...you mobile or something?"

"Sounds like a hand-held rig."

"I don't hear no hum."

"Back off your derivation control."

"Sounds okay here."

"My scope says too many highs."

"Get closer to the mike."

"Break, break..."

"The first mike was better."

"Nothing wrong with your new mike. It sounds great. Can always recognize your voice, Sam."

"Break, break...ahhhh...I haven't even plugged in the new mike yet... QRX. I'll plug it in now...ahhhhh... testing...ha-one...how does this sound now? Oh, and this is *not* Sam...this is *Italian* String Quartets!"

"Sounds the same."

"Back away from your mike."

"Sounds like your rig is fming."

"It is a fm rig, Old Man."

"Too noisy...you got a fan going?"

"You gotta change that derivation control."

"Hum is worst."

"My scope says too many lows."

"I don't hear no hum."

"It ain't full quieting."

"Still sounds okay here."

"Break, break..."

"Naw, bad distortion."

"If I was you I'd get off the air."

"Aw, naw...don't pay no attention to all them reports. Everybody can still recognize your voice, Sam." — *John G. Troster, W6ISQ, Atherton, California*

I would like to get in touch with...

☐ anyone who has information on a homebrew RTTY device constructed in 1963 by K5EXW and labeled as designed by W2JAV with modifications by W5ANW. William Hulette, KA5BWU, 4331 Deer Creek Dr., Jackson, MS 39211.

A Beginner's Look at RF-Power Measurement

Measuring transmitter output power is simpler than you may think. We need not purchase expensive rf-power meters. Ordinary test equipment or homemade instruments open the door to this routine ham-shack procedure.

By Doug DeMaw,* W1FB/8

Are you tempted to buy a commercially made device for measuring your transmitter output power? That's fine if you're willing to part company with those dollars you've been sequestering in one of the parts drawers in your workshop! But,there are simple and inexpensive ways to measure transmitter output power if you're willing to learn how. All that's needed is an rf ammeter (surplus) or a VTVM (vacuum-tube voltmeter) and a 50-ohm dummy load. Of course, an FET voltmeter can be used in place of a VTVM, but whichever instrument is chosen, it should have a very high input impedance in order to work with the rf probe we shall describe. More about that later.

We'll also discuss in-line power metering for the situation when our antenna is connected to the transmitter. But, no matter what we contemplate doing with our ham station, a dummy load is always an item that we should have on hand for all manner of equipment testing. The Heath Co. Cantenna is entirely adequate for amateur needs, and it is reasonably priced. If you don't have one in your shack, perhaps now is the time to obtain one, or at least a dummy load with equivalent characteristics.

Generally, we are more interested in knowing what the SWR (standing-wave ratio) on our feed line is than we are with the absolute rf output power of our transmitter. But there are times when we need to know something about the latter — especially when we repair an existing rig or test a new piece of homemade transmitting equipment. Therefore, it may not be prudent to spend a large amount of money for a power-monitoring instrument. Less-expensive alternatives are available to us if only occasional power measurements are to be made.

The Old Way — RF Ammeters

Many years ago, we used a standard technique when measuring transmitter output power in the medium- and high-frequency spectrum. The device was an rf ammeter. They are still available at reasonable prices from those dealers who sell military surplus. Rf ammeters show up frequently at ham radio flea markets. Some may have full-scale movements that are meant only to accommodate rf milliamperes, while others may be capable of reading rf current up to 10 A. Rf ammeters contain a bimetal device known as a "thermocouple." The rf energy heats this element and causes the meter to deflect. Thermocouple elements are sensitive to excessive current, and will burn out (become open) quickly if too much current is fed through the meter. Therefore, if you buy an rf meter, be sure it is rated for more current than your transmitter can generate at maximum rated output power.

Rf ammeters are usually meant for use from 50 MHz down, respective to operating frequency. Therefore, they aren't recommended for vhf use. Although they can provide a meter reading at vhf, the meter indication will not be accurate. Furthermore, the mere presence of the meter in the feed line can cause an SWR "bump," and disturb an otherwise low SWR. This is not apt to be a problem at 30 MHz and lower.

Fig. 1 shows the usual method for utilizing an rf ammeter to measure rf power. The instrument is placed between the transmitter and a 50-ohm dummy load. We need only to understand grade-school algebra to determine the transmitter power when we know the amount of rf current that flows into a known-value resistive load. In Fig. 1, we shall assume our transmitter puts out enough power to cause 3 A of rf current to flow through the meter. Using Ohm's law, we can readily calculate the power into R_L as 450 W. The same procedure would be followed if our feed line to R_L was 75 ohms or some other value. Of course, the dummy load would need to have a resistance that equaled the impedance of the coaxial cable. Our rf ammeter could remain in the transmission line during normal on-the-air activity, and it

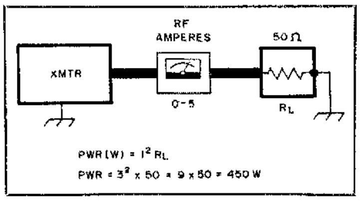


Fig. 1 — Method of rf-power measurement when an rf ammeter is used.

*P.O. Box 250, Luther, MI 49656

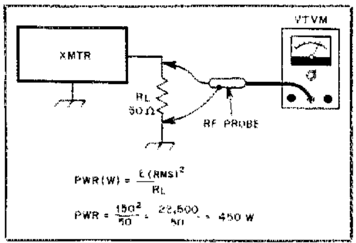


Fig. 2 — A vacuum-tube voltmeter can be used with an rf probe to measure rf power across a known load resistance.

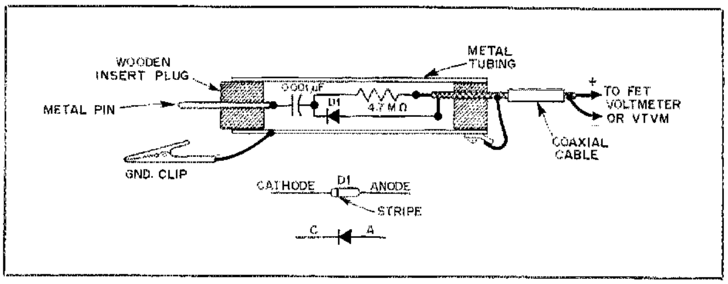


Fig. 3 — Details for building a homemade rf probe. D1 is a small-signal germanium diode.

would yield an accurate power indication as long as the antenna presented a 50-ohm impedance to the transmitter.

Power Determination with RF Voltage

If current is allowed to flow through a resistance, a voltage will appear across that resistance. This is illustrated in Fig. 2, where a 50-ohm resistor serves as our dummy antenna. Here again we need only to apply Ohm's law ($P = E^2/R$, where P is in watts, E is in rms volts and R is in ohms). A VTVM or FET VM and an rf probe are used to learn the rf voltage across R_L. In the example, we are able to read 150 V across R_L, our 50-ohm load. This is equivalent to a power output of 450 W. The rf probe rectifies the rf (ac) voltage and converts it to dc voltage so that we can read it directly on the meter of our VTVM.

If our readings are to be accurate, we must be certain the rf energy from the transmitter appears as a classic sine wave. That is, the wave form must not be distorted or contain a high amount of harmonic current. Most modern transmitters are very clean with respect to the output wave form, so no need to be alarmed about the accuracy of our readings.

Details for building our own rf probe are seen in Fig. 3. This circuit has appeared for many years in the ARRL *Handbook* and other League publications. It is being repeated here for the benefit of those who are new to Amateur Radio and may not have the reference material from which to obtain this information. The probe is contained in a metal case (brass or aluminum tubing), and the cable between it and the VTVM can be small-diameter coaxial line. RG-174/U subminiature coaxial cable is lightweight and flexible, and is excellent for this purpose. The larger cables are stiff and heavy, and may easily become the "tail that wagged the dog" if they are used! The voltmeter with which the rf probe is used should have an input characteristic of 10 megohms or greater for best accuracy of rf-power readings. One of the older Heath VTVMs or an RCA Voltohmist™ an be purchased inexpensively if you don't have a meter of this variety. Keep an eye on the *QST* Ham Ads for possible bargains, and don't forget the flea markets at hamfests.

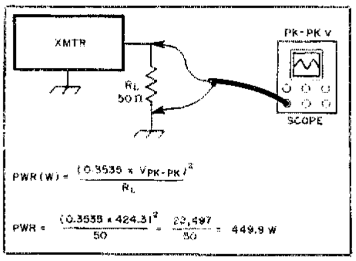


Fig. 4 — Procedure for measuring rf power with a scope.

The measurements chapter of recent editions of the *Handbook* contains directions for building your own FET VOM, should you be a "purist" of the first magnitude!

Using a Scope to Measure Power

We're in an entirely different badminton court when it comes to using a scope for rf-power measurements. Why? Because a scope responds to peak-to-peak (pk-pk) ac voltage rather than rms (root mean square) voltage. This must be taken into account when performing the type of test that can be made with a VTVM. The rf sampling technique remains unchanged, but the math we must employ is a trifle different. Fig. 4 illustrates the general method for using an oscilloscope.

Before we continue, a word or two about scopes is important. The scope must be capable of displaying the operating-frequency wave form of our transmitter accurately. Therefore, the rated *bandwidth* of the scope is important to know. If the real-time bandwidth is specified as 30 MHz, we're all set for hf-band power measurements. But, if the specs state that the instrument is for dc to, say, 5 MHz, our highest ham band for measurement of power will be 75/80 meters. This does not mean that we won't see a wave form above 5 MHz; it merely implies that the accuracy of our measurements will have "flown out the window" above 5 MHz.

Another point worth considering is that a scope tube can't display the voltage with as fine a resolution as we can obtain with the analog meter on our VTVM. The deflection may fall between or beyond the lines on the scope-tube graticule, which calls for interpolation of the precise pk-pk voltage. For most amateur purposes, this will be satisfactory (depending on how good your "interpolator" is!).

Calculations based on the pk-pk voltage are still as simple as those founded on rms voltage. The difference is that we must add a decimal factor to convert the pk-pk voltage to an rms value. All we need do is multiply the pk-pk voltage by 0.3535 to obtain that value. The equation in Fig. 4 shows how this is done when the transmitter power is being measured. Here we assume that the pk-pk voltage measured is 424.3, which becomes 150 V (rms) when multiplied by 0.3535. From this point on, we follow the same calculation procedure called out in Fig. 2. Simple, eh?

Measurements with the Antenna Connected

We can follow these same routes to power measurement, even when we have the antenna connected to the transmitter. How? Well, all that's necessary is to have the system look like 50 ohms during the measurements. If an SWR condition exists, we may insert a Transmatch (or antenna coupler or tuner, as some call them) between the sampling point and the antenna. This will require an SWR indicator in the line, as shown in Fig. 5. The Transmatch is adjusted to disguise the SWR condition, providing an SWR of 1:1. The rf ammeter can now be relied on to render meaningful data for our power-output calculations. In other words, even though the antenna may not allow the system to look like 50 ohms down in the shack, the transmitter and rf ammeter "think" they are in a 50-ohm system, and that's all we care about. It must be remembered that any SWR meter we use should be contained in a feed line that has the same impedance as that for which the instrument was designed. That is, we don't put a 50-ohm SWR indicator in a 75-ohm feed line — not if we want our SWR readings to be accurate.

The same rule applies when sampling the 50-ohm line with an rf probe or scope (Fig. 6). One method for providing a sampling point is the use of a coaxial T-connector.

This permits a through connection to the SWR indicator, while making a test point available for power measurements. Again, the Transmatch must be adjusted to disguise any reflected power on the feed line. Power measurements can be made when the SWR is 1:1.

The Classic RF-Power Bridge

A potentially accurate device for measuring rf-power output and SWR was designed by Warren Bruene (pronounced "brine"), W5OLY, of Collins Radio. It was described in April 1959 *QST*, and has remained the standard circuit in most commercial directional wattmeters for Amateur Radio use to this day. A circuit diagram of that instrument is shown in Fig. 7. Various adaptations of the basic design can be found in December 1969 *QST*.

A toroidal transformer serves as the heart of the instrument (T1 of Fig. 7). The transmission line is routed through the center of the toroid, minus the shield braid, and it serves as the primary of the transformer. D1 and D2 sample forward and reflected rf components on the feed line to enable the user to take SWR readings. C1 and C2 are used to balance the bridge when it is terminated in 50 ohms. R3 through R6, inclusive, provide calibration for the desired power ranges in the forward and reflected positions of S2. The calibration chart at the right in Fig. 7 will yield fairly accurate rf-power readings if a 200-μA dc meter is used at M1. Precise calibration can be effected by placing a laboratory-grade rf meter in series with the circuit of Fig. 7 and noting the reading on M1 at various power levels. A Bird Thruline wattmeter is excellent for the purpose, if you can borrow one for the tests.

Fig. 8 contains a nomograph for determining the SWR when using the circuit of Fig. 7. Alternatively, if you know the values of your forward and relected powers you can determine the SWR from

$$SWR = \frac{1 + x}{1 - x} \qquad \text{(Eq. 1)}$$

where

$$x = \sqrt{\frac{PWR\ (ref.)}{PWR\ (fwd.)}} \qquad \text{(Eq. 2)}$$

This grade-school algebra should be easy to work with, especially in this day of calculators! In fact, it is by no means formidable when doing it longhand.

A Practical Power Meter

The QRP enthusiast is usually forgotten when commercial power meters are designed. That is, we seldom find one with a low-power scale.

Our workshop project this month is a resistive rf power bridge that will safely handle power levels up to 10 W without damage to the bridge resistors. Momentary measurements (very momentary!) of powers up to 20 W are possible. The circuit of the power meter/SWR indicator is given in Fig. 9. Circuit boards or a kit version of this instrument are available.[1]

Parallel combinations of noninductive 2-W resistors are used for R1 through R6, inclusive. This provides ample power capability in the bridge arms without damage to the resistors. Each of these resistors is 100 ohms, which in the parallel format yield 50 ohms.

A 50-ohm dummy load is needed for calibration of the instrument. We may use an available 50-ohm load, such as a Cantenna, or we can build one from a group of 1-W composition resistors. An example of such a resistive load is shown in the photograph of Fig. 10. Here, we have a "resistor sandwich" composed of two pieces of pc board and 20 resistors. Ten 1200-ohm and ten 820-ohm resistors can be used to develop a 10-W, 48.7-ohm load. This is close enough to 50 ohms for our purpose. A coaxial connector of your choice can be connected to the two wires of the load (keep the leads short).

A pc-board potentiometer, R9, is used to calibrate the meter (M1) for 20 W at full scale. Other full-scale power levels can be

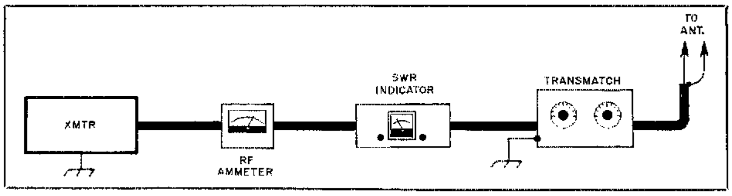


Fig. 5 — If the antenna does not present an SWR of 1:1 to the transmitter, a Transmatch can be used to disguise the SWR and accurate rf-power readings will be possible.

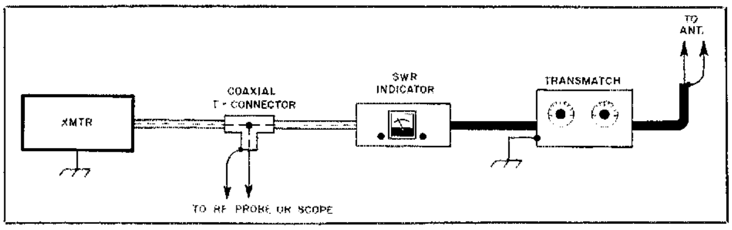
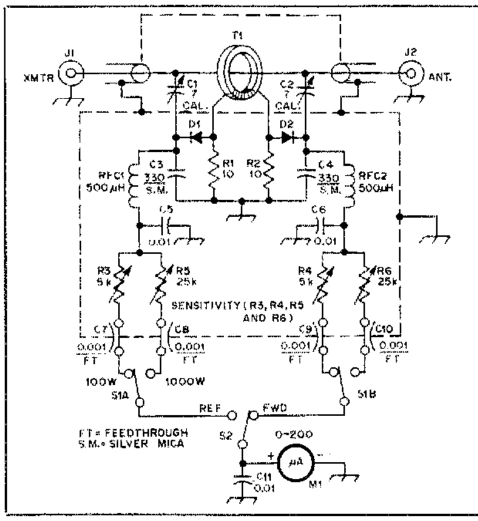


Fig. 6 — Sampling method for power measurement with a scope or rf probe. A coaxial T-connector is inserted in the 50-ohm line between the transmitter and the SWR indicator.



WATTS	M1	WATTS
100	200	1000
90	180	900
80	170	800
70	155	700
60	145	600
50	125	500
40	105	400
30	85	300
20	65	200
10	40	100
5	20	50

EXCEPT AS INDICATED, DECIMAL VALUES OF CAPACITANCE ARE IN MICROFARADS (μF); OTHERS ARE IN PICOFARADS (pF OR μμF); RESISTANCES ARE IN OHMS; k = 1000

Fig. 7 — Circuit of one version of the Bruene power meter. A toroidal transformer serves as a sampling device. The calibration scale at the right is representative of the meter readings versus power when a 200-μA dc meter is used. R3 through R6, inclusive, are thumbwheel-adjust controls inside the instrument case. D1 and D2 are 1N914 silicon diodes.

[1]Notes appear on page 39.

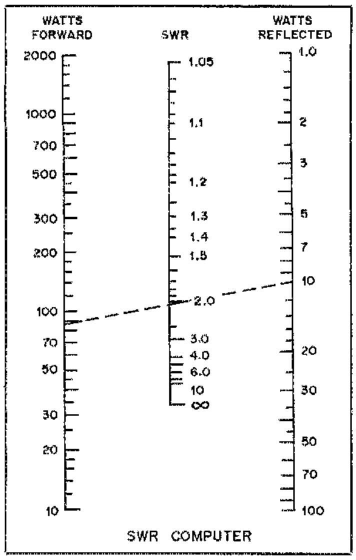


Fig. 8 — Nomograph for determining SWR when the forward and reflected power amounts are known. The dashed line shows an SWR of 2:1 when the forward power is 90 W and the reflected power is 10 W.

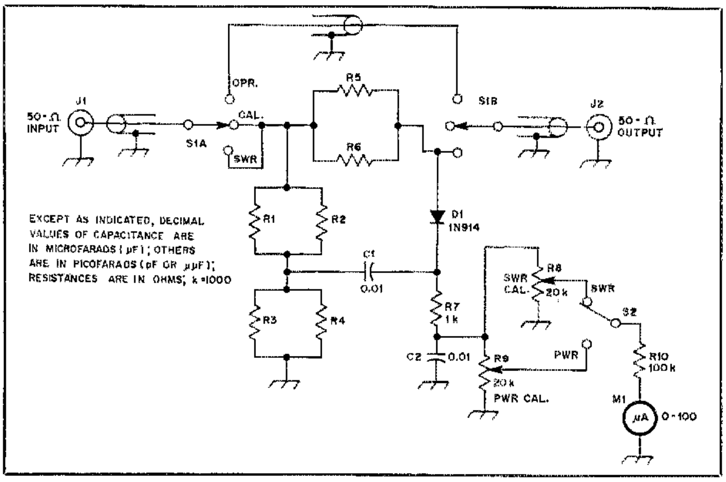


Fig. 9 — Schematic diagram of the resistive power bridge. Capacitors are disc or chip ceramic types. R1 through R6, inclusive, are 100-ohm, 2-W composition (noninductive) resistors. S1 is a two-pole, three-position phenolic wafer switch. S2 is a spdt toggle switch. R8 is a 20-kΩ linear-taper control with shaft (panel mounted). A pc-board control is used for R9. R7 and R10 are 1/2-W composition types. R10 aids the meter linearity. It can be removed if greater sensitivity is needed.

had by adjusting R9 accordingly. I chose a 20-W scale because of the surplus meter I obtained; it had a 0-20 W scale, plus an SWR scale.[2]

The SWR function of the instrument in Fig. 9 can be calibrated by means of a panel-mounted control, R8. The operating modes — OPR, CALIBRATE and SWR — are chosen by switch positioning of S1. The bridge is bypassed in the OPR mode. We may expect accurate power readings (within 10%) from 1.8 through 30 MHz with this circuit. Tests with a Bird Thruline wattmeter confirmed the accuracy. Two less-expensive meters are suitable for use in this bridge (Fig. 11) (see note 2), but you will have to develop your own calibration scale if one of them is used.

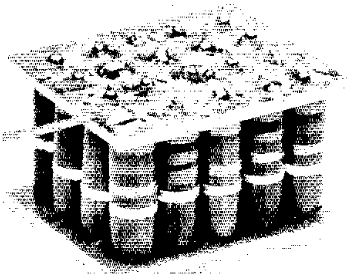


Fig. 10 — A homemade low-power 50-ohm load (see text).

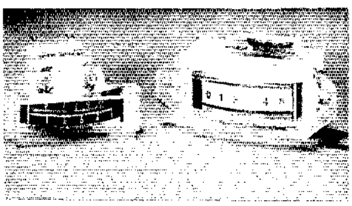


Fig. 11 — Low-cost surplus fm tuning meters can be used at M1 of Fig. 9.

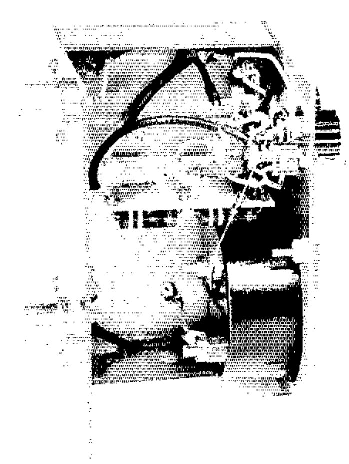


Fig. 12 — Interior view of the QRP power meter. W7ZOI "ugly construction" is used.

Construction Notes

The photograph on the title page of this article shows how the instrument is laid out. A green panel was chosen (pc board panel coated with Hunter Green spray-can paint) for use with green Dymo tape labels. This provides a reasonably professional appearance for the finished product. The U-shaped cover is fashioned from aluminum sheet and is painted gray. The kit version of this instrument is somewhat smaller than the W1FB version. Dimensions for the latter unit are (HWD) 3 × 4-1/2 × 3 inches (76 × 114 × 76 mm).

An interior view of the meter is seen in Fig. 12. The bulk of the components are contained on a pc board that is mounted vertically in the center of the box (Fig. 13). Miniature coaxial cable (RG-174/U) is used for all rf leads between the circuit points that we must join. The shield braid of the cables should be grounded at both ends. A pair of no. 6 spade bolts is used to secure the pc board to the base of the cabinet.

Our cabinet is made entirely of pc-board sections (single-sided board). The walls and bottom are joined by means of soldered seams. A 40-W or greater pencil type of iron is best for this work. It will ensure ample heating of the surfaces and will result in an even flow of solder.

Although SO-239 uhf connectors are used in my model of the bridge, other styles of rf connector can be used. It would be entirely suitable to use single-hole-mount RCA-style phono jacks, BNC connectors or CATV hardware. The use of phono jacks would certainly reduce the cost of this project. A parts-placement guide for the pc-board module of the instrument is provided in Fig. 15.

Bridge Setup and Use

A 50-ohm dummy load is connected to J2 of Fig. 9 during our testing of the assembled unit. A low-power transmitter, or a medium-power one with the drive turned down to provide a few watts of output, is connected to J1. S1 is placed in the CAL position, and R8 is set at midrange. Rf power is applied slowly to the bridge until a full-scale reading is noted at M1. Next, we will switch S1 to the SWR setting. If the circuit is functioning properly, the needle of M1 will drop to zero or nearly so. In the OPR position of S1, we can bypass the bridge during the operating period. Irrespective of the power output of your transmitter (within the safe rating of the bridge), the SWR adjustment remains as just described. In other words, set R8 for a full-scale reading of M1, then switch to the SWR mode and note the meter indication. If the meter does not display a zero reading, adjust the antenna matching device or Transmatch until the meter reads zero. That will indicate an SWR of 1:1.

Our next step calls for calibration of the instrument for reading rf power. First we must decide what the full-scale power range will be. It can be 0-1, 0-5 or 0-10 W, if we choose. If a calibrated power meter is available from a friend, the task will be a simple one. That instrument will be connected between our transmitter and J1 of Fig. 9. A 50-ohm dummy load will be attached to J2. The desired full-scale power is applied to the instrument, and the internal potentiometer, R9, is adjusted to yield a full-scale reading at M1. The transmitter power is next reduced in increments of 1 or 2 W, and the meter face is marked accordingly. Upon completion of the calibration exercise, we can, if we wish, make a new meter face and paste it over the original one. Fortunately, the meter shown in the photograph of my unit tracks very nicely with the circuit of Fig. 9 for a 20-W full-scale reading. The bridge resistors can't take a sustained 20-W power level without damage, however, so I make measurements in that range with short key-down bursts. A steady power of 10 W is not suggested either, because the resistors can change value over time at that power level. It is best to make our measurement periods short (a few seconds) when the power exceeds 5 W.

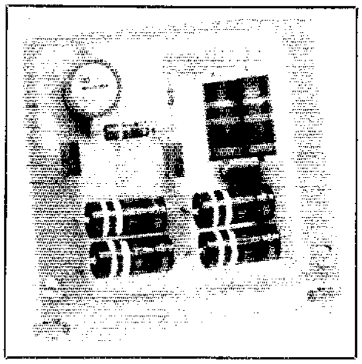


Fig. 13 — The pc board that contains most of the parts for the bridge portion of the instrument.

If you're unable to borrow a calibrated rf-power meter, you may plot your meter scale by following one of the simple procedures described earlier in this article: A dummy load can be connected to J2 of Fig. 9 and an rf probe with VTVM used to calculate the power through the load.

Some Final Comments

The characteristics of the diode used at D1 of Fig. 9 can have an effect on the relative linearity of the meter scale. Different types of small-signal diodes may have dissimilar conduction profiles, thereby resulting in slightly different conduction curves — especially at very low power levels.

A meter with a movement of 500-μA dc or less will be best for use in this circuit. Many of the surplus fm tuning meters (Fig. 11) have movements of approximately 200μA, making them suitable for use in this circuit. They can be taken apart to permit pasting on a new meter face. If the meter movement is greater than 500 μA, we may be unable to obtain full-scale deflection at very low power levels.

For the time being, at least, the QRP person can make accurate power measurements by building his or her own inexpensive instrument. The techniques are uncomplicated. If you haven't measured the output power of your rig, this article will show you how it's done. Good luck with your workshop project!

Notes

[1]Pc boards or complete parts kits for the QRP power meter are available from Circuit Board Specialists, P.O. Box 969, Pueblo, CO 81002, tel. 303-542-5083. Catalog available.

[2]The meters shown in this article are available from Surplus Electronics Corp., 7294 N.W. 54th St., Miami, FL 33166, tel. 305-887-8228. Catalog available.

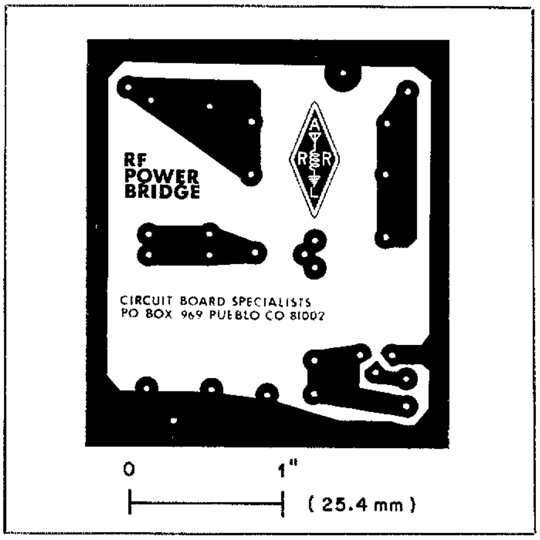


Fig. 14 — Circuit-board etching pattern for the RF Power Bridge. The pattern is shown full size from the foil side of the board. Black areas represent unetched copper foil.

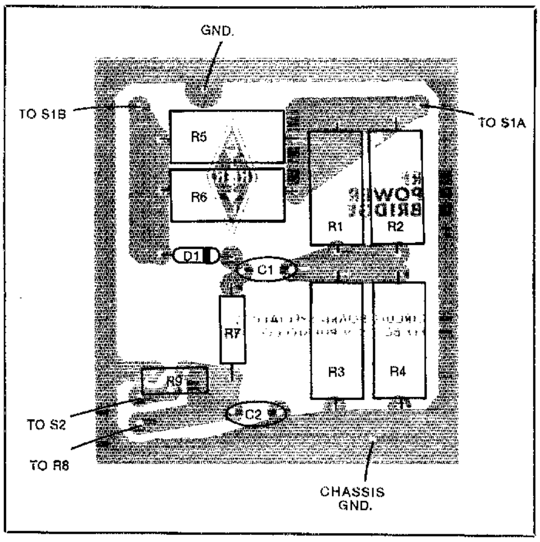


Fig. 15 — Parts-placement guide for the pc board of Fig. 14. Components are mounted on the non-foil side. Gray areas represent unetched copper.

Hints and Kinks

Conducted By Larry D. Wolfgang,* WA3VIL

A SIMPLE LOGIC PROBE

☐ Many projects use logic circuits to perform a variety of tasks. Studying the operation of such circuits, or troubleshooting them, does not require a measurement of the actual circuit voltages. It is only necessary to know if a logic 1 or a logic 0 is present. Fig. 1 shows a simple but effective logic probe that I built as a weekend project. The components are easy to find, and the cost for all new parts should be less than $10. Most hams will be able to find many of the parts in their junk box.

This is a high-impedance probe. It will not load down the circuit under test. Input to the probe is through a 400-kΩ resistor to two transistors wired as a Darlington pair. Output from the Darlington amplifier is fed to a 7404 hex inverter, which is used to drive appropriate segments of a common-anode, 7-segment LED display. The complete schematic diagram is shown in Fig. 2.

Power for the probe is obtained from the circuit under test. Batteries could be used, but they would add to the size and weight of the unit.[1] I used an etched circuit board for my probe, but the layout is not critical; any construction method should work fine. Fig. 3A is a circuit-board etching pattern and Fig. 3B is the parts-placement diagram. Most small-signal npn transistors will work for Q1 and Q2. I made the probe tip from a 2-1/2 inch length of no. 12 copper wire soldered to the circuit board.[2,3] I covered this with the plastic barrel from an old ballpoint pen.

If the probe detects a ground (logic 0) or no connection, the top half of the readout will light, forming a small 0. If a positive voltage is applied to the probe tip, a 1 will be displayed. If the probe tip detects a pulse, a P will light up. The P will appear to flash at the pulse rate, but you should be aware that if the pulse duty cycle is much different from 50% a true indication will not be given.

I have found this logic probe to be a useful device for checking digital-circuit operation. I am sure it will be a welcome addition on your test bench. — *Robert Crawford, WA6RYZ, La Mesa, California*

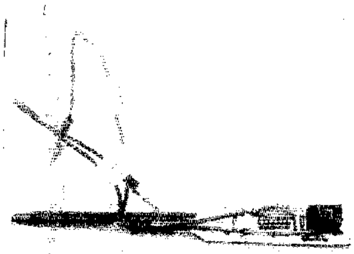


Fig. 1 — Photo of a simple logic probe built by WA6RYZ.

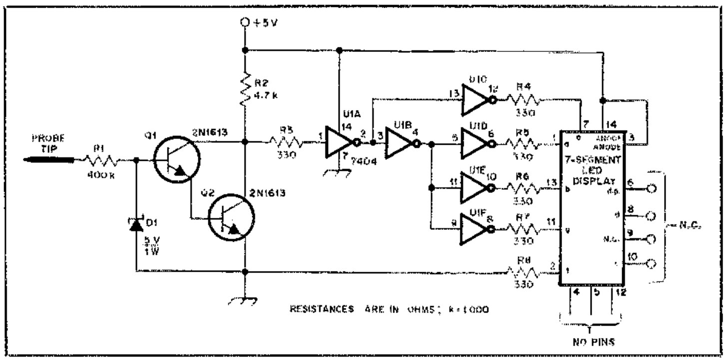


Fig. 2 — Schematic diagram of the simple logic probe. Operating power is taken from the circuit under test.

Q1, Q2 — 2N1613 or other small-signal npn transistor, such as 2N2222.

U2 — Common-anode, 7-segment LED display, such as Radio Shack part no. 276-053.

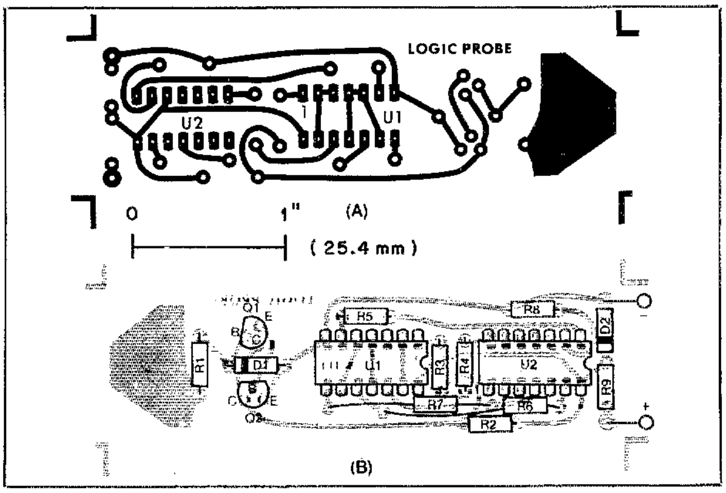


Fig. 3 — A full-size circuit-board etching pattern is shown at A. Black represents unetched copper, viewed from the foil side of the board. A parts-placement diagram is shown at B. This view is from the component side of the board, with gray areas representing an X-ray view of the remaining copper. R7 and R8 are mounted on the foil side of the board. This pattern includes a 5-V Zener diode (D2) and 150-Ω current-limiting resistor (R9) for the power leads.

[1][Editor's Note: As shown, the probe can only be used with circuits that have a 5-V supply. In the unit I built I wired a 5-V Zener diode across the power connection, and brought the positive supply lead to the board through a 150-Ω current-limiting resistor. This enables me to use the probe on circuits with up to a 12-V supply. The two extra parts are soldered to the appropriate circuit pads on the foil side of the board.]

[2]mm = in. × 25.4.

[3]Etched circuit boards and complete parts kits are available from Circuit Board Specialists, P.O. Box 969, Pueblo, CO 81002 and from RADIOKIT, Box 411, Greenville, NH 03048.

*Assistant Technical Editor

LOGIC LEVELS FROM THE TRS-80® MICROCOMPUTER CASSETTE OUTPUT PORT

☐ The circuit for generating TTL levels from the Radio Shack TRS-80® Model I computer cassette output port, shown in the September 1982 Hints and Kinks column, has a significant drawback. The open-collector nature of the LM339 comparator may not properly force a TTL high. This can be solved by adding a 2.2-kΩ pull-up resistor between the output (pin 1) and V_{cc} (pin 3). Fig. 4A shows an alternative circuit

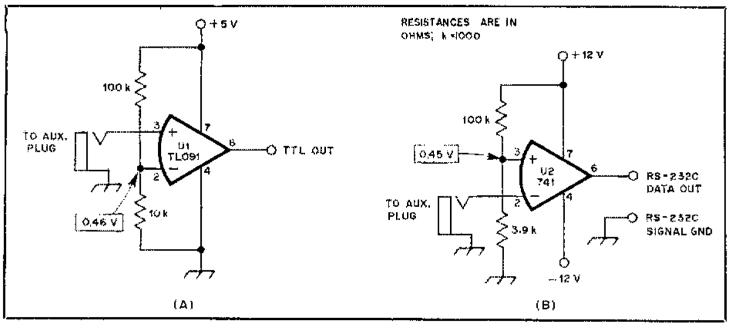


Fig. 4 — A circuit for developing TTL levels from the TRS-80® microcomputer cassette output port is shown at A. A similar circuit to develop RS-232C levels is shown at B.
U1 — TL091 op amp, Radio Shack part no. 276-1745.

using a TL091 op amp. It does not need a pull-up resistor.

A similar circuit can be made to drive an RS-232C line, if a dual-voltage power supply is available. Fig. 4B shows this arrangement. If you do not have a dual-voltage supply, two 9-V transistor-radio batteries (connected in series), with the common terminal between batteries as ground, will work with no modifications to the circuit. If batteries are used, changing the 3.9-kΩ resistor to a 4.7-kΩ unit will give slightly improved performance. Note that the LM741 is wired in the inverting mode because the RS-232C convention is for a voltage between −5 and −15 to represent logic 1 (mark) and for a voltage between +5 and +15 to represent logic 0 (space).

With both circuits, the statements "OUT 255,1" and "OUT 255,2" can be used to develop a logic 1 and logic 0, respectively. The logic states set by these statements will not change unless another OUT statement is executed, or the program in which the original statements are executed stops running. — *Michael Bilow, N1BEE, Cranston, Rhode Island*

INCREASING TUBE LIFE WITH A THERMOMETER

☐ Some operators find that the final amplifier tubes in their rigs will last for many years, while less-experienced hams will have to buy several sets of tubes in the same time. One important aspect to maintaining long tube life is to prevent the tube temperature from getting too high. My solution to this problem is to install a thermometer in the final-amplifier compartment. Now I can check the ambient temperature of the tubes. This gives an indication of how long I can hold the key down during tune up, and how effective the cooling fan is. Others who have tried this idea report that their tubes now last up to 100 times as long as before! I hope this information may help other amateurs keep their tubes operating longer. — *Jan Martin Noeding, LA8AK, Vaagsbygd, Norway*

MAGNETIC SWITCH FOR CW TUNE UP

☐ I use a Curtis Lil' Bugger keyer for cw operation. This keyer does not have a tune position, so there is no convenient method to produce a steady signal to tune my transceiver. Fig. 5 shows how I use a reed switch and a small magnet to provide a tune function. My transceiver requires a 1/4-inch phone plug for the key line, so I wired the switch into that. You can use whatever type of plug your rig needs. By holding a small magnet next to the plug, you can key the transmitter. Remove the magnet when you complete the tune-up operation and you are ready to go. — *Rick Lucas, WBØNQM, Lawrence, Kansas*

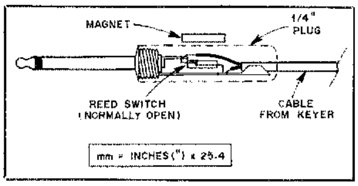


Fig. 5 — WBØNQM uses a reed switch and a small magnet to provide a tune function with his electronic keyer.

LIDS FOR PC-BOARD PROJECT CASES

☐ Adding a screw-down lid to a project case made from pc-board material can be a problem, especially when you try soldering a nut into the corner of the box. I find that a better way to fasten the lid is to use a piece of 1/8-inch brass or copper tubing. Hold one end in a vise and thread about a 1/2-inch length using a no. 4-40 tap. A little tap fluid may help. (The 4-40 screw is a bit small for 1/8-inch tubing. A no. 5-40 screw is the right size. This size tap will cut a full-depth thread in the tubing. — Ed.] You can cut off this section with a utility knife, by rolling the tubing on your workbench. A 1-foot piece of tubing will provide more than 20 "nuts." They line up easily in small corners, solder quickly in place and take up little space (Fig. 6). — *Robert Dixon, W8HGH, Rochester, Michigan*

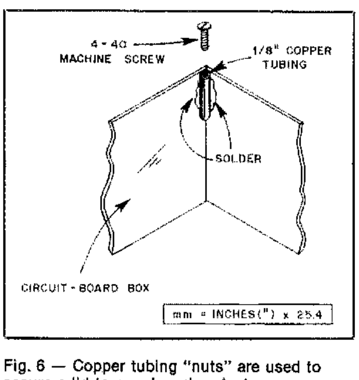


Fig. 6 — Copper tubing "nuts" are used to secure a lid to a pc-board project case.

WEIGHTED TUNING DIAL FOR SMOOTHER OPERATION

☐ My Kenwood TS-830S tuning mechanism had a nice feel, but I performed a simple modification that made tuning it even smoother. I removed the tuning knob and found a large groove around the inside edge. I simply added weight to the knob by carefully melting solder and letting it drip off the tip of the soldering iron into the groove in the knob (Fig. 7). I built up several layers of solder, allowing each one to cool before adding the next. Be careful that you do not block the set-screw holes with solder in the process! [Lead shot or BBs and white glue or epoxy also work well. — Ed.]

I reinstalled the knob after the solder had thoroughly cooled. The feel of the tuning dial is even nicer now, and a spin of the knob sends the VFO rapidly up or down the band. I am sure this modification can be made to many other rigs just as easily. — *Donald Cottingham, VE3HXY, Burlington, Ontario*

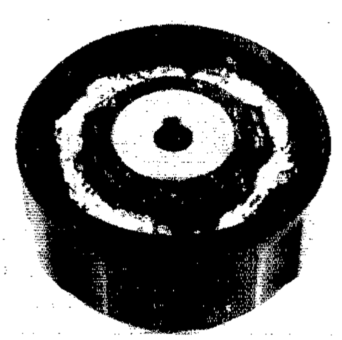


Fig. 7 — Photo showing the solder that VE3HXY added to the tuning dial of his TS-830S to provide smoother tuning.

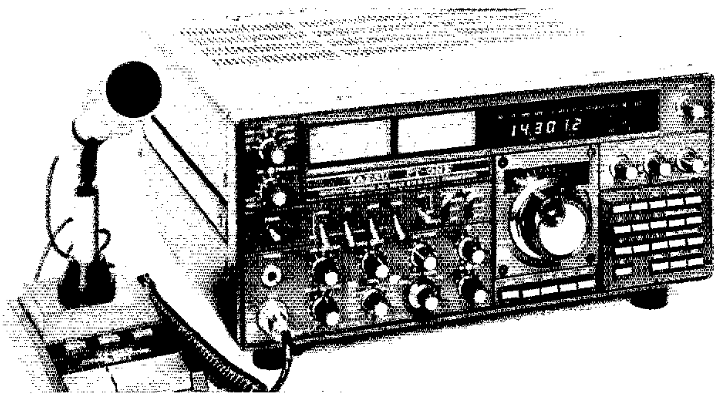
RFI AND CAPACITIVE-TOUCH PADDLES

☐ I built the CMOS Iambic Keyer and added an Ash-Proof Keyer Paddle.[4,5] I soon found that being within 100 feet of an antenna, using anything more than 50 W, caused erratic operation. I added a 0.75-mH rf choke to each of the dot and dash lines from the touch pads. (Ferrite beads on these lines might also work.) The chokes eliminated the problem except during 20-meter operation. I also connected 100-pF capacitors from the pads to ground. Sensitivity adjustment is more critical if larger values of capacitance are used. These changes have solved my RFI problems. — *Rolla Wade, KF4TF, Kingsport, Tennessee*

[4] T. Theroux, "A Digital CMOS Iambic Keyer," *QST*, June 1982, pp. 26-28.
[5] R. Lewallen, "An Ash-Proof Keyer Paddle — Something New for CW Operators!," *QST*, Aug. 1981, pp. 30-31.

Conducted By Paul K. Pagel,* N1FB

Yaesu FT-ONE HF Transceiver



One look at the FT-ONE and you know that it is not the "average" transceiver! There is such an array of buttons, knobs, switches and displays that there is no empty space on the front panel. My first thought upon receiving the review model was to hook up an antenna, plug it in and start tuning around. After a second thought, I decided to read the operating manual instead; I'm glad I did!

The clearly written manual does an excellent job of telling you what you need to know. For example, almost one page is dedicated to the operation of the variable bandwidth and i-f shift features. By studying the large accompanying diagram, one can easily grasp how these controls operate and how to use them to best advantage. I spent about an hour reading and digesting the contents. The time was well spent! I noticed later that people unfamiliar with the FT-ONE were not able to take advantage of the many features that make it a joy to use.

A Peek at the Panel

Two meters are found on the front panel. The one on the right is an S meter during receive; in the transmit mode, it measures alc level. The other meter is controlled by a switch in the upper left corner of the panel; with it you can measure the final amplifier collector supply voltage and current, discriminator voltage (fm mode), compression level of the built-in speech processor, relative forward output power and SWR. Below the meter switch is a pair of concentric controls that set VOX and relative forward power sensitivity.

Below the meters is a row of toggle switches. Those labeled POWER, PROC and NB are used to turn on and off the main power, the rf speech processor and the noise blanker, respectively. The automatic microphone gain control (AMGC) switch enables a microphone gain threshold circuit. This circuit acts as a "microphone squelch" system. The microphone amplifier is not enabled until a minimum input level is present. Speech monitoring and cw sidetone are controlled by the MONI switch. An audio filter is controlled by the APF/NOTCH switch. In the APF position, it behaves as a cw audio filter; in NOTCH, an interfering carrier can be greatly reduced in level. Filter center frequency is controlled in the 300- to 1500-Hz range by a knob just below the switch. The AGC switch selects recovery time for the receiver agc circuit; you can choose fast or slow, or turn off the agc if you desire. The SCAN switch selects the desired stop mode during scanner operation. When this switch is placed in the AUTO mode, the scanner will halt on any signal that is strong enough (S1 or greater) to cause agc action. To stop the scan manually while in the AUTO mode, place the SCAN switch in the STOP position momentarily. Place the SCAN switch in the MAN position for manual scanner operation; the scanner will be activated when the UP or DOWN switches on the keyboard or microphone are pushed. Release the switch to halt the scan.

The lower-left portion of the front panel contains two jacks. When a standard two-conductor phone plug is inserted into the PHONES jack, the internal speaker is disconnected automatically. The headphone impedance should be 4-8 ohms. An eight-pin MIC connector accepts the microphone audio input, as well as the push-to-talk (PTT) and scanning control lines. Nominal microphone impedance is 600 ohms.

To the right of the jacks is a series of knobs. Gain of the microphone amplifier during ssb and a-m operation is set by the MIC control; a concentric control labeled COMP is used to adjust the compression level of the rf speech processor. Another concentric control is used to set the speed of an optional internal electronic KEYER and to adjust VOX hang time DELAY. When the DELAY control is rotated fully counterclockwise into the click-stop, the transceiver is ready for QSK (break-in) cw operation. The DRIVE control is used to set rf levels on a-m, fm, cw, fsk and on ssb when using the rf speech processor. (When the processor is switched out, the MIC control determines rf levels.) When the noise blanker is switched on, turning the NB knob clockwise lowers the threshold, causing the blanker to be more sensitive to impulse noise.

The MODE switch has 11 positions for the selection of operating mode and optional filters.

Yaesu FT-ONE HF Transceiver, Serial No. 040145

Manufacturer's Claimed Specifications	*Measured in ARRL Lab*
Frequency coverage: Receive — 150 kHz to 29.9999 MHz; transmit — 1.8 to 2.0, 3.0 to 4.0, 7.0 to 8.0, 10.0 to 11.0, 14.0 to 15.0, 18.0 to 19.0, 21.0 to 22.0, 24.0 to 25.0, 28.0 to 29.99 MHz.	As specified.
Modes of operation: cw, ssb, a-m (fm optional).	As specified.
Frequency display: Six digit.	3/8-inch high, six-digit yellow LED.†
kHz/turn of knob: 10 MHz/20/2.	As specified.
Frequency resolution: 100 Hz.	As specified.
Backlash: Not specified.	Nil.
S-meter sensitivity (μV/S9 reading): Not specified.	160 m, 9; 80 m, 8.5; 40 m, 8.5; 30 m, 8.4; 20 m, 10; 17 m, 8.8; 15 m, 9.1; 12 m, 9.0; 10 m, 9.1.
Transmitter output: 160-15 m, 100-W PEP; 10 m, 90 W.	160 m, 120; 80 m, 125; 40 m, 120; 20 m, 120; 15 m, 125; 10 m, 110.
Harmonic suppression: Better than 50 dB.	−53 dB (see photo).
Spurious suppression; Better than 40 dB.	−65 dB (see photo).
3rd-order IMD: Better than 31 dB.	−38 dB (see photo).
Receiver sensitivity: Less than 0.3 μV for 10-dB S + N/N.	Receiver dynamics measured with optional 300-Hz filter installed:
	80 m / *20 m*
	Noise floor (MDS) dBm: −133 / −138
	Blocking DR (dB): noise limited
	Two-tone 3rd-order IMD DR (dB): noise limited
	3rd-order intercept: noise limited
Size (HWD): 6.2 × 14.6 × 13.8 in.	As specified.
Weight: 37.5 lb.††	As specified.

†mm = in. × 25.4. ††kg = lb × 0.4536.

*Assistant Technical Editor

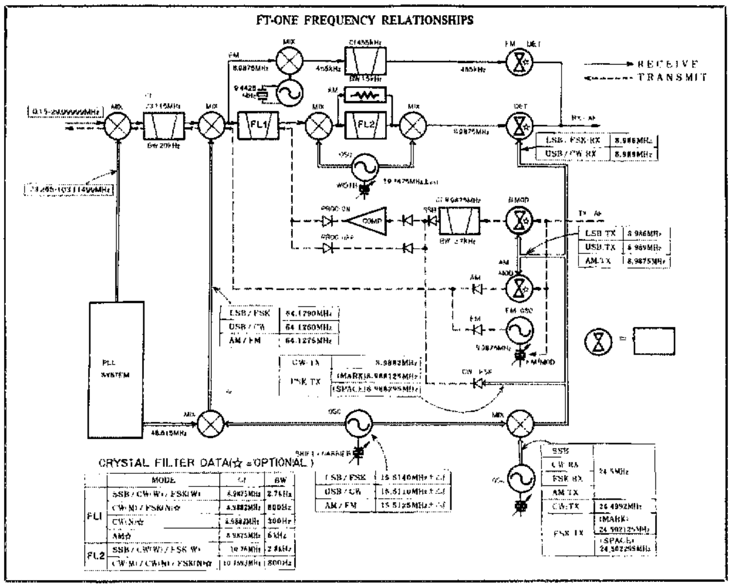


Fig. 1 — Block diagram of the FT-ONE showing frequency relationships.

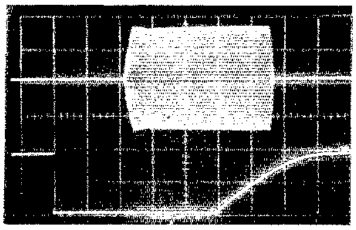


Fig. 4 — FT-ONE cw keying waveform. Upper trace is the rf envelope; lower trace is the dc level at the key jack. Each horizontal division is 5 ms.

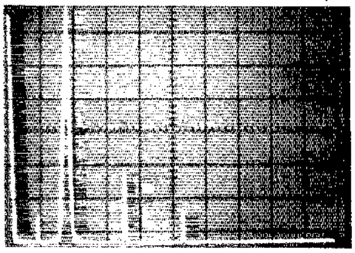


Fig. 2 — Worst-case spectral display of the FT-ONE. Vertical divisions are each 10 dB; horizontal divisions are each 2 MHz. Output power is 120 W at 1.8 MHz. All spurious emissions are at least 53 dB below the fundamental output. The FT-ONE complies with current FCC specifications for spectral purity.

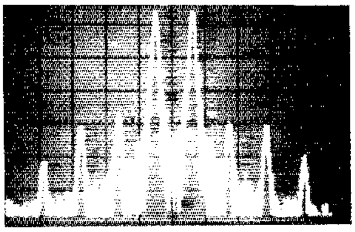


Fig. 3 — Spectral display of the FT-ONE during transmitter two-tone IMD test. Third-order products are 38 dB below PEP output, and the fifth-order products are 40 dB down. Vertical divisions are each 10 dB; horizontal divisions are each 1 kHz. The transceiver was operated at rated output power on the 160-meter band.

You can even choose RX-U/TX-L (receive usb, transmit lsb) or RX-L/TX-U. The value of that feature may not be obvious unless you plan to operate through the AMSAT Phase IIIB satellite (OSCAR 10). What goes up on usb (Mode J or L) comes down as lsb. A tip of the hat to the folks at Yaesu for that little extra that will add much to the value of the FT-ONE.

The SHIFT and WIDTH controls are mounted on concentric shafts. Moderate friction between them allows fingertip adjustment of the i-f shift and variable bandwidth features. The controls are normally aligned so that the black zone on the inner dial is fully within the cutout area of the outer (paddle) control. If interference is heard on ssb, narrow the bandwidth by holding the center knob and rotating the paddle control. A narrower bandwidth has now been set. Use the paddle to align the passband for optimum reception and interference reduction.

Receiver rf and i-f amplifier gain is varied by the RF GAIN control. Audio output level is set by the AF control. In the fm mode, the SQL control keeps the receiver silent until a signal is present.

Frequency Control

The transceiver main tuning knob can be used to control the VFO or the clarifier (RIT) frequency. Five push buttons, located directly below, are used in conjunction with the main tuning knob. When the alternate-action FINE switch is in, the main tuning knob will vary the transceiver frequency at a rate of 2 kHz per revolution; in the out position, the rate is 20 kHz per revolution. When the MHz switch is held depressed and the main tuning knob is turned, the tuning rate is 10 MHz per revolution.

Pressing the CLAR switch activates the clarifier, allowing an offset in receive frequency of up to ± 9.9 kHz. Once the clarifier is activated, the transmit frequency cannot be changed. My first impression was that I did not like that feature. After using the FT-ONE for a few weeks, I changed my mind. While this feature can be used for split-frequency operation, it is not the best way to operate "split" — but more on that later. When the clarifier is operating, pressing the RX-TRX button causes both the transmit and the receive frequencies to offset. A second button depression causes only the receive frequency to offset. Pressing the button marked LOCK disables frequency changes by the main tuning knob.

To the right of the tuning knob is the keyboard and three VFO switches. This combination makes a versatile and powerful frequency-selection system. Fig. 5, taken from the FT-ONE operating manual, will give you an idea of how many ways you can select a given frequency. Once you become familiar with the equipment, you will be surprised at how quickly and easily you can change frequency, set up a split or scan a portion of a band.

The RF ATT control, located in the upper-right corner of the front panel, adjusts the receiver front end attenuation. PIN diodes are used in the attenuator. This control is particularly useful on crowded and noisy bands.

A yellow LED display directly over the tuning knob shows the operating frequency. A smaller red LED display to the right shows which of the 10 VFO memories is in use and the amount of clarifier offset (when the clarifier is switched in).

A set of 10 yellow LED indicators is located above the frequency displays. These indicators provide quick verification of various operating functions. When a particular LED is illuminated, the respective function is in operation. The functions include PROC (speech processor), NB (noise blanker), AMGC (automatic microphone gain control), MONI (ssb voice monitor/cw sidetone), APF (audio peak filter), NOTCH (audio notch filter), A. SCAN (automatic scan), RX (receiver clarifier), TRX (receive/transmit clarifier), D. LOCK (dial lock), TRCV (Transceive mode) and TX. DIS (transmitter disabled — lights when attempting to transmit outside an amateur band).

Versatility is the word that describes the front panel; it also describes the rear panel. Antenna connections are designed so you can use an external receiver, a separate receive antenna, a vhf transverter or other external equipment (see Fig. 6).

Other rear-panel jacks can be used to connect an external speaker, an fsk terminal (170-Hz shift), a tape recorder (400 mV at 50 kΩ, fixed level), anti-trip audio (from external receiver), a phone patch, a PTT switch (I use a foot

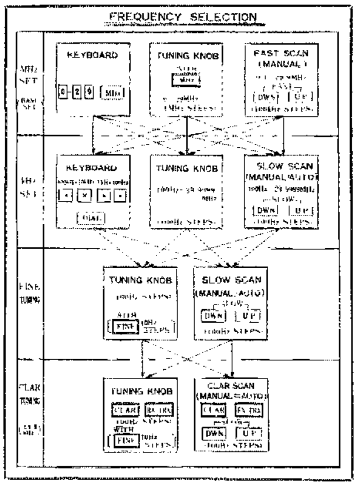


Fig. 5 — This diagram is used in the FT-ONE operating manual to explain the various ways of tuning to a given frequency.

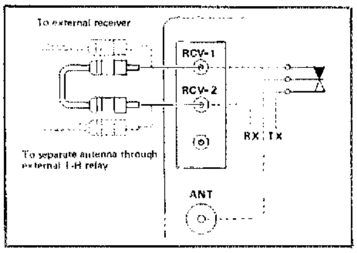


Fig. 6 — Antenna and rf connections on the rear panel of the FT-ONE. In normal operation, a jumper connects RCV-1 to RCV-2. The RF OUT jack provides transmitter output at −6 dBm into 50 ohms for use with a transverter or other accessory equipment.

switch), an i-f signal waveform monitor or sidetone (for external receiver). A pair of key jacks are used to key the transmitter directly (KEY-1) or to operate the optional internal electronic keyer unit (KEY-2). Two DIN-type jacks are used for making interconnections to linear amplifiers, transverters or other accessories.

Personal Observations

I wanted to change the cw offset on the FT-ONE. With other transceivers it has been as simple as setting the RIT control for the desired offset. That won't work with the FT-ONE! A telephone call confirmed the proper procedure; VR 3005 (TX CW) on the LO board sets the cw offset. Complete the job by adjusting the sidetone frequency as described on page 26 of the operating manual.

My linear amplifier has an electronic bias switch. When no rf energy is applied, the tubes are biased to cut off. When rf is first applied, it takes a couple of cycles for the switch to activate. The FT-ONE did not "like" that. When I tried to send a series of dots, the first few were severely level-limited by the transceiver alc. It took a bit of pondering to solve the problem (the FT-ONE worked fine with other amplifiers!). When I disabled the electronic bias switch, the transceiver and amplifier worked together nicely.

I have had the FT-ONE on the air for several weeks, and in a couple of contests. I am pleased with the results. Most reports have been complimentary (twice I was told that the keying was slightly harsh). The speech processor and AMGC help deliver clean, crisp ssb that is clear of background clutter. The joys of QSK must be experienced to be fully appreciated. When I need the amplifier (not very often!), it is only necessary to insert the appropriate delay — my amplifier is not equipped for QSK.

Split-frequency operation is handled with ease by using the VFO select switch in the RA-TB position. Set the A and B VFO channel selector switches for different channels. Tune in the station you want to contact. Set the VFO select switch in the B position. Tune in the frequency you want to transmit on. Return the VFO select switch to the RA-TB position and you are ready to call.

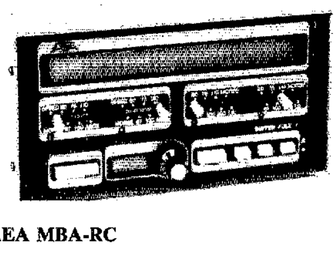
From my rural location, I noticed reciprocal mixing problems only a couple of times — and then not bad. In this regard, the FT-ONE is no different than most of the frequency-synthesized, hf transceivers that have been tested in the ARRL lab.

At the end of the review period, the FT-ONE developed a malady: It would not transmit in the low end of the cw bands. The transceiver was returned to the Yaesu service center. The problem, a bad solder joint on the PLL board, was repaired under warranty.

It is easy to summarize my opinion of the FT-ONE — I like it! It has the flexibility and features that appeal to DXers and contesters alike. Those same features make net and casual operating fun, too. The FT-ONE deserves to be called "top of the line."

Yaesu has two modification kits available for the FT-ONE. The NB-ONE is a noise blanker modification that the manufacturer says will eliminate the "woodpecker." A second kit, the FAN-ONE, changes the operation of the PA heat-sink fan from continuously running to on-demand. That is, the fan operates only during transmit periods or when the heat-sink temperature has risen beyond a specific point.

Price class of the FT-ONE with the 300-Hz cw filter, the 600-Hz cw filter, the 6-kHz a-m filter, the 3rd i-f cw filter, a memory back-up board and the fm board: $2900; KY-ONE keyer price class: $45; dc cable, $15. Price class of the NB-ONE is $10; FAN-ONE, $5.40. — *Chuck Hutchinson, K8CH*



AEA MBA-RC

☐ Probably the best word to describe this μP-based unit is "multifaceted." With it, you can receive and transmit Morse code and Baudot/ASCII RTTY. So, what makes this unit any different from many others on the market? The fact that it can perform code conversions and allow cross-mode operation. You can transmit in any of the modes and speeds available on the 'RC and receive a station using a different mode or speed. You cannot input 300-baud ASCII from your keyboard to the 'RC, but the unit can transmit 300-baud ASCII. (It speed converts 110 baud to 300 baud to do this.) The Baudot speed range is 60, 67, 75 and 100 wpm, while Morse speeds of from 2 to over 80 wpm can be handled.

At the top of the front panel is a 32-character vacuum fluorescent-blue display that offers a built-in monitoring system for reception and transmission. Input and output mode/speed selection switches occupy left and right positions immediately below the display. The bottom row of controls includes RTTY/cw filter selecton switches, a signal-tuning LED bar graph, a variable shift control, carriage return/line feed control switches, an i-d/cw speed set switch, mark/space inversion controls and a power ON/OFF switch. A BUFFER FULL LED sits just above these last switches.

All I/O connections are made at the rear panel. These include power, audio, speaker, tape, scope, keying and printer ports. TTL level output, parallel ASCII, 20/60 mA current loop keyer and an optional RS-232 hookup are available. Some less-often-used controls, such as the speaker and afsk output level potentiometers, and selection switches for downshift on space, printer lockup and 170/850-Hz shift, are there, too. A required external power supply delivering 13-V dc (±3 V) at about 1.2 A connects to the 'RC by means of a miniature coaxial power plug.

Two 3870 single-chip, 8-bit microcomputers are at the core of the MBA-RC. A pair of 74150 16-input multiplexers handle the speed/mode selector switch decoding, and a pair of 2114s provide 1K of RAM. An XR-2206 generates the afsk tones.

Positive and negative key lines are controlled by high-voltage transistors (MPSA42, MPSA92). Speaker tones are created by the ever-faithful 555 IC. Current-loop input and output circuits are optoisolated.

The demodulator section decodes both mark and space tones using op-amp filters. Demodulator performance is good. While there are frequency adjustment potentiometers to set the filters to their proper frequency, the manual contains no maintenance or alignment instructions. It does have three schematic diagrams and three board layout drawings, but that's it.

Three pc boards (including the display board) hold the circuit components. The boards are glass-epoxy types, double-sided and solder masked. Component quality is excellent.

A complete account of the multitude of control functions is beyond the scope of this review. With this machine, you'll want to read the instruction manual first. It's well written and contains plenty of drawings to aid you in equipment setup and operation.

Perhaps I'm spoiled by 24-line, 40- and 80-character displays, but I found I needed some time to get used to the "Times Square" scroll and the character style of the display. Along with the incoming text to be read, there are the special characters used for carriage returns, line feeds, etc. After a while, I learned to disregard (for the most part) those irrelevant characters. The 32-character display certainly makes for an all-in-one unit, and it has its applications, but my eyes found the display difficult to follow except at slower speeds. Reading the display at rates of up to the commonly used 60-wpm Baudot speed was tolerable, but anything higher than that was a bit too fast for comfort.

After a bit of practice, tuning in a signal using the bar graph tuning indicator becomes easy. Of course, there is the option of using the scope outputs to provide the familiar RTTY cross pattern used for tuning.

The most fun I had with the 'RC was using it with my newly acquired (old) model 33 110-baud ASCII printer. The 'RC built-in current loop keyer provided a means of hooking up the '33 and getting some "hard copy." Because of the conversion abilities of the 'RC, I could print anything from cw to 110-baud ASCII. I got a kick out of tuning in a cw signal at, say, 15 wpm and seeing the copy placed on paper as if by a slow (very slow!) typist. Copying RTTY signals at any of the available speeds was no problem for the 'RC. For the more affluent, there's a Centronics compatible parallel printer port for use with the more modern printers. (That sort of printer is on my "wish list.")

At the other end of the copying spectrum, I tried receiving 60-wpm Baudot and 110-baud ASCII while using Morse code output. Sure it works, but the limitations of operating this way should be obvious. You can also reverse the roles, and input Morse while transmitting RTTY. If you're entertaining that idea, remember that if your Morse transmission is slow the receiving operator is going to have to exercise a lot of patience! On the other hand, if you can rattle the keys of a keyboard fairly well, you can make the transition more palatable.

Selecting the VAR position of the filter selection switch permits using the TUNE control to vary the space tone filter frequency from about 2225 to 3125 Hz. This provides a received frequency-shift range of approximately 100 Hz to 1 kHz to allow reception of other than standard RTTY shifts. The TUNE control also adjusts the cw filter center frequency within a range of 900 to 950 Hz.

Cw reception suffers from the same maladies that affect most other automated cw receiving machines, the primary problem being sloppy fists. As is said, "garbage in, garbage out." During Morse operation, the operator can monitor received and sending speeds. The speed indicator appears in the extreme right side of the display.

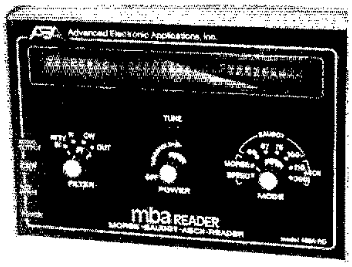
Cw transmission is best accomplished with a Baudot or ASCII keyboard. I tried using a hand key and a keyer to input Morse, but the buffered delay makes this type of operation confusing.

The I-D function performs two tasks: It is used to load an i-d message and to change the cw output speed. Unless you select a different speed, the 'RC will default to 20 wpm. I-d message length is limited to 40 characters. When that limit is reached, the BUFFER FULL LED will light and no more characters are accepted.

During transmission, the i-d message is sent by depressing the appropriate control button; it is not an automatic function. During reception, the BUFFER FULL LED illuminates when 25 characters are left in the 1024-character block. Should the buffer be forced into overflow, the display will extinguish, but data in the buffer at the time will not be lost. A.E.A. recommends using the 300-baud ASCII position of the receive output selector to avoid display blanking.

Manual and automatic control of the carriage return/line feed functions are switch selectable. In the automatic mode, a CR/LF is generated at the first space following the 60th received character or at the end of a 71-character line. Most operations will use the automatic mode.

The MBA-RC measures (HWD) 5 × 9-1/4 × 7-1/4 in. and weighs in at 4 lb 6 oz.[1] The price class is $470. It is available from Advanced Electronics Applications, Inc., 2006-196th SW, Lynwood, WA 98036. — *Paul K. Pagel, N1FB*



MBA-RO CODE READER

☐ I certainly found the AEA MBA-RO to be an interesting "appliance." It provided many hours of observing how poorly most of our cw operators handle the Morse, and it also showed the telltale evidence of my own lack of proper (cw) character formation!

One might wonder how effective a code reader can be in a crowded amateur band. Well, the circuit depends on being fed properly formed characters. It does not "like" to copy fists that are erratic or those with a "Lake Erie" or "banana boat" swing. But then, not many of us enjoy listening to that kind of sending, either! Generally, the MBA-RO will display gibberish if it is locked onto a signal that contains poor cw sending. Similarly, if the operator uses incorrect spacing between letters and words, the code reader will display exactly what is being fed into it — assuming the individual code characters are sent properly. It is distracting to have a cw message run together as though it were a horrendously long word, or to have enormous gaps between some letters or words. In practice, the human ear and mind can copy rotten cw much better than a machine can (I found myself taking my eyes off the digital readout of the code reader and relying on the coherence that my brain and ears could supply after so many years of cw copying).

This does not mean that the AEA product is deficient or poorly designed. Rather, it suggests that there aren't too many fists that such a device can copy. Indeed, it's a sad testimonial for the quality of the cw we find in our amateur bands, but it's a reality we must accept. QRM and heavy QSB tend to negate the good qualities of the instrument as well, despite the selectable built-in audio filter.

In a more positive vein, the MBA-RO does a fine job of copying keyboard-generated cw (allowing for the poor typing ability of some ops). Similarly, the RTTY copy was a delight. W1AW bulletins came through in perfect shape on both the cw and RTTY modes.

I would not recommend this device to beginners. It takes a fair amount of practice to be able to tune in a signal for reliable copy. It is not a casual operation, and one that could easily frustrate a newcomer to Amateur Radio. Conversely, I would recommend the system to anyone who hangs out on the keyboard frequencies or who works RTTY and ASCII. It would also be a suitable accessory for those who work stations that send proper keyer-generated Morse. If the operator uses his or her paddle correctly, the code reader will copy the data perfectly (I did run across a number of operators who were properly acquainted with the business end of their paddles or bug keys).

For many years, I fancied myself as a reasonable cw operator while using a quality paddle and keyer. I was gripped with despair when I hooked the MBA-RO to the sidetone output of my keyer and perceived what was being printed out on the display! My spacing wasn't all that bad, but there were a few letters that the code reader just couldn't digest. A few hours of practice, while monitoring my sending on the digital display, helped me to correct my not-so-good character formation. The code reader may be well worth the price for that application alone. If you want to be proud of your fist, you may want to give the MBA-RO a try.

The digital display has 3/8-in. characters (32 in all), and the color of the illuminated display is blue. The mode switch provides for straight Morse readout, or Morse plus the cw speed of the incoming message. In the speed-indicating mode the last two right-hand digits are used to display the sending rate of the other station. There are four Baudot positions: 60, 67, 75 and 100. The ASCII position takes care of 110 and 300 rates. Another panel switch permits wide or narrow RTTY selectivity; likewise for standard cw. The MBA-RO measures (HWD) 5-3/4 × 9 × 2-1/2 in. It is powered by a small external dc supply. The price class is $300. — *Doug DeMaw, W1FB*

CUSHCRAFT CORPORATION 220B 220-MHz "BOOMER"

☐ During the past few years, growing interest in vhf and uhf operation among amateurs in the U.S. and Canada has led manufacturers to develop high-performance antennas for the bands above 50 MHz. High performance is usually equated with forward gain, and forward gain is directly related to boom length. Cushcraft's entry in the high-performance antenna market is the "Boomer" line, aptly named in light of their relatively long booms. In addition to the 220B reviewed here, Cushcraft also markets "Boomers" for 50, 144 and 432 MHz.[2,3]

The 220B is based on the gain-optimized 4.2-λ NBS design.[4] This design calls for 15 elements on an 18-ft 9-in. boom.[5] The 220B is designated a 17-element antenna because of Cushcraft's "Trigon" reflector assembly, which employs two additional reflector elements for some additional gain. A T-match is used to feed the driven element, and the antenna is designed to be used with 50-ohm cable.

The 220B is made from high-quality aluminum stock. The machine work on the review anten-

[1] mm = in. × 25.4; kg = lb × 0.4536.

[2] G. Hull, "Cushcraft 617-6B 'Boomer' 6-Meter Yagi," *QST*, Sept. 1982, p. 41.

[3] D. Sumner, "Cushcraft 32-19 'Boomer' and 324-QK Stacking Kit," *QST*, Nov. 1980, pp. 48-49.

[4] P. Viezbicke, "Yagi Antenna Design," *NBS Technical Note 688*, U.S. Department of Commerce, Washington, DC, Dec. 1976. A discussion of NBS antennas also appears in D. Lusis, "Go for the Gain, NBS Style," *QST*, Aug. 1982, pp. 34-38.

[5] m = ft × 0.3048.

na was generally good, although some of the drilled holes were not deburred (causing one minor flesh wound).

The main portion of the boom is made from two 6-ft sections of 1-1/8 in. tubing joined by a 1-ft section of 1-1/4 in. tubing. Two pieces of 1-in. tubing complete the boom. The various boom sections are secured with worm-gear clamps and machine screws. Masts of up to 2-in OD will fit the boom-to-mast bracket. A rigid brace made from 3/4-in. tubing supports the boom. This brace may be mounted above or below the boom with no effect on antenna performance.

All elements are made from 3/16-in. aluminum rod, except for the driven element, which is made from 1/2-in. tubing. Elements are mounted to the boom by machine screws, which go through holes in the elements, then through small aluminum brackets to keep the elements from moving, and then through holes in the boom.

Cushcraft supplies first-rate hardware. All bolts, screws, washers and nuts (including the boom-to-mast U bolts and hardware) are stainless steel. Plastic end caps are provided for the boom and driven-element assemblies. Silicon grease and vinyl boots are supplied to waterproof the balun and feed-line connectors. It's obvious that Cushcraft has taken great care to provide the right parts for long and reliable antenna life.

Assembly took about one hour. The hardest part is making sure each of the 17 elements is in the correct spot along the boom. Cushcraft helps by bundling together directors number 8 through 13; they are all the same length. The instruction manual is brief and to the point. Using a minimum of text, the manual relies on clear, detailed drawings to guide you through assembly. Parts shown in the illustrations are numbered according to the parts list, leaving no questions about where each piece goes. The manual also gives instructions on stacking two or four of these antennas to provide additional gain.

Upon completion of assembly, care should be taken to align the elements. Because of the light weight of the 220B, installation can be a solo job. Remember to keep a watchful eye on the elements (especially the Trigon reflector elements) so they don't get caught in the tower. They will bend if pushed hard enough.

Chances are good that if you're reading this review you're a hard-core vhfer, or at least interested in vhf contesting. I fall into the latter category. The review 220B was originally set up at WA2OMY/3 in the Philadelphia area for the ARRL January VHF Sweepstakes contest. We mounted the antenna on a 50-ft self-supporting tower and fed it with about 70 feet of high-quality foam-type RG-8/U. Hardline is highly recommended at this frequency, but the coaxial cable was all we had available.

The antenna proved to be a real performer,

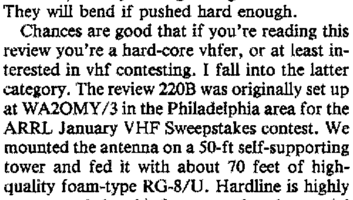


Fig. 7 — SWR curve of the Cushcraft 220B measured at the antenna with a Bird model 4410 wattmeter.

Cushcraft Corporation 220B 220-MHz "Boomer"

Manufacturer's Claimed Specifications	*ARRL Evaluation*
Boom length: 18 ft 9 in.	As specified.
Longest element: 26-1/16 in.	As specified.
Turning radius: 8 ft 8 in.	As specified.
Assembled weight: 10.5 lbs (4.77 kg).	As specified.
Windload: 2.6 sq. ft.	Not measured.
Frequency coverage: 220-223 MHz.	Confirmed.
VSWR: Less than 1.2:1 at 220.100 MHz.	Confirmed.

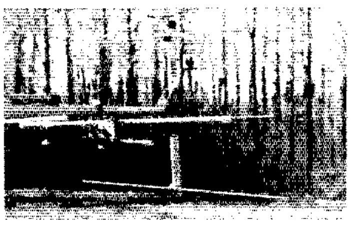


Fig. 8 — Close-up of the Cushcraft 220B Trigon reflector assembly.

despite the marginal feed line. With fair-to-average conditions, we were able to work every station we heard. Contacts north into New England and south into Maryland and Virginia were easy, whether we ran the 500-W amplifier or just the 10-W exciter. We ended the contest with the highest 220-MHz score among the Pack Rats. There was simply no comparison between the Boomer and an old-style 11-element Yagi we used previously.

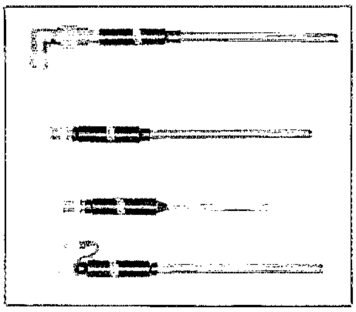
If you are looking for a reliable, well-built, high-performance antenna for the low end of 220 MHz, the 220B is worth serious consideration. It lists for $110. Manufacturer: Cushcraft Corporation, P.O. Box 4680, Manchester, NH 03108. — *Mark Wilson, AA2Z*

RF PRODUCTS 5/8-λ 220-MHz AND 450-MHz ANTENNAS

☐ In the last couple of years, 5/8-λ antennas for 2-meter "hand-helds" have proved their worth. With the increasing number of 220-MHz and 450-MHz hand-held transceivers in use each day, it was only a matter of time before someone began manufacturing similar antennas for these bands. RF Products, a manufacturer of antennas for the commercial services, has met the challenge.

Fully extended, the 220-MHz version measures slightly over 33 inches; collapsed, just under 8 inches. It (as well as the 144-MHz and 450-MHz versions) is available with BNC, PL-259 and no. 5/16-32 thread fittings. A heat-shrink-covered, copper-clad-steel spring between the whip and connector serves as a shock absorber and as part of the matching network. SWR is less than 1.5:1 across the 220-MHz band.

Using the 220-MHz 5/8-λ antenna, I am able to put a full-quieting signal into my favorite repeater from my living room with a hand-held unit, which I cannot do with a standard 1/4-λ whip. In fact, with the 1/4-λ whip, my received signal is noisy and will "drop out" when I move the whip a fraction of an inch. I am very pleased with the performance.



Diversity is the name of the game at RF Products. In addition to making gain antennas for portable equipment with a wide variety of base connectors, RF Products makes these antennas for 220, 450 MHz and 144 MHz.

Tests on the 450-MHz antenna indicated the SWR to be less than 2.5:1 across the fm band (440 to 450 MHz). Although we were unable to subject this unit to as extensive field tests as we did for the 220-MHz version, our experiments did show it to be a good performer. Most users should find a significant improvement over a 1/4-λ whip or "duck." Overall length ranges from 6-1/4 inches collapsed to 17 inches fully extended.

Price class for each antenna is $20. Additional information can be obtained from RF Products, P.O. Box 33, Rockledge, FL 32955, tel. 305-631-0775. — *Peter O'Dell, KB1N*

TWIN OAKS MORSE CODE TRAINING PROGRAMS

☐ A new series of tapes for learning Morse code, incorporating modern psychological theories, is being offered by Twin Oaks Associates. The firm, a partnership of amateurs who are also mental-health professionals, claims that its systems "represent careful interfacing of clinical psychology and Amateur Radio."

Twin Oaks markets three Morse code training programs: the System 12 Alphabet Book (designed to take the student to over 5 wpm), the System 12 (for speeds to over 13 wpm) and the System 24 (to take the student over 20 wpm). Each system has several common denominators. Emphasis is placed on learning through auditory processing — listening then mentally or verbally recognizing what was sent. The student must be willing to commit at least *30* minutes each day to code practice. The goal is to enable the student to recognize/process ever-increasing amounts of auditory material at an automatic level.

Each system asks that the student never rewind a tape to pick up something that was missed. Just listen to the first side and then the second side, repeating this process until everything on the tape is readily understood. After comfortably mastering the first tape, the student may proceed to the next tape, and use the same study method. Succeeding tapes are increasingly demanding. A written study guide details the methods and theories used to design the tapes. According to Twin Oaks, "this method has been proven under many test situations. It works if used as directed."

As the saying goes, "The proof is in the pudding." ARRL staff member Andrew Tripp tested the System 12 Alphabet Book (and is now

KA1JGG, as a result!). Here's what he had to say about it:

"Like many methods of teaching Morse code, the Alphabet Book uses an ear-training system. That is, it teaches one's brain to associate the sounds of code with the number of dots and dashes. This system, however, has a few interesting twists.

"First, one is instructed to go through all 12 sides *without* writing anything down. Only after students have mastered all the material aurally should they practice writing down what is heard. By doing this, the system's author says, you allow your brain and ear to work together without also having to be concerned with writing.

"Another interesting feature is the use of reinforcement, a technique no doubt gleaned from the author's vocation, psychology. After each character is sounded, it is identified by a narrator. This helps shorten the learning process, as well as giving the student an immediate progress report. Side 2 of each tape reviews all previous material without the narrator, simulating on-the-air conditions more closely.

"The material is presented in a unique way. Groups of characters are presented according to the number of dots and dashes they have, instead of at random. For example, Tape A introduces letters E, T, I, A, M and N, which have one or two dots/dashes. This builds, with the addition of a dot/dash per tape, to Tape E and the presentation of numerals and punctuation, which have the longest code equivalents. Again, this method helps cut down the time needed to master the material, as the repetition of sounds has a cadence to which the ear can easily become accustomed. However, this technique falls somewhat short of preparing one for on-the-air conditions, in which code is far more random in nature.

"Any problems encountered were with the audio quality of the tapes, not with the actual mechanics of the code course, which succeeded in preparing me for the Novice code test. For instance, at a couple of points in the program a microphone "thump" is heard, or the narrator's voice fades out. Also, one of the tapes had an irritating tendency to slow down, causing the code to be slurred. Gaps, created by missing code or narration, and a misplaced letter (K in with H) occurred, but were not prevalent. But these problems were minor, and could be remedied easily."

Others used the System 12 and System 24 to sharpen their cw skills after having been away from code for a while. The consensus agrees with Andrew's opinion: The only minor problems were with audio quality on the tapes. All agree that by following directions and being committed to practicing 30 minutes each day, success is *inevitable.* This is no magical path to conquering cw, but it is a viable, realistic method for success.

The Systems are produced by Twin Oaks Associates, Rte. 5, Box 37, Knoxville, IA 50138. Price class: System 12 Alphabet Book, $15; Systems 12 and 24, $30 each — *Carol L. Smith, AJ2I*

N9CR CONTEST RADIO OPERATING SYSTEM

☐ I am a contester, but I hate to shuffle paperwork. Operating a contest with an arsenal of aluminum of the "death-ray" variety can be a very enjoyable experience, but compiling the paperwork associated with a 24- or 48-hour operating stint makes me feel more like a *victim* than a *victor.* I shudder when I think of how many times I've put a serious effort into a contest operation, then haven't bothered to submit an entry for fear of the drudgery of duplicate checking.

I've often thought that it would be nice to operate a contest, push a few keys on a computer keyboard and be left with legible, *error-free* log and dupe sheet printouts! A foolish dream? A few years ago, perhaps, but with modern microcomputer technology, a very reasonable one.

With this idea in mind, N9CR has developed a real-time program for the APPLE® II microcomputer that puts *all* contest paperwork where it belongs — in the closet with your old vacuum tube handbooks! The system not only compiles a running log and checksheet while you operate, it *sends cw contest exchanges as well!*

Computer Requirements

To use the Contest Radio Operating System, an APPLE II Plus or APPLE II microcomputer with 48K of Random Access Memory (RAM) and at least one 16-sector disk drive are needed. With the APPLE II (Integer BASIC), Ramcard APPLESOFT® is an additional requirement.

Log and checksheet printout is accomplished with any of the available Apple-compatible printers. For log printout, a 67-column printer is required, but a unit capable of printing a 92-column line is necessary for dupe sheets.

N9CR lists a clock card as an option but I would have listed it as a definite requirement if a serious effort is planned. Without such a clock card, it is necessary to input a four-digit time entry with each QSO. Depending on the contest and exchange, this could slow the operator down considerably. After approximately 20 entries without the clock card, I chose to borrow one from a friend to complete the review.

System Operation

Operating-system software is supplied on a single 5-1/4 inch diskette. To utilize the program for logging and duping duties, it is necessary to use the supplied software diskette to format another, on which the log and duplicate check sheet(s) are maintained. A single log diskette formatted with the software diskette can store up to 2350 QSOs, or the results of 17 contests, whichever occurs first.

When the user first runs the program, the software requests contest-rules information. System provisions enable a consecutive serial number to be sent and incremented (for ARRL SS, for instance). The software can also be used to keep a dupe sheet for each band/mode; for Field Day or IARU Radiosport competitions, this is a must.

While the system is in operation, the dupe-search function performs almost as quickly as I can type in the call sign. I've seen many programs that have taken as much as a minute or two before completing a single check! In addition, provisions have been made to allow editing of the call sign and received report during operation. A call sign or report can be edited after the contest, also.

The software contains approximately 20 abbreviated commands, all designed to speed up the system and virtually eliminate operator-generated "cockpit" errors. An example of one such command is used to change cw transmitting speed. Instead of being forced to "break" the program and restart, the keystrokes CTRL-S, a numeric input between 4 and 100 (WPM), and the "RETURN"(ENTER) will instantly change the cw routine speed.

Documentation

With each software diskette, the manufacturer supplies 14 pages of information. A detailed, 13-page instruction manual presents the prospective user with information on software usage as well as interfacing instructions. In addition, valuable information on user modification of the software is given. This information enables the user to modify several functions, such as logging in local time (instead of UTC), or to locate the real-time clock in a slot other than 4.

While the program is in operation, the software writes information to the disk every four QSOs. This could cause a few QSOs to be "lost" if shack power should ever fail. If this possible loss of information annoys you, the software can be modified to force a "write to disk" after every QSO. Modification provisions for control of various printers is also addressed.

For the operator (like myself) who has difficulty remembering his or her *own* call at 3 A.M., a single-page, quick-reference guide is provided. Posted near the machine, this sheet will remind the user to achieve the desired system function.

Comments

I found the software a joy to use after becoming familiar with the numerous commands. My first attempt was to use the operation on an old contest log. Just as with a new electronic keyer, I would suggest several off-the-air practice sessions before using the system on the air, as one slip of the keys can confuse the unskilled user.

I would prefer an optional two-digit time input format to the required one. This would allow entry of the time every five minutes or so, and allow a four-digit input at the change of the hour. This would eliminate the added expense of a clock card.

Overall, I would recommend this operating system for anyone who really loves to contest, but hates paperwork. Think of it — no more nasty letters from the ARRL Contest Branch about dupes or sloppy logs!

The N9CR Contest Radio Operating system is available from CR Software, 2512 James Dr., Dyer, IN 46311. Price class: $35. — *Michael B. Kaczynski, W1OD*

Strays

Visitors to the Eastern India Science Camp in Calcutta pause to watch and listen as VU2MKI makes a contact. Members of the Amateur Radio Department of the Birla Industrial and Technological Museum (curator VU2BMT) manned a booth at the Camp for a week last February/March.

VOLTAGE-SENSITIVE BRIDGES

☐ Voltage measurements can be important. Often, the most important consideration is whether a particular voltage level is exceeded. The voltage-sensitive bridge is useful in this type of detection. I've reviewed a few types of voltage-sensitive bridges and will provide hints for their use.

Resistors connected in a Wheatstone bridge arrangement (Fig. 1) produce no difference in voltage between points A and B when the ratio of voltage divider R1 and R2 is the same as at divider R3 and R4. If the resistance of R1 is reduced, A will become positive with respect to B, while increasing R1 will in turn make A negative. If the supply voltage is ac, the A-B voltage will reverse polarity with every input voltage reversal. This polarity reversal can be used as a control signal.

Bridge balance is independent of the supply voltage as long as all of the resistors will increase current *proportionally* to the applied voltage. The bridge balance can be made sensitive to applied voltage by using one or more resistors whose current is *not* proportional to applied voltage.

One of the simplest voltage-sensitive bridges can be made by substituting incandescent pilot lamps for one or two of the resistors (Fig. 2). Tungsten lamps will show an approximate 10:1 increase in resistance when hot, as compared to cold. This resistance depends on the rms voltage applied; therefore, this circuit will work on both dc and ac above approximately 60 Hz. (The metal filament has thermal inertia and tends to stay at a nearly constant temperature across the ac cycle.)

In the circuits of Fig. 2, as the applied voltage is increased from zero, the A-B voltage will exhibit a polarity reversal as the lamp filament temperature (resistance) passes the balance point. This principle was used more than 50 years ago to provide "volume-expander" action in entertainment receivers. A more modern embodiment uses Zener diodes (Fig. 3). The circuit shown in Fig. 3B is commonly used in solid-state voltage regulators, and when using glow regulator tubes or neon lamps, was used previously as the sensing element in vacuum-tube, high-voltage regulators. Corona tubes have also been used as the breakdown element in cathode-ray-tube supply controls. Note that the voltage sensitivity of the full bridge (Fig. 3A) is twice that of the simplified bridge at Fig. 3B. One disadvantage, however, is that the balance voltage has to be approximately twice that of the regulator element.

Both Zener bridges are designed for ac use, and are sensitive to the instantaneous value of the waveform, rather than the rms value. Also, glow regulators must have an initially high

*Assistant Technical Editor

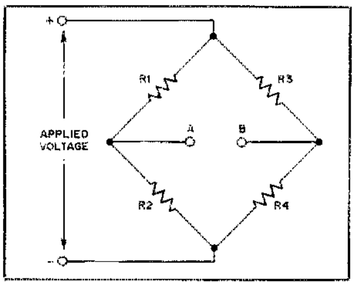


Fig. 1 — The Wheatstone bridge has no potential between A and B when R1R4 = R2R3. Different voltages can unbalance the bridge when one or more "nonlinear" resistances are used.

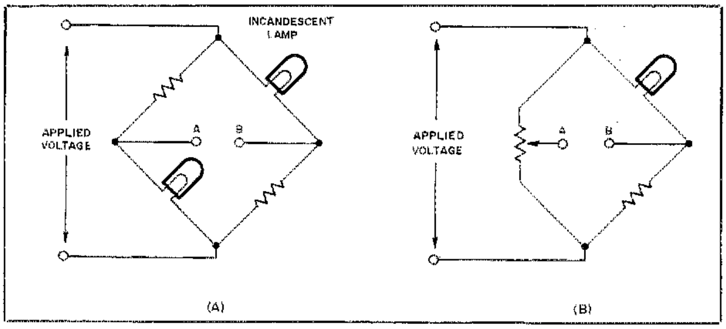


Fig. 2 — Low-impedance, voltage-sensitive bridges using incandescent pilot lamps. Heat inertia allows these to respond to rms.

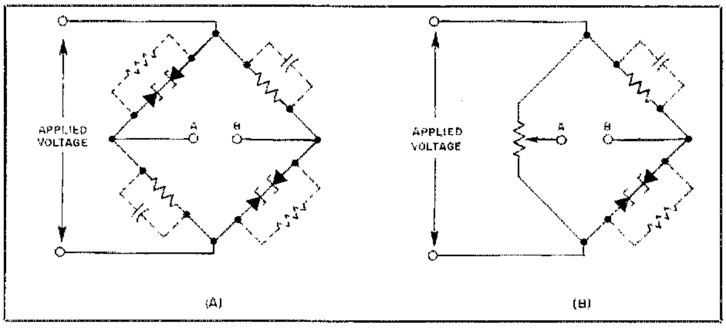


Fig. 3 — Zener bridges, which respond to instantaneous voltages. Use of balance resistors (dashed lines) eliminates response to less than twice the Zener voltage in the full bridge (A) and less than the Zener voltage at B. Small capacitors (dashed lines) help minimize spurious responses.

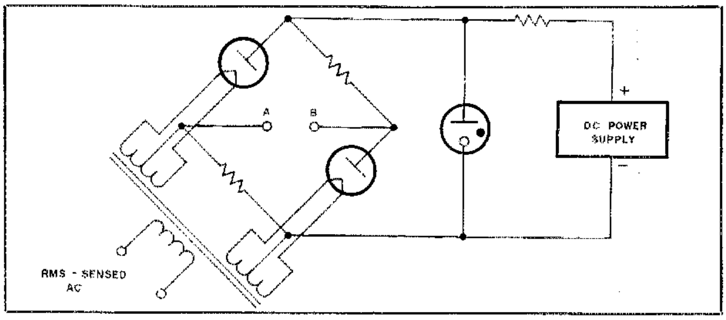


Fig. 4 — Special temperature-limited diodes act as variable resistors, which respond to heater rms voltage.

voltage to start the glow and, practically speaking, are much better for dc than ac.

Sorenson (a Division of Raytheon) solved many of the problems of rms voltage sensing by using temperature-limited diodes with high heater inertia (Fig. 4). Such diodes emit very few electrons from a cool filament and appear as a high resistance, which decreases as the filament receives more power. As the thermal intertia is controlled by design, the response to changes in rms level can be made fairly independent of waveform.

A practical transient sensor (Fig. 5) can be made from a dc Zener bridge and a bridge rectifier. This sensor produces voltage between points A and B whenever the sensed voltage exceeds 400 V, either polarity. While most major appliances are supposed to withstand occasional 1200-V surges, surges of 400 V or more often occur in residences and may damage underdesigned equipment. Capacitor C (usually under 1000 pF) is adjusted to minimize either overshoot or undershoot at A when less than 200-V dc is suddenly applied to the "sensed voltage" terminals. The fuse is used to prevent a line fault if surges more than 1000 V short a 1N4007 rectifier. — *David T. Geiser, WA2ANU, ARRL TA, New Hartford, New York*

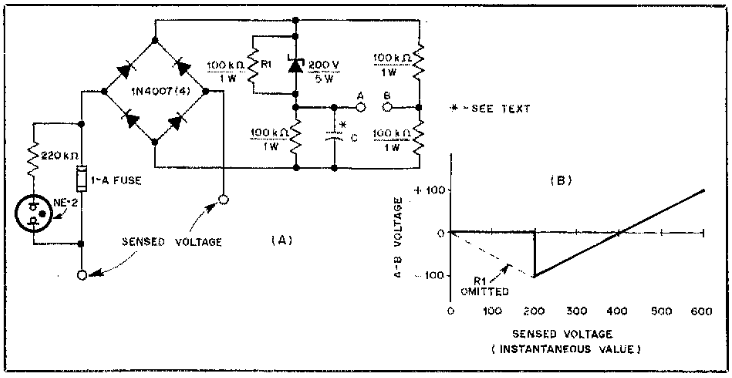


Fig. 5 — A practical transient line-voltage sensor providing indication of surges exceeding 400 V. The neon lamp indicates that a surge exceeding 1000 V has shorted one or more rectifiers. At B, a graph represents the response of the line-voltage sensor circuitry.

HEADPHONE FREQUENCY RESPONSE

☐ It appears the most popular brands of headphones provide rather broad frequency response. What is the use of a headphone with a frequency response from 300 Hz to 20 kHz when receiver filtering leaves us a 300-Hz to 3-kHz range? For cw or ssb operation, should one look for a narrow or a broad response in headphones? It is difficult to answer these questions, for there are no references in the literature on headphones or speakers and their responses. — *David J. Schoenwald, WB2KJL, Huntington, New York*

CMOS PLL NOTES

☐ A variety of CMOS phase-locked loop (PLL) integrated circuits is currently available, including the Motorola MC145145 and MC145146, and other members of this series.[1] These ICs easily interface to a microcomputer bus and can provide a low-cost, easy-to-build, high-performance frequency synthesizer.

Unfortunately, certain aspects of using these PLLs with varactor-tuned VCOs are not clear from the data sheets; in some cases, the data-sheet formulas will lead to a design that will not achieve lock when the tri-state phase detector output (PD) is used rather than the ϕ_R and ϕ_V outputs. Often, it is desirable to use the PD output because a passive loop filter can be constructed, which will give low phase noise and fast settling time.

The loop-filter equations given in the MC145146 data sheet are based on the presumption that the short-term average (dc) output level from PD_{out} is proportional to the phase error. This is true if the VCO input can be represented as a resistance to ground or to V_{DD}, and the loop will obey the data sheet equations. However, if the VCO dc input resistance is extremely high, as in a FET gate or a varactor, the PD output acts more like an integrator than a phase detector. That is, if the input phase lags behind the reference phase by a small amount, the PD output will charge up any capacitors in the loop filter until it reaches very nearly V_{DD}. This is not in agreement with the stated $K_\phi = V_{DD}/4\pi$ (in volts/radian) for PD, but is in agreement with the tri-state timing model given in the data sheet.

Furthermore, when the input and reference phases are almost perfectly in step, PD_{out} changes to a "doublet mode" in which it puts out a positive and negative pulse before returning to the high-impedance state. In this mode, the sum of the durations of the two pulses is approximately 200 ns, and their relative widths depend linearly on the phase error; in the doublet mode $K_\phi = V_{DD}/(f_R \cdot 2 \times 10^{-7})$, where f_R = the reference frequency. Note that the doublet mode is a very desirable way to operate the PLL, as the doublet at equilibrium has no f_R content but only harmonics of it — and those are at low levels.

Returning to the "normal" mode of the PD output, the integrator action adds an extra 90° phase lag to the PLL, destabilizing it. If the phase error can be made to go through zero slowly enough for the PD to "catch" the loop in the doublet mode (and stay in the doublet mode), stable, low-phase-noise operation will result. For this to happen, the loop filter must be relatively fast and should resemble a Type 2 Second-Order sytem.[2] A circuit I have used (Fig. 6), acquires frequency lock to within < 200 Hz in ≤ 50 ms, with 5-kHz steps at 2 meters. Fm noise in the audio range is less than 25-Hz peak deviation. The circuit covers a bit more than 3 MHz of the band. I wish to thank Patscenter International, Inc., Princeton, New Jersey, for allowing me to use their facilities in this research. — *Raymond Simpson, WA2PYX, Princeton, New Jersey*

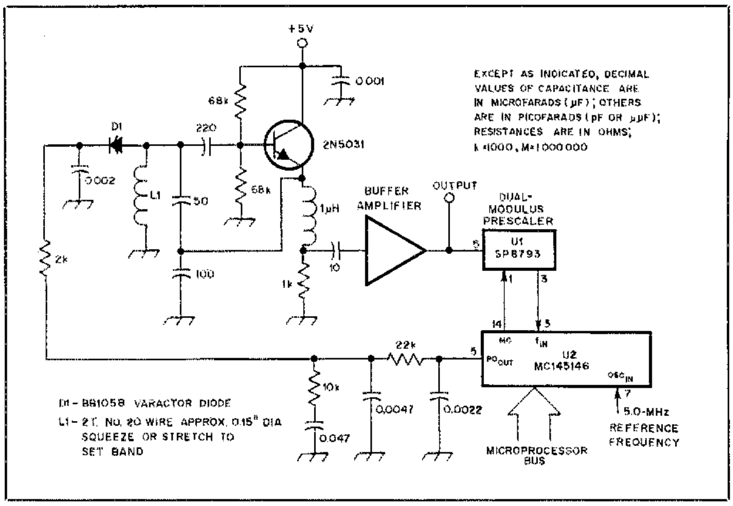


Fig. 6 — CMOS PLL frequency synthesizer described in the text.

[1] Motorola Semiconductor Products, Inc., Box 20912, Phoenix, AZ 85036.

[2] G. Nash, *Phase-Locked Loop Design Fundamentals (AN535)*, Motorola Semiconductor Products, Inc., Box 20912, Phoenix, AZ 85036.

Space Shuttle *Columbia* Calling All Radio Amateurs

Will NASA Astronaut W5LFL, the first ham in space, put you in his STS-9 log?

By Bernie Glassmeyer,* W9KDR, Peter R. O'Dell,** KB1N and Roy Neal,*** K6DUE

Electrifying excitement is building rapidly as word is spread around the world of the official NASA approval of Owen Garriott's bid to become the first Amateur Radio operator in space. On September 30, the Space Shuttle *Columbia* will, if all goes according to plan, lift off the launch pad at Cape Canaveral, Florida, carrying a specially built 2-meter fm transceiver. Dr. Garriott will establish two-way communications with as many amateurs as he can while he circles the globe.

All Amateur Radio operators will have an equal chance to work W5LFL from *Columbia*, but his operating time will be limited. It will be impossible for everyone who would like a QSL card from space to complete a two-way contact, since Owen will be allowed a total of only six hours of operating (one hour per day). Before delving into the mechanics of this historic event, let's look at it from a public information standpoint.

You and the Press

You don't have to make a contact to participate in this history-making event. Simply monitoring the operation and passing up-to-the-minute information to your local media will enable you to become involved personally. But if you do make a QSO, so much the better. As Rich Moseson, N2BFG, of the CBS-TV program "In the News," explains, "A local person actually talking to an astronaut could easily be front-page stuff in a newspaper or a near-to-lead story on a radio or TV newscast." Rich also points out the other side of the coin: "No reporter or editor can cover a story he or she doesn't know about."

It's not too early to set up an appointment with your local news reporters or editors. Make your arrangements now. This is the first major national story about Amateur Radio that doesn't involve a catastrophic event. Let's make the most of a perfect opportunity to promote the Amateur Radio Service.

Dr. Owen Garriott, W5LFL (left), and Bernie Glassmeyer, W9KDR, discuss some antenna-mounting possibilities inside the Space Shuttle Trainer at the Johnson Space Center in Houston. (*NASA photo*)

Begin by making a survey of your local media; in addition to the obvious daily newspapers, radio stations and TV stations, remember the not-too-obvious things such as company newsletters, cable-TV stations and weekly "shoppers." Send the ARRL Public Information Coordinator (ARRL Hq.) a list of the reporters/editors you will be in contact with, including the type of an organization they represent. We'll send you a Press Information Kit explaining what the Garriott mission and Amateur Radio are all about. You can pass these kits on to these reporters and editors when you contact them. (This should be done at least several weeks before the launch.)

The next thing is to think about your station, the image it presents to the "uninitiated" public and what you can do to make the most favorable impression on visitors. First of all, it just makes good sense to tape-record everything on a mission like this. You should experiment now to determine the best method of hard-wiring the recorder into your station. (There are too many variables to give you a "universal" interface, but start with some 1-kΩ isolation resistors.)

The next thing to consider is what your station is going to look like to the outsider — particularly one with a camera. Ask unlicensed members of your family how it looks to them. If you have an instant camera, take a few photos and go over the prints carefully. (You don't have to dispose of the box of surplus RTTY gears you bought at the hamfest; honest. Just hide them.) Dispose of clutter until your station conveys an image of "professionalism."

Suppose you are trying to work the Shuttle and a reporter asks you a question; what would you do? The best bet is to team up with at least two other hams. Two can keep track of the station equipment and operations, watch the clock and, generally, double-check each other. The third should act as spokesman for the team. The spokesman should be completely familiar with the station and its operation, because he or she will need to answer questions about what is going on. (You'll also need a second set of handouts for the reporters who cover the story. ARRL Hq. will provide these to the people who make the advance contacts we suggested.)

This is a once-in-a-lifetime opportunity for the local ham to be a part of a national story — and it's a story that doesn't involve pain and suffering. Make your preparations early. *Do it now!*

Preparing for W5LFL's Flight

How can you have the best chance to work W5LFL aboard the Orbiter? Completing a two-way contact is going to take some advance preparation — and some luck!

Starting on day three of the nine-day mission, Dr. Garriott will provide about one or two hours advance notice of his intention to operate the Amateur Radio transceiver. He will announce this to Mission Control on the Orbiter's normal air-to-ground frequency. The announcements will then be disseminated through AMSAT nets, W1AW bulletins and a special "900" telephone number that will be announced before launch.

Operation will be limited to a maximum of about one hour per day, when no other flight activities (or sleep periods) are

*OSCAR Program Manager, ARRL
**Public Information Coordinator, ARRL
***c/o NBC News, 3000 West Alameda Ave., Burbank, CA 91523

Table 1
STS-9 Operating Frequencies

Space to Earth: 145.510 to 145.770 MHz
Earth to Space: 144.910 to 145.470 MHz
(20-kHz steps)
All operations will be F3.

This range of frequencies will allow operation from most parts of the world. Although some frequencies fall on some repeater inputs and outputs, operation through repeaters is not planned.

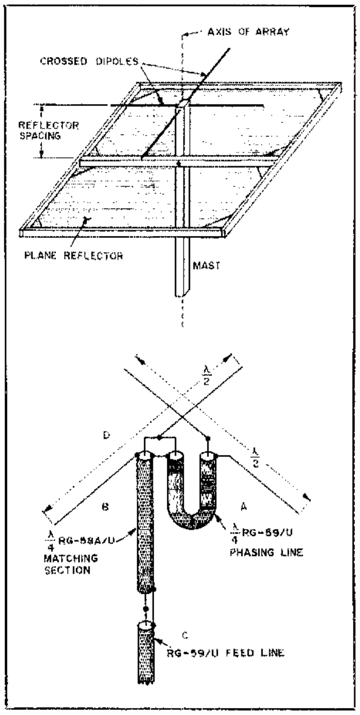


Fig. 1 — Dimensions and connections for the turnstile antenna. The phasing line is 13.3 inches of RG-59/U coax. A similar length of RG-58/U cable is used as a matching section between the turnstile and the feed line.

Table 2
STS-9 Orbital Parameters

Period: 90 minutes
Altitude: 155 nautical miles (250 km)
Inclination (angle measured north from equator): 57°
Increment (equator degrees that the earth turns during one complete spacecraft orbit): 22°

This data, with an equator-crossing time and coordinate (in degrees, west longitude) after launch, will "plug in" to most computer or calculator programs.

scheduled. Orbit numbers and ground tracks for potential Amateur Radio operations are being prepared and will be identified prior to flight.

As the Orbiter approaches the portion of the ground track where Amateur Radio operations are planned, Owen will call and listen on alternate minutes. He will transmit continuously for one minute, beginning on the even minutes, and will receive continuously for one minute, beginning on the odd minutes. Be sure to synchronize your station clock to WWV, plus or minus two seconds.

The transceiver will have the capacity to transmit and receive on channels 20-kHz apart within the planned operating range shown in Table 1. Discrete downlink (space to earth) frequencies and uplink (earth to space) frequencies will be announced before the flight.

During a typical even-minute transmission period, Dr. Garriott will identify a geographical area or call district that he will listen for. He will also announce the frequency range and, as time permits, describe crew activity or views of the earth.

During the odd-minute receive period, Dr. Garriott will scan the announced uplink frequencies for call signs from the designated area only. To establish contact, you will send your full *call sign only*, repeating it several times during the scanning period.

During the next transmission period, on the even minute, Dr. Garriott will acknowledge all call signs he has heard during the scan period. No other report will be needed; call-sign identification constitutes a two-way contact. This procedure will give more operators a chance to make a contact. If time permits, some stations may be called on for short transmissions to fill the time period.

How to Track the Orbiter

Maximum communication time as the Shuttle passes directly over your QTH is eight minutes. Accurate timing is essential, so calibrate your clock to the most accurate time standard available to within two seconds. Orbital information will be updated daily on the "900" number, on AMSAT nets and on W1AW bulletins.

Tracking with the OSCARLOCATOR is possible, but you will need to make a simple modification: You can modify your existing OSCAR 7 or OSCAR 8 overlay by tracing the full-size ground track and range circle shown on page 79. Use a permanent-ink fine-point pen, like the Pilot SC-UF, to trace your curve, and you're all set.

Those who have computer tracking programs will also be able to track the Orbiter. The elements needed for most programs are shown in Table 2. If you have a computer or programmable calculator and need a program, write to AMSAT Software Exchange, Box 27, Washington, DC 20044.

Equipment You'll Need

Making a 2-meter contact will not require an elaborate station setup. Actually, it will take less than is required for normal OSCAR-type contacts. AMSAT and ARRL recommend around 40 W of output power to a turnstile antenna (see Fig. 2) to have the best chance of being heard. Construction details are available in recent editions of *The Radio Amateur's Handbook* and *The ARRL Antenna Book*. Most 2-meter nbfm equipment will be able to make the earth-to-space connection. Dr. Garriott will be running only 5 W to a loop antenna, and it will be interesting to see how little power it takes to make a contact.

Operating procedure and timing will be key factors in getting your call sign in the W5LFL log.

Since the planned operation calls for separate listen and transmit periods, you won't need special split-frequency capability. Having the exact frequencies of operation, you can simply monitor the space-to-earth frequencies during the even minutes and switch to the earth-to-space frequency of your choice for your transmission.

Do not try to track the Shuttle with azimuth and elevation control unless you have precision, computerized equipment; with only a few minutes of access time, you'll find it almost impossible, especially if your antennas have a narrow beamwidth. If you use a gain antenna, we recommend the old, reliable "Arm-strong Rotator" method — holding the antenna and aiming it by hand for best reception.

QSL Cards for All

Specially designed QSL cards will be available to everyone who sends a reception report of the Amateur Radio operation from Space Shuttle *Columbia*. Pass the word to your nonamateur friends so they can monitor on scanners, or let them listen in on your receiver so they can qualify for the special QSL. Send all reception and confirmed contact reports to ARRL, STS-9, 225 Main St., Newington, CT 06111 USA.

Operation Protocol

The first attempt to communicate with the Space Shuttle will present a challenge to most Amateur Radio operators. Beyond the achievement of being in the right place at the right time is another challenge: public relations. The public at large may have a little trouble relating to a cold piece of electronics hardware orbiting the earth, but they can relate to a human aboard the Space Shuttle. What we do and how we conduct our operations can bring more prestige to Amateur Radio and promote international goodwill.

If we can keep this historic event simple and remember our rules of diplomatic etiquette, we will have a better chance of gaining future Space Shuttle Amateur Radio opportunities. Join in the fun of this operating "Event of the Decade." QST

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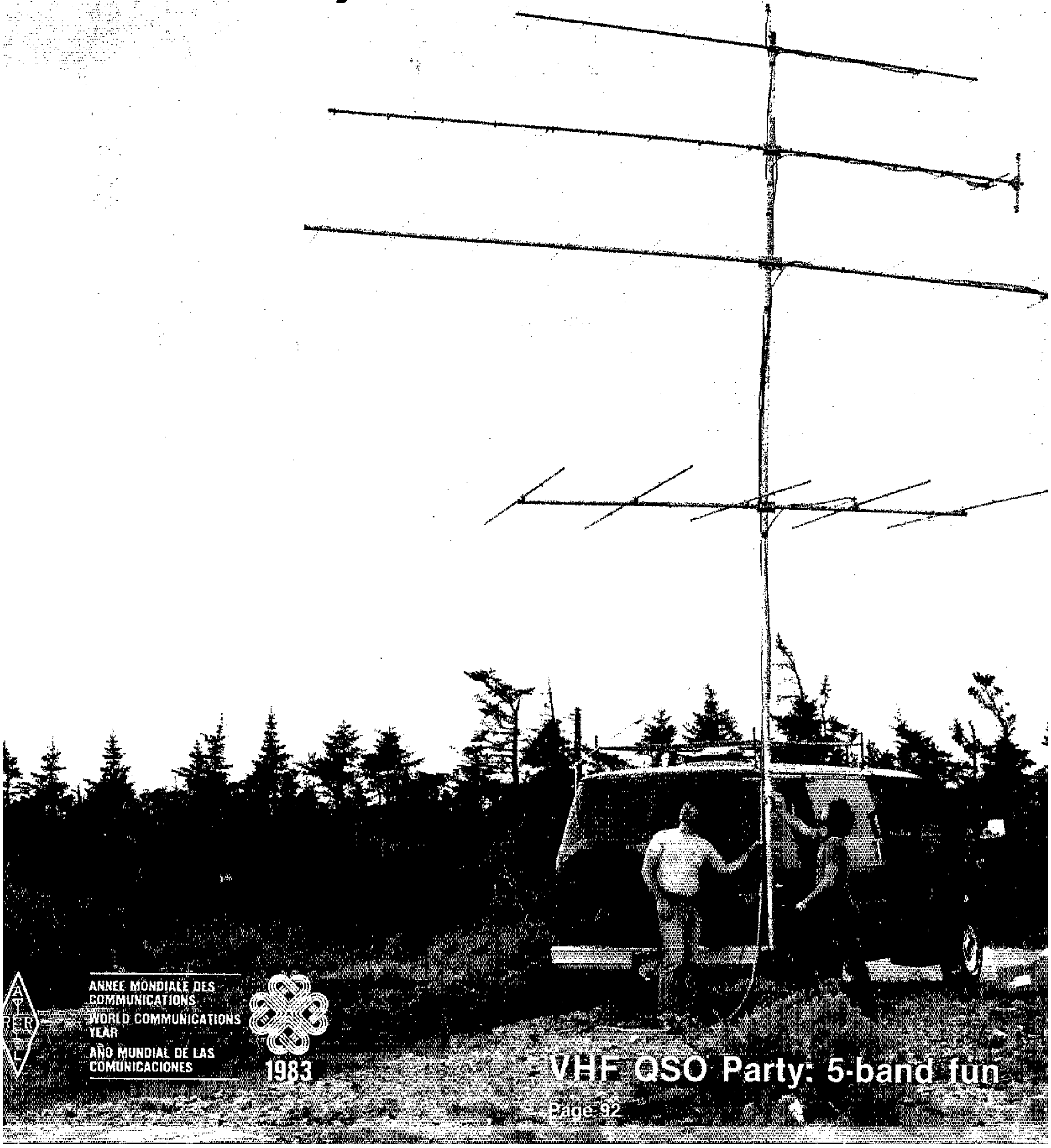

ANNEE MONDIALE DES COMMUNICATIONS
WORLD COMMUNICATIONS YEAR
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VHF QSO Party: 5-band fun

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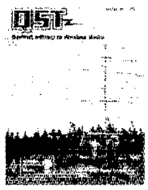


OUR COVER

It's Saturday, June 11, and the WA2VUN/1 VHF QSO Party operation is underway atop Cadillac Mountain in Maine. The 3-person, 4-band effort netted a respectable score, some decent DX — and an unforgettable weekend! To see how you did, see page 92.

TECHNICAL

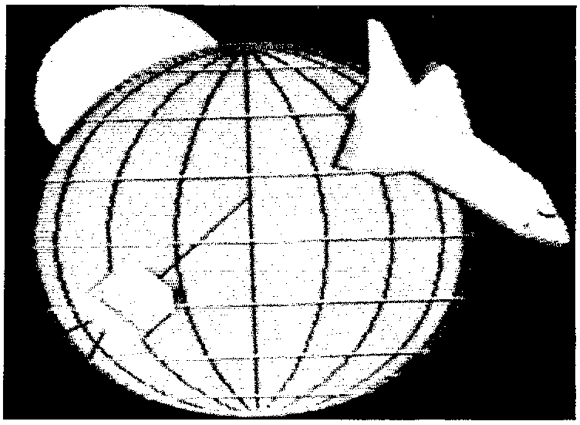
BEGINNER'S BENCH

NEWS AND FEATURES

OPERATING

DEPARTMENTS

Graphics on RTTY



Sending schematics over the air may not be as difficult as you think. Read about a computer-graphics technique that may be your opportunity to help advance the state of the art!

By Ted Thompson,* VE3FTT

This is an account of how Bob Howard, VE3GQW, and I combined Amateur Radio with the latest computer-graphics technology. We conducted picture-transmission experiments on hf using a coding method that has not been used previously on Amateur Radio. Techniques have existed for years to send pictures from one Amateur Radio station to another. Amateurs have used, and at times have been the first to use, such methods as uhf fast-scan television, vhf and hf slow-scan television, and facsimile, and have even sent greeting cards using asterisks or other characters.

A Better Way

Bob and I have been active in Amateur Radio and computering for several years. We had come to the point of wanting to combine the two hobbies and try experimenting beyond normal RTTY communications. There is a definite limit to the quality of graphic information that can be sent using RTTY characters (notwithstanding the fine pictures that appear every so often on the air). While trying to describe a circuit diagram on ssb one evening, we realized that there must be a better way. *That* was a real exercise in frustration!

We searched the available literature to see what methods had been used to send highly detailed pictures, such as circuit drawings, on hf. The distance between our locations restricted operations, for practical purposes, to 80 and 40 meters with the attendant bandwidth and mode restrictions. If we had been closer, we could have tried 220-MHz packet radio and used a higher transmission data rate, but we felt that whatever we finally decided on should be applicable to all the hf bands for the widest possible use.

Our literature search revealed only two methods in common use; slow-scan television and facsimile. Either of these would have meant the purchase of equipment for each end. In addition, neither method was directly related to our interests in computer graphics.

The search moved into the computer graphics field, and we were quickly overwhelmed with reports of methods for storing detailed picture data in computer memory. We soon found that several methods had been developed for transmitting pictures over coaxial cables, telephone lines and even broadcast television signals. This seemed more like it! There was no reference to any use on hf; in fact, several reports implied that the slow data rates we are required to use would be impractical. We were not aware of any data available about the error rates that could be expected on hf RTTY. This was important because we had decided to try to make use of some modern computer-communication method for sending graphic data.

Which Method?

Of the systems we researched at the time,[1] three stood out as possible methods to meet our requirements: Prestel, Antiope and Telidon. These three methods differ in the picture-storage space required, the amount of data to be sent to define a picture, the method of drawing the picture and most important, the picture quality. We had the opportunity to see a Telidon demonstration. This technique provided higher quality and greater versatility than the others, so we decided to look into it further.

At this point we had a fairly good idea of just what we wanted to do on the air: Select one of the available standards and try sending picture data, find out if error rates are low enough for useful work and see if sophisticated error-correcting techniques would help. There was one remaining problem, however. In Canada, as in most other countries, the rules specify that all Amateur Radio communications must be in "plain language." Unfortunately none of the techniques under consideration are "plain language," so we realized there would be difficulties, whichever one we chose. In Canada, the Department of Communications (DOC) controls Amateur Radio. Coincidentally, another branch of this Department had developed the Telidon picture-transmission techniques. It was this fact that finally led to our selection of Telidon as the experimental technique.

The DOC allows the use of Baudot or ASCII for amateur RTTY, but Telidon falls into a gray area. It is an internationally recognized extension of the ASCII character set but can hardly be called "plain language." Bob and I applied for special permission to use Telidon codes on the air. After further discussion with the Department, we were given permission, subject to several restrictions and the requirement for a full report to the Department on the results of our tests.

What is NAPLPS?

A few months later the choice would have been easier: Telidon is now recognized by the Canadian Standards Association —

*61 Sample Rd., Ottawa, ON K1V 9T9

[1]Notes appear on page 14.

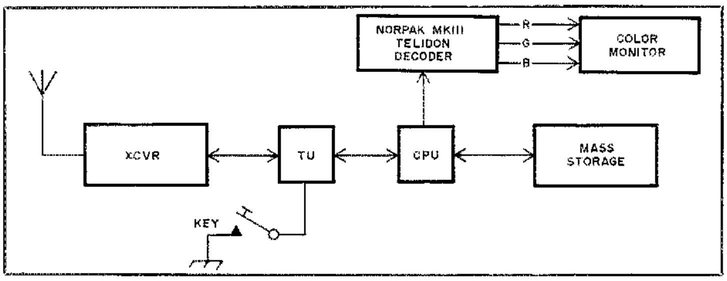


Fig. 1 — Block diagram of the station set up used for the Telidon tests.

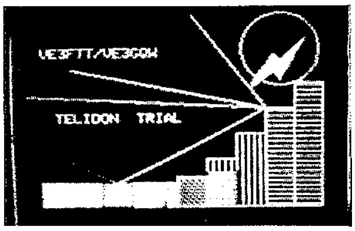


Fig. 2 — Photograph of one trial picture, as received by VE3FTT. The picture was received error free.

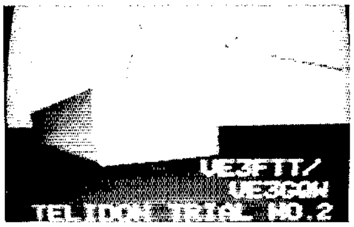


Fig. 3 — An error-free picture received during the Telidon testing.

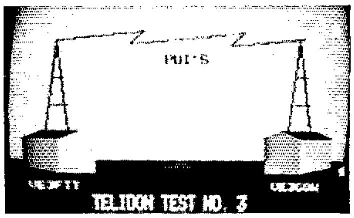


Fig. 4 — An error-free Telidon picture, illustrating the picture detail possible with this system.

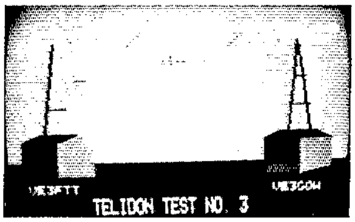


Fig. 5 — A test picture received with errors. Compare this with Fig. 4.

CSA(T500), the American National Standards Institute — ANSI(X3L2), and the American Telephone & Telegraph Co. — AT&T(PLP). This graphics-transmission technique has become known as the North American Presentation Level Protocol Syntax (or NAPLPS) and is being supported by such industry giants as AT&T, DEC and Intel.[2,3]

NAPLPS is a method of using an international standard code set (the International Organization for Standardization — ISO 2022) to encode text and pictures using a scheme that is independent of the particular terminal equipment used. Amateurs are already familiar with the idea of terminal independence in that the text code (ASCII) has been used for years on all sorts of terminals and hard-copy devices. The same idea is now extended to sending graphics data.

Pictures are defined under NAPLPS as points, lines, arcs, rectangles and polygons in a manner that does not depend on such terminal characteristics as resolution or color capabilities. The picture is produced by the receive terminal based on the drawing commands received.

This standard has found support from several major hardware manufacturers and could become as important to communications, and in particular computer communications, as ASCII is today. Full details of the standard can be found in a series of articles starting in the Feb. 1983 issue of *BYTE* magazine.[4]

Technical Details

Equipment

The equipment configuration at each station is shown in Fig. 1. Each station used a transmitter capable of producing 130 to 200 W on RTTY and a receiver with a homemade RTTY demodulator or terminal unit (TU). VE3FTT used a demodulator based on the Exar 2211 chip using Exar application notes for the design.[5,6,7] VE3GQW used a DT-600 demodulator described in the Feb. 1976 issue of *Ham Radio*.[8] Both units were optimized for the somewhat unusual baud rates used in the tests.

Each station had a computer connected to the transceiver for picture file generation and storage. One computer was a Heath H89, and the other was an S-100-based machine. Each computer was also connected to a Norpak Mark III Telidon decoder unit. This unit drove a modified color television used as an RGB (red, green, blue) type monitor.

RTTY Standards

All RTTY operation used standards that have developed on the amateur bands. We used 170-Hz shift with the mark frequency high. Data rates of 45, 75, 110 and 150 bauds were tried.

Propagation

A restriction forced on us by our station locations was the limitation to 80-meter operation. The path length was about 150 miles, but 40-meter QRM was always so high that 80 meters was the only band usable most of the time.[9] We operated in typical autumn/early winter conditions with low atmospheric-noise levels but fairly heavy band usage in early evening. There was generally strong fading with the signals going from S9+ down to noise level within seconds. Such fades would last for several seconds and do a lot of damage to a transmitted data stream.

We found very good cooperation from other operators. Only once did we fall prey to malicious interference in several evenings of operation. All coordinating communications were carried out in ASCII at the baud rate in use at the time. The Telidon files were each identified as such before transmission. This was a DOC requirement because the files were not "plain language."

What We Sent

Figs. 2 through 5 are examples of the pictures we used in our experiments. All the photographs are as drawn by the Norpak decoder on a modified Heath GR-169 14-inch color television using received data. All are error-free except Fig. 5, which is included to illustrate the effect of errors on a display file. Table 1 is a listing of the data file that will produce the display in Fig. 4. The underlined characters represent the corresponding control characters, used to switch the decoder from text to graphics mode and back. The rest of the characters are as sent, and appear as "non-plain language" when received by a normal ASCII terminal.

The time required to transmit a Telidon file will vary depending on the actual picture being transmitted. A more detailed picture requires an increased number of characters. The data for Fig. 4 will take 68.5 seconds to be transmitted at 45 bauds, and 28 seconds at 110 bauds.

Error-Rate Results

The one word that might best describe the results of our tests is "surprising." The experience we had with RTTY seemed to

Table 1
Sample Telidon Data File

```
N=M=@<P3@@@Xv<h3@GFZxz<`3@nB@[V<X7@nB@CFxhHGE@GWz<`7Aj@@XPxp^wZ<@5@H
nxhHwZ@Px<X3P~B@[V<`5@P~@xhHGDA<@7YRP@Pxxp^xhHwZ<X5x`h@HnGOj<@*Ab@Bk
F)FO@+YJNAI@)FO@+Ae[@H`+B`g@@x+YMk@H`+ZPA@@x<x/Z1I@@¦xxL<h7ZkF@@gxxg
$HB@<x=acOTELIDON TEST NO. 3N=aC$@TcOVE3FTTN$P¦sOVE3GQWN=J+BkF@xg)vv)
HQ@)vv)@pl]$Jg\OPDI'SN=H
```

indicate much higher error rates than we found. It may be that there are a lot of poor typists out there but we think rather that it is a matter of controlled conditions producing better results. There is not a lot of ASCII operation on the air at the lower baud rates that we were using, except for ARRL bulletins. Error counts on the bulletins seem to indicate the same order of magnitude as our 110-baud error rates.

Before we started we were not encouraged by those who are using Telidon every day. A normal Telidon decoder runs at 1200 bauds, so our low baud rates were deemed unworkably slow. However, that point of view is influenced by Telidon design objectives. Our application does not require the speed (although it would be nice) but rather the flexibility and image quality available.

We conducted extensive tests using Telidon files at 45, 75 and 110 bauds by saving each received file and doing automated and manual error counts. As I mentioned earlier, fading tends to introduce an extended burst of errors and missed characters. The calculated error rates include all these burst errors as well as single-character errors. Counting a burst as a single error gives an order of magnitude improvement in error rates, but that is a biased figure.

There is another effect that can introduce error bursts at the receiver. We found that it is possible for a terminal to lose synchronism with the data stream if the wrong sort of noise pulse comes along at just the right time. RTTY is an asynchronous communications method, requiring start bits and stop bits. In a continuous data stream a noise pulse can cover up a start bit, delaying the start of a character. The data stream would then be out of sync for a random period. We found that it sometimes took up to 30 character times at 45 bauds for the system to become synchronized again. This loss of sync did not happen very often so it has been left out of the error rate calculations but it will be considered further.

Table 2 shows the details of the error rates that we obtained over the entire test period. Remember that these error-rate numbers are statistical. There were evenings during which good conditions resulted in only two character errors in over 2400 characters received (at all baud rates). The tabulated error rates combine the good days with the not-so-good ones to give the averages shown.

Table 2
Error Rate Results

Baud Rate	*Errors/ Chars. Sent*	*Error Rate*
45	72/4890	1.5%
75	57/2103	3.1%
110	81/1815	5.1%

These statistical numbers are fine but what do errors do to the pictures? Fig. 5 is a photograph of a reproduced display file with typical error effects. This picture requires 308 characters to define, and there were 10 errors for an error rate of 3.2%. As can be seen, the errors show up not as noise spots (snow) in the picture but rather as obvious errors of omission or positioning of the picture details. This shows two things: first, that single errors can cause large picture faults and second, that even this high error rate can produce quality pictures.

Comments

There are several comments that can be made about our test results:

1) Error rates decrease with decreasing transmission speed as would be expected, but never approached rates that are normally encountered in telephone-line data communications (typically 1 error in 10,000 characters at 300 bauds).

2) Propagation conditions were average to good for normal amateur RTTY communications.

3) We could always produce an error-free file by using the results of three separate transmissions of a picture file to build one perfect copy. This is not a "high tech" way to do things but was good enough for our purposes.

4) Transmitting a full character time of constant mark every 25 characters or so helped reduce the length of out-of-sync bursts. The spacing required between mark characters depends on path conditions, with frequent static bursts requiring more frequent characters.

5) The original idea was to use this technique to transmit schematic diagrams but the error rates are somewhat high for that type of picture. There is, however, no problem in using NAPLPS coding for general-purpose graphic communications on hf RTTY such as QSL cards, pictures or cartoons. The use of more sophisticated error detection and correction schemes (such as longitudinal parity, over say 25 characters, and packet techniques) could improve the error rate so the transmission of schematic diagrams would be feasible.

As a result of our successful tests, we have forwarded a recommendation to the Department of Communications that NAPLPS be allowed as an Amateur Radio RTTY code set. Because it is a North American standard it seems logical that the FCC might also allow NAPLPS on the air.

There is still a lot of experimental work to be done on the technique, such as error-correcting schemes, packet-radio trials, tests on other bands and over longer paths, tests through amateur satellite links, and tests at higher baud rates. Work on any of these areas depends on being permitted to use the NAPLPS on the air.

Equipment Availability

During our tests we used Telidon decoders supplied by Norpak Ltd.[10] These decoders are available ready to use or as a board-level product that requires power supply and cabinet ($499 U.S. funds). A more advanced Mark IV unit is also available; details can be obtained from Norpak. Others sell decoders either as a board, a black box or a complete color monitor. For example, a decoder board is available for use with Apple® computers. Companies such as Electrohome, GTE and AT&T are selling or developing NAPLPS equipment. Development work is proceeding in Japan as well. In the very near future chip-level integration should bring the price of NAPLPS decoders below $200, and access to this new communications technique will be readily available to anyone who is interested.

Conclusion

In trying to find out if a picture can be described in a thousand words, we finally found ourselves using a communications technqiue that is new to Amateur Radio. After some experimentation we concluded that the transmission of pictorial information using the NAPLPS/Telidon coding method is suitable for the type of work amateurs are interested in and that further work will refine the technique to be an even more useful one. Here is an exciting new field to test the traditional ingenuity of Amateur Radio operators.

We would like to thank Norpak Ltd. for the use of the Telidon decoders and the Canadian Department of Communications for the ready approval we received when we first requested permission to use Telidon codes on the hf bands.

Ted Thompson has been a radio amateur for over 19 years, and he currently holds a Canadian Advanced Amateur Certificate (VE3FTT). He also held the call G4JCV during a year spent in England.

A professional engineer, Ted works on air-traffic-control radar and display equipment for the Canadian Department of Transport. He holds a BAS degree in electrical engineering from the University of Waterloo and an MS in aviation electronic systems from the Cranfield Institute of Technology, Bedford, England. He has been working with computers and computer graphics for 12 years.

Notes

[1]IEEE Transactions on Consumer Electronics, Vol. CE25, No. 3, July 1979, a special issue on Teletex and Videotex systems.

[2]Videotex Standards, *Computer Data*, Nov. 1982, pp. 50-54.

[3]Telidon — Videotex Presentation Level Protocol, *CRC Technical Note No. 709-E*, Department of Communications, Canada, Feb. 1982.

[4]NAPLPS: A New Standard for Text and Graphics, *BYTE*, Feb. 1983, pp. 203-255.

[5]Product Review, *Computerworld*, Dec. 13, 1982, pp. 12-14.

[6]Stable FSK Modems Featuring the XR-2207, XR-2206 and XR-2211, Application note AN-01, Exar Integrated Systems, Inc., Sunnyvale, California.

[7]"Active Filter Design with IC Op-Amps," Application note AN-03, Exar Integrated Systems, Inc., Sunnyvale, CA.

[8]"The DT-600 Terminal Unit," *Ham Radio*, Feb. 1976, pp. 8-15.

[9]km = miles × 1.6; mm = in. × 25.4

[10]Norpak Ltd., Pakenham, Ontario

New Books

HEARST BUSINESS COMMUNICATIONS IC MASTER

IC Master, Vols. 1 and 2, *edited by Dave Howell. Published by Hearst Business Communications, Inc./UTP Division, Garden City, NY. Hard-bound, 8 × 11 inches, 3677 pages, $95.*

☐ In the early days of radio, amateur gear was not nearly as complex as it is today. The typical amateur normally constructed equipment with hollow-state devices (vacuum tubes), resistors, condensers (capacitors) and a coil or two.

These simple rigs were replaced by compact, solid-state transceivers, and servicing units constructed on printed-circuit boards became commonplace. Integrated circuits have replaced much of the circuitry found in the gear of the '60s and '70s. With these ICs, the equipment became more compact and complex, to the point where the number of servicing do-it-yourselfers has dwindled to almost nil.

Why doesn't the ham down the street service his own equipment anymore? Surely, the compact layout of the circuitry is playing a role in this, but most likely it is the *lack of information* on individual IC packages that is causing the bulk of the problem.

Just what is the LJ-8764583 chip, how does it function and who makes it? With the *IC Master*, you need never ask these questions again. The *Master* is a compilation of data from over 150 IC, single-board-computer, microprocessor-development-system and PROM-programmer manufacturers, designed to provide quick, detailed information on over 55,000 commonly used ICs.

Data is compiled several ways: by application category, type, specific function, part number and manufacturer. In addition, devices are listed numerically by part number, alphabetically by application, and pin-for-pin as equivalents.

For those who like finding sources of generic ICs, the *Master* lists the generic number (555, for instance), the various manufacturers who make the chip, and the specific manufacturer's part number. Did you realize that the 555 is available from nine different sources and that 19 different part-number designators identify a 555 timer?

With such detailed information, it should be obvious that the *IC Master* is not a minor publication; in fact, the 3585-page guide is supplied in two hard-bound volumes. Hundreds of manufacturers' data sheets, applications notes and directories are included.

While not *all* ICs are contained in the *IC Master*, most of the commonly used chips are documented sufficiently. During the review period, only one of the components I needed information on was nowhere to be found. That is a very good track record, especially in the fast-paced electronics field.

Volume 1

This text consists of 14 divisions, including a master selection guide, a part-number index and a part-number guide. These indexes are organized for quick location of any IC contained in this reference manual.

A guide to manufacturer logos is a welcome section in the manual. With over 150 different logos listed, the user should have no trouble locating any manufacturer.

An extensive application-note directory follows the logo section of the *IC Master*. This section lists various available applications notes from many manufacturers, and is invaluable for anyone who is interested in using ICs for a specific application.

A military-parts directory includes a master selection guide for components used in military/aerospace electronics. This section also includes a military-to-commercial cross-reference.

Digital devices are referenced next, along with microprocessors, μP-development systems, μP boards and support boards. For the computer-oriented do-it-yourselfer, these sections prove important.

Abbreviations frequently used in IC work are listed, along with an alternate-source directory. If a component is unavailable from the local TI distributor, for instance, this section will direct the user to another manufacturer, *including* that manufacturer's exact component designation for that particular device.

Volume 1 closes with a manufacturers-and-distributors directory, to help the user locate both a manufacturer's main office address and local distribution offices.

Volume 2

This book begins with a section on interface devices, including an index of digital-to-analog and analog-to-digital converters. It also contains manufacturers' application notes for some of the more advanced devices.

Linear devices are covered in Chapter 11. As in other chapters, this section is complete — a generous helping of manufacturers' application notes are supplied.

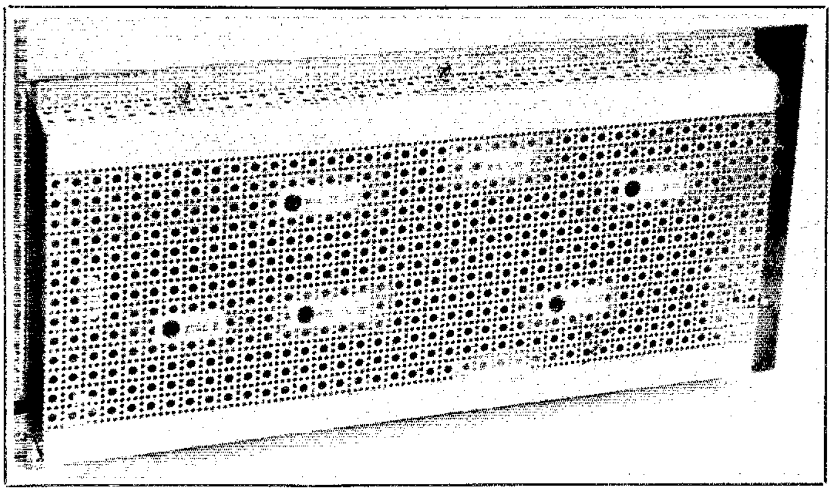
Chapter 12 is devoted to memory devices. This section contains listings and notes from several manufacturers, and is the most complete I've ever seen.

Chapter 13 contains a chart of PROM programmers from several manufacturers. Specifications and safety features are addressed, and the user is directed to the appropriate page in the chapter for more detailed information.

What manual would be complete without a chapter on special devices? Chapter 14 lists companies that supply custom digital, linear and combined digital/linear circuits. Twenty different manufacturers have supplied detailed information for the compilation of this section.

I recommended the *IC Master* for the serious "homebrewer" or the professional engineer. This reference is *not* for everyone, although 99% of hams would at one time or another find information in the books useful. — *Michael B. Kaczynski, W1OD*

Build a Satellite Transceiver Adapter



Do you need six hands to run the equipment in your satellite station? W6IOJ may have the cure for you!

By John Reed,* W6IOJ

Making AMSAT-OSCAR 8 or RS satellite contacts is similar to DXing on the hf bands — communications between stations are on or about the same frequency, and exchanges are generally limited to signal reports and station information (equipment, location, etc.).[1] Limited satellite access time and variable propagation conditions during a satellite pass encourage a similar short-QSO format. The ability to operate in the transceive mode during satellite operation would improve operating efficiency and enjoyment.

This article describes an adapter that makes use of an existing communications receiver and a variable-frequency 2-meter transmitter to provide transceive operation on OSCAR 8 Mode A or the Soviet RS satellites. The design has excellent dynamic range performance, and a converter can be added for OSCAR 8 Mode J operation.

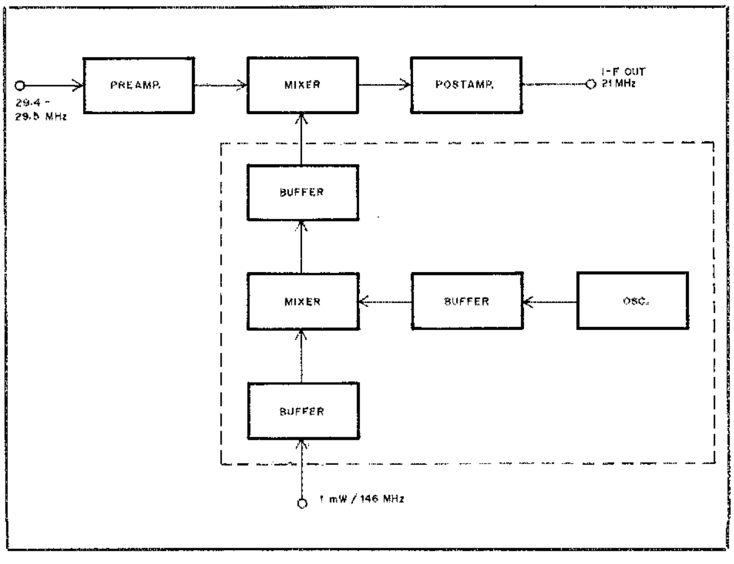


Fig. 1 — Block diagram of the Satellite Transceiver Adapter.

[1]At presstime, OSCAR 8 is not operational, although it may again become active (at unpredictable times) when it is in periods of greater exposure to sunlight.

*770 La Buena Tierra, Santa Barbara, CA 93111

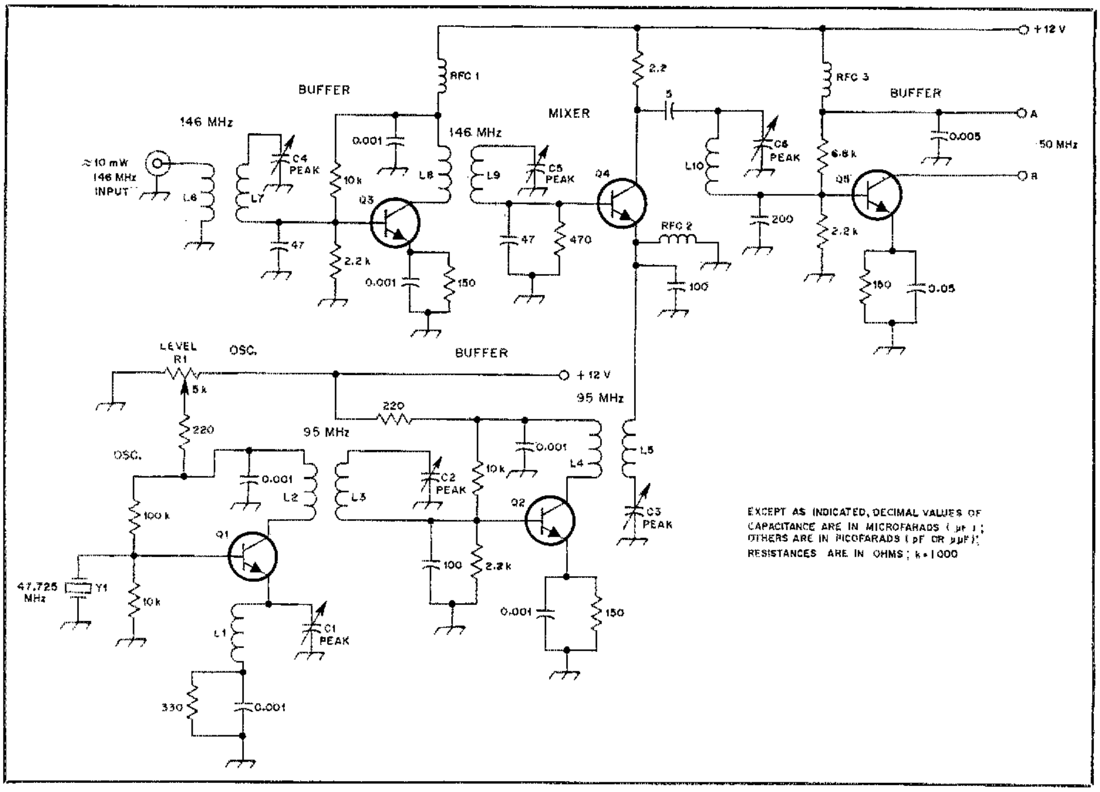


Fig. 2 — Schematic diagram of the LO portion of the adapter. Fixed-value capacitors are disc ceramic, unless otherwise noted. Polarized capacitors are electrolytic. Fixed-value resistors are 1/4- or 1/2-W carbon types, unless indicated otherwise. All inductors are air-core units wound with no. 22 bare solid wire, 3/8-inch diameter by 3/4-inch length, unless specified otherwise. Numbers in parentheses are Radio Shack catalog numbers.

C1-C6, incl. — 5-60 pF trimmer capacitor (272-1340).
L1 — 14 turns.
L2, L4 — 6 turns.
L3 — 2 turn link of no. 22 stranded hookup wire over L2.
L5 — 2 turn link of no. 22 stranded hookup wire over L4.
L6, L8 — 7 turns.
L7 — 2 turn link of no. 22 stranded hookup wire over L6.
L9 — 3 turn link of no. 22 stranded hookup wire over L8.
L10 — 9 turns.
Q1-Q5, incl. — 2N2222A npn silicon general-purpose transistor (276-2009).
Y1 — 47.725-MHz crystal (International Crystal type 031081).

Many of the parts are available from Radio Shack; all are available from mail-order houses.

Concept

The block diagram of the transceiver adapter is shown in Fig. 1. One milliwatt of 2-meter drive is required from the transmitter; this signal is mixed with a crystal-controlled first LO to produce a second LO signal tuning the range of 50.4 to 50.55 MHz. A second mixer combines the 50-MHz LO and the 29-MHz satellite signals to produce an i-f at 21 MHz. An alternate choice for the i-f would be 28 MHz; tests show performance to be equal to the 21-MHz version, except that the rf signals are isolated from the i-f by only 60 dB, introducing the possibility of interference from strong 28-MHz signals.

The advantage of this system is that one knob controls both the transmit and receive frequency. Coarse tuning is provided by the transmitter dial, and the receiver dial is adjusted to account for Doppler shift (an RIT-type function.)

Circuit Description

Local Oscillator

The LO schematic diagram is shown in Fig. 2. The crystal oscillator operates as a doubler, with the emitter circuit tuned to the *third* harmonic of the *marked* crystal frequency. A buffer stage (Q2) cleans up the oscillator signal and provides isolation. Drive from the 146-MHz transmitter is processed through a buffer stage (Q3) and injected into the base of bipolar mixer Q4. Output from this first mixer is lightly coupled to the next stage. LO output is approximately 20 mW, more than enough to drive the following passive mixer. The output level from the LO is adjustable using R1.

Signal Mixer

In this circuit application, the proximity of the i-f to the signal frequency requires the use of a doubly balanced mixer (DBM). A DBM provides superior port-to-port isolation and can be built with inexpensive components. The diode-ring DBM shown in Fig. 3 provides aproximately 40 dB of port-to-port isolation using ordinary, unmatched diodes. To ensure proper operation of the DBM, 4-dB pads are used at the rf and LO ports to provide a constant-impedance load.

Pre- and Postamplifiers

The preamplifier-stage schematic diagram is shown in Fig. 4A. Maximum

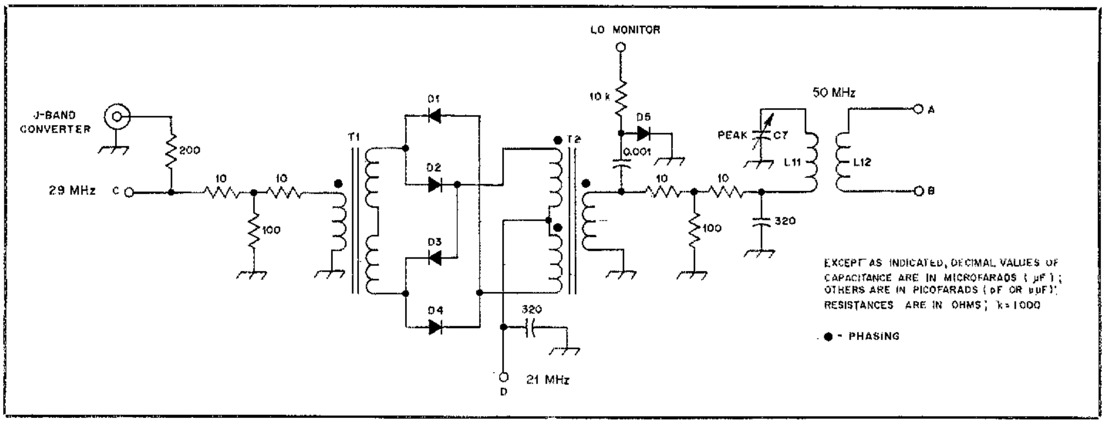


Fig. 3 — Schematic diagram of the DBM portion of the adapter. Fixed-value capacitors are disc ceramic. Fixed-value resistors are 1/4- or 1/2-W carbon types. Numbers in parentheses are Radio Shack catalog numbers.

C7 — 5-60 pF trimmer capacitor (272-1340).
D1-D4, incl. — 5082-2835 silicon barrier diodes (276-1124).
L11 — 9 turns of no. 22 solid bare wire wound on a 3/8-inch-diameter, 3/4-inch-long form.
L12 — 3 turns of no. 22 stranded hookup wire wound over L11
T1, T2 — 8 turns of no. 26 enameled wire trifilar wound on an FT37-43 toroidal core.

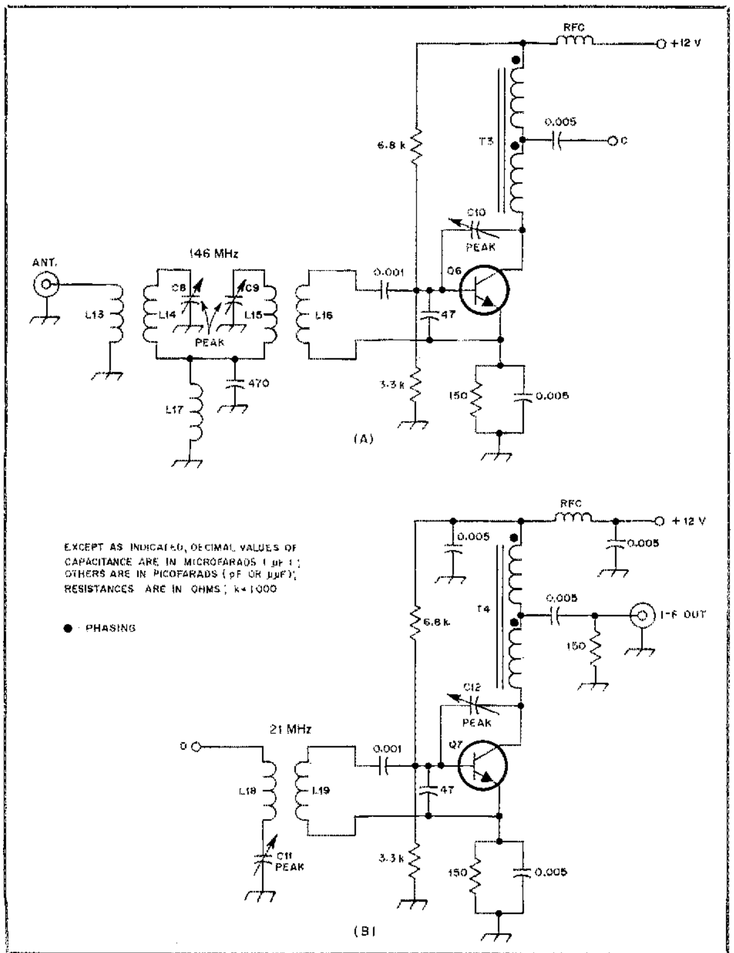


Fig. 4 — Schematic diagram of the (A) preamplifier and (B) postamplifier stages of the adapter. Fixed-value capacitors are disc ceramic. Fixed-value resistors are 1/4- or 1/2-W carbon types. Numbers in parentheses are Radio Shack catalog numbers.

C8-C12, incl. — 5-60 pF trimmer capacitors (272-1340).
L13, L16 — 2-turn links of no. 22 stranded hookup wire wound over L14 and L15, respectively.
L14, L15, — 25 turns of no. 22 enameled wire, 3/8-inch diameter by 3/4 inch long.
Q6 — MRF-901 npn silicon vhf amplifier transistor (276-2044).
Q7 — 2N2222A npn silicon general-purpose transistor (276-2009).
RFC3, RFC4 — 30 turns of no. 30 wire close wound on a high-value 1/2-W resistor.
T3, T4 — 10 turns of no. 26 enameled wire bifilar wound on an FT37-43 toroidal core.

gain is aproximately 20 dB, and the noise figure should be less than 2 dB. If large signals are present at the preamp input, gain can be reduced by decreasing the value of C10.

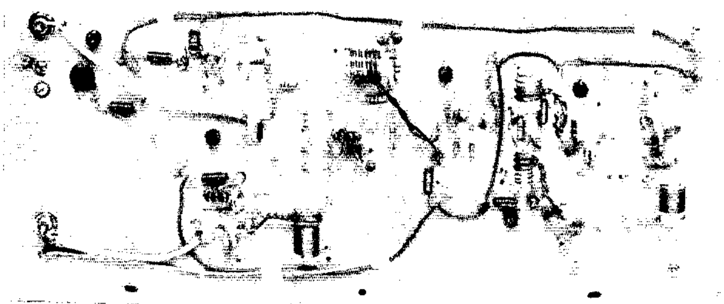
Post-mixer amplification is provided by a single bipolar stage using a 2N2222A in a circuit similar to that of the preamplifier (Fig. 4B). The gain of this stage is variable from 3 to 17 dB.

Construction

The transceiver adapter is built on a 5-1/2 × 12-in. single-sided pc board.[2] Components mount on both sides of the board using push-in terminals. Each ter-

[2]mm = in. × 25.4.

(A)



(B)

Fig. 5 — Top (A) and bottom (B) views of the transceiver adapter, illustrating construction techniques.

minal is isolated from the ground-plane foil by removing 1/8 inch of copper from around the terminal with a drill bit. It is *very* important to observe good vhf construction practices while assembling the unit — keep lead lengths as short as possible, especially between stages. Layout details can be seen in Fig. 5. Two shields are used on the top and bottom of the pc board; they are made from aluminum cane stock. This light, perforated material is easy to bend and drill.

Table 1
Adapter Performance Summary

Circuit	*Gain*	*Maximum Output*
Input	0 dB	0.1 mW
Preamp	20 dB	10 mW
Pad	– 4 dB	4 mW
DBM	– 6 dB	1 mW
Postamp	17 dB	50 mW
Total gain	27 dB	

Alignment

The initial alignment and troubleshooting work does not require the shields to be in place; the unit is very stable. Alignment is straightforward. A dc voltmeter, attached to the LO MONITOR test point and set to the 1-V range, is used to measure LO output. First, peak the crystal-oscillator trimmer capacitors for maximum output. Next, peak the remaining stages with 1 mW of 2-m drive applied. The oscillator drive control, R1, should be adjusted so that the dc meter reads 0.7 V when all stages are peaked. Tuning the preamp and postamplifier stages requires a signal source. Either a signal generator or dip meter coupled to the antenna circuit will work. Peak each tuned circuit for maximum output at the i-f port.

A performance summary for the transceiver adapter is given in Table 1. The unit has performed well at W6IOJ and continues to add enjoyment to my satellite operation. The ease of construction, alignment and operation should encourage the newcomer to try (single-handed) satellite operation.

Next Month in QST

There's some fine reading ahead for those of us into contesting and/or computers. In the October issue you'll be able to see how you (and your competition) did in the International DX Contest. And you laid-back types can prepare for the next one with an article describing a BASIC program that generates contest QSOs as well as random code (for practice, not contesting!).

Elsewhere in the issue you'll find out how to join the fun on 10 MHz in an article surveying several simple but effective antenna designs. Another article shows how to build an Amateur Radio modem. What's an Amateur Radio modem? Quite simply, it's the interface between a computer and a transceiver.

All this, and a great deal more, in October *QST*.

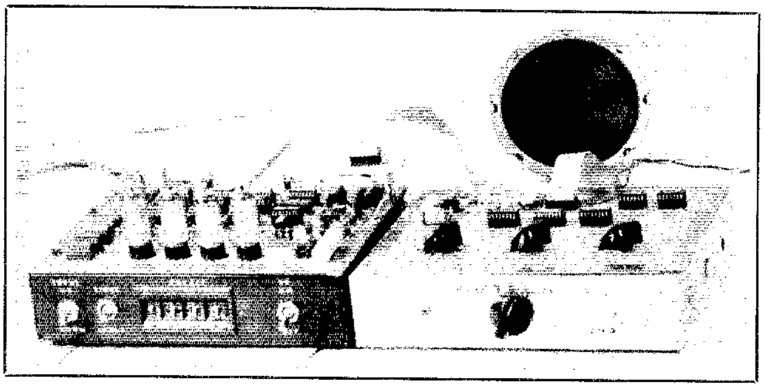
Strays

WORKING CMOS KEYBOARD MEMORY UNIT SOUGHT

☐ There have been many inquiries to the ARRL Technical Information Service about the interfacing of "An Inexpensive Morse Keyboard" (Jan. 1978 *QST*) and "Memory for the K2BLA CMOS Keyboard" (Dec. 1980 *QST*). Problems with the design have so far proved the two units incompatible. If anyone has a working memory unit for the CMOS keyboard, please contact Bob Schetgen, KU7G, Technical Information Service, ARRL, 225 Main St., Newington, CT 06111.

Amateurs at the Collins Radio Company in Cedar Rapids, Iowa, are celebrating the company's 50th anniversary by operating special event station ADØC through the end of the year. Among those manning ADØC, which will be operating before and after work hours and on weekends, are station committee members (l-r) WØNGL, WA9YZN and KØDAS. See July *QST* Special Events for operating frequencies.

The "Beeper": An Audible Frequency Readout for the Blind Amateur



A BCD-output frequency counter, a decoder and an oscillator produce a tone to tell you when your transceiver is tuned to a predetermined frequency.

By Philip S. Rand,* W1DBM

When a blind ham wishes to tune a transceiver to a given frequency, he or she must rely on the 100-kHz crystal calibrator that is usually built into the rig. He will turn the tuning dial fully counterclockwise to the low-frequency end of the band. Next, he switches on the 100-kHz calibrator and, turning the dial clockwise, counts the 100-kHz beat notes until coming to the marker nearest the desired frequency. Finally, he counts dial revolutions, having previously found the number of revolutions per 100 kHz, and from that, the number of kHz per revolution (16 kHz per revolution on my Yaesu FT-101ZD).

By this method anyone can find, for example, 3716 kHz quickly and accurately by counting two beat notes plus one dial revolution, turning the dial in a clockwise direction. Some transceivers have frequency markers every 25 kHz, and also tune almost 50 kHz out of the band at each end. With such a rig a blind ham must count 10 beat notes to reach 3700 kHz.

Suppose he wants to tune into a net on 3923 kHz. He must count 18 beat notes and continue tuning clockwise about 1-1/2 dial revolutions. He tries to locate the net by tuning around and listening for a familiar voice.

Some blind hams use an external 1-MHz crystal marker generator for finding band edges accurately. This is followed by several ICs to divide the frequency down to 100, 50, 25 or 10 kHz, as selected by a switch. Others use specially made crystals to serve as markers for their favorite frequencies.

A Better Way

Having designed a visual frequency display previously,[1] I decided to convert this unit to an audible frequency display that could be used by visually handicapped amateurs. This device allows a ham with impaired vision to tune his transceiver quickly and accurately to any frequency in any amateur band without counting beat notes or dial revolutions. The lead photo shows my "Beeper" along with the frequency counter described in note 1.

Theory of Operation

Basically, the unit consists of 5 sections, as shown in Fig. 1. These are:

1) The up/down presettable BCD-output counter.

2) A decoder to change the BCD data into decimal form.

3) A set of three program switches for setting the desired frequency.

4) Two NOR gates and one NAND gate to provide the logic that determines when you have tuned to the programmed frequency.

5) A one-shot multivibrator that triggers an audio-frequency oscillator driving a miniature loud-speaker.

The Up/Down Counter

The up/down, presettable counter is almost identical to my previous design ex-

*P.O. Box 8, Haverhill, NH 03765

[1]Notes appear on page 24.

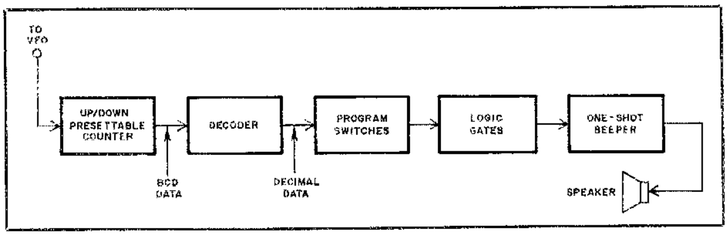
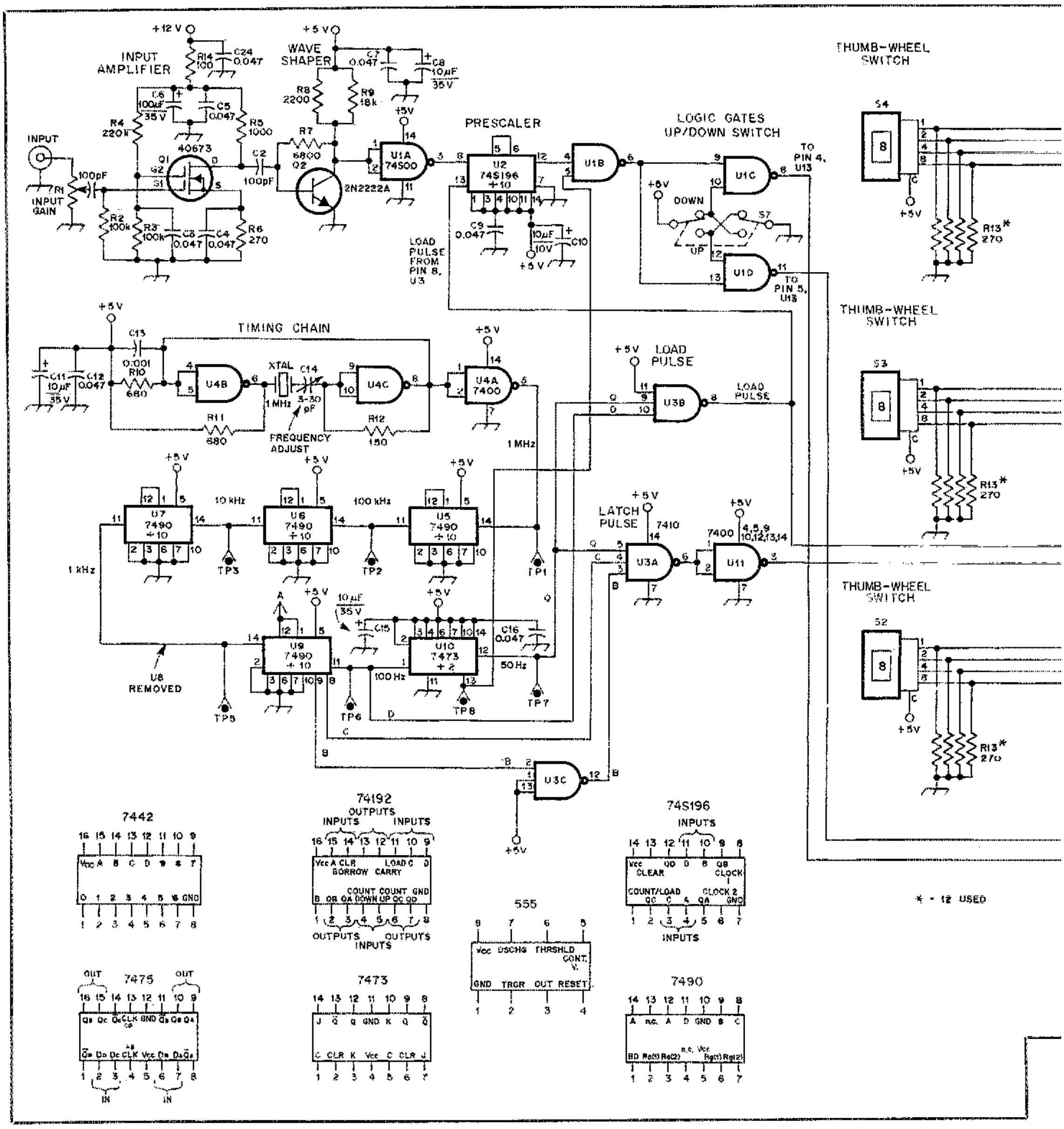


Fig. 1 — A block diagram of the audible frequency readout unit.

cept that to allow for a faster tuning rate, one divide-by-10 IC has been eliminated in the clock and a preset up/down counter has been eliminated from the counting chain. I can now turn the dial at a rate of 50 kHz per second instead of only 5 kHz per second.

As explained in note 1, an up/down presettable counter is necessary so the unit can be used with almost any transceiver regardless of VFO frequency or i-f. The only connection to the transceiver is a piece



of miniature coaxial cable to the VFO. A complete schematic diagram of the counter and readout is given in Fig. 2.

The Decoder

The decoder consists of three 7442s that decode the BCD data for the three least-significant figures of the frequency. For example, a frequency of 14,303 kHz would be decoded as 303. You know that you are tuned to 14 MHz because of the position of the band switch on the rig. The counter does not read hundreds of hertz because it would slow down the tuning rate and really is not necessary. Most nets and round tables operate plus or minus a few kilohertz as band conditions and QRM dictate.

Each 7442 has four BCD inputs and 10 decimal outputs. All outputs are high except the decoded one, which is low. A high is represented by +3.5 V or higher on the output pin of the IC, while a low is usually +0.3 V or less.

Program Switches

The three frequency-programming switches are one-pole, 10-position, non-shorting rotary switches. The output pins of each IC are connected to the 0 through 9 positions on a switch. Fig. 3 shows a means by which shorting-type switches can be used for the programming switches. The 0 through 9 positions of each switch are marked on the top panel with no. 2-56 machine screws so the blind ham can tell the position of the switch by feeling with his or her finger (Fig. 4). The left-most switch programs the 100s of kHz, the middle switch programs the 10s of kHz while

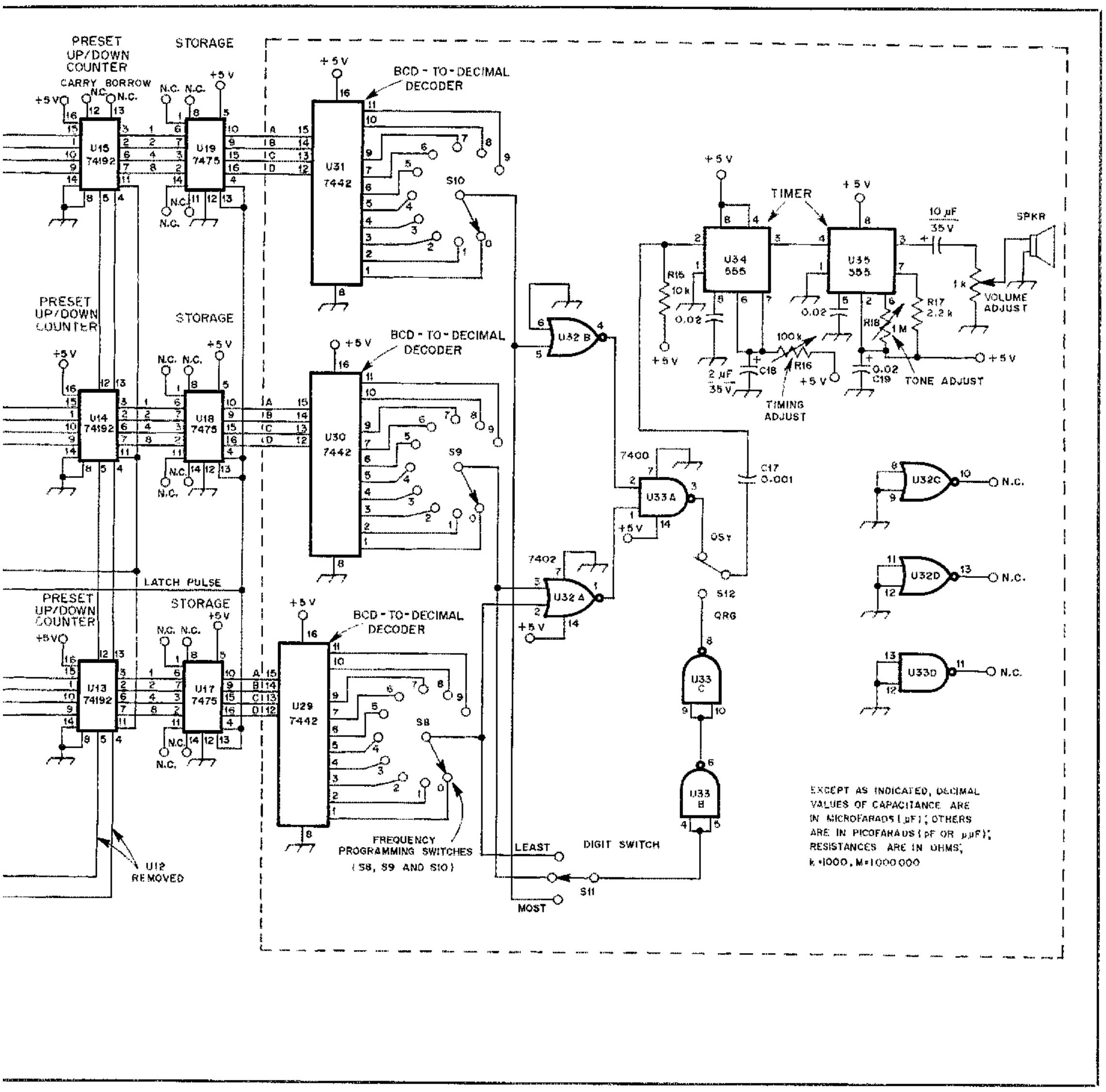


Fig. 2 — Complete schematic diagram of the frequency counter and audible frequency readout.
S2-S4 — Thumb-wheel switches, 0 to 9 BCD output.
S8-S10 — Single-pole, 10-position, non-shorting, rotary-type switches.
S11 — Single-pole, three-position rotary-type switch.
S12 — Spdt toggle switch.

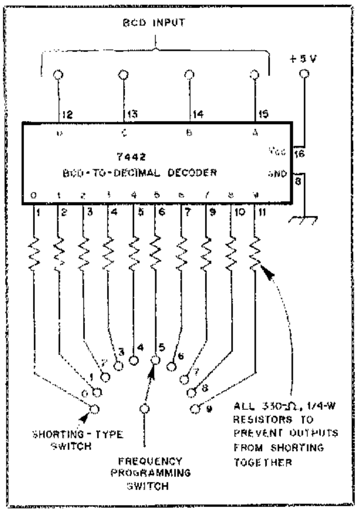


Fig. 3 — A method of using shorting-type rotary switches for the frequency-programming switches.

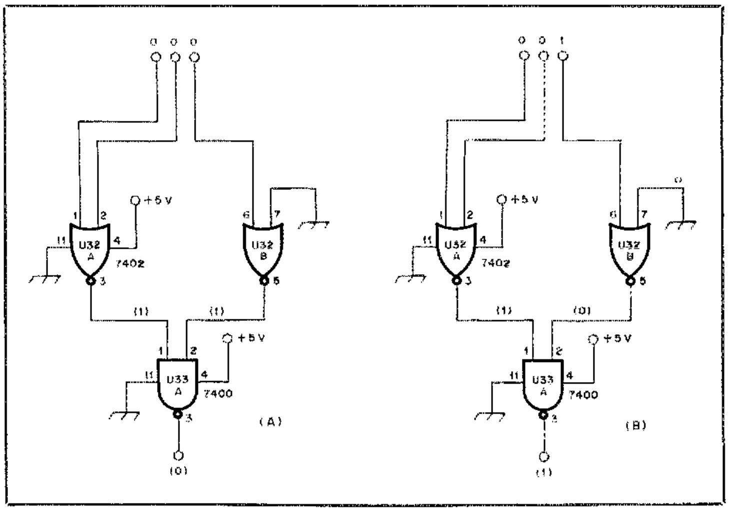


Fig. 5 — A three-input OR gate can be wired from two NOR gates and a NAND gate. The output from this gate when the transceiver is tuned to the programmed frequency is shown at A; B shows the output when the tuning is off by 9 kHz or less.

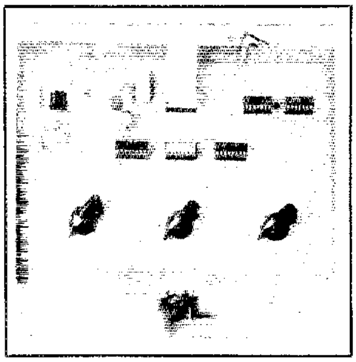


Fig. 4 — Photograph of the top panel of the decoder and beeper chassis. The three frequency-programming switches, with no. 2-56 machine-screw position markers, can be seen. No etched-circuit pattern is given because the wiring is simple, and experienced builders should be able to lay out a board if they desire to use one.

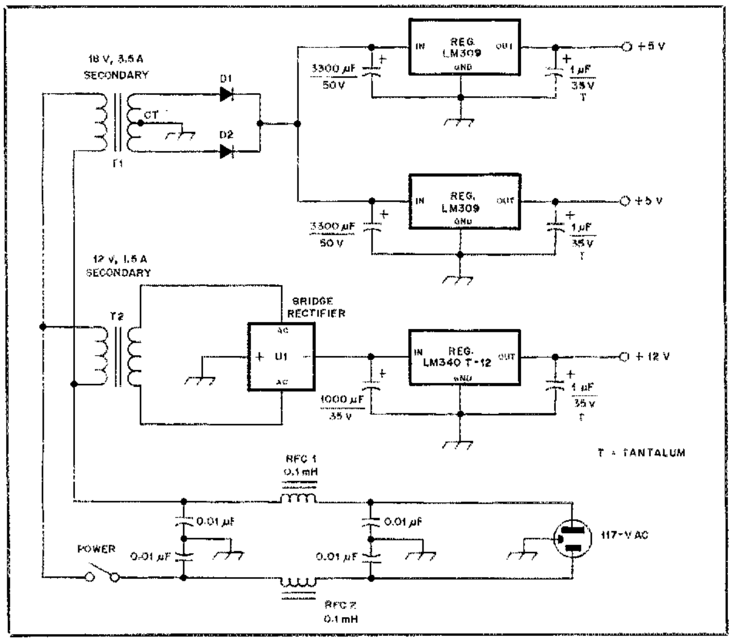


Fig. 6 — A schematic diagram of a power supply suitable for use with the complete audible frequency readout unit.

the right-most switch programs the units digit. The programmed output is taken from the movable contact on each switch and fed to the three gates.

Logic Gates

A circuit is needed that will give a low output only when three lows are fed into it. This requires a three-input OR gate. Any output containing a high will be ignored by the decoder. The OR gate is easily wired as shown in Fig. 5. A NOR gate output will be high only when both inputs are low. If one or both inputs are high, the output is low. I have wired a three-input OR gate by using two 2-input NOR gates with one input tied to ground and then combining their outputs with a NAND gate. A NAND gate only

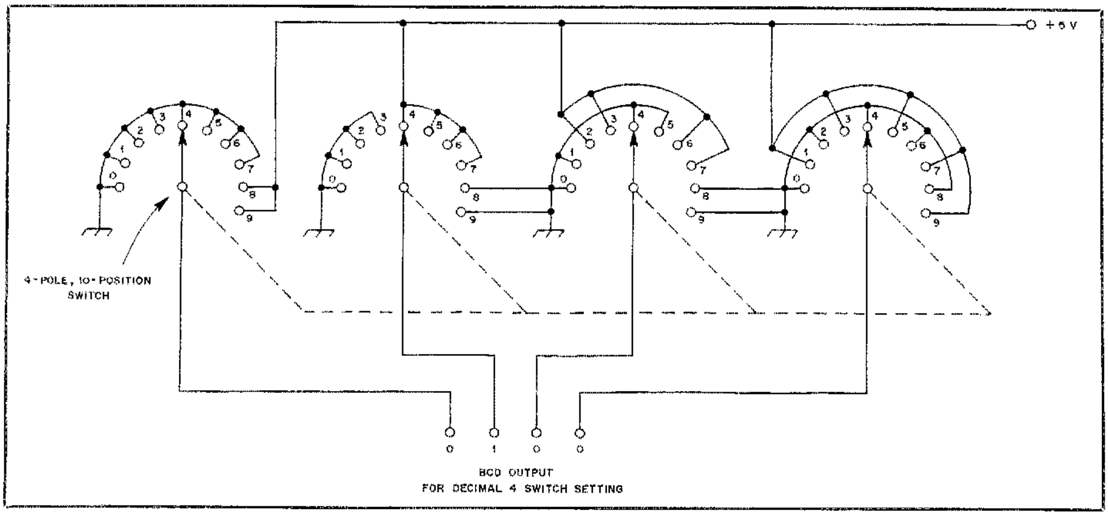


Fig. 7 — A method of wiring a four-pole, 10-position rotary switch to provide a BCD output is shown.

produces a low output when both inputs are high. Referring to Fig. 5A, note that when you have tuned to the programmed frequency, three lows are present at the input to the NOR gates, producing two highs at the input of the NAND gate and forcing its output low. Fig. 5B illustrates the situation when you have correctly tuned to the left and middle digits of your programmed frequency but are still as much as 9 kHz away from the exact frequency.

The Beeper

We need a low output from the NAND gate to trigger the 555 one-shot multivibrator shown in Fig. 2. Since the output from the NAND gate is high up to the moment that you tune to the exact programmed frequency, and then falls to 0, C17 and R15 are used to differentiate this wave form to produce the necessary pulse for the one-shot. The output of the one-shot is timed by R16 and C18 for about 1/10 of a second and is fed to the control pin of the second 555, an audio-frequency oscillator that generates a tone in a miniature loud speaker. The tone of the beep is controlled by the values of R17, R18 and C19. R15 ensures that the voltage on pin 2 of the one-shot is always high except during the short interval of the trigger pulse. The trigger pulse always must be shorter than the one-shot output pulse.

Construction

You have several choices in building this audible frequency readout. If you own a rig with a digital frequency display you may omit the counter portion and build only the decoder and beeper sections. In this case you must tap into the BCD lines with a 14-wire cable. This may be difficult because of the compactness of modern solid-state transceivers and the fact that some use single-chip counters with no BCD output available. Generally, it is much easier to connect a short length of miniature coaxial cable to the VFO output and use an external counter.

If your rig does not have a built-in frequency counter, then you will need to build or acquire a suitable counter. If you elect to build the counter from scratch, the entire unit will fit on one 7- × 9-inch circuit board.[3] All the switches go on the front panel.

A power supply can be included in the cabinet. See Fig. 6 for the circuit diagram of a suitable supply. The power-supply requirements are reduced by not using the visual readout. A brute-force line filter is included to help prevent RFI problems.

If you already built the counter described in note 1, you may do as I did, and build just the decoder/beeper on a circuit board to be mounted on a separate 7- × 9- × 2-inch chassis. The Beeper is connected to the counter by means of a flat 16-wire cable with 16-pin DIP plugs on each end (a standard Radio Shack item). The new circuit board and switches could also be mounted directly over the counter. If you are using a circuit board from Circuit Board Specialists[3], be sure to put jumpers from input to output across the spaces for the two divide-by-10 ICs that are not used. The BCD outputs are available at the input pins of the old 7448 sockets.

Thumb-wheel Switches

Since we are using only three digits in the audible readout, three thumb-wheel switches are needed for presetting the up/down counters instead of the four used in the original counter. If you cannot locate 0-to-9, BCD-output, thumb-wheel switches, you can make substitute switches from 4-pole, 10-position rotary switches by connecting them as shown in Fig. 7.

Calibration

Calibration of the up/down presettable counter is necessary because the unit is designed to work with any rig that you may have, with or without a frequency display. When shifting from band to band, from upper to lower sideband or to cw, an additional adjustment of the thumb-wheel switches will be necessary.

To calibrate, simply tune to a known frequency containing three 0s (such as 7000 kHz) using the 100-kHz crystal marker. Now set the program switches fully counterclockwise to 000, place the QSY/QRG switch in the QRG position and set the three-position "digit" switch to the right, the least significant digit. Next, rotate the right-hand thumb-wheel switch until you hear a beep. Now set the digit switch to the middle position and rotate the middle thumb-wheel switch until you hear another beep. Finally, with the digit switch in the most significant position, fully counterclockwise, rotate the left thumb-wheel switch for a third beep. The counter is now calibrated for that particular band and mode of operation.

Next it is necessary to check for up/down counting. To do this, set the program switches to read 100 and then tune the rig to the next 100-kHz marker, turning the dial clockwise. If the unit beeps as you reach zero-beat, it means you are counting in the right direction. If not, you must change the up/down switch and repeat the calibration. Rigs such as the Heathkit

HW-101, Yaesu FT-101 and others using the same conversion logic require the switch to be in the count-down position. This need only be checked once for your rig.

Using the "Beeper" to QSY

To tune rapidly from one part of a band to a net frequency, simply program the desired frequency with the three switches, set the QSY switch and crank the dial one way or the other until you hear a beep. If you were tuning fast you probably overshot the mark, so tune slowly in the other direction a few kilohertz until you hear another beep. You are now on frequency, plus or minus 1 kHz if you calibrated the up/down counter for the band and mode of operation in your particular transceiver.

QRG? What Is My Frequency?

To read the frequency you are tuned to, it is necessary to question each digit, one at a time. This is not as hard as it sounds because you usually know about where you are tuned: You may only have to question the least significant digit. For example, set the QSY/QRG switch to QRG, set the digit switch to the right (the least significant digit) and rotate the right-hand program switch until you hear a beep. Let's assume you are working in the General class portion of the 20-meter phone band and you think your frequency is somewhere between 14,300 and 14,310 kHz. You determine by feeling the screw heads with your finger that the switch is in the 3 position. It's a pretty good bet that you are on 14,303 kHz. It could not be 14,403 or 14,203 but perhaps it is 14,313. To be sure, set the digit switch to the middle position and rotate the middle program switch for a beep. Your finger tells you that the second digit is in fact a 0. If you want, you can also confirm that the left-hand digit is a 3. Of course the 14 MHz is determined by the band switch on the rig.

Conclusions

I gave some thought to using a voice synthesizer to produce an audible frequency readout, but this seemed to be a needless complication. Most operators would not want to move a few kilohertz, wait for a voice announcement of the frequency and tune a few more kilohertz, repeating this procedure until finally reaching the desired frequency. The circuit required for a voice synthesizer would be more complicated, and construction of the unit would be more difficult. The device just described is simple to build and easy to use. It enables a ham with impaired vision to tune to any desired frequency quickly and accurately.

Notes

[1]Rand, "A Versatile Digital Frequency Display," *QST*, Nov. 1977, page 21.

[2]mm = in. × 25.4

[3]Circuit boards and complete parts kits for the original frequency counter and the audible frequency readout are available from Circuit Board Specialists, P.O. Box 969, Pueblo, CO 81002.

Strays

TEACHING THROUGH AMATEUR RADIO

☐ Have you ever participated in a historical event, or met a famous person? Why not tune to 21.395 MHz between 1200 and 2100 UTC Monday through Friday, and share that experience with members of the NYC Junior High School No. 22 ARC? See the difference you can make in a youngster's education. — *Joseph J. Fairclough, WB2JKJ, New York, New York*

Wayland "Soupy" Groves, W5NW (left), of Odessa, Texas, was honored recently with a party put on by the members of the West Texas ARC in recognition of his 60 years of membership in the ARRL, 35 of which were spent on the Board of Directors. Helping Soupy celebrate were club President K5IID (center) and ARRL West Gulf Division Director W5EDZ. *(photo by Rainey)*

MOVING, CHANGING CALL?

☐ When you change your address or call sign, be sure to notify the Circulation Department at ARRL Hq. Enclose a recent address label from a *QST* wrapper if at all possible. Address your letter to Circulation Department, ARRL, 225 Main St., Newington, CT 06111. Please allow six weeks for the change to take effect. Once we have the information, we'll make sure your records are kept up-to-date so you'll be sure to receive *QST* without interruption. If you're writing to Hq. about something else, please use a separate piece of paper for each request.

IHN HAS NEW MANAGER

☐ Cal Burt, KAØDFN, is the new manager of the International Handicappers' Net, succeeding Ray Meyers, W6MLZ, sponsor of the IHN for the past 25 years. KAØDFN, a retired Air Force colonel, was recently appointed to the President's Committee for the Employment of the Handicapped. Membership in the IHN costs but a letter stating your call sign and the nature of your handicap, and a business-size s.a.s.e. to hold your certificate. For more information, write to Cal Burt, KAØDFN, P.O. Box 59, Oregon, MO 64473.

INTERESTED IN LEAGUE-SPONSORED INSURANCE?

☐ The Membership Affairs Committee is studying the feasibility of offering League members additional insurance programs beyond the present ARRL Ham Radio Equipment and Club Liability programs. Participation in the plans would be voluntary. Please take a few moments to complete the following questionnaire and send it to the Membership Services Department, ARRL, 225 Main St., Newington, CT 06111. Make a photocopy if you'd like, or simply answer the questions on a separate piece of paper. Thanks.

I would be interested in the following types of voluntary insurance plans, should the ARRL offer them:

	I am interested	I am not interested
1) Life Insurance	☐	☐
2) In-Hospital Insurance (to pay a fixed amount each day during hospitalization)	☐	☐
3) High-Limit Accident Insurance (to pay specified amounts for specific levels of injury due to accident)	☐	☐
4) Disability Income Protection	☐	☐
5) Excess Major Medical (to pay medical charges above a deductible of $15,000 to $25,000)	☐	☐
6) Major Medical (to pay medical expenses above a deductible of $500 to $1000 but with a maximum benefit of $20,000)	☐	☐

I do not believe the ARRL should sponsor insurance programs that have no direct relation to Amateur Radio. ☐

Please note: An expression of interest in no way binds you to a particular course of action. This survey is designed only to gauge membership interest in these insurance programs.

A Top-Fed Vertical Antenna for 1.8 MHz — Plus 3

Mathematical analysis unlocks the secret behind this unusual antenna system.

By Carl Eichenauer,* W2QIP

Does your antenna act strangely at times? Does it resonate where it is not supposed to? Well, mine does. To understand why, I developed a BASIC computer program to "crunch" through various mathematical calculations. The program is listed in the Appendix. My "strange" installation consists of 80- and 40-meter inverted-Vs, fed from a single coaxial cable and mounted on a common support.

There is nothing unusual about this system; the approach is described in *The ARRL Antenna Book.*[1] The antenna shown in Fig. 1 provides satisfactory performance on 80, 40 and 15 meters. It also provides — strange as it may seem — effective radiation on 160 meters! No traps, switches or antenna matching networks are required to accomplish 4-band operation.

A certain amount of skepticism is in order at this point — in fact, I could not have been more surprised. After trying to tune the antenna on 160 m with the aid of a Transmatch, I eventually found that the system had a SWR of 1.2:1 on 1805 kHz with *no matching network* at all!

The following description and analysis of my antenna system should present ideas useful to other antenna builders. Those familiar with BASIC programming may eliminate many hours of number crunching on a calculator by using a program similar to mine.

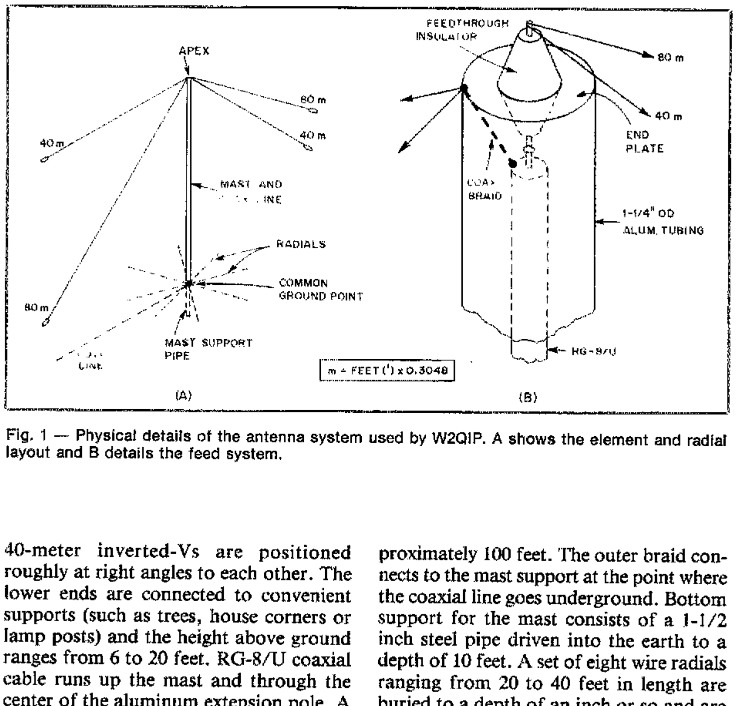


Fig. 1 — Physical details of the antenna system used by W2QIP. A shows the element and radial layout and B details the feed system.

Physical Details of System

A 34-foot wooden mast with a 12-foot aluminum-pole extension provides a center support for both antennas.[2] The 80- and 40-meter inverted-Vs are positioned roughly at right angles to each other. The lower ends are connected to convenient supports (such as trees, house corners or lamp posts) and the height above ground ranges from 6 to 20 feet. RG-8/U coaxial cable runs up the mast and through the center of the aluminum extension pole. A small aluminum plate covers the top end of the aluminum pole and a ceramic feed-through insulator passes through the plate. The four antenna wires, in addition to an insulated guy wire, support the top of the antenna. The wooden mast is also guyed.

From the feed point, the coaxial line runs down the mast to ground level and then continues underground (at a depth of several inches) to the house, through the basement, and eventually reaches the operating position. Total line length is approximately 100 feet. The outer braid connects to the mast support at the point where the coaxial line goes underground. Bottom support for the mast consists of a 1-1/2 inch steel pipe driven into the earth to a depth of 10 feet. A set of eight wire radials ranging from 20 to 40 feet in length are buried to a depth of an inch or so and are connected to the coaxial-cable braid where it goes underground. This ground network turns out to be an important part of the 160-meter radiation system.

System Modeling

In an attempt to understand the mysterious antenna resonance at 160 meters, I investigated several analytic techniques. The technique that gave the closest agreement with measurements is based on principles set forth in Jordan's

[1]Notes appear on page 27.

*205 Lathrop Rd., Syracuse, NY 13219

book.[3] He points out that while lumped-constant circuit elements can be used to model an antenna system over a narrow band of frequencies, transmission-line simulation of the antenna elements is a more accurate approach when analyzing an antenna over a broad frequency range.

The following 10 steps and the BASIC program outline the analysis of my antenna system. Each step is based either on transmission-line or radiation-resistance formulas. In each case, the formulas are functions of frequency.

The six conductors represented in Fig. 2 consist of two 80-meter dipole halves, two 40-meter dipole halves, and the center conductor and braid of the coaxial feed line. For the purpose of mathematical modeling, we will first assume that the six antenna conductors are suspended in free space.

1) Fig. 2A shows the two 80-meter antenna elements. Each wire can be analyzed as a sloping transmission line in which one conductor is a combination of the earth and antenna radial system. Inspection of the actual antenna shows that the input end of the line (I) is 46 feet high, the output end (K) is 6 feet high and the output load impedance is infinite. This information along with wire diameter and frequency-of-operation data is placed into an appropriate formula and the reactance looking into each half of the antenna (X8) is calculated. Next, the same data is used in a different formula to calculate the radiation resistance looking into the two elements acting as a dipole. One-half of this value is assigned to each element (R8). The key electrical properties of these two "radiating transmission lines" are now defined.

2) The same procedure used in step 1 is used to calculate R6 and X6.

3) Next, the 40- and 80-meter elements are connected together and their total reactance is calculated. This value (X9) is simple — the parallel value of X6 and X8 (Fig. 2C).

4) Fig. 2D represents the analysis of the coaxial line. Since the bottom end is connected to ground, only the 46-foot section shown is relevant to the model. Furthermore, since the line is open-circuited at the top end (for the moment), nothing but a source of voltage appears at the upper end (assuming, of course, that the signal generator is turned on and that it can operate into an infinite-impedance load).

5) As an alternative equivalent, the transmission line can be represented by a solid conductor whose diameter is equal to the O.D. of the transmission-line outer braid (Fig. 2E). The voltage from the transmitter is represented simply by a two-terminal sine-wave generator with one terminal connected to the vertical conductor and one terminal open circuited (for the moment).

6) Next, the two sets of conductors from step 2C are "connected" to the vertical conductor top. From the reactance of this combination, X9, the equivalent electrical length (in degrees) of the elements can be determined for any given frequency. The vertical conductor electrical length (in degrees) is calculated from the physical length. These two sections form a top-loaded vertical antenna. By "lifting" the bottom connection off ground, the base radiation resistance of the top-loaded vertical can be determined (Fig. 2F).

7) An unfortunate fact of nature must be accounted for — no ground system is perfect. Therefore, a resistor (RG, the ground loss) is inserted in series with the radiation resistance of the vertical, as

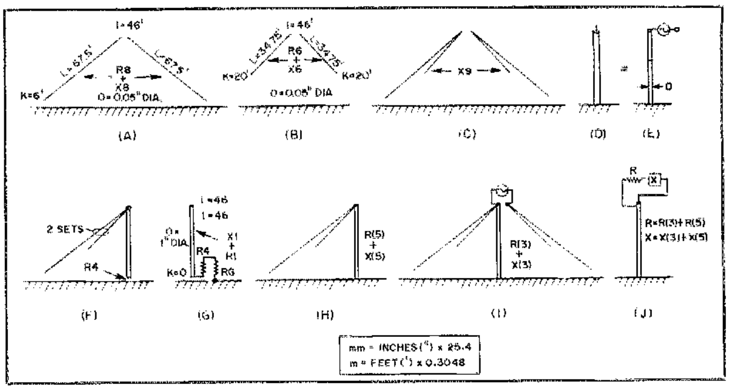


Fig. 2 — Step-by-step development for the mathematical model of the antenna. See text for details.

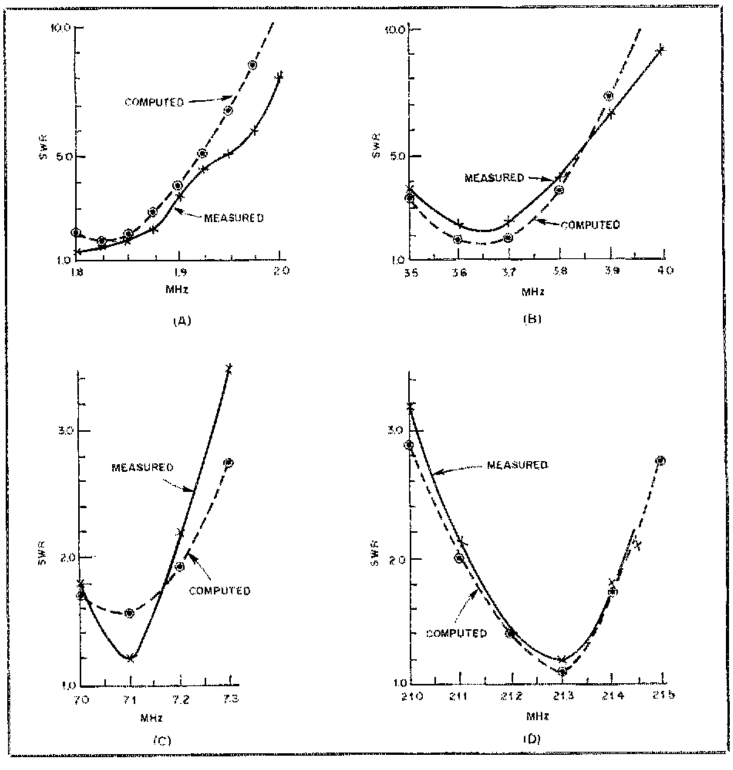


Fig. 3 — Measured and calculated SWR curves for the antenna on (A) 160 meters, (B) 80 meters, (C) 40 meters and (D) 15 meters.

shown in Fig. 2G. Consider the 46-foot vertical section as another transmission line in which the input end is at 46 feet above ground and the output end is terminated in a resistor of value R1 + RG. This information, along with the vertical-section outside diameter and operating frequency is used to determine the values of X1 and R1, the reactive and resistive components of the vertical transmission line.

8) In Fig. 2H one set of sloping elements is reattached. Now, the series equivalent value for the whole system is calculated.

9) In Fig. 2I one terminal of a sine-wave generator is attached to the vertical element and the set of wires and the other set of sloping wires is connected to the other terminal of the generator. This allows the equivalent series reactance and resistance for the complete set of wires to be calculated.

10) Fig. 2J displays the equivalent result of the modeling. The resistances and reactances of steps 8 and 9 are added together to give the effective input impedance for the complete antenna system. Since this representation is broadband in nature, the system SWR can be checked at any frequency by "sweeping" the signal source!

Calculated vs. Measured Results

An SWR meter at the transmitter end of the transmission line was used for SWR measurements in all cases except one; because SWR values from below the 160-meter band were desired, a home-made RX bridge was employed.

Fig. 3 shows the measured and calculated SWR curves. Resonance occurs near the bottom edge of the 160-meter band. Fortunately, virtually all of my operation on 160 is cw. For operation higher in the band, the computer program showed that lowering the height of the apex by a foot or two should provide satisfactory performance.

Note that the calculated SWR is consistently higher than the measured values. This characteristic can be modified by selection of the value of ground resistance used in the calculations. In the case shown, a value of 5 ohms has been arbitrarily selected. If a lower value is used, the lowest calculated SWR coincides almost exactly with the lowest measured value, but the off-resonance SWR rises at a more rapid rate than the measured values. This simply points out the fact that the model is representative, but not exact. In defense of the model, it should be pointed out that SWR measurements are seldom exact either!

The SWR curves for the other bands present few surprises. Over the years, similarly installed dipoles and inverted Vs have produced essentially the same response characteristics.

Conclusions

If you wonder whether this antenna really works, I must confess that it has not been responsible for a 160-meter DXCC award. However, in a recent ARRL 160-meter contest it was used with a QRP rig running three watts input. Over a three-hour period 50 stations in 22 states were worked. If nothing else, this should prove that the system does radiate on 1.8 MHz!

Appendix

The BASIC program listed below was developed on a Timex/Sinclair 1000® computer with 16K of RAM. [The program will run on the 2K T/S 1000 if the REM statements are removed from the program. — Ed.] The program follows the analysis algorithm presented in the text. Readers familiar with BASIC programming should have no trouble in adapting the program to any BASIC-equipped machine.

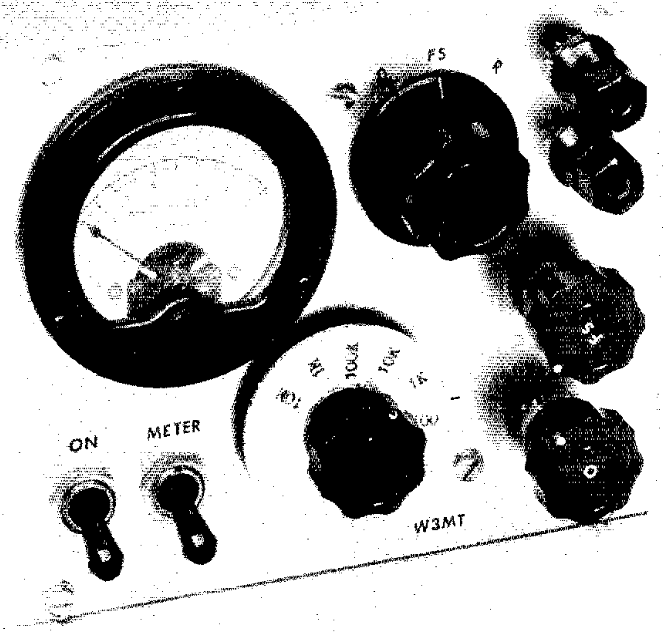
Notes

[1]G. Hall, Ed., *The ARRL Antenna Book*, 14th ed. (Newington: ARRL, 1982), chapter 8.
[2]m = ft × 0.3048; mm = in. × 25.4.
[3]Jordan, *Electromagnetic Waves and Radiating Systems* (Englewood Cliffs, NJ: Prentice-Hall, 1950), chapter 13.

```
10 REM FILE NAME IS ANTENNA2
20 FOR F=1.8 TO 2.0 STEP .025
30 LET P=3.1415927
40 DIMR(10)
50 DIMX(10)
60 REM 80 M.SLOPER REL.TO GND.
70 LET L=[illegible]
80 LET Q=.05
90 LET I=46
100 LET K=5
110 LET D=(I+K)/2
120 LET R4=0
130 LET X4=-1E6
140 GOSUB 400
150 LET X8=X5
160 LET R8=.01215*(ABS(1-COS(ATN(B)))*180/P)↑2
170 REM 40 M.SLOPER REL.TO GND.
180 LET L=24.75
190 LET Q=.05
200 LET I=46
210 LET K=20
220 LET D=(I+K)/2
230 LET R4=0
240 LET X4=-1E6
250 GOSUB 400
260 LET X6=X5
270 LET R6=.01215*(ABS(1-COS(ATN(B)))*180/P)↑2
280 REM VERTICAL REL.TO GND.
290 LET L=46
300 LET Q=1
310 LET I=46
320 LET K=0
330 LET D=(I+K)/2
340 LET X4=0
350 LET R4=0
360 GOSUB 400
370 LET R1=R5
380 LET X1=X5
390 GOTO 550
400 REM Z SUBROUTINE
410 LET B=TAN(2*P*L*F/984)
420 IF X4=0 GOTO 450
430 LET Z0=60*LOG(48*D/Q)
440 GOTO 510
450 LET Z0=60*(LOG(24*L/Q)-1)
460 LET X9=X8*X6/(X8+X6)
470 LET B1=ATN(Z0/ABS(X9/2))
480 LET B2=ATN(B)
490 LET R4=.01215*(ABS(COS(B1)-COS(B1+B2))*180/P)↑2
500 LET R4=R4+5
510 LET R3=R4/Z0
520 LET X3=X4/Z0
530 GOSUB 900
540 RETURN
550 REM PAR.R1 AND X1 VALUES
560 LET R=(R1↑2+X1↑2)/R1
570 LET X=(R1↑2+X1↑2)/X1
580 REM PAR.EQUIV.OF 80 SLOPER
590 LET R(6)=(R8↑2+X8↑2)/R8
600 LET X(6)=(R8↑2+X8↑2)/X8
610 REM PAR.EQUIV.OF 40 SLOPER
620 LET R(1)=(R6↑2+X6↑2)/R6
630 LET X(1)=(R6↑2+X6↑2)/X6
640 REM PAR.EQUIV.OF 80&40 SLOPERS
650 LET R(2)=R(6)*R(1)/(R(6)+R(1))
660 LET X(2)=X(6)*X(1)/(X(6)+X(1))
670 REM SER.EQUIV.OF 80&40 SLOPERS
680 LET R(3)=X(2)/(X(2)/R(2)+R(2)/X(2))
690 LET X(3)=R(2)/(X(2)/R(2)+R(2)/X(2))
700 REM PAR.EQUIV.OF80&40 SLOPERS+VERT.
710 LET R(4)=R(2)*R/(R(2)+R)
720 LET X(4)=X(2)*X/(X(2)+X)
730 REM SER.EQUIV.OF 80&40 SLOPERS+VERT.
740 LET R(5)=X(4)/(R(4)/X(4)+X(4)/R(4))
750 LET X(5)=R(4)/(R(4)/X(4)+X(4)/R(4))
760 REM SER.VALUES FOR ANTENNA
770 LET R1=R(5)+R(3)
780 LET X1=X(5)+X(3)
790 REM REFLECTION COEF.
800 LET V1=SQR((R1-50)↑2+X1↑2)
810 LET V2=SQR((R1+50)↑2+X1↑2)
820 LET V=V1/V2
830 REM VSWR
840 LET V=(1+V)/(1-V)
850 PRINT"F=";F;" X=";X1
860 PRINT"R=";R1;" VSWR=";V
870 PRINT
880 NEXT F
890 STOP
900 REM LINE R AND X SUBROUTINE
910 LET C=(1-X3*B)↑2+(R3*B)↑2
920 LET R=R3*(1+B↑2)/C
930 LET X=(X3*(1-B↑2)+B*(1-R3↑2-X3↑2))/C
940 LET R5=R*Z0
950 LET X5=X*Z0
960 RETURN
970 END
```

```
*RUN

F=  1.8    X= -36.79183
R=  77.79574  VSWR=  2.060922

F=  1.825  X= -18.213753
R=  86.00515  VSWR=  1.743639

F=  1.85   X=  21.77462
R=  95.5599   VSWR=  2.046966

F=  1.875  X=  54.04112
R=  106.7074  VSWR=  2.791916

F=  1.9    X=  88.28508
R=  119.8171  VSWR=  3.855286

F=  1.925  X=  125.0536
R=  135.3697  VSWR=  5.124723

F=  1.95   X=  164.747
R=  153.9985  VSWR=  6.782116

F=  1.975  X=  207.8211
R=  176.5483  VSWR=  8.590436

F=  2      X=  254.7821
R=  204.1618  VSWR=  10.5928

ready
*20 FOR F=3.5 TO 4.0 STEP .1
*RUN

F=  3.5    X=-73.5251
R=  66.48304  VSWR=  2.415184

F=  3.6    X= -24.44687
R=  74.61266  VSWR=  1.751711

F=  3.7    X=  26.76289
R=  77.6977   VSWR=  1.837677

F=  3.8    X=  81.90717
R=  76.59321  VSWR=  3.56362

F=  3.9    X=  142.457
R=  77.41974  VSWR=  7.299825

F=  4      X=  210.934
R=  80.77132  VSWR=  13.17314

ready
*20 FOR F=7.0 TO 7.3 STEP .1
*RUN

F=  7      X= -24.09847
R=  75.45659  VSWR=  1.722581

F=  7.1    X=  4.48773
R=  73.68868  VSWR=  1.582331

F=  7.2    X=  31.80422
R=  75.99452  VSWR=  1.905327

F=  7.3    X=  60.01251
R=  76.04982  VSWR=  2.763779

ready
*20 FOR F=21 TO 21.5 STEP .1
*RUN

F=  21     X= -50.86379
R=  48.20148  VSWR=  2.854866

F=  21.1   X= -24.9547
R=  49.26755  VSWR=  1.994922

F=  21.2   X= -12.06762
R=  49.21908  VSWR=  1.408788

F=  21.3   X=  5.17015
R=  45.08509  VSWR=  1.119743

F=  21.4   X=  24.11262
R=  40.9635   VSWR=  1.75343

F=  21.5   X=  45.62159
R=  39.40734  VSWR=  2.750102
```

A Linear, Self-Calibrating Ohmmeter



Do you want an accurate, easy-to-use ohmmeter, but can't afford a digital unit? Try your hand at putting together this easy-to-build project, and you'll have one.

By Frank Noble,* W3MT

A conventional ohmmeter has a nonlinear scale. At high resistance values the meter is hard to read because the scale graduations are so close together. It may be difficult to interpolate between the calibration points at the low-resistance end of the scale. The meter accuracy is degraded as the battery approaches the end of its life. While most of these disadvantages can be overcome by providing additional ranges, the construction of a meter scale remains a major problem if you want to build your own ohmmeter. Finally, most multimeters pass a relatively large current through the device under test, and this may be sufficient to destroy some components.

The nonlinear scale of the conventional ohmmeter causes larger reading errors with increasing resistance. For best accuracy the instrument should be calibrated against standards in the range of the unknown each time it is used. This article describes a linear-scale meter, self-calibrating against internal standards, which dissipates very little power in the unknown resistor. It has six decade ranges from 100 ohms to 10 megohms, full scale, so that any resistance in the range 10 ohms to 10 megohms will yield a meter deflection of at least 10% of the scale. Neben presented a similar meter, but the circuit described here includes several design improvements.[1]

The Scheme

An understanding of the system requires a quick review of op-amp theory. In Fig. 1, the amplifier is assumed to have infinite gain, infinite input impedance, zero output impedance and exact phase inversion. Because the gain is infinite and the output voltage is finite, the voltage at the amplifier input is zero. Hence the current in R_A is just

$$I = E_i/R_A \qquad \text{(Eq. 1)}$$

None of this current can enter the amplifier because any current flowing through an infinite impedance will produce infinite voltage, saturating the amplifier. The only recourse is for the same current to flow in R_B, forced there by E_o.

$$I = E_o/R_B \qquad \text{(Eq. 2)}$$

Setting Eq. 1 and 2 equal, we have the usual

*10004 Belhaven Rd., Bethesda, MD 20817

[1]Notes appear on page 30.

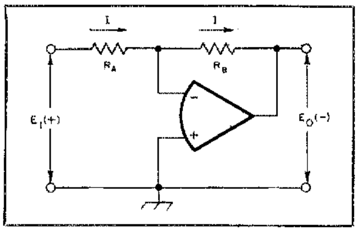


Fig. 1 — Diagram of a simple inverting amplifier. The theory of operation is explained in the text.

formula for the gain of an inverting op amp.

$$G = E_o/E_i = R_B/R_A \qquad \text{(Eq. 3)}$$

Design Considerations

To minimize the power dissipation in the unknown resistor, E_o must be as small as possible, but still drive the meter to full scale. Although almost any meter movement could be arranged to read low voltages, a sensitive meter is indicated because it must be protected from transients that can occur when you switch ranges, disconnect the unknown, or turn the meter on with no unknown connected. Diodes are not suitable elements for this protection because they will degrade the scale linearity. A large R-C time constant is used instead. (Fig. 2). While this does not eliminate the meter overload, it does slow the deflection rate so the meter will not be damaged mechanically. The internal resistance of my 100-μA meter is 1800 ohms, as compared to 50 ohms for a typical 1-mA meter. The larger meter resistance makes it possible to slow the meter movement with a much smaller capacitor for C6.

Optimum sensitivity combined with long time constant is achieved by coupling the capacitor/meter combination to the amplifier through a resistance equal to the meter resistance, in this case 1800 ohms. The meter now reads 360 mV full scale. This is the maximum on-scale voltage that can appear across an unknown resistor; a maximum power dissipation of about 1.3 mW will occur in a 100-ohm resistor. The maximum voltage that can be applied to the meter is 4.5 volts, producing less than 12-mW dissipation in the meter winding. This will not burn it out, even over long periods. In addition to these precautions, a switch is provided to shunt the meter whenever transients are expected. A small resistor limits the capacitor discharge current, protecting the capacitor and switch.

The 360-mV maximum output is large compared to the expected offset drift, and small enough to expect good amplifier linearity and low current consumption. Fig. 2 shows that a dual op amp is used, the first section being a voltage follower and the second section an inverting amplifier as shown in Fig. 1. The voltage follower supplies sizable currents to R_A from an output impedance of much less than an ohm, providing good voltage and gain regulation against range changes. Since the input impedance of this amplifier is virtually infinite, it can be supplied by a high-impedance source requiring little battery current. The net result is that E_i to the in-

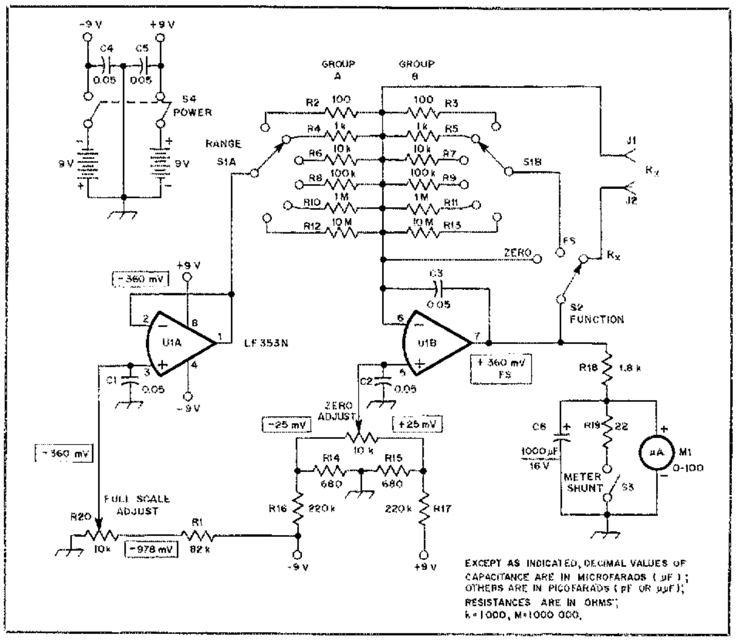


Fig. 2 — Schematic diagram of the Linear, Self-Calibrating Ohmmeter. Fixed-value resistors are 1/2 W composition. R2-R13, inclusive, are 5% tolerance or better. Capacitors are disc ceramic unless otherwise specified.

J1, J2 — Binding posts.
S1 — 2 pole 6 position nonshorting rotary.
S2 — 1 pole 3 position nonshorting rotary.
S3 — Spst toggle.
S4 — Dpst toggle.
U1 — Dual JFET op amp, LF353N (Radio Shack part no. 276-1715).

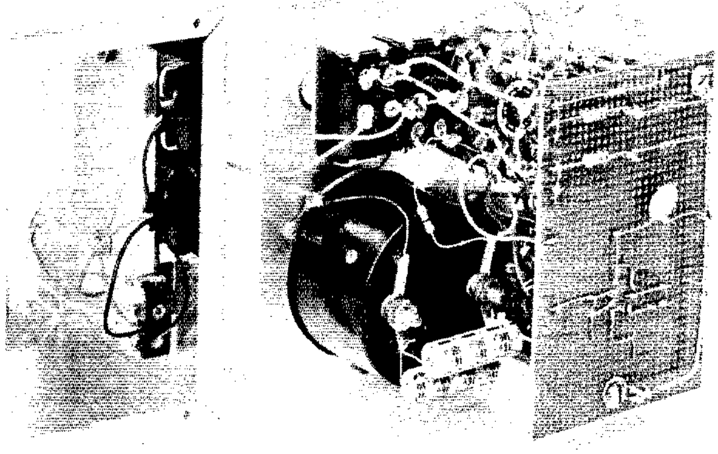


Fig. 3 — Construction details of the ohmmeter. A piece of perf board and point-to-point wiring is used for those components not connected directly to panel-mounted parts. Note that C6, wired directly across the meter terminals, consists of two 500-μF capacitors in parallel.

verting amplifier does not change materially with range switching, and the control is always linear.

The question of gain (R_B/R_A) now arises. Suppose we choose a gain of 1/10. This would require the follower to operate at 3.6 volts in and out, which would certainly reduce drift problems. But the maximum R_A value for a 10-megohm range would be 100 megohms, which is not commonly available and not likely to be stable in the presence of moisture and dirt. Suppose we consider a gain of 10. Then the follower would operate at 36 mV in and out, which is too small in relation to offset drift. So the vicinity of unity gain is indicated. The availability of inexpensive pairs of 5% resistors from Radio Shack influenced my final choice.

From Eq. 3 we see that E_o will be 0 V when $R_B = 0$, and $E_o = E_{in}$ when $R_B = R_A$. The ohmmeter in this article is constructed so the meter reads zero with R_B shorted, and full scale when $R_B = R_A$. An unknown with a resistance of R_B or less can be read on a linear scale with this meter.

Construction

I built my unit in an aluminum box measuring 5 × 6 × 4 inches (HWD).[2] The meter and all controls are mounted on the front panel for convenience when using the ohmmeter. The 9-V transistor-radio batteries are mounted on the side of the box by means of an aluminum strip. The op-amp IC and associated components are mounted on a piece of perf board, and the circuit is wired using point-to-point techniques. There is nothing critical about the circuit layout or the lead lengths. Fig. 3 shows the construction details.

Operation

The range switch, S1, is turned to the position that will put R_x on scale. The function switch, S2, is rotated to zero, and the ZERO control, R21, is adjusted to set the meter needle at zero. The function switch is then turned to the full scale position and the meter is deflected to full scale by adjusting the FS control, R20. Finally, the function switch is set to R_x, and the meter reads the unknown resistance directly.

In practice, the zero and full-scale positions do not vary more than a few percent with range changes. For routine work it is necessary to make these settings only occasionally. But for greatest accuracy, both settings should be adjusted immediately before the measurement is made.

Accuracy

Correct readings depend on the accuracy of the resistors in Group B and on the meter quality. Resistors with 1% tolerance may be justified if the meter is an excellent one. Close-tolerance resistors in Group A would also be a convenience because the full-scale deflection would not vary with range. However, precision resistors are expensive and not readily available.[3] In the rare event that extreme accuracy is required, a standard resistor may be connected to the R_x terminals and the meter set to full scale with R20. Now when the unknown is connected to the R_x terminals, the readings will be as accurate as the meter permits.

Notes

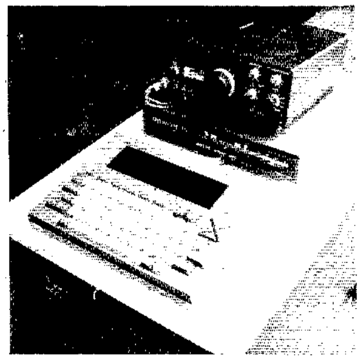
[1] H. Neben, "An Ohmmeter With a Linear Scale," *QST*, Nov. 1982, pp. 38-39.
[2] mm = in. × 25.4.
[3] [Editor's Note: 1/2-W, 1% metal-film resistors (10 Ω to 1 MΩ) are available from Mouser Electronics, 11433 Woodside Ave., Santee, CA 92071. The price is 13¢ each in single lots, with a $20 minimum order.]

Strays

SPANNING THE WORLD WITH MICRO AMTOR

☐ Pounding a key or talking into a mike may well be the traditional notions of Amateur Radio, but more and more amateurs have taken to Baudot radio teletype (RTTY). Still others have turned to AMTOR (*AM*ateur *T*eletype *O*ver *R*adio), an error-detecting system that produces a spectacular improvement in accuracy over ordinary RTTY.

This system is specified in ITU (CCIR) Recommendation 476, and it is widely used by maritime stations. Credit for the development of the amateur version must be given to Peter Martinez, G3PLX, who produced a workable microprocessor program and the AMT-1, which houses the μP unit and the terminal (see June 1981 *QST*, page 25).



AMTOR has found its way to Malaysia, where Colin Richards, 9M2CR, gets big results using little power with this station.

The final "micro" touch at 9M2CR (see photo) was achieved by using an NEC PC-8201 personal computer as a keyboard and display. Powered by four penlight batteries, it can house up to 64K of RAM and run BASIC. The standard RS-232-C port can be set with the required parameters for the AMT-1 by keying in a code, which stays in the memory even after shutoff. There are also ports for a printer, a cassette-recorder, a CRT and disk drives. And all in a 10 × 12 × 2-inch package. With the tiny TS-120V putting out 10-W maximum, the system spans the world. It's micro magic! — *Colin Richards, 9M2CR, Negri Sembilan, Malaysia*

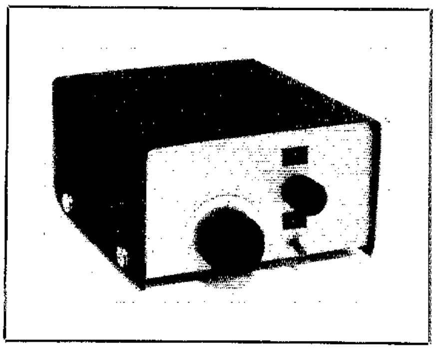
SATELLITE LINKS REPEATERS

☐ In what may be the first operation of its kind, Amateur Radio repeaters on the East and West Coasts were linked by satellite. On May 21, 1983, amateurs at the RCA Astro-Electronics Space Center at Hightstown, New Jersey, used an RCA Satcom satellite to establish a transcontinental link between their club's 450-MHz repeater, WB2JQR/R, and a 2-meter repeater, WA6OBT, in Thousand Oaks, California. During the seven-hour operation in commemoration of the 25th anniversary of the Space Center, several hundred contacts were made between hams in New Jersey and California, with more than 150 QSOs being logged by WB2JQR/R. — *Irv Seideman, KB2LG, RCA Astro-Electronics ARC, Princeton, New Jersey*

I would like to get in touch with...

☐ anyone who has the engineering drawings and installation plans for a Tristao HZ-354 tower. Marty Waite, WA6JDU, 11408 Buell St., Sante Fe Springs, CA 90670.

A Traveler's Receiver for 20 Meters



Simple, small and sensitive — three noteworthy features of this portable receiver design.

By Doug Blakeslee,* N1RM

Whether visiting the next state or halfway around the world, most hams wonder what band conditions sound like from a different location. My business provides travel to several rare spots on the globe, and so I decided to build a portable ham-band receiver to listen in from "the other side." The "Progressive Communications Receiver" circuit by Hayward and Lawson had the basic features I desired, but as in any design a series of compromises had to be made.[1]

Design Tradeoffs

For simplicity, I chose a single-band, battery powered design. Twenty meters was selected because it is the best all-around DX band. The antenna would be a quarter wavelength of hook-up wire, so the design required a "hot" front-end. Strong-signal performance was not a consideration — there would be few (strong) local signals where I travel.

A direct-conversion design was chosen as a tradeoff of size for performance. A block diagram of the receiver is shown in Fig. 1. The modules are the Hayward-Lawson designs. The VFO is modified to cover 13.9 to 14.5 MHz and a 20-meter preamplifier is added to increase the system gain. These circuit changes are shown in Fig. 2. A 6-pole ssb-bandwidth audio filter is also included; it provides adequate selectivity for casual listening on phone and cw. A pair of lightweight headphones designed for use with personal tape players are a delight to use because of their small size and their clean, sharp response.

Construction

I built the receiver using ready-made pc boards, except for the 20-meter preamplifier board, which I made.[2] Everything "played" well the first time. For those who build their own equipment, the VFO capacitor and associated dial are always difficult to locate. My junk box yielded a capacitor with a built-in reduction drive (of unknown origin) that is physically large. A dial pointer was fashioned by soldering a stiff wire to the outer brass shaft of the capacitor. Not very elegant, but inexpensive! A planar drive with a standard capacitor would work as well.

Packaging the receiver was a difficult

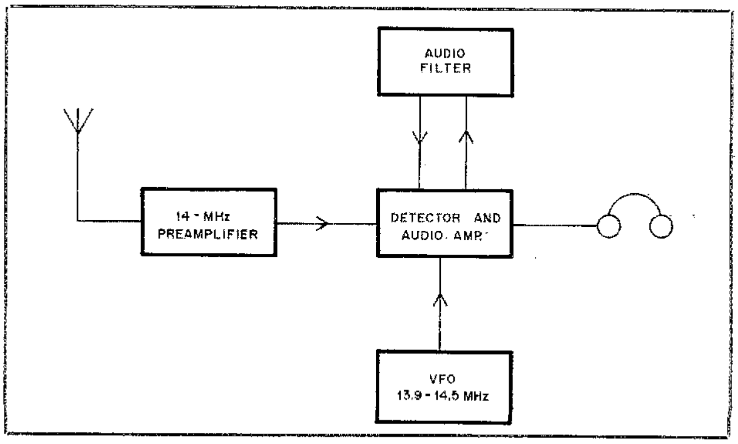


Fig. 1 — Block diagram of the portable receiver.

[1]Notes appear on page 32.

*4 Maple La., Brookfield, CT 06804

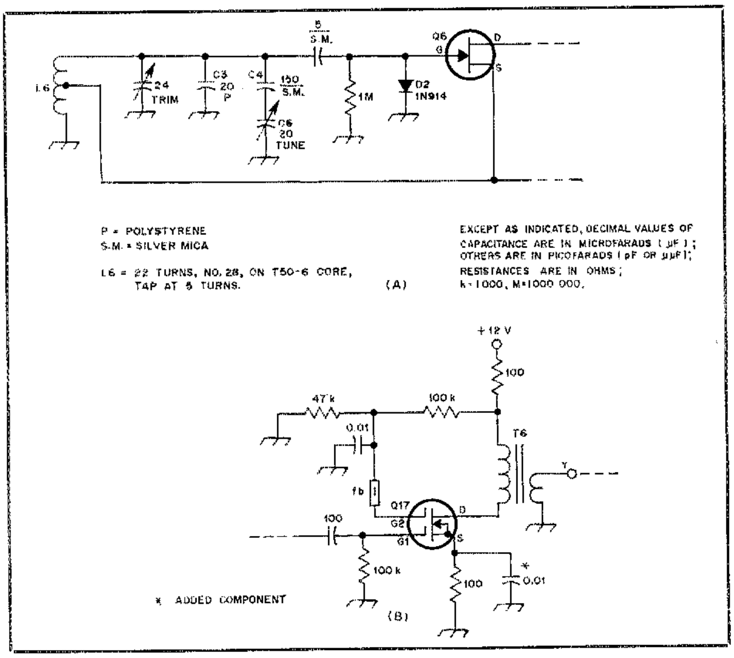


Fig. 2 — Component value and circuit changes to the VFO and rf amplifier sections of the Hayward-Lawson receiver are shown here. Component designations are those of the original circuits.

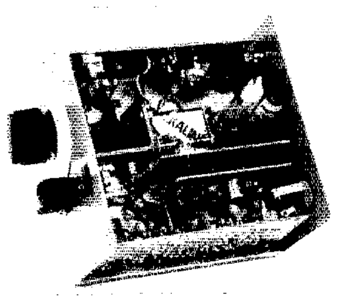


Fig. 3 — Internal view of the portable receiver.

problem. The receiver size had to be small. It had to appear simple — not similar to an aircraft control panel! The appearance had to be innocuous, as it would have to pass through Customs in a number of countries. So, the package ended up with a tuning knob, volume control and an on-off switch on the front panel. A miniature phone jack provides for antenna connection; it looks less conspicuous than a coaxial connector.

To make the smallest package, the circuit boards are mounted vertically, as shown in Fig. 3. I was concerned about coupling from the VFO into the preamplifier. So the VFO and preamplifier are placed on opposite sides of the chassis. The audio-filter board is used as a shield. Investigation with an hf oscilloscope indicated that there are no unwanted coupling problems. A Radio Shack 5-1/4 × 3 × 5-7/8 inch (270-253) cabinet houses the receiver;[3] a Bud Minibox would have been a better choice, because the front and back panels of the Radio Shack cabinet are unsupported and, thus, tend to be bent as my suitcase endures international travel.

Results

The receiver was finished only two days before I was to leave on a trip to Asia. Casual listening with an outside antenna proved that it was indeed "hot." Using a 1/4-wave end-fed antenna stretched across the floor, a good deal of DX was audible. I tried assorted shake and bounce tests, fearing that one bad solder joint would render the unit inoperative 9000 miles away from my soldering iron. These fears were also groundless — the unit worked perfectly for the entire trip. Many stations from around the world were heard. BY1PK was one of the loudest signals on the band — while I was listening from Singapore, of course! If only they were as loud in the northeastern U.S.!

Notes

[1]Hayward and Lawson, "A Progressive Communications Receiver," *QST*, Nov. 1981.
[2]Circuit Board Specialists, P.O. Box 969, Pueblo, CO 81002.
[3]mm = in. × 25.4.

Strays

STRAY HINTS

☐ "Strays" are those interesting fillers used when space allows in *QST*. Think you have an item with Stray potential? Here are some hints to help your submission become one. (1) Be sure the information will be of interest to most readers of *QST*. (2) Submit your material before deadline — the 8th of the second month preceding desired publication (i.e. arrive at Hq. before September 8 for November *QST*). (3) Any photographs you send should be good-quality, black-and-white glossy prints. Color prints, slides and instant photos do not usually reproduce well.

Items submitted are normally acknowledged, but that doesn't necessarily mean that your Stray will be appearing in *QST*. We receive far more material than we can find room for. If you want your material returned, please include a statement to that effect and an s.a.s.e.

Follow the above hints and maybe your Stray will find a home in *QST*. — *Andrew Tripp, KA1JGG*

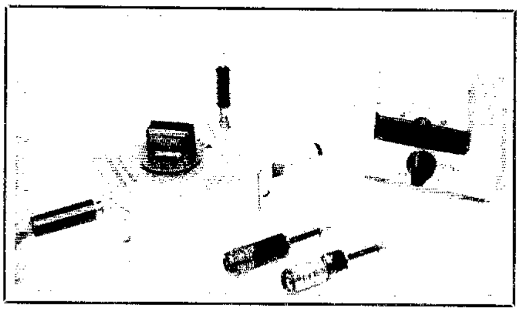
Last May, members of the Albuquerque (New Mexico) Amateur Radio Caravan Club set up a station at the 34th International Science and Engineering Fair in Albuquerque, where they passed messages through the National Traffic System for students participating in the fair. Among those manning the station were (l-r) KA5KYR, N5CFO, KC5QF and WB8VCE.

RCA ARC ADDRESS UPDATE

☐ Many amateurs who sent for the certificate and QSL card issued to commemorate the East Coast to West Coast linkage of 2-meter and 450-MHz repeaters by means of the RCA Satcom Communications Satellite on May 21, 1983 have had their requests returned undeliverable. This is because the *Callbook* address for the RCA-Astro-Electronics ARC is outdated, and the post office does not forward the mail. The club's correct address is P.O. 800, Princeton, NJ 08540. Certificates and QSL cards are being sent to all stations having a QSO entered in the club's log. — *Irv Seideman, KB2LG, RCA Astro-Electronics ARC*

The Ever-Useful Wavemeter

No amateur workshop is complete without a wavemeter. Learning a few useful tricks with these simple instruments can save time during construction-project exercises and antenna experiments.



By Doug DeMaw,* W1FB/8

You say that building homemade gear is beyond you because you lack suitable test equipment? If this has been your reason for not becoming involved in this satisfying and enjoyable side of Amateur Radio, then follow this *QST* series. Last month's Beginner's Bench column marked the start of a series on homebuilt test apparatus for the ham workshop. We have a number of simple, interesting test-equipment projects planned for the *QST* issues to come.

The wavemeter has been a basic Amateur Radio tool since the start of our communications pastime. In fact, it was perhaps the *primary* unit for frequency measurement during the early years of this hobby. Although digital frequency meters are accurate and reliable, we need not use them for coarse frequency checks. A home-constructed wavemeter will suffice for numerous routine measurements, and the cost for our own instrument will be a fraction of that for a commercial frequency counter. Furthermore, our wavemeter can be used as a relative field-strength meter or an "rf sniffer" for detecting rf-energy leaks. Still another advantage of the wavemeter is its passive nature — meaning that it requires no dc power supply to operate the circuit. This provides us with an instrument that can be carried afield for all manner of uses, without dependency on a power source.

This month, we shall learn some uses for wavemeters, how to apply them and how to build an inexpensive unit that covers the range from 1.5 to 38 MHz. An uncomplicated construction technique is described to make duplication a simple matter for even the least-experienced builder.

*ARRL Contributing Editor,
Box 250, Luther, MI 49656

What Is a Wavemeter?

A wavemeter is a tuned circuit with components of capacitance and inductance (a coil and a tuning capacitor). The tuned circuit must be capable of tuning the frequency range of interest. For example, if we want to check the relative field strength of our 80-meter antenna when adjustments are being made, our wavemeter must tune from 3.5 to at least 3.8 MHz. A short antenna is connected to the instrument so it can pick up enough radiated energy to provide a field-strength indication.

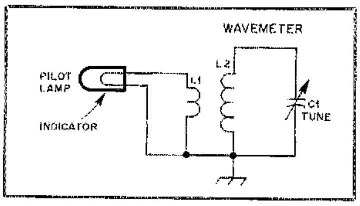


Fig. 1 — Circuit illustration of an early-day absorption wavemeter that used a lamp or neon bulb as a visual indicator.

But, we need something in addition to the basic tuned circuit. It is essential that we employ some form of visual indication to show us that we have the instrument tuned to the desired frequency. In the early days of radio, the indicator was often a small incandescent lamp or neon bulb that was coupled to the tuned circuit. If sufficient rf energy could be sampled, the lamp would glow when the wavemeter was tuned to the frequency of the energy source. Fig. 1 shows a circuit of this type. Greater sensitivity is possible by using the modern technique of rectifying the sampled energy, thereby converting it to dc. The dc is routed through a sensitive meter, and the response of the tuned circuit is indicated by an upward deflection of the needle in the meter. A circuit of this kind is depicted in Fig. 2. Considerable sensitivity is possible when a *microammeter* is used for the indicator.

Most wavemeters have a calibrated dial face that indicates the approximate frequency to which the L-C circuit is tuned. This permits us to take a direct analog reading from the dial when the peak response is obtained. Alternatively, a numbered dial can be used. If this is done, we can compile a list of frequencies versus

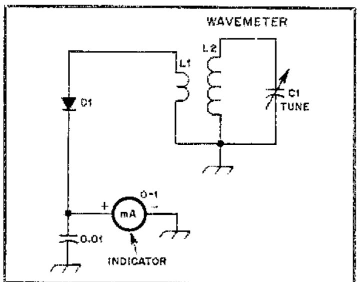


Fig. 2 — Simplified circuit of a modern wavemeter. A link is used to couple to the detector diode, thereby minimizing loading of the tuned circuit, L2/C1.

the dial numbers (a calibration chart) to permit us to be aware of the frequency to which the instrument is tuned.

A wavemeter can be used for field-strength observations when we experiment with antennas. Generally, a short whip antenna is plugged into the wavemeter to increase the pickup ability. Rf energy that reaches the sampling antenna (whip) is carried to the tuned circuit in the instrument, detected and displayed visually on the dc meter. For many years, this was the only indicator used by some amateurs for adjusting antenna matching sections, Transmatches and pruning the coils for resonance in mobile antennas; rf-power meters and SWR indicators were not common items in those days. A field-strength meter is useful for determining the approximate radiation pattern of an antenna — still another application we can take advantage of.

Some Other Uses for Wavemeters

Let's suppose we were working on a homemade transmitter and wanted to observe the operating frequency of one of the stages in the circuit. We may wonder if the tuned circuit of that stage is on the desired frequency, or if it is tuned to a harmonic of the desired frequency. We simply place our wavemeter probe (plug-in coil) in proximity to the questionable tuned circuit and adjust it for a peak reading. If all is well, our wavemeter will show the correct frequency. If not, our response will occur at some other frequency — indicating a problem with the coil or capacitor value in the transmitter.

Another trick we can play with our wavemeter is to observe the progress we are making in extracting the last acceptable vestiges of power from a particular tuned-circuit in our transmitter. We can place the wavemeter probe near the output tuned circuit of the stage in question, then tune or adjust the circuit values for maximum wavemeter indication. *Caution: Do not poke the wavemeter near any circuit that contains high operating voltages:* This could result in personal injury. There should be no need for concern, however, when working with transistorized circuits.

A wavemeter is useful in getting a general idea of the harmonic energy contained in the output of a transmitter or excitation stage. Tune the wavemeter to the second and third harmonics of the operating frequency. If a significant amount of harmonic current is present, the wavemeter will yield a pretty good indication. Compare the amount of meter deflection with that obtained at the fundamental frequency. This will provide a comparative analysis of the problem. Take corrective steps to reduce the harmonic levels, then recheck the circuit with your wavemeter until the harmonics do not deflect the indicating meter or just barely show up on the instrument.

Transmitter and receiver stages that are unstable can be detected quickly with a sensitive wavemeter. This calls for a spectrum search with our meter — tuning from medium frequencies through the upper end of the high-frequency region. If a wavemeter response is noted on some frequency that is unrelated to the operating frequency, chances are that we will know the self-oscillation frequency. Of course, such unwanted oscillations may be occurring in the vhf region or in the standard a-m broadcast band, or lower. Our wavemeter would have to cover all of those frequencies to make the search complete. Some hams build two or more wavemeters — one for low frequency, one for high frequency and perhaps another for the vhf range. It depends primarily on how complete you want your test-equipment capability to be.

VHF Wavemeters

The plug-in coil concept can be applied at vhf and uhf, just as it is for dip meters. But, most commercial vhf or uhf wavemeters are constructed in a more rugged manner. Tunable resonant cavities or strip-line resonators are the usual choice. The sampling point can be effected by means of link or probe coupling to the cavity, and output to a rectifier can be by the same means. A circuit example of a strip-line resonator for wavemeter use is given in Fig. 3. The tuned circuit, L3 and C1, are selected for the vhf range of interest.

Table 1

Wavemeter Applications

Transmitter Testing

1) Tuning indicator for maximizing stage output.
2) Testing for fundamental output or unwanted harmonic output of a transmitter stage.
3) Finding the frequency of a transmitter stage.
4) Detecting an unwanted self-oscillation and determining its frequency.
5) Comparing the relative level of harmonics against that of the desired transmitter frequency.
6) Determining whether an oscillator is functioning.
7) Comparing the relative output power of transmitter stages to ensure that each low-level stage is providing gain.

Receiver Testing

1) Establishing that oscillator circuits are operating.
2) Investigating stages for self-oscillations.
3) Coarse frequency measurements of oscillators.

Rf Sniffing

1) Checking for ground loops (rf) on pc boards and chassis.
2) Sampling cabinet openings for rf leaks (meters, louvers, etc.).
3) Checking for rf leakage on line cords, microphone cables and control lines.

Antennas

1) Making relative field-strength measurements.
2) Adjusting antennas for maximum gain.
3) Observing approximate radiation patterns.

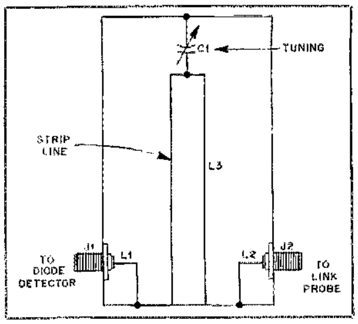


Fig. 3 — Hybrid diagram of a vhf or uhf style of wavemeter. A strip-line tuned circuit is indicated. L3 and the inner surface of the box form a transmission-line section.

For example, we might use a 25-pF variable at C1 to work with a strip line (L3) that is, say, 15 inches (381 mm) long. This should give us coverage from roughly 100 to 160 MHz. L1 and L2 are adjusted in size and proximity to L3 in accordance with the degree of coupling we desire. Too much coupling will lower the tuned-circuit Q. Too little coupling will cause the tuned circuit to present an insertion loss. The tuned circuit is contained in a metal box made from copper, aluminum or silver-plated brass. The impedance of the strip line and box, considering them as a section of transmission line, can be on the order of 75 ohms. J1 is connected to a diode detector and indicating meter, similar to that shown in Fig. 2. J2 is mated with a suitable length (short) of 50-ohm coaxial cable that has a one-turn small loop at the far end. The loop is used to probe the circuit of interest. A short whip antenna can be attached at J2 for field-strength use when adjusting antennas.

Table 1 provides a digest of the many uses to which we may put a homemade wavemeter. Certainly there are additional applications that did not come to mind as this article is being written. Perhaps you will think of them in the course of your circuit testing.

A Practical Wavemeter

Let's dispense with the whys and wherefores of wavemeters and consider constructing one of these instruments. I'm sure each of us enjoys short-term, simple projects that can be built during an evening. This instrument fits that description. The design and assembly of this wavemeter took six hours. It may require more or less time to complete the task, depending on the variety of hand tools you have in your shop and how much skill you may possess in assembling circuits. At any rate, you should be able to complete your work in a weekend.

The wavemeter is built in two pieces. It

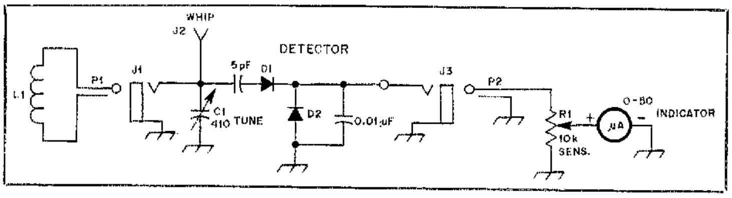


Fig. 4 — Schematic diagram of the wavemeter described in the construction section of this article. D1 and D2 are small-signal diodes. See text for a discussion of the components in this circuit. R1 is a linear-taper composition control.

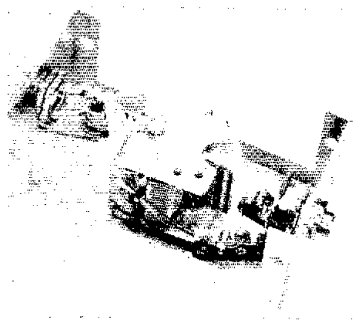


Fig. 6 — Interior view of the wavemeter tuning head. Large-diameter bus wire is used to join C1 to J1. This will help to reduce unwanted stray inductance.

Table 2

Wavemeter Calibration

Coil Dial-Skirt Numbers Versus Frequency (MHz)

Coil	0	1	2	3	4	5	6	7	8	9	10
A	38.0	27.0	22.0	18.0	16.0	14.5	13.5	12.5	11.75	11.0	10.5
B	12.0	8.9	7.0	5.9	5.25	4.7	4.4	4.0	3.8	3.6	3.4
C	5.4	4.0	3.2	2.8	2.4	2.2	1.92	1.79	1.67	1.57	1.5

Coil information: Coil A contains 7 turns no. 20 enam. wire (0.7 µH) space wound to fill 0.5-inch length. Coil B has 30 turns no. 22 enam. wire (6 µH) close wound. Coil C contains 65 turns no. 28 enam. wire (27.5 µH) close wound. Note: Set C1 at minimum mesh with the tuning dial at 0.

could be assembled as a one-piece unit, but that would make it unwieldy for use with compact circuits. The two-piece format makes the unit more convenient to use, especially when it is necessary to probe into tight places during a test. The tuning head is built around a 4 × 2 × 1-1/2 inch (in. × 25.4 = mm) Minibox. A homemade enclosure can be fashioned from pieces of pc board if that is your pleasure. A piece of aluminum sheet (3-1/2 × 3-3/4 inches) is bent to provide a sloping panel. This serves as the base for the microampere meter (2-1/4 × 2-3/4 inches) and the sensitivity control. A fancier indicator unit would result if we were to use a sloping-panel meter case, but the cost would be greater.

Fig. 4 shows the circuit of the wavemeter. Only three plug-in coils are required for coverage from 1.5 to 38 MHz. P1 is the base of the plug-in coil, and J1 is the jack into which it is inserted. C1 is a single-section broadcast-radio tuning capacitor.[1] It has a capacitance range of 15 to 410 pF, as measured in the ARRL lab. A 365-pF variable capacitor can be used, but this will result in a sacrifice of the tuning capability at the low end of each plug-in coil range.

Light coupling from the high end of the tuned circuit to the rectifier is ensured by our use of a 5-pF capacitor. This permits ample rf energy to reach the rectifier without lowering the Q of the tuned circuit by loading it with the low impedance presented by D1 and D2. Too low a loaded Q would cause broad tuning and reduced sensitivity.

[1]Notes appear on page 36.

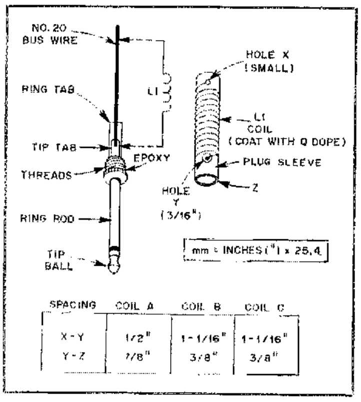


Fig. 5 — Assembly and modification details of the PL-55 style of phone plug used for plug-in coils A, B and C. Do not exert undue pressure on the plug insulating sleeve when drilling holes X and Y, lest the plastic cracks. Drill a no. 60 pilot hole at each point before applying a larger drill bit. See text for additional data.

A voltage doubler is used for the rectifier. This results in somewhat greater meter deflection than would be realized with a single diode in a half-wave rectifier. Germanium diodes, such as the 1N34A type, are recommended for best sensitivity at low signal levels. I used silicon small-signal diodes (1N914) with good results.

P2 plugs into J3, thereby connecting our meter and sensitivity control to the tuning head. J2 is a pin jack that accepts a short whip antenna (18 inches for my unit) made from brass brazing rod. The lower end of the rod is soldered into a pin plug. A small loop is formed at the top of the whip to prevent eye damage through accidental contact with it.

Construction

In an effort to provide a small plug-in coil of good mechanical strength, I decided to try a PL-55 type of phone jack. This seemed like an inexpensive solution to an otherwise perplexing problem. The notion was a good one, for the coils are easy to wind on the insulating sleeve of the jacks, and the resultant unloaded Q is acceptable. The Qs for coils A, B and C are 130, 120 and 90, respectively.

Fig. 5 shows the details for preparing the plug-in coils. First, remove the sleeve from the plug and note where the solder tab for the ring connection is. Place a mark on the ring rod so that it's aligned with the center of the ring tab. Next, screw on the insulating sleeve until the threads are snug. Now, make a mark on the sleeve so that it is lined up with the mark on the ring rod of the plug. This will indicate the line along which you will drill hole "Y." The line can be extended up the sleeve to show where hole "X" must be drilled.

Use a 3/16-inch drill bit when making hole "Y." This will allow sufficient access for soldering the lower coil lead to the ring tab. A length of no. 20 bus wire is soldered to the tip tab and extended through the upper opening in the insulating sleeve of the phone plug. The upper coil lead will be soldered to this wire, flush with the top of the sleeve. Poke the solder joint into the sleeve slightly so that it is not exposed. An interior view of the tuning head is shown in Fig. 6.

Coil A has the ring tab at its full length. The tab is snipped off (see chart in Fig. 5) adjacent to hole "Y" for coils B and C. This prevents the ring tab from protruding into the coil and degrading the Q. A drop of epoxy cement can be placed on the plug threads just before you screw the sleeve into place for the last time. The coil is wound after the sleeve is in place. Start with the lower end, and solder it to the ring tab. A coating or two of coil cement or Q Dope can be applied to the completed coil windings to protect them and to hold the turns in place. Shrink tubing might offer an even better solution to this problem. I did not try it.

Calibration

If the coil dimensions in this article are followed closely, and if the variable capacitor specified for C1 is used (or one of the same capacitance range), a 0-10 dial face will yield the calibration specified in Table 2. If you depart from the circuit and component values suggested here, calibration can be effected by using a dip meter to plot the megahertz points on your tuning dial. The dipper signal can be checked for accuracy by listening to it with a calibrated general-coverage receiver.

The readout resolution can be made better if a smaller amount of capacitance is used at C1 — such as 100-pF variable capacitor. If this is done, it will be necessary to wind more coils in order to cover the desired tuning range. My intention was to keep this circuit as simple and inexpensive as possible, hence only three coils.[2]

Summary Comments

Care needs to be exercised when sampling rf near a high-power part of a circuit. Too much rf energy will destroy the detector diodes. Fortunately, diodes are inexpensive and easy to replace, so there's no need to be deeply concerned about this potential problem. When in doubt, start your tests with the wavemeter a considerable distance from the circuit being tested. Move it close enough to get a full-scale meter reading with the sensitivity control set for maximum meter response. That procedure will ensure you of safety to the diodes.

If you constructed last month's project, and will build this one, you'll find yourself well down the road toward having some useful gadgets in your workshop for designing, building and testing homemade amateur equipment. Perhaps now is the time to get involved!

A parting comment: Almost any microammeter will serve in this project. Many of the surplus edgewise fm tuning meters can be used. See note 1 for a supplier of low-cost meters of that variety.

Notes

[1]400-pF variable capacitors and low-cost fm tuning meters are available from Surplus Electronics Corp., 7294 N.W. 54th St., Miami, FL 33166, tel. 305-887-8228. Catalog available.

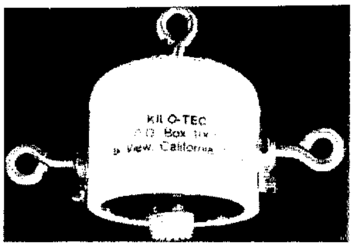
[2]A kit of parts for this project is available from Circuit Board Specialists, P.O. Box 969, Pueblo, CO 81002.

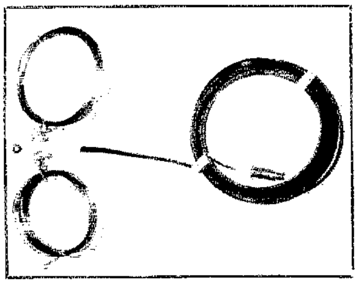
New Products

KILO-TEC ANTENNA AND DIPOLE CENTER CONNECTOR

☐ The KT5B is a multiband antenna designed for use on the 1.8-30 MHz Amateur Radio bands. It uses no loading coils or traps and the manufacturer claims it will handle 2 kW PEP. Price: $59.95.

The DP-1 antenna center connector is supplied with an SO-239 fitting and can be used with dipoles, quads or V antennas where a weather-resistant coaxial connection is needed. The manufacturer rates the unit at a full 2 kW and states it features a Mil-type potting material to prevent rf flashover. Price: $8.95. For further information on these products, contact Kilo-Tec, P.O. Box 1001, Oak View, CA 93022 or call 805-646-9645. — *Paul K. Pagel, N1FB*





DX PREDICTOR

☐ A *QST* article describing MINIMUF, an hf propagation prediction program, raised considerable interest within the amateur community.[1] A program currently being marketed by Flynn's Business Services, P.O. Box 903, Mountain View, CA 94043, embodies not only the NOSC (Naval Ocean Systems Center) MINIMUF program but also algorithms that predict fot (frequency of optimum transmission) and luf (lowest usable frequency). This package relies upon NOSC algorithms and has a user-friendly input routine and a graphic display. The program was written to provide amateurs with a ready-to-go program that can be of use as an operating and, possibly, an educational tool.

The DX Predictor is priced at $40 and includes a 5¼-inch diskette (or tape) and documentation. It is available for the Apple II® (48K), Commodore PET® and Commodore VIC 20® (16K RAM card required) personal computers. The program is also compatible with the Apple IIe.® — *Paul K. Pagel, N1FB*

[1]Rose, R., "MINIMUF: A Simplified MUF-Prediction Program for Microcomputers," *QST*, December 1982, p. 36.

ICM CRYSTAL CATALOG

☐ A new, 32-page crossover list and custom-crystal catalog is available from International Crystal Mfg. Co., Inc. The crossover list provides the user with the ICM crystal catalog number, referenced to the manufacturer's equipment number for the most popular types of commercial two-way equipment. The custom crystal list features charts that establish the ICM catalog number and available holder dimensions for their crystals.

If you're a builder or you like to modify commercial equipment, you'll probably find the catalog to be a valuable addition to your stash of reference material. To obtain your free copy, write to International Crystal Mfg. Co., Inc., Dept. A, P.O. Box 26330, Oklahoma City, OK 73126. — *Paul K. Pagel, N1FB*

Hints and Kinks

Conducted By Larry D. Wolfgang,* WA3VIL

CONVERTING A HUSTLER MOBILE ANTENNA FOR 30-METER OPERATION

☐ A Hustler RM-40 resonator can be trimmed easily for operation on the 30-meter band by removing a few turns from the coil. There are two versions of the 40-meter resonators made by Hustler. One has a coil diameter of about 1 inch, and the other has a diameter of about 1-1/2 inches.[1]

I moved the smaller coil higher in frequency by removing turns from each end, leaving 55 turns in the middle of the form. With the larger coil, I had to leave 30 turns on the form. These numbers were the end result of my experimentation. I would suggest that you leave a few more turns on the coil at first. Check the resonant frequency of your system, and remove turns gradually until the antenna resonates at 10.125 MHz with the adjustable whip extended half its length.

My SWR measurements with both coils were less than 1.4 at resonance. On-the-air results have been good. — *Bob Douglas, W5GEL, Corpus Christi, Texas*

A TRIPLE-MODE POWER SUPPLY

☐ A low-voltage utility power supply can serve triple duty if a double-pole, triple-throw switch is included in the circuit, as shown in Fig. 1. You may choose to build a separate supply or add a switch to one you already have. This supply provides a switch-selectable output: either as a full-wave center-tapped circuit, a bridge-rectifier circuit or as a full-wave voltage-doubler circuit.

When the mode switch, S1, is in the 1/2 position, D1 and D3 are effectively out of the circuit. The center tap of the transformer is connected to ground, and the supply is in a conventional full-wave, center-tapped configuration. The loaded output voltage is nominally about half the total secondary rms voltage. To configure the supply as a full-wave bridge circuit, move the switch to the 1 position. The ground connection is transferred to the junction of D1 and D3, and the loaded output voltage is approximately equal to the transformer rms voltage. With S1 in the 2 position, D1 and D2 form a full-wave voltage-doubler circuit. D3 and D4 are reverse biased in series across the output and have no effect on circuit operation. Nominal loaded output voltage is twice the rms voltage of the transformer.

The voltage and current ratings for the components in your supply must be chosen for the worst case of the three modes. Diode and filter-capacitor voltage ratings are determined by the voltage-doubling mode. The diodes must have a PIV rating of at least 2.83 × V_{rms} plus some safety factor. A rating of about 4 × V_{rms} would be about right for most applications. C1 and C2 each have a maximum voltage of 1.41 × V_{rms} applied to them, so they should be rated at about 2 × V_{rms}. Power ratings for the equalizer/bleeder resistors, R1 and R2, are also determined in the doubling mode. This rating can be found by

$$P = \frac{E_o^2}{R} \qquad \text{(Eq. 1)}$$

where

P = the minimum power rating for the bleeder resistors

E_o = the nominal loaded output voltage of the supply in the voltage-doubling mode

R = the total resistance of the bleeder resistors, R1 + R2

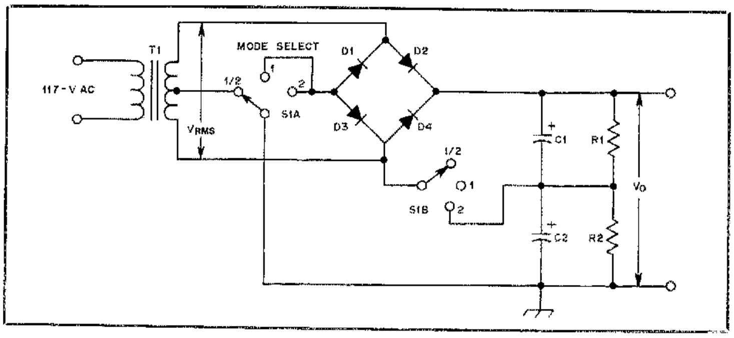


Fig. 1 — Schematic diagram of a triple-mode power supply. Component values and ratings are discussed in the text.

The filter capacitance and the diode current ratings are determined by the maximum current to be taken from the supply in the full-wave, center-tapped configuration.

This design offers a degree of flexibility for an unregulated power supply, and it can also provide increased operating efficiency with a supply that uses an adjustable three-terminal regulator. You can select the appropriate mode to reduce the voltage drop in the regulator for different output voltages. This will permit the regulator to handle more current at low voltages, while remaining within power dissipation limits. — *R. B. Gibson, KE5E, Los Alamos, New Mexico*

BEWARE THE GROUND CONNECTION!

☐ Many mobile-radio manufacturers recommend that you make all power connections as close to the battery as possible. This will reduce noise pickup and other problems, but it is imperative that *both* the positive and negative leads be fused.

Most cars are wired so that a short, heavy negative battery lead goes directly to the engine block, on or near the starter motor. Another smaller-diameter lead completes the circuit to the car chassis, which is used as the power return (or negative) connection for the car accessories. Since the starter motor is the biggest current hog (taking several hundred amperes while cranking the engine), it is essential that the voltage drop on the cables to the battery be minimal.

On my car, the bolt attaching the negative lead to the engine block loosened. One morning when I tried to start the car, it cranked slowly and erratically. I traced the problem to the loose bolt. I tightened it, and the car started right up. But when I turned on my Clegg FM-76 220-MHz mobile rig, it did not seem to work right. The output power was down, there was ignition noise on the receiver and the squelch did not operate properly.

Mike Santana, WB6TEB, helped me troubleshoot the radio. We found four burned-out ground traces on the printed-circuit board, with charred debris around them. Amazingly, bridging these spots with short pieces of wire was all that was needed to restore the radio to full performance. But what had caused this near-total disaster?

After some thought, I realized that the radio was grounded through the chassis and my antenna; this had provided the ground return path for the starter motor when the bolt loosened. The power cord for my rig had a fuse only in the positive lead, even though the manufacturer recommended connecting both leads to the battery. With a fuse in both leads, if this problem ever recurs, the starter-motor current will blow the negative-lead fuse before any damage can be done to my radio.

Fortunately, some of the more recent radios being produced have fuses in both leads. A single fuse would be safe if you connect the negative lead to the car chassis, but if you want to minimize noise pickup by going directly to the battery, make sure you have fuses in both leads! [The fuses should also be as close to the battery as possible. — Ed.] — *Merv MacMedan, N6NO, Arcadia, California* (adapted from an article of the same title in *W6VIO CALLING*, Dec. 1982, the Jet Propulsion Laboratory ARC newsletter)

COAXIAL CABLES FOR MINIATURE PROJECTS

☐ In the construction of small projects it is often desirable to use miniature coaxial cables. The most popular type is RG-174/U. The major drawback to this cable is that the insulation is made of polyethylene. This material will melt with the heat of soldering the cable, making it difficult to work with. One good thing about it is the price, about 15 or 20 cents per foot in small quantities.

[1]mm = in. × 25.4

*Assistant Technical Editor

I find it far better to use Teflon-insulated cable for these projects. This type of cable is expensive, costing as much as a dollar or two per foot, but it is much easier to work with. It is no problem to do a nice looking job of wiring a project with this cable. The price is acceptable when you realize that only a foot or two will be needed for most projects. RG-178/BU is one type of Teflon cable.[1] It features silver-coated copper braid and a stranded center conductor.

Try this method for making a neat wiring job on your project: Cut all the way around the outer insulation with a sharp knife. Start about 1/2 inch from the end and be careful not to nick the braid. Tin the exposed braid. When this is done properly, the braid weave should be visible. If you apply too much solder, heat it again and flip the excess off. Be careful if you do this, and wear safety glasses so you don't flip hot solder in your eyes. Now use a sharp knife to score the braid all the way around about 1/4 inch from the end. Bend the braid back and forth at the score mark with your fingers until the piece breaks off. Trim a piece of the Teflon insulation away from the center conductor using your knife. Be careful not to nick the wire. You may need a short length of wire to connect the outer braid to your circuit. Wrap a few turns of bare hook-up wire around the braid and solder it. Fig. 2 illustrates this technique. — *Robert Shriner, WA0UZO, Pueblo, Colorado*

Fig. 2 — You can make a neat connection to the shield braid on a piece of Teflon-dielectric miniature coaxial cable by wrapping a few turns of wire around the braid and soldering it.

A FLEXIBLE MOBILE-ANTENNA MOUNT

☐ I have found an easy way to build a completely flexible and inexpensive mount for my 1/4-λ 2-meter antenna. Overhead obstructions, such as garage doors and low tree branches, no longer damage my antenna.

To build this mount, you will need to cut two 1-1/2 × 2-3/4 inch pieces of galvanized sheet metal. These must be bent into a right-angle bracket, one side being 1-1/4 inches long and the other 1-1/2 inches. Drill a 5/8-inch hole in the 1-1/2 inch side of one bracket to accommodate an SO-239 connector. Cut a small piece of sidewall from a discarded non-steel-belted tire. The piece of tire should be 1-1/2 × 3-1/4 inches. Fig. 3 gives the assembly details. You will have to drill the required bolt holes in the brackets and tire sidewall. I drilled matching holes in the lower bracket and my car roof for mounting the antenna. You may want to find an alternate means of attaching the bracket to your car.

My antenna consists of a 17-inch-long piece of 1/16-inch brass brazing rod. I soldered this into the SO-239 connector on my antenna mount, and attached a short piece of braided wire between the two brackets to ensure a good ground connection. A little extra length is needed

[1]RG-178/BU Teflon cable is available from Circuit Board Specialists, P.O. Box 969, Pueblo, CO 81002.

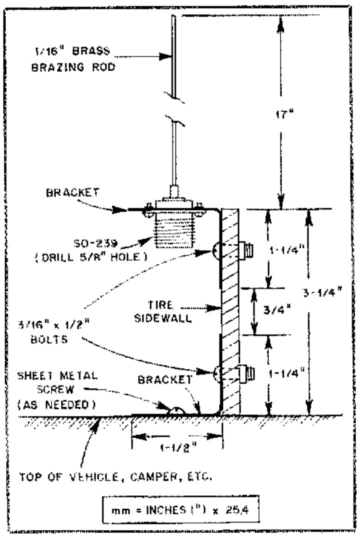


Fig. 3 — Construction details of a flexible mount built by N5DHN for his 2-meter mobile antenna.

for flexibility. Finally, I placed an SWR meter in the antenna feed line and trimmed the antenna to resonance. — *P. K. Hurlbut, N5DHN, Midland, Texas*

REPEATER INPUT FREQUENCIES WITH THE AZDEN PCS-300

☐ Owners of an AZDEN PCS-300 2-meter hand-held transceiver can listen to the input or output frequency of a repeater at the touch of a few keys. First, lock the keyboard while receiving on the repeater output frequency. Now, press any number key in the top row, and simultaneously press the corresponding key (in the same column) in the bottom row. You are now receiving on the repeater input frequency. Releasing either key puts the receiver back on the repeater output frequency. — *Dick Shongut, W2QFR, New Rochelle, New York*

OLD TIMER'S NOTEBOOK

Making Inductance Clips

☐ It is hard to do a neat job of making small clips to go on tubing, etc., with a pair of pliers. This kink may help.

A strip of phosphor bronze or other suitable material of the proper width is bent in the middle to form the two sides of the clip. A nail or rod of approximately the same diameter as the tubing on which the clip is to work is then inserted between the blades at the point where the bends should occur. The assembly is then put in a vise between two small blocks of very hard wood (Fig. 4), the nail carefully set at right angles to the length of the material, and the vise screwed home. Fast and neat — and the two halves come opposite each other! The outer lips are then bent back slightly with pliers. A small bolt may be put through the clip about mid-way along the length if necessary. — *K. B. Warner, W1EH (reprinted from* Hints and Kinks for the Radio Amateur, *2nd ed. [Newington: ARRL, 1937], p. 6)*

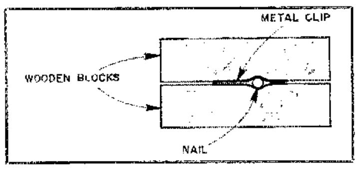


Fig. 4 — Spring clips for transmitting inductances can be formed readily with the help of a pair of wood blocks and a vise.

Tapping Transmitter Coils

☐ A satisfactory method of making taps on heavy wire coils is shown in Fig. 5. A short piece of bare wire, bent as shown, is soldered directly to the turn to be tapped. In soldering, both the coil wire and the tap should be tinned separately and then sweated together without additional solder. This method helps prevent shorting the adjacent turns.

The insulated "clip," also shown in Fig. 5, consists of a jaw taken from a miniature knife switch, the handle being the top of an old binding post. A machine screw of suitable length is first run firmly into the binding-post top, then the head of the screw is clipped off and the assembly fastened to the switch jaw with a small nut. The connecting wire is fastened between the jaw and handle. The resultant clip is positive, easy to handle, has plenty of contact surface and is neat in appearance. *(reprinted from* Hints and Kinks for the Radio Amateur, *2nd ed. [Newington: ARRL, 1937], pp. 6-7)*

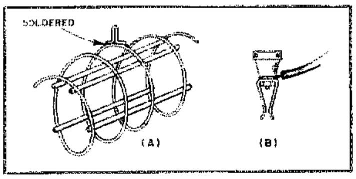


Fig. 5 — Details of a method of tapping a coil are shown at A. A suitable clip for attaching to this tap is shown at B.

Coil-Tapping Aid

☐ Finding the correct point for the tap on a small air-wound coil can be greatly simplified by using a paper clip that has been equipped with a flexible wire lead. The smaller paper clips can be easily slipped over turns that usually cannot be gotten at with regular commercial clips. — *John J. Schultz, W2EEY (reprinted from* Hints and Kinks for the Radio Amateur, *5th ed. [Newington: ARRL, 1954], p. 8)*

Tapping Miniature Coils

☐ It is always difficult to make a movable tap arrangement for small coils, where the turns are so close together that almost any of the usual clip arrangements merely short out several adjacent turns.

This problem can be solved easily by using a shortened bobby pin. Cut off all but the last "wiggle," clean the enamel off, and solder a flexible lead to the top. This gadget makes a swell movable tap that will fit between the turns of any of the small coils, and will make a good contact without shorting turns. — *Don Geary, VE3BTS (reprinted from* Hints and Kinks for the Radio Amateur, *4th ed. [Newington: ARRL, 1949], p. 7)* QST

Product Review

Conducted By Paul K. Pagel,* N1FB

Icom IC-740 HF Transceiver

Amateur Radio equipment has changed drastically in the past decade! Back in 1973, equipment designers were toiling over how to improve such things as oscillator stability and image rejection. Also, their efforts included trying to produce high-performance, low-cost transceivers for the limited-size amateur market. In the late '70s, this market mushroomed, largely because of the influx of disheartened CB operators. This increase in market size allowed manufacturers to develop quality equipment for sale at a reasonable price. Also, during this period, the state of the art in communications electronics equipment improved significantly, but there were still deficiencies, specifically in receiver performance.

In much of the Amateur Radio literature there was talk of improving receiver dynamic range.[1] The ability of a receiver to tolerate large signal levels is important to the consumer. Manufacturers responded to this market demand and, by 1980, were producing equipment with excellent dynamic-range figures.

The challenge of the '80s will be to produce a fully synthesized hf transceiver that is low in cost and has a low-noise frequency synthesizer. Much of today's amateur equipment uses digital frequency synthesis; the advent of LSI synthesizer and microcontroller (dedicated computer) ICs greatly simplifies circuit designs. But the new problems associated with synthesizer circuitry are by no means simple.[2] Digital frequency synthesizers inherently generate noise that is superimposed on the output signal; this noise can mask weak signals present in a receiver passband or cause QRM to nearby stations while transmitting.

The IC-740 features most of the "bells and whistles" found on competitive products, small size and all-band coverage. The most important feature, in my opinion, is the high-performance, low-noise receiver section.

Basic Features

Receiver

Frequency coverage of the transceiver includes all the amateur bands (WARC bands, too!) from 1.8 through 30 MHz. Passband tuning (PBT) and i-f shift (IF) functions are adjusted by front-panel controls, although they are not available independently. The agc decay time is set by a potentiometer rather than a switch; this should satisfy those who are not happy with a "two-speed" control. Two noise-blanker switches control the blanking level and the pulse width — narrow for ignition-type noise and wide for "woodpecker-type" QRM. Other unique receiver functions include an rf notch filter, an audio squelch control that works in all modes, an audio tone control and a switchable 10-dB gain broadband preamplifier. The remaining controls and jacks are listed in Table 1.

Transmitter

Emission type is selected by the MODE switch, which can select between NORmal or REVerse sideband (lsb on 1.8 through 10 MHz, usb on 14 through 28 MHz), cw, fm (with the optional fm unit installed) or RTTY. True fsk is used during RTTY operation; the user must supply a pair of "dry contacts" to key the radio. On ssb, an rf-type speech compressor circuit can be activated by the COMP button. A new feature cropping up these days is an rf-power control that works in all modes; the '740 includes this

ICOM IC-740 HF Transceiver, Serial No. 01721

Manufacturer's Claimed Specifications	*Measured in ARRL Lab*
Frequency coverage: 1.8-2.0, 3.5-4.0, 7-7.3, 10.0-10.5, 14.0-14.35, 18.0-18.5 (receive only), 21.0-21.45, 24.5-25.0 (receive only), 28.0-29.7 MHz.	As specified.
Modes of operation: Ssb, cw, RTTY and fm (with an optional fm unit installed).	As specified; the fm unit was not supplied with the review unit.
Readout: 6-digit, 100-Hz resolution.	1/2-in. high, 6-digit fluorescent-blue display.
kHz/turn of knob: Not specified.	100/10/1.
Frequency resolution: 100 Hz.	As specified.
Backlash: Not specified.	Nil.
RIT range: Not specified.	± 800 Hz.
S-meter sensitivity: Not specified.	W/o preamp (band, μV) — 80 m, 9.6; 40 m, 15; 20 m, 16; 15 m, 13.5; 10 m, 11.
Transmitter rf power input: 200 W cw; 200-W PEP, ssb; adjustable.	Greater than 100-W output, except on 160 m (90 W), 80 m (95 W) and 40 m (95 W).
Harmonic suppression: Better than 50 dB.	− 57 dB (see photo).
Third-order IMD: Not specified.	− 30 dB (see photo).
Spurious suppression: Better than 50 dB.	− 63 dB (see photo).
Receiver sensitivity: Less than 0.15 μV for 10 dB S + N/N with preamp on; 0.3 μV for 10 dB S + N/N with preamp off.	Receiver dynamics measured with optional FL-45 500-Hz cw filter:

	Preamp Off	*Preamp On*
Noise floor (MDS) dBm:	− 133	− 141
Blocking DR (dB):	130	125
Two-tone, 3rd-order IMD DR, worst case (dB):	95	94
Third-order intercept:	+ 9.5	− 0.5

Color: Black and gray.
Size (HWD): 4.2 × 11.7 × 14. in. (107 × 297 × 356 mm).
Weight: 17.6 lb (8 kg).

[1] W. Hayward, "More Thoughts on Receiver Performance Specification," Technical Correspondence, *QST*, Nov. 1979, p. 48, and Burwasser, "Reducing Intermodulation Distortion in High-Frequency Receivers," *Ham Radio*, March 1977.

[2] A. Helfrick, "The Universal Synthesizer," *QST*, Sept. 1981.

*Assistant Technical Editor

Table 1

Additional Front- and Rear-Panel Controls and Connections

Front Panel	Rear Panel
METER switch	Accessory socket (Molex)
VOX switch	Power socket (Molex)
VOX DELAY control	Antenna connector (PL-259)
VOX GAIN/KEYER speed control	Memory backup terminal
POWER switch	T-R control jack
T-R switch (MOX)	External alc jack
+0.5-MHz switch (for 10-meter operation)	Transverter jack (T-R)
AF GAIN control	Receiver input jack
RF GAIN control	Receiver output jack
MIC GAIN control	RTTY keying jack
PHONES jack (1/4-inch phone)	Key jack (1/4-inch phone)
MIC jack (multi-pin)	External speaker jack (1/8-inch phone)
Incremental tuning control	
RIT and XIT switches	

Jacks are phono types unless otherwise specified.

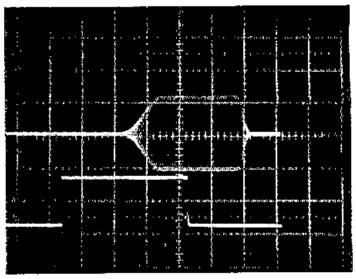


Fig. 3 — Cw keying waveform of the IC-740. Upper trace is the actual key closure; lower trace is the rf envelope. Each horizontal division is 5 ms. This keyed wave has a rapid fall time, which *may* cause key clicks.

feature. Another front-panel control sets the speed of a built-in electronic keyer (which is an option).

Several other functions are controlled by knobs mounted flush with the top of the cabinet. These include a MONITOR switch, which lets the user hear the transmitted signal; a MARKER switch and calibration control for an optional crystal calibrator; a FREQUENCY SET control that allows minor adjustment of the dial calibration; and the ANTI-VOX potentiometer.

Frequency Control

In keeping with the ICOM tradition, this transceiver incorporates a chopper-type optical counter dial and three speed-selector push buttons to select an operating frequency. Each pulse from the counter dial will step the frequency synthesizer 10 Hz, 100 Hz or 1 kHz, depending on the selected speed. An additional button provides a lock fuction that electronically turns off the counter dial. Two "VFOs" are built into the transceiver; they may be used independently or combined for split-frequency operation. A one-frequency memory is available.

Circuit Highlights

To describe the entire transceiver circuitry is beyond the scope of this column. Rather, I will highlight specific design areas that influence the overall performance.

The receiver front-end design is typical of today's high-performance equipment — band-pass filters followed by a doubly balanced diode-ring mixer. A switchable 10-dB-gain MOSFET preamplifier improves the receiver noise figure (sensitivity) on the higher hf bands. Note in the specification table that the receiver input intercept (P_i) drops 10 dB when the preamp is switched in; this means the ability to tolerate large input signals has decreased. This switchable preamp feature allows the user to choose between best sensitivity and best strong-signal performance — a choice not found in equipment that uses a simple attenuator.

A dual-conversion system allows the user to select several i-f filters. In the standard model, a 2.3-kHz crystal filter is used at the first i-f (9 MHz) and a ceramic 3-kHz filter is used at the second i-f (455 kHz). Several optional filters are available that increase the ssb-filter shape factor (by changing the second i-f filter to a crystal type) and narrow the cw bandwidth (by replacing the first i-f cw filter with a narrower bandwidth crystal type unit).

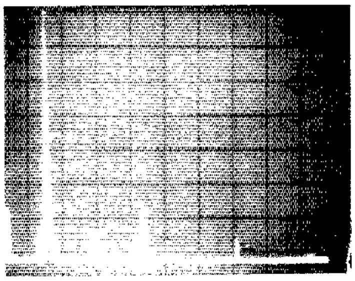


Fig. 1 — Spectral display of the IC-740. This is the worst-case condition, operating with 105-W output on 14 MHz. Each vertical division is 10 dB, and each horizontal division is 10 MHz. The IC-740 complies with current FCC specifications for spectral purity.

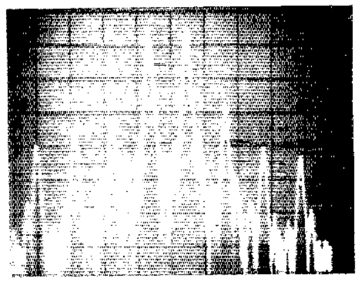


Fig. 2 — Spectral display of the IC-740 output during two-tone IMD test. Third-order products are 30 dB below PEP, and fifth-order products are 40 dB down. Vertical divisions are each 10 dB, and horizontal divisions are each 1 kHz. The transceiver was operated at the rated input power on the 14-MHz band.

Probably the most significant portion of the transceiver design is the frequency synthesizer. In this unit there are two PLL systems: one is used to generate the HFO injection signals, and the other generates the VFO signals.

Checkout and On-the-Air Performance

When the '740 arrived at Hq., I was anxious to see how it performed in the lab (other transceivers with frequency synthesizers have been reviewed and many have a poor synthesizer design, which results in a noisy receiver). To my delight, the receiver performance is impressive; in fact, the dynamic range figures shown in the specification table are better than those of some analog VFO transceivers. Receiver mixing and synthesizer products are nonevident. Of course, the supreme test for any piece of equipment is an on-the-air evaluation.

The first performance test came during the ARRL November SS cw contest. As always, signals were loud. No evidence of blocking or overload was noted. I noticed several signals with noise modulation on their carriers. These stations were probably using synthesized equipment; I'm sure the IC-740 was not at fault.

Although the "electronic" performance of the radio is impressive, the human engineering leaves much to be desired. The front-panel knobs are small; my large fingers found them hard to manipulate. Many of the functions are mounted using concentric control knobs; this makes the situation more difficult. The cw filter is available only when passband tuning is engaged; this requires the operator to press two buttons to change one function, and since there are no indicator lights, it is sometimes hard to discern why the cw filter is not working.

The cw waveform photograph looks great, but the sidetone is so mushy sounding that dots and dashes run together at speeds above 50 wpm. Most of the time I used my keyer sidetone for monitoring. Also the S-meter sensitivity is quite high.

Overall Thoughts

ICOM has done a superb job with the '740. Their experience in designing synthesized equipment has paid off; I have seen very few amateur products that are comparable. Don't let the small size of the unit fool you; it packs a lot of electronics in a small space. In fact, an optional *built-in* ac-operated power supply is available. Those potential buyers with large hands should play with the radio before purchasing one. Price classes: IC-740, $1099; FL-44 2.3-kHz second i-f ssb filter, $159; FL-45 500-Hz cw filter, $59.50; PS15 external power supply, $149; PS-740 internal power supply, $160. For more information on the IC-740 HF Transceiver, contact ICOM America, Inc., 2112 116th Ave. N.E., Bellevue, WA 98004 — *Gerry Hull, AK4L*

Yaesu FT-730R 440-MHz FM Transceiver

Manufacturer's Claimed Specifications	*Measured in ARRL Lab*
Frequency coverage: 440-449.975 MHz.	440-449.975 MHz.
Synthesizer steps: 25 and 100 kHz.	25 and 100 kHz.
Power output: 10 W.	15 W.
Mode of operation: F3 (fm).	Fm.
Spurious emissions: −60 dB or better.	Better than −64 dB.
Antenna connector: Type N.	Type N.
Receiver type: Double-conversion superheterodyne.	
First i-f: 46.255 MHz.	
Second i-f: 455 kHz.	
Sensitivity: 0.25 μV for 12-dB SINAD.	0.2 μV for 20-dB quieting.
Squelch sensitivity: 1 μV for 30 dB S/N.	0.56 μV for 30-dB quieting; 0.085 μV minimum, 0.34 μV maximum.
Selectivity: ± 7.5 kHz (−6 dB), ± 15 kHz (−60 dB).	
Audio output: 1 W into 8 ohms.	2 W @ 10% THD.
Audio output impedance: 8 ohms.	
Power requirements: 13.8-V dc (negative ground).	
Current consumption: 3 A on transmit (10-W rf output); 0.3 A on receive.	3.6 A (15-W output). 265 mA (squelched).
Case size (HWD): 2 × 5.9 × 7.1 in.†	2 × 5.9 × 7.1 in.
Weight: 3 lb 5 oz.	

†mm = in. × 25.4; kg = lb × 0.454

YAESU FT-730R 440-MHz FM TRANSCEIVER

☐ Fm transceivers for the vhf and uhf amateur bands are becoming more compact and versatile with advances in the state of the technical art. Rigs not much larger than you would expect for hand-held use are available for mobile (and fixed-station) operation, providing power outputs in the 10- to 15-W class. And they are packed with features that will amaze you if you haven't kept up with what has become available recently. The FT-730R is one such rig.

At first glance, the '730 may look like just another fm transceiver. The case is finished in a dark gray, almost black enamel. The rig sports a 1.4-in. edgewise D'Arsonval type of S meter that doubles as a relative-power-output meter.[1] In operation, the meter scale, with its red and black markings, is illuminated from behind. To the left of the meter is the digital frequency readout. This is a liquid-crystal type of display, with five 1/4-in. numerals displaying frequency to the nearest 100 Hz. This is a newly developed readout device with wide-angle viewing, and is also illuminated from behind. When power is removed, the black numerals disappear, leaving the frosty-gray area clear. Between the readout and the S meter are two LEDs, one above the other. The upper red LED indicates ON AIR, and the lower green LED indicates BUSY when the squelch has been opened.

The usual ± transmit frequency offset switch is located at the upper left of the front panel. Directly below are concentric VOLume and SQUeLch controls. To the right, below the frequency readout, is a large knob, obviously for frequency tuning. A multiposition MEMORY switch indicates some type of memory capability. The seven-pin microphone connector may seem a bit strange if you consider the need for only audio, PTT and common lines. There are also eight push buttons with various labels on the front panel, a clue that there are additional features. A 3-inch built-in speaker is located at the bottom of the enclosure, with a miniature phone jack on the rear panel for an external speaker.

[1]mm = in. × 25.4

Operating Features

But behind the neatly arranged front panel are all kinds of features that are not immediately apparent at first glance. The '730R is microprocessor controlled. In part, that's why the manufacturer can pack so many features into such a small space. The rig has 10 memories and two VFOs for frequency selection. A small lithium battery provides power to hold all 12 frequencies in memory when the rig is turned off or disconnected from the 13.8-V supply. (Life expectancy of the battery is five years or more.)

The frequency displayed on the indicator is that of actual operation, except that the first two digits (44) are omitted. In other words, add 440 MHz to the readout indication to get the actual operating frequency. A three-position rotary switch labeled RPT controls the transmitting-frequency offset. With simplex operation (center position of the switch), the readout remains unchanged when going from receive to transmit. At the right and left positions of the switch, the transmitter offset is plus and minus 5 MHz, respectively. The readout changes to indicate the transmitting frequency during offset transmissions. Provisions are also made for split-frequency operation, by receiving on a memory channel and transmitting on a VFO frequency (while following the instruction you set up with the offset switch). In this way, an almost unlimited number of frequency combinations may be used in split operation.

The '730R has a "goof-proof" feature, too. If by chance you have selected a frequency and an offset combination that would place your transmitted signal *outside* the 440- to 449.975-MHz frequency range, depressing the PTT switch on the microphone does nothing except display an E on the readout. Meanwhile, the transceiver remains in the receive mode.

Frequency Selection

With 10 memories and two VFOs, you might think that selecting desired frequencies would be a confusing or perhaps cumbersome process. Not so! In manual VFO operation, the appropriate VFO (A or B) is selected with a push switch. When pushed once, this switch locks itself in the IN position for one VFO. When pushed again, the switch unlocks and returns to the OUT position for the other VFO. It's as simple as that. Meanwhile, the frequency that was last selected on the unused VFO is held in memory, and is available instantly, merely by returning to that VFO.

VFO tuning is done with the main tuning knob. This is not a continuously variable control, but has closely spaced detents — 50 per knob revolution, to be exact. There is no

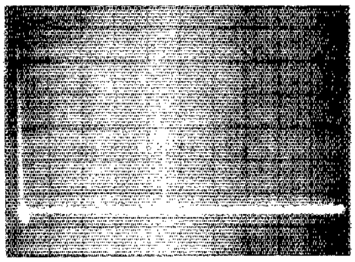


Fig. 4 — Spectral display of the Yaesu FT-730R. Vertical divisions are each 10 dB; horizontal divisions are each 100 MHz. Output power is approximately 15 W at a frequency of 448.775 MHz. The fundamental has been reduced in amplitude approximately 40 dB by means of notch cavities; this prevents analyzer overload.

mechanical stop on this control; it may be turned indefinitely in either direction. Switching to the next detent position changes the frequency by one step. The frequency steps are selectable at the front panel with a STEP push switch, either 25-kHz or 100-kHz increments. This switch is a momentary type, not a locking switch. For rapid frequency changes from one end of the band to the other, the 100-kHz steps are useful. Then, pushing the STEP switch toggles operation to provide 25-kHz steps for "fine tuning." The instrument continues to step in 25-kHz increments until the STEP switch is again depressed, when it returns to 100. The direction of frequency tuning is conventional — clockwise rotation of the knob to select higher frequencies, and counterclockwise for lower frequencies. There are no provisions to tune to frequencies that are not multiples of 25 kHz.

The 10 memories are selected with a rotary switch. Any VFO frequency may be stored in memory, merely by pushing a switch labeled M. (The previous content of that memory is lost when a frequency is stored.) Recalling a frequency from memory requires only pushing a "memory recall" switch and selecting the appropriate memory with the rotary selector switch. The transmitting-frequency offset functions are the same with memory frequencies as with VFO operation. This allows you to set up the memories on the output frequencies of repeaters in your area, and transmit in normal offset operation to hit the repeater input frequencies. With this feature, the operator who spends most of his time on only a few repeaters will have little use for the VFOs except for storing frequencies in memory. But, of course, two VFOs are nice to have.

Earlier, I mentioned a seven-pin microphone connector. Those extra pins provide for frequency-scanning operation, controlled from the microphone. Two switches, labeled UP and DWN, control this feature. During VFO operation, a momentary push of either switch shifts the frequency up or down by one frequency step. Operation is exactly the same as moving the main tuning knob by one detent position. But if either switch is held down for approximately one second, the entire frequency range of the rig is scanned automatically, either in 25- or 100-kHz steps, as controlled by the STEP switch.

A three-position slide switch on the rear panel permits you to select momentary interruption of the scanning on either a busy channel or else a clear channel, or to scan continuously without interruption. Let's say you set the '730R up to scan in 25-kHz increments for a busy channel. Scanning will start from the VFO frequency that is displayed when you initiate scanning, and proceed in the frequency direction according to which control switch you pushed. The scanning will continue until a frequency is tuned where a signal of sufficient strength to open the receiver squelch is found. Scanning will pause automatically for approximately five seconds, and then resume.

During that five-second period, you can halt the scanning operation by depressing the UP or the DWN switch or the microphone PTT switch. Now you are ready to transmit on that frequency (or an offset from that frequency) immediately, if you so desire. This is a handy feature if you are looking for signals on the band. Scanning of the entire 10-MHz range without momentary halts requires about 40 seconds. If no signals are found, the scanning progresses from the bottom of the frequency range to the top, and returns to the bottom of the range without delay to continue upward again (or vice versa, if you are scanning downward in frequency).

But that's not all the scanning you can do. If you wish, you can scan only the frequencies retained in the 10 memories. All the features mentioned above, except frequency steps, are available in memory scanning, too. This is superb if you want to search only your favorite repeater frequencies for activity.

VFO A or B, or readiness for memory scanning, is selected from the front panel. Actual scanning is controlled from the microphone only; there is no provision to initiate or terminate scanning from the front panel.

The '730R has yet another useful feature called priority channel operation. Using a VFO and a memory, this feature lets you listen on a frequency you select with the VFO, but provides automatic checking of the memory frequency for activity every five seconds. If the memory frequency is busy (or clear, as selected by the three-position slide switch), the scanner will halt on the memory channel. Otherwise, it will return immediately to the VFO frequency, and hardly a word of what is being said will be missed.

The FT-730R comes with yet other features. One is a tone-burst generator. When activated with a two position rear-panel slide switch (ON and OFF), an 1800-Hz tone having a duration of approximately 0.5 second is keyed each time the PTT switch is activated. For longer bursts, a front-panel CALL switch provides for manual control of the burst duration.

An optional feature contained in the review unit is a tone-squelch decoder. Many repeaters operate with such decoders on their input frequencies, perhaps more commonly known as PL® operation. A front-panel push switch on the '730R activates this feature to prevent unwanted signals from breaking the receiver squelch. An encoder in the equipment provides the subaudible tone at all times during transmission.

As shown in the title photo, a mounting stand for using the '730R at a base station is available. An accessory FP-80 ac-operated power supply is also offered. For mobile operation, a universal bracket is supplied for under-dash mounting. The transceiver may be removed from the bracket merely by unsnapping a locking bar and sliding the rig forward. Mobile installation requires a vehicle with a negative ground. Power connection is made through a keyed, two-pin plug, supplied with leads for interconnection to the power source.

The FT-730R is manufactured by Yaesu Electronics Corp., P.O. Box 49, Paramount, CA 90723, tel. 213-633-4007. The price class is $399. Various versions of the FT-730 are manufactured for use in countries other than the U.S. and Canada. These models offer various features, such as a frequency coverage that is 10 MHz lower and a different tone call frequency. — *Jerry Hall, K1TD*

FOX-TANGO 2.1-kHz KENWOOD TS-830S TRANSCEIVER FILTER MODIFICATION

☐ For the past two years, I have used a TS-830S transceiver at my home station. The variable bandwidth tuning (VBT) and IF SHIFT controls are extremely valuable for the elimination of adjacent-channel interference. On cw, these controls eliminate just about any QRM encountered.

On ssb, however, the stock Kenwood filters didn't perform as well as I expected. In fact, while operating on a crowded band, I noticed many instances in which using the VBT and IF SHIFT controls together wouldn't put a dent in the QRM! Since most of my operating time is spent on cw, this problem didn't disturb me too much. When a serious operation in November Sweepstakes was undertaken, however, the shortcoming of the standard filters became apparent.

A few weeks after the contest, I noticed a Fox-Tango advertisement in *QST*. A single Fox-Tango 1.8-kHz filter worked very well in my TS-520S several years ago, so the twin 2.1-kHz filter TS-830 modification was extremely appealing to me.

The filter modification comes with several photocopied pages of instructions to assist in the installation of the filters. The documentation is rather extensive; in fact, *too* much information is supplied. A prospective modifier *may* become discouraged after reading the complete package. The modification is relatively simple to perform, so don't feel intimidated by the pages of paperwork!

Modification installation is fairly straightforward, but care should be used during the operation. Screws that hold the i-f board in place must be removed and the board tilted 90° to allow unsoldering and removal of the stock filters. Exercise care during this operation, as the circuit board traces in this area are narrow and close together. The filters supplied for this modification are *not* direct plug-in replacements for the stock filters, so it is necessary to mount them elsewhere in the transceiver.

Instructions suggest mounting the new filters on top of the existing cw filter (if installed) with double-sided tape. Suitable tape is supplied with the kit. This approach didn't appeal to me, so I placed them on a piece of perforated board. This board is mounted on the chassis, just to the left of the power transformer.

Connections between the filter terminals and the i-f board were made with RG-174/U coaxial cable (supplied). The supplied cable length is sufficient for filter installation atop the cw filters, per instructions, but more was required for my installation.

Filter installation took approximately four hours. Much of this time was spent fabricating the filter mounting board and realigning the i-f after installation. Realignment is not required, but highly recommended. If an rf signal generator is available, follow the i-f alignment

Table 2
Receiver Performance

Specification	*Before*	*After*
MDS		
80 Meters	−129 dBm	−133 dBm
20 Meters	−130 dBm	−133.5 dBm
IMD D.R.		
80 Meters (low)	85 dB	88 dB
80 Meters (high)	78 dB	81 dB
20 Meters (low)	83 dB	88.5 dB
20 Meters (high)	77 dB	81 dB
Third-Order Intercept		
80 Meters (low)	− 1.5 dBm	− 1 dBm
80 Meters (high)	−12 dBm	−11.5 dBm
20 Meters (low)	− 5.5 dBm	− 1 dBm
20 Meters (high)	−14.5 dBm	−12 dBm

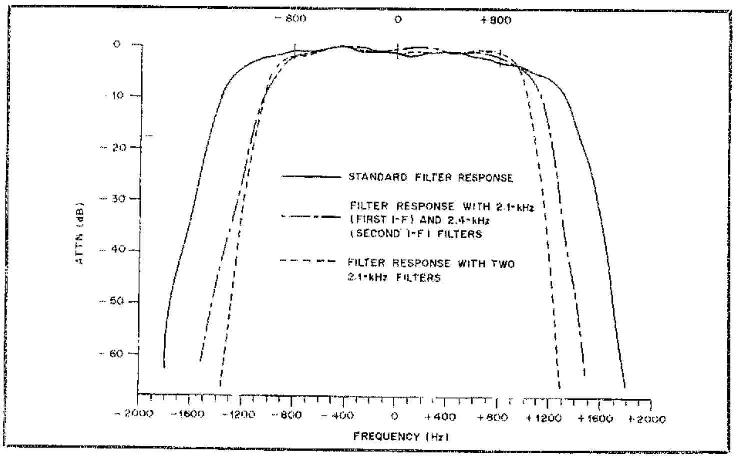


Fig. 5 — Filter response curves for the TS-830S with and without the Fox-Tango filters.

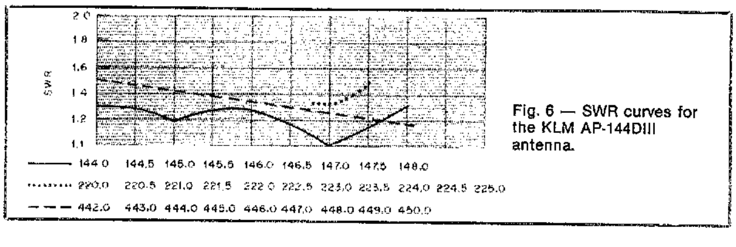


Fig. 6 — SWR curves for the KLM AP-144DIII antenna.

procedure in the TS-830 *service* manual. If either is not available, the procedure in the user's manual will suffice.

Overall, I'm pleased with the performance of the "new" '830. The band-pass response curves of a stock TS-830, one with a new 2.1 kHz *first* i-f filter (standard second i-f filter) and one with *both* new 2.1-kHz filters, are shown in Fig. 5. Measured −60 dB bandwidth of the unmodified '830 was 3.4 kHz. After modification, the measured bandwidth at −60 dB is 2.7 kHz — a 700-Hz improvement.

Receiver performance characteristics (before and after) are shown in Table 2. An approximate 3-dB improvement in two-tone dynamic range and third-order intercept is noted. This, in addition to the narrowed bandwidth, made the necessary work worthwhile.

The FTK-830 modification is available from Fox-Tango Corporation, Box 15944T, West Palm Beach, FL 33406. Price Class: $150. — *Michael B. Kaczynski, W1OD*

KLM AP-144DIII BASE STATION VHF ANTENNA

☐ Some hams may think that once they have seen a few base-station antennas for 2 meters, all such antennas are basically similar. The KLM AP-144DIII is different.

This antenna is a collinear, double 5/8-wavelength vertical for the 2-meter band that also incorporates a discone antenna for 70-cm (440-450 MHz) operation. The feed point is weatherproof, being housed within a machined aluminum hub. A sealed static-charge arrestor is fitted to the feed point. KLM claims the arrestor will offer some degree of protection to the transceiver in the event of static buildup.

All the parts and the machining of the KLM AP-144DIII are of a high standard, and everything fit together beautifully during construction. The clearly written instructions include excellent exploded drawings of the antenna, especially of the feed-point housing. Several reducing adapters are provided for the feed point to accommodate different sizes of coaxial cable. Coaxial cable ranging in size from RG-58/U to slightly larger than RG-8/U can be used. I am feeding the review model with RG-8/U. Two small tubes of silicone sealant are provided in the kit, along with specific details for weatherproofing the antenna. I used only half of one tube to seal the antenna; this leaves plenty of sealant for other uses around the shack. In all, complete assembly of the antenna took about one hour, with an overnight wait for the sealant to set.

At my QTH, the antenna is mounted atop the tower at a height of 75 feet. The antenna performs better than another commercially manufactured collinear that it replaced. For purposes of comparison, I used a distant (95 mi) 2-meter repeater in northwestern Massachusetts. Before I installed the new antenna, my signals into that repeater were marginal, and the repeater output would scarcely open the squelch of my transceiver. The first QSO with the new antenna netted me a signal report of almost full quieting; I now could copy the repeater S 3 to S 4. Overall, the improvement was immediate and gratifying. SWR checked across the 2-meter fm band is shown in Fig. 6. Worst case, it is less than 1.5:1. SWR on 440-450 MHz slopes from 1.5:1 to near unity. Signal reports on 440-450 MHz were excellent through the local repeater, as well. Although the antenna instructions provide for some adjustment of antenna length for minimizing SWR, none was necessary.

The SWR chart for typical operation, supplied by KLM, suggests that the AP-144DIII can be used on the 220-MHz fm band with a reasonable SWR, although KLM makes no claims to that effect. I decided to try operation on 220 MHz, and I'm pleased to report that this antenna appears to operate on that band with an SWR of approximately 1.5:1. The frequencies used were three standard 220-MHz repeater pairs and 223.50-MHz simplex. The entire 220-MHz fm band was not checked, inasmuch as the rig I used is a crystal-controlled unit. This three-band capability should definitely interest many active vhf ops.

The KLM AP-144DIII appears to be a well-engineered antenna. Materials and finish are first-rate; performance has been excellent on 144, 220 and 440-450 MHz. It certainly is nice to put up a two-band vhf vertical antenna and find you've got operation on an extra band (220 MHz) along for the ride! The AP-144DIII is imported by KLM Electronics, P.O. Box 816, Morgan Hill, CA 95037. Price class: $112. The same antenna, less the uhf discone, is available in the $99 price class as the AP-144DII. — *Sandy Gerli, AC1Y*

WEST JERSEY COMMUNICATIONS PRODUCTS 80-METER "BN CAGE" ANTENNA

☐ Many amateurs would like to find an antenna that provides coverage of the entire 80-meter band with an SWR of 1.5:1 or less. Solid-state rigs reduce power output if the SWR is much higher than this. Of course, you can use a Transmatch to present the required low SWR to your rig, but that seems to be an unnecessary set of controls to manipulate. The advantage of a "no tune" rig is, after all, being able to tune from one end of the band to the other without adjusting any controls.

The cage antenna has been around since the early days of radio, and was discussed in *QST* articles in recent times.[4,5] The idea behind the cage antenna is really quite simple. As the length-to-diameter ratio of an antenna element decreases, the resonant length for a given frequency will decrease. The antenna bandwidth will also increase. This principle is more evident in higher-frequency designs, where aluminum tubing is often used for the radiating elements. It is impractical, however, to use large-enough diameter tubing on 160, 80 or 40 meters. One

[4]A. Harbach, "Broad-Band 80-Meter Antenna," *QST*, Dec. 1980, p. 36.
[5]J. Hall, "The Search for a Simple, Broadband 80-Meter Dipole," *QST*, April 1983, p. 22.

method of producing an antenna element of large effective diameter is to space a number of wires around the perimeter of a cylinder. At the ends of the cylinder, or cage, the wires can be brought to a point in a conical taper, and joined.

There are a number of parameters that will affect the feed-point impedance and bandwidth of the cage. The size and number of wires, the diameter of the cage and the length of the taper are all important. WJCP has experimented with all of these parameters, and appears to have balanced them in a workable design.

The 80-meter antenna kit comes in a box 10-1/2 inches on a side and weighing about 18 pounds.* It includes everything you will need to put the antenna together, except some basic tools. In the box you will find eight molded-plastic spreaders, a spool of no. 14 enameled wire (approximately 800 feet) and 100 feet of coaxial cable with a PL-259 on one end and pigtails ready to connect to the antenna on the other. You will also find 50 feet of 1/8-inch Nylon rope, center and end insulators, and all required hardware. Even a small piece of emery cloth for removing the enamel insulation from the ends of the antenna wire and a small packet of Noalox® compound for use with the aluminum cable connectors are included. The coaxial cable is about the size of RG-59/U, but is sold by Channel Master and is specified to have a characteristic impedance of 67.2 ohms.

The instructions that come with this antenna are detailed and seem to cover all aspects of construction. The manufacturer recommends that you read them through at least once before beginning work. I certainly agree that this is a good idea. It is much easier to understand each step when you are aware of what will have to be done next.

One person could build the antenna, but a helper sure is nice, at least for measuring and cutting the wires. Mike Kaczynski, W1OD, assisted me with most of the construction. At one point you are directed to secure a collet 11 inches from the end of each wire. This is to prevent the last spacer from sliding on the wire when the conical taper is formed at the end. The instructions suggest that you use a crimp tool or ordinary gas (slip-joint) pliers to crimp the collet in place. (These collets look like crimp-on wire connectors.) We found it impossible to dent the ones supplied, even with a large pair of pliers. Ultimately, we had to resort to using a hammer and a piece of scrap iron to crimp the collets in place.

I found the insulation coating on the wire to be quite heavy, and I needed about three times as much emery cloth as had been supplied to complete the job of stripping the wire ends. A pocket knife could be used to scrape the enamel off, but the danger with this procedure is that you may nick the wire, causing it to break after the antenna is up.

As the four ends of the cage sections are completed, you are instructed to twist the six wires together, apply a *thin* coating of Noalox and tighten an aluminum cable clamp over them. The manufacturer suggests that you may want to solder the wires together first. I would say this is a required step if you want the electrical connection to be maintained. The amount of Noalox supplied really is insufficient to cover the wires. A few years ago, I purchased a container of this compound at an electrical supply store. It is inexpensive, and I apply a liberal coat to all exposed connections. The investment for a container of Noalox would be worthwhile for most hams involved in antenna work. Total assembly time for the cage was about three hours.

*mm = in. × 25.4; m = ft × 0.3048; kg = lb × 0.454.

Deployment and Performance

The first real snag in the operation came when Mike and I went to move the entire assembly into position to put it in the air. The antenna weighs only about 15 pounds, but to hold the entire length off the ground by pulling on the ends, 100 pounds or more force is needed! If any part of the antenna is allowed to touch the ground while you are moving it, the spreaders will roll, twisting the cage wires together. Believe me, it is very difficult to get all of the twists out once this happens! A third person to hold the center insulator would be a great help for moving the antenna.

Initially, one end of the dipole was attached to a 40-foot tower atop the ARRL Headquarters building and the other end sloped down to a tree. The feed line was brought in to Hq. club station W1INF, and the SWR curve was plotted. A Bird Model 43 wattmeter was used to make these measurements. We found it necessary to prop the center insulator with a wooden pole to hold it about 10 feet above the corner of the building. This resulted in the SWR curve shown in Fig. 7. The curve compares favorably with the manufacturer's claim of an SWR of 1.5:1 or less over the entire band.

The amount of sideways pull exerted on a tower with the antenna in this configuration is quite large. Any additional tug on the antenna, or a slight breeze, would cause the tower to sway noticeably. We did not leave the cage up this way for very long! If you want to support the antenna from your tower, I would recommend that you hang the center insulator from an arm off the tower.

After carefully coiling the antenna wires to prevent kinks or tangles, I took it home for further evaluation. My yard has two tall maple trees in the center. There is no way to support an antenna above those trees, so I decided to use a branch of one of them as a center support. About half of one side of the dipole goes through the top branches of this tree, about 30 feet off the ground. One end is tied off to the peak of my house roof and the other is supported by another maple tree. One problem that developed is a tendency for the wires to twist once the cage is in the air. This may be a result of having "rolled" the antenna earlier when we tried to move it. Even though this arrangement would have to be considered less than ideal, a check of the SWR curve revealed it to still be less than 2:1 over the entire 80-meter band. Not bad!

The real question about any new antenna is, "How does it perform compared to the old antenna?" In many cases, such a comparison is difficult because the old antenna is taken down to make room for the new one. My comparison antenna was a 135-foot inverted L fed through a Transmatch. I did not have to take this antenna down, and was able to switch between the two conveniently. The inverted L is mostly horizontal, also goes through part of a maple tree, and is about the same height as the cage dipole. This should permit a reasonable comparison to be made. Received signals are two to three "S" units stronger on the cage antenna. Transmitted signal reports indicate the same results. For a while, I had a hard time convincing some of my friends on the Eastern Pennsylvania Emergency Phone and Traffic Net that I had not moved back to PA! The cage antenna made a definite improvement in my 80-meter antenna system.

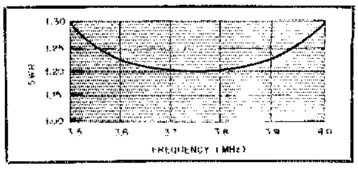


Fig. 7 — SWR curve for the 80-meter "BN Cage" antenna.

One ordeal I was not able to put this antenna through during the review period was the test of a typical New England ice storm. Several Headquarters staffers have virtually guaranteed me that this antenna will come down in such a storm. Their feeling is that the weight of the ice build-up on the cage wires will break either the support rope or the wires themselves. My own feeling is that ice is not going to build up on the wires. The insulating coat is quite smooth, and any water or ice just seems to fall off. One snowstorm did stick to the coaxial cables, the inverted L wire and even to my Cushcraft R3 vertical. The snow built up to about two inches in diameter. The cage wires remained clear, however. This seems to support my feeling, but I'll have to wait for a real ice storm to know for sure.

Specifications

A few things about the antenna specifications are worth mentioning. Our measured SWR curve does not go as low at the center of the band as the manufacturer suggests it should. After some thought, I realized that the 67.2-Ω coaxial-cable impedance would provide some mismatch to the Bird wattmeter and transmitter, which are designed to operate into a 50-Ω resistive load. Under these conditions, the feed line will act like an impedance transformer, and the impedance seen by the transmitter and wattmeter will depend on line length and the degree of mismatch at both ends. The measured SWR will not necessarily be the actual SWR caused by a mismatch at the antenna. Of course, there are many other factors that will influence the SWR measurement, but I found these few thoughts shed some light on one aspect of the problem.

The 80-meter cage antenna has a wind-loading surface area of about 2.5 ft^2. In an 80-mph wind, this would produce about 60 pounds of force on the antenna. Add this to the 100 pounds or so of force needed just to hold the antenna up without unreasonable sag, and you will realize that a couple of very sturdy end supports are required. This should reinforce what I said earlier about not using your tower as an end support.

The manufacturer claims the coaxial cable supplied with the antenna will handle 1000 W, but I did not test this capability because I do not have an amplifier. A few calculations indicate that this should be reasonable power for a cable this size to handle.

Besides the 80-meter antenna, WJCP also has cage antennas for 160 and 40 meters. If you like to operate at both ends of these bands and want an antenna that will cover the entire band with a low SWR, then I recommend the "BN Cage" antennas. They are available from West Jersey Communications Products, 932 Oakland Ave., Burlington, NJ 08016. Price class: 160-meter cage, $180; 80-meter cage, $100; 40/15-meter cage, $80. — *Larry Wolfgang, WA3VIL*

Technical Correspondence

Conducted By
Dennis J. Lusis,* W1LJ

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THE QUAD-J-COLLINEAR ANTENNA

☐ This new antenna is essentially a four-element, wide-spaced, collinear broad-side array. However, it has been reduced to the minimum possible overall size without diminishing its performance. The Quad-J-Collinear antenna is a highly efficient, bi-directional array that produces a moderate 6-dB gain over a 1/2-λ dipole. If it's installed so the lower element is at least 1/2 λ above ground, this array will provide low-angle, elliptically polarized radiation useful for DXing. The elliptical-polarization feature provides a desirable polarization-diversity effect, which overcomes signal fading caused by constant variations in ionospheric refraction.

Fig. 1 illustrates the basic principle of operation, and may also be used as a design guide. Fig. 2 provides the dimensions for a 10-m antenna. The array is constructed from copper wire for light weight and low wind resistance. It may be suspended by nylon ropes from existing towers, masts or even trees — provided that sufficient separation between supports exists.

This is a fairly broadband type of array and provided the dimensional proportions are maintained, it will work effectively over a large frequency range. Overall bandwidth is primarily governed by the "J" type phasing stubs and the method employed in transmission-line matching.

Feed methods are left to your particular station requirements. However, it should be known that the feed point has a high impedance (>1 kΩ), and is balanced. Because of these characteristics, either tuned feeders or a 1/4-λ matching section with a balun (Fig. 2) may be used. The length of coaxial cable from the balun to the transceiver is non critical in this application. — *Richard Schellenbach, W1JF, Reading, Massachusetts*

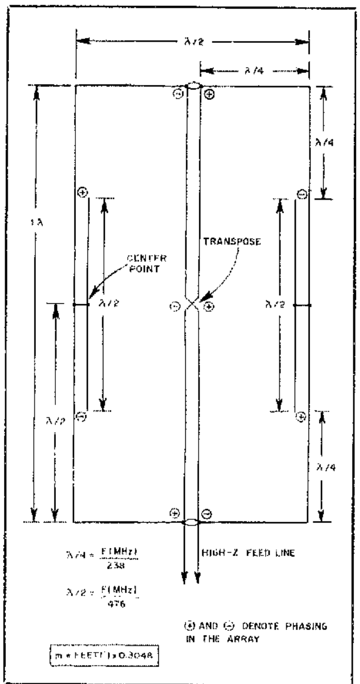


Fig. 1 — Drawing of the Quad-J-Collinear antenna, showing principal phase and element-length relationships.

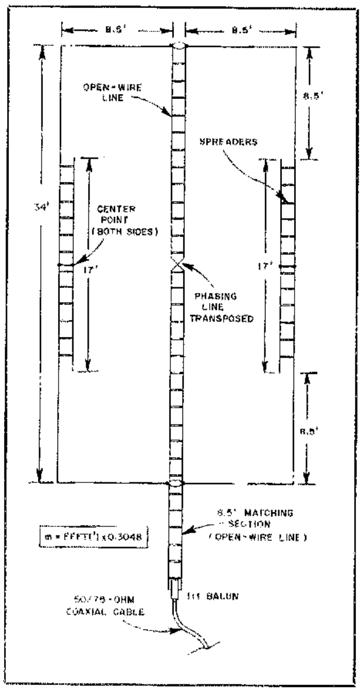


Fig. 2 — Dimensional drawing of a Quad-J-Collinear antenna for the 10-m band. Resonance is set for 28.0 MHz.

IMPROVEMENTS FOR THE AA6PZ POWER CHARGER

☐ The AA6PZ Power Charger (Dec. 1982 *QST*, p. 17) has proven most useful in maintaining my HT battery in both mobile and home operation. Zander's article is informative as well as descriptive, and the charger fills the need to quick-charge a battery safely. I have a few comments that will correct a minor circuit error, and help the builder ensure that his battery is charged to maximum capacity.

Correspondence received with the power-charger pc board indicates that the emitter of Q4 is connected to the wrong side of the ammeter jumper. A voltage drop in the meter could turn on Q4 and limit the current to a lower-than-normal value. Cut the pc trace and reconnect the emitter of Q4 to the correct side of the jumper.

The text implies that R5 should be left out for a 700-mA charge-current limit; however, any variation because of the tolerance of R1 and other resistors will affect the current limit value markedly. In my charger, R1 was actually 1.1 Ω and the limit was only about 250 mA! A value of 1 kΩ for R5 was required (as indicated by Eq. 1) to bring the limit to 700 mA.

Calibration of the output voltage is extremely critical, and most inexpensive VOMs are definitely not accurate enough to do the job. Furthermore, typical milliammeters have a relatively high series resistance and will affect the charge current when removed from the circuit. Here's why: To charge the battery to 100% of capacity, the individual cell voltage must reach a minimum of 1.42 V at a charge rate of 0.1C, or 11.36 V for an 8-cell battery without an internal diode.[1] A set point of 11.20 V, or 1.40 V per cell (as suggested), will result in only 15-37% of full charge. I experienced an apparent loss of battery capacity, which was actually due to insufficient charge current at 11.20 V. Similarly, voltmeter errors as little as 0.16 V or milliammeter-induced voltage drops of the same amount will result in failure to charge the battery properly.

The charge rate must taper to the normal charge rate of 0.05C to 0.1C at 1.42 to 1.44 V per cell, or 22.5 to 45 mA for a 450-mAh battery.[2] This charge rate may be left on indefinitely. A charge rate of 0.02C is a trickle charge, and it is used only to maintain a battery which previously has been fully charged. The power charger should not taper below 0.05C, or the battery will never reach 100% of capacity; 0.1C is a better minimum value.[3]

You may wish to calibrate the charger with a digital voltmeter as follows:

1) Using the slow-rate "wall charger," give your battery a 16-hour charge.

2) Connect the battery to the power charger for at least one-half hour. Monitor the current with a meter having a resistance of less than 0.2 Ω. You may use a current-sampling resistor of 0.1 Ω in series with the charger, and measure the voltage drop across it with a *good* DVM. For example, a reading of 0.0045 V indicates a charging current of 45 mA. (This measurement is usually beyond the means of anything but laboratory-grade instruments.)

3) If a laboratory-grade DVM is not available,

*Assistant Technical Editor

[1] *Nickel-Cadmium Application Engineering Handbook*, 2nd ed., General Electric Co. (Battery Business Dept., Box 861, Gainesville, FL 32602), pp. 4-22, 4-33.

[2] *Ibid.*, p. 4-11.

[3] *Ibid.*, p. 4-12.

connect the charger directly to the battery, without any series resistor or ammeter, using the leads intended to be permanent. Measure the battery voltage with a DVM or laboratory-grade analog meter.[4] Adjust R2 for a battery voltage of 1.42-1.44 V per cell, adding 0.7 V if your battery pack has an internal diode. Adjust R3 so that the yellow LED barely glows, then readjust R2 (because of interaction). Alternatively, measure the voltage across R1 and adjust R2 and R3 to give a reading corresponding to 0.05C-0.1C. V = R1 × 0.1C, so for a 450-mAh, 8-cell battery, V = 1.0 × 0.045 A = 0.045 V.

As a final thought, I strongly suggest the use of a multifingered or other effective heat sink and thermally conductive grease together on Q1. Without them, excessive temperatures and thermal cycling will eventually damage the transistor, possibly damaging your battery from overcharge. — *Sam Bases, K2IUV, Yonkers, New York*

HYSTERESIS LOSS AND METAL DETERIORATION

□ With reference to the article "A Homemade High-Power Tuning Capacitor" (June 1983 *QST*), I suggest that brass washers be used for this application. From my experience with high-frequency rf circuits, I have found that hysteresis loss will eventually cause steel washers to disintegrate where high currents are present. — *Bill Barnard, Jr., WB6ACV, Flintridge, California*

MORE ON REFLECTED POWER

□ Since Walter Maxwell, W2DU, was gracious in suggesting that my own article be consulted,[5,6] I thought it propitious to add a few comments. Maxwell's most recent effort quite correctly states the reflected-power case.[7] My own article also attempted, in a heuristic fashion, to describe the mechanism of reflection. I took into account matters seemingly most contradictory to experience: That power is actually reflected and should not be confused with standing waves; and that power "flowing" backward doesn't necessarily cause the source to suffer for it. I'm sure that more than one amateur has suggested the possibility that reflected power must somehow dissipate in the output tank circuit. The corollary, that "real" power is not reflected, as suggested by some authors, must therefore come as a relief to those who don't understand the mechanics involved.

The practical result is that an understanding of this subject has little to do with the everyday operation of most stations. We have all been warned to avoid high SWR. If you operate a modern solid-state, broad-band rig, protective feedback circuits simply reduce or eliminate the power output if feeding into a gross mismatch. Thus, new amateurs might conclude that operating into a high SWR limits the ability of the amplifier to transfer power to the load. Vacuum tube users are more likely to conclude that while it is entirely possible to transfer maximum (or nearly so) power, reflected power would cause all kinds of havoc to the final tubes and tank circuit. Of course, it's all nonsense (although some types of damage can result when the SWR is excessive), but most hams avoid trouble by resorting to a Transmatch circuit. Thus, as Maxwell long ago advised, SWR really doesn't mean all that much.

I think the most interesting of the arguments suggested by Maxwell is the analog to visible light reflected from a mirror. Those amateurs who have seen light both reflected from, and transmitted through, a half-silvered mirror must appreciate, *a priori*, that reflected power is real. You can do this with a flashlight as long as you are willing to accept that the essential difference is one of wavelength. — *Martin R. Kramer, K2KGF, Brooklyn, New York*

[4]Mercuric-oxide "mercury" cells (used in photographic light meters) can be used as precision voltage-calibration standards. They are either 1.35 or 1.4 V, depending on type.

[5]"Feedback," *QST*, April 1983, p. 40.

[6]M. Kramer, "Reflected Waves and Mismatched Loads," *CQ*, June 1978.

[7]W. Maxwell, "The Reality of Reflected Power," Technical Correspondence, *QST*, Feb. 1983.

LESSONS LEARNED

□ I recently had a potentially disastrous electrical fire in my car. Here's what happened: A short developed in the line carrying 12-V dc to the slide-in mount for my 2-meter rig. (Everything was burned so badly I couldn't pinpoint the exact cause.) Although the rig had an in-line fuse, the short was on the battery side of it. Consequently, the 12-V power line got so hot that the insulation started to burn and parts of the car were also ignited.

At 55 mi/h this was quite a thrill. I stopped the car and pulled (ripped was more like it!) the rig from the slide-in mount. That of course didn't help a bit. I ran back to the trunk, where the battery is located, grabbed a pair of diagonal cutters and cut the wire coming off the positive battery terminal. I then ran back to the front of the car and extinguished the fire by blowing and swatting it out.

All this was bad enough, but I was very lucky not to lose the whole car. I learned the following lessons:

1) It's a good idea to check your mobile installation periodically for possible loose connections and cut and chafed wires.

2) In addition to the fuse that protects your rig, make sure *every* wire you install is adequately fused *at the power source*. Alternatively, rely on the car fusing by making sure the wires you install going to the rig are at least the same or bigger gauge than the wires of the 12-V source you are tapping. (The typical 2-meter rig or car radio/stereo can trap you. The in-line fuse is normally installed close to the unit, with 4 or 5 feet of cord left for attachment.[8] Connecting that wire to a source like an ignition switch exposes you to exactly the situation I had!)

3) Carry a fire extinguisher.

All of this may seem like overkill, but, believe me, all you need is one fire at 55 mi/h to become a believer. An ounce of prevention... — *Hank Garretson, W6SX, Edwards, Californa*

KEEPING RF OUT

□ You *can't* keep rf out of the shack! Even the amateur who uses coaxial cable to the antenna often weeps that the microphone burns his lip, the vfo chirps, the keyer or computer doesn't work, or that sparks can be drawn off of the equipment cabinets. The poor guy mutters that he's "gotta get a better ground" and tries to figure out the best way of running a wire to earth.

His real problem isn't ground, but that all the pieces of equipment (including the person holding the mike) are at *different* rf voltages. That potential difference causes feedback and sparks.

First, *all* equipment cabinets should be wired together with the shortest possible connections. The cabinets can be strung together in a line, with a good dc ground somewhere around the center. (This also is a good shock and fire preventative.) Sometimes the connections are made to an inch-wide copper strap running under the back lip of the operating desk.

If the desk isn't metal, the underside can be turned into a ground plane by gluing aluminum foil or tacking copper screen to it. If grounded to the equipment, this plane will help all equipment resting on it (including the amateur, through his arms) to stay at the same rf potential. This markedly reduces sparks, feedback, and weeping. — *David T. Geiser, WA2ANU, ARRL TA, New Hartford, New York*

[8]mm = in. × 25.4; m = feet × 0.3048

Feedback

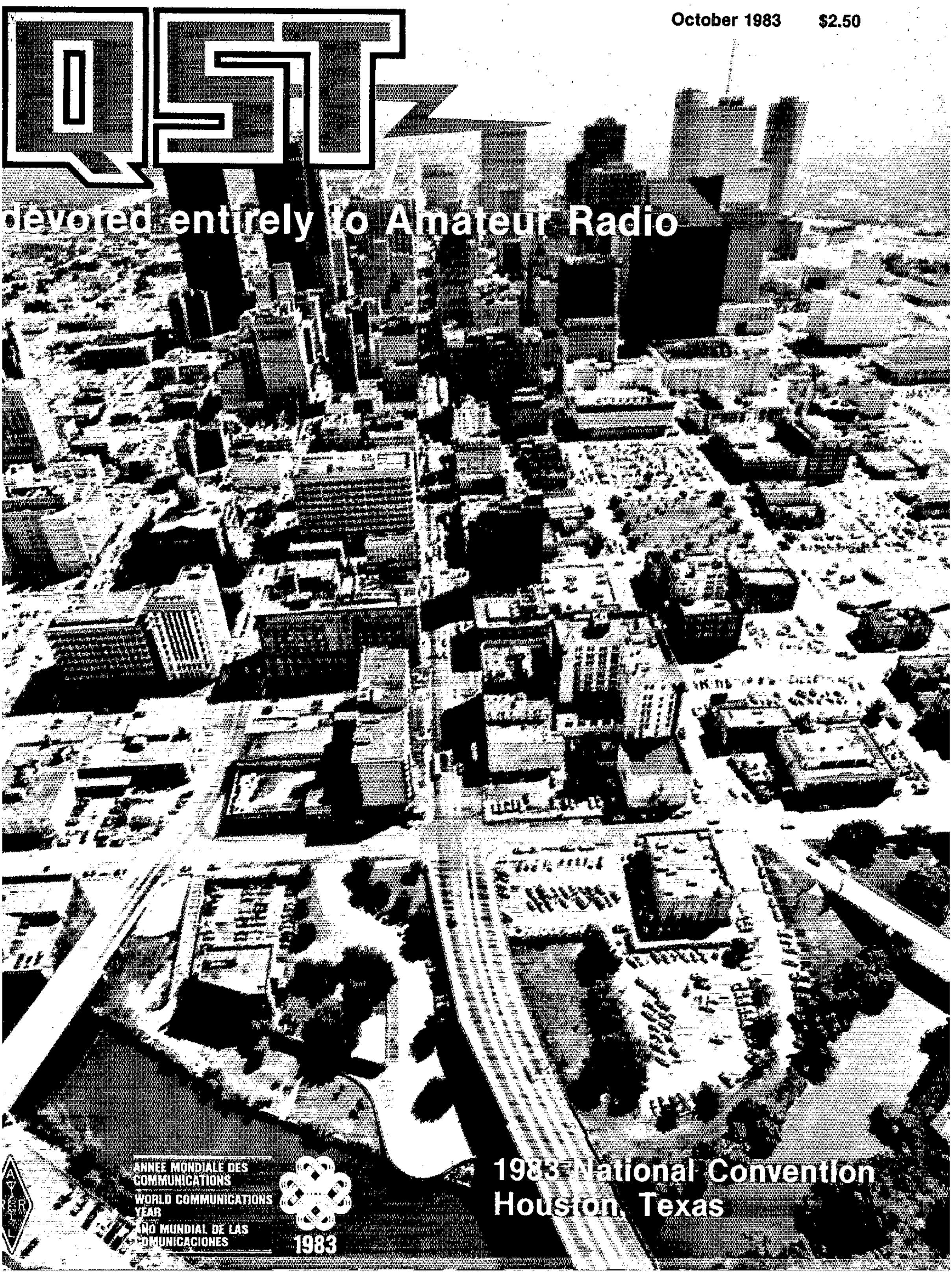
□ Please note: Pc boards and kits of parts for the ASCII/Baudot character generator in the July issue of *QST* ("A Serial ASCII/Baudot Character Generator You Can Build") are available from Circuit Board Specialists, P.O. Box 969, Pueblo, CO 81002.

□ A number of readers have expressed dismay over Fig. 1 of "Wire Antennas for the Beginner," June 1983 *QST*, p. 34. Some type of stress relief should be used at either end of the antenna support cable to prevent damage to the vent pipe from tree movement. Thanks to the many who wrote us!

□ In "An Introduction to AMTOR," July 1983 *QST*, there is an error in the bibliography. The reference to A. Daley should read: .... *IEEE Transactions on Vehicular Technology*, VT-26, No. 3...

Strays

In recognition of the 7600 radio amateurs in Maryland, Governor Harry Hughes proclaimed June 20-26 as Amateur Radio Week. Participating in the presentation are (l-r) AK3B, KK3C, Governor Hughes, WB3ELV and KB3OB.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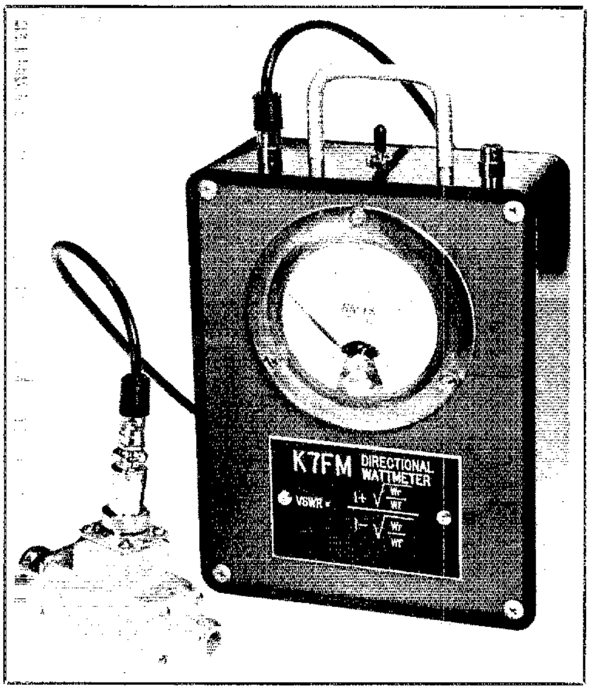

You say that good vhf/uhf wattmeters are expensive? Not this one!

By Colin Lamb,* K7FM

A calibrated, directional wattmeter is treasured by the vhf/uhf enthusiast. Commercial equipment is accurate, but expensive — and most often cannot be used remotely.

The line sampler described over a decade ago in *QST* by McMullen, W1SL, has become a classic in homebuilt directional couplers.[1] It can be built from inexpensive, common plumbing parts, but I decided to construct it differently, avoiding the need for a blowtorch and also to allow for a reversible sensing element. Before attempting to build the unit, you should review McMullen's original article (or a subsequent description in *The ARRL Antenna Book* or *FM and Repeaters*) for construction technique and theory.

Construction

This wattmeter can be built using only simple hand tools, and calibrated to the same accuracy as many commercial units. The line sampler is built around a 1/2-in. *square stock* brass "T" fitting (Fig. 1).[2] This common plumbing item has exactly the same width as the flange of a type N or uhf connector. Since the ID of the T fitting is actually 3/4 in., a piece of 5/16-in.-OD copper tubing is used as the inner conductor, which turns the T into a 50-ohm section of transmission line.

The T should be drilled and tapped to a 4-40 thread for the coaxial connectors (Fig. 1). Use a punch to line up the holes before drilling. When tapping, do not turn the tap more than 1/2 turn without backing it off to clear the threads. After the holes

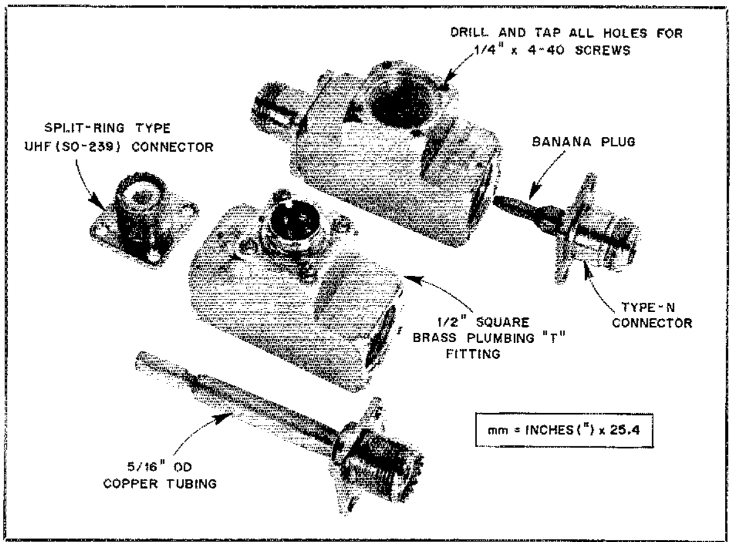

Fig. 1 — Exploded view of two line samplers, illustrating various construction techniques. The sampler at the top uses type-N coaxial connectors, and a banana plug/jack combination for the center conductor. In the bottom sampler, split-ring uhf (SO-239) connectors are used.

[1]Notes appear on page 13.

*Rte. 1, Box 65A, Newberg, OR 97132

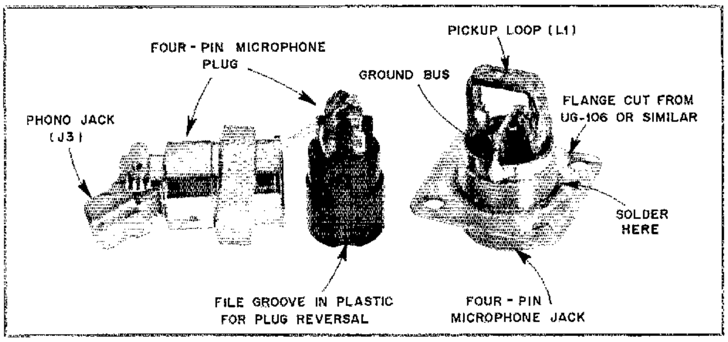

Fig. 2 — Detail of the pickup loop and detector assembly. See text for details.

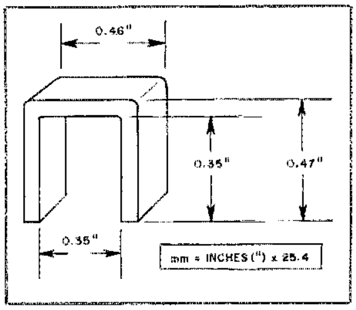

Fig. 3 — Dimensions of the pickup loop, L1. It is constructed from tinned shield braid taken from RG-59/U cable.

are completed, 1/4-in. long 4-40 brass screws are used to hold the connectors in place.

Since the coaxial connectors cannot be soldered after assembly, it is necessary to secure the inner conductor by an alternative method. I prefer to use an N connector that has a removable center pin retained by a split-ring washer. This is the method described in McMullen's article. However, most N connectors available today do not use a split-ring retainer. Fortunately, Amphenol still makes one; the part no. is 82-368. This item is approximately twice the price of a regular N connector, because it is silver plated and has a gold-plated center conductor. If you use the split-ring method of construction, it is only necessary to use one of the special connectors per sampler.

I have found that many of the more common uhf (SO-239) connectors have split-ring retainers. These have been found in surplus stores, and seem to be salvaged from imported CB equipment. If you wish to use these, remove the retainer and center pin. Ream out the dielectric just enough so that the pin can be reinserted from the back. Since the center pin is soldered to the 5/16-in. tubing, this provides a rigid method of construction.

A third method of assembly uses a banana plug and jack along the center conductor. By using this method of construction, any type of coaxial connector can be employed. The banana plug should be installed near one of the coaxial connectors, and you should try to use one with a diameter of approximately 5/16 in. to avoid an impedance bump in the line. I tried this method, and at first considered it inferior because it contained an unsoldered joint. Upon afterthought, I realized that *every* coaxial connector relies on an unsoldered joint, so this method is not such a bad one after all!

Probe and Pickup Loop Assembly

The diode and pickup loop (L1) termination are located inside a nonpolarized plug that allows forward and reflected power readings by merely reversing it (Fig. 2). My first probe assembly was housed in a twin-contact uhf plug and jack set, and worked perfectly. Externally, these plugs and jacks look like standard uhf fittings, but have two small-diameter pins inside. Amphenol still manufactures them, but they must be specially ordered. Because they are difficult to purchase, I substituted a four-conductor microphone connector set — and it also works fine. As purchased, the plug is polarized. To allow reversal, remove the plastic insert and file a groove opposite the existing one. It is wise to use a plug that secures the metal shield to the plastic with a setscrew, avoiding rotation of the shield after assembly. The socket half should be soldered onto a coaxial hood (UG-106, UG-372, or UG-177) that has had the flared portion sawed off.

I constructed the pickup loop (L1) by stretching and flattening out a piece of RG-59/U shield braid and filling it with solder (Fig. 3). This makes a loop that is just a bit wider than necessary, but easy to file down for calibration purposes. The loop is soldered between pins 1 and 3 of the socket, while pins 2 and 4 are soldered to ground (Fig. 4). The plug must be wired with sufficient care so that all the components will fit inside the shield. Miniature capacitors are necessary, and 1/4-W resistors easily fit, although there is sufficient room for 1/2-W types. Pin 2 of the plug must be removed to provide space for the 1-kΩ resistor.

A panel-mount phono jack designed for a 1/4-in. mounting hole is used as J3, facilitating connection of the shielded cable to the meter. All the components are first soldered to the plug, and an insulated wire from C2 is left to connect to the phono jack. The shield is installed and the wire is soldered to J3. Tighten the microphone-

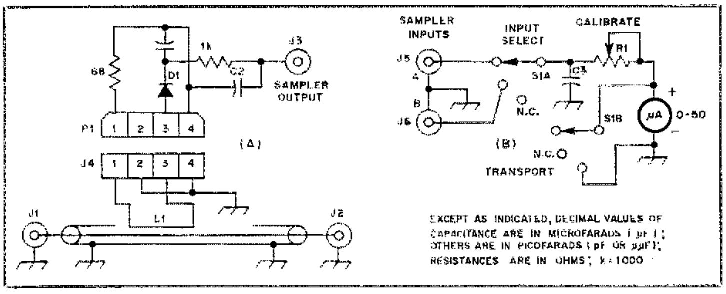

Fig. 4 — Schematic diagram of the complete wattmeter. As wired, the meter will show ***reflected*** power when the load is at J2.

C1 — 200-500 pF miniature disc-ceramic or Mylar® capacitor.
C2 — 500-1000 pF miniature disc-ceramic or Mylar® capacitor.
C3 — 0.005-μF disc-ceramic capacitor.
D1 — 1N34, 1N60, 1N270 or equiv. germanium diode.
J1-3, J5, J6 — Phono jack (Radio Shack 274-346).
J4 — Four-pin, chassis-mount, microphone socket (Radio Shack 274-002).
L1 — Pickup loop (see text).
P1 — Four-pin microphone plug with setscrew.
R1 — 50-kΩ, linear-taper potentiometer.

plug cable clamp around the threads of the phono jack for a perfect fit.

Calibration

Install the pickup-loop assembly into the T fitting, oriented so that it is parallel to the copper tube. Connect a 50-Ω dummy load to J2 and a vhf transmitter to J1. Orient the microphone plug to read *reflected* power. Set R3 to mid-position, and apply between 10 and 25 W of drive. Note the reflected power reading, and remove the pickup-loop assembly from the T. File the pickup loop a bit narrower, and reassemble. Check the reflected power again, and continue this procedure until the meter reads zero. This operation matches the impedance of L1 to that of the terminating resistor R1.

The easiest way to calibrate for power level is to place a wattmeter of known accuracy in series with your homemade unit. Apply power and adjust R3 until both meters coincide. R3 may then be replaced by a fixed-value resistor. You may calibrate your entire meter scale in this manner. A 2-meter multimode rig is ideal for this operation, since most usually have continuously variable output power.

My meter scale was made photographically, but you can produce one by any method you prefer. I chose 5 W for full-scale deflection, as this permits me to do all my antenna work with a portable 2-meter transceiver. It is a simple matter to increase the meter power range by inserting a spacer between the probe assembly and the T fitting. To determine the proper spacing, thread studs into the 4-40 holes and place nuts on them. Use the nuts to change the spacing until the desired meter power range is reached. Replace the studs and nuts with a permanent spacer. Since voltage at the diode is constant, meter calibration should remain the same, regardless of the power level applied. Just remember that the power reading of this wattmeter will only be correct for the band in which it was originally calibrated, although the linearity will be correct. To change bands, a scale multiplier can be used, or a spacer can be inserted to decrease the sensitivity of the instrument to the next higher calibrated range. The meter can also be used on the hf bands, but sensitivity decreases so that the minimum full scale reading will be considerably greater than 5 W.

Final Notes

The remote meter assembly includes a switch that should be used to short out the meter during transportation. I have also provided switching for the use of two different inputs. By constructing two samplers, you can use both of them simultaneously in remote locations, and switch between them at the meter. When both samplers are used in the forward mode, you can, for example, easily determine transmission-line loss.

The engraved plastic label was obtained because I can never remember the formula for SWR! One can be purchased for a few dollars from an engraver. This adds a final, professional touch to a project that I'm sure you will enjoy. Also, my thanks to Roy Mather, K7DFV, who took the photographs for this article.

Notes

[1]T. McMullen, "The Line Sampler," *QST*, April 1972, p. 21.
[2]mm = in. × 25.4.

Colin Lamb, K7FM, was first licensed as K7GYF at the age of 13. By profession, he is a lawyer and holds BA and JD degrees from Willamette University. Colin serves as president of the Chehalem Valley Amateur Radio Club, and maintains a homebuilt, wind-powdered fm repeater. The rest of Colin's station is also homebuilt and powered by two additional wind generators.

New Books

DISKGUIDE™ SERIES

published by Osborne/McGraw-Hill, Berkeley, CA. First edition, 1983. Softbound volumes, 5¼ × 5¼ inches, 32 pages.

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For user convenience, each DiskGuide is organized by function, and is designed for compact storage and easy computer-side use. The aim of this series of booklets is to save a computer user from lengthy searches through a number of different reference manuals by providing all the essential information in one volume. If you can't find the guides at your favorite book, software or computer store, you can order them directly from Osborne/McGraw-Hill, 2600 Tenth St., Berkeley, CA 94710.

Apple II DiskGuide by Zelda Gifford, $7.95; *IBM PC DiskGuide* by David A. Wilson, $8.95; *CP/M DiskGuide* by Curtis Ingraham, $8.95; *VisiCalc DiskGuide* by David A. Wilson, $6.95; *ATARI 400/800 DiskGuide* by John Taylor, $7.95. — *Paul K. Pagel, N1FB*

A Tunable CW Filter

Cascaded band-pass filters with voltage controlled center frequencies yield a cw filter with high selectivity and low ringing.

Richard A. Nelson,* WBØIKN

The benefits of an outboard audio filter in improving cw reception have been documented during recent years, and several designs have been examined in the literature.[1] Unfortunately, most of these filters suffer from at least one of two drawbacks: ringing and a fixed center frequency. Among the simple band-pass filter designs, the state-variable filter (also known as the biquad filter) is particularly useful because of simple construction, ease of tuning and stability (even at high Q).[2] High values of Q result in a tendency toward ringing, thus limiting the maximum usable selectivity of the filter. Ringing may be minimized by connecting low-Q filter sections in series. But to tune the filter requires the use of ganged, closely matched potentiometers.

I built a filter that utilizes a pair of two-pole, state-variable sections with voltage-controlled center frequencies. This design permits tuning with a single potentiometer while providing four filter poles for high selectivity with reduced ringing. An audio amplifier stage allows the filter to be used as an outboard accessory to an existing receiver, or as a complete audio section for a home-built receiver.

At the heart of this filter is the National Semiconductor LM13600 dual operational transconductance amplifier (OTA). OTAs are a specialized family of op amps that exhibit a transconductance (conductance being the inverse of resistance) that is programmable, generally by means of an external bias current. The OTA has a current-controlled gain stage that may be incorporated into a variety of useful circuits. Current-controlled amplifiers, multipliers, multiplexers, oscillators and a variety of filter designs are easily realized through the use of OTAs.

The LM13600 (Fig. 1) contains a pair of identical OTAs, each with an associated Darlington buffer stage and linearizing diodes at the input. The linearizing diodes, (not used in this design), are included to compensate for the logarithmic characteristics of the OTAs, enabling them to pass larger signals without distortion. With the output buffers, they represent a significant improvement over earlier OTA ICs.

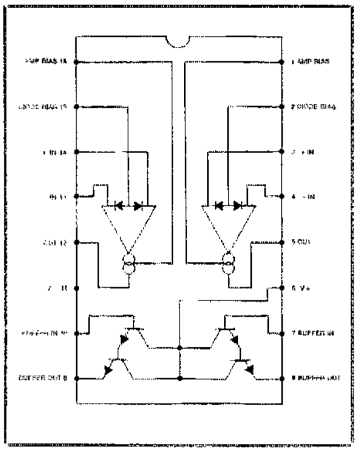

Fig. 1 — Pinout and internal block diagram of the LM13600. Each IC contains two identical sections, each with an operational transconductance amplifier stage and its associated buffer.

The tracking linearity of the LM13600 gain stages is accurate over a range of several decades. This allows two or more LM13600 gain stages to be configured as current-controlled integrators in a series of cascaded, state-variable filters. The filters, therefore, will track within close tolerance over a wide frequency range in response to a control current connected to each OTA bias input pin. The result is a tunable filter with high selectivity and low ringing.

Circuit Description

Refer to the block diagram (Fig. 2). You will notice that the current-controlled filter circuit consists of two major sections: a signal path and a control-current path. The audio signal is fed through the two series connected state-variable filter sections, while the control current pins are connected in parallel to a common variable current source. Although two filter sections are used in this design, any number of similarly connected filters may be used, limited only by the signal-to-noise characteristics of the circuit.

With this design it is possible to obtain two bandwidths simply by tapping the circuit at the output of each filter section. The output from the selected tap is fed to an audio power-amplifier stage that is capable of driving a loudspeaker or headphones. A bypass switch has been included to allow the operator to remove the circuit from the audio line when desired.

The schematic diagram (Fig. 3) shows the filter and amplifier circuits. The filter design is adapted from information found in the LM13600 data sheet.[3] Varying the bias current applied to the transconductance control pin changes the integrator time constant, shifting the band-pass center frequency. Since the frequency-control potentiometer will be supplying a variable voltage, it is necessary to convert this voltage to a variable current capable of biasing the transconductance stages. This is achieved by a voltage-to-current converter, R5, which is connected directly to the wiper of R7. Connect all control-current pins in parallel and route them to R5, which is located off the pc board. Note that the value of R5 will have to be scaled to maintain the same frequency range if more sections are added. The frequency range with the values shown is 350 Hz to 2800 Hz.

The Q of each filter section is determined by R8 (or R9). Larger values at R8 (or R9) will yield higher Q, but the gain at resonance will also increase. To maintain unity stage gain you must increase the value of R1 (R3) or decrease the value of R2 (R4), or both. I found that a Q of about 5 provides the best compromise between selectivity and ringing. R8 and R9 could be replaced by a ganged potentiometer to provide variable Q if desired. I made the selectivity switchable by means of a panel-mounted switch (S1). This switch taps the output of the first or second filter section and provides bandwidths of approximately 200 Hz and 120 Hz, respectively (−3 dB with a center frequency of 1000 Hz), with the values shown.

The audio signal from the selectivity switch is fed through volume control poten-

[1]Notes appear on page 16.

*3640 Juanita Rd., Fort Collins, CO 80524 .

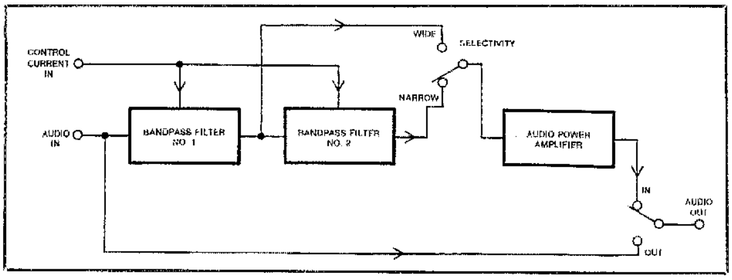

Fig. 2 — Block diagram of the tunable cw filter.

tiometer R10 to the audio amplifier. The amplifier uses the LM386 IC, chosen for its low external parts requirement and low idling current. It can supply up to 600 mW of audio — enough to drive most loudspeakers to a reasonable volume level. The amplifier audio output is connected to a spdt switch on the front panel (S2) that selects the filter circuit or the audio signal from the receiver.

Construction

A pc-board etching pattern (Fig. 4) is provided to speed construction and minimize wiring errors. Notice that the pc pattern is divided into two functional areas. One side of the pattern contains two state-variable filter sections, while the other side of the pattern contains the audio amplifier section. This allows the filter circuit or the amplifier to be used alone or incorporated into other designs, as well as permitting the use of an additional board to provide two more filter sections. The pattern fits nicely on a 3- × 4-inch pc blank. However, you may increase the width to 3.6 inches to conform to the standard suggested by Grabowski.[4,5]

Assembly of the board is straightforward, and with the exception of four jumpers on the foil side, all parts are top-mounted. These jumpers are used to connect the transconductance control pins on each board in parallel, and to route the output of one filter into the input of the next. Regardless of how many filter sections you use, remember to parallel all of the control-current pins, and connect them to the control-voltage source through a voltage-to-current converting resistor.

To assure proper matching of the filter sections, use only high-quality components. Use 5%-tolerance or better resistors and capacitors, if available, and avoid parts of unknown reliability. In particular, be sure the capacitors in the integrators (C1 through C4) are high-quality mica or polystyrene types and are as closely matched as possible.

A parts-placement diagram is shown in Fig. 5; Fig. 6 shows the locations of the foil-side jumpers. Use of IC sockets is recommended; it prevents damage during soldering and greatly speeds trouble-shooting. Be sure to double-check IC orientation when inserting them into the sockets.

The printed-circuit assembly should be mounted in a shielded enclosure to avoid rf pickup during transmit. All leads entering the enclosure should be bypassed as shown in the schematic diagram. If you plan to include a power supply, be sure to

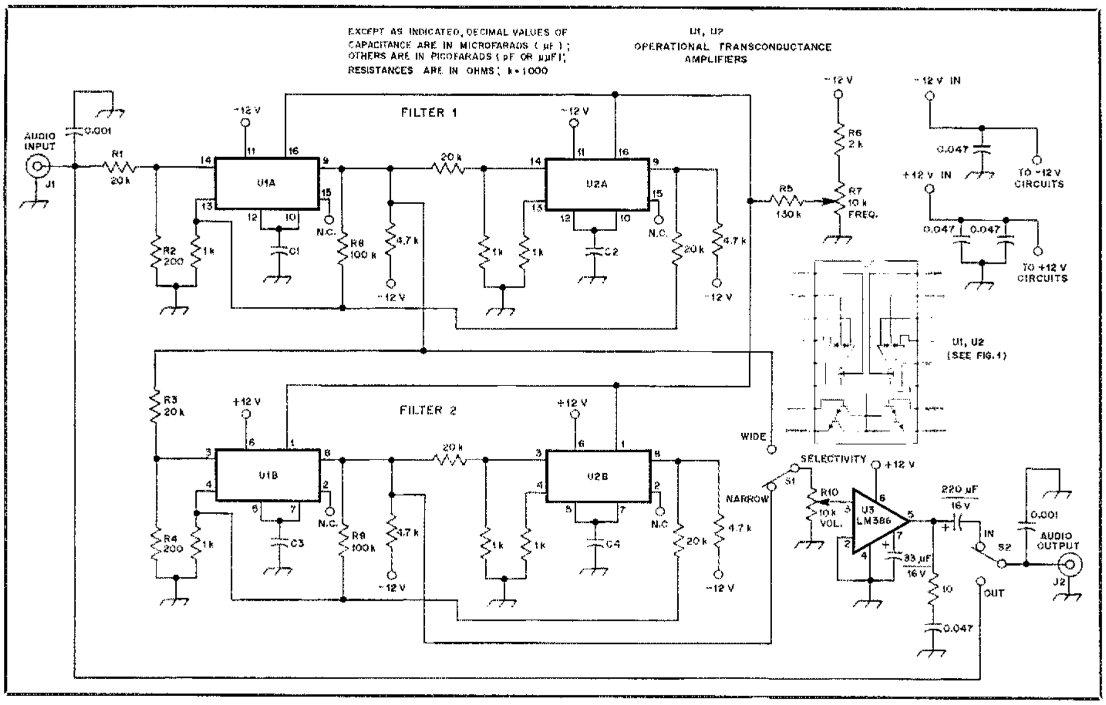

Fig. 3 — Schematic diagram of the tunable cw filter. All fixed-value resistors are 5%-tolerance types.

C1-C4 — 0.001-μ F, 5%-tolerance polystyrene or mica capacitor.

R7 — Linear-taper potentiometer.

R10 — Audio-taper potentiometer.

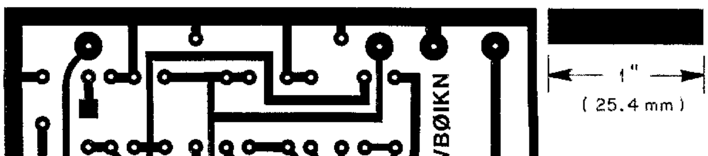

Fig. 4 — Circuit-board etching pattern for the tunable cw filter. Black represents copper. The pattern is shown actual size from the foil side of the circuit board.

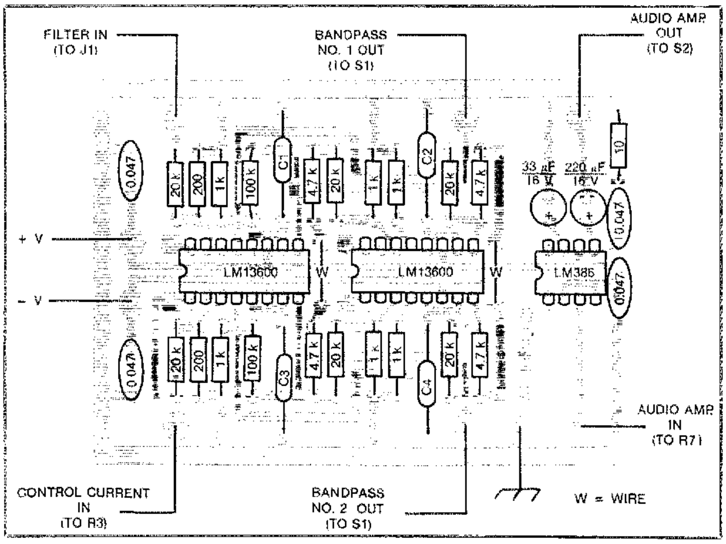

Fig. 5 — Parts-placement guide for the tunable cw filter. Parts are placed on the nonfoil side of the board; the shaded area represents an X-ray view of the copper pattern.

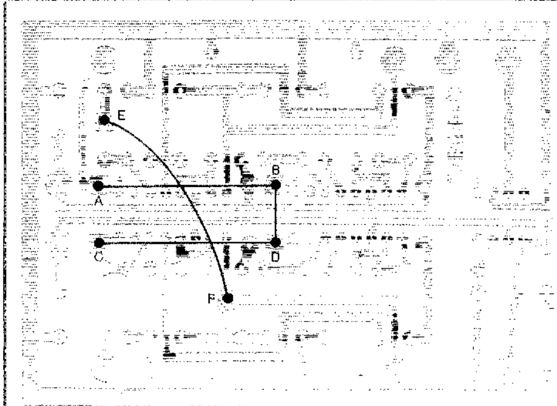

Fig. 6 — Jumper-placement diagram for the tunable cw filter. Jumpers are placed on the foil side of the board; the shaded area represents the copper pattern.

allow sufficient room and provide for air circulation: The transconductance stages will drift with temperature changes.

The layout and dress of interconnections is not critical, although some basic precautions should be observed. Be sure to keep all inputs and outputs as far apart as possible, particularly those associated with the amplifier circuit. Pay special attention to the possibility of ground loops. It is advisable to use a "star ground" technique wherever possible, returning all ground connections to a common point on the chassis. This will help to keep hum from creeping into the audio.[6]

Operation

To install the filter, connect the input jack to the loudspeaker or headphone jack on your receiver using a patch cord. Connect the filter to a regulated power supply capable of providing +12 V at 250 mA and −12 V at 100 mA (I use a pair of rechargeable batteries to eliminate any chance of hum in the audio amplifier). Then connect a loudspeaker or headphones to the output jack and apply power to the filter. Flip the IN/OUT switch to OUT (direct audio) and tune in a cw signal on your receiver. Now flip the IN/OUT switch to IN and the SELECTIVITY switch to wide and tune the filter frequency control until the desired signal is peaked. Adjust the VOLUME control to equalize the direct and filtered amplitudes, and then try the NARROW selectivity position. You will notice that off-center signals are reduced significantly, with no increase in ringing on the desired signal. You may now peak any signal within the receiver passband without changing the receive frequency, simply by tuning the audio filter.

This filter has brought new life to my aging Drake R4B receiver. Not only does it provide a significant improvement in selectivity, but it removes hiss (wideband noise) and hum from the audio (much needed). Whether your receiver is a vintage tube-type, or is state-of-the-art, this filter will help you dig out the weak ones.

First licensed in 1969 as WB2IQF, Richard A. Nelson joined the ARRL that same year. Richard has worked as an audio recording engineer, and as chief engineer and station manager of KCMK-FM, Glenwood Springs, Colorado. He is the founder of Analog Technology, Fort Collins, Colorado, manufacturers of Amateur Radio and professional audio products. Currently, Richard is majoring in engineering physics at Colorado State University in Fort Collins. His Amateur Radio interests include receive and transmit signal processing, and contesting. Other hobby interests include microcomputers, sports cars, building and playing electronic music synthesizers, and collecting gemstones.

Notes

[1]Bloom, "Active Filters," *QST*, July, 1980, p. 17.
[2]Berlin, "The State Variable Filter," *QST*, April 1978, p. 14.
[3]"LM13600/LM13600A/LM11600A Dual Operational Transconductance Amplifier with Linearizing Diodes and Buffers," National Semiconductor, July 1978.
[4]Grabowski, "PC Board Standards Can Speed Experiments," *QEX* (ARRL), May 1982.
[5]mm = in. × 25.4
[6]The author will answer your questions about this article; please include a business-size s.a.s.e. when you write.

High-Pass Filters for Receiving Applications

Is your 160-meter hearing impaired by strong local broadcast stations? Here are some filter designs to help keep your receiver from being "crunched."

By John K. Webb,* W1ETC

High-pass filters are often needed for eliminating broadcast-band signals (0.54 to 1.6 MHz) from the input circuits of amateur receivers. Strong signals may come from local stations, or be propagated over long distances at night. Interference and other effects of circuit nonlinearity can result, especially in receivers without preselectors. In an earlier article, I described an interference-reduction system that employs a phase-canceling technique.[1] This system involves an antenna-pattern null that can be directed toward the interfering signal. My null steerer required the addition of a high-pass filter to reject strong broadcast signals.

A 3.4-MHz-cutoff, five-section high-pass filter for the broadband null steerer is shown in Fig. 1. Correspondence about 160-meter operation of the null steering device led me to consider high-pass-filter designs for that band. This is a more stringent filter requirement than that of the 80-meter filter because the separation between the passband and stopband of the 160-meter filter is only 200 kHz.

The five-section, elliptic, 3.4-MHz filter design achieves 60 to 70 dB of attenuation below 1.6 MHz with 1.0 dB of passband ripple. The stopband is shown in Fig. 2, and the passband is shown in Fig. 3. Photographs are from a Hewlett-Packard network analyzer. This work shows that readily available powdered-iron toroidal cores are very suitable for this type of filter, and that adjustment of shunt-branch resonant frequencies assures achieving desired filter characteristics without a complete dependence on component tolerances.

Filter designs using five sections and providing lower ripple and SWR are available, but these provide less attenuation of the bc signals. Attenuation would be about 10 dB less with a passband ripple of 0.1 dB and a SWR of 1.3 (see Technical Correspondence, March 1983 *QST*).

A filter stopband of about 70 dB starting at 1.6 MHz is required for receiver front-end protection against strong broadcast-band signals. If the five-section elliptic filter is scaled from 80 to 160 meters, the 70-dB stopband does not start until 1.0 MHz, and only 30 dB is available at 1.6 MHz. This prompted me to examine more-complex filters for 160-meter operation.

My reference for elliptic filter parameters was "Simplified Modern Filter Design" by Philip R. Geffe, John F. Rider Publications, 1963. Tables for elliptic filters are given in Appendix 4 of that book. These filters provide extremely sharp transitions from the passband to the stopband, as well as resonant frequencies at which the attenuation exceeds the minimum stopband value. They are well-suited for amateur builders because filter properties are assured by measuring and trimming branch resonant frequencies. Tests on these filters using powdered-iron toroidal inductors showed that the predicted performance can

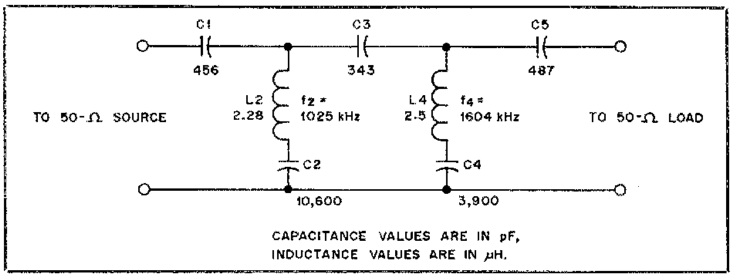

Fig. 1 — Schematic diagram of a five-section, 80-meter filter. L2 = 24 turns no. 26 wire on a T37-2 or T50-6 core. L4 = 25 turns no. 26 wire on a T37-2 or T50-6 core.

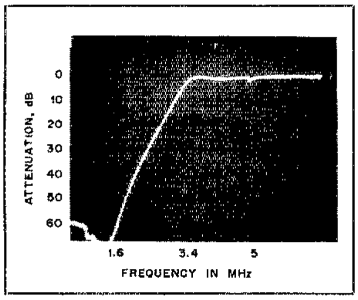

Fig. 2 — Stopband of the five-section filter, as measured on a Hewlett-Packard network analyzer.

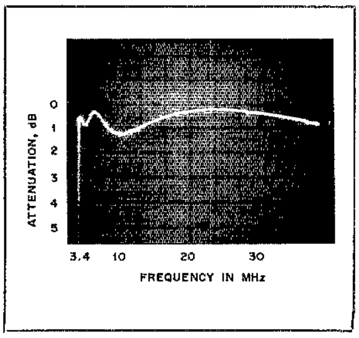

Fig. 3 — Passband of the five-section filter.

[1]Notes appear on page 19.

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Table 1

160-Meter High-Pass Filter Alternatives

From Geffe's Tables, Appendix 4

Table	*Sections*	Ω_s	*Components*	*Calculated Min. Atten., A_s, Below 1.6 MHz*
A4-8	6	1.124	8	32 dB
A4-9	7	1.122	9	45 dB
A4-11	8	1.120	11	52 dB
A4-12	9	1.122	13	65 dB
A4-14	10	1.124	14	73 dB

be approximated in practice, but as more sections are added the deviation from theoretical results will increase. It should be noted, however, that the theoretical filter performance for these or other filter designs cannot be guaranteed simply by buying components and assembling the filter. At the very least, it is important to select actual component values that are as close as possible to the theoretical values. To optimize filter performance, more sophisticated measuring techniques may be required. You should also be aware that there can be some variation from the published inductance-versus-turns values for a given batch of iron-powder material used to make the toroidal cores. A few turns more or less may be required on the core to achieve the desired inductance.

Table 1 lists the number of sections, the parts count and the calculated minimum attenuation below 1.6 MHz for several filters, all with 0.18 dB of passband ripple, an SWR of 1.5 and a cutoff frequency of 1.8 MHz. The normalized frequency (Ω_s) at the start of the stopband is equal to 1.8/1.6 = 1.125. Geffe's designs having Ω_s values closest to 1.125 are listed. Geffe does not provide prototype tables for the more complex filters with larger ripple values; however, as the complexity of the filter grows, component-tolerance variations will moderately increase the ripple from the nominal value.

Table 2 lists calculated inductance and capacitance values for seven- and nine-section filters. It includes the number of turns of wire for Micrometals powdered-iron toroidal cores.[2] Inductance turns were rounded up or down to the nearest integer. I chose toroids because of their self-shielding properties and good Q with small size. The wire size is not critical, but should allow neat, single-layer windings. No. 30 wire is suitable for the size 37 core while No. 26 wire fits well on the size 50 core. One inductor required 75 turns. I used a size 68 core with No. 30 wire. Winding the coils is not difficult, but good lighting and vision aid quality workmanship. Capacitors should be selected to resonate with the companion inductance at the frequency specified for each resonant branch. Filter characteristics are best assured by using measured capacitors and inductors. A small

Table 2

Parameters for Seven- and Nine-Section High-Pass Filters

f_{co} = 1.80 MHz; Ω_s = 1.122; R = 50Ω

Figure	*4*	*4*		*5*
A_s (dB)	*45*	*65*		*65*
Number of Sections	*7*	*9*		*9*
C1	1530 pF	1433 pF	L1	3.58 µH 27 Turns (no. 26 wire)
C2†	7950 pF	13041 pF	L2	32.6 µH 75 Turns (no. 30 wire on 68-2 core)
L2	3.8 µH 31 Turns	3.50 µH 30 Turns (no. 28 wire)	C2†	1400 pF
F2	918 kHz	745 kHz	F2	745 kHz
C3	1230 pF	1090 pF	L3	2.72 µH 24 Turns (no. 26 wire)
C4†	1480 pF	2097 pF	L4	5.24 µH 33 Turns (no. 26 wire)
L4	6.8 µH 41 Turns	5.38 µH 37 Turns (no. 28 wire)	C4†	2152 pF
F4	1587 kHz	1498 kHz	F4	1498 kHz
C5	1490 pF	1532 pF	L5	3.83 µH 28 Turns (no. 26 wire)
C6†	2110 pF	1531 pF	L6	3.83 µH 28 Turns (no. 26 wire)
L6	6.0 µH 39 Turns	6.52 µH 40 Turns (no. 30 wire)	C6†	2606 pF
F6	1418 kHz	1593 kHz	F6	1593 kHz
C7	2270 pF	1355 pF	L7	3.39 µH 26 Turns (no. 26 wire)
C8†	U	3430 pF	L8	8.58 µH 42 Turns (no. 26 wire)
L8	N U	4.75 µH 34 Turns (no. 28 wire)	C8†	1898 pF
F8	S E	1247 kHz	F8	1247 kHz
C9	D	1860 pF	L9	4.64 µH 31 Turns (no. 26 wire)
Cores	50-6	50-6		50-2 except L2

†resonate with L at f

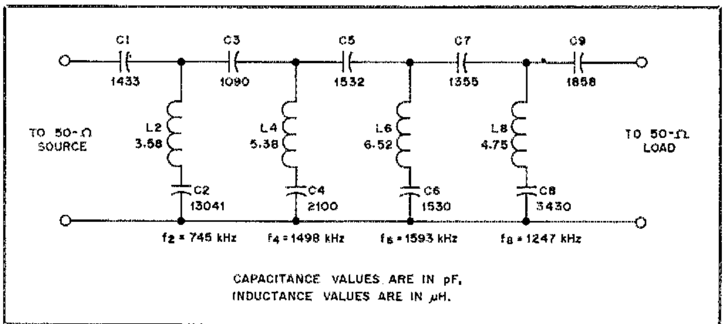

Fig. 4 — Schematic diagram of a nine-section 160-meter filter using four inductors.

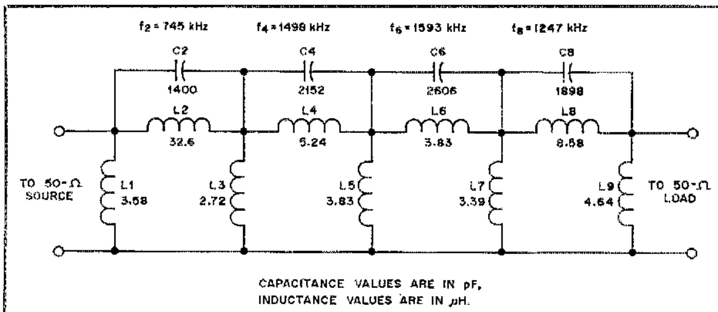

Fig. 5 — Schematic diagram of a nine-section 160-meter filter using nine inductors.

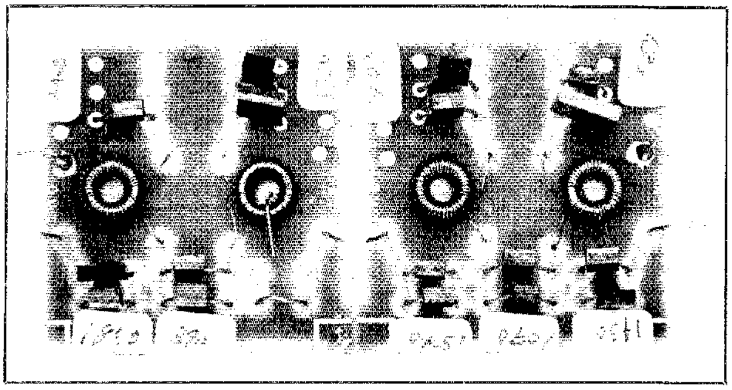

Fig. 6 — The nine-section, 160-meter filter with four inductors. The ground foils on the two boards are connected by means of copper strips soldered across the foils on the underside. This provides mechanical strength as well as good electrical contact.

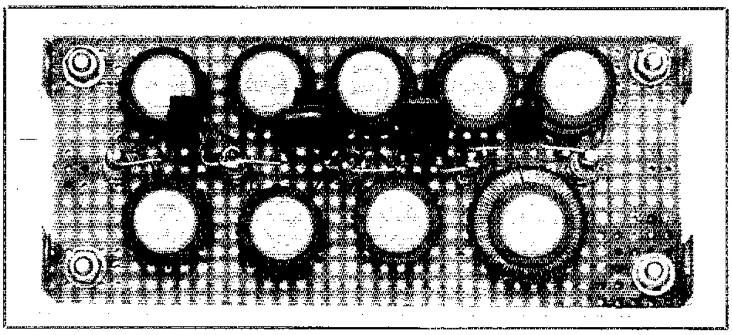

Fig. 7 — The nine-section, 160-meter filter with nine inductors.

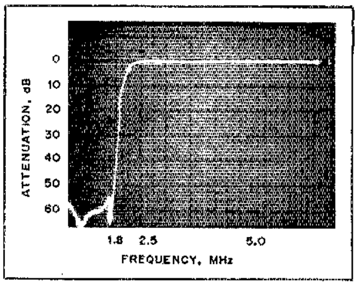

Fig. 8 — Stopband spectral display of the nine-section, 160-meter filter using four inductors.

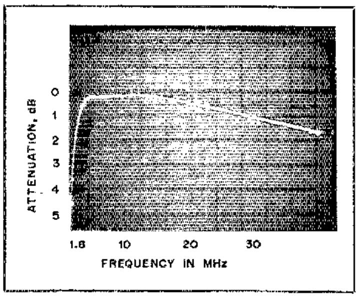

Fig. 9 — Passband spectral display of the nine-section, 160-meter filter using four inductors.

degree of inductance adjustment can be effected by compressing or spreading turns on a core. However, I did not use this fine-adjustment technique with the filters I built. These filters use paralleled standard-value NP0 ceramic capacitors selected to within 1% of the target values.

Two nine-section filters are listed in Table 2. They are "duals," which means they have an alternate configuration that theoretically has the same attenuation response. The first, shown in Fig. 4, uses nine capacitors and four inductors; the second, in Fig. 5, uses nine inductors and four capacitors. The second configuration was built after the first because I found that the accuracy of the calculated toroid inductance values was very good. When the target value required a fractional turn, the measured inductance value was below or above the target value, depending on whether the number of turns was rounded up or down. Each prototype filter has a theoretical passband ripple (A_p) of 0.18 dB and a stopband attenuation (A_s) of 65 dB. The stopband frequency (f_s) is 1604 kHz, and the cutoff frequency (f_{co}) is 1800 kHz. Each filter has a rejection notch just below f_s, and there is a different rejection-notch notch frequency for each shunt branch.

The nine-section filter in Fig. 4 was constructed on two circuit boards, each designed for a five-section filter with three parts locations for series capacitors and two for capacitors in the shunt branches. This filter is illustrated in Fig. 6. I used T50-6 cores for the inductors in this filter. T37-2 cores are specified to have the same A_L value, so they should work with the same number of turns. The second nine-section filter (Fig. 5) was built on a perforated circuit board with standoff terminals. The construction is shown in Fig. 7.

The passband and stopband data for the filter of Fig. 4, as measured on a Hewlett-Packard network analyzer, are shown in Figs. 8 and 9. The filter shown in Fig. 5 provides about 10 dB less stopband attenuation and about the same passband attenuation at 30 MHz. Both filters have a more gradual cutoff than expected, probably a result of the Q of the inductors. Attenuation at 1.8 MHz is about 3 dB, and a passband ripple of 0.18 dB is not realized below about 2.0 MHz.

The filter of Figs. 4 and 6 provides the better stopband performance of the two versions. The ease of building the two types of filters is about the same; one requires more inductors, while the other requires more capacitors. Either filter has its characteristics determined largely by resonant frequencies of L-C branches, which may be measured and adjusted.

I would like to thank ARRL TA Ed Wetherhold, W3NQN, for his correspondence and suggestions. This led to further review of the subject and construction of the nine-pole, 160-meter filters.

Notes

[1]J. Webb, "Electrical Antenna Null Steering," *QST*, Oct. 1982, pp. 28-32.

[2]Micrometals toroidal cores are available from Amidon Associations, 12033 Otsego St., North Hollywood, CA 91607, and from Palomar Engineers, 1924-F West Mission Rd, Escondido, CA 92025.

References

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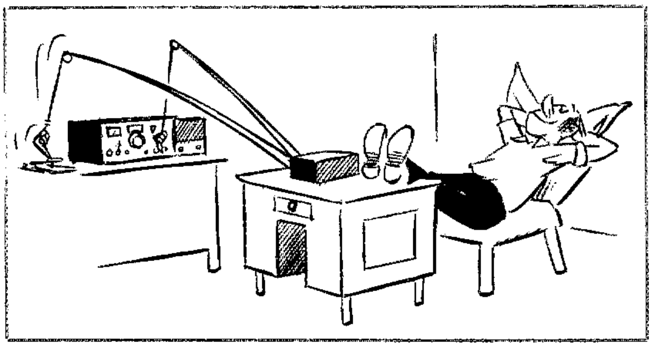
The Would-Be Contest Killer



Use your "micro" to gain an edge over the hotshot contesters.

By Jerry Hess,* W9KTP

Until 1982, the only contest I had entered was Field Day. Contests didn't interest me; when I encountered one, I changed frequency or went QRT. During the 1981 Field Day one of our contest pros issued a challenge: "You oughta try some contestin', kid."

That challenge was still echoing in my head a few weeks later, so I decided to give contesting a try. Working DX was lots of fun, particularly using cw, so I decided to take a shot at the ARRL International DX competition. Boy, what a shock! There are *expert* operators out there! A comparison of past scores with my effort was a real ego deflater. Those operators with thousands of QSOs must have four arms and two heads!

After a good sulk, I formulated a plan for the upcoming CQ World Wide DX contest. Why not outsmart those hotshots by

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Table 1

BASIC Program Listing

```
1 DIM A (50), B (50)
2 DIM C$(100), E$(50)
3 Q = 0
4 TI$ = "000000"
5 DATA 6, 204, 5, 136, 6, 64, 5, 144
6 DATA 5, 248, 5, 120, 5, 56, 5, 24
7 DATA 5, 8, 5, 0, 5, 128, 5, 192
8 DATA 5, 224, 5, 240, 1, 0, 1, 0
9 DATA 1, 0, 1, 0, 1, 0, 6, 48
10 DATA 1, 0, 2, 64, 4, 128, 4, 160
11 DATA 3, 128, 1, 0, 4, 32, 3, 192
12 DATA 4, 0, 4, 0, 4, 112, 3, 160
13 DATA 4, 64, 4, 192, 4, 128, 5, 224
14 DATA 4, 96, 4, 208, 3, 64, 3, 0
15 DATA 1, 128, 3, 32, 4, 16, 3, 96
16 DATA 4, 144, 4, 176, 4, 192, 1, 0
17 DATA 1, 0, 1, 0
20 FOR I=1 TO 50
21 READ A(I), B(I)
22 NEXT I
50 PRINT:CLS
51 PRINT
52 PRINT "                OPTIONS"
53 PRINT
54 PRINT " E=ENTER CALL           D=DELETE CALL"
55 PRINT " C=CALL HIM             S=SEND EXCHG"
56 PRINT " A=CALL CQ              Q=QSL HIM"
57 PRINT " R=REPEAT               Z=QRZ"
58 PRINT " L=LOG HIM              K=KEYBOARD"
59 PRINT " P=PRINT                X=CODE PRAC"
60 PRINT " N=STO QSO              B=SET CLOCK"
61 PRINT
62 PRINT
63 PRINT "              WHICH OPTION=";
65 INPUT A$
66 PRINT
70 IF A$="E" THEN 100
71 IF A$="D" THEN 110
72 IF A$="A" THEN 120
73 IF A$="C" THEN 130
74 IF A$="K" THEN 135
75 IF A$="Z" THEN 145
76 IF A$="Q" THEN 150
77 IF A$="S" THEN 155
78 IF A$="L" THEN 160
79 IF A$="R" THEN 165
80 IF A$="P" THEN 170
81 IF A$="X" THEN 180
82 IF A$="N" THEN 190
83 IF A$="B" THEN 195
99 GOTO 50
100 PRINT
101 PRINT
102 PRINT "HIS CALL-10 CHARS";
103 INPUT B$
104 GOTO 50
110 B$=""
111 GOTO 50
120 C$="CQ CQ CQ DE W9KTP/3 W9KTP/3 K"
121 M=LEN(C$)
122 FOR N=1 TO M
123 D$=MID$(C$,N,N)
124 GOSUB 1000
125 D$=""
126 NEXT N
127 GOTO 50
130 C$=B$+" DE W9KTP/3 K"
131 GOTO 121
135 Y=INP(8)
136 IF Y=164 THEN 50
137 IF Y=32 THEN Q$=" "
138 IF Y > 221 THEN 135
139 IF Y < 172 THEN 135
140 C$=Q$+CHR$(Y-128)
141 GOSUB 6000
142 GOTO 135
145 C$="QRZ DE W9KTP/3 K"
146 GOTO 121
150 C$="BK QSL 5 NN 5 NN 05 05 BK"
151 GOTO 121
155 C$=B$+" DE W9KTP/3 5NN 5NN 05 05 BK"
156 GOTO 121
160 C$="QSL 73 DE W9KTP/3 QRZ K"
161 GOTO 121
165 C$="? BK"
166 GOTO 121
170 PRINT " SIGNAL REPORT ";
171 INPUT H$
172 PRINT "ZONE";
173 INPUT J$
174 E$=TI$+" "+B$+" "+H$+" "+J$+" "
175 GOSUB 5000
176 GOTO 50
180 T=INT(100*RND(0))
181 IF T < 44 THEN 180
182 IF T > 90 THEN 180
183 IF T > 64 THEN 185
184 IF T > 57 THEN 180
185 D$=CHR$(T)
186 GOSUB 1000
187 GOTO 180
190 C$="TNX FER RPT-UR RST 579 ES QTH JOHNSTOWN, PA-NAME
     JERRY-HW?"
191 GOTO 121
195 GOSUB 7000
196 GOTO 50
1000 L=ASC(D$)
1001 GOSUB 4000
1002 GOSUB 4000
1003 P=128
1004 R=L-43
1005 IF R < 1.0 THEN 1018
1006 C=A(R)
1007 D=B(R)
1008 FOR J=1 TO C
1009 T=INT(D/P)
1010 IF T < 1.0 THEN 1012
1011 D=D-T*P
1012 P=P/2
1013 IF T < 1.0 THEN 1016
1014 GOSUB 3000
1015 GOTO 1017
1016 GOSUB 2000
1017 NEXT J
1018 PRINT D$;
1019 RETURN
2000 OUT 3, 1
2001 GOSUB 4000
2002 OUT 3, 0
2003 GOSUB 4000
2004 RETURN
3000 OUT 3, 1
3001 GOSUB 4000
3002 GOSUB 4000
3003 GOSUB 4000
3004 GOSUB 4000
3005 OUT 3, 0
3006 GOSUB 4000
3007 RETURN
4000 FOR K=1 TO 3
4001 U=1
4002 NEXT K
4003 RETURN
5000 LPRINT E$
5012 RETURN
6000 M=LEN(C$)
6001 FOR N=1 TO M
6002 D$=MID$(C$,N,N)
6003 GOSUB 1000
6004 D$=""
6005 NEXT N
6006 Q$=""
6007 RETURN
7000 PRINT "INPUT HOURS";
7001 INPUT H$
7002 PRINT "INPUT MINS";
7003 INPUT E$
7004 TI$=H$+E$+"00"
7009 RETURN
```

Note: Program line 135 is a call to the keyboard; line 1014 calls the dash routine, while line 1016 calls the dot routine. Lines 2000, 2002, 3000 and 3005 are machine specific (see text).

Table 2
Assembly-Language Routines

Loc. Data			*Comments*
7172-076	Timer:	LDA 20	; major loop constant
7173-020			
7174-006	Y:	LDB 377	; minor loop constant
7175-377			
7176-000	X:	NOP	
7177-000		NOP	
7200-020		DJNZ X	; dec B until zero
7201-374			
7202-107		LDB, A	; dec A until zero
7203-075		DEC A	
7204-020		DJNZ Y	
7205-366			
7206-311		RTN	
7207-076	Dots:	LDA 2	; set port-3 bit 1 to 1
7210-002			
7211-323		OUT 3	
7212-003			
7213-315		CALL TIMER;	one time interval for dot
7214-172			
7215-007			
7216-076		LDA 0	; reset port
7217-000			

Loc. Data			*Comments*
7220-323		OUT 3	
7221-003			
7222-311		RTN	
7223-076	Dash:	LDA 2	; set port-3 bit 1 to 1
7224-002			
7225-323		OUT 3	
7226-003			
7227-315		CALL TIMER;	3 time intervals for dash
7230-172			
7231-007			
7232-315		CALL TIMER	
7233-172			
7234-007			
7235-315		CALL TIMER	
7236-172			
7237-007			
7240-076		LDA 0	; reset port
7241-000			
7242-323		OUT 3	
7243-003			
7244-311		RTN	

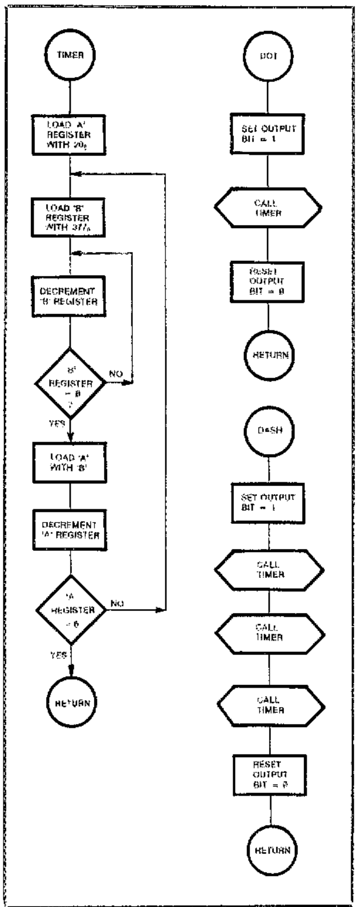


Fig. 1 — Flowchart of the assembly-language cw-timing subroutine.

logging and sending the contest information with my microcomputer? This article describes the result of my efforts. Now that I have the program operating, the only thing holding back my QSO rate is my ability to locate new stations quickly.

Background

My contesting program is written in BASIC, with the cw-timing subroutines written in Z80 assembly language. The BASIC dialect should be adaptable to any microcomputer BASIC, and the machine-language subroutine uses no absolute addresses, so they should be adaptable (relocatable) to any hardware configuration or processor. A parallel I/O port or peripheral interface adapter (PIA) is the only program hardware requirement — it is required for sending cw.

Basically, the program generates several "canned" messages and sends them, using the machine-language subroutine, to one bit of a parallel port. In my version of the program, exchange information for the CQ World Wide DX contest is included. The user can change messages easily as a particular contest dictates. In addition, a subroutine generates random code for practice purposes.

Inside the Program

The BASIC program listing is given in Table 1; refer to it while reading the text. Morse code (cw) elements — dots and dashes — are generated with the use of a *lookup table.* The lookup table is located on lines 5 to 17 of the listing. Dots are represented by "0"s and dashes are represented by "1"s using the binary number system (base 2). There are actually two lookup tables in the program, A and B (line 21). Table A values represent the length of the Morse character in elements, and Table B values represent the decimal (base 10) equivalent of the binary value of the Morse character. For example, the letter "A" has a length of 2 elements — a dot and a dash — therefore, the value in Table A for the letter "A" will be 2. The equivalent binary number for the letter "A" is 0100 0000, or 64 in decimal form.

A simple index for the two tables allows the ASCII value of any character to be used as a pointer for the tables. The table index value is equal to the ASCII value of a character minus 43. For example, the ASCII value for the letter "A" is 65; the data for the letter "A" will be in position 22 in Tables A and B.

Machine-Language Subroutine

BASIC is an interpreted computer language, which means that it is very slow (as far as computing is concerned). The timing routines for my program were originally written in BASIC (as they appear in the listing), which limits the maximum sending speed to approximately 15 wpm. During contests, the cw sending speed varies between 20 and 50 wpm; therefore, I decided to write a machine-language subroutine to generate the dot and dash timing elements. The Z80 assembly listing is given in Table 2. Address and data values are given in octal (base 8); memory locations 7172 through 7244 were used because my machine (a Digital Group unit) had free memory in that area. All of the program is relocatable; find a free space in your computer's memory and go to it! Of course, the number used in the OUT statement will depend on your hardware con-

Table 3
Commands

D-Delete call	Deletes the entered call after transmitting and recording or if you decide to work someone else.
S-Send Exchange	This is used to answer a call to your station. It sends the contest exchange.
A-Call CQ	Calls CQ and signs your call.
Q-QSL Him	Does a break and gives the contest exchange.
R-Repeat	Sends a "?" and "K."
Z-QRZ	Sends your call, "QRZ" and "K."
L-Log Him	Sends "QSL," your call and "QRZ."
K-Keyboard	Allows free-form input from the keyboard. A "$" will return you to the menu.
P-Print	Request signal report and the other station's exchange, and prints the time, call, report and received exchange on the printer.
X-Code Practice	Randomly generates code practice.
N-Standard QSO	Sends a standard QSO message.
B-Set Clock	Allows you to input the current time.

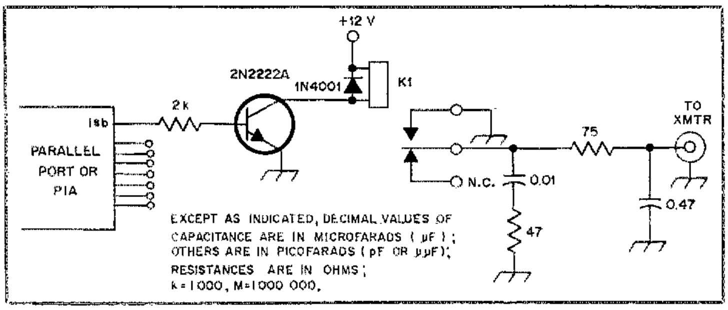


Fig. 2 — Schematic diagram of the interface circuit. The relay has a 12-V coil and 1-A contacts.

figuration. Flowcharts for the machine-language routines are given in Fig. 1. If your computer does not use the Z80 µP, these should aid in designing a subroutine for your machine. Cw sending speed can be adjusted by altering the constant in location 7173. My computer has a clock rate of 2.5 MHz, which yields a speed of approximately 25 wpm with the constant given.

Hardware Requirements

The machine-language program assumes that the parallel port latches to a particular state (0 or 1; "on" or "off") when an OUT statement is executed. Many computers do not have a latched port, but a simple RS flip-flop IC can provide the function. My transceiver-interface circuit is shown in Fig. 2.

Running the Program

Lines 52 through 60 in the program are the "menu"; the program is "menu driven." Each command is executed by typing one letter; after each command has been executed, the menu reappears on the screen. Table 3 describes each command function in detail.

I've learned a lot about operating in the past few months. It will be a long time before I'm able to beat the contest pros — even with a computer! Please drop me a note if you have any comments on the program. Special thanks to my wife who helped with the editing, and to John Rogers, W3IW, for his assistance with DX contests. QST

Strays

THE FOX FIXER

☐ Are you tired of perennial second-place finishes in your club's fox hunts? Like the Boston Red Sox, do you disappoint your fans, family and friends when you snatch defeat from the jaws of victory? If this description fits you, consider building the Fox Fixer, and throw your competitors for a loop!

The brainchild of Dave Baysinger, WBØBAE, the Fox Fixer doubles as a Turkey Tracker. Malicious interference on local repeaters is a serious problem affecting many public-service-minded fm clubs. This loop antenna is the perfect answer to an interference committee looking for effective methods of direction finding.

For the benefit of the uninitiated, a fox (or bunny, in some circles) hunt is a competition or exercise where a member of a ham group is designated "the fox." The fox makes its way to an undisclosed position hidden from the hunters' view. The fox then makes intermittent transmissions for the purpose of providing clues as to whereabouts. The hunters employ various contraptions to be the first to find the fox's lair. The Fox Fixer is one such contraption, and a good one!

Construction

The mast is a piece of steel tubing with a diameter of 5/8 in. and a length of 35 in. The bottom 12 in. are covered with electrical tape to provide a comfortable handle for the antenna. A small hole drilled 1 in. below the top of the mast provides access for the feed line, RG-58/U. The upper terminus of the mast should be slotted and crimped in the manner shown in Fig. 1 to allow attachment of the antenna loop.

The antenna loop consists of two separate pieces of 1/4-in. copper tubing — the first 9 in. long, and the second 9½ in. long.

The loop attaches to the mast as shown in Fig. 1, and is tuned by C1, a 15-20 pF variable capacitor. The antenna is fed with the inner conductor soldered at approximately 1½ in. from the mast on the 9-in. element, and the outer conductor braid is soldered to the mast itself. An 8-ft (2.4-m) length of RG-58/U is sufficient for most applications. A PL-259 connector makes the feed line compatible with most 2-meter rigs. For others, a BNC adaptor may be required.

Operation

In the field, most fox hunters will use a hand-held 2-meter fm rig with an S meter that indicates relative signal strength. With the Fox Fixer connected to the rig, listen for the fox on its transmitting frequency. Note the signal strength level. A 10-70 dB step attenuator should be used.[1] While keeping your eyes on the S meter, slowly rotate the loop to determine signal peaks and nulls. Signal peaks will occur when the fox is in the same plane as the loop. A null will occur when the fox is in a plane perpendicular to the loop.

Choose the direction of maximum signal and make your way toward the fox. The closer you get, of course, the stronger the signal will be, so you may have to step up the attenuation with the use of a 10-70 dB step attenuator to allow the S meter needle to retreat from the "pinned" position. A little practice will make you an expert, and you'll be able to invite your fans, family and friends into your trophy room to show off your collection of *vulpeus fulva!* — *Richard Palm, K1CE*

[1]See "Knock-It-Down and Lock-It-Out Boxes for DF," April 1981 *QST*, p. 41.

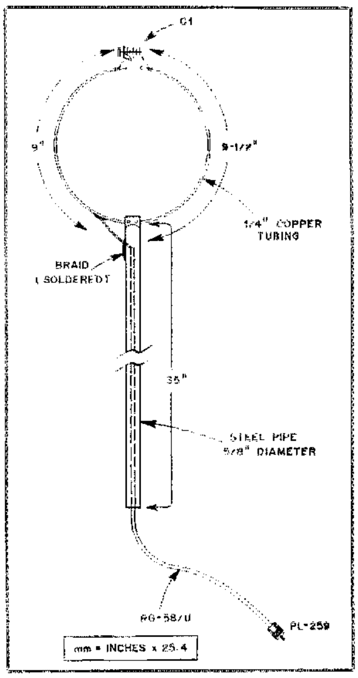
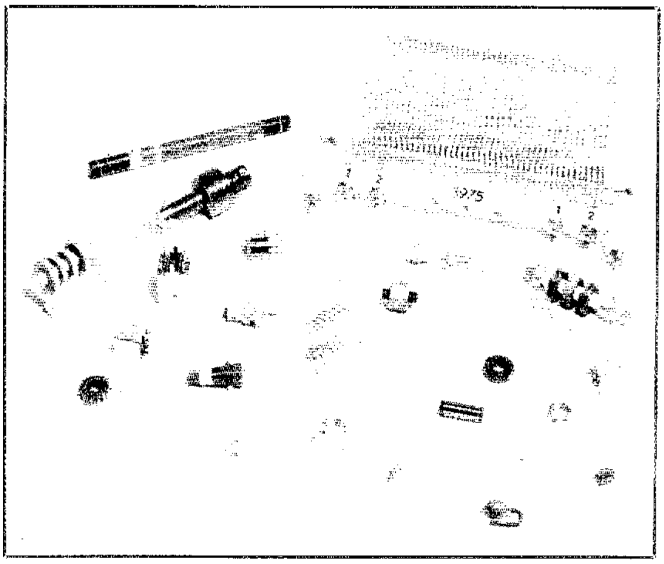


Fig. 1 — Construction of the Fox Fixer.

Understanding Coils and Measuring Their Inductance



Winding your own rf coils or selecting good ones from the surplus market is easy when you understand the basics of inductors. Let's discuss selection and simple methods for measuring inductance in the ham workshop.

By Doug DeMaw,* W1FB

How many times have you bought a surplus or unidentified coil for use in a homemade circuit, only to discover it wouldn't function properly in your project? Or, how about all the head-scratching you've done while perusing bargain inductors at ham radio flea markets? It's not uncommon to see a dozen or more nice-looking slug-tuned inductors or small rf chokes for as little as $1. But, what are the coil characteristics? Is the Q high enough for our proposed circuits? What is the fixed-value inductance, or what might be the tuning range of the slug-adjusted inductor?

Unfortunately, that valuable information can be gotten only by means of measuring equipment. If no measurements are possible in your workshop, you could be buying the oft-mentioned "pig in a poke" when acquiring assorted coils.

I would never discourage anyone from buying bargain components. The cost of any commercial coil of recent production (new, that is) is enough these days to make one consider a less-expensive hobby — such as carving artifacts from wood! The economical alternative to purchasing new, expensive coils is to wind our own, or to acquire them at flea markets or from surplus outlets. The various coils or rf chokes can be graded out at home and labeled with their approximate characteristics for use later on.

We can apply simple techniques in our workshops to learn easily what the coil inductance is, and what the relative Q of the inductor might be. We'll discuss those approaches later. But first, let's talk about coils and chokes in general.

The Fundamental Nature of Coils

The word "coil" is self-explanatory. That is, it is a length of conductor — generally wire — that has been formed into a multiturn coil. Some coils are solenoidal (single-layer winding), while others have two or more layers of wire. These are often referred to as "pi-wound" or "bank-wound" inductors. The layers of wire are laid rather precisely on top of each other by means of machinery. We hams might simulate this winding style by using what is called "scramble winding." The coil Q (figure of merit) may not be quite as high when we scramble-wind, at least with respect to a machine-wound multilayer coil, but it will usually be adequate for our needs.

A thorough treatment of the subject of inductors and inductance can be found in *The Radio Amateur's Handbook* (ARRL), Chapter 2. We will not delve into the theory of inductance here, but shall look at some practical aspects of the subject.

For example, a given length of conductor will have a particular inductance value when stretched out. All conductors have *some* value of inductance (Fig. 1). That same length of wire will exhibit increased inductance when it is formed into a coil. The greater the number of turns, the higher the inductance. We can observe also that the greater the spacing between the coil turns, the lower the inductance. Maximum inductance for a given coil winding will occur when the coil turns are immediately

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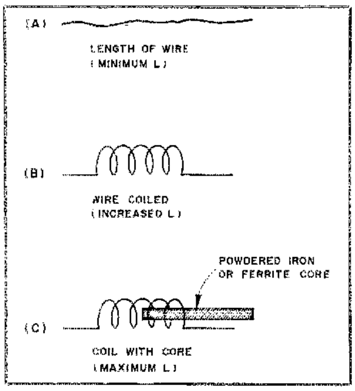

Fig. 1 — Even a straight piece of wire has an inductance value (A). The same wire formed into a coil (B) will have considerably more inductance. When a powdered-iron or ferrite core is inserted into the same coil (C), the inductance increases markedly.

adjacent to one another, or when the windings are stacked atop one another.

The inductance of a coil can be increased further by inserting a magnetic core inside it. Such core materials (suitable for rf work) are ferrite and powdered iron. Some coils contain brass cores or slugs rather than ferrite or iron ones. Brass has the opposite effect on coils: it decreases the coil inductance over that which exists when no core is used. Slug-tuned coils that contain brass cores are used mainly at vhf. The iron or ferrite cores are employed in circuits from vhf down to the low frequencies (broadcast band and lower). We must be aware, however, that the right core material must be used for a particular operating range. This and the general subject of ferrite and iron-core materials are discussed in plain language in the reference cited at note 1.[1]

The Wrong Core Material

I have witnessed a lot of frustration among inexperienced amateurs when they bought slug-tuned coils, blank forms and toroids at flea markets or from surplus dealers. Such vendors rarely provide the electrical characteristics of the coils or the core material. We may have the "gut feeling" that a particular coil looks great for that neat circuit we're putting together, but after installing it (even though the inductance value may be correct), that part of the circuit performs poorly, or not at all! How can this be? Ah, ha! The core material is wrong for the operating frequency!

Let's suppose our circuit is an oscillator and the operating frequency is 7 MHz. Our bargain coil has a core meant for use at, say, 500 kHz. At 7 MHz, the core has spoiled the Q of the inductor. Because of this low Q, our oscillator won't function, or if it does it will operate in a very sluggish manner (hard to start and keep oscillating). The irony is, if I may resort to still another pun, that the coil might have a marvelous Q — perhaps 400 — at 500 kHz, whereas it could drop to as low as 10 at 7 MHz. This phenomenon can occur with ferrite or powdered-iron cores.

The lower the intended operating frequency of the proper core, the greater its permeability. Permeability (μ_e) is the trait that determines the effect of the core material on the inductance. That is, the higher the permeability of the core, the greater the inductance for a specified number of coil turns. This can provide an excellent clue to the relative characteristics of an unknown core. Specifically, if a coil has only a few turns but has a large value of inductance, the core material is probably designed for use at medium or low frequencies. Conversely, if the inductor has many turns but the inductance seems low for a coil that contains a core, the slug is probably ideal for high-frequency use. In fact, it may be suitable into the vhf region. Many cores and toroids have a color coding that indentifies their optimum frequency ranges. The J. W. Miller Co. coil catalog lists the color codes versus frequency (given also in the reference cited in note 1).[2]

The improper core material can be identified readily when it is used in an rf-amplifier stage. If the stage gain seems too low and if the tuning is very broad, it's likely that the Q of the tuned circuit has been degraded seriously by the core in the coil. A sharp-tuning circuit is indicative of high Q (usually desirable).

Other factors that affect the Q of a coil are the wire size, the form factor, the insulating material used in the coil form and the proximity of the completed coil to nearby conductive objects. The larger the cross-sectional area of the wire, the greater the coil Q. This is because the current will flow more easily through the larger conductor (reduced I^2R losses). The form factor is the ratio of the coil diameter to the winding length (Fig. 2). For example, if our coil was 3 inches long and 1½ inches in diameter, the form factor would be 2.[3] I have found during laboratory measurements that a form factor of 1 to 1.5 yields the best unloaded Q for a particular coil, although the Q does not deteriorate seriously with form factors as great as 3. These values of Q are for an "unloaded" condition, or more commonly specified as the Q_u value of a coil. Once the inductor is installed in a circuit, this value of Q becomes rather insignificant, since the other circuit elements change the Q to some lower value through loading effects. The value of greatest interest to us is the "loaded Q," or Q_L. The Q_u value is of importance when designing a circuit, however. For instance, we may design a VFO tuned circuit for a Q_L of 15. The Q_u of the coil should be somewhat higher to ensure that the design Q is realized. I like to select a Q_u that is three times or greater than the required Q_L.

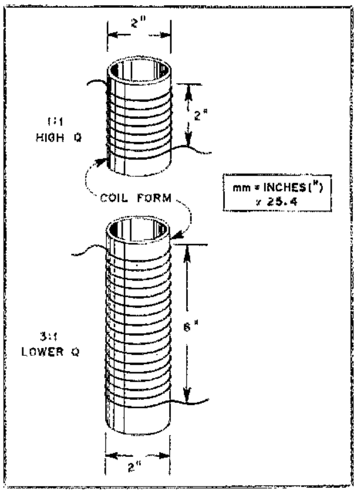

Fig. 2 — Illustration of form factor for coils. The highest Q should result with a factor of 1 (A). As the coil is lengthened (B), the Q will decrease somewhat.

The coil form can affect the Q of the inductor, also. The better the dielectric material (insulating properties), the higher the Q. An air-wound coil will have the best Q, of course, but coils of this type are seldom practical below vhf because they are too floppy and unstable. So, our choice is to use a coil form. Ceramic or steatite forms are very good in terms of Q. Phenolic forms are fine for hf and lower, and there are a variety of modern materials (plastics) that are satisfactory, also. PVC tubing is not recommended in circuits that handle rf power at high impedances. In the presence of high levels of rf voltage, the PVC tubing may heat, break down and melt; likewise with nylon.

Earlier, we mentioned the proximity of nearby conductive objects and their effect on the coil Q. A good rule of thumb is to keep your coil (toroids excepted) at least one coil diameter away from nearby conductors, such as chassis, shield plates, variable capacitors, and the like. Iron will have a significant effect on the coil Q and inductance. Iron will increase the inductance, whereas aluminum and brass will decrease the inductance. Since a toroidal coil is self-shielding in nature, it will not be affected particularly by nearby metal objects and other coils.

Types of Coil Wire

You may have wondered why some coils, especially the smaller ones, have stranded wire that has insulation (silk, cotton or enamel) on each strand. Other coils simply used plain enameled wire, which was a lot easier to solder! Well, the multistrand wire is capable of providing a much higher Q than the single strand winding of enameled

[1]Notes appear on page 26.

wire. The former is known as *Litz wire*. By combining many individual insulated strands of wire to form a single conductor effectively, the I^2R losses are reduced dramatically and the Q increases accordingly. This is because the rf currents flow on the surface of a conductor, for the most part. They do penetrate the conductor to some degree, and the penetration becomes more pronounced as the operating frequency is lowered. The higher the operating frequency, the more troublesome the I^2R losses. This phenomenon is called "skin effect." The use of Litz wire greatly increases the effective conducting surface of the winding (many skins, so to speak, in parallel) and enhances the Q.

When you are shopping for surplus coils, it is a good idea to watch for Litz wire inductors. This wire can be taken from the coil form and used when winding new coils. Litz wire is especially useful in the hf range, where high Q becomes more difficult to realize with small wire. Litz wire (new) is *very* expensive. We can fabricate our own Litz wire by weaving several strands of no. 30 enamel wire together with a hand drill and a vise. Larger gauges of enameled wire can be treated in a like manner for heavier "Litz" conductors. Each strand of wire must be stripped properly at the ends of the coil winding so that they may be soldered in parallel. Otherwise, the good effects of the Litz winding will be lost.

Measuring Unknown Inductance Values

Perhaps one of the most common methods of learning the inductance value of an unknown coil is to use a dip meter. Although this is a rather unsophisticated procedure, compared to the use of high-cost lab equipment for precise measurements, it is suitable for us hams. When using a dip meter, we must also have a calibrated receiver with which to monitor the dip-meter signal (unless your dipper has a very accurate dial scale). The remaining ingredient is a collection of high-quality, known-value capacitors. Silver-mica capacitors are my preference for doing this type of testing.

Fig. 2 illustrates the test procedure when using a dip meter. A known-value capacitor is placed in parallel with the unknown value of inductance. The dipper is swept through its tuning range until a dip (resonance) is observed on the meter. Next, the dip meter is moved farther and farther from the coil until the dip is just discernible. Now we remove our hands from the dipper and tune the monitor receiver to find the dipper signal. This will give us the approximate resonant frequency of the coil and capacitor. The inductance value can then be determined quickly by using the ARRL L/C/F Calculator, Type A. If you prefer to find the inductance value mathematically, you can use the following equation:

$$L(\mu H) = \frac{X_L}{2\pi f(MHz)} \qquad \text{(Eq. 1)}$$

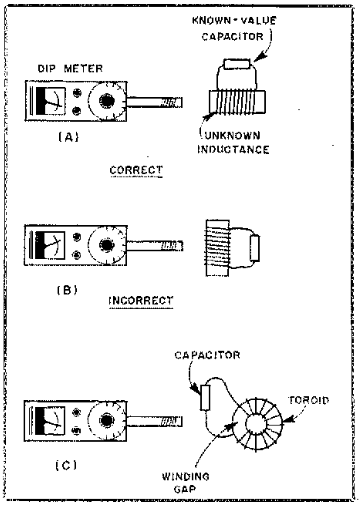

Fig. 3 — Correct coupling between the dip meter and the test coil is shown at A, where the axis of the dipper coil matches that of the test coil. Incorrect placement of the test coil is shown at B. The toroid at C can be dipped if the dipper probe coil is inserted into the capacitor leads at the winding gap.

where X_L is equal to X_c (resonance) and is expressed in ohms.

The reactance (X) of the capacitor connected across the coil can be found from

$$X_c = \frac{1}{2\pi f(MHz) \times C(\mu F)} \qquad \text{(Eq. 2)}$$

Let's take an example and work it out to help you understand the progression of the equations. Suppose we placed a 100-pF silver-mica capacitor across our unknown inductance. The dip meter and the receiver told us that the resonant frequency of the LC combination was 3.9 MHz. First, we will find the reactance of the 100-pF capacitor.

$$X_c = \frac{1}{\underset{(2\pi)}{6.28} \times \underset{(MHz)}{3.9} \times \underset{(\mu F)}{0.0001}} \qquad \text{(Eq. 3)}$$

hence $X_c = 408\Omega$

Note that pF has been changed to μF for this equation.

Now that we know the X_c we can proceed with Eq. 1 to learn the inductance of our coil. Remember that the coil reactance (X_L) is equal to the X_c (408 ohms) because X_c and X_L are equal at resonance.

$$L(\mu H) = \frac{408}{6.28 \times 3.9} \qquad \text{(Eq. 4)}$$

hence $(L\mu H) = 16.65$

If the coil were the type that has a movable slug, we could learn the tuning range by obtaining a dip reading with the core at each end of its travel. The Q would be highest with the core material inside the coil winding.

When using a dip meter, we can get a feel for the relative Q of a coil by noticing how far from the coil we must put the dip-meter probe to obtain a meter reading (dip). The farther we move from the coil under test (still observing a dip), the higher the Q. If we must couple the dipper probe very tightly to the test coil, the Q is quite low. Fig. 3 shows the correct way to couple to an unknown coil. The axis of the dipper-probe coil must match that of the coil being tested. No dip will occur if we approach the test coil from its side.

We will experience a similar difficulty when attempting to measure a toroidal inductor with a dip meter. This is because of the inherent self-shielding nature of the toroid, as mentioned earlier. However, by probing the gap on the toroidal coil (Fig. 3C) it should be possible to find a dip. This is because the discontinuity of the self-shielding at the winding gap permits some coupling to the dipper. The capacitor leads become part of the overall inductance and provide a coupling point.

A Homemade Inductance Meter

Wouldn't it be convenient to have a homemade instrument that could be used to measure small values of inductance? Such a device can be made from ordinary components, and it need not be complex. The two most important parts of the unit would be the tuning capacitor and the indicating meter. The variable capacitor (C1 of Fig. 4) needs to be calibrated in picofarads through its tuning range. This can be done with the aid of a dip meter, a calibrated receiver and a known-value inductance. Coarse calibration can be achieved by checking the resonant frequency of the test circuit (coil and variable capacitor) at 5° settings of C1. A dial scale can be plotted as we proceed with the resonance checks. A vernier drive with a 0-100 dial scale is ideal for use during this exercise. Once the capacitance per 5° increment is known, the value of an unknown inductor can be calculated by means of the equations given earlier, or we may use the ARRL L/C/F slide-rule calculator. Greater sophistication can be had by developing a chart that compares the tuning-dial readings with inductance values.

A suitable circuit is given in Fig. 4. Q1 is a crystal oscillator that operates at 5 MHz. Y1 can be for some other frequency in this general range; it is not a magic number. Q2 is lightly coupled to the output of Q1, and it operates as a broadband amplifier with feedback. Output from Q2 is sampled lightly by means of a 5-pF capacitor, then fed to a voltage-doubler diode detector, D1 and D2. The resultant dc voltage is supplied to a microampere meter, M1. As C1 is tuned through resonance with the test inductor (plugged in at J1 and J2), a peak meter reading will

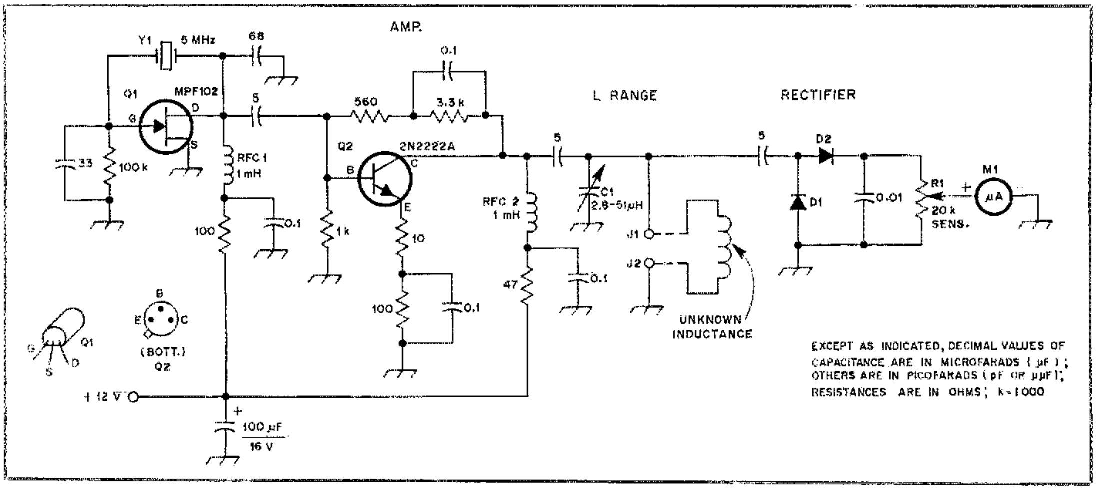

Fig. 4 — Schematic diagram of a test instrument for measuring inductance and relative Q. Fixed-value capacitors are disc-ceramic. The capacitor with polarity marked is tantalum or electrolytic. Fixed-value resistors are 1/4- or 1/2-W carbon composition. J1 and J2 can be plastic binding posts. RFC1 and RFC2 are miniature rf chokes (2.5 mH also suitable). R1 is a linear-taper carbon control.

be noted. The inductance or capacitance can then be read from the C1 tuning dial, depending on whether you calibrate your dial for picofarads or microhenrys.

The circuit shown will measure inductances from approximately 2.8 to 51 µH when C1 is a 365-pF broadcast-radio variable capacitor. Minimum capacitance for these units is approximately 20 pF, according to measurements I have made on a number of them. But, other values of tuning capacitance can be used if desired. The lower the maximum capacitance of C1, the more restricted the inductance range of the instrument.

The range of the circuit in Fig. 4 can be extended upward and downward by switching the crystal at Y1 to a higher or lower range. This could be done manually or by means of a wafer switch. For example, if we change Y1 to 10 MHz, the circuit of Fig. 4 will have an inductance-measuring range of 1.4 to 25 µH. Changing Y1 to 2 MHz, the range would be 7 to 127 µH.

The relative Q of the coil under test can be observed by the meter deflection versus the setting of the sensitivity control, R1. The higher the meter reading as R1 is turned down, the greater the unloaded Q.

For best sensitivity of the test instrument, I would suggest using germanium diodes at D1 and D2. But, good results can be obtained when silicon switching diodes, such as 1N914s, are used. M1 can be a 50- or 100-µA meter. The 50-µA movement will provide the better sensitivity. You may want to try one of the low-cost edgewise fm tuning meters that are available on the surplus market. Most of them are in the low microampere range. Meters and tuning capacitors are available from Surplus Electronics Corp.[4]

A Way to Measure the Q

A colleague of mine, W7ZOI, who is an ARRL Technical Advisor, described a simple method for the measurement of tuned-circuit Q. It is a technique he uses in his home workshop. The circuit under test, which consists of a parallel combination of L and C, with L being the unknown inductor in our case, is swept with a signal generator or low-power amateur transmitter. The tuned circuit is sampled lightly with a scope or VTVM (vacuum-tube voltmeter or FET VOM) and rf probe. The 3-dB points on the tuned-circuit response curve are noted in terms of signal frequency. From this information we can calculate the unloaded Q of the circuit. This is explained in detail on page 29 of the *ARRL Electronics Data Book*. A circuit diagram and pertinent equations for Q calculations are included.

Some Final Thoughts

Simple measurements of the kind discussed in this article are not beyond the beginner. The methods and circuits for simple measurement of inductance should not be difficult for you, either. We can learn far more by *doing* than by simply reading or thinking about these projects. A basic knowledge of coils and their properties will help you to save a lot of money when buying parts for new circuits.[5] You will know how to select surplus or flea-market inductors if you remember the simple guidelines of this article. Winding your own coils should be easier, too!

Notes

[1] D. DeMaw, *Ferromagnetic Core Design & Application Handbook* (Englewood Cliffs, NJ: Prentice-Hall, Inc.).
[2] 19070 Reyes Ave., Compton, CA 90221.
[3] mm = in. × 25.4.
[4] 7294 N.W. 54th St., Miami, FL 33166. Catalog available.
[5] Circuit boards and parts kit are available from Circuit Board Specialists, P.O. Box 969, Pueblo, CO 81002.

Mini Directory

As a convenience to our readers, here is a list of items of particular interest and when they most recently appeared in *QST*.

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Amateur Radio Station Call Sign Assignment System	June 1983, p. 61
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FCC Wrap-Up	Sept. 1983, p. 65
License Renewal Information	Jan. 1983, p. 53
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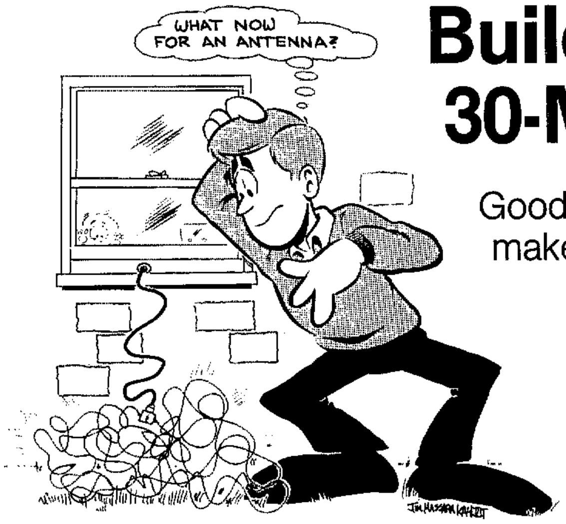
U.S. Amateur Frequency and Mode Allocations	Jan. 1983, p. 53



Building and Using 30-Meter Antennas

Good propagation and minimum QRM make 30 meters a great band for the DX hunter and ragchewer. Good results are possible without elaborate antennas. Let's review some simple ones and scale them to the new WARC band.

By Doug DeMaw,* W1FB

What makes 30 meters so interesting and unique? Perhaps the most outstanding virtue of 10.1 MHz is the propagation characteristic found there. In a nutshell, it is a pleasant combination of the good features we've learned to enjoy on 40 and 20 meters. For the most part, 30 meters is open to some part of the USA or world most of the time. The skip is generally longer than it is on 40 meters, and when 20 is dead, 30 is often open. I'll always recall how amazed I was during my first weeks on the new band, when in Connecticut I was able to hear JAs, ZLs, Europeans and U.S. stations at the same time! I don't remember having that happen on the other hf bands.

Another nice aspect of 10.1 MHz is the present lack of recognition for states or countries worked. At this time, there are no WAS or DXCC awards being issued for 30 meters, and there are no contests being staged. That, plus the 250-W input power limit, cw and RTTY only, make the band a delight to work if you enjoy chewing the rag or operating QRP. The band is not infested with operators who seek to pile up high contest scores or garner new states and countries for the purpose of acquiring an award. The 10.1-MHz band has proven ideal for many of us who have regular schedules with friends.

The U.S. frequency allocation is 10.100 to 10.109 MHz. A second segment exists from 10.115 to 10.150 MHz. *Beware: 10.109 to 10.115 MHz is a "no-man's land" for U.S. operators!* It is illegal for us to operate in that portion of the band. Don't be misled by VE signals in that part of the band: Canadians are allowed to use 10.109 to 10.115 MHz.

The Fine Art of Antenna Selection

I have observed some interesting practices, respective to antenna use, since becoming active on 30 meters. Many amateurs who are trying the band for the first time are using antennas that are cut for other bands — at least until they decide if they wish to stay active on 10.1 MHz. I went through a similar exercise during my first month on the band. I was surprised to find that any hf band antenna would bring reasonable results if the SWR was disguised by means of a Transmatch. Of course, the feed-line losses varied with the antenna being used, since different lengths and types were involved. For example, I had good luck with a 40-meter 5/8-λ sloping vertical, even though I was brute-forcing rf energy through a tapped-coil matching network at the base of the antenna. I was pleased to learn that my Cushcraft A4 triband Yagi would radiate well when using a Transmatch at the transmitter end of the feeder. I had reasonable success when feeding my shunt-fed 50-ft tower, which was resonant at 80 meters.[1] Each antenna enabled me to work DX and local (USA) stations, but the tribander for some reason had the advantage of reduced QSB. I hasten to state that there appeared to be no directional characteristics when using the A4.

If you have a center- or end-fed Zepp antenna, or a dipole with tuned feeders, chances are that you will find that antenna to be ideal on 30 meters. Many stations I have worked displayed strong signals when using that style of antenna. I had no antennas with tuned feeders, so I finally erected a sloping dipole with 50-ohm coaxial feed line, and it worked very nicely for all-round QSOing. But, let's consider some antennas that are designed for 30-meter dedicated use.

Effective Wire Antennas

I find it surprising that a number of amateurs have written to me and asked, "What shall I put up for 30 meters? Can you give me a set of dimensions?" Cutting an antenna to the proper length is perhaps the easiest task we hams must face. Yet, there are many who lack the confidence or knowledge to tackle that simple job. *The ARRL Antenna Book* should become a standard reference for those amateurs.[2]

Fig. 1 shows some basic wire types of antennas that should yield good results on 30 meters. The antenna at A is the popular, easy-to-erect drooping doublet, drooping dipole or inverted-V. This antenna and the others we will consider in this article can each be supported by a single mast or tower. That simplifies the installation. The inverted-V exhibits an omnidirectional radiation pattern. The enclosed angle should be about 90 degrees, but angles up to 120 degrees provide good results.

The length of the inverted-V legs do not become the exact length obtained from the ft = 468/f(MHz) formula. Cut them a trifle long to permit trimming the length for an SWR of 1:1. The final length will depend on the proximity of the wires to nearby conducting objects. The height of the lower ends above ground will also affect

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[1]Notes appear on page 29.

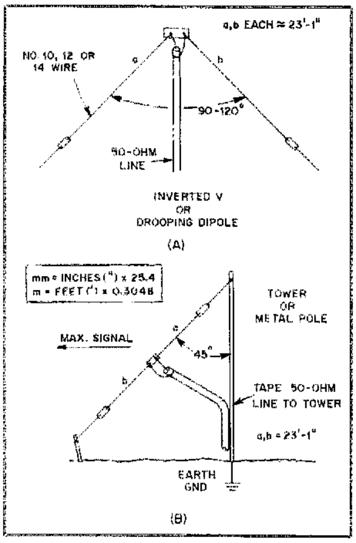

Fig. 1 — Good all-round results can be expected when using the inverted-V antenna shown at A. For successful DX work, the center of the dipole should be at least 0.5 wavelength above ground. The sloper seen at B will serve nicely for local and DX work on 30 meters. The arrow indicates directivity off the slope of the antenna when the supporting structure is metal.

the final length. Trim equal amounts of wire (1/2 inch at a time) from the ends of the dipole until the SWR will decrease no further. Although the SWR may not drop to as low as 1:1, it should be close to that ratio when using 50-ohm feed line. The height above ground for the center of the antenna should be as high as practical. Ideally, the feed point would be a half wavelength (48 feet) above ground, or greater. But, lower heights will permit good operation, too. The higher the antenna, the better the low-angle radiation and, of course, the better the DX results. This rule applies to operation on any hf band.

The antenna seen at B of Fig. 1 is excellent for DX and local use. By "local," we may consider a distance of a few miles out to a couple of thousand miles or more. This sloping half-wave dipole is commonly referred to as a "sloper" or "full sloper." If the antenna is supported from a grounded metal structure as shown, there will be directivity (not gain) in the slope of the dipole, as we have indicated by the arrow in Fig. 1B. A slope angle of 45 degrees is generally used for this radiator, but larger angles can be employed without significant change in performance. Any convenient size of wire can be used for the legs, a and b, provided the wire is strong enough to support the overall dipole. If you have a short mast, causing the lower end of antenna-leg b to be near the ground, some trimming of the antenna length may be required in order to reduce the SWR to near 1:1. Best results will be had if the lower end of the dipole is at least 6 feet above ground. If a tall wooden support is used for the antenna, you may want to erect the radiator as a vertical dipole (perpendicular to the ground). This will yield a good low-angle radiation lobe, which will be ideal for DX work. The pattern will then be omnidirectional. As many as four slopers are used by some amateurs (spaced at 90-degree intervals around the tower). The feed lines are switched by remote means to provide directivity in four directions.

Effective Wire Loop Antennas

Some advantages are possible if we configure our 30-meter wire antenna as a closed loop. If a full-wave loop is erected (Fig. 2A), we will realize increased aperture (called *capture area* by some hams), and there can be a marked reduction in local man-made noise during receive. A full-wave loop is seen as a triangle in Fig. 2A. Again, a single support is specified. The triangle has sides of equal length, although some disparity in the equality will not spoil the antenna performance. As shown, the loop is equal to the driven element of a Delta Loop beam, but with the feed point inverted.

Although top feed is shown in Fig. 2A, the antenna could be fed at the center of one of the sides or at one of the lower corners. You may wish to experiment with the feed point to learn which one provides the best results for your type of operation. Shifting the feed point will change the polarization of the signal and will also have some effect on the radiation angle. I prefer top feed, but have also fed this type of antenna at the center of the bottom leg.

Tuned feeders (open-wire line or 300-ohm TV ribbon) are specified in the interest of reducing feed-line losses (especially if a long run of coaxial cable would be required). The tuned transmission line will permit this loop to do a good job on 10 meters, too. Although it is not exactly related in terms of the third harmonic, it will radiate well on 28 MHz. It also should work well on 20 and 15 meters.

Coaxial cable can be used to feed the loop in Fig. 2A, but a quarter-wave matching transformer will be needed in order to obtain a low SWR. The feed impedance of the full-wave loop is approximately 115 ohms. A section of 72-ohm coaxial cable, 16 feet 1/2 inch long, can be inserted between the loop feed point and the 50-ohm line to the shack when coaxial-cable feed is desired. The piece of RG-59/U or RG-11/U cable will then serve as a matching transformer to convert the feed impedance to that of the 50-ohm line. A low SWR should result. The size specified for the 72-ohm section is a quarter wavelength for 10.125 MHz, and the 0.66 velocity factor of the cable has been included in the dimension. The exact length will depend on the type of 72-ohm line you use; the velocity factors differ slightly with respect to the insulation used. The ARRL's *Handbook* and *Antenna Book* list the velocity factors for the various popular coaxial cables. The formula is 246/f(MHz) × 0.66 when using solid polyethylene insulation.

Height factor h of Fig. 2A is of importance if we are to have good DX results with the loop. Ideally, dimension h would be 35 feet or greater to ensure a low angle of radiation. But, good results can be had with much lower antenna height. I enjoyed very good DX results with a 40-meter version of this antenna, even though the lower leg was just 4 feet above ground. I was able to work 22 countries with a 2-W QRP cw

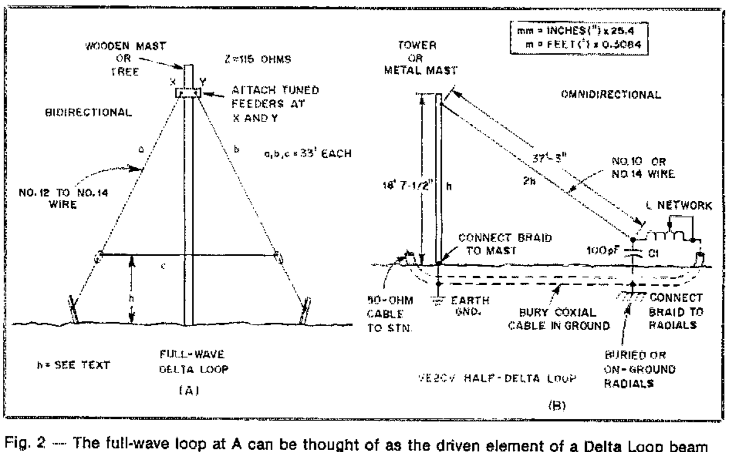

Fig. 2 — The full-wave loop at A can be thought of as the driven element of a Delta Loop beam antenna. It is capable of yielding excellent results on 30 meters, but maximum directivity is at right angles to the plane of the loop, making it a bidirectional radiator. A 30-meter version of the VE2CV Half-Delta Loop is shown at B. This is a highly effective DX antenna, but is practical also for "local" communications. Ground radials are needed with this antenna.

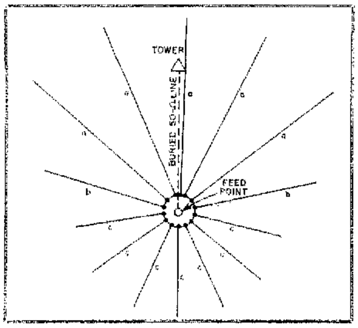

Fig. 3 — Looking down on the Half-Delta Loop, we envision the layout for a suitable radial system. The greater the number of radials, the better the effective ground under the antenna. Radials "a" should extend slightly beyond the vertical element of the loop, as shown. Depending on the ground conductivity in a given region, fewer radials than those shown may be ample. Good results have been reported by some amateurs who used only ground rods at each end of the loop.

rig during a three-month period when I lived in Connecticut. The loop was broadside east and west.

There is no reason why the loop of Fig. 2A could not be erected in a square, or even a slightly rectangular format if your property requires it. Irrespective of the antenna shape, the radiator will be a full-wave loop if the overall length is taken from 1005/f(MHz), which provides the answer in feet. The sides of the loop should not be too close together (long, narrow rectangle), lest performance be impaired. Such a format would approach that of a folded dipole. The radiation from the loop will be bidirectional off the broad side of the antenna.

The Half-Delta Loop for 30 Meters

The complete story about the Half-Delta Loop was carried in *QST* and other literature.[3] Therefore, we won't delve too deeply into the theory of operation. In effect, our loop conductor (tower and slant wire) is 0.5 wavelength long, plus a k factor of 1.15. Again, a single support structure is needed (h of Fig. 2B), and that section of the system must be conductive, since it is an electrical part of the loop. There should be *nothing* atop the mast or tower, because any additional conductors will seriously affect the antenna resonance and performance. This means that a tower with guy wires or additional antennas can't be used for the Half-Delta Loop. A tree or wooden mast can be used, however, provided a vertical wire (h) is added to take the place of the missing tower or metal mast. So, you need not have a tower to use this excellent antenna.

Electrically, the antenna of Fig. 2B functions as a full-wave loop by virtue of the missing half of the loop appearing in the ground as an image. The polarization is vertical, and the radiation pattern is principally omnidirectional. The angle of radiation is very low, which makes the loop excellent for ground-wave communication and DX.

A feed impedance of approximately 100 ohms will result if the dimensions are correct and when a reasonable ground system is used with the loop (radials). A matching transformer made from 72-ohm coaxial cable (described for the antenna of Fig. 2A) could be used if the L network of Fig. 2B was not desired. A low SWR would result. The simple L network will, however, permit precise matching to 50- or 72-ohm line.

The feed line is buried a few inches in the ground to keep it from being a physical hazard to people. The performance will be just as good if the feed line is simply laid on the ground. In either event, the shield braid of the feeder should be connected to the radial wires at one end of the loop and to the base of the tower at the remaining end.

Good results can be obtained with a few in-ground or on-ground radial wires. A suggested minimum number is illustrated in Fig. 3. A ground rod, or a group of rods tied together electrically and spaced 3 feet apart, should be located at the feed point and at the base of the tower or mast.

The L network of Fig. 2B may contain a coil that can be tapped during the SWR adjustment. For 30 meters, I use a handwound inductor that has 20 turns of no. 14 copper wire. The coil diameter is 1-1/2 inches, and the length is 6 inches. C1 should have 1/8-inch plate spacing for power levels up to 250 W — the maximum dc-power input allowed in the USA for 30 meters. C1 and L1 are varied until the SWR is 1:1. The bandwidth of the L network will be great enough for coverage of the entire 30-meter band when the SWR is adjusted at 10.125 MHz.

Vertical Antennas

A ground-plane vertical will generally give better performance results when it is used in preference to a ground-mounted vertical at 30 meters. As we approach the upper end of the hf spectrum, the signal is affected more and more by nearby conductive objects (power lines, downspouts, plumbing, fences and even large, dense trees with foliage). Absorption and pattern distortion can result when the antenna is obstructed by such things. Better performance will result if the vertical is elevated above the nearby clutter. Another advantage to the above-ground installation is the reduced number of radial wires required. A system of four "floating" radials will permit the system to do a good job for DX work.

A simple ground-plane vertical is shown in Fig. 4. The vertical portion consists of aluminum tubing. To ensure ample rigidity and longevity, I use telescoping sections of tubing (3), starting with a 1-1/2 inch diameter piece at the bottom. The sections

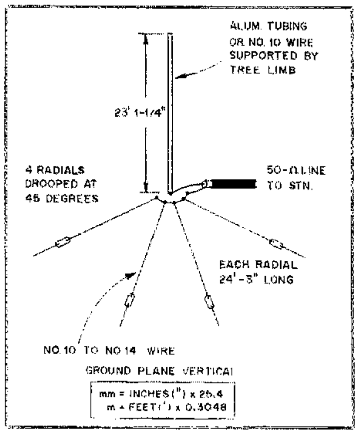

Fig. 4 — Details of a ground-plane vertical antenna for 30 meters. This omnidirectional, low-angle radiator will do a fine job for DX and local operations. See the text for information on construction methods.

slide into one another for a length of 12 inches. Four slots are sawed in the outer tubing at each joint to permit a good bond between the sections when using a stainless-steel hose clamp. A short piece of dowel rod inside the inner tubing at each joint will permit the clamps to be drawn up tight without distorting the tubing.

The radials can serve as guy wires for the support mast. If they are drooped as shown in Fig. 4, the feed impedance will closely approach 50 ohms. If they are at right angles to the vertical element, the impedance will be on the order of 30 ohms, thereby requiring a matching section.

An antenna of this type can be made entirely from wire if you have a tall tree from which to hang it. The vertical element would then be made of wire and would require an insulator at the top end. Irrespective of the construction method used, this style of antenna provides low-angle radiation and has an omnidirectional pattern.

Tag Ends

We did not get into a treatment of gain types of directional antennas in this article. Certainly, Yagis, quads, ZL Specials and other types of antennas are worth considering if you want a single-band beam antenna. The intent here was to describe some simple, inexpensive wire antennas that are capable of delivering good performance at 30 meters. Needless to say, these antennas can be scaled to other bands of operation, and should provide equally good results. Good luck and happy DXing!

Notes

[1] m = ft × 0.3048; mm = in. × 25.4.

[2] The greatly revised 14th edition is now available from ARRL.

[3] J. Belrose, "The Half-Delta Loop: A Grounded, Vertically Polarized Antenna," *Ham Radio,* May 1982. Also, J. Belrose and D. DeMaw, "The Half-Delta Loop: A Critical Analysis and Practical Deployment," *QST,* Sept. 1982. QST

The Noise Maker — An Aid to Learning the Morse Code

Using a keyed white-noise generator in place of a conventional code-practice oscillator may prove to be an advantage when learning to copy cw.

By Don E. Hildreth,* W6NRW

Code-practice oscillators have always been pretty much the same, it seems. I was about 12 when I first used one to learn the code — and that was 50 years ago! More often than not, the tone employed is somewhere between 500 and 1000 Hz, and is relatively pure (narrow band). In addition, well-designed tone generators provide smooth keying with a minimum of starting and stopping transients, or "clicks" (clicks are not only annoying to listen to, but over time they can cause hearing damage). Thousands, perhaps millions, of people have learned the Morse code using these simple and effective devices.

So why change something that has worked so well, for so many, for so long? Actually, what I am describing here is not a change but an alternative. Why bother? Well, consider the fact that there are many would be "brass pounders" who just can't seem to learn the code. Why should this be? The spectrum of speech frequencies is immensely more complex than the simple on-and-off pattern of Morse code. Still, there are those who have great difficulty with cw communication. Could it be that the hearing process for some of us just "doesn't like" a single tone going on and off all the time?

While teaching the code, I have encountered those who would start making headway when the keyed tone was lowered to around 300 Hz or less. Some of these people had hearing problems, usually a high-frequency hearing loss, that also made speech comprehension difficult. But this was not always the case. I have also encountered those with normal hearing who claim to experience a ringing that seems to fill the off periods in a code letter. For these people, learning the code is difficult indeed. Others with normal hearing, as measured by an audiologist, just can't seem to learn code even though they have an intense desire to do so, and they cannot describe their problem. Since none of us can experience the senses of another, it is very difficult to learn how to be of assistance.

Keyed Noise Generator

Recently, I tried keying a noise generator in place of an oscillator for code practice, and the results were good. Based on past experience in teaching the code, I feel certain that this technique could help some of those with defined hearing loss — and perhaps others, as well. I don't have a hearing problem for code, but I just like the sound of this keyed noise generator. I wouldn't think of going back to the old tone-generator system.

But what would be the use of learning with this device when almost all ham gear produces a beat note for cw? The answer to this would depend on your specific problem. If you have a high-frequency hearing loss that makes the typical 750-Hz beat note difficult to hear, you can use the electronic switch included in this article. This will work in cases where QRM is not too bad and for casual use. There is also a more sophisticated process called Tone-Tag, which is designed to tone modulate a cw signal with 100 Hz only when it is tuned to a beat note range of 700 to 800 Hz.[1] This can operate, and even provide benefit, under extreme QRM and poor signal-to-noise ratio conditions. But in many cases (perhaps most), as you become proficient with the code, you will develop the ability to handle adverse conditions unaided. It seems clear that it is mostly when you are first learning that you can use all of the help you can get! The Noise Maker may help you get over the hump.

Circuit Description

The basic circuit, which uses very low cost components, is shown in Fig. 1. It is designed so that low-impedance stereo headsets or the older, 2-kΩ monaural headsets can be used without circuit changes. Although the output level is partially dependent on the headset efficiency, most units will provide sound levels of 60-70 dB (A), which is usually adequate for comfortable listening. [Editor's Note: The dB (A) is a unit of sound-level measurement that uses "A weighting." This is a frequency-dependent scale on a sound-level meter, adjusted to approximate the way your ear responds to different frequencies. The 0-dB reference level is the minimum sound level that your ear can detect at its most sensitive frequency, about 1 to 2 kHz. Normal conversation is in the range of 50-70 dB (A).]

Q1, connected as a Zener diode through R6 to +9 V, is the noise source. The noise signal is coupled through C2 to the high-impedance noninverting input of the opera-

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[1]Notes appear on page 31.

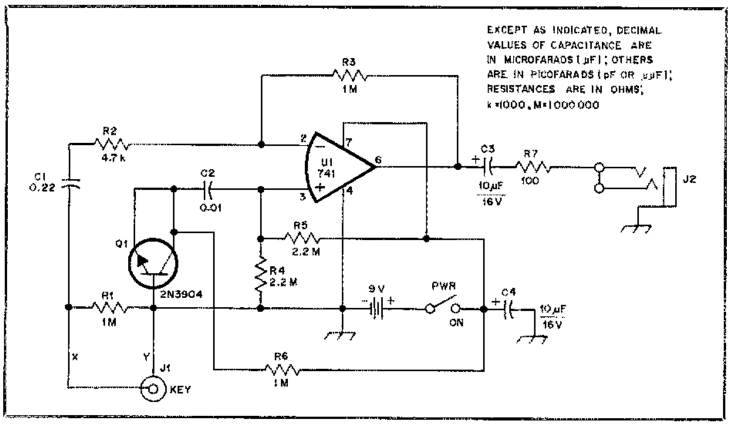


Fig. 1 — Schematic diagram of the Noise Maker. All resistors are 1/4-W, 5%, carbon types. Capacitors are disc ceramic unless otherwise noted. Polarized capacitors are electrolytic.

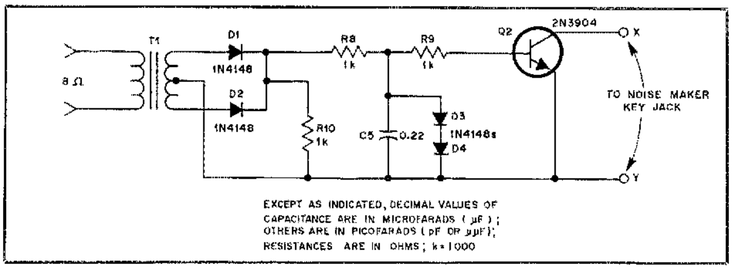


Fig. 2 — Transistor switch for keying the Noise Maker from a receiver speaker output. T1 is a miniature 8-Ω to 1-kΩ audio transformer. All resistors are 1/4-W, 5% carbon types.

tional amplifier (op amp), U1. R4 and R5 serve as a load resistor and set the op amp output to one half the supply voltage. The amplifier voltage gain is approximately 2 when the key is open, as determined by the gain equation 1 + R3/(R1 + R2). When the key is closed, the gain goes up to about 1 + R3/R2, which results in a 34-dB on/off ratio. Using this circuit arrangement, C1 is forced to mirror the nominal +4.5-V level at the amplifier inverting input, thus eliminating occasional keying transients. Using this design, the key is making and breaking only a few millivolts of noise signal. This results in smooth and clickless keying. C3 eliminates dc from the load, and R7 is included to avoid overloading the op amp when 8-Ω headsets are used, and to establish the condition such that the sound level is about the same whether you plug in high- or low-impedance headsets. With the component values given, the noise band produced will be from approximately 150 Hz to 10 kHz.

Construction

In general, construction is not critical. It may be prudent, however, not to use key leads of more than two or three feet in length and to keep the leads associated with pin three of the op amp to an inch or less in length.[2] This will reduce the possibility of picking up a local broadcast station or other strong local fields.

I just stuck the parts down on the back of the panel with Silastic® and then used point-to-point wiring. Avoiding a circuit board cuts the parts cost almost in half, and direct wiring only takes a few minutes.[3]

Alternative Uses

Quite naturally the question will arise: What if I want to be able to listen to code from my receiver or code practice tapes, etc. through this thing?

Fortunately, the switching needed to make Noise Maker function without a key is simple. Adding an ordinary transistor switch assembly, as shown in Fig. 2, will allow your receiver or other code producing device to operate the Noise Maker very nicely. With this circuit addition, T1 connects to your receiver speaker output and the Q2 collector and emitter leads connect to the Noise Maker key terminals. When a cw signal is tuned in, the nominal 750-Hz signal is increased in voltage by the transformer and rectified by D1 and D2. R10 is a load resistor. The combination of R8, C5, D3 and D4 filters and limits the resulting signal to a suitable level for driving the base of the transistor switch (Q2) through current-limiting resistor R9. An electronic switch of this type is far from perfect, but it is more than adequate in this case.

There are other uses for white noise. Just short the key terminals and use headsets plus noise to reduce and mask outside interference when trying to study, concentrate, meditate, or whatever. Most white-noise generators on the market cost a lot more than this unit — and you can't key them very well either.

Notes

[1]D. E. Hildreth, "Communications Audio Processor for Reception," *Ham Radio*, Jan. 1980, pp. 71-79.
[2]mm = in. × 25.4; m = ft × 0.3048.
[3]Complete units are available from the author for $12.95 plus $2 to cover shipping. The ARRL and *QST* in no way warrant this offer.

Next Month in QST

November's the month for a full report on June 1983 Field Day and rules for some future operating activities. You'll also read about a new nationwide program designed to attract youngsters into ham radio.

For technical articles, "A DTMF 'Easy-Controller' — With Security" describes a security system for repeaters, and a Beginner's Bench article gives some simple ways to test your transmitter.

This is just a taste of all of the interesting tidbits coming in the next issue; it promises to be the largest issue so far this year.

Strays

SSC LOGO CONTEST

☐ Are you or any of your club members artists? If so, you have an opportunity to design a logo for the new Special Services Club program. Here's what's involved.

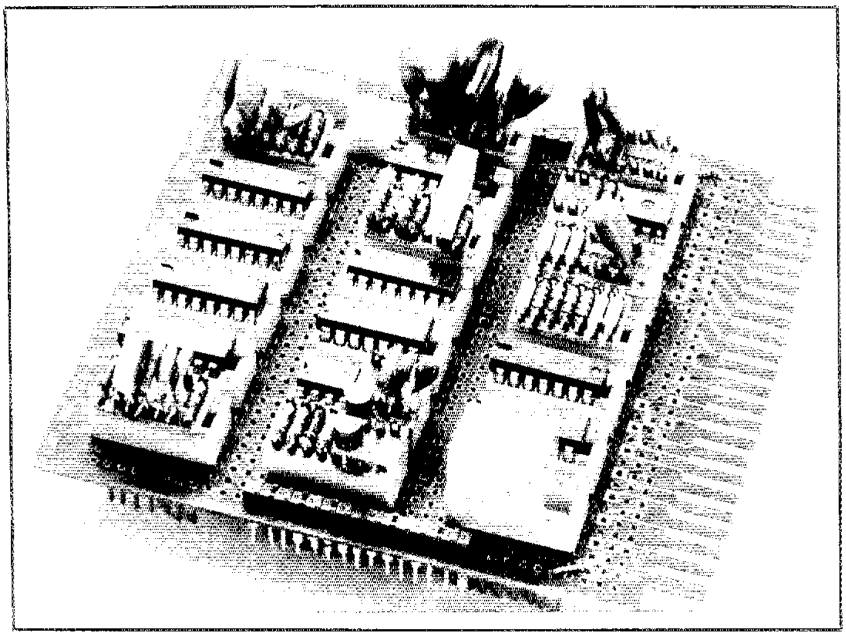
Your work will be judged on how well it
a) symbolizes Amateur Radio clubs
b) expresses superior achievement
c) is aesthetically pleasing.

Your design must
a) incorporate the ARRL logo in a prominent location
b) spell out "Special Service Club"
c) be legible when reduced and embroidered in a size of roughly 4-6 inches.

All designs will be submitted anonymously to the ARRL Membership Affairs Committee and the Board of Directors for a final decision. The winning designer will receive an engraved plaque, a cloth-bound *Handbook* and photo coverage in *QST*.

Submit your original design on 8½- × 11-in. white paper to Special Service Club Logo Contest, ARRL Hq., 225 Main St., Newington, CT 06111. Include your name, call sign and address. Your entry must arrive at ARRL Hq. no later than January 15, 1984. *This contest is for ARRL members only.*

Build an Amateur Radio Modem



Want to use your computer on the air? This explanation of the RS-232 interface, and a modem you can build, will help you do it.

By Reed Valleau,* N1BEG

Amateurs have used digital data for years. Until recently, this meant attaching an RTTY (radioteletype) unit to a transceiver and using the 5-bit Baudot code. A couple of changes in recent years have altered this picture. Because of integrated circuits (ICs), sophisticated communications and computer equipment is available at reasonable prices. In fact, computers have become so inexpensive that many are dedicated to trivial tasks — something unheard of a few years ago. Also, the FCC now allows amateurs to use ASCII in addition to the Baudot code. Since this is the code used by most computer hobbyists, the FCC has made it easier to send computer data over the air.

A typical modern digital communications setup is shown in Fig. 1. The source of digital data is no longer a teleprinter unit, but could be an ASCII terminal or a terminal and computer combination. If only the terminal is used, ASCII data would be transmitted as the operator typed at the keyboard. Add a microprocessor to this setup and more options become available, such as:

1) ASCII-to-Baudot conversion: The operator enters ASCII data; the computer changes this data to Baudot and sends it to the transceiver at any speed.

2) Protocol implementation and error detection: Messages can be communicated between computers within an envelope. This envelope can contain enough redundant data to let the computer detect errors. If errors are detected, the computer can request retransmission. Only correct messages would be displayed to the operator.

3) Packet Switching: Once an error-free protocol is implemented, the computer can act as a store-and-forward device to transfer error-free packets within a network.

Similarly, the computer could be programmed to receive all messages, but save only those sent to a particular station. Most home computers can be programmed to implement any of these functions.[1-4]

Whatever the source of data, one problem must be addressed before digital data can be sent over the airwaves. Computer information is stored digitally and transferred in serial binary format, while radio transceivers can communicate analog signals only. The device that merges these two incongruous worlds is the *modem* (for MOdulator/DEModulator). Modems have application in areas other than communications over the airwaves. They are used in telecommunications to provide data-links between computers over telephone lines. Another type of modem is used to store and retrieve digital data using an analog tape recorder. Modems may be purchased from a number of manufacturers or they may be built easily.

The modem presented here conforms to the EIA RS-232-C standard for half-duplex communication over a dedicated line and will operate reliably at speeds up to 300 bauds. Half duplex is the mode normally used in amateur communications. It means that only one party is transmitting at a time. (Full duplex allows for simultaneous bidirectional communications, such as

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[1]Notes appear on page 36.

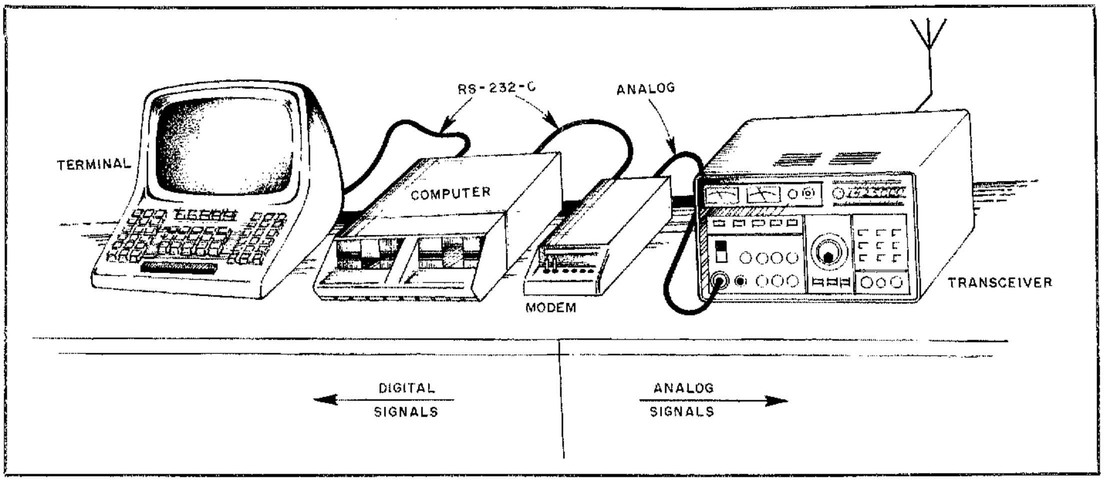


Fig. 1 — A modern set up for communicating digital data might use an ASCII terminal, a computer, or both to generate the digital data. The digital signals are converted to analog signals by the modem and are fed into the transmitter microphone input. Received data goes from the receiver to the modem, where the analog tones are changed into digital data that the terminal or computer can use.

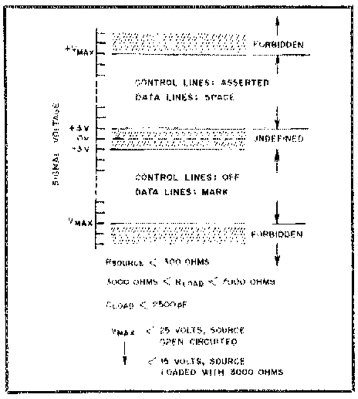


Fig. 2 — Voltage levels on RS-232-C data and control lines are between 3 and 15 V or between −3 and −15 V, as shown in this figure. Also shown is the meaning of these two ranges for the data lines (RxD and TxD) and the control lines (RTS, CTS, DCD, and DSR).

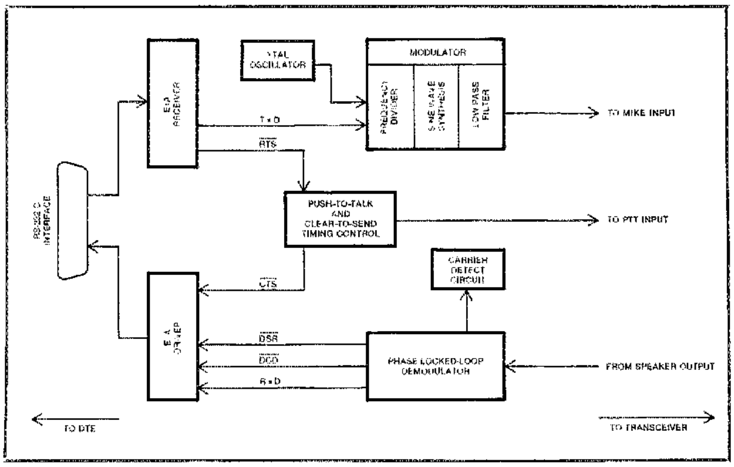


Fig. 3 — A block diagram of the modem. Connected to the left side is a terminal or computer with an RS-232-C connection. The right side connects to a transceiver.

normal telephone communications.) This modem can be used to connect any standard terminal or computer, equipped with an RS-232-C interface, to a transceiver or transmitter/receiver pair. The modem changes serial digital data to audio tones and audio tones into digital data. The tones produced by the modem are fed into the transmitter microphone input. This produces F1 or A2 emissions when used with an ssb or an a-m transmitter, respectively. Tones received from the transceiver speaker output are demodulated by a phase-locked-loop circuit to produce a serial data stream that is sent to the computer or terminal for interpretation.

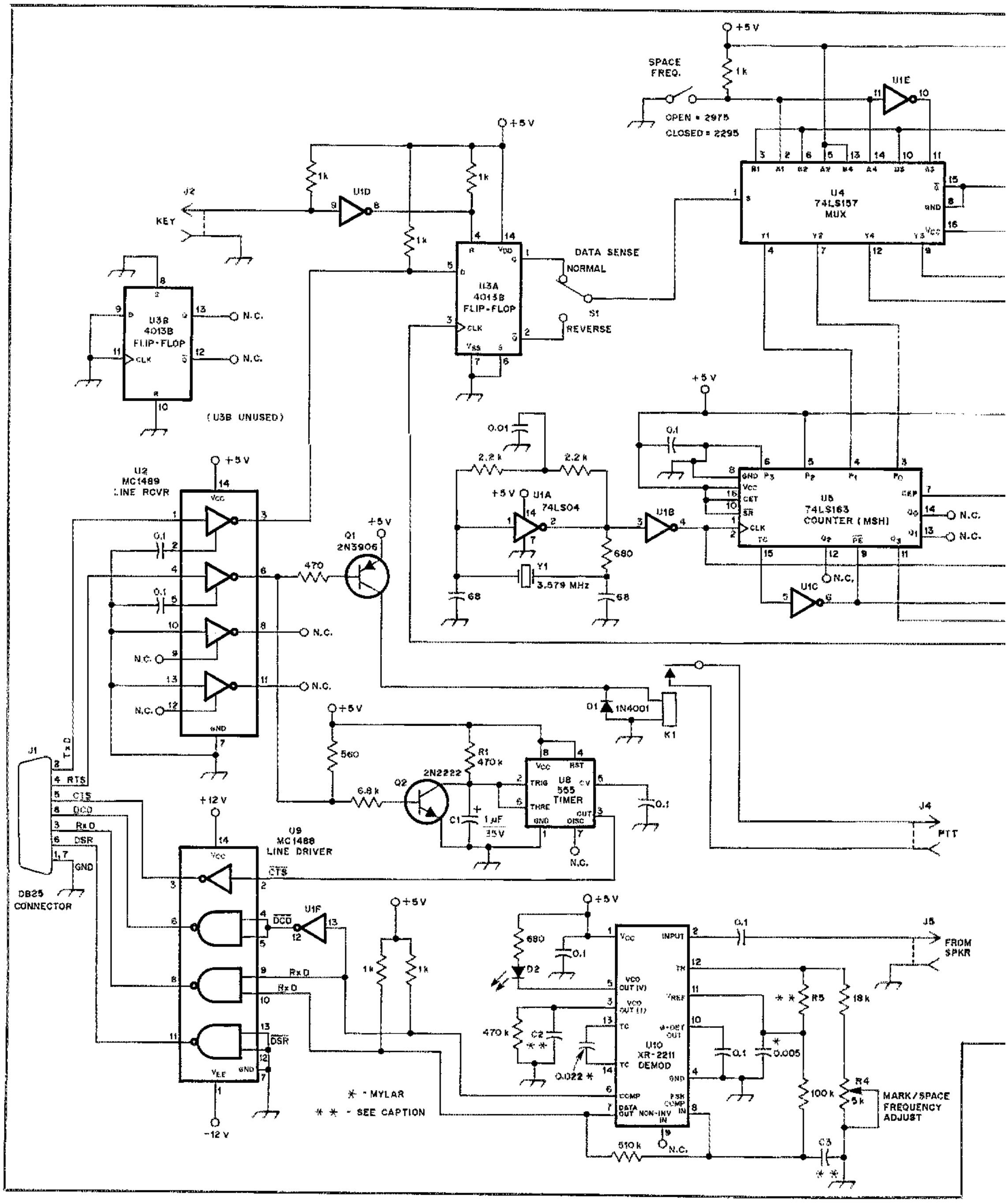
What is RS-232-C?

The RS-232-C interface standard was designed years ago to reduce the problem of interconnecting computer equipment made by different manufacturers. The standard separates equipment into two broad categories: Hardware that is the source and/or destination of communicated data, and hardware used to perform the communication process. The first category is known as Data-Terminal Equipment (DTE) and is usually a terminal or a computer. The second category is referred to as Data-Communication Equipment (DCE), and generally refers to modems. In Amateur Radio applications, DCE includes not only the modem but also the transceiver.

The RS-232-C standard is comprehensive in scope, but remains general-purpose in nature — which accounts for its wide acceptance. This standard describes the electrical characteristics of the interface, including voltage levels (Fig. 2), impedances and capacitances. It also specifies which data and handshaking signals are required for different types of communication. The electrical specifications are numerous and have caused many headaches for engineers throughout the years. Fortunately, Motorola has designed a pair of moderately priced ICs that take care of these electrical requirements for most applications.

Handshaking Signals

The data and handshaking signals for a half-duplex line may seem complicated at first, but are easy to understand with a little scrutiny. For a half-duplex communication channel, the standard requires two lines into, and four lines out of, the modem. Since the computer "sees" these connections as the communications channel (and



since many applications don't follow the letter of the standard), it is worth spending some time discussing these signals and their relationship to one another.

The first signal required by the standard is known as Data-Set-Ready (DSR). It is used by DTE to determine whether or not there is DCE attached to the RS-232-C port and if that equipment is ready to accept control and data information. When the DTE is ready to send data, it "tells" the modem to go into transmit mode by asserting the Request-To-Send (RTS) control line.

This causes two things to happen: First, the modem places the transceiver in transmit by activating the push-to-talk (PTT) relay, causing the mark frequency to be sent. RTS is also used to start a programmable timer, the output of which is used as the Clear-To-Send (CTS) signal.

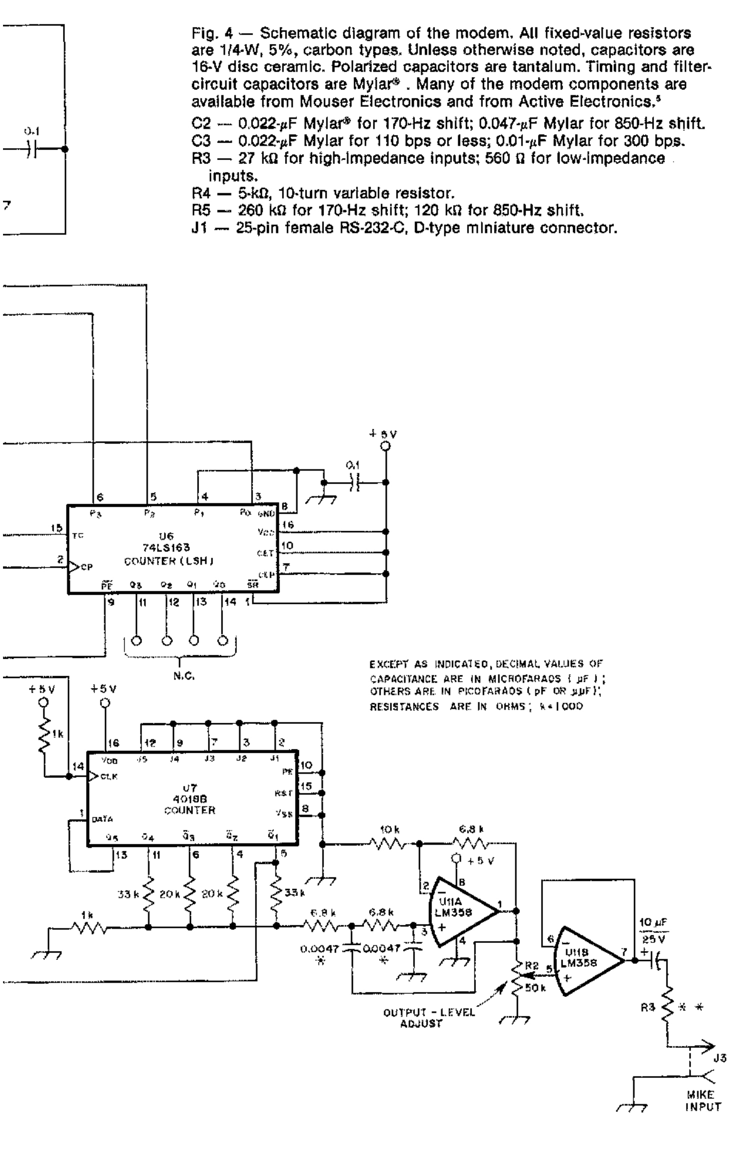


Fig. 4 — Schematic diagram of the modem. All fixed-value resistors are 1/4-W, 5%, carbon types. Unless otherwise noted, capacitors are 16-V disc ceramic. Polarized capacitors are tantalum. Timing and filter-circuit capacitors are Mylar®. Many of the modem components are available from Mouser Electronics and from Active Electronics.[5]

C2 — 0.022-µF Mylar® for 170-Hz shift; 0.047-µF Mylar for 850-Hz shift.
C3 — 0.022-µF Mylar for 110 bps or less; 0.01-µF Mylar for 300 bps.
R3 — 27 kΩ for high-impedance inputs; 560 Ω for low-impedance inputs.
R4 — 5-kΩ, 10-turn variable resistor.
R5 — 260 kΩ for 170-Hz shift; 120 kΩ for 850-Hz shift.
J1 — 25-pin female RS-232-C, D-type miniature connector.

The RTS-to-CTS delay ensures that the transmitter relays have enough time to switch, and that the receiver has time to lock onto the carrier before data is sent. The exact length of the RTS-to-CTS delay is not specified by the standard and may vary from almost no delay at all to a second or more, depending on the application. A delay of 500 ms is reasonable for most amateur work, but it will depend on the circumstances. For example, if two computers are connected through all solid-state equipment, with fast switching times, the delay can be reduced. When the DTE receives CTS, it is allowed to start sending data immediately. As long as the CTS control line remains asserted, the DTE is free to send data in serial form to the modem over the Transmit-Data line (TxD).

When the DTE has completed sending the data, it puts the modem into the receive mode by dropping the RTS line. The modem responds by turning off the PTT relay and by immediately dropping the CTS control line. The transceiver then goes to the receive mode and the modem monitors the speaker output for a carrier sent by another modem. When a carrier is detected, the modem activates the Data-Carrier-Detect (DCD) control line to the DTE. It then allows any received data to be transferred from the demodulator circuit to the DTE by way of the Received-Data line (RxD).

While the preceding description is technically accurate, not all control lines will be used for all applications. For example, many terminals sold today assert the RTS line as long as they are powered. This problem can be overcome by disabling the push-to-talk control circuit and switching the transceiver manually.

About the Circuit

Fig. 3 is a block diagram of the modem, and the schematic diagram is shown in Fig. 4. The modulator circuit is a sine-wave synthesizer driven by a crystal-controlled, 3.579-MHz oscillator. The oscillator signal is divided by one of the two preset divisor values before being applied to the synthesizer. One divisor results in the generation of the mark frequency while the other produces the space frequency. A multiplexer (U4) is used to select the correct divisor, depending on the state of the serial TxD input. The oscillator output is divided by 168 for the 2125-Hz mark frequency, by 156 for the 2295-Hz narrow-shift space frequency or by 120 for the wide-shift space frequency of 2975 Hz. The oscillator signal is further divided by 10 in the CMOS (complimentary-symmetry metal-oxide semiconductor) sine-wave synthesizer (U7). The synthesizer is a 10-step Johnson ring counter that generates a five-level sine-wave approximation.

The synthesized sine-wave is routed through a low-pass filter to remove high-frequency harmonics, and buffered before being applied to the transceiver microphone input. The first significant harmonic from the synthesizer is the ninth (about 19 kHz), so the two-section, low-pass filter has been designed to attenuate frequencies higher than about 5 kHz. This means that the audio signals produced by the synthesizer will experience no appreciable phase shift, but harmonics are effectively removed. The result is a pure sine wave with little phase distortion. R2 is used to set the output level to the transmitter and should be adjusted so the modem does not overload the transmitter audio amplifier. U3 ensures that data transitions occur only at sine-wave zero crossings, thus reducing unwanted adjacent-channel splatter. The modulator circuit requires no frequency adjustments and output frequencies are accurate to within 0.25%.

The demodulator circuit is designed around the EXAR-2211 IC (U10), which is specifically designed to demodulate afsk tones. This IC uses phase-locked-loop techniques and incorporates many circuit enhancements to ensure reliable operation. The IC outputs are used to drive the DCD control line and to provide the demodulated data to the RS-232 interface

IC. As mentioned earlier, the RS-232-C drivers (U2 and U9) are a pair of Motorola ICs that translate the EIA-specified levels to TTL-compatible levels. The RTS-to-CTS delay circuit is a one transistor unit that charges C1 through R1 when the RTS line is asserted and discharges the same capacitor, through Q2 when RTS is dropped. The RTS-to-CTS delay time can be adjusted from a few milliseconds to several seconds by changing the value of R1. The greater the resistance the longer the delay time. The exact value of R22 can be found for this time range by using the formula

$$\text{R1 (ohms)} = \text{delay time (seconds)} \times 945{,}000 \text{ (ohms/second)} \qquad \text{(Eq. 1)}$$

A 555 timer (U8) is used as a threshold detector and comparator. The U8 output serves as the CTS signal.

Construction

The construction of this modem is generally noncritical, but a few precautions should be followed to ensure proper and long-lasting operation. My prototype unit was constructed from wire-wrap components on a Radio Shack 4.5- × 6.0-inch wire-wrap board and cost about $30 (using all new parts).

The crystal-oscillator components should be mounted close to each other to ensure frequency stability. Make sure that the unit is housed in a metal container to prevent RFI from being generated by the oscillator. The crystal is a standard television color-burst type and should be available at any TV repair shop or Radio Shack store. All TTL parts should be of the LS or HC variety for low power consumption.

Adjustments

After the unit is constructed, remove C2 and connect U10 pin 2 to pin 10. Attach a frequency counter to U10 pin 3 and adjust R4 so that the frequency counter indicates a frequency halfway between the mark and space frequencies (2210 Hz for 170-Hz shift, or 2575 Hz for 850-Hz shift). Reconnect capacitor C2. Check the output of the Johnson ring counter (U11, pin 7) with a frequency counter or an oscilloscope to ensure proper operation. As a final check, connect the modulator output to the demodulator input and attach an ASCII terminal to J1. Asserting RTS should cause CTS to be asserted after the proper delay. Data sent to the modem on the TxD line should be echoed to the terminal on the RxD line.

For final adjustment, connect the modem to a transceiver and tune to a RTTY transmission. The Carrier-Detect indicator will be brightest when the signal is tuned correctly. An oscilloscope connected to the RxD output will show the received data as it would be presented to the DTE. Remember that this type of unit will produce a 100% duty cycle when transmitting, compared to a duty cycle of about 25% for unprocessed ssb voice communication. Be sure your final amplifier can stand this type of workout, or reduce the drive to the amplifier so that it will not overheat.

Televideo 920, 925, Most ASCII Terminals

(Input)	*Pin*	*Name*		*Name*	*Pin*	*(Modem)*
	2	TxD		TxD	2	
	3	RxD		RxD	3	
	1,7	GND		GND	1,7	
	4	RTS	┐	RTS	4	
	5	CTS	┘	CTS	5	
	6	DSR		DSR	6	
	8	DCD		DCD	8	

(A)

Altos 8000-15D Computer, or a computer whose RS-232-C connections are configured as DCE

(Input)	*Pin*	*Name*		*Name*	*Pin*	*(Modem)*
	3	RxD		TxD	2	
	2	TxD		RxD	3	
	5	CTS		RTS	4	
	4	RTS		CTS	5	
				DSR	6	
	20	DTR		DCD	8	
	1,7	GND		GND	1,7	

(B)

IBM Personal Computer, or a computer whose RS-232-C connections are configured as DTE

(Input)	*Pin*	*(Name)*		*Name*	*Pin*	*(Modem)*
	2	TxD		TxD	2	
	3	RxD		RxD	3	
	4	RTS		RTS	4	
	5	CTS		CTS	5	
	6	DSR		DSR	6	
	8	DCD		DCD	8	
	1,7	GND		GND	1,7	

(C)

Fig. 5 — Typical connecting cables to go between three types of DTE and the modem.

Typical Connections

Three cable configurations are shown in Fig. 5 for connecting the modem to:

1) A typical ASCII terminal (Televideo 920 or 925).

2) A computer RS-232-C port configured as DCE (ALTOS 8000-15D).

3) A computer RS-232-C port configured as DTE (IBM Personal Computer).

I chose these three connections because they cover about 95% of the cases encountered when connecting DTE to DCE. In the first case the output of the Televideo, like most ASCII terminals, is wired as DTE, so the connection is straight through on all lines except RTS (Fig. 5A). As mentioned earlier, RTS is usually always held asserted on terminals. This problem is overcome easily by simply routing RTS to CTS on the terminal and switching the transceiver manually. The second case shows a computer RS-232-C connection that looks like DCE, even though computers are often thought of as DTE.

This situation is a little confusing but very common because it requires the simplest cable to connect a terminal to the computer. The cable shown in Fig. 5B is known as a crossover cable, because it crosses TxD to RxD and RTS to CTS. The last case (Fig. 5C) shows a computer output that is wired as DTE. Note that this cable configuration is identical to 5A with the exception that RTS and CTS are fully implemented and under computer control.

A word of caution: As anyone who has ever connected two pieces of computer equipment together will tell you, there is usually some snag that prevents the connection from working the first time. The best way to get your connection working is to read the manuals for the equipment you are using, understand just what is required by both sides of the connection, and work on one signal at a time. A patch box or "break-out box" is frequently helpful for stubborn connections. This method takes time, but persistence will always solve the problem. Start building your modem now and soon you'll be able to enjoy using your computer for RTTY communications.

Notes

[1]P. L. Rinaldo and D. W. Borden, "The Making of an Amateur Packet-Radio Network," *QST*, Oct. 1981, pp. 28-30.

[2]J. P. Martinez, "Amtor, An Improved Error-Free RTTY System," *QST*, June 1981, pp. 25-27.

[3]G. McIntire, "Designing a Microprocessor-Based RTTY Speed and Code Converter," *QST*, Jan. 1982, pp. 18-21.

[4]E. Kalin, "A Programmable Serial-Communication Interface," *QST*, Sept. 1982, pp. 18-24.

[5]Mouser Electronics, 11433 Woodside Ave., Santee, CA 92071; Active Electronics, P.O. Box 8000, Westborough, MA 01581.

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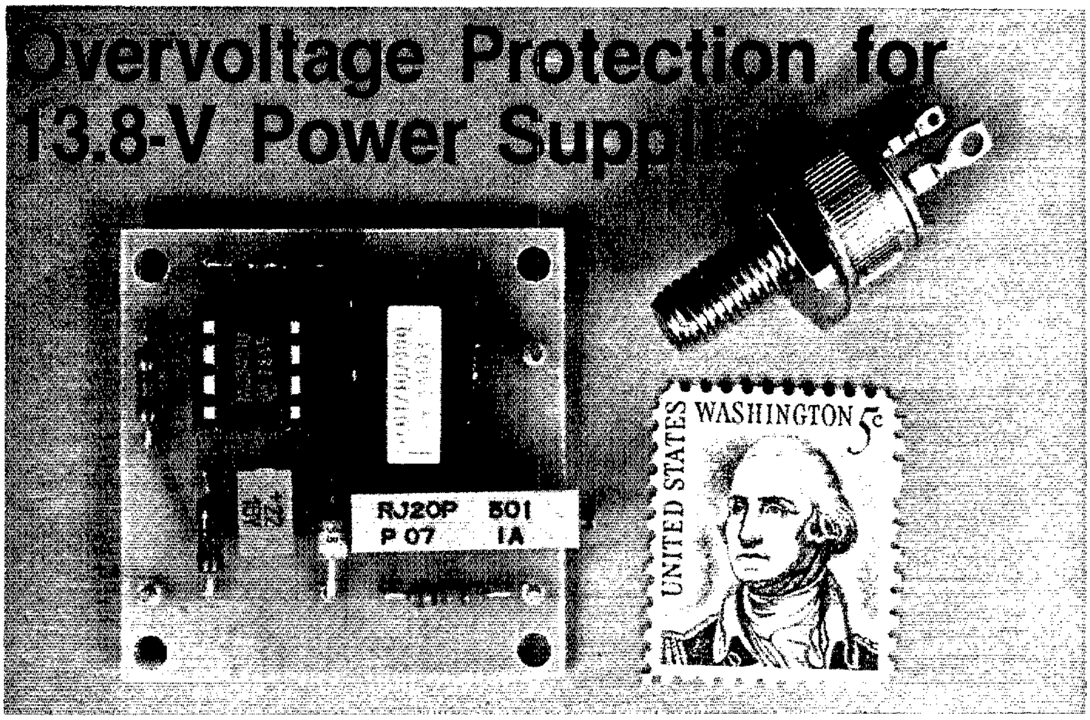
Kuo, F. F., ed. *Protocols and Techniques for Data Communications Networks*. New York: Prentice Hall, 1981.

McNamara, J. E. *Technical Aspects of Data Communications*. Maynard, MA: Digital Equipment Corp., 1977.

Overvoltage Protection for 13.8-V Power Supplies



Failure of the series pass transistor in regulated power supplies can put a destructively high voltage across the load — your rig! Here's how you can avoid that situation.

By Ian N. Cousins,* VK5IK

I used to think that power supply faults causing an overvoltage at the output terminals were simply the result of poor components or workmanship. Time has proved me wrong!

My first introduction to the problem came when I received a call telling me that a power supply had "gone high" and worked a nasty mischief on a remote base-station transceiver operated by our State Police. To further compound the indignity, the service technician hadn't checked the power-supply output voltage before he connected a brand new transceiver to the power source! The power supply very quickly administered the coup de grace to the second transceiver. Both radios required extensive rebuilding of the final and driver stages. In this case, the failure was caused by a Darlington driver transistor that developed

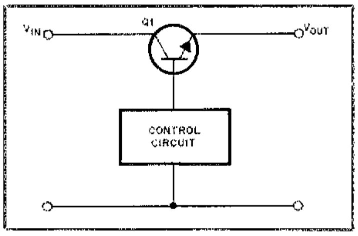


Fig. 1 — The basic series voltage regulator used in most power supplies.

gross collector-to-emitter leakage, effectively turning the series transistors "full on," regardless of the control signals from the regulator IC.

To prove that my troubles were not isolated, we had two other base-station-supply failures caused by lightning strikes. In each case, the output voltage went high, gassed the parallel battery dry and eventually damaged the base station.

What Causes This Problem?

Refer to Fig. 1. This is the basic diagram of a series voltage regulator. Q1 stands between the input voltage and the regulated output voltage. The control circuit drives Q1 so as to "soak up" the difference in voltage between V_{in} and V_{out}. It's easy to see that a short circuit in the transistor will put unregulated voltage where there should be regulated voltage! (Most practical supplies will have a low-value resistive component or two between V_{in} and V_{out}, but their contribution to minimizing damage is negligible.)

Since the potential results are likely to be so nasty, what's the chance of a short circuit in the series transistor? To make a start in finding the answer, I consulted that archive of failures — my transistor graveyard. Of 44 silicon power transistors collected during power-supply work over the past decade, one was completely open-

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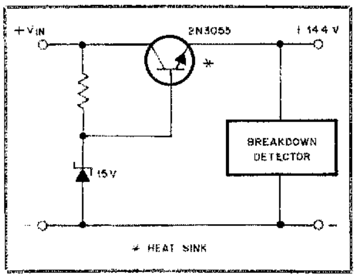


Fig. 2 — A simple regulator circuit used for transistor stressing.

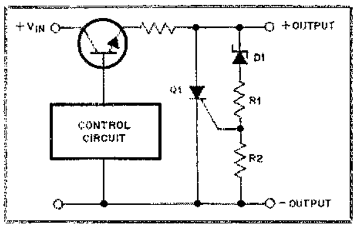


Fig. 3 — A common overvoltage-protection circuit.

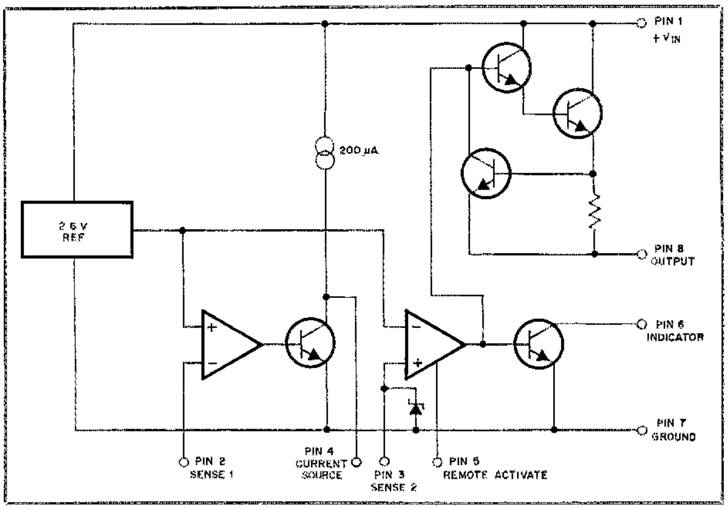


Fig. 4 — Block diagram of the MC3423 overvoltage- protection IC.

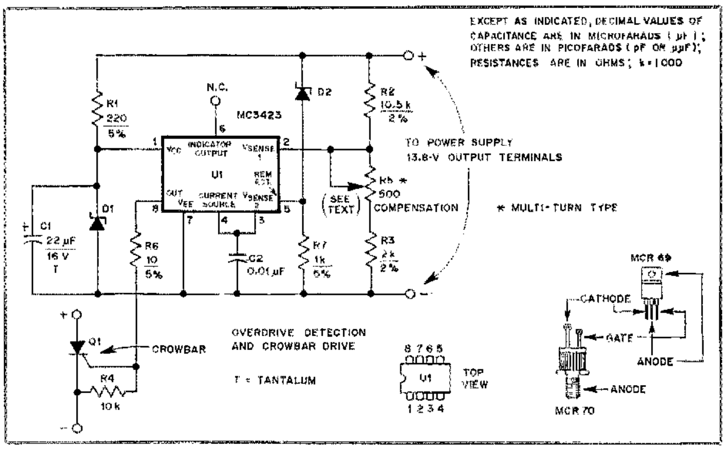


Fig. 5 — Schematic diagram of the overvoltage-protection circuit described in the text. Resistors are 1/4-W, 5% film types, except for R2 and R3, which are 2% metal-film units. R4 is mounted close to the SCR terminals. It reduces the possibility of Q1 triggering because of noise or leakage.

C2 — 0.01-μF, 50-V polyester film.
D1 — 8.2-V, 1-W, 5%-tolerance Zener diode. Motorola 1N4738A, Philips BZX87-C8V2 or equiv.
D2 — 15-V, 400-mW, 5%-tolerance Zener diode. Motorola 1N5245B or Philips BZX79-C15 or equiv.
Q1 — Crowbar SCR, Motorola MCR70-2, 50-V, 850-A peak discharge current metal-can SCR or equiv. If the power supply unregulated input voltage exceeds 50 V, use an MCR70-3, which is rated at 100 V. Alternatively, the MCR69-2 (50-V, 750-A peak) or the MCR69-3, rated at 100 V, can be used. A more expensive alternative is the MCR71-3, rated at 100-V, 1700-A peak discharge current.
U1 — Overvoltage-protection IC. The MC3423P (Motorola and Texas Instruments) and SG3423M (Silicon General) are in plastic eight-pin DIL packages. The MC3423U (Motorola), MC3423JG (Texas Instruments) and SG3423Y (Silicon General) are in ceramic eight-pin packages.

circuited, nine medium-power transistors used in inverters were shorted between collector and base, and the remaining 34 devices were shorted between collector and emitter. This was an indication that short-circuit failures from collector to emitter were fairly common.

To try to better understand the likely nature of transistor failure in amateur service, I constructed the simple regulator circuit shown in Fig. 2. The transistor is an amateur favorite (the 2N3055), although it seems reasonable to expect that other high-power, low-frequency devices would behave similarly.

The 2N3055 was mounted centrally on a piece of 14-gauge (1.6 mm) aluminum measuring 8.3 × 4.8 inches (210 × 123 mm), suspended on nylon line. I then set out to thermally overload the transistor to the point of failure. Clearly, overvoltage punch-through will produce a short circuit, so the collector-to-emitter voltage was kept low for these tests. I was more interested in what happens when the transistor fails than the processes leading to failure.

Four 2N3055 transistors of different manufacture were abused. All failed with collector-to-emitter short circuits after periods ranging from two minutes to over five hours. Interestingly, during the test several devices tripped the detector circuit, indicating failure. But the transistors tested okay after cooling to room temperature. Complete failure occurred shortly after resuming operation.

It seems reasonable to conclude that if a series transistor fails, it is likely to fail as a short circuit, putting unregulated input voltage on the supply output terminals. In addition, failure of voltage regulator ICs or driver transistors can also lead to an output overvoltage condition.

How Do We Protect Our Radio Equipment?

The common ("crowbar") overvoltage-protection circuit (OVP) is shown in Fig. 3. D1 is a low-power Zener diode, chosen to have a voltage rating a little above the power-supply output, say 15 V. R1 limits the current through the Zener diode, and R2 bleeds off any leakage through it. If the power-supply output voltage exceeds the Zener diode voltage plus the gate-to-cathode voltage of Q1, Q1 fires and short circuits the output, eliminating the overvoltage by clamping the output terminals. A typical Zener diode/SCR combination will give a trip level of about 15.5 V, a fairly satisfactory figure, being well above the normal 13.5- to 13.8-V power-supply output level and a little below the normal damage threshold for radio gear. Unfortunately, the difficulty with this circuit is that of predicting the voltage level at which all likely combinations of Zener diodes and SCRs will cause a "trip." Even assuming the tightest commercially available tolerance for the Zener diode (5%), the permissible variation of SCR trigger voltage means that some combinations will trip as low as 14.0 V, probably operating every time the power-supply load changes, to as high as 19.2 V — so high that it may not even operate in the event of a shorted series transistor, providing only falsely based peace of mind. We have not considered the effects of the variation of SCR trigger current or turn-on speed, so this circuit is probably best avoided unless the load to be protected is of little value, or if equipment exists to measure actual operation.

A Comprehensive Overvoltage-Protection IC

The MC3423 has been around for a few years, but seems to have been largely overlooked in amateur applications. This IC was developed specifically as an overvoltage-protection device. Fig. 4 shows a block-diagram of the MC3423. It contains a precision 2.6-V reference, two comparators and an SCR "crowbar" driver.

Fig. 5 is a circuit diagram of a complete overvoltage-protection system designed around the MC3423. R1, C1 and D1 develop a regulated and transient-free supply of about 8.2 V to the IC, so that the internal reference sees a stable input voltage. The internal reference voltage of the IC, the point at which it starts to "do its stuff," is nominally 2.6. So we divide the power supply terminal voltage with R2 and R3 so the normal level at pin 2 is slightly less than 2.6 V. R5 allows compensation for the IC reference voltage production tolerance.

We have to select a voltage at which the protection circuit will operate. This voltage should be sufficiently above the normal power-supply level so that the circuit doesn't operate every time there's a "hiccup" — turning the main switch on or off, etc. — but the voltage has to be sufficiently low to provide meaningful protection.

For a power supply with a 13.8-V output, an overvoltage protection operating point of 14.75 V is ideal. If the sample of the output voltage applied to pin 2 exceeds the reference voltage, indicating an output voltage in excess of 14.75, an internal circuit starts a time delay, determined by C2. This time delay gives the circuit an amount of noise immunity — it ignores insignificant hiccups. A value of 0.01 μF for C2 gives a delay of 0.1 ms, 0.1 μF provides 1 ms, and so on. If the overvoltage condition disappears during this time delay, the circuit resets itself.

If the overvoltage is still present after the time delay, a fast-rising, high-current pulse is sent from pin 8 to the gate of a crowbar SCR. The SCR is capable of passing large pulse currents and, once triggered, will stay turned on as long as current flows through it.

D2, connected to pin 5, provides a bypass of the time-delay function. If the power-supply voltage reaches approximately 16.4, the circuit operates immediately, firing the crowbar.

Component Selection

This circuit performs an important protective function, so use the best available components during construction. Metal-film, 2%-tolerance resistors for R2 and R3 and a good quality Cermet or wire-wound trimmer for R5 should be used; avoid old-style surplus trimmers. Supplying a socket for IC1 will prevent exposing it to soldering temperatures.

A number of SCR types can be used, but the Motorola MCR69, MCR70 and MCR71 devices were developed specifically to handle the very high peak currents experienced in crowbar service. Check that any alternative device used has equivalent trigger-voltage and current requirements. The SCR does not require a heat sink, since it should conduct for only milliseconds before the fuse opens, but a small heat-sink bracket may provide a convenient mounting arrangement.

Construction and Setup

The circuit is built easily on perf board or a pc board.[1] Before installation, the overvoltage trip point must be set at 14.75 V. A test setup is shown in Fig. 6. A digital multimeter will provide optimum accuracy; if you don't have one, use a good analog instrument. Set the variable power supply for its lowest voltage level and switch it on. Slowly increase the voltage, while carefully watching the meter. If the dial lamp lights below 14.75 V, readjust the trimming potentiometer (R5) so that the lamp lights as the voltage reaches 14.75. If the power-supply voltage reaches 14.75 and the lamp is still out, adjust R5 until the lamp lights with the power supply held at 14.75 V. Check the operation several times to ensure the circuit operates as increasing voltage reaches the 14.75-V level.

[1]A professionally made glass-epoxy pc board and one each 10.5-kΩ, 1% and 2-kΩ, 1% metal-film resistors are available from the author for $5 U.S. currency, including return Air Mail postage. Personal checks are not accepted. Please allow 2 to 4 weeks for delivery. The ARRL and *QST* in no way warrant this offer.

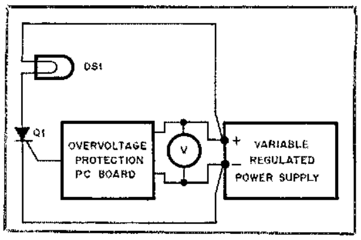


Fig. 6 — Test setup for adjusting the overvoltage-protection-circuit operating point. DS1 is a 12-V dial lamp. Q1 is identified in the parts list of Fig. 5. The voltmeter is discussed in the text.

If the circuit operates at less than 14.75 V, regardless of adjustment of R5, it is possible that the actual voltage of D2 is on the low side. Lift one end of D2 and repeat the adjustment procedure. The circuit should now act normally. Replace D2 with another 15-V device.

How Do We Use the Protection Circuit?

Since the power for our equipment comes from the power-supply output terminals, that's the correct place to connect the IC and related components. However, positioning the crowbar SCR requires more care. Putting the SCR across the output terminals will stop damage to equipment connected to the supply, but what happens after the SCR fires? Over-voltage on the output is most likely to be a result of failure of the series pass transistor or the voltage-regulator IC. In either case, it is unlikely that the normal power-supply current-limiting function will still operate. So, we would have a power supply with inoperative overload protection looking into the near short circuit of the conducting SCR. Under these circumstances we would have to rely on blowing the transformer primary mains fuse. Unfortunately, this fuse has to be selected to withstand repeated turn-on inrush current of the power supply. It is thus unlikely to blow instantly, given the wiring resistance and current-limiting resistance in the power supply. My experience is that a rectifier shorts first and that may, or may not, blow the fuse before the transformer burns out!

The circuit of Fig. 7 overcomes these problems. The power-supply voltage is sensed at the output terminals. A fuse is placed between the main filter capacitor and the voltage regulator. The crowbar SCR is placed on the output terminal side of the fuse. If the SCR fires, a large amount of energy is available from the transformer to blow the fuse immediately and so open the positive line to the regulator. In this position, the fuse does not have to withstand significant inrush current because most regulator circuits turn on slowly and the fuse can be chosen to handle the normal full-load current of the supply, plus a small safety margin.

D1 discharges any capacitance on the output of the regulator. If your power supply uses a three-terminal regulator with the adjustment pin bypassed, D2 will also be necessary to discharge the bypass capacitor.

If you "float" a battery across the output of your power supply, a diode will be needed between the output of the power supply and the battery, so the battery is not shorted via D1 and Q1. A Schottky diode will have the lowest voltage drop.

Installation

Install the new components in the circuit position shown in Fig. 7. Put the crowbar SCR near the main filter capacitor and the series fuse. Ensure that the positive supply for all regulator devices is taken from the output side of the fuse, including collector voltage for any Darlington-connected driver transistors and the supply voltage to the regulator IC. The overvoltage-protection IC and its associated components can be tucked away with sensing leads running to the output terminals. Keep the circuit away from hot areas.

Make sure the fuse holder is able to carry the required current. Most small plastic units can handle little more than 5A. Do not use a fuse with too large a current rating. When the SCR fires and short circuits the transformer secondary, the fastest possible clearing of the fault current is desired to minimize any possible damage to the transformer and rectifiers. Above approximately 7.5 A, a fast-acting circuit breaker is an attractive alternative to the fuse.

Troubles

If the protection circuit operates when your power supply is switched on or off, you most likely have transient-overshoot problems in the supply. The time-delay bypass Zener diode, D2, is very quick to actuate the circuit under transient overvoltage conditions, but it is better to improve the power supply than to reduce the level of protection.

Here's an effective fix for transient overshoot, especially in switching supplies. Series-connect a 1N5342B 6.8-V, 5-W Zener diode and a 1N4343B 7.5-V, 5-W Zener diode. Check that the reference voltage of the combination is above about 14.1 V and below the 14.75-V operating point of the overvoltage-protection circuit. Connected across the output of a power supply — observing the correct polarity — this combination limits transient overshoot to a safe level.

In repeater installations, where lightning strikes are common, transient sensitivity can be reduced by omitting D2 and increasing C2 to 0.1 μF, but *only* if a storage battery is floated across the power-supply output to dampen momentary transients. Refer to the comments made earlier regarding diode isolation of a floating battery.

I have used this overvoltage-protection circuit since late 1977 in applications ranging from the State Police base stations that fueled my original interest, to dc-to-dc converters in mining equipment. I'm sure you'll find it to be a valuable addition to your power supply.

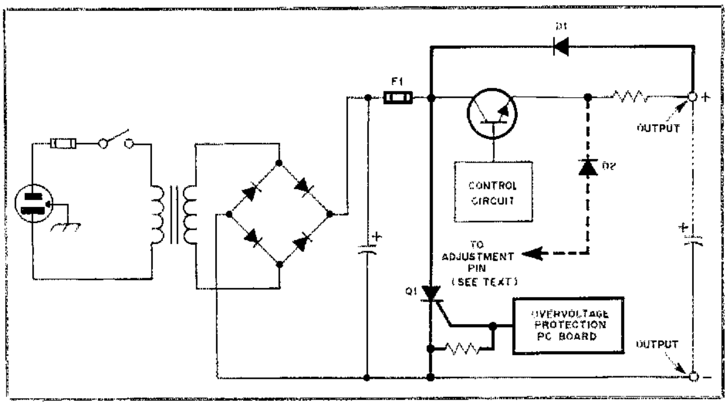


Fig. 7 — Overvoltage-protection circuit component position within a power supply. Added components are shown boldface. D1 serves to discharge the power-supply output capacitor. D2 discharges the adjustment pin capacitor if a three-terminal voltage regulator IC is used with the adjustment (or ground) pin bypassed. F1 is a standard 3AG fast-blow fuse with a current rating equal to the steady state rating of the power supply, plus a small margin; e.g., a 5-A fuse for a 4-A power supply, and a 10-A fuse for a 7.5-A supply. For power supplies with higher current ratings, a small circuit breaker can be substituted for F1.

D1 — MR751 (Motorola, 100-V, 400-A surge), 1N5401 (various manufacturers, 100-V, 200-A surge).
D2 — 1N4002 or equiv.
F1 — See text and caption. Circuit breakers such as the Airpax Electronics T11-1-XXXA-02-11AL or Heinemann Electric JA1-B3-XX-3 series may be used. The required current rating is inserted in place of the Xs; e.g., a T11-1-12.5A-02-11AL is a 12.5-A unit, and a JA1-B3-15-3 is a 15-A circuit breaker.
Q1 — Refer to Fig. 5.

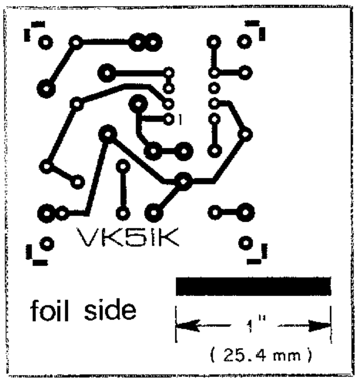


Fig. 8 — Pc-board etching pattern for the overvoltage-protection circuit. Black represents copper foil. The pattern is shown full-size from the foil side of the board. The parts-placement guide appears in Fig. 9.

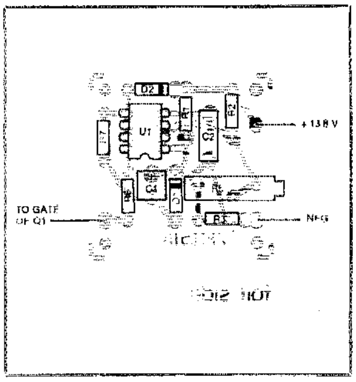


Fig. 9 — Parts-placement guide for the overvoltage-protection circuit. Components are mounted on the nonfoil side of the board. Shaded areas represent copper on the foil side of the board.

Ian Cousins was first licensed in 1960. He attended Adelaide University, leaving to work in electronic equipment and component sales for four years. Ian then started his own business, designing and building audio amplifiers and radio tuners for long-distance passenger locomotives. Since 1971, he has been involved in the design and manufacture of power-supply systems for radio equipment, primarily dc-to-dc converters that convert the high battery voltage of heavy traction vehicles to 13.8-V dc.

Hints and Kinks

Conducted By Larry D. Wolfgang,* WA3VIL

A NEW MIXER FOR THE ASTRO 103 RECEIVER

☐ When I purchased a Cubic Astro 103, I compared its performance with that reported in the *QST* Product Review column for the Astro 102BXA.[1] My results were much the same (−13 dBm third-order intercept and the blocking DR not measurable because of noise limiting), and indicated to me that some improvement to the receiver front end should be possible.

When I studied the circuit diagrams, I found that the rf amplifier, the first mixer and the post-mixer amplifier were also part of the transmitter. Everything seemed okay on transmit, so I decided to switch in a new circuit for receive only. My plan was to add a new mixer and a post-mixer amplifier, along with the required power supply and switching scheme.

After reading everything I could find about mixers, and making some performance checks, I decided that the Plessey SL6440C device would suit my application quite nicely. It boasts a third-order intercept of +30 dBm and requires an LO signal level of about 300 mV. I used a balanced circuit to reduce the LO leakage to a minimum, and used 1:4 input/4:1 output transformers to

[1]D. DeMaw, "Cubic Astro 102BXA," Product Review, *QST*, Dec. 1981, pp. 48-49.

*Assistant Technical Editor

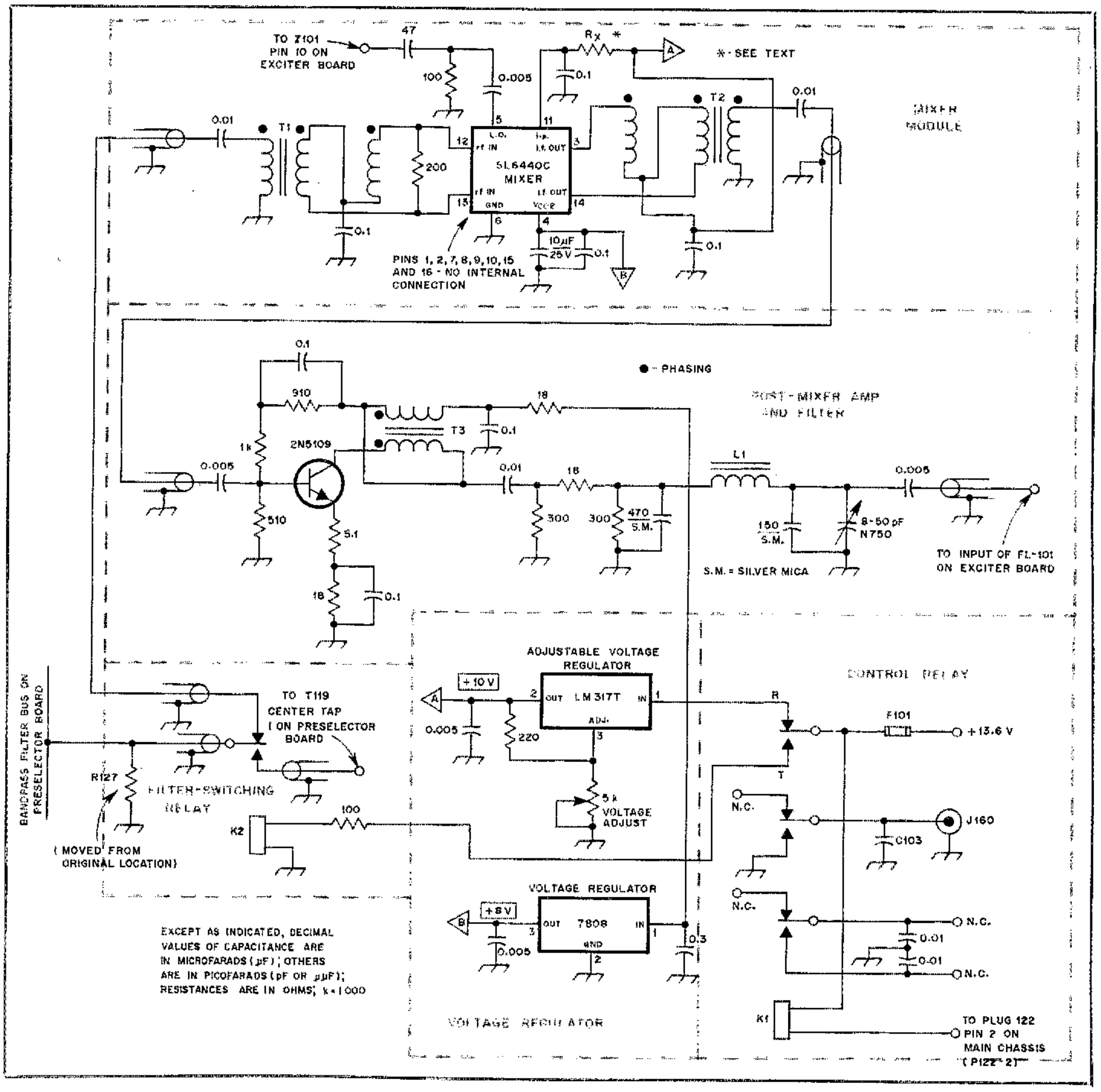


Fig. 1 — Schematic diagram of the receiver-circuit modifications that AG4R made to his Astro 103. The circuit is built as four separate modules, as indicated.

K1 — 3pdt relay, 12-V, 300-Ω coil.
K2 — spdt relay, 5-V, 50-Ω coil (Radio Shack part no. 275-215).
L1 — 22 turns of no. 28 wire on a T50-6 toroid core.
T1, T2 — Nine trifilar turns of no. 32 wire on an FB-801-43 ferrite bead.
T3 — Ten bifilar turns of no. 32 wire on an FB-801-43 bead.

provide the necessary gain and impedance matching. Fig. 1 is a schematic diagram of the circuit that I added to my receiver. R_X should be adjusted for the best IMD level using a two-tone, 0-dBm input signal to the mixer. After this adjustment is completed, you can substitute a fixed resistor of the proper value. I used a 200-Ω resistor here. I found that about 23 mA at pin 11 of the SL6440C gave the lowest IMD level for the voltages in my circuit and the mixer samples I checked. I found it necessary to test several mixer chips to select one with the best characteristics. I also found it important to hand select the 2N5109 transistor in the post-mixer amplifier. The tested IMD had wide variations from chip to chip and from one transistor to another!

The post-mixer amplifier and 50-Ω input to 500-Ω output low-pass filter designs were borrowed from previous designs, and so took no special engineering skills.[2,3,4] A 3-dB attenuator pad was included to provide a gain of 1, as compared to the original circuit. This attenuation can be varied to adjust the overall gain of the modified circuit. A step attenuator is a handy test item for this experimentation. The pad and low-pass filter are designed to present a 50-Ω impedance to match the amplifier output impedance. This is important for best overall operation of the amplifier.

My voltage regulator and control-relay circuit are straightforward and easy to build. My radio came with an spst relay in the line to control an external relay, even though the schematic diagram gives no indication of this. It was a simple matter to replace this relay with a 3pdt unit (K1) that I had on hand, to provide the desired internal and external switching and relay control features. The extra set of contacts can be brought out to another jack and used for antenna or amplifier switching.

K2 is used to switch between the new mixer circuitry on receive and the original circuit on transmit. This is accomplished easily by breaking the lead to T119 on the preselector board and inserting the relay contacts. One other change is required. I found it necessary to move the location of R127 as shown in Fig. 1, because this resistor is part of the PIN diode switching arrangement to select the proper band-pass filter.

I built the circuit on four pieces of circuit-board material. The mixer, the post-mixer amplifier and filter, the voltage regulator and the control-relay/filter-switching unit were each constructed and tested as separate modules. I used double-sided pc-board material for the mixer module and single-sided material for the others. I drew the layouts on the boards freehand, with consideration for the parts that I had available. The mixer IC needs a heat sink for this application. Mine is made from flashing copper and is attached to the IC by means of Super Glue®. The construction details are given in Fig. 2.

Each module is mounted by means of a single angle bracket and an existing mounting bolt. This provides a good ground connection between each module and the main circuit. The modules are mounted as close to the boards they connect to as is practical. You can see the mixer and amplifier/filter modules in Fig. 3A. The voltage regulator and control-relay modules are shown

[2]D. DeMaw and G. Collins, "Modern Receiver Mixers for High Dynamic Range," *QST*, Jan. 1981, pp. 19-23.
[3]P. Chadwick and D. DeMaw, "Receiving With Plessey ICs," *QST*, April 1981, pp. 13-15.
[4]W. Hayward and J. Lawson, " A Progressive Communications Receiver," *QST*, Nov. 1981, pp. 11-21.

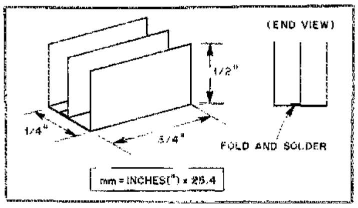
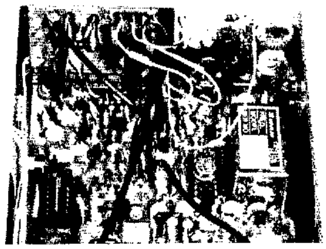
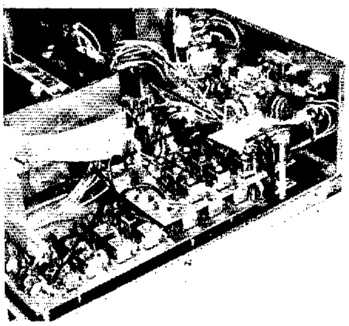


Fig. 2 — Construction details of a heat sink suitable for attaching to the mixer IC with Super Glue®.



(A)



(B)

Fig. 3 — The mixer and post-mixer amplifier modules can be seen at the top of A. Notice the homemade heat sink on the mixer IC and the conventional heatsink on the post-mixer-amplifier transistor. The voltage regulator and control-relay circuitry can be seen as mounted in the transceiver at B.

in Fig. 3B. All rf interconnections are made with RG-174/U miniature coaxial cable.

These simple modifications seem to have made quite an improvement to the receiver. I measured a third-order intercept point of +17 dBm and a blocking DR of 116 dB. My test measurements were made following the method described in *The Radio Amateur's Handbook* so that a comparison could be made with results published in the *QST* Product Review column over the last few years.[5] This method uses two signals 20 kHz apart at the receiver input. My measurements were made on the 20-meter band.

I don't claim to have the "Ultimate Receiver," but I am pleased with the results of my modification. For my application, I have not found an rf-gain control or input attenuator to be necessary with this receiver. These options may suggest further experimentation for another

[5]G. Woodward, ed., *The Radio Amateur's Handbook* (Newington: ARRL, 1982), Chs. 8 and 16.

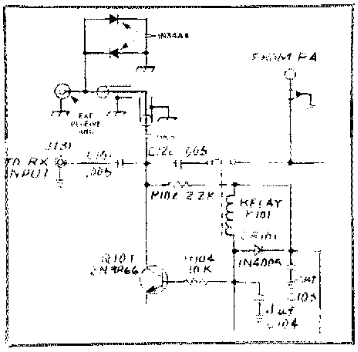


Fig. 4 — Part of the Astro 102BX Low-Pass Filter/Bandswitch Board is shown. The additional components required to add external receive antenna capability are shown darker than the main diagram.

amateur. — *Wayne Cooper, AG4R, Miami Shores, Florida*

SEPARATE RECEIVE ANTENNA FOR THE CUBIC ASTRO 102BX

☐ One feature that my Astro 102BX did not have, but which the newer Astro 103 did, was a jack for an external receive antenna. I checked with Cubic about adding this feature, and was told that it was not possible because the preselector board had been redesigned for the 103. I decided to study the schematic diagrams myself to see how I might modify my 102BX.

My investigation led to the discovery that I should be able to add an external receive antenna quite easily. All that is necessary is to tack solder a 0.005-μF capacitor to the top of R102 on the Bandswitch/Low-Pass Filter Board. Use a short length of RG-174/U miniature coaxial cable to connect to the new jack that is installed on the back panel, and solder two diodes from the center conductor to ground for protection against overload. See Fig. 4 for a section of the transceiver diagram. The new components appear a bit darker than the rest of the circuit.

This modification bypasses the low-pass filter section, but these filters are not required on receive anyway. I have found no ill effects on the normal operation of the radio, and I am pleased to be able to use the rig with vhf receiving converters. — *Robert Whitford, WA7STA, Ellensburg, Washington*

PROTECTIVE COATING FOR DTMF KEYPADS

☐ Several of the hams in my area who purchased ICOM hand-held transceivers began to have a similar problem. The DTMF keypad identification markings were wearing off. Worn keypads can be replaced, but this requires sending the unit back to the manufacturer or taking it apart yourself.

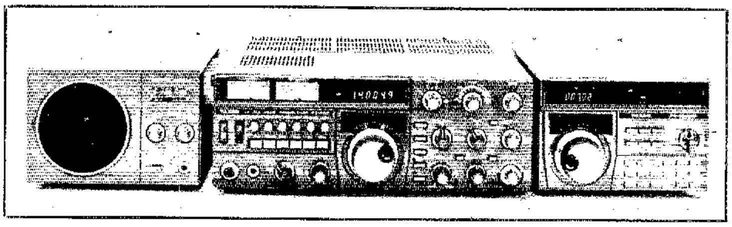
We tried various methods to protect the markings, but the best one seems to be simply coating the keypad with a clear lacquer spray. Mask off the area around the pad with tape to protect the rest of the rig. Two light coats of spray provide adequate protection. I have experienced no ill effects to the operation of the keypad, and the finish is like new after months of use. — *W. E. Wiehe, WD9BBI, Plainfield, Illinois*

Product Review

Conducted By Paul K. Pagel,* N1FB

Yaesu Electronics Corp. FT-102 HF Transceiver

I have observed a steady and impressive improvement in the performance of each new model of Yaesu transceiver. The receiver dynamic range has increased markedly, and the transmitter spectral purity has improved similarly. The FT-102 compares favorably in this respect to some other present-day, high-performance commercial transceivers. But most importantly, there is no performance similarity between the FT-102 and the earlier FT-101B and E models. The '102 provides fully competitive performance!



Yaesu Electronics Corp. FT-102 Transceiver, Serial No. 2J-050152

Manufacturer's Claimed Specifications	*Measured in ARRL Lab*
Frequency coverage: 160 through 10 meters, plus WARC bands.	As specified, plus additional coverage above and below each band: 1453-2032, 3453-4032, 6953-7532, 9953-10,532, 13,953-14,532, 17,953-18,532, 20,953-21,532, 24,453-25,032 and 27,953-30,032 kHz.
Operating modes: Cw and ssb (a-m and fm optional).	As specified.
Readout: Blue 5/16-inch digital-display numerals.	As specified.
Resolution: 100 Hz.	As specified, 18 kHz per 360° dial rotation.
Power requirements: 100, 117, 200 or 234-V ac, 50/60 Hz.	As specified.
Power consumption: 95 VA receive and 440 VA transmit.	As specified.
Transmitter rf power output (cw with 50-ohm load): 100 W or greater.	150 W on 160-15 meters; 100 W on 10 meters
Transmitter third-order IMD: Better than −40 dB at 14 MHz.	−40 dB
Spurious suppression: Better than −40 dB.	Approximately −44 dB worst case (10.1 MHz).
Frequency stability: Less than 300 Hz during first 30 minutes and less than 100 Hz each 10 minutes thereafter.	Less than 200 Hz during first hour of operation.
Audio output (receiver): 1.5-W minimum at 8 ohms, 10% THD.	As specified.
RIT range: Not specified.	± 3 kHz.
Receiver notch depth: Better than −40 dB.	As specified.
S meter (μV for S9): Not specified.	160 m — 50; 80 m — 43; 40 m — 43; 30 m — 42.5; 20 m — 42.5; 15 m — 39.5; 10 m — 30.
Receiver dynamic range (preamp off): 95 to 102 dB, depending on filter used.	Receiver dynamics measured with optional narrow cw filter installed.

	80 m	*20 m*
Noise floor (MDS) dBm:	−127	−127
Blocking DR (dB)†:		
Two-tone, third-order IMD DR (dB):	96.5	97.5
Third-order input intercept (dBm):	18	19.5

Size (HWD): 5 × 14.5 × 12 inches (129 × 368 × 309 mm).	As specified.
Weight: 33 pounds (15 kg).	As specified.
Color: Not specified.	Dark gray.

†unmeasured — noise limited

Some of the Features

There seems to be no important feature missing. A notable feature is the fm-mode capability. An accessory a-m/fm module can be installed to permit operation in these two modes. The fm function allows the operator to employ fm on 10 meters. It also provides for vhf and uhf fm operation by means of a suitable transverter. A squelch control is included as a standard feature of the FT-102.

Not two, but *three* 6146B tubes are used in the transmitter PA! This permits driving external power amplifiers at the prescribed excitation level without operating the FT-102 PA at or beyond the nonlinear range. This is a shortcoming with many commercial exciters, causing wide signals of poor quality on our amateur bands. The dc input power to the 6146Bs is rated at 240 W (ssb or cw) from 1.8 to 25 MHz. The limit is 160 W from 28 to 29.9 MHz. During SSTV and fm operation, the dc input power is restricted to 120 W on all frequencies. It is reduced to 80 W for a-m transmissions.

The stock i-f filter has a 2.7-kHz bandwidth at the −6 dB points on the response curve. This can be reduced to as low as 500 Hz by means of the SHIFT/WIDTH control. An ssb "narrow" filter is available as an accessory. It yields a 1.8-kHz bandwidth. Additional filters are offered for cw reception in a variety of bandwidths: 600, 500, 300 or 270 Hz. The accessory a-m filter bandwidth is 6 kHz.

An i-f NOTCH control is included. It provides a notch depth greater than 40 dB for reducing the effects of QRM. This, when used separately or in combination with the built-in R-C active audio filter (frequency-variable), has proven to be a tremendous asset when dealing with interference on the cw bands.

Two panel meters are employed in the FT-102. One of them indicates the relative signal strength (S meter) and has a scale for use when setting the alc level correctly. The remaining meter, by virtue of a front-panel switch, can be used to read the high voltage, PA current, relative output power and speech-compression level. If the a-m/fm accessory adapter is installed in the transceiver, this meter serves as a discriminator-tuning indicator.

Some of the controls that are used infrequently are accessible from the front panel, but once adjusted, they are recessed into the panel, where they are safe from accidental "readjustment." A slight inward pressure pops the control knobs out so that adjustment can be accomplished. Pushing gently upon the tips of the knobs will cause the controls to recess and lock in that position. The functions controlled by these knobs are VOX/GAIN, VOX DELAY, MIC GAIN, NB LEVEL and SQUELCH. A row of push-button switches below them permits operator selection of MOX, RF AMP, NARROW filter, PROCESSOR, NB and CW MONITOR.

A vertical row of push buttons at the right of the main-tuning knob provides control of the AGC ON/OFF, AGC FAST/SLOW, ALC metering, RX RIT and TX RIT. Another push-button switch can be used to provide a 500-kHz "upshift" for use

*Assistant Technical Editor

during 10-meter operation. The audio filter and notching features are actuated by means of two additional push-button switches.

Transmitter Tuning

Four tuning controls are involved when adjusting the transmitter to the operating frequency. They are labeled PLATE, LOADING, DRIVE and PRESELECT. The latter is used also for receiver front-end peaking when the 10-dB switchable preamplifier is activated. A definite tuning procedure for ssb operation is spelled out in the operating manual. If the method is followed as prescribed, signal purity will be assured. Generally, it calls for keeping the PA plate current below 300 mA at resonance. This will yield up to 100 W of rf output. During cw operation, the plate current may be increased to 350 mA, which results in substantially greater output power. The maximum output obtained in the ARRL lab was 175 W on the bands below 10 meters.

Receiver-Performance Observations

Real-life testing of the FT-102 was done at the usual severe (former) W1FB proving grounds — two blocks from the simultaneous multiband high-power onslaught of W1AW, which has its 20- and 40-meter Yagis bore-sighted over my house. If ever a receiver will collapse from strong signals, it's at that location! My worst band for survival is 80 meters: I have measured a W1AW signal level of 5-V peak to peak across a 50-ohm termination at the transmitter end of my 80-meter antenna! Only a few commercial receivers could handle this without 10 or 20 dB of front-end attenuation switched in. The FT-102 fared well under these conditions. No problems were noted when the preamplifier was not in use. In fact, I was able to copy weak signals within 5 kHz of the W1AW frequency. When the preamplifier was actuated, there was some evidence of cross-modulation on signals lower than S9, but they were still readable. For the most part, the preamplifier is not needed on the bands below 15 meters, since the atmospheric noise usually exceeds that of the receiver anyway.

I detected no dynamic-range problems on the bands above 80 meters, even with the preamplifier operating. The notable exception was when I pointed my 20-meter beam antenna directly at the stacked 20-meter Yagis of W1AW. The problem was resolved when the preamp was turned off.

The audio quality of the receiver is good, even at relatively high output levels into a speaker. There is no evidence (by ear) of reciprocal mixing (buzz between signals when a strong signal is within the receiver passband) when using the FT-102 local oscillator or the outboard FV-102DM memory/synthesizer LO unit. This is important when strong signals are being handled by a receiver, irrespective of how high the receiver dynamic range might be.

Performance of the built-in audio filter is similarly good. There is no ringing evident, and the frequency peaking is sharp, although easy to adjust. Use of this filter greatly reduces the wide-band noise from the overall system, which provides a useful noise-reduction benefit. The audio filter can be switched in during all cw operation. It can't be used for phone operation, and is automatically disabled when changing from the cw to the voice mode.

Noise Blanker

As is the case with most blankers, this one works well on sharp impulse noise, but is ineffective in reducing ordinary QRN. It is not useful when the blanking level is advanced full on, at which time all signals become limited and distorted. The best level setting seems to be about "12 o'clock."

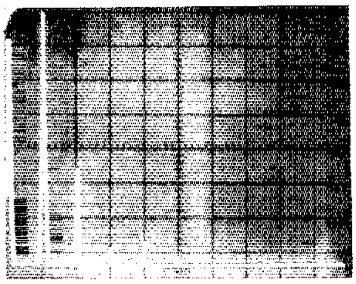


Fig. 1 — Worst-case spectral display of the Yaesu FT-102. Vertical divisions are each 10 dB; horizontal divisions are each 10 MHz. Output power is approximately 150 W at a frequency of 10.105 MHz. All spurious output is at least 45 dB below peak fundamental output. The FT-102 complies with current FCC specifications for spectral purity.

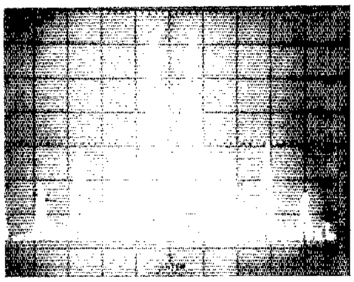


Fig. 2 — Spectral display of the FT-102 output during transmitter two-tone IMD testing. Third-order and fifth-order products are about 40 dB below PEP output. Vertical divisions are each 10 dB; horizontal divisions are each 1 kHz. The transceiver was being operated at rated input power on the 20-meter band.

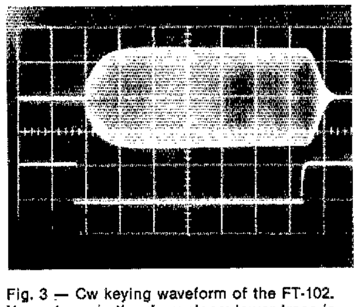


Fig. 3 — Cw keying waveform of the FT-102. Upper trace is the rf envelope; lower trace is the actual key closure. Each horizontal division is 5 ms.

Rear-Panel Ports and Switches

A SEP/NORM antenna switch is located on the rear wall of the transceiver. When in the SEP mode, an outboard receiver may be bridged to the FT-102, permitting the main station antenna to be used with the transmitter and outboard receiver. The FT-102 receiver has no antenna connected under this condition. But, a separate antenna can be attached to the FT-102 receiver if it is connected to the phono jack labeled ANT, near the switch. In the NORM mode, there is no internal connection to the EXT RCVR jack, and the main antenna is used both for transmit and receive with the FT-102.

An RF OUT jack provides low-level transmitter output for use with a transverter. The output level of the energy is 0.1-V rms (-7 dBm) at 50 ohms. There is also a jack for connecting the FV-102DM outboard VFO (synthesizer) to the transceiver.

A seven-pin DIN jack permits muting an external receiver and supplying sidetone to it. The FT-102 scanning signals can be picked off at this jack for external use. Another jack, the ACC-1, is a six-pin DIN type that permits the operator to utilize the FT-102 control circuits to be used with a transverter.

The ACC-2 socket is a five-pin DIN unit that allows T-R switching and alc input-control connections for use with a linear amplifier. In addition to this socket are a number of phono jacks that provide phone-patch input, wide-band i-f output, constant-level af output (for a recorder), foot-switch control and narrow-band i-f output for use with a monitor scope. There is a 12-V, low-current dc output jack for interface with auxiliary equipment that requires that operating voltage.

Inside a removable bottom cover on the '102 are controls for adjusting the side-tone level and pitch. There is another control that can be adjusted to boost the high-frequency response from the microphone before it is routed to the transmitter modulator. A low-frequency boost control is also available for shaping the audio response.

Comments on Performance

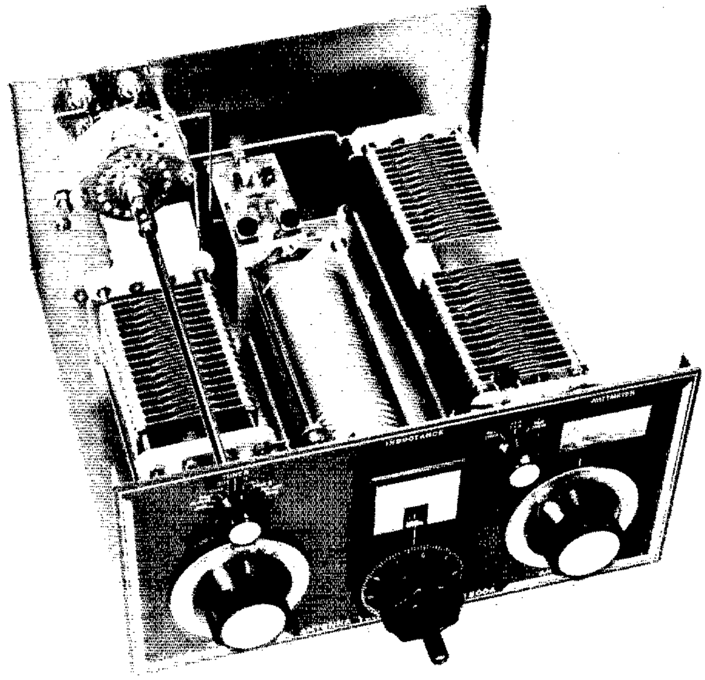
A frequency-jumping problem was observed (30 to 50 Hz) at random intervals when the unit was new. This affected the transmit frequency as well as that of the receive mode. It did not occur when using the outboard VFO. The malady ceased with time (about a week), and it appears that it was caused by a tight tuning mechanism (backlash) on the FT-102, which loosened up after being used for awhile.

Some of the earlier releases of the FT-102 had severe key clicks on cw. Apparently, this has been resolved. Yaesu states that owners with this problem should contact the company for instructions concerning a cure. Not all FT-102s had the problem. It was a sporadic type of anomaly that was cured easily. The review unit has a good cw wave form.

One might be baffled by the excellent receiver dynamic range after examining the receiver front-end circuit. There is nothing unusual about the mixer, for it contains only a pair of source-driven FETs in a singly balanced mixer arrangement. The rf amplifier is similarly mundane upon cursory examination: It employs two more JFETs, this time in series. The reason for the strong front-end performance comes clear when you trace the dc supply line to the pc-board terminal strip: Yaesu uses + 24 V for the rf amplifier and mixer rather than the usual 12-V dc! The higher operating voltage greatly enhances dynamic range.

FV-102DM Outboard LO

This unit is a remarkable accessory. It contains



a synthesizer and CPU designed especially for use with the FT-102. This product allows tuning by the dial, scanner, keyboard or memory 100 kHz beyond the band edges during transmit, receive, transceiver or RIT operation.

A dual-function, 17-button keyboard on the panel permits push-button frequency entry along with 5- or 20-kHz stepping, four-speed scanning, frequency lock, last-digit blanking and RIT operation. There are six extra keys to allow convenient receive and transmit frequency-source selection of the FT-102 internal VFO, the FV-102DM dial or one of the 12 memory channels of the FV-102DM. LEDs indicate the operating status.

The FV-102DM includes an internal-battery holder to provide dc backup for retaining the data stored in the memories when the transceiver is turned off, or if a power failure occurs. The five-digit frequency display indicates kilohertz with resolution to 10 Hz, or 100 Hz if the last digit is blanked. Scanning can be controlled from the front panel of the FV-102DM or by means of the FT-102 scanning microphone, if the latter is used. No outward indications of reciprocal mixing caused by synthesizer noise in the FV-102DM were observed. Similarly, nearby stations were unable to detect wide-band noise on the transmitter signal.

This transceiver is fully modern, and the performance is outstanding. Price class: $1150 (stock FT-102). Manufacturer: Yaesu Electronics Corp., 6851 Walthall Way, Paramount, CA 90723. — *Doug DeMaw, W1FB/8*

VIEWSTAR VS 1500A TRANSMATCH

☐ The VS 1500A Transmatch looks functional from the front, even at a quick glance. However, it isn't until one gets the cover off that the cleanliness of design and layout strikes the eye.

Truly a first-class job of design and construction, the matching network employs some of the highest-quality components available. This is also the first commercially made Transmatch I have seen that makes use of the SPC configuration. The SPC arrangement, developed by Doug DeMaw, W1FB, is the result of an effort to design a circuit that would maintain a band-pass response under all load conditions. Some readers may be interested in reviewing W1FB's explanation of the SPC circuit, which appeared in Technical Correspondence, *QST*, July 1980.

Features and Flexibility

The VS 1500A will handle the "legal limit" and a bit more. It is generously rated to handle 1500 W continuously. A 1:4 balun is incorporated to handle balanced feed lines. The range of antennas that may be connected to the VS 1500A includes dipoles, inverted Vs, Yagis, whips and random-length wires.

The built-in wattmeter will read 300 W or 3 kW in the FORWARD position, and 300 W in the REVERSE position; it is always in the circuit. A front panel switch selects one of two coaxial-cable-fed antennas (direct or through the tuner), a balanced line or a random-wire antenna. The LOAD and BYPASS positions provide for connection to an external dummy load, and to a coaxial-cable-fed antenna that is connected to, but not *through*, the Transmatch. In the LOAD, BYPASS, COAX 1/OUT and COAX 2/OUT positions, the tuner is bypassed.

Instructions

The manual that accompanies the VS 1500A is detailed and complete, and includes a large schematic diagram of the unit. There are a few minor errors in the manual, and some terminology doesn't quite agree with the labels on the equipment. These have been called to the attention of the manufacturer, who has assured me that they will be corrected. However, none of them is sufficient to cause any problem with understanding the operation of the VS 1500A, or in preventing one from placing the unit in use.

Lab Tests

As measured in the ARRL Lab, the insertion loss of the Transmatch was only 0.5 dB. It handled the rated input power into a 50-ohm dummy load with no difficulty.

The VS 1500A Transmatch has been in use in my station for several months, and has done a thorough job of helping to match to a variety of antennas from 160 to 10 meters. Operation is simple, straightforward and effective in that it combines all antenna switching and feed-line matching into one compact unit. As with any device of this type, it is desirable to obtain a matched condition at low power, prior to the application of high power.

The VS 1500A measures 5-3/4 × 11-1/4 × 13-1/2 inches (HWD) and weighs 6-1/4 lb.[1] It is distributed by Unadilla/Reyco Division, Microwave Filter Co., 6743 Kinne St., East Syracuse, NY 13057. Price class: $490. — *Lee Aurick, W1SE*

MORSE CODE TRAINER II

☐ Written by Joe Morris, N4EU, this product is a versatile Morse training program for users of the Radio Shack TRS-80® Models I or III microcomputer. This software package, available for 16K cassette or 32K disk systems, tutors and drills you in the 26 letters, 10 numerals and five common punctuation marks (. , ? / —), and helps you increase your code speed, at your own pace, to 31 words per minute (wpm). The 16K cassette version was reviewed on a TRS-80 Model III.

The manufacturer has intentionally omitted a detailed instruction manual to keep the cost down, knowing that even novice TRS-80 users will use the CLOAD command in the absence of other directions. The instructions can be called easily, and the program will automatically lead you in the right direction with clear-cut menus, self prompting and effective error trapping.

Not until you've unwrapped the package, loaded the program and begun running the Morse Code Trainer II will you realize that additional equipment is needed. Early in your first encounter with the program, you'll read on the video screen: "Requirements: In addition to the computer and software you will need a code oscillator or an audio amplifier similar to those available at your local Radio Shack." You'll also need wire, clip leads and possibly a speaker or batteries, depending on whether the oscillator or amplifier comes equipped with them.

The code-practice oscillator and speaker are attached to the computer by running a clip lead between the oscillator and the smallest cassette cable plug. This cable is normally used to turn the cassette player on and off remotely; here it is used by the computer to key the oscillator. Alternatively, you can take advantage of the keying tone generated by the computer at the large auxiliary (AUX) plug by using an audio

[1] mm = in. × 25.4; kg = lb × 0.454.

amplifier. Suitable oscillators and amplifiers are available from Radio Shack under part numbers listed in the program.

Program operation is straightforward, self-explanatory and effective. On starting up, you're presented with a choice of creating or loading from a cassette (previously created and saved) word/phrase data file that will be used under later menu options. The beginner can bypass this simply by pressing C, * and ENTER. Once this has been completed, you are given the main menu options: (W) random words, (M) mixed groups, (L) letter groups, (N) number groups, (B) Beginner's Club and (E) end cw practice. You simply respond with the letter of your choice.

If you're a newcomer, you'll select (B) to enter the Beginner's Club, and will be asked to respond to several questions. First, you'll be asked to list those characters (if any) already known and then to list a few characters you'd like to learn during the session. The program initializes at 5 wpm (characters are formed at 13 wpm, spaced to yield 5 wpm overall) and begins the instruction sequence.

The first of the new characters to be learned is sent 10 times at 5 wpm in perfectly formed code, and the character is displayed on the screen. The beginner is then asked if "You got it . . . (Y/N)?" Responding with a "no" causes the sequence to be repeated; responding with a "yes" causes the computer to offer congratulations and gives the option of reviewing the characters one more time. Then, the sequence recycles for the next new character to be learned. When all new characters are learned to the *user's* satisfaction, you'll move from the instruction sequence to the drill sequence.

Under program control, you're drilled on both the characters just learned and the ones listed earlier as already known; a character is sent and you're asked to type the appropriate response. If the response is correct, another character is sent. If the response is incorrect, the phrase "No, I sent . . ." (followed by the character) is displayed as the character is sent three more times; then, the incorrect response is displayed and sent three times. If you don't respond at all, the character being sent is eventually displayed on the screen.

When you have had enough drill, or if you already know the code at 5 wpm, you can get to the other options through the main menu. Typing W, M, L or N will bring up the speed-setting routine for subsequent practice. Code speed is adjusted from 5 to 31 wpm by holding down the F key to increase, or the S key to decrease the speed; the speed, which changes in 2-wpm increments, is displayed on the screen. At code speeds of less than 13 wpm, each character is sent at 13 wpm with between-character spacing adjusted for the proper overall speed; at 13 wpm and above, the code is formed and spaced in "real time." You also have the option of changing the within-character dot/dash/space ratios, that is, the character weighting. Ratios are initialized at the proper relative values.

The remaining options provide a variety of specialized practice routines to get you up to your desired code speed quickly. Selecting L results in five-character groups of random numerals; M mixes letters and numerals in five-character random groups; W yields randomly selected words and phrases from the user-generated text files; and E ends the code practice session. Under all of these options (except E), the characters being sent are displayed on the screen so you can check your accuracy.

How effective is the Morse Code Trainer II? Its approach, from aural-only drill, immediate feedback and user-specified drill sequences to the slow-code-speed 13-wpm character generation, is an adaptation of many successful techniques. These, with the operating flexibility, make the Morse Code Trainer II attractive. Micro 80, Inc., claims that a newcomer who spends 1/2 hour daily using the program will reach the 13-wpm level in 120 days, and within 60 more days, 20 wpm. For most people, this estimate is likely conservative, especially if they also *use* their newly learned skill *on the air*. Though many other techniques that don't require a computer are also effective in teaching the code and increasing one's speed, if you have a TRS-80 Model I or III microcomputer, Morse Code Trainer II is certainly an effective and enjoyable way to go. Morse Code Trainer II is available from Micro 80, Inc., 2665 North Busby Rd., Oak Harbor, WA 98277. Price class: $17 (plus $2 shipping). — *Steve Place, WB1EYI*

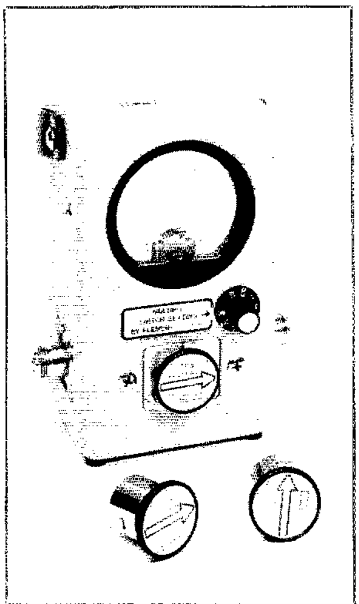
BIRD MODEL 4410 THRULINE® WATTMETER

☐ Bird wattmeters have long enjoyed a reputation for accuracy and reliability. The model 4410, designed for use in 50-ohm transmission lines, upholds the tradition.

The 4410 is housed in a sturdy grey 6-7/8 × 4 × 3-3/8 inch (HWD) aluminum enclosure, complete with leather carrying strap and rubber feet on both the bottom and rear panels.[1] It weighs three pounds. In size and appearance, the 4410 is very similar to the venerable Bird 43. The unit is normally supplied with female type N connectors, but a wide variety of male and female Bird quick-change fittings is available.

Depending on the plug-in element selected, this instrument will measure 0 to 10 kW from 200 kHz to 30 MHz, and 0 to 1 kW from 30 to 1000 MHz. Like the Bird 43, the 4410 samples rf flowing in the precisely machined Thruline, a short section of air-type line with a characteristic impedance of 50 ohms. The coupling circuit, which samples the traveling waves, is in the plug-in element. The element may be rotated to measure either forward or reflected power.

[1]m = ft × 0.3048.



Unlike the Bird 43, the 4410 incorporates circuitry to allow each plug-in element to measure a wide range of full-scale power values. The 4410 meter face has two scales: zero to one, and zero to three. A switch above the slug socket allows the user to select the full-scale power range needed for the job at hand. The switch has the following settings: 100, 30, 10, 3, 1, 0.3 and 0.1. Each element has a "factor" (100 for the 10-kW slugs, and 10 for the 1-kW slugs). The power-range-switch setting multiplied by the element-factor number gives the full-scale-power value. For example, with the 1-kW, 144-520 MHz slug installed (factor = 10) and the switch set on 1, the full-scale power reading is 10 W. Move the switch to 30, and the full-scale value becomes 300 W. Depending on the switch setting, each 1-kW slug will measure 0 to 1000, 300, 100, 30, 10, 3 or 1 W. For the 10-kW slugs, the full-scale values are 10,000, 3000, 1000, 300, 100, 30 and 10 W.

A 9-V alkaline battery powers the 4410. With the range selector in any position but OFF, there is a slight battery drain. Battery life is rated at 24 hours of continuous use. Among the settings on the range switch is a battery test position, which should be used before making any measurements. A weak battery will affect accuracy.

To find out if the same slug could really measure 0 to 1 and 0 to 1000 W with accuracy, we compared readings made with the 4410 to some made with the ARRL laboratory unit, a recently factory-calibrated Bird 43. We tried various powers at different frequencies, from 1 kW on 3.5 MHz, to 120 W on 14 MHz, to 10 W on 144 MHz, to 500 W on 432 MHz. Although the 4410 and the 43 did not always exactly agree, the readings were always within Bird's claimed specifications of ± 5% of reading above 20% of full scale. Readings on the 4410 were consistent from setting to setting (i.e. 100 W on the 100 W setting was also 100 W on the 300- and 1000-W settings).

The only complaint I have with the 4410 is that the meter movement is slower in responding to power changes than most other wattmeters I've used. It's so slow that I sometimes found it difficult to tune up a transmitter with "touchy" controls when using the 4410 to monitor output power. I would tune through the output power peak before the meter responded. Other than that, the 4410 is a dandy little unit, able to measure QRP or QRO accurately at the flick of a switch.

The instruction manual is exceptionally comprehensive for a device as simple as a wattmeter. Included are such goodies as nomographs for determining VSWR based on forward vs. reflected power readings, and complete calibration information. Slugs of interest available to amateurs include 1 or 10 kW 2-30 MHz, and 1 kW 25-80, 50-200, 144-520 and 200-1000 MHz units.

Price class of the Bird 4410: $495; plug-in elements: $150 each for the 2-30 MHz units, $125 for the 144-520 MHz unit. The manufacturer is Bird Electronics Corp., 30303 Aurora Rd., Solon, OH 44139. — *Mark Wilson, AA2Z* QST

Technical Correspondence

Conducted By
Dennis J. Lusis,* W1LJ

The publishers of QST assume no responsibility for statements made herein by correspondents.

NOTES ON NOISE FIGURE

☐ Recently, I was reading about diode mixers and noise figure in Microwave Associates' *Receiving Diode Handbook.*[1] Thinking over what I had read, I realized there are a number of ways to confuse people when discussing noise figure and noise factor. I am going to try to clarify these terms.

The *Receiving Diode Handbook* gives the following definition of overall receiver *noise figure* (p. 8): "The noise at the output of a receiver is the sum of the noise arising from the input termination (source) and noise contributed by the receiver itself (i.e., the i-f amplifier and mixer diode). The *noise factor* is the ratio of the actual output noise power of the device to the noise power which would be available if the device were perfect and merely amplified the thermal noise of the input termination without contributing any noise of its own."

The ARRL *Handbook* also gives the following equation for noise factor:

$$NF = \frac{S_i/N_i}{S_o/N_o} \qquad \text{(Eq. 1)}$$

Look at Fig. 1, which illustrates the noise factor relationship. We can see that

$$S_o = G \cdot S_i, \text{ and } N_o = G(N_{amp} + N_i) \qquad \text{(Eq. 2)}$$

The noise factor is derived as follows:

$$NF = \frac{S_i/N_i}{G \cdot S_i/G(N_{amp} + N_i)}$$

$$= \frac{S_i \cdot G(N_{amp} + N_i)}{G \cdot S_i \cdot N_i} = \frac{N_{amp} + N_i}{N_i}$$

$$= \frac{N_{amp}}{N_i} + 1 \qquad \text{(Eq. 3)}$$

This equation agrees with the definition. However, it is distinctly different from the popular noise-factor and noise-figure equations, which represent the equivalent noise power looking into the system (N_{amp}) compared to the noise power of a 293 K source.

The two concepts are separable. The first case (Eq. 1) doesn't seem particularly useful, because the noise factor changes with generator temperature. The second case (Eq. 2) represents the effective input temperature of the system referenced to a known temperature (290 K, 293 K and 298 K are used). This "standardization" allows meaningful comparisons among different equipment.

I've shifted from the term "noise power" to "temperature" because it is more useful and easier to use. Noise power is the product of bandwidth, Boltzmann's constant (1.38×10^{-23} J/K) and the temperature in Kelvins. The bandwidth and the constant cancel themselves in the ratios used to derive noise factor, while only the temperature ratio remains.

[1]Microwave Associates, South Avenue, Burlington, MA 01803.

*Assistant Technical Editor

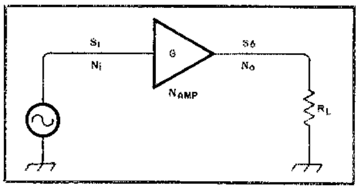


Fig. 1 — Model used to illustrate noise figure and noise factor.

The third source of confusion about noise factor is that it may refer to either

$$\text{(A)}\ \frac{T_{amp}}{293} + 1 \quad \text{or} \quad \text{(B)}\ \frac{T_{amp}}{293}$$

The A form is the more common of the two; however, the B form is needed for calculations. In the *Receiving Diode Handbook* both forms are mixed in one equation (Eq. 13, p. 187): $NF = L(t_m + F_{IF} - 1)$.

In that equation, F_{IF} is apparently the noise-factor form A, since 1 is subtracted from it in the equation. Inconsistently, NF is the noise-factor form B, since 1 is not added back to it.

Another source of confusion comes from mixing the logarithmic form with an arithmetic form in one equation. Noise figure is the logarithmic "dB" form, and noise factor is the arithmetic form. It is the noise-factor (B) form that is divided by the gain of preceding stages — *not* the noise figure when calculating the combined noise factor of several amplifier stages.

We could benefit by standardizing the terms in several ways. Noise figure could always be in the decibel form, and noise factor the arithmetic form, $\frac{T_{in}}{298} + 1$.

Temperature ratio (Tr) would generally be $\frac{T_{in}}{298}$, or $\frac{T_{out}}{298}$ for diode specs.

I've chosen 298 K because it equals 25°C, which is the commonly used reference for semiconductor ratings. I think *one* arbitrary reference temperature is enough! Actually, we could do away with NF entirely. I don't know of any use for the logarithmic form that can't be better served by the temperature ratio, or simply the temperature. — *Martin Sample, WA6JTD, Tuolumne, California*

MY DIPOLE DOESN'T WORK RIGHT

☐ Have you ever cut a dipole to the dimensions in the ARRL *Handbook* and ended up with a horrible mismatch to the feed line? Take heart. It happens for a reason that is fairly simple but rarely mentioned in the literature. I've had it happen a number of times to me.

An item that is discussed very briefly at best is *ground effect* on antenna impedance. When an antenna is strung over the ground, regardless of height, ground reflections cause changes in the feedpoint impedance. Table 1 gives the minimum height for a dipole where ground effects cause less than a 20% change in impedance. It's pretty terrifying for the 160-meter buff! Regardless, let's look at the overall picture and see what we can learn.

Table 1

Minimum Antenna Height for Less Than 20% Impedance Shift from Ground Effects

Band	*Height*
160 m	260 ft†
80 m	130 ft
40 m	66 ft
30 m	47 ft
20 m	34 ft

†m = ft × 0.3048.

The model I used for comparison is an antenna mounted over a *perfectly conducting* ground plane. When a dipole is strung horizontally above this perfect ground, the entire system appears as if there were no ground, but instead an active antenna located below the real one, spaced twice the distance to ground level (Fig. 2).

The subject of mutual impedance for parallel antennas is covered in the literature, but usually in an unintelligible manner. Kraus contains a fairly good section and Terman has several useful charts on different configurations.[2,3] Fig. 3 is a graph of the mutual impedance between two antennas at spacings of up to 1.8 λ (0.9 λ above ground for the "real" antenna). Note that the mutual reactance and resistance varies in a damped sine wave fashion. This function is described by Kraus. By studying Fig. 3 one can observe several interesting facts. If the antenna is mounted close to the ground, there is severe mutual coupling between it and the ground image "antenna." The resultant resistance and reactance of the antenna pair is given by

$$R = R_{11} - R_{21} \quad X = X_{11} - X_{21} \qquad \text{(Eq. 4)}$$

where

R_{11} and X_{11} represent the antenna self impedance

R_{21} and X_{21} represent the mutual impedance between the antenna and its image.

The importance of this relationship is that the mutual resistance and reactance subtract from the free-space antenna characteristics. Thus, at resonance (where the antenna is about 68 ohms in free space) and at 0.1 λ high, the center im-

[2]J. Kraus, *Antennas* (New York: McGraw-Hill Book Co., 1950), pp. 262-268.

[3]F. Terman, *Radio Engineer's Handbook* (New York: McGraw-Hill Book Co., 1943), pp. 774-780.

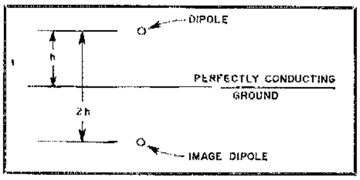


Fig. 2 — Illustration of the relationship between a dipole and the image dipole "formed" within the ground.

Table 2

40-Meter Dipole Characteristics at Different Heights

Height	Free Space			0.2 λ Above Ground			0.48 λ Above Ground		
(MHz)	R	X	SWR	R	X	SWR	R	X	SWR
6.9	63.6	−16.9	1.34	63.6	−16.9	1.34	63.6	−71.2	2.72
7.0	65.8	−29.2	1.54	65.8	6.7	1.18	65.8	−47.6	1.97
7.1	68.0	−5.7	1.14	68.0	30.2	1.54	68.0	−24.1	1.42
7.2	70.2	17.8	1.29	70.2	53.7	2.07	70.2	0.6	1.07
7.3	72.4	41.2	1.74	72.4	77.1	2.73	72.4	22.8	1.36
7.4	74.5	64.6	2.32	74.5	100.5	3.52	74.5	46.2	1.84

[1]*Note: Antenna is 67.45 feet long and fed with 75-Ω feed line.*

pedance is actually R = 68 − 50, or 18 ohms with X = 0 − (−24) = +24 ohms reactance. Where we had an SWR of 1.1:1 with 75-ohm feed line in free space, we now have an SWR of 4.62 with 75-ohm feed line, which will drop to 3.49:1 when using 50-ohm feedline. This is for a 160-meter dipole at 50 feet, or an 80-meter dipole at 25 feet.[4] A Transmatch will make these impedances acceptable to the transmitter, but I do not like them!

To illustrate this more graphically, some calculations were made for a 40-meter antenna. The results are shown in Table 2 and plotted in Fig. 4. You can see the shift in apparent resonant frequency as the height above ground is changed. For this example, I have chosen heights at which the mutual resistance is zero. Choosing other heights gives much poorer SWR curves, and muddies the water in seeing what happens as a function of height. You can estimate the effects of height on your antenna by assuming a free-space center impedance of 68 ohms, and subtracting the mutual R and X as read on the curve in Fig. 3. To illustrate this, assume we have an 80-meter dipole cut to 3750 kHz, which is hung at 62 feet above the ground. In free space the impedance would be about 68 ohms. Sixty-two feet is 0.236 λ at this frequency (0.472 λ between antennas), so the mutual impedance from the curves is −9.8 − j30.2 ohms. Subtracting the mutual terms from the self impedance, the antenna now appears as 77.8 + j30.2 ohms. Fed with 75-ohm line, the SWR would be 1.48:1, while the lowest SWR point has dropped down about 100 kHz to 3650 kHz. This example illustrates why antennas in general must be shortened when they are installed nearer the ground. Granted, the shift is only 100 kHz, but haven't you ever wondered why this is so?

What about the "real" world where there are

[4]m = ft × 0.3048.

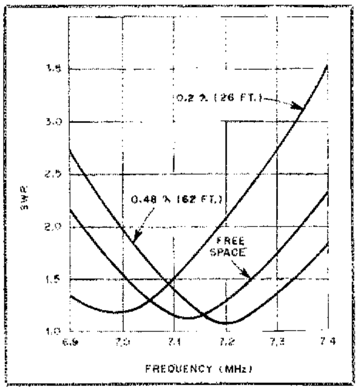


Fig. 4 — Graph showing the effects of antenna height on resonant frequency.

no perfectly conducting ground planes? I don't live on a field that used to be a broadcast station antenna farm! Less than perfect ground causes two effects. The first effect is that the image dipole spacing is greater than twice the physical height above ground. I've found the impedance of an 80-meter dipole mounted over 60 feet of coral rock (in the Caicos Is.) to be that of a dipole 80 feet above perfect ground. The coral rock is a very poor ground plane! The second effect is that there is less mutual coupling because of lower illumination of the image antenna through ground losses. This reduces the change in antenna impedance. (It is difficult to find information on these effects!)

What does all this add up to and what can be done about the situation? From Fig. 3, calculate the mutual coupling with the antenna ground image. Subtract these values from the free-space antenna impedance. This impedance may be calculated with the formulas developed by Hall or myself.[5,6] The SWR for a given feed line is determined by the formula in the ARRL *Handbook*.[7] If the SWR is acceptable, then all is fine. However, even if it is as bad as the example shown earlier, all is not lost. A matching system at the feed point can be used to make this low dipole appear as though it were in free space. This all sums up to the fact that one should string his or her antenna as high as possible, unless you desire some particular angle of radiation for point-to-point communications.[8] — *Alan B. Harbach, WA4DRU, Melbourne, Florida*

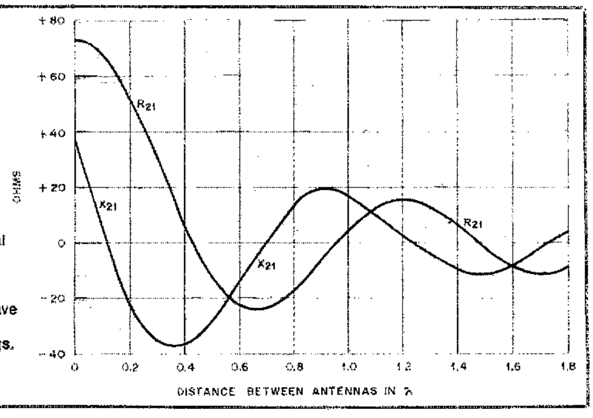


Fig. 3 — Mutual resistance (R_{21}) and reactance (X_{21}) between parallel, half-wave dipoles at various spacings.

MINIMUF FOR POLAR PATHS

☐ Users of MINIMUF (Dec. 1982 *QST*) should be suspicious of muf predictions on any path through a polar region that is in the summer season. MINIMUF uses the control-point method of muf prediction. For paths longer than 4000 km, the control points are located along the path at points 2000 km from each of the two terminals.[9] The path muf is the lesser of the two found at the control points.

MINIMUF computes the muf at each control point as a function of several parameters, including the number of hours of sunlight at the control point on the specified date and the elapsed time since sunrise or sunset at each UTC hour. If a control point is located in a polar region and a winter season polar night condition exists, there is no sunrise or sunset; MINIMUF detects this condition and uses an appropriate algorithm to compute the control point muf. It does *not* contain a separate algorithm for the summer season, midnight-sun condition in which there are 24 hours of sunlight at a control point. MINIMUF will instead compute an erroneous muf. I have observed this failing on two paths from Los Angeles to Europe in June and July.

There is no simple way to correct this problem. However, MINIMUF can be modified so that it will generate an error message instead of incorrect mufs if the midnight sun condition exists at a control point. It is only necessary to change line 1530 to

```
K9 = 12-ATN(K9/SQR(1-K9*K9))*7.639437
```

The argument of the SQR function is negative under the midnight-sun condition. The wording of the resulting error message will depend upon the computer system. — *Sheldon C. Shallon, W6EL, Los Angeles, California*

[5]J. Hall, "The Search for a Simple, Broadband 80-Meter Dipole, *QST*, April, 1983
[6]A. Harbach, "Broadband 80-Meter Antenna," *QST*, Dec. 1980.
[7]G. Woodward, ed., *The Radio Amateur's Handbook*, 60th ed. (Newington: ARRL, 1983), p. 19-2.
[8]G. Hall, ed., *The ARRL Antenna Book*, 14th ed. (Newington: ARRL, 1982), Chapter 2.
[9]mi = km x 0.6215.

Feedback

☐ In "A Top-Fed Vertical Antenna for 1.8 MHz — Plus 3" (Sept. 1983 *QST*), there are two errors pointed out by author Eichenauer. On page 27, the first paragraph should read: ". . . resistor of value R4 + RG."

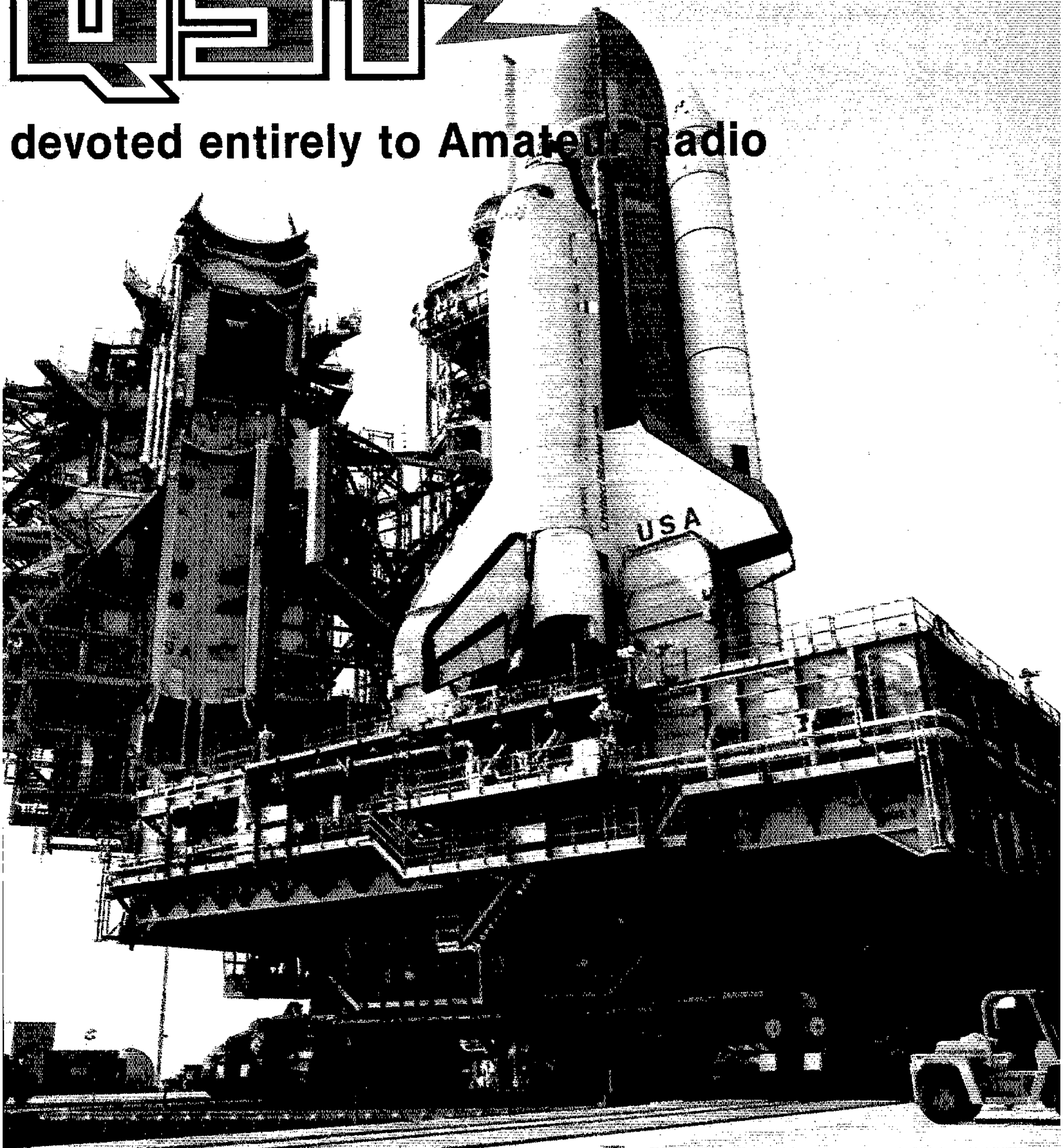
Also on the same page, computer program line 490 should read:

```
490 LET R4 = .01215*(ABS((SIN ATN (B))*180/PI))↑2
```

November 1983 $2.50

QST

devoted entirely to Amateur Radio



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WORLD COMMUNICATIONS YEAR
AÑO MUNDIAL DE LAS COMUNICACIONES
1983

The Space Shuttle *Columbia* rollout
Are you ready for STS-9?

QST

November 1983 *Volume LXVII Number 11*

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OUR COVER
The Space Shuttle *Columbia* takes a final earthbound ride to the launch pad before it carries our first Ham in Space — W5LFL. For the latest STS-9 flight information, call 203-666-0688. *(photo courtesy NASA)*

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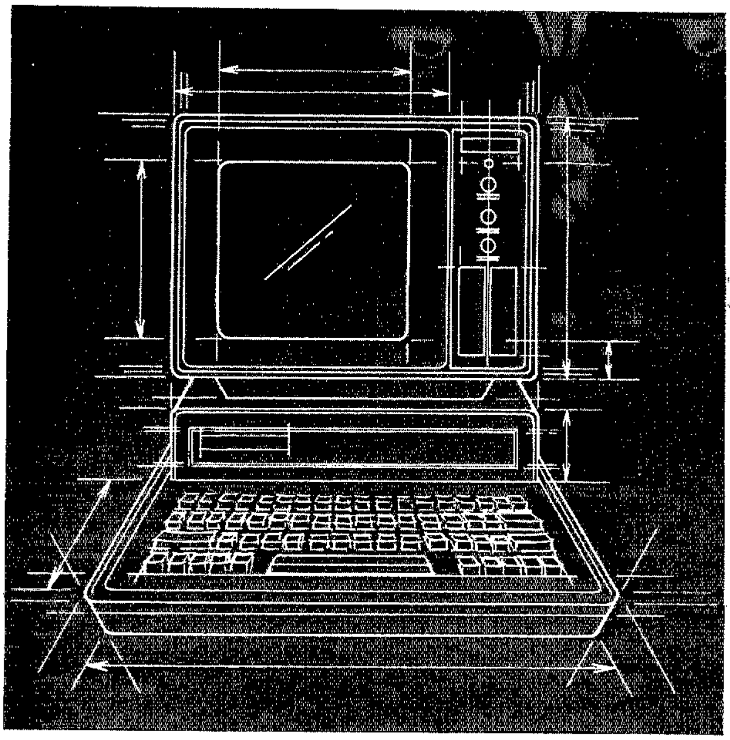


The Personal Computer†

Part 1: Has the desire to acquire a personal computer struck you? Are you baffled by "computerese"? Perhaps the following information will help you avoid making a costly mistake.

The personal computer industry is enjoying a booming business. No matter where you go, talk of computers is bound to come up. There are many who would like to own a personal computer, but feel out of place in a computer store and are baffled by computer lingo. The following information should help you understand a little more about computers.

Buzzwords

The personal computer industry is a spin-off of scientific and industrial efforts spanning four decades. One result of this historical background is a plethora of buzzwords used to describe or explain a computer or peripheral device operation. To converse in the computer world, a vocabulary of "computerese" must be developed. The accompanying computerese mini-dictionary is by no means complete, but should provide you with the basics of computer jargon.

†Adapted from *Personal Computer,* a Combustion Engineering corporate technology newsletter published by the Electronics Technology Applications Center, Combustion Engineering, Inc., 44 South 122 East Ave., Tulsa, OK 74128. The individual articles were written by several C-E employees as an internal newsletter for the benefit of the corporation.

Bit — a unit used to describe the smallest active element in a computer operation. A bit can have one of only two values, 1 or 0. These two binary values can be used to represent logical conditions (1 = true, 0 = false), physical states (1 = on, 0 = off) or actual quantities of 1 and 0.

Byte — a group of eight bits used as a single unit to express a value within a computer operation. Since a single bit can represent two values, 0 and 1, if bits are in groups of eight, they could be used to express numbers as powers of 2. If we line these eight bits up in descending powers of 2 as in Fig. 1A, we can see that if we give each bit a value of 0 that the resultant byte would look like that of Fig. 1B and have a value of 0. This is the minimum value of the byte. If we then replace all the 0 bits with values of 1, then the byte would look like that of Fig. 1C. It would have a value of $2^7 + 2^6 + 2^5 + 2^4 + 2^3 + 2^2 + 2^1 + 2^0 = 255$, which is the largest value of the byte. By altering the proper bits in a byte, values from 0 to 255 can be expressed.

The most common use for the term byte is to express the memory size of a computer. Most personal computers operate on eight-bit values, and thus store quantities in their memory as bytes. Generally, personal computer memories range in size from less than 1024 bytes to as much as 64,512 bytes or more. The term "kilo" is often used to shorten the expression of size to 64 kilobytes or 64 K.[1]

Baud — a unit generally used to express the rate at which data bits are moved in a serial manner. For personal computer use, 1 baud = 1 bit/second. Common signaling rates encountered in personal computing range from 110 to over 19,200 bauds.

BASIC — an acronym for Beginner's All-purpose Symbolic Instruction Code. Perhaps the most common personal-computing programming language, the original BASIC language was developed at Dartmouth to aid in teaching students computer programming. The use of BASIC helps remove the programmer from worrying about how the computer handles the manipulation of bytes and bits at the machine level, and allows the programmer to concentrate on the task to be performed. Virtually all personal computers offer some form of BASIC as their primary programming language. Most of these BASIC implementations are extended or modified versions of the Dartmouth standard version. Programs written in one BASIC

[1]Notes appear on page 14.

dialect can generally be made to run on a different machine with another version of BASIC if special extensions and machine dependent statements are avoided.

BASIC can be self-taught, and many good BASIC programming books are available. As with any programming language, you have to know what you want to do and *how to do it* before BASIC (or any other programming language) can be made to solve problems.

CP/M® — an acronym for Control Program/Microcomputer. This control program was developed by Digital Research Corporation of California to provide a disk operating system for microcomputers. CP/M has become the de facto standard for eight-bit microcomputers based on the Intel 8080 microprocessor or the Zilog Z80 microprocessor. CP/M provides full file management and interface abilities for custom machine environments. Some versions are now available to run on non-8080 or -Z80 systems. There are several hundred programs commercially available that will run under CP/M, including over 15 different BASIC dialects.

Cassette — a magnetic-tape storage device used by some personal computers to save programs for future use. Most cassette-based systems store data on the cassette at speeds of from 300 to 1200 bauds. This means that a typical program of 4 K in length (4000 × 8 bits = 32,000 bits) would take 106 seconds to load into the computer at a rate of 300 bauds and 27 seconds at 1200 bauds. This is not too fast, but much better than reentering an entire program by hand each time you want to run it!

Many personal computers that use cassette storage can use a standard audio cassette recorder in the $40 to $100 price range and moderate-quality audio cassette tapes for data storage. Cassette systems do not generally have supervisory programs that provide a directory of the programs contained on the tape, so you must keep track of what is on each cassette.

Disk — a magnetic media device that allows storage of programs for later use. The medium for use in a disk drive is called a *diskette* and is often referred to as a *floppy disk* since it is quite flexible and looks like a very thin phonograph record.

There are three disk sizes in use on personal computers: 5¼ inch, 8 inch and a 3½-inch version that is gaining in popularity.[2] A typical 5¼-inch diskette can store about 81 K of data or program material, and some special formats can store up to 500 K. The rate at which data can be stored is much higher with a disk unit than with a cassette tape. A typical transfer rate for a 5¼-inch disk is 125,000 bits/second or 125k bauds. To load a 4-K program would take about ¼ of a second for the actual transfer. (This doesn't include the various mechanical delays and processing overhead time to put the data away in memory.)

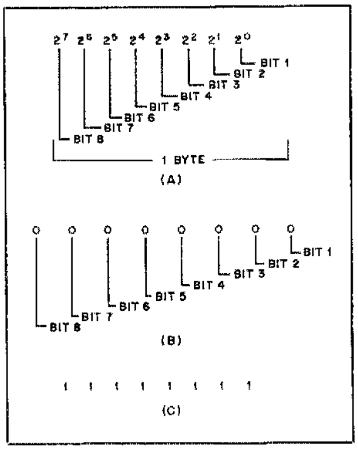


Fig. 1 — At A, bit positions and their respective powers of the base 2. A byte value of 0 is represented in B; the maximum byte value of 255 is depicted at C.

An 8-inch diskette can store about 140 K of program material and some special units store over 1 million bytes! The transfer rate for data is also faster, about 250k bits/second.

Density — a term generally applied to the method of recording data on a diskette. Single density (SD) is the most common format available to the personal-computer user. For a 5¼-inch disk, this works out to about 80 K per diskette; for an 8-inch disk, about 240 K. Double density (DD) is roughly twice the storage capacity of single density providing about 160 K on a 5¼-inch diskette or 500 K on an 8-inch diskette. Quad density is actually a misnomer. It is usually a special disk drive that can use both sides of a double-density diskette, thus providing 330-360 K for a 5¼-inch disk and over 1 megabyte for an 8-inch disk.

Format — a term related to the method of storing data on a diskette. The generally accepted "standard" (watch out for the use of this word) was adopted from IBM®. For single-density systems, it is generally patterned as follows:

A) The diskette is divided into concentric circles called *tracks*, usually likened to grooves on a phonograph record. Each track has a unique position relative to the center of the diskette. Each track is given a number or *address* to represent this position. For 5¼-inch diskettes of single density, this is usually from 0 to 34 or 0 to 39, with 0 being the outermost track from the center. For 8-inch diskettes of single density, it is 0 to 76.

B) The diskette is further divided into smaller subdivisions called *sectors*. Sectors also have addresses that are relative to their track address. A typical track has 16 sectors on a 5¼-inch SD diskette and 26 sectors on an 8-inch diskette. The sector size can range from 128 to 256 bytes.

Hard Disk — a magnetic recording device that uses relatively nonflexible disks, as compared to a floppy disk. The hard disk differs from the floppy disk in other, more important ways:

A) The pickup heads actually float a few micro-inches above the surface of the disk. The heads do not touch the surface of the disk as they do with a floppy disk. This means little disk or head wear occurs.

B) It has more tracks and sectors than a floppy disk and has a faster access to a given track or sector. Even a 5¼-inch hard disk typically has a capacity of 6.5 megabytes.

C) It is much more sensitive to damage from vibration or temperature extremes because the heads float above the surface. Hence, most hard disks are sealed against the environment. There are replaceable media "pancake stacks" containing multiple disk surfaces, but this arrangement is usually reserved for large disk systems (greater than 20 megabytes).

D) The cost of a hard-disk drive is from three to 10 times that of a floppy-disk drive, but the difference is steadily declining.

Modem — a device that allows a computer to communicate with a terminal or another computer, e.g., over a telephone line. "Modem" is a contraction of "modulator/demodulator." Standard modems use 300- to 1200-baud transmission rates. They use either a shift in frequency or phase to represent the two possible bit states of 1 and 0. For instance, to send a bit value of 1, 2225 Hz might be used, and 2025 Hz used for a bit value of 0.

In the Bell 103-series modems, there are two basic modem types, depending on who places the call: (1) *Originate* modems send 1270-Hz and 1070-Hz tones to a remote (answer) modem; (2) *Answer* modems send 2225-Hz and 2025-Hz tones to the originate modem. There are also combined Originate/Answer modems. This scheme allows for full duplex operation; that is, both modems can "talk" at the same time. A modem is simply a conversion device to make the computer binary output bits suitable for transmission over standard telephone lines.

Operating System — a supervisory program that interprets operator input and operates on it. An operating system can be as simple as a *system monitor* that can only provide basic I/O functions, memory display or alteration, and a means to execute assembly-language programs. Or, it can be as complex as CP/M with full file-handling facilities and multiple-language capabilities. Some low-end (small, less expensive) personal computers and even some more expensive units do not have a readily visible operating system. Rather, it is made a part of a BASIC interpreter, and all commands are entered while in BASIC.

RAM — an acronym for random-access memory that refers to the changeable memory available within a given machine. RAM is used to store user programs in most systems. The stated memory size of a personal computer in advertisements usually refers to RAM. RAM is volatile, which means the data stored in RAM are lost if power is removed. RAM can be written to (store data) or read from (retrieve data) and is called "read/write" memory.

ROM — An acronym for read-only memory. It refers to memory that is not alterable by the user. ROM usually contains such things as the BASIC interpreter and is nonvolatile; it retains data when power is removed. Most types of ROM cannot be written to by the user. There are three basic types of ROMs that appear in personal computers today:

A) ROM. This IC has the required data programmed into it at the place of manufacture by a process called *mask programming*. Once programmed, the data cannot be changed.

B) EPROM. This type of ROM is erasable and programmable. The information is erased by exposure to ultraviolet light and the IC can be programmed in a special device. After programming, the EPROM is operationally identical to a standard ROM. EPROMs can be erased and reprogrammed many times.

C) EAROM. This is an electrically alterable, read-only memory. A special type of ROM, it has the nonvolatile capabilities of a true ROM, but it can be altered like RAM while in the user's system. It is usually slower to write to than normal RAM and there is a limit as to the number of times the data can be altered. The IC will retain the data after power is removed.

Serial — a term usually applied to a method of transferring data over a wire from a computer to a peripheral device. In serial operation, data is sent out 1 bit at a time to the peripheral device. Some printers, disk drives and CRT terminals use serial communication to transmit data to and receive data from the computer.

The most used serial hookup is the EIA RS-232-C. This designation refers to a standard interface specification for serial communication in which the signal levels to and from the computer are strictly defined. In addition, the connector type and its pin connections are standardized so that virtually any device specified for RS-232-C connection can be made to function by simply plugging it in.

Computer Shopping

What do you expect a personal computer to do for you? What are you willing to pay for it? These questions are not easy to answer. If you don't have the answers before you shop for a personal computer, you leave yourself open to all sorts of disappointment.

The general public tends to think of all computers in a sort of stereotyped manner; that is, if you tell a computer what you want, it will do it for you. Some sales people, for various reasons, try to reinforce this view by selling you a machine and telling you that you can do your taxes, keep personal records, catalog a record collection, etc. — all on the $200 special 4-K machine. Of course, what the salesperson failed to mention is that another 60 K of memory, two disk drives and several software packages must be added, at a net increase of $2000 or more!

Many people become soured to personal computing because the machine they purchased cannot do everything that it was expected to do. As with any job, if you need a tool to do a job, the better you define the job before you make or buy the tool, the better performance you can expect from the tool.

Let's look at a few possible answers that might be given to the two previously asked questions.

A) *I want it to play games, maybe even write a few games myself.*

This answer fits a fairly large number of first-time buyers and, fortunately, doesn't cost too much to satisfy. The machines available that can fill this requirement range in cost from less than $100 to about $400. Memory capacity is about 4 K or more, and a cassette interface is provided for use with a tape recorder as a program-storage device. The display unit is typically a modified TV or a standard TV with an external rf modulator inserted between the computer and TV set. The rf modulator converts the video signal from the computer into a TV signal that can be displayed. BASIC is usually provided in ROM for user programming.

The primary limitations of this class of machine are: (1) Memory size may be too small for significant program development or use with large games; (2) the cassette interface allows saving only program files written in BASIC without any directory of contents; (3) memory expansion may be limited, and the BASIC interpretor provided may be limited in scope; (4) the amount of information (lines and columns) that can be presented on the screen is limited by having to use a TV set as a video monitor. Formats of 22 lines and 22 columns, 16 lines and 32 columns, or 20 to 24 lines of 40 columns are provided. These formats cannot display a typical 8½ × 11-inch "page" of information.

B) *I want to do some financial/accounting work and keep my personal book records on the computer, but I can't afford to spend too much money.*

This response is probably applicable to a large segment of potential buyers. The purchaser, who sees the growing use of computers in home appliances, the automobile, on the ham bands and in the office, feels that he or she has to get involved in this computer business or fall behind. Caveat emptor (let the buyer beware) is applicable to this situation. The danger here is twofold. On one hand is the danger of over buying and spending far more than is necessary. On the other hand is the danger of underbuying and then becoming upset because the machine won't cooperate and produce the desired results.

Disappointed computer buyers are usually those who underbuy. That results from a lack of understanding of just what is to be done with the computer and/or by an enthusiastic salesperson who made excessive claims for the "Super-8" computer selling for only $99.95 plus tax.

The primary requirements of a system that will fulfill the needs mentioned earlier are

1) some means of data storage that allows data to be recalled by name. A simple cassette-storage capability will not suffice. There are some cassette systems that provide these features, but the vast majority of systems now use one or more disk drives, usually 5¼-inch types.

2) enough RAM to handle a reasonable amount of data when the control program is also in memory; 16 K is typically a minimum.

3) a BASIC interpreter that features floating-point arithmetic; that is, the ability to deal with fractional numbers, not just whole numbers (integers).

While these requirements are a minimum, there are a few desirable features that would minimize the required programming knowledge of the user and enhance the final results:

1) A program such as VisiCalc® would greatly simplify developing an accounting system. Most systems have some form of electronic spread sheet available. Cost: $100 to $300.

2) A printer enables you to maintain a printed record of financial information and spread sheets. Most systems will support a serial or parallel printer. Cost: $300 to $1500 or more. A minimum setup might consist of a game system (as in example A) with an added disk drive and memory expansion unit for costs ranging from $1000 to $1800.

C) *I want to learn programming.*

Buyers in this category are generally the prospective software hobbyist or those who have a desire to learn or improve their computer skills and do not have on-the-job access to the equipment to do so. A personal computer can be a valuable aid in learning programming techniques. However, it is only an aid; it cannot teach you to program. It can help you to learn to code. There is a difference:

Code — to write a sequence of steps, in a given computer language, that follows a predefined algorithm (plan) of program flow.

Program — to develop a workable algorithm to accomplish a given task.

There are many good books available for self-learning programming techniques, and some use the BASIC language for coding purposes. Remember: You must know what you want to do and how to do it before you can program a computer to do a given task! Just learning a language such as BASIC or FORTRAN will not enable you to write anything more than relatively simple programs. However, if the proper *way* to write a program is learned, then the language used to code the program becomes far less important.

A system for learning programming techniques must be configured according to the complexity of the desired skills. A simple 4-K memory machine and cassette interface are quite enough as long as you are not trying to develop file-handling skills. If you do desire to learn file structures, then a system like that of example B is required with at least one disk drive.

D) *I want a machine as a teaching aid for my children.*

Computer-aided instruction (CAI) is a rapidly growing field that is beginning to appear at all levels of education. Areas previously restricted to the college engineering or computer-science level are now appearing in elementary school science classes. Children are being exposed to the electronic media on an ever-increasing basis both in school and outside the classroom. The personal computer can be a great help in preparing children to meet the tremendous volume of electronic data that is sure to be a part of their lives in the next 10 to 20 years.

To accomplish this goal, a minimal system is required. A 4-K machine, preferably with color capabilities and a BASIC interpreter, will provide enough computing power to create math drills, word exercises, pattern recognition drills, etc. In addition, many prepackaged programs are available for teaching purposes on some of the more popular small machines. A good estimate would be about $250 to $600 for such a system.

E) *I'm an engineer and I want a machine that can handle some of the work I do so that I can work at home when necessary.*

This class of buyer is perhaps the most demanding of the personal computer users. To provide support for these requirements, a personal computer would need

1) sufficient memory to handle large user programs;

2) disk-storage capability for saving data and programs;

3) a versatile disk operating system (DOS) capable of supporting many different languages;

4) a system that has a large software inventory commercially available;

5) sufficient speed to rapidly perform complex calculations.

Fortunately, there is a good variety of systems available that will support these requirements, but they are not inexpensive. A typical system with 64 K of RAM, two 5¼-inch disk drives, CP/M operating system (or UNIX), a BASIC compiler and a FORTRAN compiler costs from $2200 to $6000 and up. The addition of 8-inch disk drives and hard disk capability can push the cost up to over $10,000.

Obviously these five situations do not describe all the possible reasons for buying a personal computer. Hopefully, they present enough angles of the problem to show that the purchase of a personal computer requires thought and planning if disappointment is to be avoided. So, to itemize some of the dos and don'ts:

Do

- know what you expect the computer to do for you.
- find out what those expectations should cost you.
- examine several systems before buying. Little things can make a difference in the long run.
- ask questions about a prospective system — lots of questions.
- think about what you may want to do with the system later on (and are willing to pay for now).
- check into the availability of repair service.
- read some of the personal-computer periodicals. *Byte, Popular Computing* and *Personal Computing*, for example, contain many equipment reviews and hints on buying a system, as well as advertising such systems.[3,4,5]

Don't

- believe everything you hear. Investigate sales claims.
- expect the computer to be able to do something you don't know how to do unless you are using a commercially available program.
- buy used or out-of-warranty equipment unless you are willing (and able) to fix it yourself or can afford to have it fixed by someone else.
- buy the first system you look at just because the salesperson was nice; check it out!
- buy the "Super Whiz 80" computer with all the bells and whistles unless you plan to use them in the future. They cost money.
- be afraid to ask for full explanations of things you don't understand about a given machine; it's your money.

Conclusion

Personal computers can be a useful tool or an expensive toy, depending on how well you prepare yourself before buying one. The average person can learn to use a personal computer and not spend a fortune doing it. The watchword is caution. There are a lot of computers on the market, and many of them are highly specialized. Select a machine that will fit your desires and your pocketbook.

Notes

[1]One kilobyte is equivalent to 1024 bytes.

[2]mm = in. × 25.4.

[3]*Byte* Subscriptions, P.O. Box 590, Martinsville, NJ 08836.

[4]*Popular Computing*, P.O. Box 307, Martinsville, NJ 08836.

[5]*Personal Computing*, P.O. Box 2941, Boulder, CO 80321.

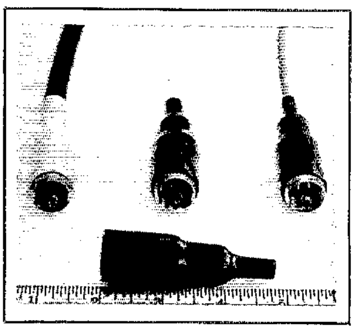
New Products

KILO-TEC WEATHER BOOT

☐ Kilo-Tec is supplying a custom weather boot for use with RG-8X, RG-59/U, RG-58/U and RG-8/U coaxial cables and PL-259/SO-239 combinations. During cable installation, simply slip the weather boot over the coaxial cable before soldering on the connector. Once the connector is attached, slide the boot over the PL-259 for a good weather seal.

The boots are manufactured from a flexible vinyl material that resists moisture and breakdown from the sun's rays. They are designed to keep the connections clean and dry, and should help to keep moisture out of the coaxial cable.

Kilo-Tec offers three models from which to choose: KTB-58 for RG-58/U, KTB-8 for RG-8/U, and KTB-8X for RG-8X or RG-59/U cables. (Custom-made boots can be made for other types of cables and connectors; contact Kilo-Tec for a quote for special requirements.) If the boots are not available from your local Amateur Radio equipment supplier, you can order directly from Kilo-Tec, P.O. Box 1001, Oak View, CA 93022; tel. 805-646-9645. A kit of two of each type costs $5.95 plus shipping and handling charges. — *Paul K. Pagel, N1FB*



Looking *Down* on the Aurora

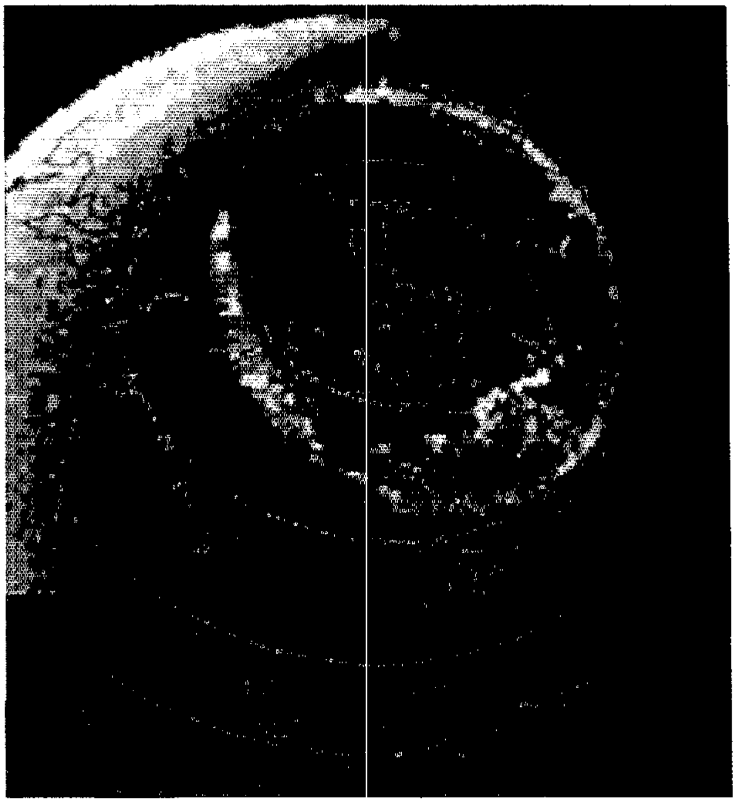
By Tom Frenaye,* K1KI

On August 3, 1981, a pair of research satellites were placed in orbit from the Western Test Range at Vandenberg, California, with a McDonnell-Douglas Thor/Delta 3914 launch vehicle. One of the two NASA/GSFC spacecraft, Dynamics Explorer 1, was placed into an elliptical, polar orbit, with initial perigee and apogee altitudes of 570 km and 23,280 km.[1] What makes the DE-1 satellite of interest to radio amateurs is the auroral imaging instrumentation carried on board that was developed by and is now used by the Physics and Astronomy Department of the University of Iowa. The instruments can record pictures of the entire auroral oval as taken from high altitudes, after centuries of ground-based studies and speculation.

The initial latitude of apogee was 78.2°N and the precession is about 0.328 degrees/day, moving the apogee to 90° N 36 days after launch. Apogee was located over the equator in early June 1982 and over the South Pole in early March 1983. With apogee at polar latitudes, continuous imaging of the aurora for more than five hours can be made during a single orbit (orbital period is 6.83 hours).

The instrumentation carried aboard the DE-1 spacecraft includes two imaging photometers able to "see" light from ultraviolet wavelengths (110-165 nm), through most of the visible light spectrum (175-700 nm). Protective circuits prevent sensor destruction as the dayside of the earth is scanned.

While Amateur Radio operators, in addition to scientists around the world, have been experimenting to learn more about the aurora for many years, the DE-1 spacecraft, and those to follow in the future, may allow current images of the aurora to be received, as easily as television pictures are now relayed by stationary satellites. The complete amateur station in the 1990s may include a color monitor that



This image, the first view of the entire auroral oval from high altitudes, was taken on September 25, 1981 with the spacecraft located approximately 3.27 earth radii (20,860 km) above the North Pole. The light colors are the areas with the highest intensities. Although quite faint along some of its circumference, the auroral oval extends across the terminator into the sunlit hemisphere of the earth. Half of the earth is completely dark except for the auroral oval. The sun is illuminating the earth from the left-hand side of the picture.

displays a picture of the current aurora, including true colors!

Should you think that the aurora doesn't have much effect on communications, remember that while minor auroral activity can block communications across polar latitudes (and black out communications for those in polar latitudes), a major storm can black out communications for the entire radio spectrum, including satellite transmission. In addition, the entire power distribution grid in North America could be disrupted severely or even be completely useless. Ground-induced currents of more than 1000 A have been measured in the Alaska pipeline from more moderate auroral activity. Satellite-borne computers are extremely vulnerable, as is early warning radar in our defense system. Major solar substorms typically happen on the decay phase of the solar cycle. The present decline began in December 1979, and is expected to continue until mid-1987, after which activity will begin to increase. There are 20 solar superstorms on record since 1880, and none since 1960. As with California earthquakes, we are overdue for a big one.

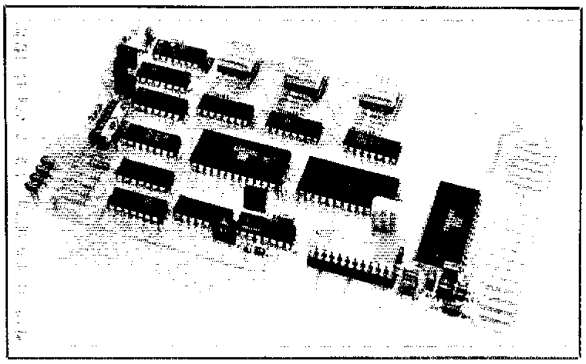
[1]The orbit of Dynamics Explorer 1 is an elliptical polar orbit, meaning that each orbit crosses both the North and South Poles. The highest point of the orbit (apogee) gradually changes from directly over the North Pole to directly over the South Pole over a time period of about 549 days. While over polar latitudes, DE-1 provides excellent coverage of the complete auroral oval.

*P.O. Box 62, Unionville, CT 06085

A DTMF Easy-Controller — With Security

Are unauthorized persons accessing your controller? This unit features an electronic combination lock that should keep "Tone-Pad Charlies" out.

By Phil Czerkies,* WA8KPY



This 16-function control package features a decoder patterned after "The DTMF 'Easy-Ceiver.'"[1] The system was originally designed for remote control of lights in a large auditorium, but it soon became apparent that this controller might be well suited for use as a repeater or remote-base controller.

The system shown in the accompanying photographs consists of a mother board supporting four identical output-control (relay) boards and the decoder board. All the boards in the controller system shown here are coated with a clear, conformal coating to prevent moisture and corrosion problems from occurring. A smaller, two-slot mother board can be used for systems requiring only four control functions.[2]

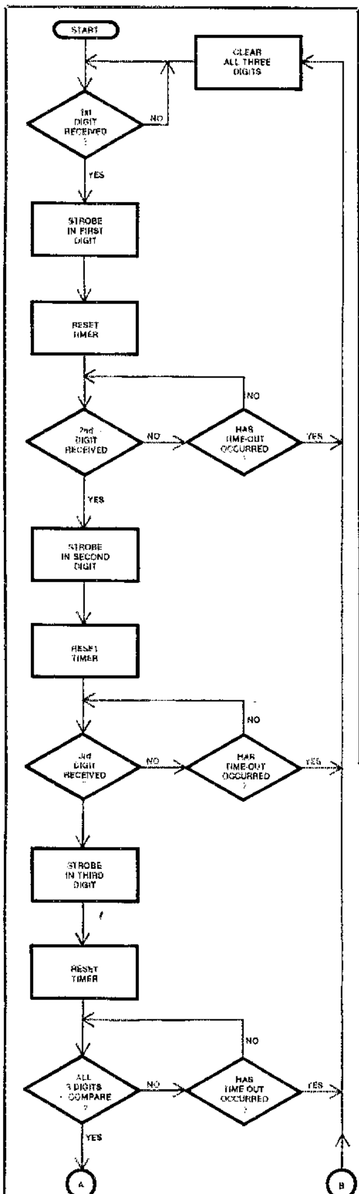
The decoder provides two levels of security. The first level requires the reception of the correct three-digit sequential security code, while the second level requires that each digit be entered within a given time period. Over 4000 code combinations are provided at the first level. Code combinations are selected on the board by means of three binary-coded DIP switches. At the second level, the time-out duration (N) is selected by choosing a single resistor value to provide delays of one, three or five seconds, or longer, if desired.

Security Mode

Each time a digit is entered, the timer is reset and you have N number of seconds to enter the next digit. If there is a pause greater than the time allowed, the previously entered security codes are

[1]Notes appear on page 21.

*11831 E. 62nd Pl., Tulsa, OK 74133



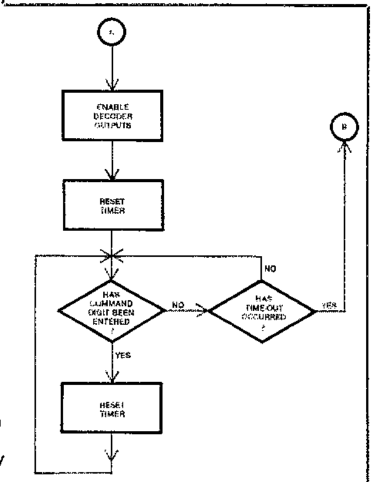


Fig. 1 — Flow chart of the controller security system.

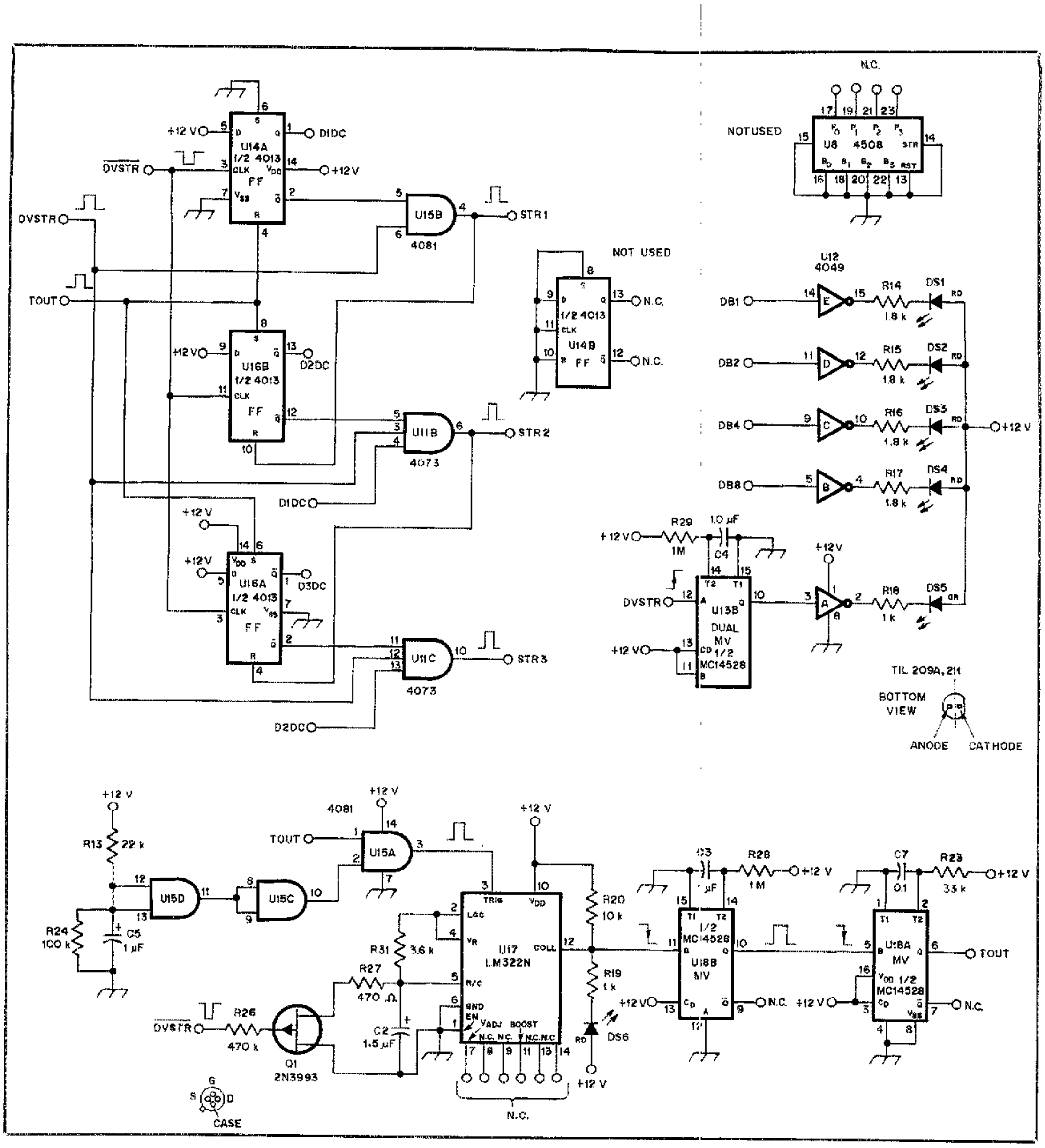


Fig. 2 — Partial schematic diagram of the DTMF decoder. The input section is shown here.

cleared and you start all over again. It's like playing "Beat the Clock."

Entering the wrong security code is like playing "Reverse Beat the Clock." The security codes are compared sequentially. If a bad code is entered, you have to wait until time-out occurs before a new security code can be entered; otherwise, you will never access the system.

User Function Mode

Once the correct security code has been entered, you can issue a full set of 16 commands to the controller. The time-out function still applies. You have N seconds for each digit entered. Let's say N is equal to five seconds. You enter a digit and three seconds pass before another digit is entered. You now have five seconds to enter the next control command.

When you're finished issuing control

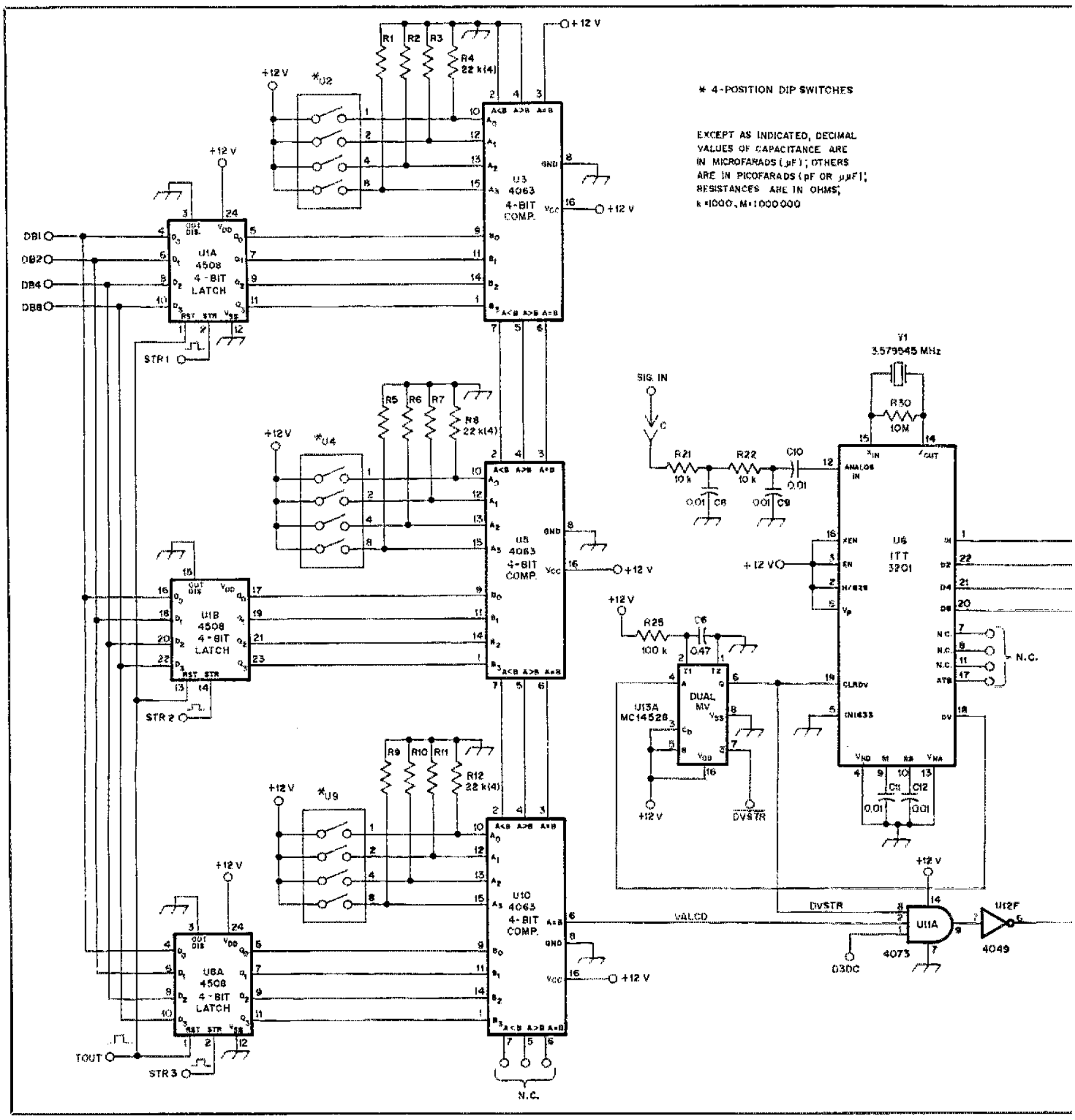


Fig. 3 — The DTMF controller decoding-section schematic diagram. U2, U4 and U9 are four-section DIP switches.

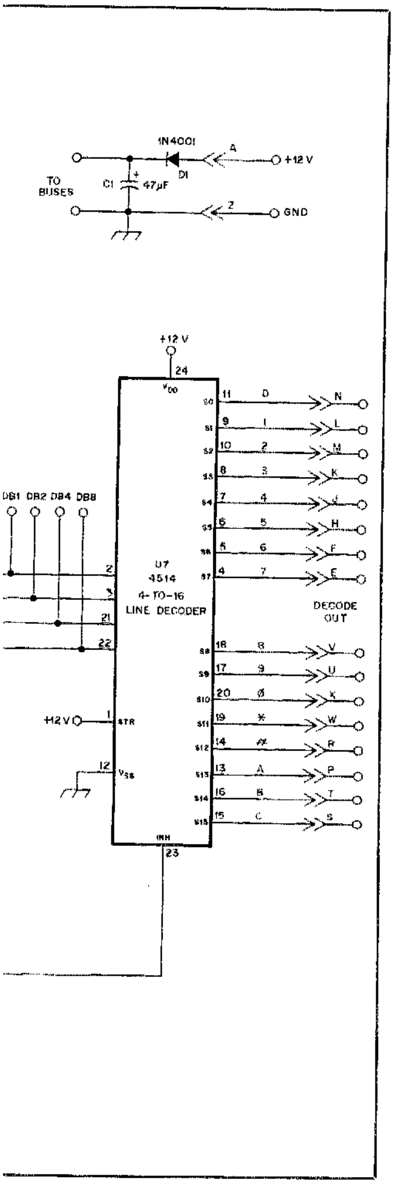
commands, all you have to do is let the system time itself out. Once that happens, you must again pass through security before any further control commands can be issued.

Other Decoder Features

Operation, setup and system troubleshooting are easy. The decoder board uses an ITT 3201 tone-decoder chip that requires a minimum of external parts for operation: a 3.58-MHz (color-burst frequency) crystal and an input filter to keep out the tone harmonics.

There are no adjustments to make except for an initial receiver-audio-level setting done at the radio receiver. LEDs display the binary representation of the decoded digits and presence of the strobe and timer signals. The LEDs are useful indicators, especially the timer LED. It blinks when each time-out occurs, giving a "system heartbeat" effect. So, if you see that the timer LED is not blinking during operation, you know you're in trouble!

The decoder operates on 12-V dc and uses CMOS ICs throughout. For those of us who occasionally take a notion to install boards backwards or to apply voltage of the incorrect polarity to a board, rest easy — the controller system boards have "idiot" diodes installed on each V_{CC} card pin. A pin-for-pin replacement for the



DTMF Easy-Ceiver, the decoder board is about 6.5 inches in length and 3.5 inches wide, just right for a 22-pin card-edge connector.[3]

Decoder Logic

The ITT 3201 IC contains all the high- and low-group filters and other circuits needed for DTMF decoding. It even has an option for selection of hexadecimal or binary output at pin 2.

For those of you who wish to go through the logic of the system, I have included a flow chart (Fig. 1) to make it easier to understand. Also refer to Figs. 2 through 4. The desired three-digit security code is chosen by means of the three four-section DIP switches. When U6 detects a valid digit, pin 19 Digit Valid goes high. This causes U13 to generate two pulses of opposite polarity called Digit Valid Strobe (DVSTR) and Digit Valid Strobe NOT ($\overline{\text{DVSTR}}$). These logic signals, along with the time-out signal (TOUT), are what make things happen and they serve many functions.

The DVSTR signal is used to clear the '3201 via pin 19 of U6 (CLRDV). This allows the '3201 to decode the next valid digit faster than is normal. Usually, the IC detects and decodes a valid digit, and then there is a pause of about 30 ms before it will be ready to decode the next digit. Resetting the IC (DV) from pin 19 (CLRDV) allows you to override this delay time. At U11, the DVSTR signal is ANDed with valid compare digits (VALCD) and Digit 3 Decode (D3DC) from U16 to enable the 4- to-16 line decoder (U7) for actual command functions to occur. DVSTR is also used to trigger U13 and turn on the STROBE LED for 500 ms, making it easier to see.

Timer

One of the functions of $\overline{\text{DVSTR}}$ is to reset the timer after each valid tone is decoded. $\overline{\text{DVSTR}}$ turns on an FET switch that discharges a 1.5-μF capacitor (C2) and resets the timer. U17 is a retriggerable type of timer. It does not time out until C2 charges to about 3 V; only then does its output go low. If the capacitor charge never reaches 3 V (having been reset by newly entered digits), the output will never go low until N number of seconds after the last valid digit was decoded.

The output stage of U17 has a current sink capability of about 20 mA, and will drive the timer LED directly. When the system is first turned on, U15 and the time-out signal (TOUT) form a power-on reset circuit. All the relay boards also have power-on resets. This keeps everything from coming up in a random state when first turned on.

Security Logic

Here's how the DVSTR and $\overline{\text{DVSTR}}$ signals are used to sequentially strobe the three received security digits into latches U1 and U8 and compare them. $\overline{\text{DVSTR}}$ is used as a clock input to U14 and U16. At the same time, DVSTR is input to U11 and U15. When the first digit is received, the $\overline{Q}$ output of U14 changes state, allowing both inputs of U15 to be ANDed, giving a valid strobe pulse (STR1). STR1 is used to latch the first digit into U1 for the compare operation. At this point, the first decoded digit is available for comparison at the inputs to a four-bit magnitude comparator, U3. Here it is compared to the code programmed into the DIP switch, U2. If there is a match, then A = B is true and it is cascaded down to U5. If the second and third digits match in like manner, then the output of U10 (pin 6) generates a Valid Compare Digit (VALCD) signal that will eventually enable U7 via U11 and U12.

Notice that there is some interlocking of this logic with U11, U14, U15 and both sections of U16. For example, when the first digit is decoded and U14 toggles and STR1 occurs, STR1 is used to clear the first section of U16, making it ready for the second digit. Also note that the D1DC signal is routed to pin 4 of U11 and unlocks it for the second digit only after a valid first digit has been received. The same type of interlocking is used on the second and third digits. The only difference is that when the third digit is received and STR3 strobes the third digit into U8 for a compare operation, D3DC is now an AND input to U11 and finally unlocks the 4- to 16-line decoder (U7) for 16 function outputs.

Output-Control Boards

The output-control boards are of a straightforward design and are identical physically and electrically, allowing them to be placed in any one of the four slots on the mother board. This also allows easy duplication when more expansion is necessary. Each board is comprised of identical control circuits and shares a common power-on reset circuit. The output control boards also have LED indicators on them. This aids in verifying that the correct function was selected, as comparison can be made with the decoded-output LEDs on the decoder board.

Plug-in 12-V relays are used in this design. Some might ask, "Why use a relay?" Well, relays do have a lot going for them, especially in the areas of isolation, power-handling capabilities and price! Relays are generally easier to acquire, and in this installation, if one does go bad, just pop off the retainer clip, plug in another relay and you're back in business.

Although the relay contacts are rated for about 7 A, they shouldn't be required to operate at their maximum ratings. This is because inductive devices, such as motors, can draw much more than their normal running current when switched on. The high current can weld contacts or otherwise damage them. If you must switch high-power devices, use the on-board relays to activate a contactor or an SCR controller that is capable of handling the load.

That inductive device mentioned earlier will also generate a high-voltage spike. That spike will radiate like a radio wave and may find its way back into the logic circuits and disrupt control settings. When switching other than signal sources, it's always a good idea to fuse the controller leads. This pre-

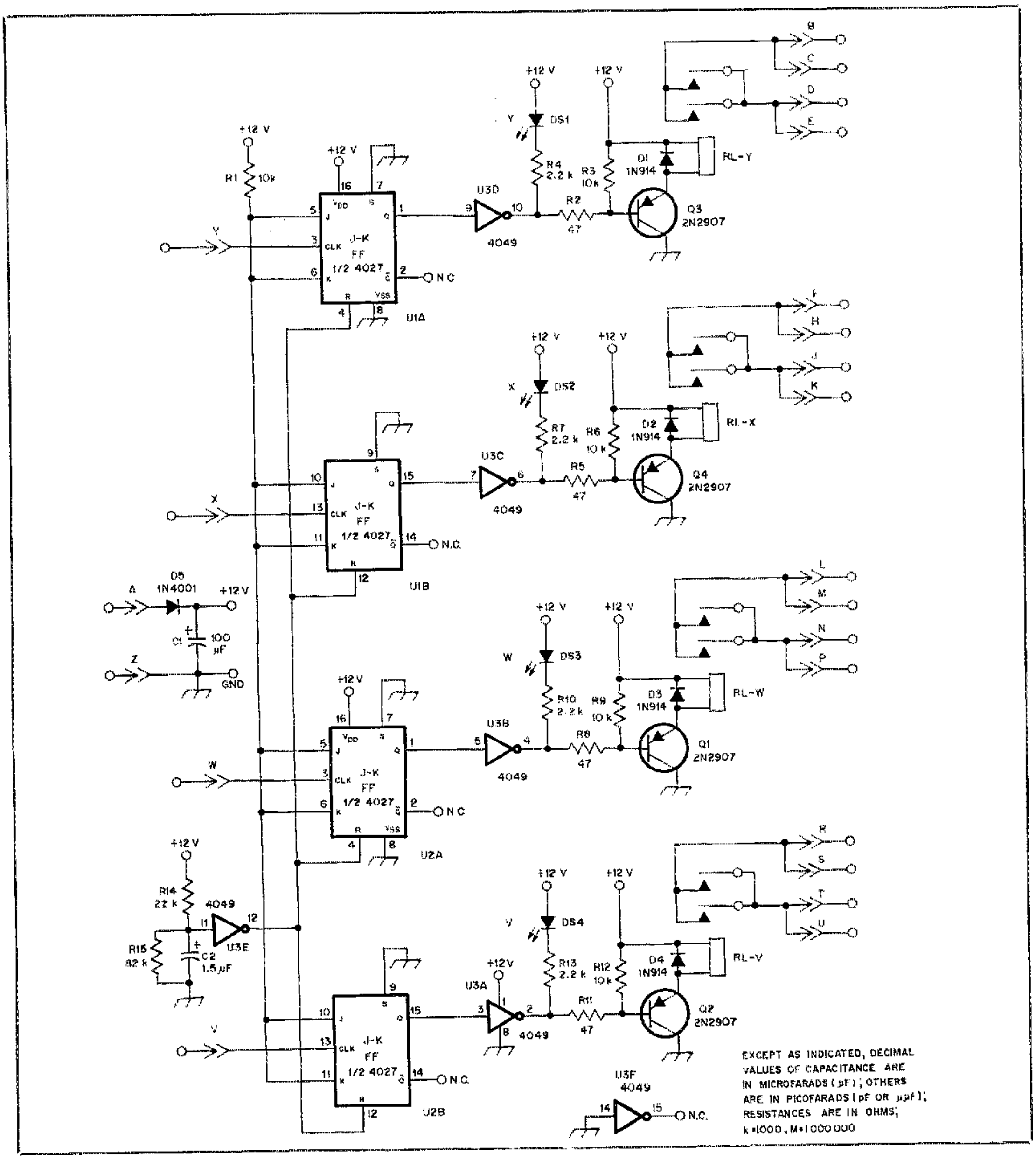


Fig. 4 — Schematic diagram of the controller output board. The relays used by the author are Potter & Brumfield R10-E1-W2-V185 types with a 12-V dc coil.

caution may keep a few boards from being burned up should something short.

Output-Control-Board Logic

The output signals from this board are routed along the mother board to each relay board in groups of four signals each. When they arrive at the output-control board, they are latched into respective latches. The Q output of each 4027 section (U1, U2) is fed to one section of a 4049 high-current driver IC (U3). The driver turns on a LED indicator and a transistor (Q1 through Q4) that finally energizes a 12-V relay. The relay contacts are paralleled to handle large currents. Connections to the relay contacts are routed to terminal strips on the lower edge of the mother board. The operational sequence of events is simple. One input pulse from the decoder board energizes a relay and another, identical pulse from the same line will de-energize the relay.

Mother Board

This board is designed for heavy-duty use. It is 1/8 inch thick and has a two-ounce

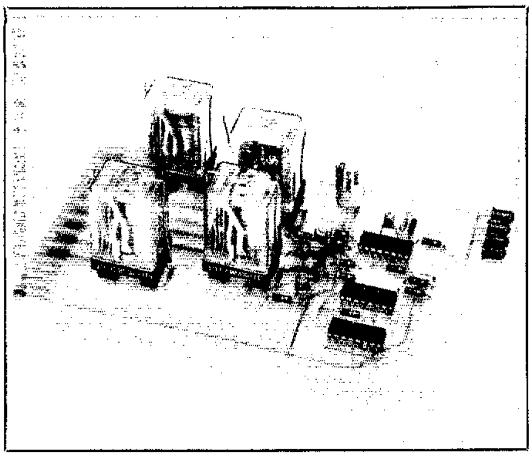


Fig. 5 — One of the output boards used in the controller.

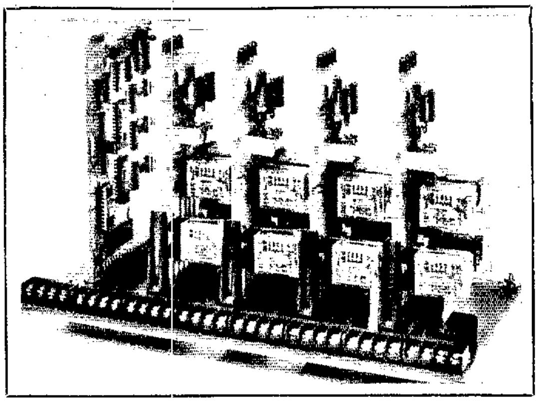


Fig. 6 — The completed controller. At the left is the decoder board; four identical output boards follow. A 1/8-inch-thick mother board supports the assembly. For applications requiring only four control functions, a two-slot mother board is available.

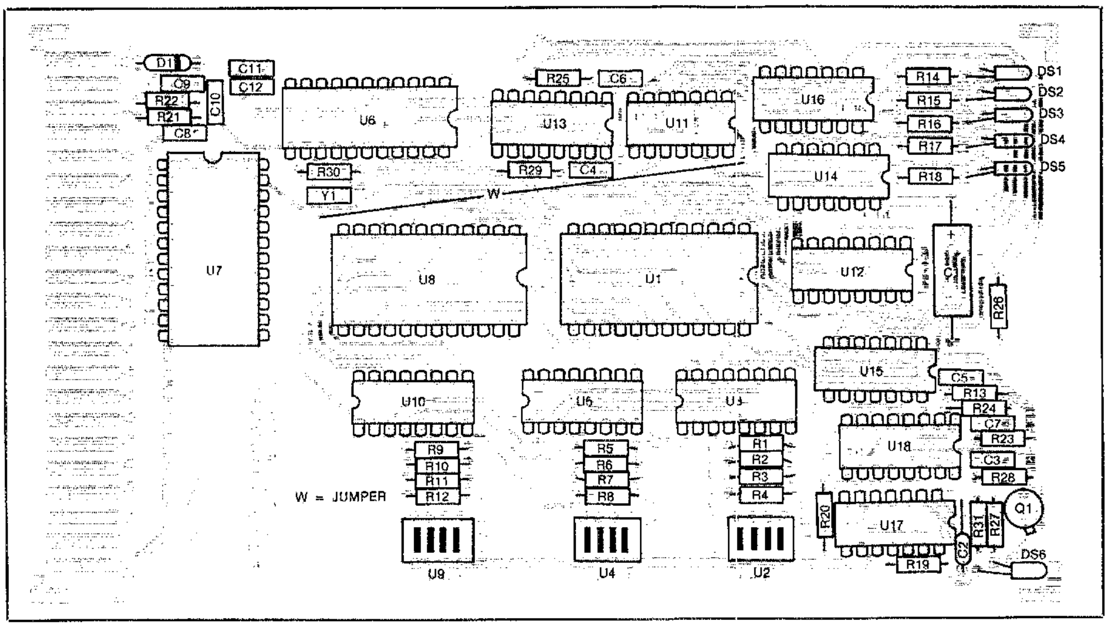


Fig. 7 — Parts-placement guide for the DTMF decoder board. Parts are placed on the top (lettered) side of the board. The shaded areas represent an X-ray view of the copper pattern. See Hints and Kinks, this issue for the etching-board patterns.

copper foil layer. The board thickness should help reduce flexing and cracking that may occur when inserting or removing the daughter boards. There are five 22-pin connectors on the board and screw-type terminals along the bottom and right-hand side of the board. The first four connections on the left side of the mother board are used for tone inputs and dc power to run the controller. The remaining terminals are assigned by the user. The backplane decoder bus on the mother board can route any control digit function to any output on each of the four output-control boards by means of jumper wires.

Summary

I am sure this controller will find many applications. It is compact, rugged and easy to use. At the present time, one controller is installed and several are scheduled to be installed at different repeater sites. Why not try it yourself?

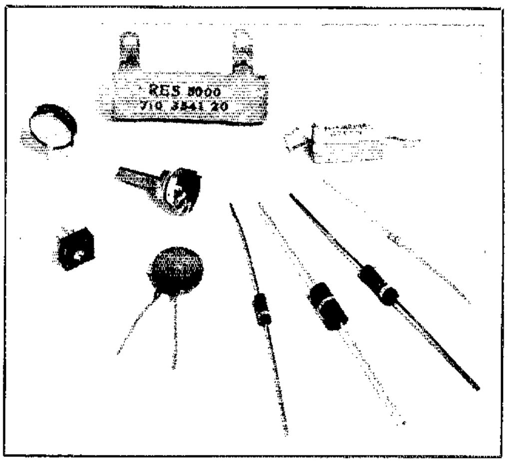
Notes

[1] J. Jarrett, "The DTMF 'Easy-Ceiver,'" *QST*, Jan. 1982.

[2] Boards, kits and assembled and tested boards of each type are available from the author. Please include an s.a.s.e. with your inquiry. Pc-board templates and overlays are available from ARRL Hq. for a large s.a.s.e. and $2.

[3] mm = in. × 25.4; g = oz × 28.35.

The Manufacture and Use of Resistors (Resistors for the Experimenter)



Have you been left wondering why a certain resistor type is specified in the latest *QST* project? What design parameters must be considered? Here are some answers to these questions and others you may not have even thought of yet!

By Larry D. Wolfgang, * WA3VIL

Most electronics construction projects use at least some resistors. Different types of components are often specified for particular sections of the project, such as metal-film resistors and Mylar® or polyester capacitors in a VFO. Most neophytes, and many seasoned veterans, become confused when it comes to substituting components with a different construction than the one specified

Why can I use this 5-W, 10-kΩ wire-wound resistor to replace a 1/2-W, 10-kΩ composition resistor in a sidetone oscillator, but not one in the local oscillator of my hf rig? There seems to be a limited amount of information available concerning these problems. In this Basic Radio installment I will present some information about the manufacturing processes and the specifications for the various resistor types commonly used in amateur projects. After reading this article, you should have an understanding of why a certain type of resistor is specified for a project, and you should be able to make informed decisions about what type of substitutions may be acceptable to ensure proper circuit operation.

Resistance

All materials exhibit some opposition to electric-current flow. The resistivity of the material is a measure of this opposition. Most metals have a fairly small value of resistivity (1.673 μΩ-cm for copper) and so are good conductors. Insulating materials have a high value of resistivity. Powdered carbon (graphite) has a resistivity of 1375 μΩ-cm.[1]
Given the resistivity (ρ) of a conductor, you can calculate the resistance by:

$$R = \rho \frac{\ell}{A} \qquad \text{(Eq. 1)}$$

where
R = resistance
ℓ = length of the conductor
A = cross-sectional area of the conductor

Resistors can be constructed by measuring a length of wire that will have the desired resistance. Sometimes, the wire is wound onto a cylindrical form to make a neat package. Materials with a higher resistivity will require shorter lengths for the resistor. Mixtures of carbon and clay are formed around pieces of wire to make carbon-composition resistors.

Most amateurs will not be interested in "rolling their own" resistors, so finding the value of resistivity, length, and cross-sectional area may not be the most practical method of determining resistance. I am sure you are all familiar with Ohm's law. Some of you have probably conducted an experiment to verify this basic principle of electronics. If you have never done this, now would be a good time to try an experiment.

A Simple Experiment

For this two-part experiment you will need a variable-voltage power supply, a VOM (or two) and an assortment of 1-W resistors. Select a 150-Ω resistor, and wire a circuit similar to that shown in Fig. 1. Starting at 0 V, increase the supply voltage in 2-V steps, recording the current through the resistor at each step. When you have reached 14 or 16 volts you can plot a graph of voltage vs. current. You should see a straight line, or linear relationship, where any applied voltage divided by the corresponding current value gives a result of approximately 150!

For the second half of the experiment you will use a variety of resistors, maintaining the same voltage drop (about 10 V) across the resistor each time. One VOM is connected to measure the voltage across the resistor, and the other measures circuit current. One meter can be used if you are willing to switch it between these circuit posi-

* Assistant Technical Editor

[1]Notes appear on page 26.

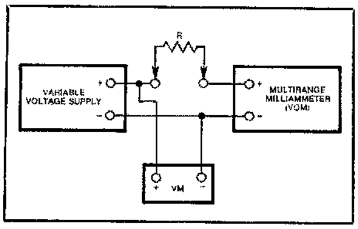


Fig. 1 — A diagram of the equipment arrangement for part (A) of the experiment to verify Ohm's law. The supply voltage is gradually increased, and the current through the resistor is measured at each step.

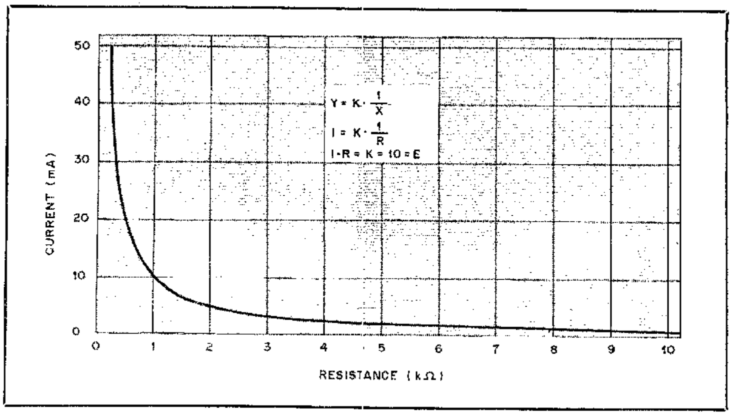


Fig. 2 — Graph of typical results from part (B) of the Ohm's law experiment.

tions. Select 10 or more resistors ranging from about 47 Ω to 10 kΩ. Be sure to record the exact values used. You may have to adjust the power-supply voltage slightly for each resistor in order to assure a constant voltage throughout the experiment. Now plot another graph, this time with resistance on the X axis and current on the Y axis.

This graph will not be a nice straight line but a curved line with very large current at low resistance and a current that is asymptotic to zero (or gets closer and closer to zero, but never actually becomes zero) as the resistance becomes very large. See Fig. 2 for a sample graph. This curve is called an inverse proportion. The product of any current value and a corresponding resistance will give the value of voltage you used. Both parts of this experiment illustrate the basic principles of Ohm's Law.

Construction

Carbon-Composition Resistors

Perhaps the most common type of resistor is the carbon-composition type. A mixture of carbon granules and clay or another bonding agent is made. The proportions of carbon and clay determine the approximate value of the resistors to be made from each batch. The compound is then formed around a pair of wire leads and baked. Finally a protective, insulating coating is put on the resistors and they are "graded out" as to resistance value (Fig. 3A). A certain batch can be made to target a particular resistance, but the actual resistors will have a range of values.

All of the resistors in a batch will be within ±20% of a standard value. Tolerances of ±10%, ±5%, and even ±2% can be obtained by sorting, or selecting individual resistors from each batch. The stability of carbon-composition resistors is not good enough over a typical operating lifetime to warrant selecting tighter tolerances. These resistors are manufactured with power ratings of 1/8, 1/4, 1/2, 1 and 2 W by controlling the diameter of the resistance element.

The standard component values, used for fixed resistors, some capacitors and other small components, are based on a number series. Each value differs from the previous value by a constant multiplier. The preferred values are rounded to two significant figures to make up the series. For tolerances of ±20%, ±10% and ±5%, the multipliers are $10^{1/6}$ (1.47), $10^{1/12}$ (1.21) and $10^{1/24}$ (1.10), respectively.

Most small calculators will find a power of ten. After you obtain the appropriate power of ten you simply multiply the number by itself, and each succeeding number by the same constant. Try generating one of these series, then compare your answers with Table 1. A complete listing of standard values can be found in *The ARRL Electronics Data Book.*[3]

Table 1
Standard Resistor Values

Tolerance

±20%	±10%	±5%
10	10	10
		11
	12	12
		13
15	15	15
		16
	18	18
		20
22	22	22
		24
	27	27
		30
33	33	33
		36
	39	39
		43
47	47	47
		51
	56	56
		62
68	68	68
		75
	82	82
		91
100	100	100

Note that only one decade of values is given here. To find larger (or smaller) values, multiply by multiples of 10 (or divide by 10).

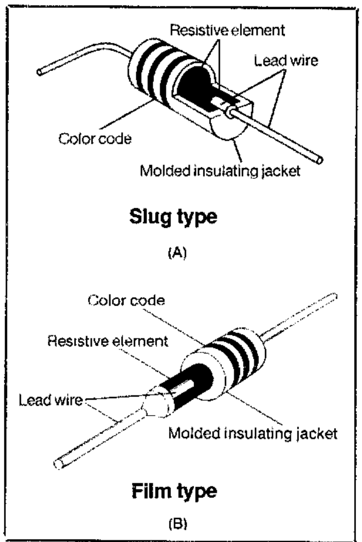


Fig. 3 — The construction of a carbon composition resistor is shown at A. B shows the construction of a carbon film resistor. (drawings courtesy Stackpole Components Co., Raleigh, North Carolina)

Some advantages of carbon-composition resistors are the wide range of available values, low inductance and capacitance, good surge-handling capability, and the ability to withstand small power overloads without being completely destroyed. The main disadvantages are large resistance changes with temperature change, operating time or on-the-shelf aging, and moisture or humidity. Several major manufacturers no longer make carbon-composition resistors, mainly because of the high cost and poor stability as compared to film-type resistors.

Operation at frequencies above 1 MHz will decrease the effective resistance of carbon-composition units, and this should be taken into account in designing circuits. Other design goals should be to provide adequate cooling, to operate the resistor at no more than half its allowable power rating, and to not mount other heat-producing components within one diameter of the resistor.[4]

Carbon-Film Resistors

Another type of carbon resistor is the carbon-film unit. These are manufactured by using high temperatures to break down certain gaseous hydrocarbons. The resulting carbon is then deposited in a thin layer or film on a cylindrical ceramic form. The resistor is sealed with a plastic or other insulating material (Fig. 3B). They are commonly available in 1/8-, 1/4- and 1/2-W sizes with tolerances of ±5% and ±2%. The thickness of the deposited film provides a means to control the final resistance.

The major advantages of carbon-film resistors are low cost and improved stability with age and temperature changes. These resistors cannot withstand electrical overloads or surges. In fact, the film will open quickly under these conditions, so they can actually be used as fuses for some applications. Corrosion under the end caps was reported to cause open circuiting with early-production units; even those kept in storage. Silver-plated end caps and improved moisture-resistive coatings seem to have cured this problem however.

Metal-Film Resistors

Metal-film resistors provide much better temperature stability than other types, but they cost a bit more. As the manufacturing process is improved, prices continue to fall. Metal-film resistors, currently priced between carbon-composition and carbon-film types, are becoming increasingly popular for many general-purpose applications.

There are a variety of manufacturing methods being used to make metal-film resistors. One technique is to vacuum deposit a thin layer of Nichrome or other resistive alloy on a glass or ceramic form. Additional methods involve screening and other coating techniques to produce a film of metal on the base or substrate. This film is trimmed in a spiral, or helical, fashion to obtain tolerances of from ±5% to ±0.1%. The amount of film that is trimmed away determines the path length and conductor size in the resistor. This trimming can be done on a mechanical lathe, or by means of a laser. Fig. 4 illustrates the much-higher-resolution track that can be cut using laser technology.

The spiral track means that metal-film resistors may tend to exhibit some inductance, especially at frequencies above about 10 MHz. This effect is increased for the higher-value resistors because a tighter spiral will be required. There is very little skin effect with film resistors because of the thin film, so the resistance will not decrease with increasing frequency.

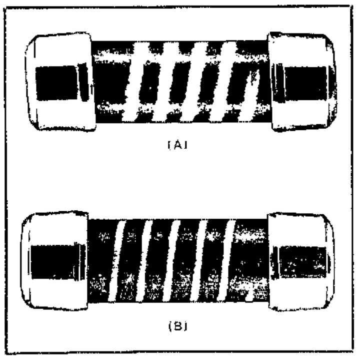


Fig. 4 — Metal-film resistors are shown in these photos. At A is a spiral track cut in the film using a mechanical lathe, and at B is a track cut using laser techniques. (photos courtesy RCD Components, Inc., Manchester, New Hampshire)

Metal-film resistors are generally low-power units (1/10 to 3 W), but they can be made in power ratings up to 20 W. After the film has been deposited and cut to specification, the entire unit is coated with a thermosetting resin to protect the resistor from mechanical damage or damage from moisture or chemicals.

Metal-oxide resistors are similar to metal-film resistors. A vapor or spray of tin-chloride solution is deposited on a heated glass substrate. Additives are used to modify the properties of the resulting tin-oxide film. This allows the manufacturer to produce specific temperature coefficients and control other resistor parameters. These resistors produce very little noise and exhibit excellent high-temperature operation. They are generally flame proof and can be made with power ratings up to 400 W, although the most common ones are in power ratings of 10 W or less.

Wirewound Resistors

Wirewound resistors are probably the most diverse type in terms of construction technique. Typically, a Nichrome wire (or other resistance alloy) is wound on an insulating form. The core may be a ceramic rod or tube, fiberglass or other material. It may be cylindrical, flat or rectangular. The completed resistor has leads welded in place, and the entire unit is coated with vitreous enamel, a protective silicone coating, cement or other insulating sealant (Fig. 5). Wirewound resistors are made in power ratings from 2 to 1500 W, with resistances of less than an ohm to a megohm. They are highly temperature stable, and can handle large power overloads if mounted in the clear to dissipate the heat.

Wirewound resistors have a significant inductance and distributed capacitance, and so are not suitable for use at radio frequencies. Some resistors are made with a so-called "non-inductive winding." These have the winding direction reversed half way through, so that the inductances will cancel. Even these may have an inductance of as much as 0.5 μH at 1 MHz, however.

Other Resistor Types

There are a variety of other resistor types on the market. Chip resistors are low-power, leadless devices for low lead-inductance rf applications. Amateurs are most interested in chip resistors for use in uhf and microwave equipment (Fig. 6). Thick-film resistor networks provide a variety of configurations in single in-line packages (SIP). These are also available in dual inline packages (DIP).

Two types of nonlinear resistors may also hold some attraction for the amateur. Thermistors exhibit a specific, reliable resistance change as the ambient temperature changes. These can be used in temperature sensing circuits and other related applications. Voltage-Dependent Resistors (VDR) or Metal-Oxide Varistors (MOV) have a significant, nonlinear change in resistance with a changing voltage applied to their terminals. When the voltage increases to a certain point, there is a sudden drop in resistance, allowing a much larger current to flow. Varistors find ap-

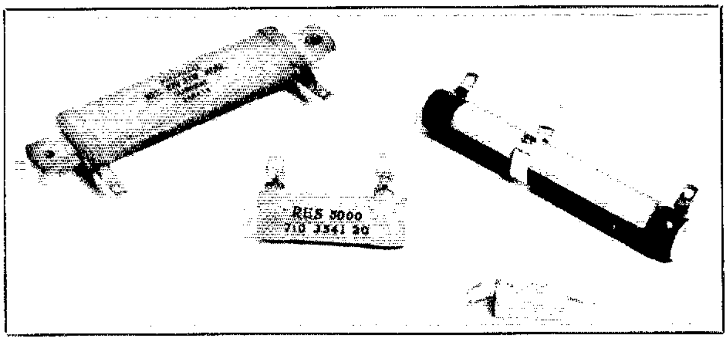


Fig. 5 — Examples of some wirewound resistor types.

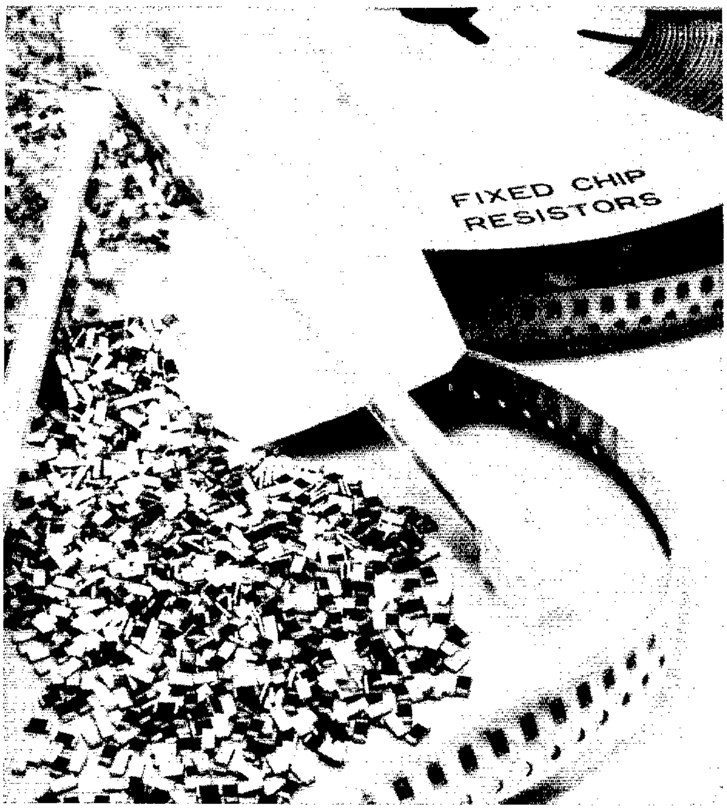


Fig. 6 — Photo of some chip resistors. (courtesy Stackpole Components Co., Raleigh, North Carolina.)

plications in RFI suppression, transient-voltage suppression, contact protection and other circuits. They resemble large ceramic capacitors, as can be seen in the title photo.

Variable Resistors

Variable resistors are available in a variety of styles and power ratings. The most common construction technique is to use a carbon-coated substrate as the resistance element, with a spring-loaded movable contact as the means of changing the resistance. Most of these require a rotation of slightly less than one full turn to cover the entire resistance range.

Another style of variable resistor is the multiturn trimmer potentiometer. These normally consist of a flat ceramic base coated with carbon material. The spring-loaded contact is moved by means of a threaded rod (see Fig. 7), and the entire unit may be sealed in plastic or have a metal cover riveted in place. Some of these small trimmers may have a wirewound resistance element, so you should be cautious about using them in rf circuits unless you can verify the construction.

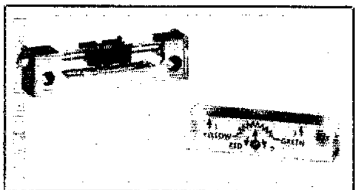


Fig. 7 — This photo shows the general construction of a multiturn trimmer potentiometer.

Wirewound resistors can include a moveable contact, either as a spring-loaded adjustment or as one that is set by hand. These can be rated to dissipate large amounts of power, and are often used as a type of autotransformer.

As you adjust the resistance of some potentiometers, you will notice that the resistance will change equal amounts for each degree of control-knob rotation. This is called a linear-taper potentiometer. A far more common situation is the one in which the resistance change varies dramatically as you turn towards the high-resistance end of the scale. This is called an audio-taper, or logarithmic potentiometer. Such a resistance change is designed to approximate the way a human ear detects changes in volume of a sound. At low volume (low resistance above ground) a fairly small change in volume will be noticeable, but at higher volumes the ear is not as sensitive to changes, so a fairly large change in resistance is required to produce a noticeable volume change.

Table 2
Nominal Resistor Dimensions

Power Rating (Watts)	*Body Length (Inches)*	*Body Diameter (Inches)*
1/8	0.145	0.062
1/4	0.248	0.094
1/2	0.375	0.142
1	0.562	0.225
2	0.688	0.313

mm = in. × 25.4.

Tolerance Ratings

Resistor manufacturers specify a number of parameters that are important if you are to select a suitable resistor for a certain application. Perhaps the most obvious is the power rating. A good design goal is to not exceed about half the rated power dissipation for the resistor. Probably the easiest way to determine power ratings is by physical size. The larger the device, the more power it can handle. Table 2 summarizes the resistor body length and diameter measurements for common-size carbon-composition resistors. Film resistors have a slightly smaller diameter. You can always identify a film resistor by the thinner section in the middle. This is caused by the end caps that are placed on the resistor before it is coated and color coded. Wirewound resistors will usually have a power rating printed on the resistor body.

The value of a resistor is normally specified to within a certain tolerance, such as ±10%. This tolerance specifies the maximum variation from the nominal value that the resistor can be expected to have. In addition, the actual resistance will change with temperature, humidity and time in use.

The Temperature Coefficient Range (TCR) rating is usually given in parts per million per degree Celsius (ppm/°C). Resistors can have a positive TCR (resistance increases with a temperature increase) or a negative TCR (resistance decreases with a temperature increase). Power dissipated in the resistor, heat radiated from nearby components and ambient air temperatures can all affect the value of a resistor, so these factors must be considered when you are designing a circuit. Electrical conditions can be controlled somewhat by specifying a certain TCR. For example, if a resistor with a positive TCR is used in a critical circuit, as the current through the resistor increases, the resistor will heat up. This will cause the resistance to rise, reducing the amount of current flowing in the circuit.

Manufacturers also specify the amount of resistance variation with relative humidity, resistance change after operation at a rated load for a certain time, such as 1000 hours, and a load derating factor that indicates the maximum ambient temperature to use the resistor at.

All amateurs should be familiar with the standard color code used to indicate values and tolerances. Table 3 summarizes this code. You will find three or four color bands on most resistors. One band will be closer to one end, and this gives the code for the first significant digit of the resistance. The second band indicates the value of the next digit. The third color band tells you how many zeros to add, and the fourth one indicates the tolerance. For carbon-composition resistors, the tolerance band will either be missing (±20%), silver (±10%) or gold (±5%). High-precision film resistors may have five color bands — three significant digits, a multiplier and a tolerance band. The tolerance band on these resistors may be gold, red, brown, green, blue or violet, as shown in Table 3.

Table 3
Resistor Color Code

Color	*Significant Digits*	*Multiplier*	*Tolerance (± %)*
Black	0	—	—
Brown	1	10	1
Red	2	100	2
Orange	3	1000	—
Yellow	4	10000	—
Green	5	100000	0.5
Blue	6	1000000	0.25
Violet	7	—	0.1
Gray	8	—	—
White	9	—	—
No Band	—	—	20
Silver	—	0.01	10
Gold	—	0.1	5

If you happen to come across a resistor with a single black stripe, don't be confused. These "zero ohm" resistors are simply a convenient way for manufacturers to place a jumper on the circuit board using automatic-insertion equipment!

Some types of resistors have actual values and ratings printed on them. To determine the other resistor specifications, you will have to consult the manufacturer's literature.

The average ham may not need to know all of these specifications in designing a simple piece of equipment, but being aware that they exist should help in making informed decisions concerning the suitability of a component for substitution.

Notes

[1] R. Weast, ed., *Handbook of Chemistry and Physics* (Cleveland: The Chemical Rubber Co., 1969).
[2] *Reference Data for Radio Engineers* (Indianapolis: Howard W. Sams & Co., Inc., 1975), Ch. 5.
[3] D. DeMaw, *ARRL Electronics Data Book*, (Newington: ARRL, 1976), p. 8.
[4] See note 2.

Strays

☐ [*Editor's Note: From time to time, we'll be publishing a profile on one of our Contributing Editors, to acquaint readers with people not on the Hq. staff who write columns that appear in* QST *month after month. Here is the first installment.*]

CE PROFILES

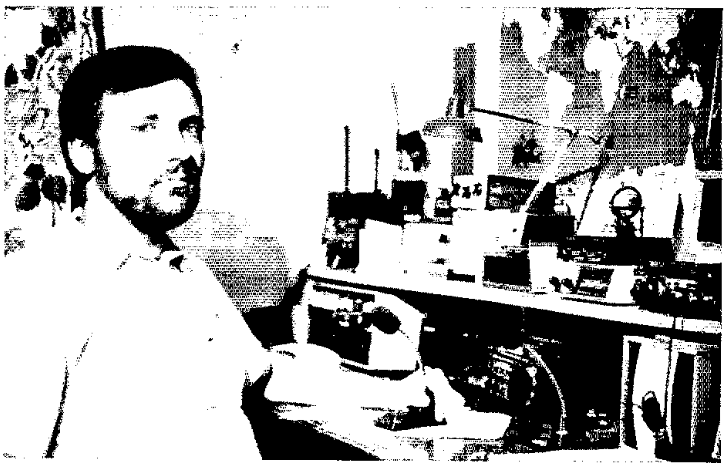
Contributing Editor Stan Horzepa, WA1LOU, has the distinction of conducting not one but two *QST* columns — FM/RPT and On Line. Stan became the conductor of FM/RPT in June 1979 having been on the ARRL Hq. staff for about two years. Two months later he left Hq. but continued with the column as contributing editor. On Line debuted in August 1981.

Stan's interests in repeaters and computers predate the multitudes that now populate these two fields of Amateur Radio. "Back in the early '70s, I had the good fortune of meeting (via 2-meter a-m) WA1NQP, who built one of the first 450 repeaters in the Northeast. That got me started in the world of 'boat anchors' and mini-rigs," says Stan. "And, when Radio Shack unveiled their first TRS-80®, I was the first one on my block (and in town) to buy one!"

First licensed in 1969, Stan is an Extra Class licensee and an ARRL Life Member. He has held a number of ARRL appointments (ASCM, NM, OBS, ORS, OVS) and is the former SCM of Connecticut. He has numerous operating awards, including A-1 Operator, BPL, DXCC, Public Service, PSHR, WAC, WAS and the Central Radio Club Cosmos silver medal for being among the first to operate through the Soviet amateur satellites.

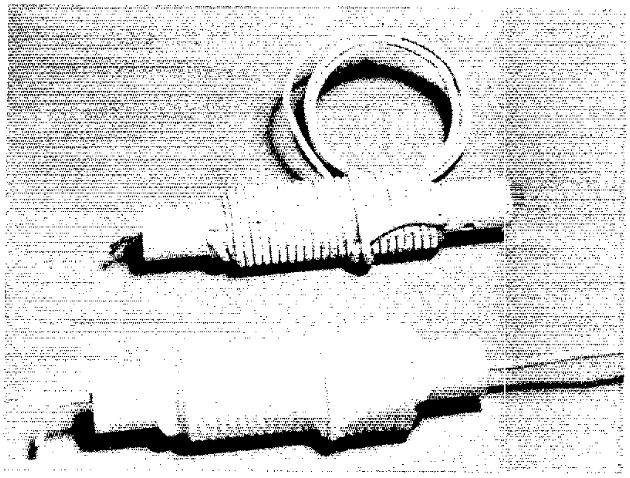
Stan lives in Wolcott, Connecticut, and is a senior technical writer for General DataComm Industries, Inc. He has a BA from the University of Connecticut and a JD from Western New England College. In 1977-78, Stan was a communications assistant in the Public Service Branch of the League's Communications Department. This sparked Stan's interest in traffic handling, and led to his founding of the Western Connecticut Net in the midst of the blizzard of 1978.

Stan operates RTTY, chases hf and vhf DX (278 countries worked on hf; 25 states on 144 MHz), and participates in contests. When not radiosporting, he raises giant squash, writes a book on UFOs and roots for the Boston Red Sox.



CE Stan Horzepa, WA1LOU, has his hands full these days alternating between columns on repeaters and computers. *(photo by Jeanette Horzepa)*

Dual-Frequency Antenna Traps



Although L-C antenna traps have been around for years, you've never seen any like these!

By Robert H. Johns,* W3JIP

Here is a new way to make antenna traps using only coils, without scarce and expensive high-voltage capacitors. An additional bonus is that these traps can be made to resonate simultaneously on two frequencies, greatly expanding their capabilities!

Cross-Linked Polyethylene (XLP) Insulation

The key to these new traps is a specially insulated wire that withstands several kilovolts. This wire is wound one layer on top of another, which produces some capacitance between the layers (Fig. 1). Enough capacitance in parallel with the coil inductance produces a resonant circuit, which can be used as an antenna trap. In transmitting service, this capacitor (formed by the insulation) will need to withstand high voltage and high currents without much dielectric loss. XLP, a recently developed type of insulation, has the excellent high-voltage and low-dissipation properties of polyethylene. It is also tough and hard from additional polymerization (or cross linking) of the molecules.

All of the traps described in this article are made from wire covered with XLP insulation. It is normally used for telephone-switchboard service and costs about 10 cents a foot for no. 14 stranded, type-SIS wire.[1,2]

How They Work

In Fig. 1, notice that the two coils are wound in bifilar fashion. The more turns added to the two coils, the lower the resulting frequency, since inductance and capacitance are increasing simultaneously. This tuned circuit is similar to the filters described by Doty, and is an extension of the coaxial cable traps of O'Neil and Johns.[3,4,5]

A second resonant frequency is achieved by winding a small *third* layer on top of the first two (Fig. 2). Another tuned circuit is formed by the capacitance between the outermost coil and the two others, together with the inductance of the small coil. This higher-frequency resonance appears only if the third coil is less than half

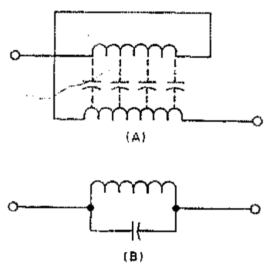


Fig. 1 — Illustration of distributed capacitance between coil turns (A). Electrical equivalent circuit of A (B).

*Chief Engineer, Barker and Williamson Co., 10 Canal St., Bristol, PA 19007

[1]Notes appear on page 30.

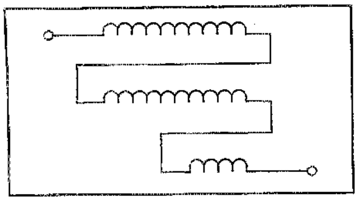


Fig. 2 — A dual-frequency antenna trap. The lower resonant frequency is determined by the entire three-layer coil and distributed capacitance. The higher-frequency resonance is determined by the smallest coil and associated capacitance. The two circuits interact, but can be adjusted individually by varying the number of coil turns.

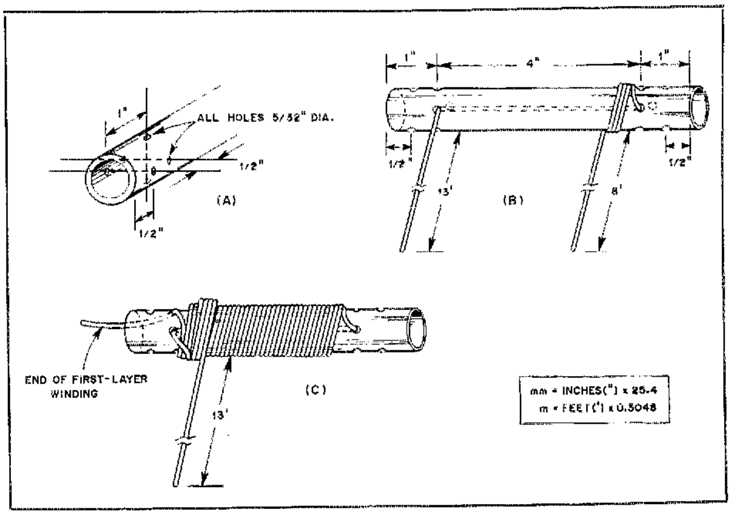


Fig. 3 — Construction details for a 40/15-meter trap. All text instructions follow this orientation of the coil form. Holes should be drilled as shown in A. First-layer coil winding is shown in B, and second-layer details in C.

the length of the larger coils. If too many turns are added, this resonance gets broader and shallower, finally disappearing when the coil becomes too large.

Dual-frequency resonance can be used to good advantage in a trap antenna. The common trap dipole for 80 and 40 meters will also work on 15 meters if the traps are tuned to 21 MHz. The half-wave section on 40 meters will function as three half-wave elements on 15 meters, and can also be fed at the center.

Construction Principles

The basic trap is a 40/15-meter version, and is made from ½-inch PVC pipe. Cut the PVC pipe to 6-inch lengths. One length is needed for each trap. Drill holes in the pipe, as shown in Fig. 3A. The form is now ready to accept the winding.

Start winding the trap by passing a 21-foot length of wire through the center of the form. Pass the ends out through a set of holes spaced 4 inches apart. One end of the wire should protrude approximately 8 feet from the form, the other approximately 13 feet. With the 8-foot wire on your right, wind it toward the left edge of the pipe for a total of 25¼ turns (Fig. 3B). Feed the end of the wire through the appropriate hole and out through the center of the trap to form a pigtail. Now coil the other wire on top of the first layer by winding in the opposite direction of rotation, laying each turn in the spaces between first-layer turns (Fig. 3C). Skip a space at the beginning of the second layer, for it has only 23 turns. This coil should finish at the right side of the trap. Feed the remainder of the wire *completely* through the trap form by passing it through two opposing holes.

Take the end of the second-layer wire (there should be about 2 feet left) and lay it perpendicular to the second-layer turns. Count 10 spaces from the right end of the outer coil and make a 90° bend in the wire at this point (Fig. 4). Lay the wire into this space and wind 9¾ turns over the second layer, making sure the coil is following the same direction of rotation as the second layer. The third layer is actually wound *over* the wire that lies perpendicular to the

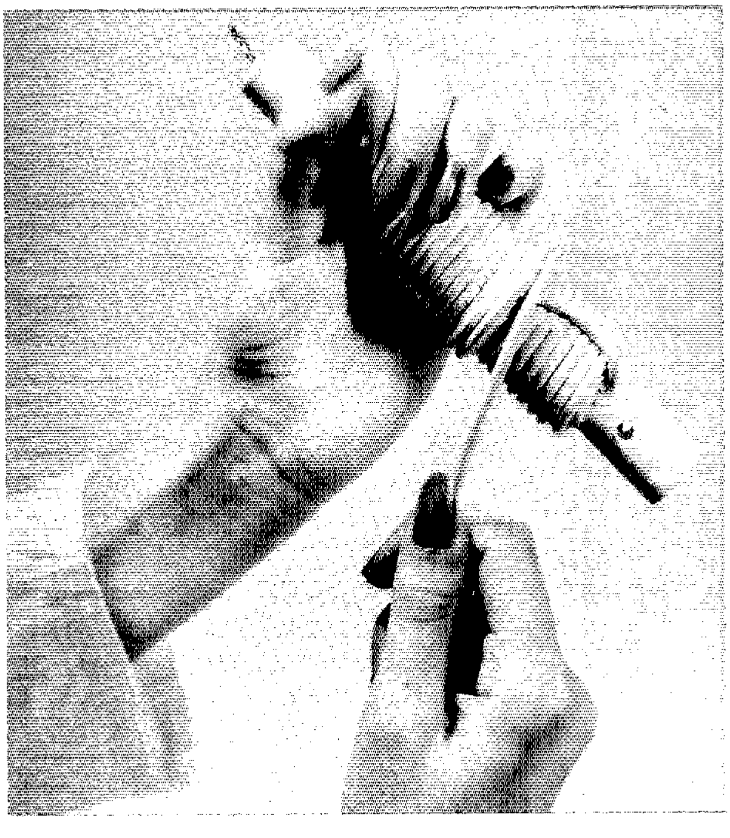


Fig. 4 — The third-layer coil winding technique for a 40/15-meter trap. This coil is wound on *top* of the wire lying over the second-layer coil.

coils. This outermost coil should end at the right side of the form. Pass the remaining wire through the appropriate hole and out through the center of the trap. You should now have a completed trap with three layers of windings and a pigtail of wire protruding from each end.

A Four-Band Antenna

Fig. 5 shows the dimensions of a trap dipole using one pair of the 40/15-meter traps described above. It is resonant on 80, 40, 15 and 10 meters.

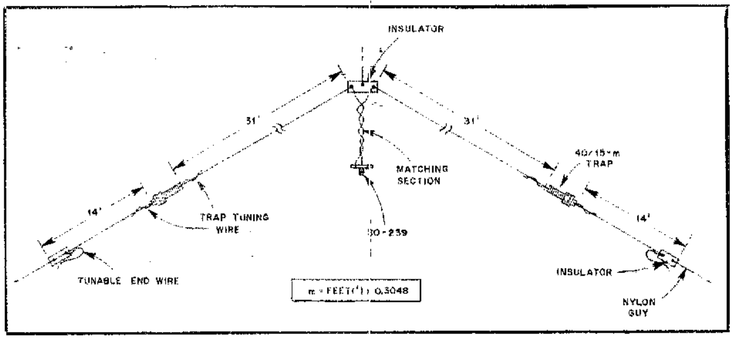


Fig. 5 — Dimensions of an 80/40/15 and 10-meter antenna using one pair of 40/15-meter traps. The center insulator can be made from ½-inch PVC pipe, and the SO-239 housing from 1-inch PVC slip caps. The 14-foot dimension includes the length of wire folded back for tuning. See text for details.

The 10-meter resonance was a pleasant surprise, and exists because the entire antenna is resonant as five half-wavelengths, capacitively loaded by the traps. The 1½-λ and 2½-λ configurations on 15 and 10 meters, respectively, are not a good match to 50-Ω coaxial cable; a short matching section is used. This is made from the no. 14 SIS wire used in the traps. Two 6-foot pieces of wire lightly twisted together makes a ¼-λ, 10-meter transformer (about 130-Ω) that also provides a good match on 15 meters. The matching section is so short that it doesn't affect the 40- and 80-meter bands significantly. The SWR curves of this antenna are shown in Fig. 6.

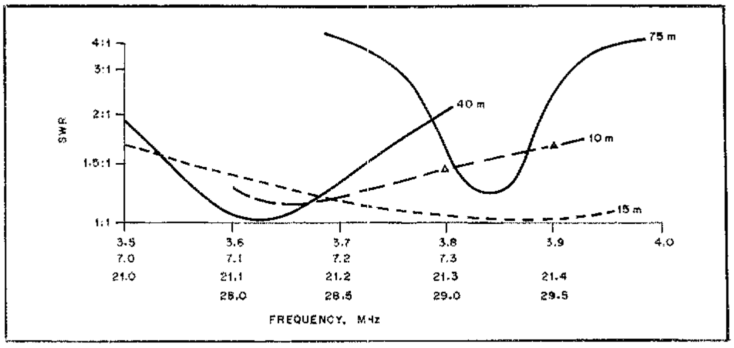


Fig. 6 — SWR curves for the antenna in Fig. 5.

Tuning

Here is a tuning trick that can be used with any trap antenna, not just the ones shown here. An insulated wire is passed through or around a trap and capacitively coupled to the antenna wire on either side of the trap (Fig. 7) This effectively places a differential pair of capacitors in parallel with the trap capacitor. The equivalent circuit is shown in Fig. 8. When the tuning wire protrudes equal amounts from both sides of the trap, additional capacitance is at a maximum. In this case, the trap (and antenna) frequency will be brought down to the lowest possible value. Sliding the tuning wire to either side causes the differential pair to decrease in series value, thereby raising the trap (and antenna) resonant frequency. The tuning wire can therefore be slid just far enough to bring the antenna up to the desired resonant frequency.

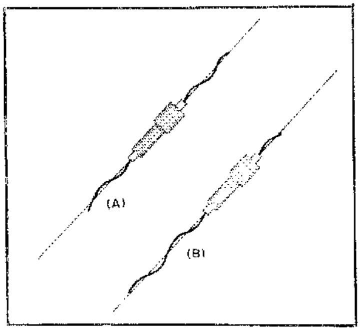


Fig. 7 — The trap-tuning element is a 3-foot length of no. 14, XLP-insulated wire. It is fed through the trap and loosely wrapped around the antenna wire on either side. At A, the capacitance is maximum and the resonant frequency will be lowest. At B, the parallel capacitance has been reduced and the resonant frequency will be higher than in A. A 3-foot wire can tune the antenna over 500 kHz on 40 meters. See text.

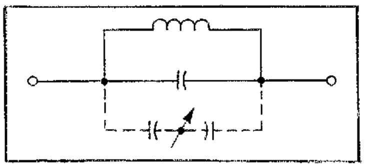


Fig. 8 — Electrical equivalent of a trap and tuning wire. The tuning wire forms a differential capacitor (dashed lines), which changes in series value as the wire is slid through the trap.

When the 40/15-meter traps are tuned in this manner, 40 meters is affected the most, with 15 meters affected to a lesser degree. This works out well, since 15-meter resonance is usually quite broad and should not need much adjustment. Although traps should normally be constructed to resonate in the middle of the band, they should resonate at the top of the band when tuning wires are to be used. They will bring resonance back down into the center of the band.

To lower the resonant frequency on 10 and 80 meters, another simple tuning scheme is used. At the ends of the antenna, some of the wire is folded back along itself and then out through a knot at the end (Fig. 9). To lower the resonant frequency on 10 and 80 meters, some of this wire is pulled out and attached to the nylon guy rope. If the antenna is set up as an inverted V, this adjustment can be made without loosening the guys.

Other Frequency Combinations

Fig. 10 shows a 10- and 15-meter trap combined with a loading coil to shorten overall antenna length. The long coil is a single layer with a self-resonant frequency of approximately 29 MHz; it also serves as the loading coil. The smaller, second-layer coil forms a 21-MHz trap. The outer portion of the antenna beyond the traps can be cut for any band lower in frequency than 15 meters (except 20 meters). Fig. 11 gives dimensions for a 40/15/10-meter antenna.

This antenna can also be tuned with insulated wires running through the traps. Two wires are used, one into each end of the trap (Fig. 10). The 15-meter tuning is quite critical. A ½-inch shift in the tuning wire position moves the resonant frequency across the entire band!

Last-Minute Hints

When constructing traps, the number of coil turns may need to be adjusted for your particular wire and layout. Typical variations in the insulation thickness are enough to change the number of turns required. A GDO (grid-dip oscillator) is essential when building traps, but don't try to measure the frequency of a trap with any wire connected to it or you will get an erroneous reading. Also, a trap cannot be dipped when it is in an antenna, for too many other resonances will appear. Once a trap is connected into an antenna, all measurements must be done from the feed point. When your traps are completed, wrapping them with electrical tape or dipping them in liquid silicone rubber to secure the turns in place is a good idea. This covering will also help protect the XLP insulation from the damaging effects of ultraviolet radiation.

Many features of the traps and antennas described here have been patented or are patent pending. Amateurs are welcome to build these for their own use, but manufacturers are cautioned that all patent rights will be strictly enforced.

Notes

[1]m = ft × 0.3048; mm = in. × 25.4.

[2]This wire is available from the local wire distributors and the Barker and Williamson Co., 10 Canal St., Bristol, PA 19007.

[3]A. Doty and A. Macnee, "Introducing the INCONS," *QST*, Feb. 1979, pp. 11-14.

[4]G. O'Neil, "Trapping the Mysteries of Trapped Antennas," *Ham Radio*, Oct. 1981, pp. 10-16.

[5]R. Johns, "Coaxial Cable Antenna Traps," *QST*, May 1981, pp. 15-17.

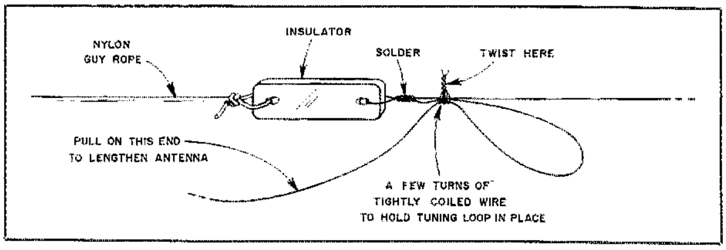


Fig. 9 — Method for tuning 80 and 10 meters on the four-band antenna in Fig. 5. Resonant frequency is lowered by pulling out the free end of the loop.

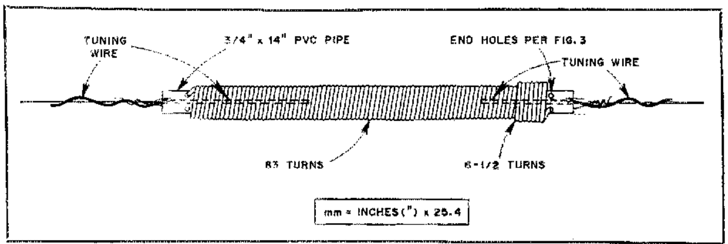


Fig. 10 — Dimensions of the 10/15-meter trap and tuning wires. The trap is wound in the same manner as the 40/15-meter version described in the text. Since this is only a two-layer coil, the end of the first layer is passed completely through the form and laid back over the windings. The second layer is then wound over this wire.

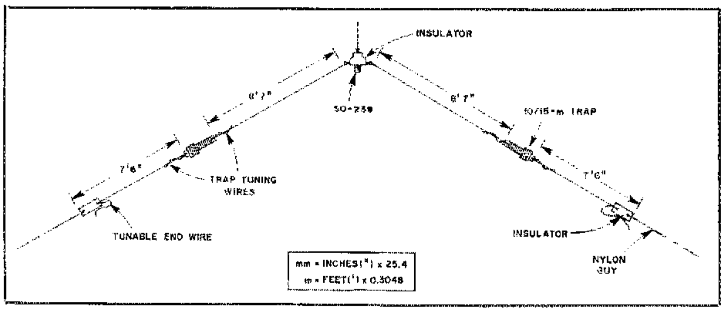


Fig. 11 — A 40/15- and 10-meter antenna using the traps shown in Fig. 10. Because of the loading coils incorporated in the traps, this antenna is approximately half the size of a standard 40-meter dipole.

New Products

AVATAR MAGNETICS "GORILLA HOOKS" TOWER CLIMBING ACCESSORY

☐ You're 100 feet in the air. It's been a long, hard day installing your new 6-element 20-meter "death ray". Sure, it was a lot of work and you're exhausted. Descending the tower with the setting sun, your mind wanders and you grin in anticipation of those pile ups you're going to breeze through. Suddenly, halfway down the tower, you lose your footing. You scream before realizing that your Gorilla Hooks have saved your life. Instead of breaking your back in a 50-foot fall, you bruise your knee after falling less than 1 foot. After your heart stops beating like a race-car engine at top speed, you finish your descent in safety and go on to flex your new-found muscles on 14.001 MHz.

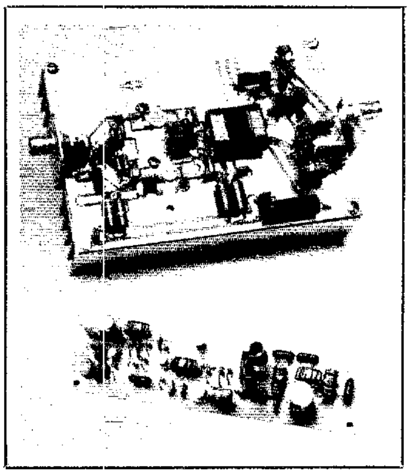
Avatar Magnetics handles a complete line of tower-climbing belts and accessories. The latest addition is the "Gorilla Hook." These hooks snap on to the D-rings of a climbing belt (the same rings the safety strap attaches to). The idea is to alternately attach the Gorilla Hooks to the next-higher tower rung as you climb. The climber is secured to the tower at all times when ascending or descending. In case of a slip, he or she cannot fall more than the Gorilla Hook's cable length.

Gorilla Hooks are made by the Klein Tool Co., a respected manufacturer of high-quality climbing belts. The hooks themselves are made of forged steel and are attached to belt snaps by a flexible steel cable. The belt snaps must be pressed on both sides to open, preventing them from accidentally unhooking from the belt. Avatar rates the Gorilla Hooks at 4000-lb overall strength.

Although we didn't have the opportunity to hang a 4000-lb gorilla from the hooks, they are easily capable of supporting the weight of a fallen climber. Workmanship is excellent, and a pair of these would be a welcome addition to a tower-climber's accessory collection.

Gorilla Hooks cost $73 per pair (postpaid to all U.S. ZIP codes), and are available from Avatar Magnetics Co. (attn: W9JVF), 1147 N. Emerson, Indianapolis, IN 46219. — *Mark Wilson, AA2Z*

A VXO CW Rig for 30 Meters



Does your present transmitter lack 30-meter provisions? Have you wanted to try this interesting amateur band? Here are the details for building your own solid-state transmitter — QRP or 60 watts.

By Doug DeMaw,* W1FB

Have you been reluctant to spend money on a new transceiver or transmitter, even though your present rig does not permit you to work 10.1 MHz? Frugality in these troubled economic times has made many of us a trifle "slow on the draw" when it comes to investing in hobby types of items. But, we can build a cw transmitter inexpensively and quickly if we wish to enjoy the interesting characteristics of the new 30-meter band. A general-coverage receiver can be used to receive on 10.1 MHz, or we might consider building a simple 30-meter converter for use with an existing ham-band-only receiver. This article describes a transmitter and includes a suggested design for a little receiving converter.

What's 30 Meters Like?

We examined the 30-meter band in general terms last month in *QST* while discussing various simple antennas for that band.[1] But briefly, it is a spectacular crossbreed of the 40- and 20-meter bands in terms of propagation. I have seldom found 30 meters closed because of poor propagation conditions. QRN is lower than on 40 meters, but slightly more prevalent than on 20 meters. Occupancy is very low, owing to the lack of contesting and "award chasing." Therefore, QRM is seldom a problem. I find 30 meters the best of the hf bands for keeping schedules beyond a 300-mile distance. Low power seems to do as well as the legal 250-W dc input limit. Many amateurs are found on the band with QRP rigs, and I have had no trouble copying stations with 1- or 2-W signals. One fellow I worked was using a modified Heath HW-7 QRP transceiver, and he was running 0.5 W. Despite the 300-mile distance, his signal was S 8, and he was using a dipole in his attic! Be sure to remember the band limits and authorized U.S. segments: 10.100 to 10.109 MHz and 10.115 to 10.150 MHz. *Do not operate from 10.109 to 10.115 MHz!*

The Transmitter Exciter Module

If you're a QRP enthusiast, it is likely that the circuit of Fig. 1 will appeal to you. It is simple and can be built in an evening. Pc boards and a parts kit are available.[2] Power output will be on the order of 1-2 W, depending on the characteristics of the transistors used from Q1 through Q3. If you are interested in having more available output power, the power-FET amplifier described earlier in *QST* can be connected to the output of this exciter/transmitter.[3] If that is done, the FETs can be changed from the MRF138s specified in the article to Motorola MRF171s, which are more readily available than the former ones. The '171 is rated at +65 V maximum, V_{DSS} and an I_D of 4.5 A continuous. Maximum power output is rated at 45 W per device from 2.0 to 200 MHz. This transistor is directly interchangeable with the MRF138.

A VXO is used in the circuit of Fig. 1. This frequency element provides excellent stability and ample frequency swing for the 30-meter band. Q1 functions as a Colpitts oscillator with feedback from source to gate. The crystal is "rubbered" (or pulled) by means of L1 and C1. A 10-kHz swing should be easy to obtain when using a quality AT-cut plated crystal in an HC-6/U style of holder. Surplus FT-243 crystals are not recommended because they may be sluggish and will not provide ample frequency shift when C1 is tuned through its range. Output for Q2 is taken from the source of Q1.

Q2, the buffer amplifier, operates Class A and is broadbanded. It has a gain of approximately 12 dB and helps to isolate the final amplifier, Q3, from the oscillator. This eliminates potential problems with chirp on the cw note. Also, Q2 gives the signal sufficient boost to drive the power FET, Q3. A broadband transformer, T1, provides the coupling from the collector of Q2 to the gate of Q3.

R1, the emitter-bias resistor of Q2, can be changed in value to control the output level of the exciter. If the little transmitter will be used as an exciter for the 60-W power-FET amplifier mentioned earlier, R1 will need to have increased resistance to limit the Q3 output to approximately 0.5 W. If not, the exciter will drive the power amplifier beyond the safe limits. You may want to add a 1-kΩ potentiometer between the low end of R1 and ground for use as a drive control.

A Siliconix VN67AF power FET is used at Q3. I chose this device because it is inexpensive and requires very little excitation to develop the desired power output. It is biased for Class B linear service, with a forward gate voltage of approximately 1.2. Class A operation, and greater linearity, can be had by changing the value of R2 to increase the gate bias to 3 V. This will lower the drive requirements from Q2. However, the idling current of Q3 will be greater with increased gate voltage.

T2 is also a broadband transformer. It transforms the Q3 drain impedance to that of the harmonic filter, FL1 (50 ohms). A simple half-wave filter ensures spectral purity, with all spurious responses being −40 dB or better, as required by FCC regulations. When driving a successive

[1]Notes appear on page 34.
*Contributing Editor, P.O. Box 250, Luther, MI 49656

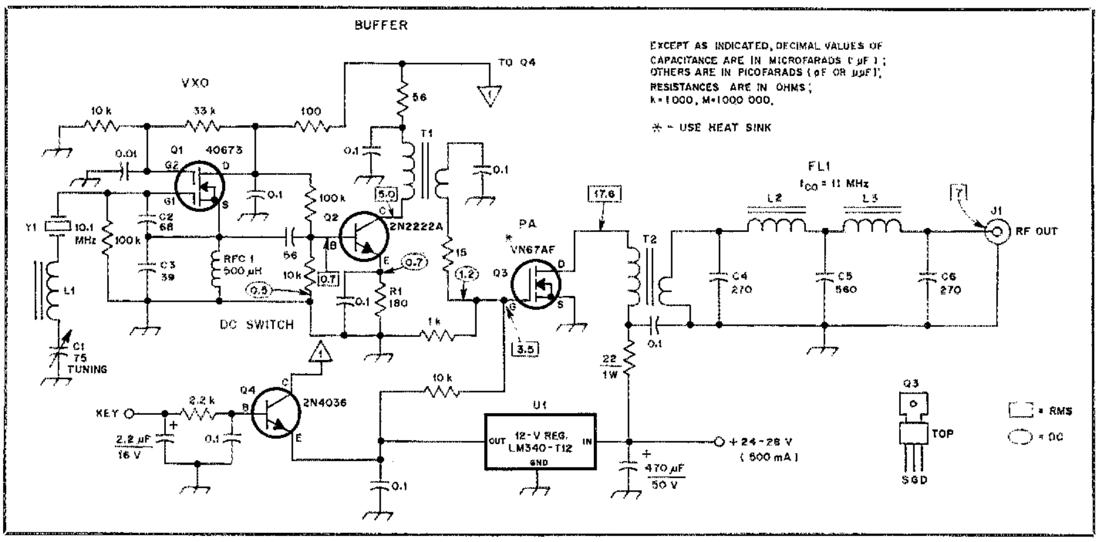


Fig. 1 — Schematic diagram of the 30-meter QRP transmitter/exciter. Fixed-value capacitors are disc-ceramic unless noted otherwise. Polarized capacitors are tantalum or electrolytic. Resistors are carbon-composition types, ¼-W, unless indicated differently.

C1 — Miniature air variable, panel mounted.
C2-C6, incl. — Silver mica or polystyrene.
J1 — Coaxial female connector of builder's choice.
L1 — 12-µH inductor. 50 turns no. 28 enam. wire on T50-2 toroid core.
L2, L3 — 0.72-µH inductor. 13 turns no. 24 enam. wire on T50-6 toroid core.
R1, R2 — See text.
T1 — Broadband transformer. Primary has 12 turns no. 26 enam. wire on an FT50-43 ferrite toroid. The secondary has six turns of the same type of wire.
T2 — Broadband transformer. Primary contains 15 turns no. 24 enam. wire on an FT50-43 ferrite toroid. Secondary has six turns of the same type of wire.
Y1 — Fundamental crystal, type AT in HC-6/U holder. Order for center of desired 10-kHz tuning range, 30-pF load capacitance.

power amplifier with the circuit of Fig. 1, clean excitation energy will be beneficial in assuring clean output from the power amplifier. It is always a good objective to supply the purest of driving power to any stage of a transmitter.

A 24-V power supply was chosen to permit Q3 to develop the desired output power easily. Although most power FETs will operate at low voltages, such as 12 V, they saturate readily at voltages below the rated potential. That is, a 28-V FET will deliver only a fraction of its rated power at 12 V. A VN46AF (available at Radio Shack stores) can be used at Q3 if we reduce the supply voltage to 18 or less. Unfortunately, the VN67AF contains a built-in Zener diode from gate to source, which can be shorted if the peak driving voltage to the gate is excessive on the negative half cycle. When this happens, the FET junction will become shorted, and that will be the end of our transistor! The safe gate swing for a VN67AF will not be exceeded in the circuit of Fig. 1.

A three-terminal regulator, U1, provides the operating voltage (regulated) for Q1, Q2, Q4 and the gate of Q3. This section can be eliminated if separate power supplies (12 and 24 V) are available, or if we choose to operate Q3 at 12 V. This reduced drain-source voltage at Q3 will require a different resistance value for R2 if we are to have the desired +1 to +3 V of forward gate voltage. Similarly, the turns ratio of T2 will need to be modified to accommodate the new drain impedance. The approximate impedance is obtained from $V_{DD}^2/2P_o$, where V is in volts and P is in watts.

Keying of the exciter is accomplished by means of Q4, which is a pnp dc switch. When the base is keyed (grounded) the transistor conducts, thereby allowing the +12 V to reach Q1 and Q2. This turns on the transmitter during key-down periods. Since Q3 operates in a linear manner, the keying is not hard (clicky) as it might be when keying the drive to a Class C amplifier. Two capacitors and a 2.2-kΩ resistor form a shaping network in the base lead of Q4. This aids in rounding off the sharp trailing edge of the cw waveform. Increasing the value of the 2.2-µF input capacitor at Q4 will further "soften" the cw note. If break-in delay is desired for the keying circuit, the module described in an earlier *QST* article can be used.[4] It is designed to prevent "hot switching" the final amplifier, which in turn protects the PA stage from no-load (momentary) damage. By avoiding hot switching of the PA (no antenna attached as the changeover relay cycles from receive to transmit), we also prevent unwanted spurious "blurps" from appearing on the air. Some commercial rigs are offenders in this regard. The referenced QSK module has switching capability for muting a receiver. It also features adjustable time constant for con-

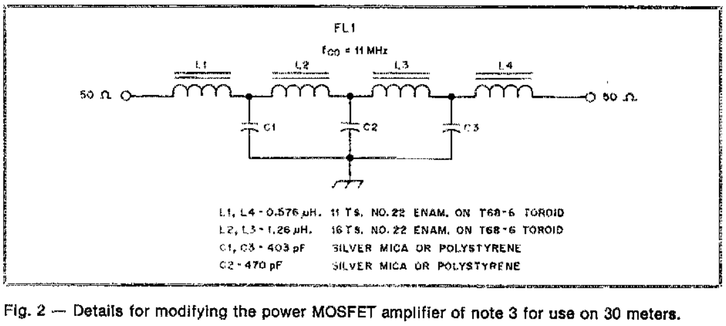


Fig. 2 — Details for modifying the power MOSFET amplifier of note 3 for use on 30 meters.

trolling the relay drop-out time. It can provide full QSK if a reed or mercury-wetted relay is used in place of the relay specified.

Interface to the Power-FET Amplifier

The exciter and the amplifier are designed for connection to 50-Ω loads. Therefore, we can use a short length of 50-Ω coaxial line to join the modules (output of the exciter to the amplifier input). A 28-V, 5-A regulated dc supply will handle the entire transmitter nicely. Many such power supplies are available from surplus houses at reasonable cost. Be sure to check the flyers and catalogs if you don't own a 24- or 28-V power supply.

We must permit only enough drive from the exciter to develop 60 or fewer watts of amplifier output power into a 50-Ω load. This will be on the order of 0.5 to 0.75 W typically. The +24 or +28 V line to the 60-W amplifier can be left operational at all times. This will avoid the need to include the amplifier in the T-R loop. The antenna lead will need to be switched, however.

Minor changes are necessary in the amplifier discussed in note 3. The only modification for 30-meter use involves changing the constants in the output filter, FL1 (Fig. 4, page 27, March 1983 *QST*). Fig. 2 of this article contains the coil and capacitor data for 30 meters. C1 and C3 may be formed by placing a 390- and a 12-pF capacitor in parallel at each spot in the filter. The powdered-iron toroid cores are available by mail from Amidon Associates, Palomar Engineers and RadioKit (see *QST* ads). The entire amplifier kit or a composite unit (also a kit) containing the amplifier and the exciter of Fig. 1 can be obtained from the supplier in note 2.

Receiving on 30 Meters

We mentioned earlier a suggested design

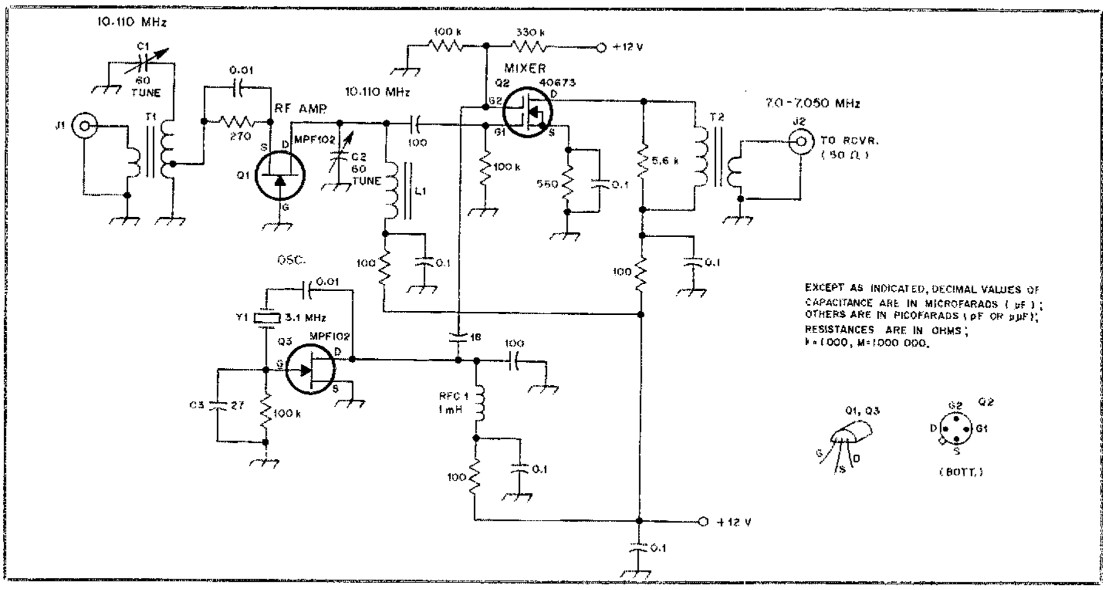


Fig. 3 — Suggested circuit for a simple 30-meter converter. The i-f is 7.000 to 7.050 MHz. C1 and C2 are mica compression trimmers or miniature ceramic or polystyrene trimmers. See text for C3. J1 and J2 are phono jacks. L1 is a 6-μH inductor (38 turns no. 30 enam. wire on a T50-6 toroid core). T1 has the same winding on the secondary as L1. Tap the source of Q1 eight turns above the grounded end. The primary has three turns of no. 30 wire. T2 has 15 primary turns of no. 24 enam. wire on an FT50-43 toroid. The secondary contains two turns of no. 24 wire.

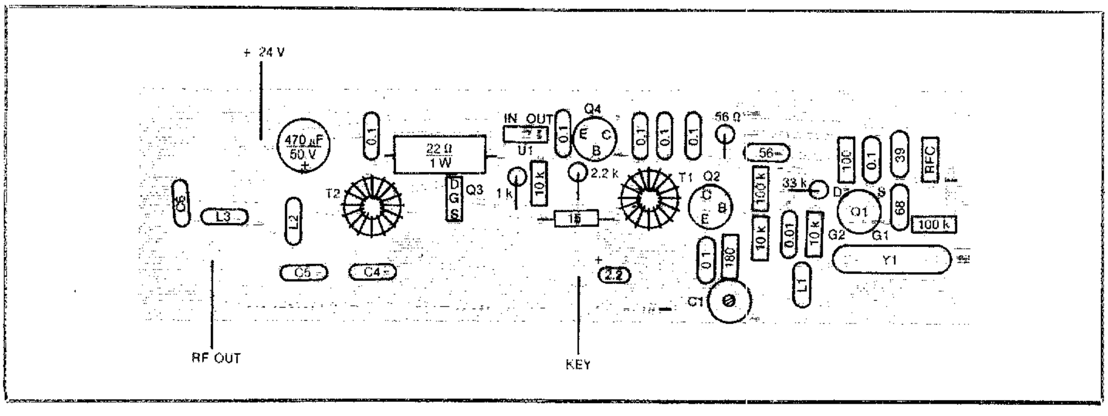


Fig. 4 — Parts-placement guide for the 30-meter exciter, as viewed from the component side of the pc board. See Hints and Kinks, this issue, for the etching-board pattern.

for a 30-meter converter. There are many designs to be found in the amateur literature, most of which could be modified easily to 30 meters. But, in the event you are not sure of what is required to perform proper alterations for a new operating frequency, the circuit of Fig. 3 should be entirely suitable. It has been proven during use on many of our hf bands.

The converter can be fed to a receiver that tunes the 40-meter band. A 3.1-MHz crystal at Y1 of Fig. 3 enables us to tune 10.100 MHz at 7.000 MHz on the receiver. When we tune our receiver to 7.050 MHz, we will be listening to the high end of 30 meters (10.150 MHz).

Q1 provides some 10 dB of gain as a common-gate rf amplifier. C1 and C2 are adjusted for a peak response at the center of the 30-meter band. Or, if you are interested mainly in the low segment of the band, you may wish to peak the trimmers at 10.105 MHz.

The mixer has a broadband output circuit to permit coverage at the i-f without need to retune the mixer output as the station receiver is tuned from 7.0 to 7.050 MHz. A 5.6-kΩ resistor is used across the primary of T2 to aid in IMD reduction and to establish a fixed-value impedance for the Q2 drain circuit.

A Pierce oscillator is used at Q3. A load capacitance of 30 pF is suggested when ordering Y1. If you use a surplus crystal and find it a bit sluggish with regard to oscillation, try increasing the capacitance of C3 slightly. This is a feedback capacitor.

Construction Data

Fig. 4 contains a parts-placement drawing of the 30-meter exciter. A scale template for the pc board appears in Hints and Kinks in this issue. Layout and construction information for the 60-W power-FET amplifier was provided in March 1983 *QST*. There is no layout data for the suggested converter. You may lay it out on perforated board, or try your hand at pc-board layout and etching. Check with the supplier in note 2 for the availability of converter boards and kits.

Closing Comments

Adventure is awaiting you on 30 meters. That, plus the thrill of going on that new band with a homemade station, will make these projects worth your while! Perhaps you will choose to start operation at the QRP level. It is logical that any newcomer would prefer to assess the band for interest and advantages before getting tooled up for the 60-W power level. That part of the project can come later if you find the band to your liking. Whatever the situation, good luck on 10.1 MHz!

Notes

[1]D. DeMaw, "Building and Using 30-Meter Antennas," Oct. 1983 *QST*.
[2]Circuit Board Specialists, P.O. Box 969, Pueblo, CO 81002.
[3]D. DeMaw, "Go Class B or C with Power MOSFETs," March 1983 *QST*.
[4]D. DeMaw, "20-Meter Hamcation Special," Nov. 1982 *QST*.

Strays

THANKSGIVING — A SPECIAL EVENT

□ It all started in 1980 when a member of the Whitman (Massachusetts) ARC approached me and several other members of the club and asked a simple question: "With all the early-American historical sites around us, why haven't we done anything to bring this to the attention of the rest of the world?" None of us had ever operated a special-event station. The closest any of us had come to that was Field Day, and what we learned from that over the years really paid off.

With this thought in mind, we contacted the administrators of Plimoth Plantation, a nonprofit living history museum in Plymouth, Massachusetts, and related our ideas to them. Their response was overwhelming, and they welcomed us with open arms. It turned out we could operate from inside the museum library. Considering the operating date was set for Thanksgiving Day and the familiar adverse weather conditions of November, it seemed like the ideal place to operate.

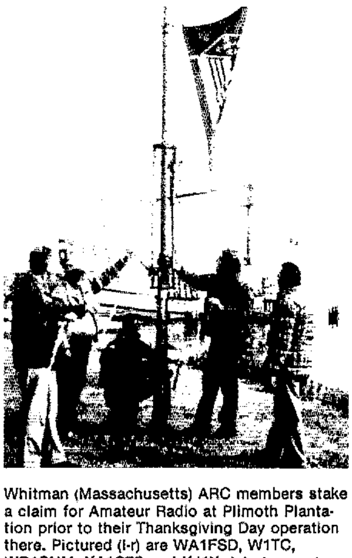
We decided to use 15 meters for distance and 40 meters for regional coverage. We had a dipole for 40 meters, a 3-element Yagi for 15 meters and an old (but in top shape) military, crank-up 50-foot field tower. The weekend before the operation, five of us went to the Plantation to set up the antennas. This is when our Field Day experience particularly paid off, as we had to brave 40-mile-per-hour winds and a downpour to get everything up and properly guyed.

On Thanksgiving Day, we worked just about every state in the U.S., all through Europe and parts of the Middle East and Africa. A special QSO occurred between a ham in Plymouth, England, and our Plantation Station.

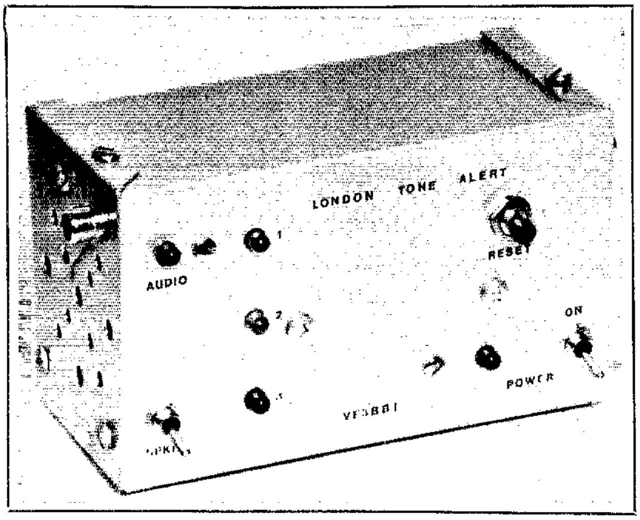
The second year of operation, conditions were poor. We did manage to work as many stations as the year before, but our DX total was only a dozen or so, with only a few in England.

This year, with propagation rapidly declining, we are going to expand our operations to include 20 meters and limited 2-meter operations (see Special Events, this issue, for more details).

This year's event will also be supported by members of the Plymouth (Devon, England) Radio Club, operating G3PRC from a site overlooking Plymouth Sound, from where the Pilgrims began their journey aboard the *Mayflower* in 1620. We are looking to break a 1000-QSO total in one day this year, so put on your rigs and be part of the holiday celebration! — *Jim Russell, WB1CNM, Brockton, Massachusetts*



Whitman (Massachusetts) ARC members stake a claim for Amateur Radio at Plimoth Plantation prior to their Thanksgiving Day operation there. Pictured (l-r) are WA1FSD, W1TC, WB1CNM, KA1CZS and KJ1X. *(photo courtesy WB1CNM)*



The London Tone Alert

Build this simple control decoder and never miss the ARES action!

By T. C. Tanner,* VE3BBI

The London, Ontario ARES needed a simple, easily constructed Touch-Tone® activated system to alert its members. One of the most important (and difficult) facets of any civil preparedness (CP) plan is activation. An ARES group needs a way of contacting its members independent of telephone lines (which will probably be out during an emergency). A unit that would allow *all* members to silently monitor the CP repeater until the Emergency Coordinator (EC) generated the "activate" command has been sought for some time.

With this in mind, Norm Ross, VE3ETJ, developed a prototype. This design contained a dual-tone, multi-frequency (DTMF) decoder and logic to drive three control relays. The system about to be described is similar to the prototype, but it uses silicon-controlled rectifiers (SCR) rather than relays. This drastically reduces the size and current demands of the device. The project is packaged in a standard 2¼ × 3 × 5-inch utility box.[1]

Design Objectives

Four basic objectives were considered in the original design:

1) three tones necessary for access, with false alert suppression;

2) automatic shutoff of the siren, and a means to disable it at any time;

3) low current drain, so the device could be left on indefinitely;

4) relatively simple and inexpensive construction, so *all* ARES members could own one.

In addition, several "bells and whistles" were incorporated to monitor proper operation and aid in the setup of the alert:

1) visual indication (LED), to let the operator know when an alert has been received;

2) LED indicator, to set the receiver audio to the correct level;

3) lights, to indicate when correct tones are received.

Circuit Description

The circuitry used in this device is simple. When a valid tone is received, it is detected by two 567 PLLs, producing a logic "low" at their outputs (pin 8). When this level is applied to one section of the 7402, a positive pulse appears at the output. This signal is transferred to the second circuit board for processing.

U4 responds to 1209 Hz, which is common to the left column on a tone pad; similarly, U1-U3 correspond to *, 1 and 7, which comprise the code selected by London-area hams. Any other vertical column of the pad could be used as well, requiring only frequency adjustments in the decoding circuitry.

Accidental triggering of the circuit by voice frequencies is prevented by a three-second time constant developed by the action of the 4.7-kΩ resistor and 220-μF capacitor between the emitter of Q1 and ground. Because of this, the first digit in the code must be held for approximately three seconds, until the capacitor is sufficiently charged to trigger Q2. This SCR lights DS3, providing a positive voltage at the anode of Q3. An SCR behaves like a latching relay. Once triggered, the device will remain on (latched), as long as current is being drawn through it. The second digit is decoded next, providing a positive pulse at the gate of Q3. This SCR lights DS4 and results in a positive voltage at the anode of Q4.

When the third number is received, Q4 turns on and lights DS5. This produces a positive voltage at the base of Q5. This transistor conducts, grounding the negative side of the audio alert circuit (U7-U8) to sound the alarm.

A 555 timer is employed to shut off the alarm. The timing cycle is triggered by acquisition of the first valid tone. After ap-

[1]Notes appear on page 37.

*548 Upper Queen St., London, ON N6C 3T9, Canada

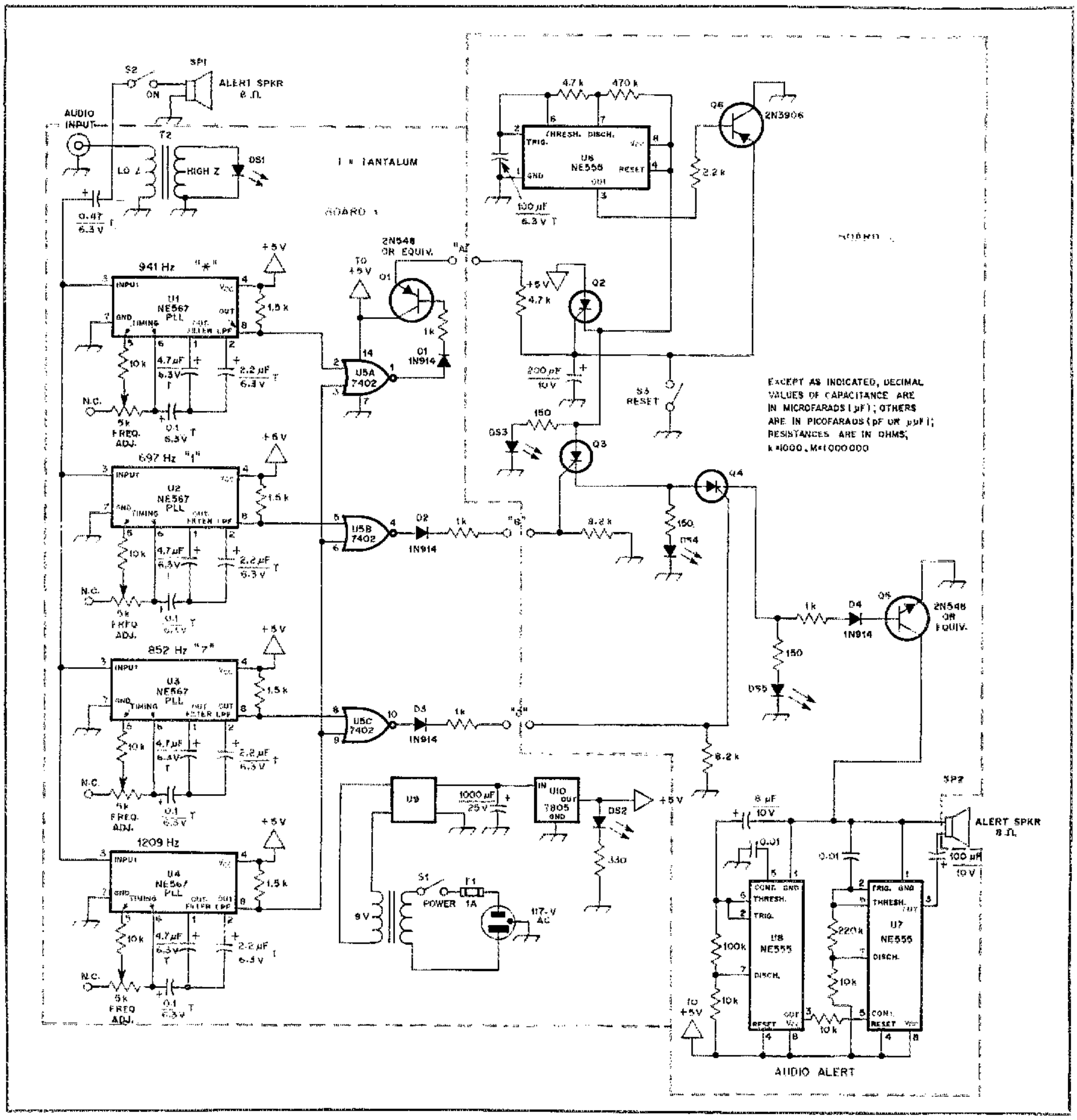


Fig. 1 — Schematic diagram of the tone alert. All resistors are ¼-W, 5% carbon types.

DS1, DS2, DS5 — Large LED (red).
DS3 — Large LED (green).
DS4 — Large LED (yellow).
Q2, Q3, Q4 — Silicon-controlled rectifier, 2N5060, 2N5061 or equiv.
S1, S2 — Spst miniature toggle switch.
S3 — Normally open push-button switch (Radio Shack 276-1811 or equiv.).
SP1, SP2 — 8-Ω speaker, 2- or 2½-inch-diameter.
T1 — Power transformer, 117-V primary; 9-V, 300-mA secondary.
T2 — Transistor-type output transformer (Radio Shack 273-1380 or equiv.).
U9 — Full-wave bridge rectifier, 50-V, 1 A.

proximately 45 seconds, the timer supplies a negative voltage to Q6. This effectively grounds the gate of Q2, shutting down most of the circuit. A small glow still remains in DS3, which brightens and dims as the timer cycles. With the given values, the dim portion of the cycle lasts for about one second, so DS3 remains lit. This serves as an indication that an alert has been received. The circuit is reset, however, just as if the RESET button on the front panel had been pushed.

A reverse-connected transistor-output transformer is across the audio input. An LED across the high-impedance side of this transformer serves as a level indicator. The receiver volume control can now be set for a level sufficient to activate the decoder;

the LED will flash when an ample voice level is received. While this is an optional part of the unit, it is highly recommended to ensure the volume is adjusted for proper operation.

When this circuit is plugged into the external speaker jack of a receiver, it will normally disable the receiver speaker. To alleviate this problem, a small speaker is provided to allow receiver monitoring. A bypass switch, S2, is included to disable the speaker for silent monitoring. The alert is still operational in this mode, even with the speaker off. A power switch and pilot LED are also provided on the front panel, to tell the operator when the unit is on.

A regulated 5-V supply powers the entire system. The bridge rectifier, filter capacitor and voltage regulator are on pc board no. 1. Four NiCd cells can also provide 5 V, if battery operation is desired. An alternative is to apply 12-V dc to the input of the regulator. One might consider a rear-panel jack and switch for this. Total current drain from the regulator is 55 mA.

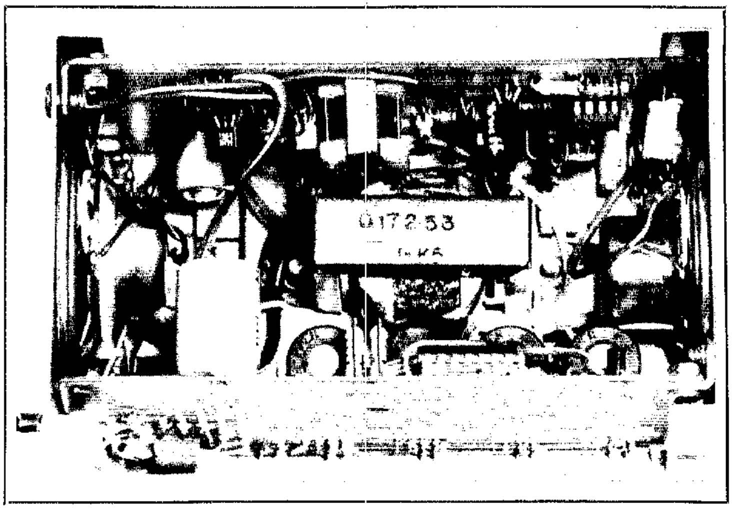


Fig. 2 — Internal view, taken from the rear of the unit.

Construction

Use ¼-W resistors,and capacitors with 6- or 10-V ratings. Use a 9-V, 300-mA transformer for power. I was able to locate almost everything at a local surplus store.

The bulk of components are contained on three pc boards. Q1, U1-U5 and their associated decoding circuitry mount on board no. 1. The LEDs, DS1-DS5 and transformer T1 are contained on board no. 3. All remaining components are housed on board no. 2. It may be necessary to adjust the mounting holes to accommodate the particular components you have. Also, if you are unable to locate a utility box of the correct size, you will have to alter the size or location of the mounting holes. This is not a "piece of cake," so give it some thought before proceeding. Don't forget that resistors can be mounted vertically to save space. The only critical components are the three tantalum capacitors in each decoder circuit.

Mount the LEDs on the Z-shaped circuit board. Using the same template, drill holes in the box to allow the LEDs to protrude up to the shoulder. No further insulation should be necessary.

Circuit boards no. 1 and no. 2 mount on each side of the utility box with spade bolts or threaded brackets. Allow enough space at each end to slip the speakers into place. Holes in both ends of the box serve as speaker grills. Use sockets for all ICs, and don't plug in the chips until the wiring has been checked. Also, note that all components are mounted in the larger section of the utility box, so it can be opened without any dangling wires.

Alignment

Alignment with a frequency counter is a snap. Begin by attaching a frequency counter to pin 6 of a 567. Adjust the corresponding 5-kΩ potentiometer until the counter reads the desired frequency to be decoded by that device. Tune each of the other decoders in the same fashion, until all four are aligned. This is all the tuning that is necessary. The device is now ready for testing.

Testing

To test the unit, it is necessary to supply amplified tones from a tone pad to the input. This can be accomplished by connecting the device to the external speaker of a 2-meter rig and sending the tones using another. I fed a Yaesu Touch-Tone microphone through a tiny audio amplifier using an LM383 chip plugged directly into the unit. Press the * button for three or four seconds, until the green LED, DS3, lights. If all is well, press 1 and 7 in succession. The yellow (DS4) and red (DS5) LEDs should light, and the siren should sound.

After approximately 45 seconds, the siren will shut down, and only DS3 will remain lit. The siren may be stopped at any time by pressing the RESET button.

Comments

This unit has been tested on the air from various locations, and worked like a charm every time. If necessary, certain modifications can be accomplished easily. For instance, if only one tone is needed to sound the alert, two 567s, Q3 and Q4 can be eliminated. However, it is a good idea to retain the delay caused by the R-C combination in the emitter circuit of Q1, to prevent triggering from voice transients.

The audio alert could also be simplified to eliminate the warbling note it produces. I prefer the distinctive note, which is like an ambulance siren.

A relay could be used at the output to control a heavier load for alerting a large group of people. To accomplish this, simply insert the relay and suitable power supply in place of the audio alert shown in the diagram. A 12-V relay could be used, and obtain its power from the same power supply, but ahead of the regulator.

Applications for this device are not limited to the functions described here. This is a practical control circuit with limitless possible uses.

Acknowledgments

Thanks for help in developing the London ARES tone alert go to Harold Delagran, VE3VL; Hugh Clark, VE3WM; Jack Strangleman, VE3GV; and especially to Norm Ross, VE3ETJ. I would also like to thank the owner of Microtronix Systems Ltd., who allowed me to produce the artwork, negatives and pc boards for this project. Microtronix Systems Ltd. will also supply pc boards for $20 (U.S.) or $24 (Canadian).[2]

Notes

[1]mm = in. × 25.4.

[2]Microtronix Systems Ltd., 120 Bessemer Rd., London, ON N6E 1R2, Canada. The ARRL and *QST* in no way warrant this offer.

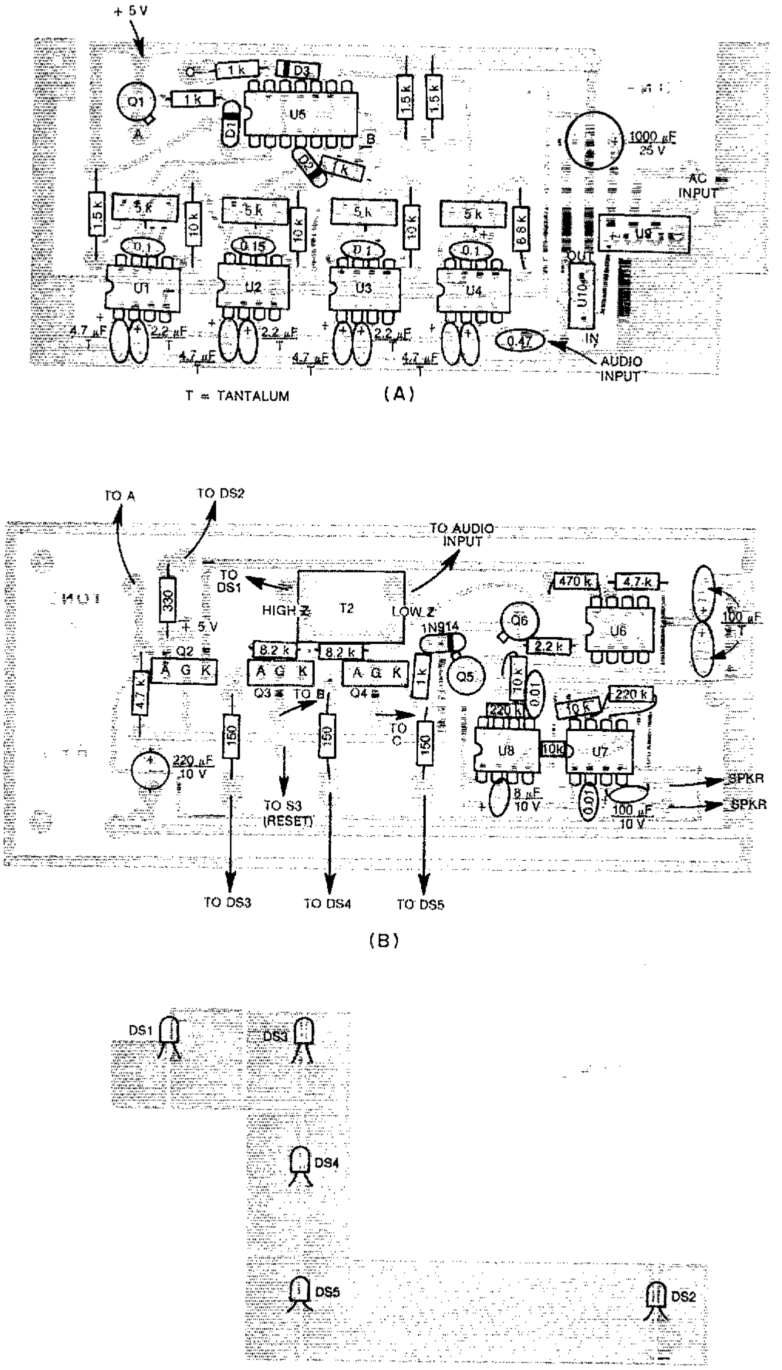
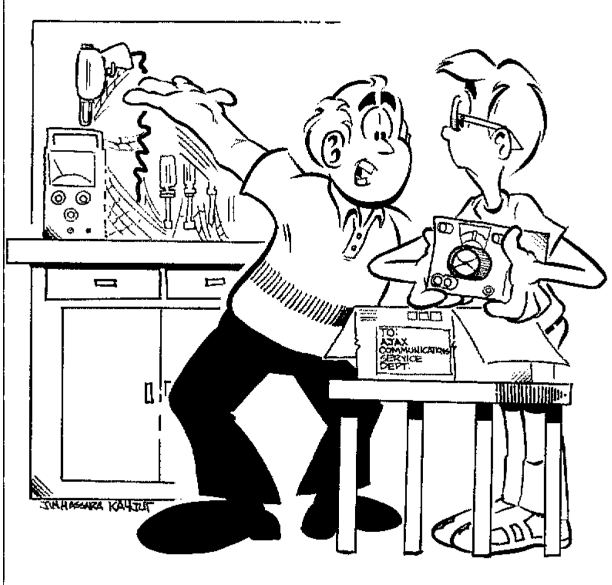


Fig. 3 — Parts-placement guide. This view is from the component side of the board. Gray areas represent unetched copper. See Hints and Kinks, this issue, for the etching patterns.



BEGINNER'S BENCH

Simple Ways to Test Your Transmitter

Is your rig inoperative? Is the rf output lower than it should be? Or, perhaps you suspect that you're radiating harmonics from your antenna? Whatever the problem, there are simple methods for checking your rig, and you won't need exotic test gear to do it!

By Doug DeMaw,* W1FB

It's easy to be confused and discouraged when your transmitter falters or refuses to deliver rf energy to the antenna. A truism, for certain. But all too many amateurs are fearful of digging into their equipment in an effort to locate and solve even the simplest of performance problems. Instead, the equipment is boxed and shipped back to the manufacturer for repair. A large percentage of the time, the amateur could service his or her transmitter and save countless dollars in shipping and repair cost, to say nothing of the inconvenience of being without the rig for weeks, or even months!

In the earlier days of Amateur Radio, it was practically unheard of to seek factory repair of one's equipment. Even the least-skilled operators had the courage to troubleshoot commercially made gear, but the unfortunate trend toward factory service seems to be ever-increasing in pace. A valuable by-product of doing our own repair work is added knowledge about circuits and how they work. Of course, the more we learn about our rigs, the more confidence we will have when the next servicing job crops up.

The intent of this article is to illustrate some simple techniques for tracking down those common ailments that befall our equipment as it ages. The test apparatus needed is neither costly nor hard to build. A common misbelief among hams these days is that one needs a laboratory and thousands of dollars worth of test gear to repair a piece of transmitting or receiving equipment. Balderdash! A volt-ohmmeter and an rf probe will serve admirably for most of the problems we may encounter.

First Things First

Learning to follow a circuit diagram is the first order of business in locating a fault. Most of today's factory-made ham radio equipment is dreadfully complex, and rightfully so if we are to enjoy the myriad operating features and conveniences that are common to the state-of-the-art transceiver or receiver. As a consequence of this circuit complexity, the schematic diagrams look bewildering and rather like a maze created by some demonic engineer. Fortunately, most of the operating manuals contain separate diagrams of the various modules in the equipment. The overall "master" diagram is even a bit much for an experienced amateur or engineer to decipher, and seldom do we find labels above the various stages to indicate what their functions might be. That, plus the countless interconnect lines that run back and forth across the drawing, can cause the strongest of heart to capitulate in favor of factory servicing! But, the individual diagrams for the various modules are generally quite easy to follow. I strongly recommend that approach when doing service work.

Most manufacturers will sell us circuit-board extenders (cables) that permit taking the modules out of the main frame while still keeping them operational. If you don't have a set for your rig, perhaps you will want to obtain one. Factory service manuals are often available for those who are willing to pay for them. They are greatly expanded versions of the basic operating manual that comes with the equipment. If you plan to keep your present rig for a period of time, you will find the service manual invaluable.

Where and How to Start

Let's consider one of the more common transmitter failures — low rf output, or no output at all. It makes no difference whether our rig is totally solid state or if it is a hybrid (tubes and semiconductors) unit; the procedure is the same when troubleshooting. We will always commence with the first stage in the transmitter chain.

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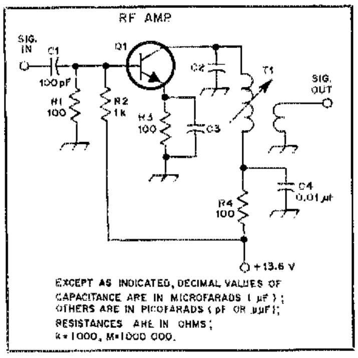

Fig. 1 — Example of a simple rf amplifier for text discussion.

These early stages are commonly referred to as "low-level stages." This means that they generate low levels of power or voltage.

1) Phase 1 calls for visual inspection of the components that are related to the stage being examined. We will inspect the resistors for discoloration or fine cracks. A magnifying glass or jeweler's loupe is practically mandatory for this part of the check. A discolored resistor usually means that excessive current has been flowing in that branch of the circuit. Discoloration or cracking is caused by overheating of the component. These defective resistors are apt to change value significantly or become open-circuited. This prevents the correct operating voltage from reaching the tube or transistor. Fig. 1 illustrates a simple rf amplifier that could have one or more defective resistors. Assume, for example, that R4 were defective. What might cause this resistor to be burnt or cracked? The culprit — excessive current flow — would have to be caused by some anomaly above the resistor, or the side of R4 opposite the 13.6-V supply line. What could cause this to happen? There are four potential sources of the difficulty: (1) Q1 could be shorted and drawing excessive current; (2) resonating capacitor C2 might be shorted. (3) decoupling capacitor C4 might be shorted internally; (4) emitter bypass capacitor C3 could also be shorted, which could permit Q1 to draw high current and become shorted. A fifth possibility would be the input blocking capacitor, C1. If it were shorted there might be too much dc voltage at the base of Q1, thereby making it draw excessive current or burn out. If this were the bad capacitor, R1 might be defective also, because of excessive current flow through it.

A good method for locating the bad capacitor (or transistor) is to check each terminal of Q1 for operating voltage after placing a new resistor at R4. Apply operating voltage only in short cycles to prevent burnout of the new resistor. If you find no voltage on the upper end of R4 (Fig. 1), or if the voltage is very low, turn off the supply voltage and remove Q1 from the pc board. Repeat the voltage check at the high side of R4. If normal operating voltage (or slightly more) appears above R4, it is safe to conclude that Q1 caused the failure. If the bias voltage at the junction of R1 and R2 is normal, the transistor is probably shorted and needs to be replaced. It is suggested that you test C3 for leakage or short-circuiting with your ohmmeter if Q1 is defective. If C3 is shorted, it will permit the emitter of Q1 to be at dc ground, thereby negating the value of R3 as an emitter-bias resistor. This will increase the current drawn by Q1.

If normal base bias is found but there is no collector voltage at Q1, perform an ohmmeter check of C2 and C4. If either is shorted, it could cause R4 to become damaged, but would not harm Q1. The most likely cause of failure (damage) at Q1 would be a shorted capacitor at C1 or C3. Also, if R2 became too low in value through aging or heat, high bias on Q1 could damage the transistor. Likewise with R3 in the emitter circuit.

I'll bet you never thought there could be so many causes for failure in so simple a circuit! That's why we must learn to analyze one stage at a time and look for all manner of causes for the malfunction. The foregoing treatment of troubleshooting a simple stage is of course applicable to audio and video amplifiers as well. The primary objective here is to find out why too much current flowed in some branch of the circuit.

2) Phase 2 will find us doing dc voltage checks with a VOM (volt-ohmmeter) or VTVM (vacuum-tube voltmeter). This test assumes that our visual inspection revealed nothing unusual about the appearance of the resistors or other associated circuit components.

Let's check the operating manual to learn whether the manufacturer was considerate enough to include a voltage and/or voltage/resistance chart for the equipment stages. Some do, and others don't. If your manual has that important information, you should use it when making the voltage checks. If not, refer to Fig. 2 for what I will call very broad ballpark voltage values for solid-state low-level stages that have a 12-14 V supply line. Radical departures from the voltage ranges indicated in Fig. 2 will suggest problem areas to look at. The nominal voltages listed in the diagram are approximately what we would expect if Q1 were drawing 10 mA during normal operation. But, your rig may be biased differently, and depending upon the power the stage must provide for the following stage, the various voltages may be different than the nominal values. The typical voltages specified in Fig. 2 should be helpful during routine testing of low-level rf or audio amplifiers.

If a voltage is too low, missing completely or too high when you check at

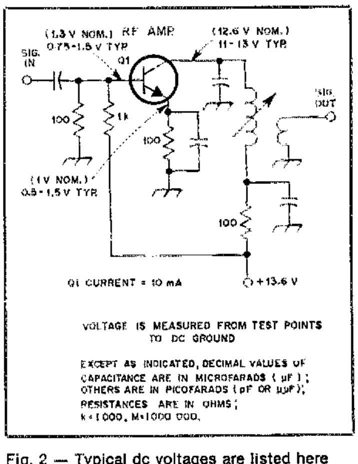

Fig. 2 — Typical dc voltages are listed here for use in locating faults in an inoperative low-level amplifier.

the base, collector and emitter of Q1, search for leaky or shorted capacitors as in Phase 1; they will cause too low a reading. If the capacitors are okay, then lift one end of each resistor from the pc board and check the resistance with your ohmmeter. Replace any resistor that is more than 30% removed from the specified value. Finally, if all components appear to be normal, replace Q1 and repeat the voltage measurements. Of course, if during your tests you found a voltage that was far too high, check for decreased resistance at R2 and R4. A high reading might also be caused if R3 of Fig. 1 had too high a resistance from the marked value.

3) Phase 3 is called "signal tracing." It may seem that we have placed the "horse after the cart," so to speak, by describing first the processes in Phases 1 and 2. After all, we did not discuss a technique for determining which part of our imaginary transmitter had the "blues." The assumption was, of course, that the failure was rather obvious, and that the affected circuit or module would be obvious to the person attempting the repair.

But, what if we have no strong thoughts about the general region of the circuit where the malfunction is taking place? Our best bet then calls for signal tracing. This can be done with an oscilloscope or an rf probe and a VTVM or FET VOM (field-effect-transistor voltmeter). The latter is perhaps the least expensive and most common ham shack method for signal tracing. An rf probe that is easy to build was described in Fig. 3 in Beginner's Bench, August 1983 *QST*. Information on this device can be found also in the measurements chapter of the ARRL *Handbook*. An rf probe is a detector that converts an rf (ac) voltage to a dc one so that the voltmeter can provide reasonably ac-

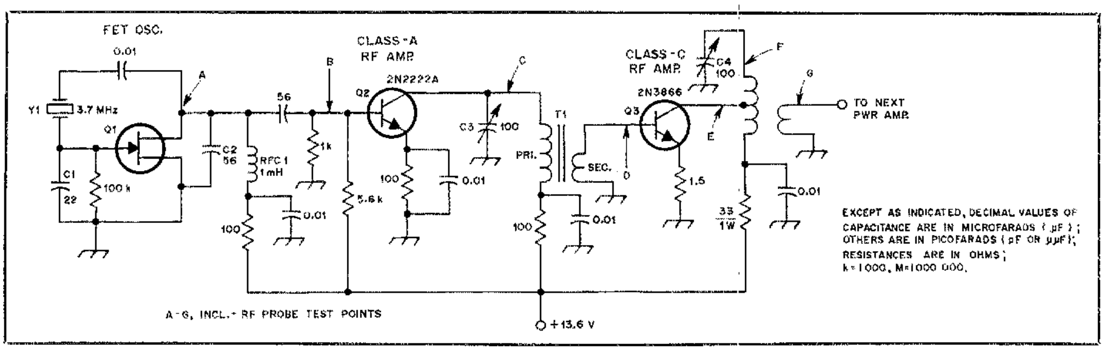

Fig. 3 — A basic transmitter rf strip showing various test points to which an rf probe can be connected during troubleshooting.

curate voltage-level readings. The ac scale of most VTVMs is not suitable for use at radio frequencies; therefore, the rf probe is required. We should remember that the accuracy of the rf-voltage measurements is dependent on the relative purity of the waveform at each test point in the circuit. Specifically, the probe is designed to read rms (root-mean-square) voltages, and if there is distortion in the waveform, it will affect the readings. In other words, our best accuracy will result when the test point contains a pure sine wave. This is seldom possible in solid-state circuits, owing to harmonic generation and distortion caused by the transistor junctions (diodes). However, we can still compare the relative voltages to determine whether a particular transmitter stage is amplifying and producing output energy. Our objective with the rf probe is to locate the stage that is inoperative or low in output. If we use a scope for this purpose, we will be able to see the waveform distortion, but our voltage readings will no longer be an rms type. Rather, they will be pk-pk (peak-to-peak) voltage measurements.

Fig. 3 shows a very basic type of transmitter circuit, such as might be found in the exciter portion of a cw transmitter. The simplicity is intentional for the purpose of easy discussion.

Let's suppose that our transmitter had ceased to deliver rf energy to the antenna. How should we pinpoint the trouble area? Well, the logical step would be to first ensure that the oscillator (Q1 of Fig. 3) was operating as it should. We'll rule out the possibility of Y1, the crystal, being defective, since it has worked satisfactorily for some time and has not been subjected to vibration or excessive rf feedback current. We will instead use the rf probe and voltmeter to check for rf energy at point A (Q1 drain). If the circuit is oscillating, we should find a few rms volts present at point A.

Next, we move to the input of the second stage, test point B of Q2. Remember that the higher the circuit impedance at the test area, the greater will be the rms voltage for a given power level. Test point A is of higher impedance than is point B, so there should be somewhat less voltage appearing at the base of Q2, since that is a lower impedance than we will find at the drain of Q1 in this particular circuit. The important thing is that we find *some* rms voltage present at B.

Next we move our probe to C, the collector of Q2. If very little rf voltage is indicated, or if there is none at all, we can conclude that Q2 or some component in that section is defective. We should find more voltage at point C than at B, since this is an amplifier stage.

If we have a normal voltage reading at C, we can then proceed to point D with our probe to learn if the last amplifier, Q3, is receiving excitation. Q3 is a higher-power amplifier than Q2, so the base impedance will be lower than that at Q2; the greater the driving power the lower the input impedance of an rf transistor. It may be on the order of 500 to 1000 ohms for a very-low-power rf amplifier, whereas it can be as low as a few ohms for a high-power amplifier. So, remembering that the lower the impedance the more diminished the rms voltage reading will be, we should find somewhat less voltage at point D than we did at C.

Our last test will take place at F, where we should find the highest of all the voltage readings. Point F is very high in impedance, hence the higher voltage appearing there. Collector tap point E of Q3 will show less voltage, owing to that being a lower-impedance area. Likewise with test point G.

We can see that finding the faulty part of the circuit is quite easy by applying the signal-tracing technique. We would proceed in the same manner if we were using an oscilloscope. Once the inoperative stage is spotted, we revert to the troubleshooting methods described in Phases 1 and 2 of this article. If no defective parts are found, we should then try a new crystal at Y1 of Fig. 3. If a spare crystal is handy at the offset, we might make that step no. 1 of our phase 3 test procedure. It could save some time, even though the odds are fairly high against a quality crystal going bad unless it is abused (dropped, or whatever).

Tracing the stages of an audio amplifier is done in a similar fashion to the foregoing. It is necessary, however, to apply an audio tone of normal level at the input to the audio chain in order to obtain an rms voltage response throughout the circuit. The rf probe can be set aside during audio tests. The ac range on your VTVM will be used for testing audio amplifiers. A scope can also be used if one is available. While speaking of scopes, I should mention that the scope used for rf testing needs to have a bandwidth (upper-frequency limit) that is as great or greater than the transmitter frequency. If not, the scope gain will be low and the voltage measurements (pk-pk) will not be accurate.

Digital Circuits

A scope or a logic probe can be used when attempting to find a defective section in a logic circuit of a complex transmitter or transceiver. An in-depth discussion of troubleshooting digital circuits is well beyond the scope of this article.

Harmonics and Spurious Responses

Let us imagine for the moment that we suddenly received a TVI (television interference) complaint from a neighbor, or that the FCC sent us a violation notice for excessive harmonic radiation. The rig seems to be working normally, but for some reason we're putting out one or more signals inside or outside of the band of operation. How could this be? Well, if we find no loose joints in our antenna system (loose joints can cause harmonic currents through rectification), and if our transmitter low-pass filter is not defective, we may assume that something has gone afoul in the transmitter. Possible causes might be instability (self-oscillation) in one or more of the rf stages, or the harmonic filter in the transmitter output could have a bad coil or capacitor. Fig. 4 shows another simplified transmitter section. We will use it for text discussion purposes.

If self-oscillation is the malady in our transmitter, we may find any number of causes for it. Refer to Fig. 4. If C1 should become short-circuited, Q1 could develop excessive gain and break into oscillation. Generally, the emitter bypass capacitor is chosen to provide gain at the high-frequency end of the transmitter operating range, but at the lower end of the spectrum and into the audio region the gain drops off gradually. Too much gain, or too large a capacitance at C1, could cause Q1 to "take off."

Some designers include a parasitic-choke resistor in the base or collector lead of a transistor to stop vhf or uhf self-oscillations. R1 of Fig. 4 is that type of resistor. If it were to become very low in value, say 1 or 2 ohms, it could allow self-oscillations or parasitics to occur. Similarly, if collector bypass capacitor C2 were to become open or too low in value, self-oscillation might take place at Q1.

The condition of C3 might also affect the stability of the transmitter. If it becomes open or too low in value, we may find Q2 to be unstable. In a like manner, base-swamping resistor R2 is used in many designs to lower the circuit Q by swamping T1 to discourage unwanted parasitics. If it became too high in value, or if it were to become open-circuited, problems could result.

C4 of Fig. 4 is also an important bypass capacitor. Often, there are two or more capacitors in parallel at this circuit point, and each is a different value from, say, 0.001 to 0.5 µF. The various values permit effective bypassing at different frequencies in the high-frequency and vhf range. A defective C4 could mean instability of Q2. C5 is a much larger capacitor. It is included to serve as a bypass capacitor for very-low frequencies down to the audio range. Many solid-state amplifiers will self-oscillate vigorously at lf or audio. Even though they are designed to work at hf or vhf. Therefore, the condition of C4 is important to assure stability of the overall Q1-Q2 rf section.

If the transmitter once had a clean bill of health with respect to harmonic output but suddenly developed "harmonic-itis," the output filter, FL1 of Fig. 4, could have a bad component. C6, C7 or C8 could have changed value, become open or perhaps shorted. Similarly, filter inductors L1 or L2 might have developed shorted turns or could have a broken toroid core. Each part should be checked if harmonics occur. Various filter configurations are found in the broad scope of designs, so your filter may have many more or fewer sections than the one depicted in Fig. 4.

Harmonic and Spurious "Sniffing"

It is absurd to consider one of us amateurs buying a $20,000 spectrum analyzer for use in checking harmonic or spurious output. So, we must make do with a low-cost alternative, or poor man's analyzer. A sensitive, calibrated wavemeter of the kind described in the measurements chapter of the *Handbook* can be built inexpensively and used to sniff for unwanted output frequencies at the output stage of the transmitter. We may also employ a dip meter to search for these "frigglies." Fig. 5 shows a simple sampling circuit we can build in a Minibox or homemade pc-board box. It contains a single loop of wire in series with the antenna feed line. Coaxial connectors permit connection to the feed line. The dip-meter coil or probe is inserted through a hole in the box wall (Fig. 5B) and placed near L1. With the transmitter operating into a dummy load (sniffer between the transmitter and dummy load), the dip meter (set in the wavemeter mode) is tuned for peak indications. These responses will tell us where the harmonics or other unwanted output energy are in terms of frequency. Such responses should be a fraction of the response at the desired operating frequency. If they are close to the same magnitude or as strong as the desired frequency, problems are definitely indicated. Troubleshooting must follow.

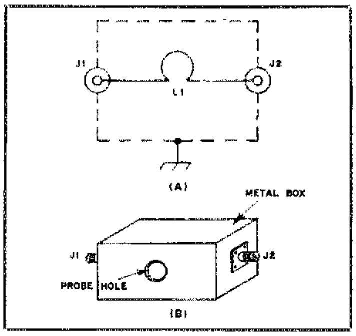

Fig. 5 — At A are the details of a one-turn loop contained in a metal box. This can be used for sampling the transmitter output when looking for unwanted spurious output. The loop can be made from no. 12 or 14 copper wire. The loop diameter is 1 inch (25.4 mm). It should be centered in the box and opposite the hole in the box wall (illustration B). See text for additional details.

In Summary

We have dealt somewhat superficially with the theme of transmitter servicing. Our objective in this article is to illustrate the general approach to finding problem areas in our equipment so that we may avoid the expense and inconvenience of sending the rig back to the factory for repair. All that's necessary for home servicing is a little courage and self-reliance that first time or two. Working on our own rigs will help us to better understand the circuits in these gray, brown or black boxes. At least, let's give it our best shot the next time the old rig lets us down! QST

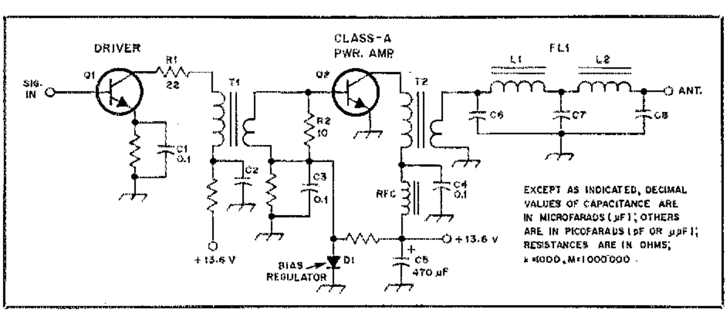

Fig. 4 — A driver and power amplifier illustration. Parts with R, C and L designators are discussed in the text. They could be the cause of instability or harmonic output.

Strays

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Tropospheric Scatter Propagation†

Work DX on vhf, uhf and microwaves using tropospheric scatter propagation. It's easy!

By J. N. Gannaway, PhD,* G3YGF

This article demonstrates the potential of communications while using tropospheric scatter (troposcatter) in the vhf, uhf and microwave bands. The troposcatter propagation mode is used regularly for working DX at vhf, but the predictability of the mode and its potential on the higher frequency bands are not widely appreciated. The nature of troposcatter propagation is explained, and an expression for the path loss is given, taking into account the characteristics of the sites at each end. By combining this expression with calculations of equipment performance at each end of the path, it is possible to predict the signal-to-noise ratio that should be obtained between the two stations under normal conditions, or to estimate the maximum range that can be expected from a given site.

Paths that involve only line-of-sight propagation are not very common, and usually the signals will have been scattered off or diffracted around several obstacles on the way. As the path length increases, so does the number of obstructions or the angles through which the signals have to be diffracted. Under these conditions signal levels will decrease very rapidly with distance, and signals arriving by other propagation mechanisms may be stronger. Propagation beyond the horizon can occur by a variety of methods, usually either by atmospheric ducting or reflecting off an object that is high enough to be visible to both stations. This object can be, for example, an aircraft, an aurora, or an ionization trail from a meteor. These phenomena are short-lived, however and a more permanent mechanism would be desirable. Satellites or moonbounce are more predictable but, apart from the case of geostationary satellites, can only be used some of the time.

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†Adapted from an article of the same title in *Radio Communication* (RSGB), August 1981.

The Mechanism of Troposcatter

Troposcatter uses the weak, but reliable, reflections that can be obtained from the dust particles, clouds and refractive index variations that occur in the atmosphere 1000 to 50,000 ft above sea level.[1] This mechanism can be used for working DX reliably over distances of many hundreds of miles. A brief illustration of the relevant part of the atmosphere is shown in Fig. 1. Air density decreases with height, and reaches one-third of its sea level value at about 30,000 ft. The refractive index of the atmosphere depends on such properties as temperature, density (pressure), humidity or the presence of water. Variations in any of these properties can scatter the signals. The scattering process is more efficient at lower altitudes where the atmosphere is denser. Turbulence associated with the weather can have marked effects on the signal levels and characteristics.

In practice, this mechanism is used by pointing both antennas along the great circle path between the two stations at as low an angle of elevation as possible. The two beams will intersect in a common volume of the atmosphere near the center of the path (Figs. 2 and 3). Propagation will be line-of-sight to the common volume from the transmitter. A very small fraction of the power passing through this volume will then be scattered in all directions by the irregularities in the atmosphere. This power then propagates by line-of-sight to the receiver. The height of the bottom of this scattering volume will depend on the path length, and to some extent on the horizons

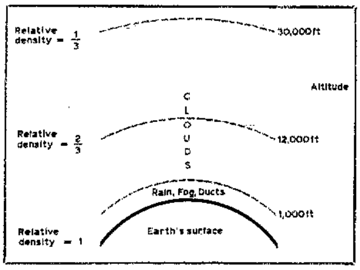


Fig. 1 — Diagram showing the structure of the lower atmosphere.

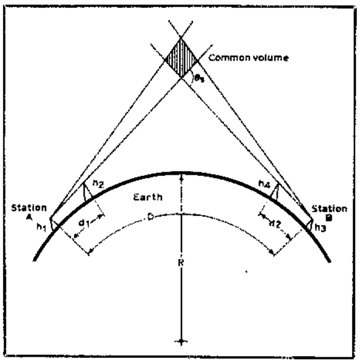


Fig. 2 — The geometry of a troposcatter path.

of the sites, but will be typically 2000 ft on a 60-mile path, and 30,000 ft on a 300-mile path.

The loss in the scattering process is usually so large that the equipment is unlikely to have enough spare capability to overcome the extra losses introduced by

[1]Notes appear on page 48.

any additional obstructions in the path. The path loss increases by about 10 dB for every degree of horizontal angle at each station. On paths over 60 miles the increase is about 9 dB for every extra 60 miles of path length. The choice of a site with a good horizon is vitally important; it can make a difference of several hundred miles in the obtainable range.

Derivation of an Expression for the Path Loss

The angle through which the signal is scattered is an important characteristic of a troposcatter path, because the loss involved increases with angle; the angle involved being usually only a few degrees. The relevant details of a troposcatter path are shown in Fig. 2. The station heights are h_1 and h_3; h_2 and h_4 are the heights of the obstructions forming the horizon at each station, at distances d_1 and d_2, respectively. All heights are with respect to sea level. R is the mean effective radio radius of the earth, 1.33 times the physical radius, i.e. 5280 miles. This allows for the amount by which the atmosphere refracts the signals toward the earth's surface under normal conditions. The scattering angle, θ_s, determined by the path geometry, consists of three terms, one depending on the overall path length and two being characteristic of the sites at each end.

$$\theta_s = \underbrace{\frac{D}{R} \times \frac{180}{\pi}}_{\text{term in path length}} + \underbrace{\left(\frac{h_2 - h_1}{d_1} - \frac{d_1}{2R}\right) \times \frac{180}{\pi}}_{\theta_A \text{ contribution from site A}}$$

$$+ \underbrace{\left(\frac{h_4 - h_3}{d_2} - \frac{d_2}{2R}\right) \times \frac{180}{\pi}}_{\theta_B \text{ contribution from site B}} \quad \text{(degrees)} \qquad \text{(Eq. 1)}$$

The units used for the various distances do not matter, provided they are the same in each term. The path loss can now be expressed as the sum of several components:

The free-space loss[2]

$$L_{fs} = 32.5 + 20 \log D + 20 \log f \quad \text{(dB, km, MHz)} \qquad \text{(Eq. 2)}$$

The loss in the scattering process

$$L_s = 21 + 10\,\theta_s + 10 \log f \quad \text{(dB, degrees, MHz)} \qquad \text{(Eq. 3)}$$

This is an empirical expression derived from observed signal levels, and shows the variation of scattering efficiency with frequency and scattering angle. The loss increases by 10 dB per degree of scattering angle.

These expressions are plotted in Fig. 4, which shows the free-space loss and the sum of the free-space and scatter losses for comparison. Much greater losses are involved in troposcatter, and they increase very rapidly with distance.

The aperture to medium coupling loss

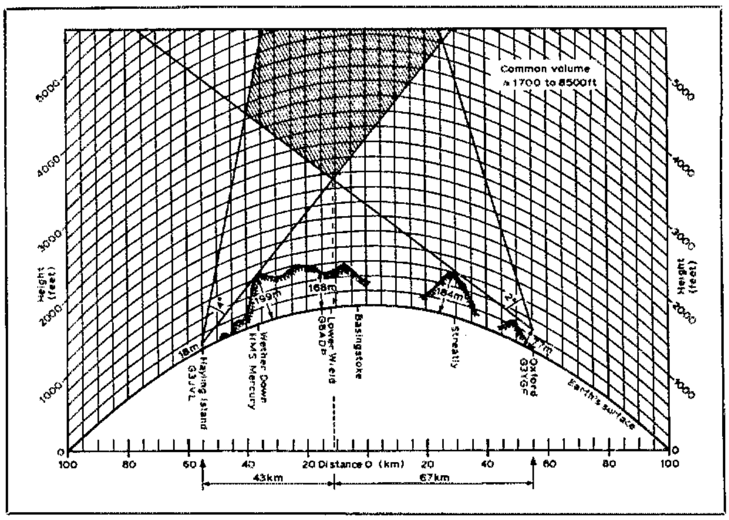


Fig. 3 — Path profile plot for the Oxford to Hayling Island path.

$$L_{am} = 2 + 2\frac{\theta_s}{\alpha} \qquad \text{(Eq. 4)}$$

where α is $\sqrt{\theta_2\theta_3}$ the geometric mean of the two antenna beamwidths.[3] This takes into account the size of the two beams and the way they cross in the atmosphere, which affects the efficiency of coupling between them. It implies that there is no point in increasing the size of the antennas above a certain gain on a given path, because the expected increases in gain will not be realized when very high gain antennas are used. This condition occurs when the antenna beamwidths approach the scattering angle, i.e., a few degrees. This term will be negligible except on the higher frequency bands, where antennas with a beamwidth of a few degrees can be realized conveniently. (E.g., a 4-ft dish on 10 GHz has a 2° beamwidth.)

Loss attributable to variation of the mean radio refractive index of the atmosphere

$$L_N = 0.2\,(N - 310)\text{(dB)} \qquad \text{(Eq. 5)}$$

where N is the refractive index expressed in millionths above unity — the nominal value is 1.000310 (310) — and will be affected by the climatic conditions mentioned earlier. If N varied by 30 units, this would affect the path loss by 6 dB, so it has a significant effect and probably accounts for the seasonal variations referred to later.

Calculation of the Path Loss

The total troposcatter loss is the sum of all these terms. It is convenient to split it into two parts: the basic loss, which only depends on the distance and frequency, and the variable losses owing to the nature of the sites used and the climatic conditions.

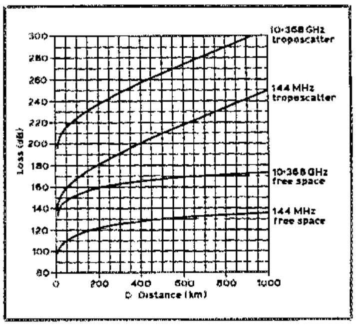


Fig. 4 — Comparison of troposcatter and free-space path losses.

The first part, the troposcatter loss between two stations on a smooth earth,[4] is obtained by taking the terms that are either constant or depend on path length or frequency.

$$L = 55.5 + 20 \log D + 30 \log f + \frac{D}{R} \times \frac{1800}{\pi} \quad \text{(dB, km, MHz)} \qquad \text{(Eq. 6)}$$

This loss is plotted against distance in Fig. 5 for the frequencies 10,368 and 144 MHz. These graphs can be used at other frequencies by adding

$$30 \log \frac{f}{10{,}368} \text{ or } 30 \log \frac{f}{144}$$

to the value obtained from the appropriate curve. Values of this term for the various amateur bands are given in Table 1. The remaining terms are the variable ones that depend on the sites or propagation conditions and weather, so these should then be added to the loss obtained from the graph:

$$L_V = 10\,\theta_A + 10\,\theta_B + \frac{2\theta_s}{\alpha} - 0.2\,(N - 310) \qquad \text{(Eq. 7)}$$

For most purposes the total loss can be taken as the loss from the graph plus the contribution from each site. The other two terms will have little effect, and the value of N is not likely to be known accurately.

Once the details of the sites are known, the values of θ_A and θ_B can be calculated using the expression for θ_s given earlier, or the loss $10\,\theta_A$ can be found directly using the graph in Fig. 6. In this, d is the distance to the first obstruction, and Δh is the height of the obstruction above the site (see Fig. 2).

$$\Delta h = h_{obstruction} - h_{site} \qquad \text{(Eq. 8)}$$

The actual height of the site does not appear explicitly in the expressions, only inasmuch as it determines where the first obstruction is and the height relative to a site. It can be seen from the original expression for θ_s that the elevation angle the obstacle presents and the distance from the site are the important parameters in determining the path loss. There is little to be gained by going higher at a site if the object forming the horizon is far away, but significant improvements are possible if the obstacle is close. A distant horizon is the key feature of a good site which, in simple terms, might be described as a place having a "good view."

It is also very useful to calculate the loss from the site-dependent terms separately, as it provides means of accurately comparing the merits of various sites and is independent of frequency. Path profile plots[5] should be performed for each direction of interest at each site to find the object causing the horizon and thus the values of d and h. A very good site can give negative values of this loss and so reduce the overall path loss. This loss is typically in the range −5 to +10 dB.

Example of Path Loss Calculation

As an example, consider the Oxford-Hayling Island path, a distance of 65 miles. Converting to metric and taking the distances from the path plot in Fig. 3, the site losses are:

Oxford —

$$\theta_A = 57.3°\left(\frac{184\text{ m} - 77\text{ m}}{26{,}000\text{ m}} - \frac{26\text{ km}}{17{,}000\text{ km}}\right)$$

$$= 0.23° - 0.09° = 0.14° \qquad \text{(Eq. 9)}$$

$$\text{Loss} = 10\,\theta_A = 1.4\text{ dB} \qquad \text{(Eq. 10)}$$

Hayling Island —

$$\theta_B = 57.3°\left(\frac{199\text{ m} - 18\text{ m}}{18{,}000\text{ m}} - \frac{18\text{ km}}{17{,}000\text{ km}}\right)$$

$$= 0.57° - 0.06° = 0.51° \qquad \text{(Eq. 11)}$$

$$\text{Loss} = 10\,\theta_B = 5.1\text{ dB} \qquad \text{(Eq. 12)}$$

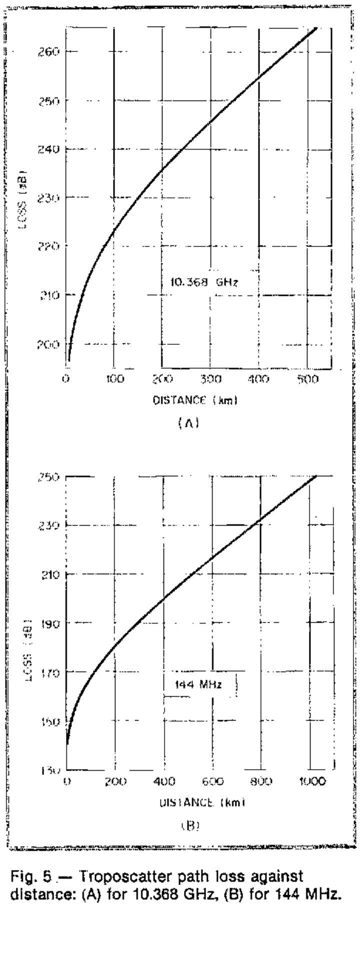


Fig. 5 — Troposcatter path loss against distance: (A) for 10.368 GHz, (B) for 144 MHz.

Table 1
Corrections to the Path Loss in Fig. 5 for Different Frequencies

Band	*Correction (dB) to be added to the loss given in* *Fig. 5A (10,368 MHz)*	*Fig. 5B (144 MHz)*
24 GHz*	+ 11	+ 67
10 GHz	− 0	+ 56
5.6 GHz	− 8	+ 48
3.4 GHz	− 15	+ 41
2.3 GHz	− 20	+ 36
1296 MHz	− 27	+ 29
432 MHz	− 41	+ 14
144 MHz	− 56	0
70 MHz	− 65	−9

*An additional allowance must be made for water vapor absorption on this band.

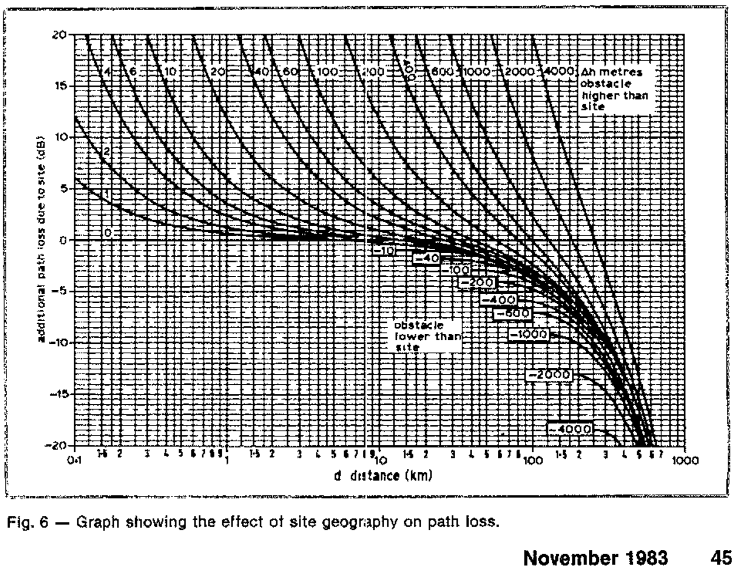


Fig. 6 — Graph showing the effect of site geography on path loss.

The total loss associated with the sites is 6.5 dB. The same results can be obtained by using these values of $\triangle h$ and d in Fig. 6.

Site A: d = 26 km, $\triangle h$ = 107 m
Site B: d = 18 km, $\triangle h$ = 181 m

Next, θ_S is needed to calculate the coupling loss. θ_S is the sum of the horizon angles at each site, plus the term in the total path length,

$$\frac{57.3\ D}{R}$$

$$\theta_S = \theta_A + \theta_B + 57.3° \times \frac{110\ km}{8497\ km}$$

$$= 0.14° + 0.51° + 0.74° = 1.4° \quad \text{(Eq. 13)}$$

Now the coupling loss can be found. The antennas, 2-ft and 4-ft dishes, have beam-widths of 4° and 2° on 10 GHz, so the mean is 2.5°. The coupling loss is then

$$\frac{2.8}{2.5}$$

(approximately 1 dB), which is negligible, as expected.

The path loss from the graph in Fig. 5 for a path of 65 miles (110 km) is 224 dB at 10 GHz, so the total loss is

$$\text{Path loss} = 224 + 6.5 + 1 = 232\ \text{dB} \quad \text{(Eq. 14)}$$

The path loss on 144 MHz can be found in a similar manner. The site losses will be the same, as they are independent of frequency, and the coupling loss will be even smaller, since θ_S is the same but the antenna beamwidths are much larger, typically 20° to 30°. The path loss from the graph is 168 dB, so the total loss is

$$\text{Path loss} = 168 + 6.5 + 0 = 175\ \text{dB} \quad \text{(Eq. 15)}$$

These values are the mean values of loss averaged over a year. There are many factors that will affect this value slightly, and these are discussed later.

Equipment Performance

Now that the path losses (pl) are known, the next step is to calculate the capability of the equipment at each end of the path — the path-loss capability (plc).[6] This is the number of decibels of loss that must be inserted between the transmitter and receiver antennas to give a 0 dB signal-to-noise ratio in the receiver. The difference between the two terms gives the signal-to-noise ratio to be expected in the receiver,

$$\text{s:n} = \text{plc} - \text{pl} \quad \text{(Eq. 16)}$$

The plc can be found for any two sets of equipment. It is the sum of the transmitted effective isotropic radiated power (eirp) and the effective receiver sensitivity (ers),

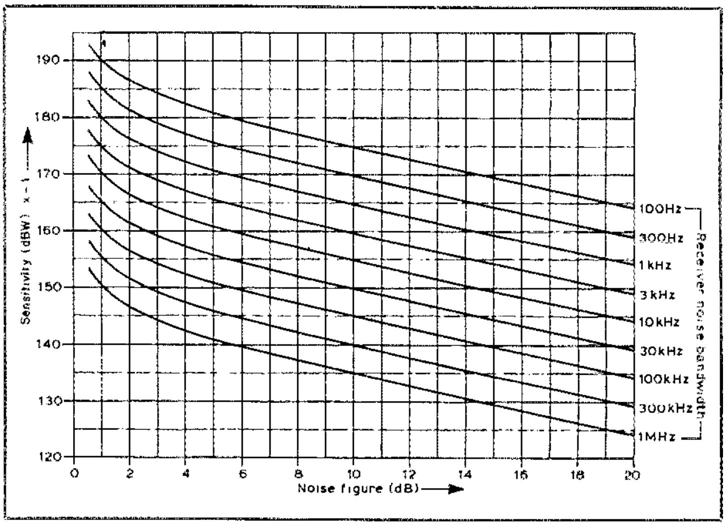


Fig. 7 — Graph of effective receiver sensitivity as a function of noise figure and bandwidth.

$$\text{plc} = \text{eirp} + \text{ers}. \quad \text{(Eq. 17)}$$

The eirp is the sum of the transmitter power in decibels relative to 1 W (dBW). Antenna gain is in decibels relative to an isotropic radiator (dBi). The feeder loss in decibels is

$$\text{eirp} = \text{transmitter power (dBW)} + \text{antenna gain (dBi)} - \text{feeder loss (dB)} \quad \text{(Eq. 18)}$$

Calculation of the ers is slightly more involved. The noise level of the receiver is expressed in decibels below 1 W, but as a positive number of decibels. The noise level can be found from the expression for thermal noise, — 10 log (kTB), where k is Boltzmann's constant,

1.38×10^{-23} W/K Hz

T is the receiver noise temperature, which is related to the more familiar noise figure by the relation

$$\text{nf} = 10 \log \left(1 + \frac{T}{290}\right)$$

and B is the receiver bandwidth in hertz. Values for this noise level can be found from the graph in Fig. 7, which is for modes such as ssb or cw, which have no detector threshold. For a-m detectors the threshold is 2.6 dB, but for fm the situation is more complicated.[7] The threshold increases with modulation index up to about 10 dB for wideband fm. These values represent reductions in the receiver sensitivity. Feeder loss and antenna gain are included:

$$\text{ers} = \text{receiver noise level (dBW, a positive number of decibels)} + \text{antenna gain (dBi)} - \text{feeder loss (dB)} - \text{threshold (dB)} \quad \text{(Eq. 19)}$$

Examples of PLC

The plc will now be calculated for the equipment used on the path mentioned earlier on both 144 and 10.368 MHz.

144 MHz

G3JVL (transmit)

Transmitter 20 W	+ 13 dBW
Feeder loss	− 1 dB
Antenna, 14 element	+ 14 dBi
eirp =	+ 26 dBW

(Eq. 20)

G3YGF (receive)

Receiver, nf 5 dB, bandwith 2.5 kHz	+166 dBW
Feeder loss	0 dB
Threshold (ssb)	0 dB
Antenna, 4 element	+ 6 dBi
ers	+ 172 dBW

(Eq. 21)

$$\text{plc} = \text{eirp} + \text{ers} = 198\ \text{dB} \quad \text{(Eq. 22)}$$

10,368 MHz

G3JVL (transmit)

Transmitter, 5 W	+ 7 dBW
Feeder loss	− 2 dB
Antenna, 2-ft dish	+ 34 dBi
eirp =	+ 39 dBW

(Eq. 23)

G3YGF (receive)

Receiver, nf 8 dB, bandwidth 500 kHz	+169 dBW
Antenna, 4-ft dish	+ 39 dBi
Feeder loss	− 2 dB
Threshold (cw)	0 dB
ers =	+206 dBW (Eq. 24)

plc = 245 dB (Eq. 25)

The path losses calculated earlier are 232 dB (10,368 MHz), and 175 dB (144 MHz). The predicted signal-to-noise ratios can now be calculated.

	Observed value	*Predicted value*
10,368 MHz	10 dB	245 − 232 = 13 dB (Eq. 26)
144 MHz	17 dB	198 − 175 = 23 dB (Eq. 27)

The 144-MHz measurement was done on only one day, so it is likely to be rather inaccurate, but the 10,368-MHz measurements were done over a period of many months and show good agreement with the predicted values. The discrepancy is probably caused by the seasonal and climatic variations, for which no allowance has been made, although there will always be a few decibels of uncertainty in the equipment parameters, antenna gains, etc.

The Potential of Troposcatter Communications

Details of various systems and the range that can be expected between two stations using them are given in Table 2. They illustrate the performance that should be expected under normal conditions from good sites. The loss contributions from the sites are assumed to be zero, and the figures are given for a signal-to-noise ratio of 0 dB in a 100-Hz bandwidth, representing a weak cw signal. The range obtained when using ssb in a bandwidth of 2 kHz, a factor of 13 dB larger, will be about 209 mi on each band. The equipment is typical of that which might be used for a serious entry in a portable contest. The noise figures given may seem rather high, but are those of the overall system, which may be significantly higher than those of the preamplifiers on their own.

Table 3 gives the troposcatter range between two systems that are capable of moonbounce communication to illustrate the relative magnitudes of the problems involved. For distances approaching 600 miles the challenge represented by the two modes of communication is comparable. These tables also show that the range attainable by troposcatter need not, in theory, vary much with frequency. In practice, however, physically smaller antennas tend to be used on the higher frequencies. It is also harder to generate comparable power levels on the higher bands. In view of the rapid advances being made in receiver and transmitter technology, however, the potential of the microwave bands for longer distance communication should not be ignored.

Characteristics of Troposcatter Signals

Several types of fading are experienced on troposcatter signals.[8] The effects are more severe at high frequencies, so are easier to observe and describe. At 10 GHz the note of the carrier can appear quite rough, being modulated by the scattering process at frequencies up to about 50 Hz. An example of this rapid fading is shown in Fig. 8 (A and B), which are oscilloscope photographs of a continuous carrier received over a 65-mile path from G3JVL to G3YGF, showing both the depth of the fading and range of frequencies over which it occurs. At times it produces a waveform that resembles 100 percent amplitude modulation. Fig. 8C shows the fading on a longer time scale, and the occasional very deep fades can corrupt cw as shown in Fig. 8D, where in the V of G3JVL the dash has been broken up into two dots.

There is also fading over a period of minutes and, in the longer term, signals tend to show a diurnal variation of about ±5 dB. These swings often peak in the afternoon when atmospheric turbulence caused by convection currents from the warm ground is at a maximum. Plots of signal level showing this effect are given in Fig. 9. There is also an annual variation of similar amplitude, with signals peaking in the summer and being at a minimum in the winter. The daily and annual variations are probably the result of corresponding variations in the average value of N over the path. These fading effects will all exist at lower frequencies, but the rates and depth of the short-term fading will be correspondingly slower. They can be seen as the slow fading on vhf DX signals, which has a period of several minutes.

The rapid fading is caused by the signal being scattered from various regions of air, each of which may be in turbulent motion, and moving relative to each other. This motion can cause both frequency and amplitude modulation of the signals. Frequency modulation results from the signals being scattered from air masses that are moving at different speeds, so there will be random Doppler shifts on the signals. At 10 GHz a speed of 30 mph will produce a shift of about 500 Hz, and this effect can spread the energy of the carrier out over 1 kHz or more; heavy rainstorms produce a sound rather similar to an auroral signal. This effect will also scale with frequency and so will be far less noticeable at vhf. These storms can also increase the signal levels by around 10 to 20 dB, as the raindrops scatter the signal more effectively. Amplitude modulation results from variations in the scattering efficiency or interference effects between signals arriving by different paths.

Several enhancements of 10 to 15 dB on 10 GHz have occurred at the same time as big lifts on the vhf bands, e.g. on November 10, 1978 and October 3, 1980. The note was T9, and very good quality ssb was obtained, showing that the clean signal that was enhanced by the ducting had swamped the normal rough troposcatter one.

Conclusion

This method of calculating troposcatter

Table 2
Range Obtainable by Troposcatter Propagation on Various Amateur Bands
(100 Hz Bandwidth Receiver)

Freq. (MHz)	Path Loss (dB)	Range (km)	Equipment	Antenna Gain
144	240	870	100 W, 3 dB nf, 2 × 16-el Yagi	18 dBi
432	247	790	100 W, 3 dB nf, 2 × 25-el loop Yagi	22 dBi
1,296	258	760	100 W, 3 dB nf, 4 × 25-el loop Yagi	24 dBi
2,304	262	720	50 W, 3 dB nf, 6-ft dish	31 dBi
10,368	234	240	100 mW, 10 dB nf, 4-ft dish	39 dBi
10,368	254	440	1 W, 3 dB nf, 4-ft dish	39 dBi

Table 3
Troposcatter Range of Equipment Capable of Moonbounce Operation

Freq. (MHz)	*EME Path Loss (dB)*	*Tropo Range (km)*	*Equipment*	*Antenna Gain*
144	252	990	500 W, 3 dB nf, 100 Hz, 4 × 16-el Yagi	21 dBi
432	262	940	500 W, 3 dB nf, 100 Hz, 20-ft dish	26 dBi
1,296	271	890	500 W, 3 dB nf, 500 Hz, 16-ft dish	34 dBi
2,304	276	860	100 W, 3 dB nf, 500 Hz, 16-ft dish	40 dBi
10,368	289	790	50 W, 3 dB nf, 1 kHz, 12-ft dish	50 dBi

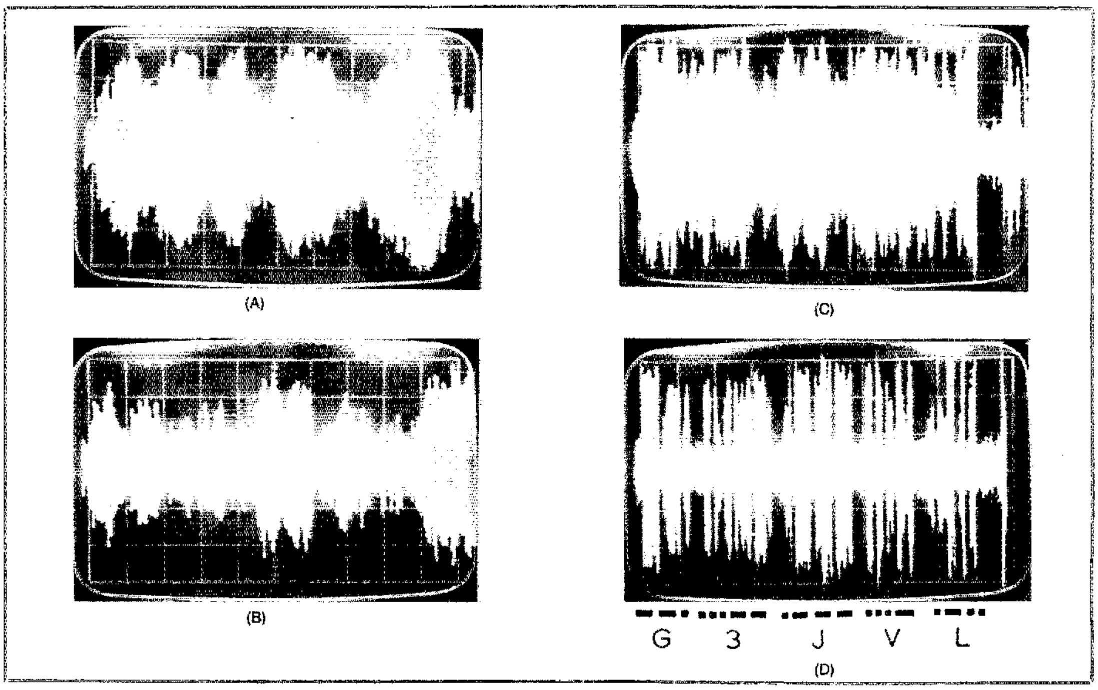


Fig. 8 — Oscilloscope traces showing rapid fading on troposcatter signals received from G3JVL on 10.368 MHz over a 65-mile (110-km) path in March 1979. At (A), continuous carrier, 50 ms/cm, showing fading at around 10 Hz. At (B), continuous carrier, 50 ms/cm, showing more random fading. At (C), continuous carrier showing fading on a longer time scale, 0.6 s/cm. At (D), Morse code from G3JVL showing occasional deep fading, 0.6 s/cm.

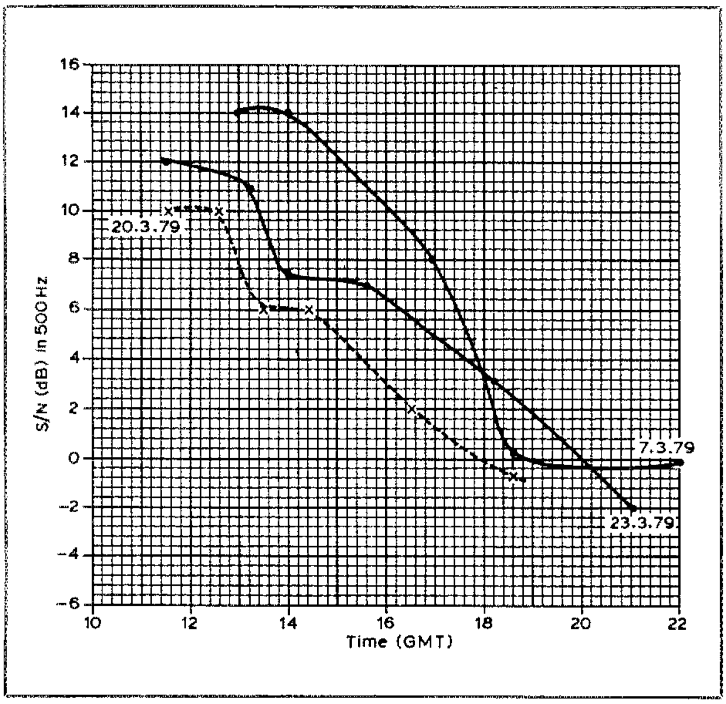


Fig. 9 — Plots of signal strength against time showing diurnal variations.

path losses has given predictions of signal levels that have been shown to be accurate to within a few decibels over a number of paths on 144 MHz and 10 GHz from 60 to 300 miles long, when used in conjunction with the calculations of the path loss capability of the equipment.

It enables the merits of various sites to be compared more scientifically, and an estimate to be given of the distances that should be workable under normal conditions. It has also demonstrated the potential of narrow-band modes on 10 GHz and has revealed several interesting propagation effects that can be investigated. The papers in the notes will provide more background information on the subject, and are quite light reading.

Notes

[1]mm = in. × 25.4; m = ft × 0.3048; and km = mi × 0.621.

[2]"Microwaves," *Radio Communication*, May 1978.

[3]Booker and DeBettencourt, "Theory of Radio Transmission by Tropospheric Scattering Using Very Narrow Beams," *Proceedings of IRE*, March 1955, p. 281.

[4]Bullington, Inkster and Durkee, "Results of Propagation Tests at 505 and 4090 MHz on Beyond the Horizon Paths," *Proceedings of IRE*, Oct. 1955, p. 1306.

[5]B. Chambers, "Microwave Path Checking," *Radio Communication*, March 1978.

[6]"Microwaves," *Radio Communication*, March 1978.

[7]D. Middleton, "On Theoretical Signal to Noise Ratios in FM Receivers," *Journal of Applied Physics*, April 1949, Vol. 20, pp. 334-351.

[8]L. Yeh, "Simple Methods for Designing Troposcatter Circuits," *IRE Transactions on Communications Systems*, Sept. 1960, p. 193.

Product Review

Conducted By Paul K. Pagel,* N1FB

Yaesu Electronics Corp. FT-77 HF Transceiver

Yaesu describes the FT-77 as a thrifty hf transceiver. This compact radio is easy to operate and well suited to mobile installations. A microphone or key, an antenna, and a power supply (20 A at 13.5 V) are all that is required to begin operation.

Features

The FT-77 is a solid-state transceiver, and no final amplifier tuning is required. It is rated to produce 100 watts (85 watts on 10 meters) of ssb or cw output power. Output power must be reduced to 50 watts when using the optional fm unit. Operating frequency is controlled by a VFO or a single crystal (not included). A scanning, synthesized, external VFO (FV-700DM or FV-707DM) is optional. Each optional accessory VFO has frequency-storage capability.

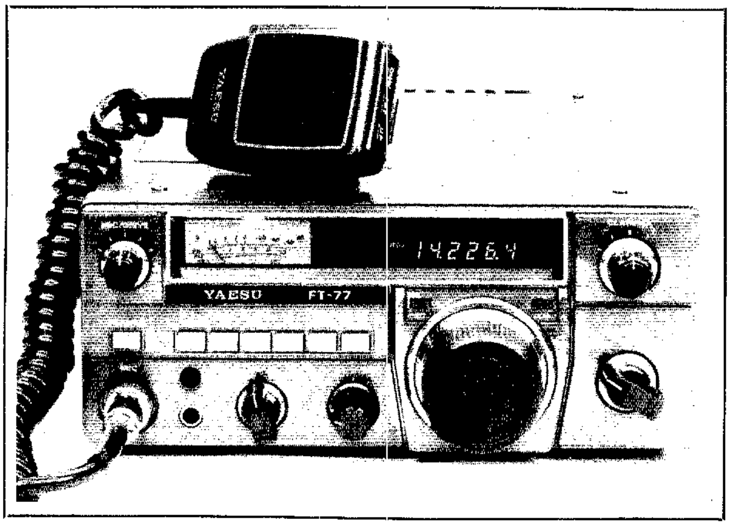
The transceiver has a single-conversion receiver section with the i-f at 8987.5 kHz. When the fm unit is operating, the i-f is shifted up 455 kHz and a second conversion is made from 455 kHz. The receiver bandwidth is 2.5 kHz for ssb and cw (wide). Ardent cw operators will probably want the optional, narrow-bandwidth filter (0.6 kHz). There are slow and fast agc settings, but no way to completely stop agc action. Two noise blanker settings are available. Set to the narrow position, the noise blanker will eliminate normal automotive ignition noise. In the wide position, longer noise pulses created by sources such as over-the-horizon radar ("woodpecker") will be reduced. A SQUELCH control is included for use with the fm unit. An optional 25-kHz marker unit is available.

I received the optional scanning microphone (MH-1B8) with the FT-77. The scanning function does not operate without the optional scanning VFO, but the microphone has a switch that allows a choice of frequency response characteristics. I found this microphone helpful with my low and "muddy" voice.

Front Panel Functions

The MIC/CAR control sets the microphone gain in the ssb mode and carrier level in the cw and fm modes. Below the MIC/CAR control is a push-on/push-off power switch. To the right of the power switch, but separate from it, is a row of six more push switches. The rf attenuator is operative when the RF ATT button is pressed. An LED at the upper left of the tuning knob lights when the attenuator is in use. Pressing the next button activates the noise blanker. AGC-F selects fast agc action for reception of cw or fast-fading ssb signals. FIX activates the crystal oscillator as the frequency control for the FT-77. MARK turns on the optional 25-kHz marker for confirmation of the frequency-display accuracy. A press of the CLAR button switches the clarifier (RIT) on; an LED to the upper right of the tuning knob indicates that it is on.

At the lower left of the front panel is an eight-pin jack with connections for audio input, PTT, up-scan, down-scan and fast-scan (up or down) controls. A ¼-inch-diameter PHONES jack allows connection of stereo or monaural headphones.[1] The RECord jack (2-conductor, "mini" phone plug) provides a fixed level (70 mV at 50 kΩ) for a tape recorder (this feature is useful if records of third-party traffic are kept on tape). LSB, USB, CW-W (wide), CW-N (narrow), and FM positions are available on the MODE switch. AF (gain) and SQL (squelch) controls are on concentric shafts, just left of the tuning knob. A hard-rubber tuning knob (similar to the focus ring of some cameras) covers about 17 kHz per revolution. The BAND switch allows selection of any one of the present hf amateur bands and the future allocations at 18 and 24 MHz. The CLARIFIER (RIT) shifts the receive frequency up to 3.4 kHz above, or 4.0 kHz below, the transmit frequency when the CLAR button is depressed.



Yaesu Electronics Corp. FT-77 HF Transceiver, Serial No. 2M010010

Manufacturer's Claimed Specifications	*Measured in ARRL Lab*		
Frequency coverage: 3.5 to 4.0, 7.0 to 7.3, 10.10 to 10.15, 14.0 to 14.35, 18.0 to 18.5, 21.0 to 21.45, 24.5 to 25.0, 28.0 to 29.7 MHz.	As specified.		
Modes of operation: cw, ssb, (fm optional).	As specified.		
Frequency display: 6 digit.	3/8 in. high, 6-digit blue vacuum tube.		
kHz/turn of knob: 15.	17.		
Frequency resolution: 100 Hz.	As specified.		
Backlash: Not specified.	Nil.		
S-meter sensitivity (μV/S9 reading): Not specified.	80 m, 24.5; 40 m, 24; 30 m, 35; 20 m, 40; 18 m, 35; 15 m, 28; 12 m, 30; 10 m, 25.		
Transmitter output: 80-15 m, 100 W PEP; 10 m, 85 W PEP.	80 - 15 m, 100 W; 10 m, 95 W.		
Spurious suppression: Better than 40 dB.	-54 dB (see photo).		
Third-order IMD: Not specified.	−35 dB (see photo).		
Receiver sensitivity: Less than 0.15 μV for 10 dB S+N/N.	Receiver dynamics measured with optional 600-Hz filter installed.		
		80 m	*20 m*
	Noise floor (MDS) dBm:	−139.5	−139.5
	Blocking DR (dB):	99	99
	Two-tone, third-order IMD DR (dB):	92	94
	Third-order intercept:	−1.5	−1.5
Size (HWD): 3.7 × 9.5 × 11.6 inches†	As specified.		
Weight: 13.2 lb.	As specified.		
Price class: FT-77, $600; FP-700, $135.			
Available from: Yaesu Electronics Corp., 6851 Walthall Way, Paramount, CA 90723.			

†mm = in. × 25.4; kg = lb × 0.454.

*Assistant Technical Editor

[1]mm = in. × 25.4.

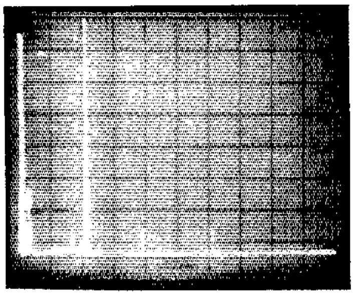


Fig. 1 — Worst-case spectral display of the FT-77. Vertical divisions are each 10 dB; horizontal divisions are each 5 MHz. Output power is approximately 100 W at a frequency of 10.105 MHz. All spurious emissions are approximately 53 dB below peak fundamental output. The FT-77 complies with current FCC specifications for spectral purity.

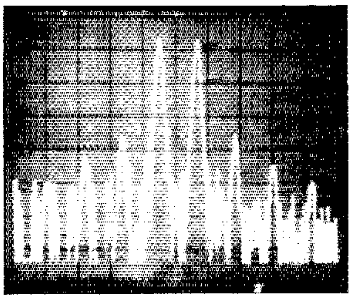


Fig. 2 — Spectral display of the FT-77 output during the transmitter two-tone IMD test. Third-order products are approximately 35 dB below PEP, and fifth-order products are about 44 dB down. Vertical divisions are each 10 dB; horizontal divisions are each 1 kHz. The transceiver was being operated at rated input power on the 20-meter band.

The vacuum-tube frequency display is blue and has 100-Hz resolution. At the left side of the frequency display are frequency control indicators: VFO-A for the internal VFO, VFO-B for the external VFO, or F (fixed). Left of the frequency display is the meter that reads S units on receive, relative forward, power, and reverse power, or alc on transmit. The meter sensitivity and function, sidetone level, cw break-in delay, a crystal socket and a trim control for the crystal frequency are all located under a hatch on top of the FT-77 case.

Rear Panel

Rear panel connections allow for a wide range of operating possibilities. RF OUT provides a low-level (220 mV at 50 ohms) rf source for a transverter (FTV-700 optional). ACC 1 is a six-pin DIN jack with switching and alc signals for a telephone interconnect (phone patch) or power amplifier. A 1/8-inch phone jack is supplied for an external 4- to 16-ohm speaker (EXT SP). There's a dc power connector (DC 13.5 V) for the FT-77 (1 A on receive and 20 A on transmit). ACC 2 is a seven-pin DIN jack with connections for scanning control, transmitter audio input, PTT, and 13.5 V-dc transmit-signals from an external VFO. An eight-pin DIN jack EXT VFO is used for frequency control input from, and power to, an external VFO. The DC 8 V jack provides power for the optional FC-700 antenna tuner. KEY accepts a ¼-inch phone plug for cw keying (0.4 mA @ 1.5 V maximum). Connections are also provided for the antenna (SO-239) and station ground.

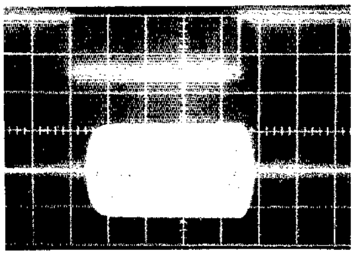


Fig. 3 — Cw keying waveform of the FT-77. Upper trace is the actual key closure; lower trace is the rf envelope. Each horizontal division is 5 ms.

Operation

This is simple: Switch on the power, and transmit! I found tuning with the rubber knob awkward at first and the tuning is a little fast for mobile operation. The scanning VFO option will make mobile operation easier. RIT offers a slower tuning rate than the main tuning knob and may be used as a bandspread control for receive only. Since the transceiver is intended for mobile use, no VOX circuit is present for voice operation. Cw operation is semi-break-in with a variable delay. The narrow noise blanker helps fight engine noise when mobile.

Comments

The FT-77 is a fine transceiver, but there is always room for improvement. The meter-function control and noise-blanker-selection control locations are awkward. To operate the controls, one must remove a hatch cover and reach inside the case to manipulate slide-switch settings. This procedure is troublesome at home, difficult and maybe impossible (dependent upon mounting location) while mobile. I am pleased to find a transceiver with an SWR meter and selectable noise blanker built in, but the controls should be on the case exterior.

The receiver "hears" the frequency counter, faintly, all across the 80-meter band. There are 20 spurious responses that are strong enough to compete with a weak cw signal in the hf bands.

The first time I used the FT-77 in my vehicle I had a problem. Each time I pressed the PTT button, the meter dropped to zero — no power at all! After an hour of searching, I found that a jumper wire had been omitted from the power connector. I installed the jumper, as shown in the manual, and when I again pressed the PTT button, the meter needle climbed the scale — and immediately fell to zero again! The fuse supplied in the dc power cord was rated for 2 A, but a 20-A fuse is required. With the proper fuse installed, the set works well. A quick QSO with a station in southern Indiana confirmed the successful installation.

The FT-77 is easy to operate. I can switch on and transmit in a few seconds. The radio is easily portable and installation goes quickly. Reports indicated a clean transmitted signal and good audio quality. Receiver performance is excellent.

Conclusion

I am happy with the FT-77. Through most of the trial period I used a mobile whip antenna. Eastern Europe and South America were worked easily. I erected a ground-mounted vertical antenna near the end of the review period and Israel was contacted the first afternoon. The FT-77 gave me many hours of enjoyable operating. The radio is fun! — *Bob Schetgen, KU7G*

ICOM IC-45A 450-MHz FM TRANSCEIVER

☐ The ICOM IC-45A is a small 450-MHz fm radio with big-rig performance. Packed with just about every conceivable feature, the IC-45A follows in the footsteps of the 2-meter IC-25A. In fact, they're matching units, and can be stacked neatly.

Not on 450 yet? The IC-45A could be your ticket to blissful solitude and quiet, away from the masses on 2 meters. Enjoy longer QSOs, and radio radio silence in the monitoring mode when you wish.

"Versatility" describes the IC-45A: a microcomputer provides many operating capabilities. Multipurpose scanning, dual VFOs, a continuous tuning system with protection against out-of-band operation, and priority/memory channel capabilities all add up to give the fm fan all he or she can handle.

Thanks for the Memories ...

Up to five frequencies can be programmed easily into the memory channels. I found this memory capability handy. A click of the switch, and you're on the next programmed frequency — simple.

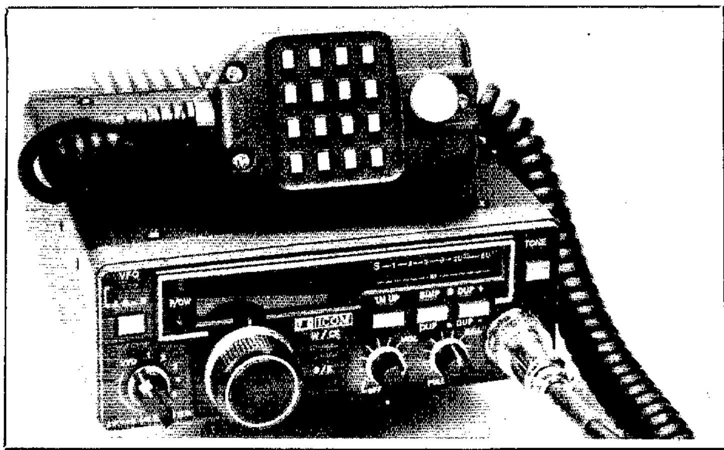
Scanning is a particularly useful function on any rig operating in the sparsely populated 450-MHz band. Multipurpose scanning capability is a feature of the IC-45A. The memory scan allows monitoring of the five memory channels and the two VFO frequencies *alone* — you don't have to wade through 10 MHz of spectrum to find a busy spot. (But, you *can* if you so desire). The program scan provides scanning between two programmed frequencies. Thus, you can search for a busy (or clear) frequency just about anywhere, with maximum efficiency.

Two VFOs can be used independently. Oddball splits can be implemented easily. VFO A changes frequency in 5-kHz steps; VFO B in 25-kHz steps. The dual VFOs allow you to set a certain frequency with one VFO, work up and down the band with the other VFO, and periodically check the set frequency simply by switching between VFOs A and B.

Another neat feature is instant monitoring of a repeater's input frequency (your transmit frequency) to determine if your friend is within simplex range. When the SIMP/DUP switch is in the "DUP" position and the memory W/CK button is pushed, you can receive the repeater's input frequency instantly.

A priority channel allows you to check your favorite channel, such as a local repeater or simplex frequency stored in a memory channel while operating on a VFO frequency. Every few seconds, the dial (receive) frequency will change to the priority channel to see if it's busy: five seconds of VFO frequency, then one second of priority channel check.

One thoughtful feature is the 1M UP switch. By pressing this switch, the dial frequency is raised by one MHz. This saves a good deal of arm-wrenching, and finger-twisting when traveling from the low end of the band to the high end.



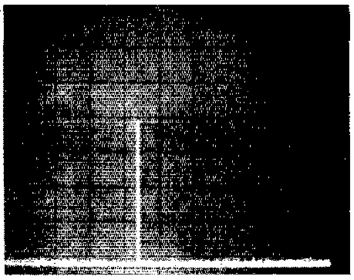


Fig. 4 — Spectral display of the IC-45A. Vertical divisions are each 10 dB; horizontal divisions are each 100 MHz. Output power is approximately 16 W at a frequency of 444.0 MHz. Spurious emissions are all greater than 68 dB below peak fundamental output. The fundamental has been reduced in amplitude approximately 31 dB by means of notch cavities; this prevents analyzer overload.

ICOM IC-45A 450 MHz FM Transceiver Serial No. 01311

Manufacturer's Claimed Specifications	*Measured in ARRL Lab*
Frequency range: 440.000-449.995 MHz.	Same.
Operating mode: 16F3	
RF Output Power: 10 W HI; 1 W LOW	16 W HI; 1.5 W LOW
Spurious emissions: −60 dB or better.	Better than −60 dB
Power requirements: 13.8-V dc ± 15%.	Same.
Current drain: at 13.8-V dc:	
rx 0.5 A (squelched).	Same.
0.7 A (full audio output).	Same.
tx 3.5 A (10 W).	3.4 A.
1.7 A (1 W).	1.6 A.
Receiver type: Double-conversion superheterodyne	
1st i-f — 21.8 MHz.	
2nd i-f — 455 kHz.	
S-meter sensitivity: Not specified.	4 µV/S9.
Sensitivity: Better than 0.3 µV for 12 dB	
SINAD; better than 0.4 µV for 20 dB noise quieting.	−119 dBm/20 dBq.
Squelch sensitivity:	0.1 µV (min) 0.5 µV (max).
Audio output power: More than 2 W.	
Dimensions (HWD): 2 × 5.5 × 7 in.†	
Weight: 3.3 lb.	
Color: Black.	

†mm = in. × 25.4 kg = lb. × 0.454

Remember, the IC-45A covers *10 MHz* of spectrum!

General

The IC-45A is durable, compact and easy to use. I found that the two power levels afforded fine performance for my mobile applications; an outboard amplifier was never required. Frequency choice is determined by the tuning control knob. Depending upon which VFO is employed, one complete rotation of the tuning knob results in a frequency change of 250 kHz or 1,250 kHz.

For home/base use, you'll need at least a 4-A power supply. The microphone has a convenient up and down switch arrangement for ease of frequency selection. A 16-key DTMF pad is also incorporated on the PTT mike.

Performance

I installed the review unit in my pick-up truck. A standard ¼-wavelength whip antenna was chosen for use with the rig. Pick-up trucks are not known for smooth rides. But this did not bother the IC-45A. It survived bumps and jolts on some of New England's ruttiest back roads.

Audio output is of sufficient level to be heard over the ambient noise in my truck. This is quite a feat considering the rig was mounted *behind* my seat. Reports of my transmitted audio were excellent — full and crisp. ICOM's reputation for good quality audio is manifested in their IC-45A.

The green color of the LEDs seemed to contribute to their readability in high ambient light situations, although LCD displays are still superior in these applications. But at night, they glowed clearly, providing for ease of reading.

I liked the feature of the low and high power option. When the truck engine was turned off, I could continue to operate at low power to avoid excessive battery drain.

The LED S-meter is easy to read even in high ambient light conditions. I also found the meter to be aesthetically pleasing to the eye. In fact, the entire rig is attractive with its avionics look. I had many admiring comments from both hams and non-amateurs alike.

The receiver is sensitive, and performs well. I experienced no difficulties with desensitization. The IC-45A worked well in cold temperatures as well as in direct sunlight and heat. Frequency stability was never a problem.

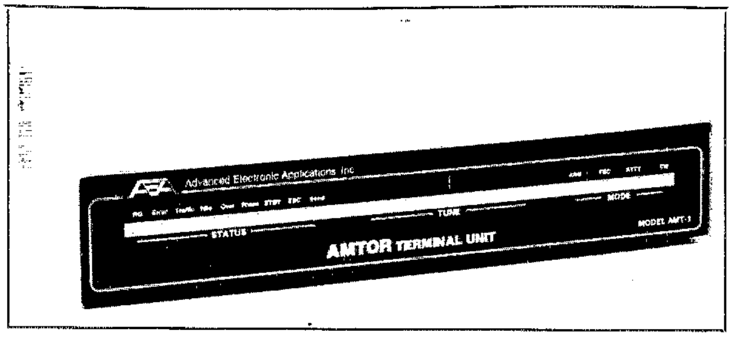
Overall, I enjoyed reliable performance with the review unit. The manual is complete, easy to read and understand. Try the "wilderness" approach to radio, and get on 450 MHz. You'll be glad you did! The IC-45A is available from ICOM America, Inc., 2112 116th Ave. NE, Bellevue, WA 98004. Price Class: $400. — *Rick Palm, K1CE*

AEA AMT-1 AMTOR TERMINAL UNIT

☐ AMTOR (Amateur Teletype Over Radio) is an almost error-free mode of RTTY operation. Although it has been used for years commercially as SITOR, it is relatively new to the U.S. amateur fraternity. It has been in use on the amateur bands, primarily in Europe, but has just recently been authorized by the FCC for use on the hf bands where F1 emission is permitted. (For a complete explanation of the system, see the articles by Martinez and Newland in the June 1981 and July 1983 issues of *QST*, respectively.)

The AEA AMTOR Terminal Unit will get you on AMTOR quickly and easily if you have a home computer or an RTTY/ASCII terminal unit with RS-232-C compatible output. In addition to your regular transceiver you need a 12-V dc source, a five-wire cable to the terminal and a 4-wire cable to your transceiver. Connections to the terminal provide for serial data transmission to and from the AMT-1, ground and, for full interfacing, CTS (clear-to-send) and RTS (ready-to-send) lines. Connections to the transceiver are audio in/out, PTT and ground. The transceiver is used in the usb mode, and 1275- and 1445-Hz tones from the AMT-1 are fed into the audio input to produce the fsk signal. Be sure you have sufficient carrier rejection so your signal is really F1. If you desire to use the AMT-1 with an existing loop circuit, an audio tone to 20- or 60-mA loop interface will be required. In addition to AMTOR, the AMT-1 will receive and send Baudot RTTY up to 100 bauds and will send cw up to 100 wpm. The AMT-1 will not receive cw and will not receive or send ASCII. It can be bypassed for ASCII operation, but then must be switched off and on again for normal operation. The manual includes programs for adapting some of the more common home computers to the AMT-1.

The AMT-1 contains no external controls and, in normal use, no internal adjustments should be required. 110-baud ASCII data from the terminal ESC and CTRL functions is used to control the AMT-1. The front panel contains a 16-LED tuning display, a 4-LED mode indicator and a 9-LED status indicator. A properly tuned-in Baudot RTTY or AMTOR signal will result in



a two-green-LED display, with the dots being spaced equidistantly from the center line of the display. A red LED will indicate whether the AMT-1 is in the ARQ, FEC, RTTY or CW mode. In the status display, white LEDs are the indicators. ERROR is lit when a received block contains an error, RQ when a distant station requests a repeat, TRAFFIC when text is being sent or received, IDLE when no traffic is being sent, OVER when a change of direction of an ARQ contact is occurring, PHASE when the AMT-1 is trying to synchronize with another station, STBY when the unit is waiting to be used, ESC when the AMT-1 is in the escape mode and the next signal from the terminal will be a command. A red LED lights in the transmit mode. Parameters such as SELCAL, AMTOR time-out, RTTY baud rate, cw speed, echo on-off and automatic new line on-off may be programmed and will remain as long as the unit is not turned off. ESC Q will list the current settings of these parameters.

You will find operating AMTOR to be different from anything you have done before. When you first turn on the AMT-1 and tune in an ARQ signal you will probably find that the copy is anything but perfect. The slave may call for repeats while you are copying perfectly and your copy will show many blocks repeated. Conversely, you may get errors when the slave doesn't and those blocks will not be repeated. However, when you are in synchronization with another station you will receive almost completely error-free copy despite fading and interference.

Another AMTOR feature is SELCAL. A self-assigned four-letter group is entered into your AMT-1, the letters usually being derived from your call. For example, W1AW uses WWAW and G3PLX is GPLX. When another station calls using your SELCAL group, your station will automatically respond. The normal way to end a transmission is with +? which automatically reverses the master/slave relationship. The last station to transmit at the end of a contact finishes with CTRL D, which switches both stations to the standby mode.

The AMT-1 has a 960-character buffer so you can enter text in advance. However, when the data rate slows down because of QRM you must be careful not to exceed the buffer's capacity. This is where the RTS connection comes in handy; the text will be held in your terminal until there is room for it. An interesting characteristic of AMTOR is that the maximum range is governed by the speed of your transceiver change-over relay. With a change-over delay of less than 15-ms, contacts almost half-way around the world are possible. Slower relay action will shorten the range.

The manual is well-written with respect to the setup and operation of the AMT-1 itself, but I found it a bit vague concerning operating procedures. Fortunately, a contact or two straightened out the confusion. At the present time, almost all AMTOR activity is near the calling frequency of 14075 kHz. Undoubtedly, when there is sufficient activity, calling frequencies will be established on the other bands as well. The AMT-1 measures 2½ × 12¼ × 9¼ inches (HWD). It is distributed by Advanced Electronic Applications, Inc., Bldg. O & P, 2006 196th SW, Lynwood, WA 98036. Price class: $590. — *Chuck Bender, W1WPR*

SUPER-RATT RTTY/CW SOFTWARE

□ Super-RATT is a popular RTTY/cw software package designed to be used with an Apple® II (or //e) computer with 48 K of RAM, Applesoft BASIC in RAM and at least one disk drive using DOS 3.3. If you've got a clock card and another disk drive, you'll find them useful, but they're not absolutely necessary to use the software effectively.

The Super-RATT package consists of a 5¼-inch floppy disk containing several programs and a 70-page manual describing their functions. The RTTY software allows the computer to send and receive Baudot or ASCII (upper and lower case) RTTY at the popular speeds from 40 to 300 bauds. Speeds of from 5 to 100 wpm are accommodated for the reception and transmission of cw.

How quickly you become familiar with the software is dependent upon how much RTTY experience you've had and how well you know your computer. Don't expect to shove the disk into the drive, push a couple of keys and be on the air in five minutes! You've got to take some off-the-air time to become acquainted with the system operation.

Makeup

You probably can't think of more for an RTTY program to do than what is already contained within Super-RATT. There are three programs and a text file of importance: Super-RATT (the operating program); Super-RATT. UTL, used to maintain and edit the system directory, add message titles, create new indices and logbooks, etc.; Super-RATT. CONFIG, used to set up program options that meet with your particular system needs, and Super-RATT. EXEC, a file created by the previous program. This file is used to insert new information into several of the main program lines. In order to modify the main program to suit different system setups and modes, several such files may be kept on the disk. It all serves to make your operation flexible. You can have things configured so that you can use more than one TU.

An RBBS (Radio Bulletin Board System) is an integral part of the software. This subsystem has over 34 commands that a user can access. All message-handling functions and log-keeping routines are included.

If you have a 16 K RAM card installed or own an Apple //e, you have at your disposal 10 K of receive and 10 K of transmit text buffer. With the receive buffer you can elect to save off-the-air copy. At 50 bauds, an empty, initialized disk will hold over five hours of copy. The receive/transmit buffers may be independently erased and full-screen display of the transmit or receive buffer is operator selectable.

When transmitting, you can elect the fill ("diddle") character (letters, figures, blank or none), word- or character-mode transmission and carriage return position. The latter permits you to select automatic carriage return (CR) at 40- or 72-character intervals or to disable the CR function entirely.

The BREAK feature is a welcome part of the transmit section of the program. It allows you to respond to a question asked by the station being received or to answer a breaking station without losing text entered into the normal transmit buffer. If you've ever been on RTTY without such a feature, you can appreciate what this addition means! When you use the BREAK feature to answer a station, you will see the break transmission text appear on the screen immediately after the last transmit buffer information. Only the break transmission information will be output. After returning to receive, you can resume filling the transmit buffer. Although the screen shows the break transmission information lodged between the interrupted transmit buffer information, the break data will not be retransmitted.

Using QUICK LOAD files, you can store messages on disk for later transmission. It is a method of message storage that is conservative of machine memory space, as opposed to the use of stored message strings in RAM. File access is rapid.

Super-RATT offers an on-line LOGBOOK, and a feature called VERIFIED FILE TRANSFER. The latter is to be used with other Super-RATT-equipped stations. It is a method of file transmission that provides for two-way handshaking and checksum computation. Once contact has been established between the two stations and the process begun, the rest of the operation is fully automatic. The system is still under development, and the program developer welcomes input from Super-RATT users who will improve the process.

The software contains a copyrighted feature known as RATT-SOFT. It is a method of using the Apple ampersand jump routine for a number of different command functions. RATT-SOFT may be called by any Applesoft® program or directly from the keyboard. Over 30 different commands are at your beck and call.

The screen display is divided into five sections, if you count the top status line and the bottom command and status line. Below the top status line is a 13-line area reserved for incoming text. That is followed by a single-line "Times Square" scrolling area that displays the transmitted text as it is being output. This single-line area is separated from the sections above and below it by screen-wide dotted lines. Beneath the "Times Square" line is a four-line transmit-buffer-text area. Transmitted text (and received text) scroll once the bottom line of the respective area is reached. Both texts can be displayed using a full-screen format, if desired.

The bottom command/status area is the place to look for help when you can't remember which keys to push to accomplish your intended purpose. The upper status line tells you your mode and speed of operation, T-R status (whether you're in receive or transmit mode), the receive buffer status, and whether or not you're saving incoming text to disk. There are also relay status and mark/space indicators. The mark/space symbols (letters M and S) flicker with incoming and outgoing text, acting as tuning/status indicators. Relay status informs you when you are re-transmitting just-received text or repeating your last transmission.

The status indicators on the top line are enclosed in brackets. This makes the line look rather busy, as far as I'm concerned. I opted to delete the brackets, and I find the line easier to read. You can modify program lines 2660 to 2880 to suit your personal tastes.

CW Software

I am somewhat disappointed with the cw section of the program. The plus it has (inclusion of most of the prosigns) is outweighed by its inflexibility. The speed algorithm is good, but it can stand some improvement. At higher transmission speeds, the weighting becomes too light and I couldn't find a way to vary it easily.

The operator must key in a cw speed for transmission. This keyed-in speed also sets up the cw receiving speed. Because I (and others) had trouble getting the program to respond properly during receive, I was told to try setting the receive speed at a figure about twice that expected. It does help, but not that much and because the receiving and transmitting speeds are interlocked, if the transmit and receive speeds differ appreciably, you'll have to change speeds as you go between transmit and receive. Ideally, the cw-receiving algorithm should be self-adjusting and adapt itself independently to the speed of the incoming code.

I would like to be able to alter the weighting of the cw characters by simply using a couple of keystrokes. It would also be nice to be able to use the computer speaker for monitoring sent cw. This feature could also be turned on and off by simple commands. As it stands, you must rely on the transmitter cw-monitor signal when transmitting.

General Comments

As I understand it, earlier versions of Super-RATT were composed of several smaller programs. The present version is, essentially, one big program. I would like to see the big program broken down into a few "bite-sized" chunks. While this would lead to code duplication, it would make the system more appealing to some, I'm sure.

One nice part of Super-RATT is that the programmer has enabled the user to modify much of what is available according to his or her tastes. When making changes to the program, ensure you make them to the operating version (Super-RATT) and not to Super•RATT REMarked. The latter version of the program is filled with REM statements that help explain the program to the user.

By all means, do as the program author suggests: Make a copy of the program disk and stick the original disk in your archives! Don't take a chance on losing the original copy. As you make changes to your backup disk, back that disk up with yet another in case something "crashes." Otherwise, you'll have to do all that work over again.

If you intend to LOAD and LIST the REMarked version of Super-RATT, you'll probably have to set MAXFILES to 1 or 2. Without doing that, I got a PROGRAM TOO LARGE error indication.

After five transmit-receive toggles, the program automatically sends a cw i-d whether or not text has been transmitted. The program purchaser's call sign is buried within the bowels of the program.

Unfortunately, Super-RATT (at this time) does not support the 80-column display capabilities of the Apple IIe. Perhaps someday this will become part of Super-RATT. The program is under constant improvement and users are kept in touch with the program developer by means of a newsletter.

I've only touched on the highlights in this review. There are other program features that are sure to be of interest to some. If you're looking for a comprehensive RTTY program for your Apple computer, you may find it in Super-RATT. The package does have some weak areas, but I've not used a piece of RTTY/cw software yet that doesn't. I hope these areas will be strengthened in the near future. Super-RATT is available from Universal Software, Inc., 9 Shields Lane, Ridgefield, CT 06877. Price class: $60. — *Paul K. Pagel, N1FB*

AEA HOT ROD ANTENNA FOR 2-METER HAND-HELD TRANSCEIVERS

☐ Fm operators have seemingly fallen in love with gain antennas for 2-meter hand-held transceivers ("HTs"), if brisk sales and high visibility can be used to judge such things. There are good reasons for this — compared to the "rubber duckie," the gain antennas (when properly designed) provide as much increase in effective transmitter power as a small amplifier. The gain antennas show the same improvement on the receiving end of things, too.

Previous efforts in this area have been limited to models designed around a 5/8-λ radiator. AEA has broken with this tradition by designing the Hot Rod around a 1/2-λ radiator. Fully extended the Hot Rod measures 39-5/8 inches, and it weighs in at a little under two ounces.[2] This makes it shorter and lighter than any of the other (5/8-λ) gain antennas on the market. This should add up to a longer life for the Hot Rod and for the BNC connector on the radio.

Does it work? Fully extended, the antenna works like a charm. I observed the same general improvement in transmitted and received signal that I've seen with the 5/8-λ variety antennas.

Collapsed, the antenna measures 8½ inches long. Although some of the AEA promotional literature suggests using the antenna in the collapsed position, you may want to check the VSWR safety rating of your rig first. VSWR measurements across the 2-meter band for the collapsed mode are included in Table 1.

If you are in the market for a gain antenna for that new 2-meter rig, you might want to give serious consideration to the AEA Hot Rod. Price class: $25. Additional information can be obtained from AEA, P.O. Box C-2160, Lynwood, WA 98036. — *Peter O'Dell, KB1N*

Table 1
Frequency Vs. SWR for AEA Hot Rod

Frequency (MHz)	*Extended*	*Collapsed*
144.0	1.0	10.8
144.4	1.0	11.2
144.8	1.0	12.7
145.2	1.0	14.9
145.6	1.0	13.9
146.0	1.0	12.9
146.4	1.25	13.9
146.8	1.35	15.9
147.2	1.47	17.9
147.6	1.55	17.14
148.0	1.70	19.9

SWR values expressed are in relationship to unity. Measurements performed in ARRL lab with an IC-2AT, a Sola Basic Directional Wattmeter and a 5-W element.

HEATH HL-2200 AMPLIFIER KIT

☐ I'm sure many readers have at least a nodding acquaintance with the popular Heath SB-220 and SB-221 amplifiers.[3,4] Therefore, a look at the exterior of the HL2200 might prompt one to ask, "How different is the '2200?" The exterior is quite a bit different. A warm brown finish has replaced the familiar green cabinet color, and other facial changes, such as the use of different styles of knobs, meters, switches and a new logo, are obvious. A couple of the exterior changes reflect modifications that have been made to the control and SWR monitoring circuits, but for the most part, the physical interior and electrical designs are identical to those of the earlier amplifiers.

Kit Assembly (and Some Hints to Avoid Kinks)

It took me approximately 22 hours to assemble the '2200. A few additions and changes to the assembly manual from supplied errata sheets are required before work is started. I found one additional minor error on page 32: The color code for R29 is given as red-violet-orange when it should be yellow-violet-orange for the 47-kΩ resistor. One assembly procedure, the installation of the pilot lamp, caused me a bit of grief. I could not get the grain-of-wheat lamp and diffuser combination to fit properly in the body of the lens. After several attempts to mate the two units, one of the lamp leads broke off. I substituted a Radio Shack 272-1140 lamp for the original bulb, and the fit is perfect.

When separating the line-cord leads, don't attempt to make two cuts of the length required, as you're bound to nick the insulation on one of the wires. A much easier method is to make a small cut either side of the green center wire and simultaneously pull the two heavy conductors away from the center wire; they'll separate easily, and no insulation will be damaged.

If you're going to operate the amplifier from a 234-V line, you'll have to purchase the proper line-cord plug. The molded plug supplied is for use only on 117-V circuits.

When installing the high-voltage transformer, I'd recommend leaving the red-yellow lead uncut until the transformer has been set in place. Once you've determined that the lead will comfortably reach the required point of attachment, you can cut it to the proper length.

Controls and Metering

One welcome addition to the new amplifier is the addition of an AMPLIFIER IN/OUT switch that allows the HL-2200 to be placed in standby. A red pilot lamp (located immediately above the POWER ON/OFF switch) illuminates when power is applied to the amplifier; it replaces that function of the lighted meters on the older amplifiers. The '2200 meters are not illuminated types.

[2] g = oz × 28.35.
[3] Recent Equipment, *QST*, Aug. 1970, p. 45.
[4] Product Review, *QST*, March 1980, p. 43.

Two meters monitor the vital amplifier functions. The left-hand panel meter measures tube plate current in increments of 50 mA to a maximum value of 1 A. To the right of the PLATE CURRENT meter is the PLATE VOLTS meter, which actually is a multifunction indicator used in conjunction with the FUNCTION switch. In the HIGH VOLTAGE position, the meter reads dc plate voltage in 400-V increments to a maximum of 4 kV. Grid current, in 25-mA increments and with a scale maximum of 400 mA, is measured with the FUNCTION switch in the GRID CURRENT position. The RELATIVE LOW and RELATIVE HIGH switch positions select taps on a voltage divider/detector network attached to the RF OUTPUT jack. In either position, relative forward power is measured, the full-scale reading dependent on the tap selected. These switch positions replace the SENSITIVITY control function present on the SB-220/221 amplifiers.

Some Observations

To comply with current FCC regulations, the '2200 has no provisions for operating on 10 meters. (At the time this review is being written, Heath has no 10-meter modification kit available for the HL-2200). The 15-meter input tuned circuit now consists of three series-connected pi matching sections, and acts as the 27-MHz filter; the main-chassis-mounted 27-MHz filter is no longer used.

The large front-panel paddle switches don't fit tightly and exhibit a small amount of front-to-back movement. These switches are snap-in types, unlike those on the older amplifiers that were bolted to the chassis. But this should not pose any problem other than one of aesthetics.

A "plug-in" high-voltage transformer arrangement is now used. Late SB-221 models have this feature. Of the eight high-voltage transformer leads, six are wired to a Molex plug. Only two easily reached wires need be unsoldered, and the Molex plug and transformer mounting screws removed, to take out the transformer for amplifier shipment or transformer replacement or substitution. Should you need to remove the front panel for any reason, taking out four screws will free the assembly.

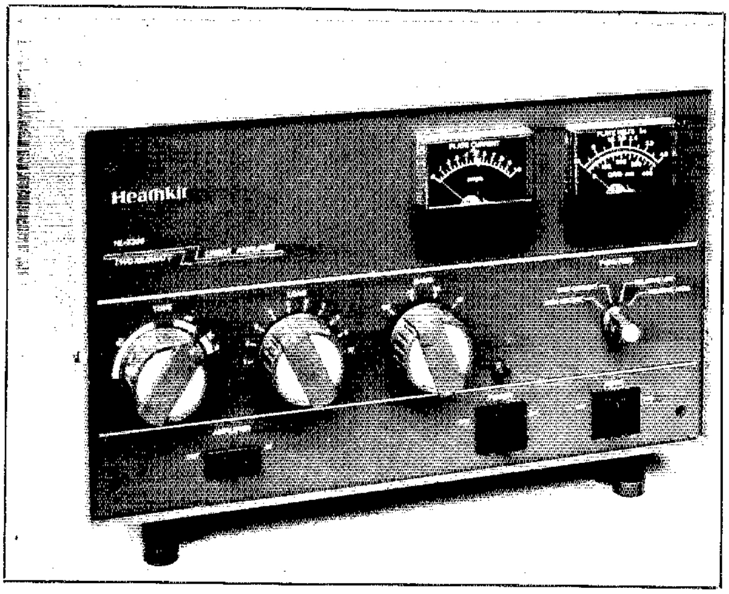
In the past, some concern had been expressed about filament inrush-current protection for the 3-500Z tubes.[5-8] I spoke with representatives of Heath and Eimac/Varian about this subject. I was told that filament inrush current poses no problem with this amplifier because of the filament transformer design. There is enough sag produced by the transformer during turn-on that the tube ratings are not exceeded. The more important consideration regarding the tubes is one of installation. When the tubes are being inserted into their sockets, they should be pushed straight down until seated properly. No rocking motion should be applied! Such action may crack the glass beads around the tube pins creating small air leaks that will prematurely destroy the tube(s). If the socket fingers are too tight to permit the tube to seat properly, bend the socket fingers slightly until the tube pins fit firmly, but don't rock the tubes!

Amplifier Testing

Results of the amplifier tests are shown in Table 2. No difficulties were encountered when the 10-minute key-down test was run.

The −30 dB third-order IMD performance of the HL-2200 is met easily. At present, the transceivers used in the ARRL lab to drive amplifiers such as the '2200 are the limiting factor when it comes to making this measurement. Because of this, new equipment is being ordered that will permit more-telling third-order IMD measurements to be made.

The HL-2200 looks like a perfect mate for the HW-5400 and SS-9000 transceivers, and should be a welcome addition to almost any amateur station. The HL-2200 is available from Heath Company, Benton Harbor, MI 49022. Price class: $850. — *Paul K. Pagel, N1FB*



Heath HL-2200 Linear Amplifier

Manufacturer's Claimed Specifications	*ARRL Lab Test Results*
Frequency coverage: 80, 40, 20 and 15 meters.	Confirmed.
Maximum power input: Ssb, 2-kW PEP; cw, 1 kW; RTTY, 1 kW.	Confirmed.
Maximum recommended key-down time at full power input 10 minutes.	Confirmed.
Third-order IMD (dB): −30 or better.	See text.
Power requirements: 117-V ac at 50/60 Hz. (20 A max.), or 234-V ac at 50/60 Hz. (10 A max.).	
Driving power required: 100 W max.	Confirmed.
Color: Medium brown.	
Size (HWD): 8-1/4 × 14-7/8 × 14-1/2 in.†	
Weight: 50 lb.	
Price class: $850.	
Manufacturer: Heath Company, Benton Harbor, MI 49022.	

†mm = in. × 25.4; kg = lb × 2.2.

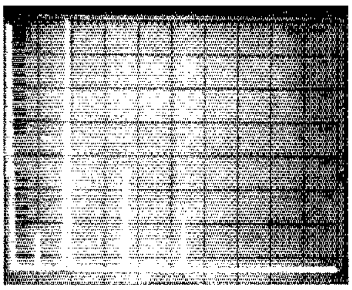


Fig. 5 — Worst-case spectral display of the HL-2200 amplifier. Vertical divisions are each 10 dB; horizontal divisions are each 2 MHz. Output power is approximately 680 W at a frequency of 3.6 MHz.

Table 2
HL-2200 HF Amplifier Performance

Band (meters)	*P_{in} (W)*	*P_{out} (W)*	*Input SWR*	*Drive Power (W)*	*Efficiency (%)*
80	1000	680	1.4:1	68	61
40	1000	720	1:1	70	65
20	1000	680	1.8:1	68	62
15	1000	680	1.9:1	100	59

[5]K. M. Gleszer, "Upgrading Your SB-220 Linear Amplifier," *QST*, Feb. 1979, p. 20; also, Feedback, *QST*, April 1979, p. 27.
[6]Hints and Kinks, "On Upgrading Your SB-220 Linear Amplifier," *QST*, Nov. 1979, p. 57.
[7]Hints and Kinks, "Comments On SB-220 Modification," *QST*, Feb. 1980, p. 44.
[8]Hints and Kinks, "Filament Inrush-Current Limiter For Linear Amplifiers," *QST*, Nov. 1982, p. 49.

Hints and Kinks

Conducted By Larry D. Wolfgang,* WA3VIL

USE YOUR 2-METER RIG FOR F2 OPERATION

□ Audio-tone-modulated telegraphy (F2) is used by most repeaters for the Morse code i-d. But even most "all-mode" 2-meter rigs are not equipped to transmit on this mode. These radios transmit cw using A1, which cannot be received without a BFO. The simple circuit described here will enable you to operate F2 while using any fm rig, without modification to the radio, and for a cost of about $15 if you buy all new parts.

FCC rules permit F2 operation anywhere on the 2-meter band, except for the lowest 100 kHz, which is reserved for A1 operation. You should avoid the repeater frequencies in your area, but this still allows plenty of room for cw operation. Small groups can use F2 on 2 meters as a means to brush up on their code skills. If one of the members has difficulty, he or she can pick up the mike and break in. This is something you can't always do while operating cw on hf.

The principle of my circuit is quite simple. An audio oscillator that will drive a speaker is needed. You can use an old code-practice oscillator, or use the built-in side-tone oscillator from your electronic keyer. The audio output can be obtained from the speaker or headphone jack in that case. Fig. 1 shows the complete schematic diagram of the unit. As shown, S2 supplies power to the oscillator module, but some oscillators may not require this switch. J1 mounts on the project case, and is a type to mate with the plug on your mike. P1 mates with the mike connector on the rig. Allow about 18 inches of cable between the radio and the F2 unit.[1] The input can be from a keyer or straight key, and you will have to provide appropriate connectors. Fig. 2 illustrates my final arrangement.

You will have to adjust the value of R1 to provide the proper deviation for your transceiver. My circuit needed a 1.2-kΩ resistor, so that may be a good starting value. Connect the circuit to your radio, set S1 to the PTT position and turn on the transceiver. Normal voice operation should be possible at this point. Place S1 in the cw position. Using a deviation meter, or another amateur listening on a separate receiver, adjust R2 for a deviation of about 5 kHz. If this control does not provide enough adjustment, then you will have to change the value of R1.

A word of caution: The key-down time for some rigs is limited by heating of the final amplifier transistors. Since the transmitter is on continuously during your transmission, you should check the heat sink periodically to be certain that proper cooling is provided. You may have to limit the length of your transmissions accordingly.

Gather a few hams into a cw net, practice

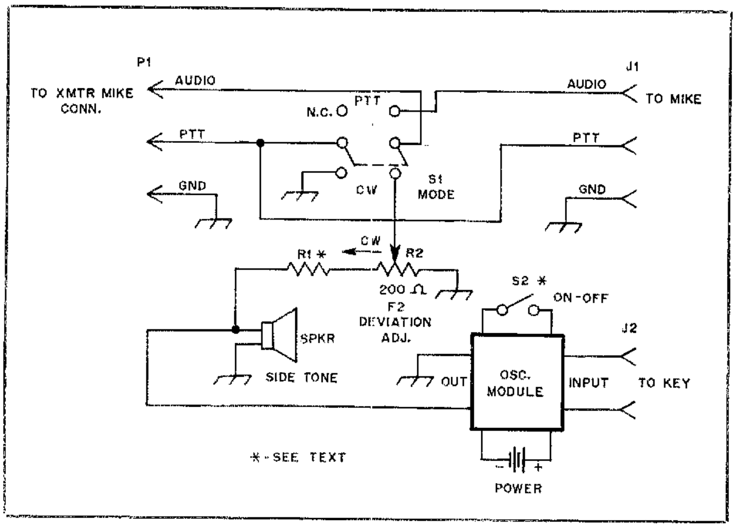


Fig. 1 — Schematic diagram of an oscillator circuit used by KA7AWD to add F2 capability to his 2-meter fm rig.

Fig. 2 — Photo showing the equipment used by KA7AWD for F2 operation. The oscillator unit is housed in the box on top of the electronic keyer.

[1]mm = in. × 25.4.
*Assistant Technical Editor

every night, and before long your code speed will begin to soar. But remember, if you get into a real bind and can't copy as fast as the others are sending, you can always pick up your microphone! — *Dennis Blum, KA7AWD, Show Low, Arizona*

SPACERS FOR MOUNTING PC BOARDS

☐ I needed some spacers to mount the pc board from a project inside a case. After looking around the shack for something to use, I came across an old Bic® pen. I cut lengths of the plastic tube with a tubing cutter and had perfect spacers for mounting my project board. Now, I save all of these pens when the ink is used up. By cutting spacers to almost any desired length, I can stack boards on top of each other or stand them side by side. A hacksaw should also work for cutting pieces of the tubing. — *Walter Yatzook, N1CJB, South Meriden, Connecticut*

KEYING THE IC-701 WITH AN FET

☐ While planning to build a new keyer for my station, I became concerned that my chosen design would not properly key my IC-701.[2] This radio (and some others) requires a key-down potential of less than 0.4 V. The output circuit of the Digital CMOS Iambic Keyer uses two silicon devices in series (a 92PU10 transistor and a 1N4004 diode), and I was concerned that this would prevent the minimum key-down voltage from being small enough. Also, I could not find the 92PU10 devices in any of the local supply stores.

I decided to use a 7404 hex inverting buffer IC between the keyer and my rig, but a friend suggested that I try a VN67AF VMOS power FET in the keyer output section instead. One manufacturer of these transistors is Siliconix Incorporated. With this transistor in the output section of my keyer, the key-down voltage is less than 0.1 V. Fig. 3 shows how I modified the keyer output circuit to suit my needs. I have also made a few turns around some ferrite toroids with the leads to my paddle and to the IC-701. This has cured a problem that I had with rf getting into the keyer. — *Darwin Bingham, N4HZC, Virginia Beach, Virginia*

[2]T. Theroux, "A Digital CMOS Iambic Keyer," *QST*, June 1982, pp. 26-28.

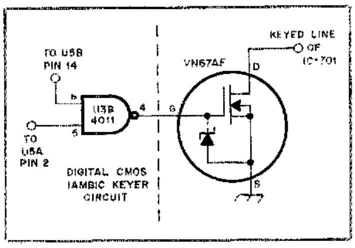


Fig. 3 — Schematic diagram of a keying circuit, used by N4HZC, with an electronic keyer and an IC-701. Some VN67AF transistors contain the built-in Zener diode to prevent overdrive, as shown here, while others do not include the diode.

A "NAVY" KNOB FOR STRAIGHT KEYS

☐ I have found that a straight key is much easier to use if it is equipped with a "Navy" style knob. This style of knob does not seem to be very common anymore. I found a way to add this style of knob to most straight keys. A small disc, cut as shown in Fig. 4, is all it takes. The material that I use is Formica®, available as a color sample or scrap from many building-supply stores. Shaping can be done with a diagonal cutters or tin snips, followed by a file or sandpaper to smooth off the rough edges. You may have to adjust the dimensions of the disc or the cut out to make it fit your key. Try this simple addition to your straight key. You might like it! — *J. A. Wright, Jr., W4UEB, Edenton, North Carolina*

SILICONE SEALER FOR MOUNTING PARTS

☐ I would like to pass along an idea that I have been using to hold batteries, large capacitors and other items in place. Simply spread a thin layer of silicone sealer on the part to be mounted, set it in place and let the sealer dry for about a half hour. Rub away any excess material with your fingers or a pencil eraser. This sealer makes a good bond to almost any clean surface, and it maintains its flexibility and strength under most conditions. If you use this material as a battery holder, you can cut or break the battery loose when it goes dead.

I have found this to be an ideal method of securing large parts to a chassis without drilling mounting holes. Occasionally you may need to isolate a capacitor or other component from ground. The silicone sealer will also serve this purpose well. Try it, and you'll probably discover even more uses for this inexpensive material. — *Hellmuth Hinz, KA8RNN, Athens, Ohio*

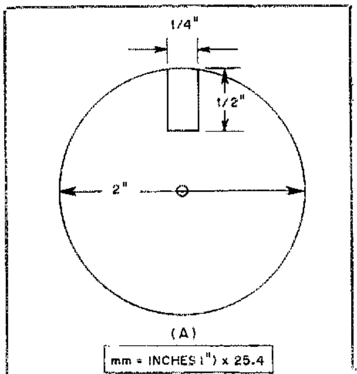
DESOLDERING HINT

☐ Most hams occasionally want to salvage a used PL-259 connector, or are faced with another difficult desoldering job. Here is a method that I have found to be quite helpful. When the unit is heated enough to melt the solder, try a blast of compressed air at about 80 lb/in^2 to blow the solder out of the joint. I have used this technique to solve a number of tasks that would have been difficult otherwise.

Several precautions must be observed in using this method, however. Be sure to wear eye protection at all times. You must also be certain that the molten solder will not hit anything that can be damaged by heat, such as a carpet or synthetic materials. Don't allow other persons to be in the area where solder may hit them, either! — *Lynn Burlingame, N7CFO, Bellevue, Washington*

SIMPLE TUNING-DIAL DRIVE REDUCTION

☐ I have a Kenwood TS-130S that I use for mobile operation. The VFO tuning is too fast for easy operation in a vehicle. I built a drive-



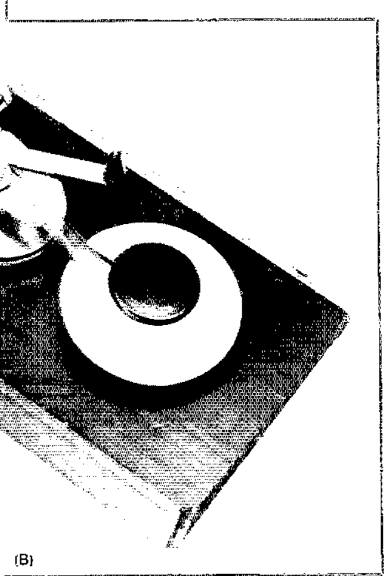


Fig. 4 — The dimensions used by W4UEB to cut a thin disc of material to form a "Navy" style knob are shown at A. At B, the disc is shown in place between the knob and arm of a straight key.

reduction adapter (for less than $2) to cure this problem. I used a ¼-inch phone plug, an enclosed jack and an angle bracket.* The bracket can be aluminum or steel, 1½ × 1 × ¾-1 inch wide.

Drill holes in the bracket as indicated in Fig. 5. The TS-130 has a screw 1½ inches from the front of the cabinet top that I used to mount the L bracket. If you are adding this drive-reduction unit to another rig, you may have to drill a hole in the cabinet to mount the bracket. I used a small piece of double-sided tape under the front of the bracket. Mount the phone jack to the angle bracket and insert the plug. Now stretch a wide rubber band over the phone-plug barrel and around the tuning knob.

You can use the phone plug for a tuning knob, or if you want a larger diameter knob, solder a short piece of ¼-inch-diameter copper tubing to the phone-plug terminals. Replace the barrel on the plug and secure a larger control knob to the extension.

The tuning rate is much slower now, but if I want to make a large frequency change I can still use the main tuning knob. — *Larry Dougherty, KN8I, Yale, Michigan*

FASTER AGC RELEASE TIME FOR THE TEN-TEC ARGOSY

☐ I have found my Ten-Tec Argosy to be a fine all-around rig. I especially like the full QSK feature. However, I believe the agc release time is too slow for cw operation. I prefer the agc to release a bit faster, even for phone work. I changed a single capacitor in the agc circuit to provide a release time more to my liking.

Remove the top cover and locate the I-F/AF circuit board. It is the large board just behind the front panel. Find C12, a 33-μF capacitor at the center rear part of the board. This capacitor is connected from the base of Q4 to ground. The rate of discharge for this capacitor controls the agc circuit release time.

Unplug all cables going to the I-F/AF board, remove the five screws that fasten it to the chassis and lift the board out. Remove C12 and replace it with a 5- to 10-μF capacitor rated at 16 V or higher. I prefer the 5-μF unit for a faster release time, but you may want to experiment. Reinstall the board and connect the cables.

I find the cw operation of my rig to be much smoother after this change. I also notice that I am able to copy more weak stations than before because strong nearby signals don't limit the receiver gain as much. — *Bill Scott, AG0O, Elizabeth, Colorado*

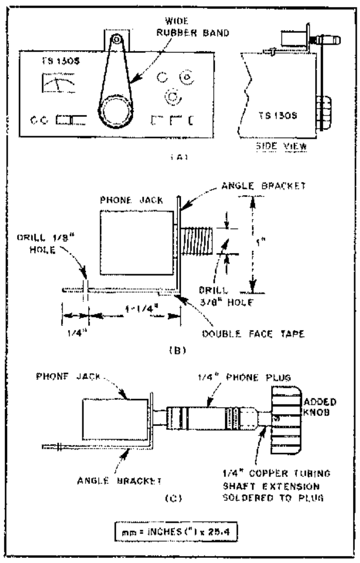


Fig. 5 — At A are shown the mounting details for a tuning-dial-drive reduction system used by KN8I. The dimensions for the mounting bracket are given at B. One method of adding a larger-diameter tuning knob is shown at C.

2-METER AMPLIFIER INSTABILITY

☐ I built a solid-state amplifier to boost the 1.5-W output from my 2-meter hand-held transceiver to about 30 W. The amplifier suffered occasional instability problems, depending on the lengths and positions of the rf input and output cables. The measured SWR, both into and out of the amplifier, was less than 1.5:1.

Some research led me to the conclusion that my problems were caused by rf energy being conducted on the outside of the RG-58/U coaxial-cable braid. The ground connections for the transceiver, amplifier and amplifier power supply were separated from each other by significant fractions of a 2-meter wavelength. There was no good ground reference for the rf energy. The chassis of each unit could "float" above rf ground, inducing rf currents to flow along the outside of the coaxial-cable shield braids. My amplifier was affected in unpredictable ways by these currents.

I found a simple cure for the problem. I looped each rf cable twice through an Amidon FT-114-43 ferrite torroid. My calculations indicate that this effectively adds about 2.4 μH of inductance in series with the outside of each braid. The result is an rf choke with over 2200-ohms reactance at 2 meters. I chose type-43 material for high loss above 100 MHz, so the toroid dissipates rf energy like a ferrite bead does. This combination of effects has completely cured the instability problems with my amplifier, regardless of cable length or orientation. — *Kenneth H. Kerwin, II, K6UXO, Londonderry, New Hampshire*

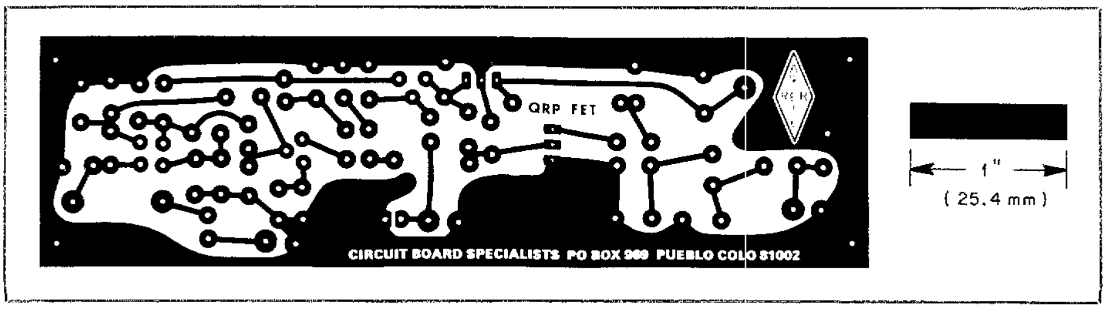
USING RUB-ON ETCH-RESIST TRANSFERS

☐ Here are a few suggestions for those who have had problems with rub-on etch-resist transfers coming loose. First, the board must be very clean. Then, warm the board before applying the transfers. I warm small boards by placing them on the shade of my desk lamp for a few minutes. Apply the transfers to the warm board, and burnish them well. Rewarm the board if it cools off during circuit layout. After all of the circuit is laid out, warm the board for several hours and then burnish all of the transfers again. Using a small piece of paper towel, gently wipe the board to remove any fingerprints or other dirt.

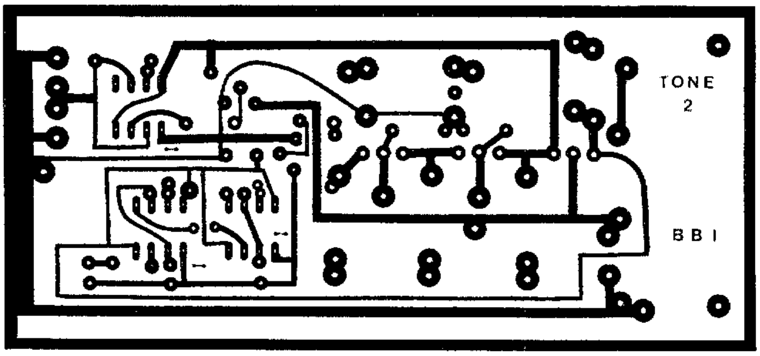
I have not had any problems with transfers coming off in the etchant since I began using this method. The warming technique also works well for applying panel-marking rub-on labels. — *J. T. Miller, N6BM, Ukiah, California*

A TUBING CUTTER AND COAXIAL CABLE

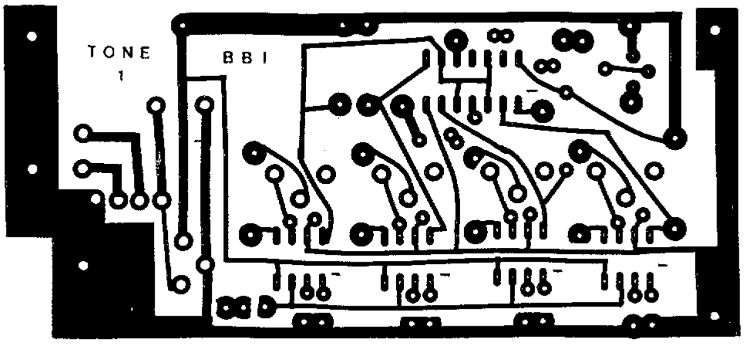
☐ I read with interest the hint by Kirk Carter in the September 1982 Hints and Kinks column about using a tubing cutter to prepare RG-8/U coaxial cable for a PL-259 connector. I have used a method like this for many years, and find that it works well. But I would add a word of caution. Before you use the tubing cutter to cut the tinned braid, be sure the cable has cooled completely! If you try to cut it while the cable is still warm, the inner dielectric will be soft. The cutter will distort the cable and mess up the job. — *George Rulffs, AA4GR, Chelmsford, Massachusetts*



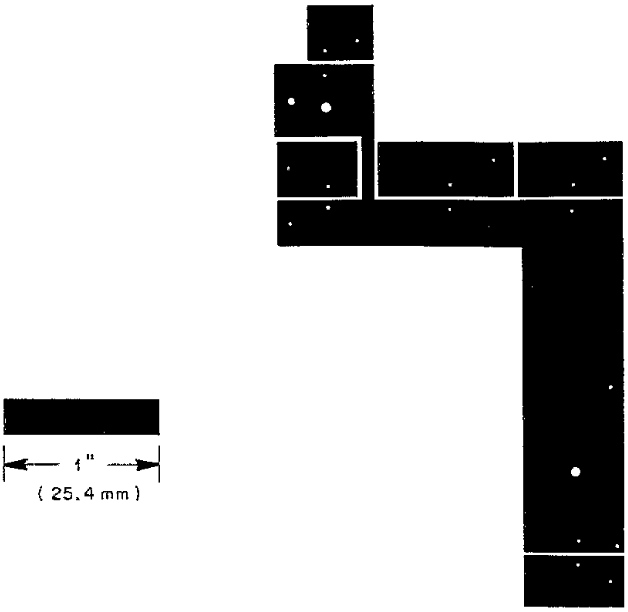
Circuit-board etching pattern for the VXO cw 30-meter exciter. The parts-placement diagram appears on page 33, this issue. Black areas represent unetched copper.



(A)

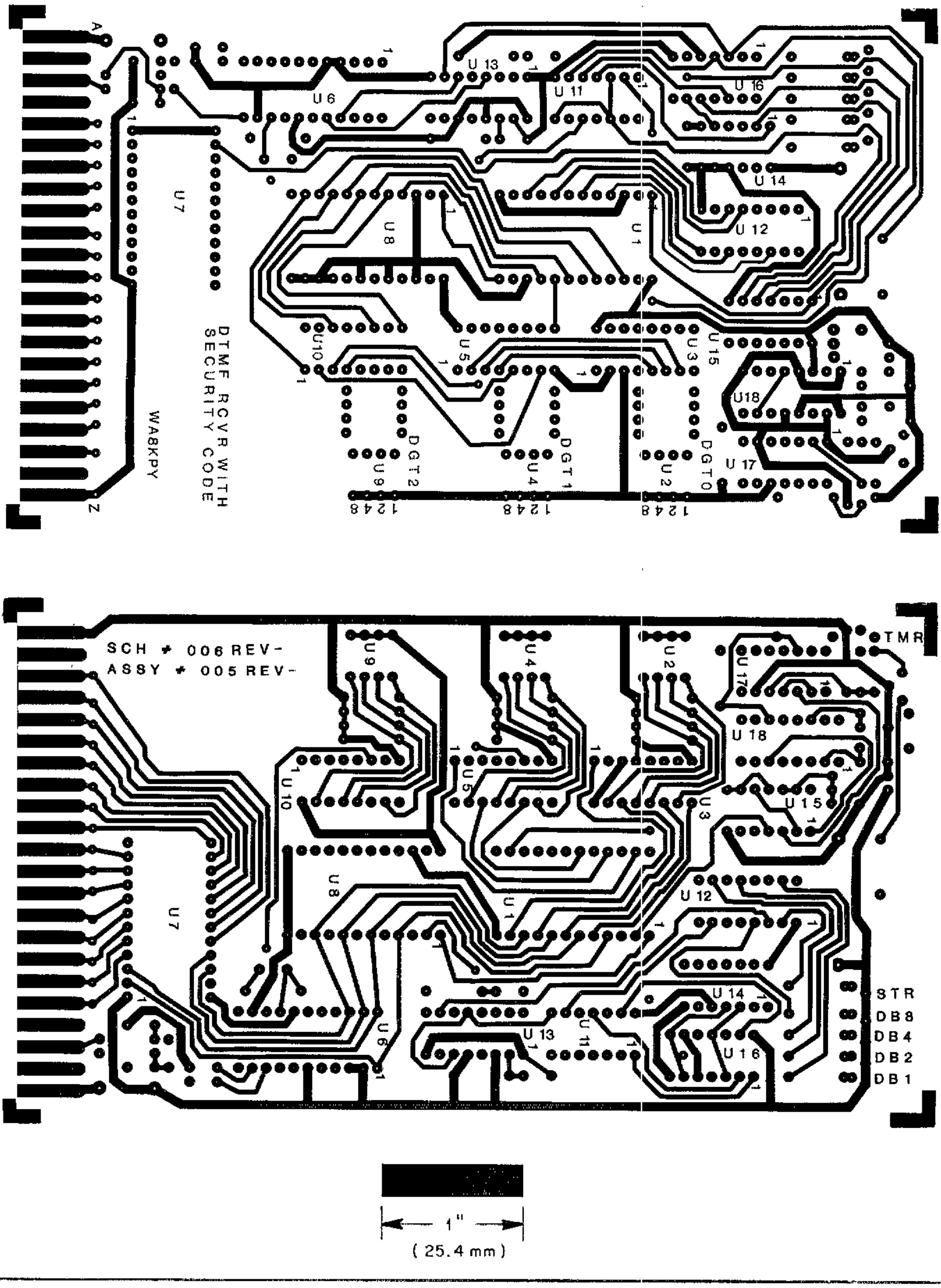


(B)



(C)

Circuit-board etching patterns for the London Tone Alert see the parts-placement diagrams of Fig. 3, on page 38 of this issue. The patterns are shown actual size. Black represents unetched copper.



Circuit-board etching patterns for the DTMF decoder board (see the parts-placement diagrams of Fig. 7, on page 26 of this issue). The patterns are shown shown actual size. Black represents unetched copper.

Technical Correspondence

Conducted By
Dennis J. Lusis,* W1LJ

The publishers of QST assume no responsibility for statements made herein by correspondents.

NØAJY cb STANDARD

☐ There are a few major problems with "The NØAJY cb Standard" (Aug. 1983 *QST*) that make it impossible to expect ". . . the ultimate in frequency-counter accuracy." The problems are not with the circuit itself, but rather with the network-affiliated television stations. Almost all of them are now using digital frame synchronizers on some, or all, network shows. These synchronizers convert network video from analog to digital form, delay it by a fraction of a frame (less than 1/30 of a second) and convert it back to analog form — in sync with the local station sync generator. Local color subcarrier frequency is required by FCC rules to be within 10 Hz of 3.579545 MHz, and in most cases is generated by a crystal oscillator no better than one found in a good frequency counter. In most stations, the color subcarrier frequency is rarely adjusted or calibrated. There are a few "major market" TV stations where the color subcarrier (and hence the sync generator) is locked to atomic standards, but that is quite rare. Additionally, many network television shows are delayed on videotape with no indication (to the viewer) of the delay. For example, in Dallas, the "CBS Late Movie" is a two- and one-half hour program that is tape-delayed 30 minutes (not an easy task)! When the videotape is played back, it is in sync with the local station, not with the network. Furthermore, all network atomic standards operate with an offset of −0.03 parts per million from the ideal 5 MHz. The 5-MHz signal is multiplied by 63/88 in a synthesizer to get a color subcarrier frequency of 3.5795454545 . . . MHz.

Anyone who is interested in achieving "the ultimate in frequency-counter accuracy" should probably leave the frequency counter on continuous standby, rather than wait for it to settle down after first being turned on. More accurate calibration of frequency counters can be achieved by phase locking an oscillator to WWVB on 60 kHz, or by simply counting the WWVB carrier frequency. Several manufacturers offer clocks and oscillators that phase lock to WWVB, but most of them are priced beyond the typical Amateur Radio budget. A few good articles have been written on the subject of calibration via WWVB.[1,2,3] Perhaps the best book available on this subject is the *Time and Frequency Users' Manual*, published by the National Bureau of Standards.[4]

WWVB is the primary standard-frequency broadcast service in the United States, whereas the network television color subcarrier comparison system is no longer viable.[5] Therefore, the horizontal sweep-frequency system in "The NØAJY cb Standard" is obsolete. It should be noted, however, that the NØAJY circuit is a clever one and it would have worked fine only 10 years ago. — *Andrew K. Dart, AE5D, Duncanville, TX 75116*

[1]D. Lancaster, "Experiment with WWVB," *Radio Electronics*, Aug. 1973.
[2]H. Isenring, "WWVB Signal Processor," *Ham Radio*, March 1976.
[3]E. Manly, "WWVB Frequency Comparator Receiver," *73*, Sept. 1972.
[4]G. Kamas and S. Howe, *Time and Frequency Users' Manual*, NBS Special Publication 559, U.S. Department of Commerce, Washington, DC, Nov. 1979.
[5]"NBS Time and Frequency Bulletin, Number 304," U.S. Department of Commerce, Washington, DC, March 1983.

*Assistant Technical Editor

TRUE ANTENNA HEIGHT

☐ "Getting the Most Out of Your Antenna" (July 1983 *QST*), by Chuck Hutchinson, K8CH, is a good article, but it omitted one aspect of antenna height that many of us on the West Coast take advantage of. To get a low vertical-takeoff angle, antenna height is not necessarily considered as the distance above ground, *but the distance above the rf reflecting surface*. This surface may be quite a horizontal distance from the antenna (see Fig. 5 in the K8CH article).

Many of us have stations on hills overlooking large bodies of water (Seattle/Puget Sound; San Francisco-Oakland/S.F. Bay; Los Angeles/Pacific Ocean). My station is about 250 feet above the surface of Puget Sound — about 3½ wavelengths high on 20 meters.[6] I get exceptionally good signal reports from DX stations in those directions my otherwise mediocre antenna "sees" Puget Sound. My antenna is only about 40 feet above the earth! — *Jack Wichels, W7YF, Edmonds, Washington*

[6]mm = in. × 25.4; m = ft × 0.3048.

NOTES ON SPREAD SPECTRUM

☐ The article by William Sabin, WØIYH, on spread-spectrum techniques (July 1983 *QST*) raised two questions about implementing spread-spectrum communications. First, except for the intrinsic charm of using a pseudo-random hop sequence, it seems that only privacy is gained by using this technique. If interfering signals occur randomly on the hop frequencies, there seems to be no additional signal-to-noise ratio gain from use of a pseudo-random sequence rather than a sequential-hop sequence. Because of limited spectrum allocations in the hf bands, a long random sequence must map into a small number of available frequencies. As an alternative hopping strategy, I suggest that the frequencies chosen by the operators be visited sequentially, rather than randomly. This technique yields advantages of simpler equipment and minimal synchronization problems.

Second, following Mr. Sabin's suggestion of using a-m rather than ssb, there should be consideration given to heterodyne problems. An a-m carrier will mix with any interfering carrier and produce audio beat notes, which can be different for each hop frequency occupied by an interfering signal. This is certain to produce annoyances. One possible solution is to use ssb, with the carrier reinserted at a level that allows phase locking at the receiver. Any phase-locked loop (PLL) frequency synthesizer suitable for transmission could then be used on reception. The acquisition range of the PLL can be restricted to prevent accidental lock to an interfering carrier. A loss-of-lock indicator used to signal a move to the next hop frequency could simplify the sync problem. For initial sync, the out-of-lock condition could allow fast jumps through the hop frequencies until lock is achieved. Once in lock, the hop rate would be controlled by the active transmitter. — *Paul Selwa, N9CZK, Brownsburg, Indiana*

☐ In response to Mr. Selwa's letter, consider 50 frequencies, sufficiently separated so that there will be no adjacent-channel interference. Fifty groups of stations, each group with two or more stations, could share these frequencies with no interference between groups. All of the groups would be part of a packet-network system. Protocol would include time and frequency management, and various control terminals would have rapid signal deployment capabilities for both domains. As an extension of this idea, consider a group of 250 frequencies. Using frequency-hop, each group could search for a clear operating frequency. When such a frequency is found, the group could elect to stay there until conditions deteriorate. Some members would want to hop on a set of frequencies in order to simultaneously address several groups.

The frequency hopping discussed above is done under protocol control. In addition, frequency hopping of the type described in my article could be used to provide increased reliability of communication during fading or interference. That is, some group(s) could hop on a set of the available frequencies mentioned above. (This hopping could be done sequentially, as Mr. Selwa suggests, rather than randomly.) If 50 frequencies were reserved for fast hopping, 50 groups could hop on these with no mutual interference when using protocol frequency management. The main advantage of non-sequential hopping is that errors from certain patterns of fading and interference (called burst errors) are more easily avoided. These hop sequences are easy to generate and supervise.

My article *does* mention ssb in a slow-hop system. Nbfm should also be considered. "Ancient modulation" is not a very attractive mode, I agree. Primarily, I would like to see an amateur data-packet network using spread spectrum. — *William E. Sabin, WØIYH, Cedar Rapids, Iowa*

ANTENNA PRUNING FOR 30 METERS

☐ I recently wasted a good bit of time because I didn't think before constructing a new antenna. It is easy to make incorrect assumptions concerning the new 30-meter band that are based on experience with "traditional" hf bands.

I was using a dipole for 80 and 40 meters with separate quarter-wave radiators fed with a single coaxial feed line. Wanting to get on the 30-meter band quickly, I decided to prune the 40-meter dipole to resonance on 10.1 MHz. This idea seemed great until I tried to load the antenna; the SWR was terrible. After a little thought, I realized that the 80-meter dipole (cut for cw at 3.525 MHz) was acting as a 3/2-wave dipole resonant on 10.575 MHz (three times 3.525). Apparently, the input impedance of the 80-meter dipole was not high enough to be insignificant at 10.1 MHz. Because I had two 30-meter antennas in parallel, the combined input impedance was no longer 50 ohms. To correct the problem, I left the 80-meter dipole alone, and trimmed approximately one foot from each end of the 30-meter dipole. This gave a combined input impedance that my solid-state rig could tolerate, but I know that the system is now way off resonance at 10.1 MHz. In short, I wouldn't recommend this combination!

By the way, the 80-meter dipole wouldn't load up worth a darn by itself on 30 meters. Progress is *so* frustrating! — *Dave Christie, WB5KFP, Bartlesville, Oklahoma*

SELECTIVE HEADPHONE FREQUENCY RESPONSE

☐ WB2KJL's letter (Technical Correspondence, Aug. 1983) raised the idea of headphones with response tailored for use by radio amateurs. Headphones of this type are available. Military surplus headsets H-140, H-227 and H-251 have tailored frequency response in the 300- to 3500-Hz range. — *David Wiesen, K2VX, Newark, New Jersey*

DIGITAL AUDIO GENERATION

☐ The *Wall Street Journal*, April 29, discussed digital audio, which is now entering the recording industry. This particular technique might benefit Amateur Radio. A reduction in voice bandwidth would effectively release many frequencies for use. A transmitter modulated by a digital representation of speech could reduce the modulated-signal bandwidth to a degree not possible with conventional methods. An ssb bandwidth of 250 Hz might be possible with sharp-cutoff filtering. Reception with a computer-enhanced and filtered system would be necessary. We have come a long way in the last 50 years; could this be in our future? — *Russ Mercer, W8JST, Union City, Michigan*

POWER INTERLOCK SAFETY

☐ I've noticed that several ARRL publications, including the *License Manual*, recommend a power-safety interlock that consists of wiring a female ac receptacle in series with the hot side of the station power line. This safety feature is suggested to ensure power isolation and prevent tampering by children and unauthorized operators who are not likely to spot or suspect the empty receptacle.

I haven't yet heard of any disasters, but I worry about the child who finds the shorted male plug and tries to install it in a live power receptacle. What happens if an unsuspecting person plugs an appliance into the interlock, energizing the station equipment and appliance?

It seems that standard U.L.-listed electrical devices should not be "jerry-rigged," whatever the intention. Please, use only approved power-isolation breakers. The life you save could be *your* child's. — *Bruce Chadbourne, KA2PAS, Vestal, New York*

CAGE ANTENNAS

☐ Jerry Hall's article, "The Search for a Simple, Broadband 80-Meter Dipole" (April 1983 *QST*), is a good one. He does, however, leave the reader "hanging" with respect to the cage antenna. My following comments are based on experience as a cage designer (bet you didn't think there was any such animal!) and on feedback I've received from various installations.

Looking at the cage, we can see that there are three variables affecting performance:

1) Size of the conductor(s).

2) Diameter or width of the array (a cage does not have to be round, but a cylindrical configuration leads to fewer mechanical problems).

3) Total length of the array.

The size of the conductors is additive. I have found the following to be good limits:

40 meters — minimum of four conductors, no. 12 to no. 18 wire.

80 meters — minimum of six conductors, no. 12 to no. 14 wire.

160 meters — minimum of 12 conductors, no. 14 to no. 16 wire.

Spacing the wire elements of a cage can pose a problem. Hall suggests that nonconductive and conductive spreaders will provide the same results. I've found that conductive spreaders make the two sides of the dipole act like the plates of a capacitor. The antenna Q goes way up, and the bandwidth gets very narrow. As a strict rule of thumb, do not use conductive spreaders! Some experiments use waxed dowels. I originallly used Frisbees® (Whamo, Inc.). They provided the necessary spacing, but were not strong enough for the winter climate. When they froze, they cracked. I bought a plastic mold, and now make my own ABS spreaders. For 80- and 40-meter antennas, the spreaders I use are less than 10 inches in diameter. On 160-meter antennas, I use 22.5-in. circular PVC plates. I have also found that there is *no significant difference* in antenna characteristics within the following range of cage diameters:

40 meters — 6 in. to 10 in.

80 meters — 9 in. to 20 in.

160 meters — 19 in. to 36 in.

The total length of a cage array will be shorter than that of a single-wire antenna. Basically, a cage is a "fat" dipole that takes advantage of a lower length-to-diameter ratio. As Hall says, there are no tables on this subject. However, Fig. 2-4 of The *ARRL Antenna Book* can be used, but does not go below a K factor of 0.90; true cage antennas have K factors below this. For 80- and 40-meter versions, I suggest a K factor of 0.885; for 160 meters use 0.832.

Mechanical problems also manifest themselves when a cage antenna gets long. Hall suggested using a spreader every couple of feet, but with good spreaders you can obtain 19-ft spacing on a 40-meter cage. Up to 29-ft spacing is possible on 80-meter cages. On 160 meters, my tests are not complete enough to finalize a good spacing, but I expect it to be approximately 35 ft.

The free-space feedpoint impedance of the antennas in this discussion is 67.8 ohms — a good match to popular coaxial cables. This directly corresponds to Hall's computer model. For a cage designed for 3750 kHz, SWR at resonance is 1.09:1; the 2.0:1 SWR points are 3200 kHz and 4300 kHz!

I hope that I have helped in the continuing search for a broadband dipole. Any questions? I will be more than glad to answer them if the need arises. — *David C. Johnson, W2BN, Burlington, New Jersey*

Feedback

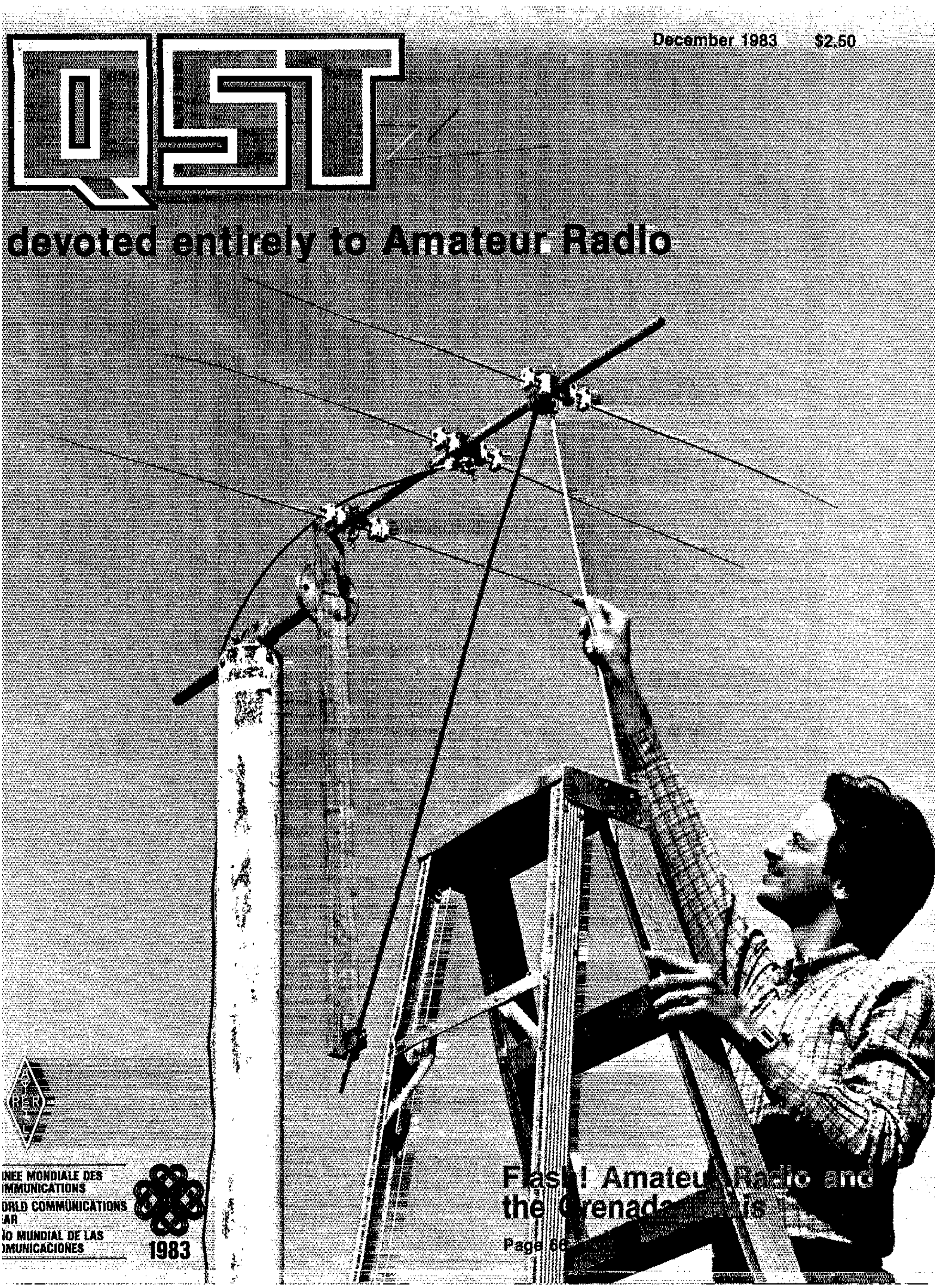
☐ Author Zander has some corrections for "Build the AA6PZ Power Charger" (Dec. 1982 *QST*). Eq. 1 requires that the current be expressed in amperes (0.7 represents 700 mA), rather than milliamperes. In Fig. 8, the component labeled R1 (at the upper right) is actually R5. The author has a list of "bugs" and application hints for the Power Charger that he will send to constructors for an s.a.s.e.

☐ Please note these corrections to the Viewstar VS 1500A Transmatch Product Review (Oct. 1982 *QST*): The price is $390, and the unit is available from the Ham Radio Outlet Stores in California, tel. 800-854-6046.

☐ There are two errors in "WARC Bands for the TS-820(S)" (Feb. 1983 *QST*). On page 41, the third line of the second paragraph under 18-MHz Modifications should read: "On the PD board . . ." On Page 42, the third line of the first paragraph under 24-MHz Modifications should read: "35-MHz crystal X10 . . ."

Strays

On August 12-14, members of the USAF MARS Los Angeles Base Support Team conducted their annual emergency communications project by operating from a century-old gold mine near Johannesburg, California. During the exercise, the amateurs contacted members of the Johannesburg Chapter of the South Africa Radio League who also were stationed in a gold mine. Among the operators who participated were (l-r) ZS6BFS/G4LRF (visiting from South Africa), KK6L and W6ULZ. (*WA6NKL photo*)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OUR COVER
Increasing the gain of a Yagi antenna is "element-ary," according to the authors of the article beginning on page 35. Check it out for yourself. *(photo courtesy Electrospace Systems, Inc.)*

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● *Basic Amateur Radio*

HF Propagation: The Basics

Say it's 10 P.M. in Savannah, and you'd like to reach out and QSO someone in southern Europe. A solid knowledge of how signals travel will help you decide if you've got a fighting chance.

By Dennis J. Lusis,* W1LJ/DL

Perhaps no other topic in Amateur Radio is as intriguing, yet confounding to the beginner as propagation — how signals travel from one station to another. Regardless of how you view it, propagation is essential to each and every QSO we make, be it a stateside ragchew or a rare DX contact. Some basic knowledge of how radio waves travel will go a long way in making your hobby a more interesting and enjoyable one. This article will introduce you to the primary modes of hf propagation. Vhf and uhf propagation is an entirely different subject.

Ground and Sky Waves

Regardless of what type of antenna you are using, the radio waves emanating from it can be categorized broadly into two types — *ground waves* and *sky waves.* Simply stated, a ground wave is one that travels directly from the transmitter to the receiver without leaving the lower atmosphere (Fig. 1). Ground-wave propagation occurs, for example, when you contact another station across town on an hf band. Amateur communications of up to 50 miles are typically possible via ground wave.[1] It is the sky wave that provides amateurs with the potential for worldwide communications.

A sky wave, just as the name implies, is one that does *not* follow the earth's surface; it travels up into the sky, away from the earth (Fig. 1). At this point, you are probably wondering how we can communicate via signals that travel out into space? After all, our receivers are down here on earth! Somehow, the signals must return here to be captured by our receiving antennas — plain and simple. And just what "persuades" our signals to come back down? Fortunately, there is a region in our upper atmosphere that is pretty good (and occasionally not so good!) at performing this task. This region is named the *ionosphere,* and it is here that we must look to understand the basic mechanisms of hf propagation.

Ionospheric Characteristics

The ionosphere derives its name from the term *ion,* which is a free electron or other charged particle. In our atmosphere, ionization (or the charging of particles) occurs in the region that lies roughly between 25 and 250 miles above the earth. In this region, air pressure is low enough so that ions can travel freely for a considerable length of time without colliding and recombining into neutral atoms. When radio waves enter the ionosphere, their courses are altered by the process of *refraction,* or bending (Fig. 2). Under proper conditions, the wave is diverted enough to head back down to earth, where it can be received.

The primary cause of atmospheric ionization is ultraviolet radiation from the sun. Therefore, solar conditions are of great importance to propagation. Exactly which solar indicators concern us will be discussed later in this article.

An Ionic "Layer Cake"

A closer look at the ionosphere reveals that it consists of dense layers stacked one on top of another, concentric with the earth's curvature. Each of the layers has maximum density in the center, with regions of gradually deteriorating density extending out toward the edges (Fig. 3). However, the absolute level of ionization is changing constantly with the time of day, season, solar conditions and other long-term variations. These variables contribute directly to the constant changes in hf propagation that often frustrate the most seasoned operators.

Each Layer Is Different

Although each ionospheric layer is comprised of free ions, their similarity (at least for our purposes) ends there. Each layer of the "cake" has special characteristics of its own, and you may be surprised to find out what effect each has on propagation.

The bottommost ionized region is known

[1]Notes appear on page 15.

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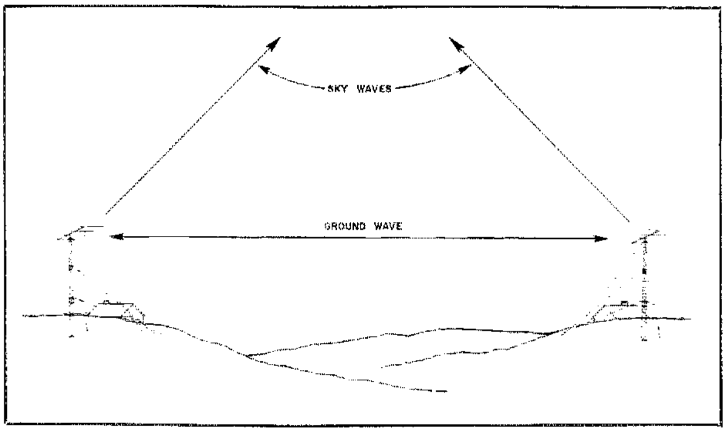

Fig. 1 — An example of the difference between ground waves and sky waves.

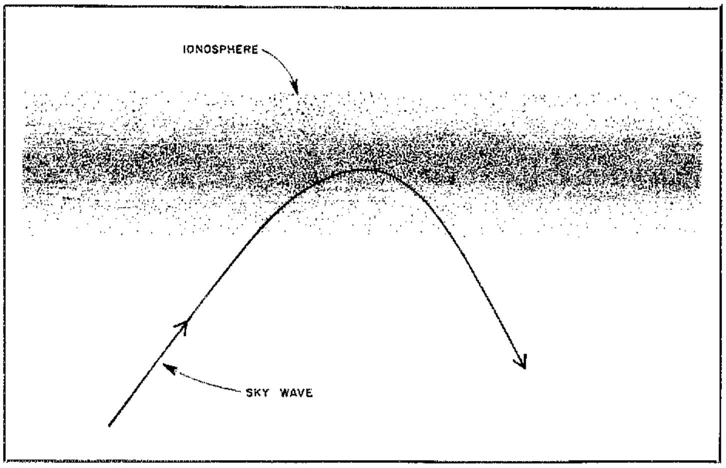

Fig. 2 — Under proper conditions, a radio wave entering the ionosphere will be refracted and follow a new course. This permits the signal to be heard on earth, perhaps thousands of miles from the transmitting antenna.

as the D layer, which lies between 37 and 57 miles above the earth (Fig. 3). This layer exists only during the daylight hours, and usually disappears within 30 minutes after sundown. Because it is located so close to the earth, the D layer is in a relatively dense part of the atmosphere. Here, ions often collide and recombine into neutral particles, which accounts for the rapid loss of this layer in darkness.

The D layer is not particularly useful to amateurs. Instead of refracting and propagating signals, it *absorbs* a great deal of them. A wave passing through this dense layer collides with a relatively large number of ions and sets them in motion. Much of the wave energy is thus used up through conversion to motion or heat energy.

Because a long wavefront will set more ions in motion than a short one, we can assume that the D layer will absorb more energy as the frequency of our signal decreases (Fig. 4). Additionally, the angle at which a wave enters the D layer has an effect on the degree of absorption. A wave going straight through has the shortest path and least absorption; a wave cutting through at a low angle has much farther to travel in the layer, and absorption will be greater. Because of these effects, the D layer is responsible for the 160, 80 and 40-meter bands being good only for short-distance communications in the daytime.

At night, when the D layer disappears, these same bands can often support DX communications of several thousand miles. This daytime absorption effect is insignificant on 20 meters and higher, which in part allows daytime DX communications on these bands.

The next higher ionospheric layer is the E layer, which is also the lowest one that will *support* radio wave propagation. This layer is located between 62 and 71 miles above the earth, and has characteristics similar to those of the D layer. For example, maximum E layer ionization occurs around noon local time, and rapidly drops off after sundown. During the period of peak ionization (midday), the E layer will absorb some energy in the lower-frequency amateur bands, but not nearly as much as the D layer. It is also interesting to note that X rays and meteors entering the atmosphere contribute to ionization of this layer. The E layer is also the scene of a spectacular type of vhf propagation known as *Sporadic E.* Because it is beyond the scope of this article, interested readers should consult the reference for an explanation of this phenomenon.[2]

Except for occasional propagation via the E layer, we rely almost exclusively on the outermost F layer to provide long-distance hf communications. Here, between 100 and 260 miles above the earth, rarification causes ions to recombine more slowly than in the other layers. Because of this, the F layer can often remain highly ionized throughout the night. As with the other layers, maximum ionization occurs around local noon time, with minimum occurring about an hour before local sunrise.

An interesting aspect of the F layer is its tendency to split up into two layers, known as the F_1 and F_2 layers. This separation occurs during the day, and causes the lower F_1 layer to take on much of the same characteristics as the E layer. Therefore, daytime propagation is largely supported by the F_2 layer. At night, the F_1 layer disperses and the F_2 layer slightly reduces its height above ground.

Refraction: The Critical Element

I mentioned briefly that refraction is the mechanism responsible for returning sky-wave signals back to earth. The most critical aspect of this phenomenon is the degree to which the waves are bent. There are two primary factors influencing this — the density of ionization and the length of the wave (or frequency). All other conditions being equal, bending will increase with higher ionization density. Also, bending increases with wavelength, or put another way, decreases as the frequency goes up. This sets up a condition in which both factors, working simultaneously, will finally determine whether a wave will be refracted back to earth.

Take a look at the example in Fig. 5. In A, we have an F layer of relatively low ionization, typical of nighttime conditions. Our 28-MHz signal is not refracted enough under these conditions to return back to earth. The 3.5-MHz wave, however, being of greater wavelength, is refracted much more and does make it back down.

In Fig. 5B, the ionosphere is more highly charged than in A, simulating typical mid-

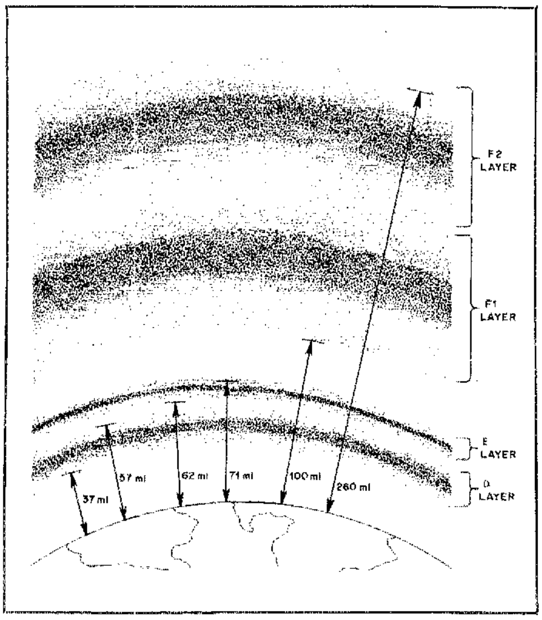

Fig. 3 — The various ionospheric layers with respect to the earth. Distances shown are not absolute, but vary with conditions as explained in the text.

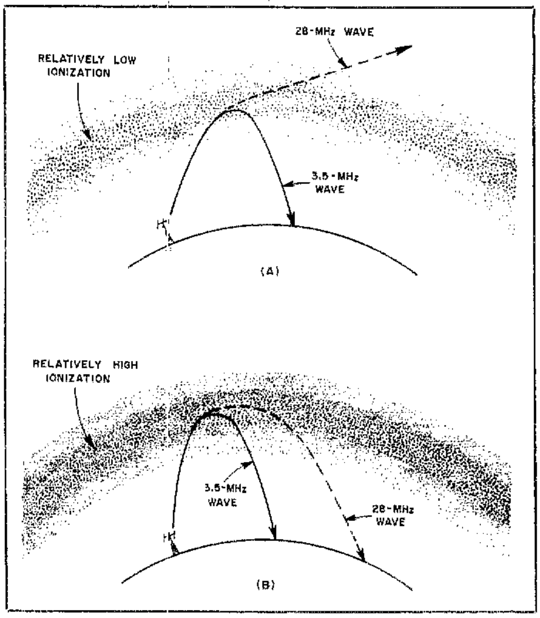

Fig. 5 — In A, the low-level ionization is insufficient to bend the 28-MHz wave back to earth; the level is high enough for 3.5-MHz propagation. Higher-level ionization in B is sufficient to refract the 28-MHz wave to earth.

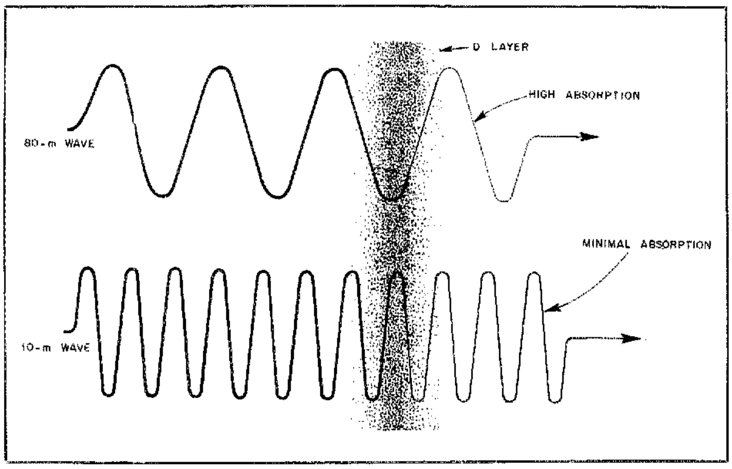

Fig. 4 — All other conditions being equal, a lower-frequency wave will undergo greater D-layer absorption than a higher-frequency one. The larger wavefront must travel a greater distance through the D layer.

day conditions. Now, with sufficient ionization we have a situation in which *both* waves are refracted back to earth. Note that the 28-MHz wave is not bent as much as the 3.5 MHz one, because of its shorter wavelength.

By now you should understand how the basic refraction process works. It's time to introduce a simple and valuable indicator that relates to our daily operating. The *maximum usable frequency (muf)* is, in the strictest sense, defined as the maximum frequency that will support communication between two specified points under existing conditions. For example, during one evening, the muf between New York City and Chicago could be 3.5 MHz, while at the same time, the muf between NYC and Denver is 28 MHz. And why is this? To answer the question, we must work one more factor into our discussion — *wave angle.*

We already know that the amount a wave is refracted depends on two factors: wavelength and the degree of ionization. But assume that for a fixed frequency and degree of ionization, waves penetrate the F layer at different angles. How does this affect propagation? Let's take a closer look.

Fig. 6A shows what typically occurs to a 28-MHz signal. Waves entering the ionosphere at high angles are not refracted back to earth, but continue out into space. As the wave angle decreases, there is a critical point where refraction causes the waves to return to earth. That angle is known as the *critical angle,* and all waves leaving at that angle or below will be propagated to earth.

The critical angle is also directly associated with a phenomenon known as the *skip distance* or *skip zone.* This zone or distance, as shown in Fig. 6, is a region where it is impossible for any regular sky-wave signals to be propagated. The length of the skip distance will vary according to the critical angle. Table 1 lists average skip distances for each band.

Fig. 6B shows the effects of the ionosphere on a lower frequency (3.5-MHz) signal. With all other conditions being the same as in Fig. 6A, we now see that the critical angle is much higher and the skip distance much shorter than on 28 MHz. Under these conditions, we would be able

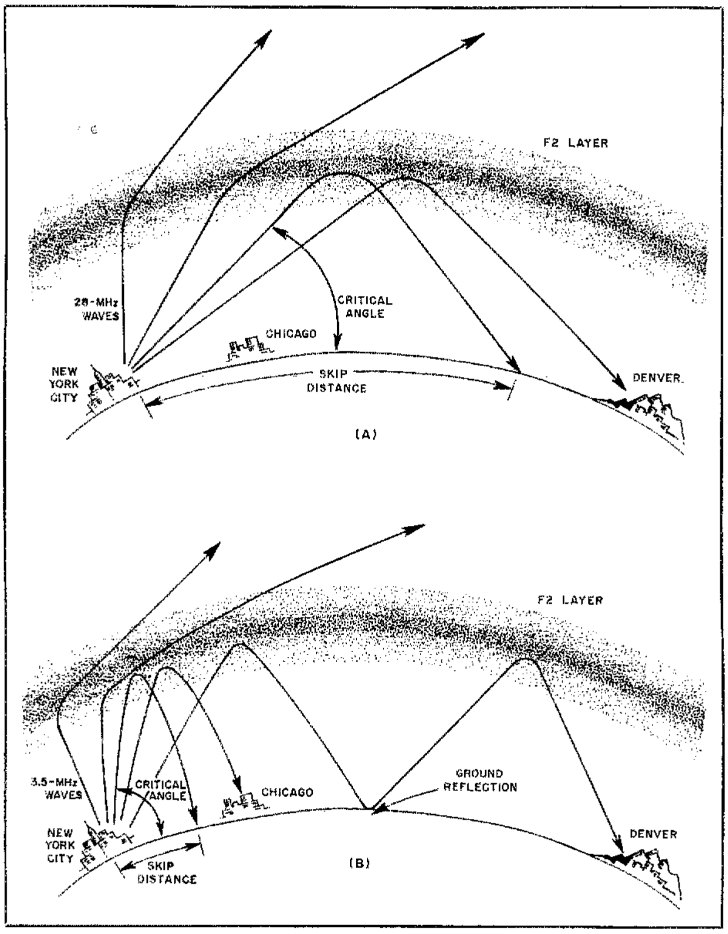

Fig. 6 — Illustration of how frequency, critical angle and skip distance are related. See text for explanation.

to QSO with Chicago from New York City. On 28 MHz, the skip distance prevents this. The reasons should now be clear as to why there is a different muf for every distance over which we wish to communicate!

Multihop Propagation

For the sake of simplicity, I have only mentioned wave propagation in terms of a single "hop" off the ionosphere. But the F_2 layer is at a certain altitude, and the maximum distance we can cover on a single hop is approximately 2500 miles. Therefore, communication over distances greater than this requires more than one hop, commonly known as *multihop propagation*. Fig. 6B shows how a wave returning to earth is reflected back up to the ionosphere, where it can be refracted again. This phenomenon can occur several times for a signal to be propagated around the earth. Because there is a considerable loss of signal strength with each hop, it is preferable to use lower-angle radiation, which takes fewer hops to reach the destination than higher-angle radiation. The factors determining radiation angle are covered in an excellent *QST* article by Hutchinson.[3]

Another factor in multihop path loss is whether signals reflect off a land mass or water. As you have probably guessed, water is the much better reflector of the two; signals will generally propagate more efficiently over it when multihop is involved. Is it any wonder that coastal stations have consistently big signals?[4]

It's Up to the Sun

We know that the sun plays a major role in the short- and long-term propagation variations we encounter. The general reason for this is quite simple: Changes in solar activity affect the sun's output of ionizing radiation. This in turn affects the degree to which our atmosphere is ionized.

Table 1
Approximate Skip Distances for the Amateur MF and HF Bands

Band	*Noon**	*Midnight**
160 m	0 mi	0 mi
80 m	0 mi	0 mi
40 m	50 mi	300 mi
30 m	300 mi	600 mi
20 m	500 mi	1000 mi
15 m	800 mi	(Daytime only)
10 m	1200 mi	(Daytime only)

*Local time at the midpoint of the path.

Logically, to predict propagation we must study solar activity. As with the weather, we are not able to predict this activity with 100% accuracy. However, we can use various solar indicators to predict band conditions with fairly good results.

By now you are undoubtedly aware that sunspots have a lot to do with band conditions. The presence of these grayish-black blemishes has been found to correlate directly with the sun's output of ionizing radiation; the more sunspots visible at one time, the more ionization we can expect.

Fortunately, sunspot behavior has been studied and well documented for the past 200 years. In this time, we have learned that sunspots (or groups thereof) move across the sun from east to west at a constant rate. This movement is caused by the sun's axial rotation, which takes about 27.5 days for a complete revolution.

Perhaps the most significant of all sunspot characteristics (at least for amateurs) is the 11-year sunspot cycle. Records indicate that a peak in sunspot activity occurs every 11 years, give or take a year. Along with this peak is a corresponding increase in the average muf, and general improvement of hf propagation conditions. Our last peak occurred in the spring of 1980, when 10 meters was open worldwide on a daily basis, and often well into the night. There were even occasions when 6-meter signals were propagated by the F_2 layer, indicating an extremely high level of ionization.

Sunspot Number and Solar Flux

These are the two primary indicators used to measure the amount of solar activity. Daily observations for sunspot count (although not the actual number of spots) are recorded, and averages determined for the month and year. The *smoothed sunspot number* for any given month is the mean for the preceding and succeeding six months. This number is also known as the Wolf number, after its inventor, or the Zurich smoothed sunspot number, because international sunspot records were stored there until recently.[5] Typical smoothed number values range from the single digits, during 11-year sunspot minimums, to over 200 during the tremendous 1957-1958 sunspot peak.

The solar flux number provides another

indication of ionospheric conditions. This number represents the amount of solar noise found on the 2800-MHz band. Research has revealed that on this frequency, noise amplitude is closely related to ionization of the F layer. This indicator has been monitored and recorded since 1947. You can obtain the daily solar flux number by listening to the WWV propagation bulletins at 18 minutes past each hour. The method of interpreting these numbers is beyond the scope of this article; those interested should consult other references for more detailed information.[6,7]

Propagation Predictions

Understanding the basics of propagation will help you avoid making gross misjudgments regarding signal paths. Normal fluctuations in conditions can catch you off guard, however, and perhaps even prevent you from keeping that schedule with a DX friend. In this respect, a little bit of forecasting can go a long way. Like weather forecasts, propagation forecasts are not 100% accurate, but can offer a bit of warning "before the storm."

The science of propagation forecasting is indeed an involved one. It is therefore a good idea for beginners to acquaint themselves with the available ready-made forecasts. It's worth mentioning again that WWV offers a propagation bulletin at 18 minutes past each hour. Every month *QST* offers easy-to-use charts in the How's DX? column. These, combined with the WWV data comprise a very useful tool for planning your operating.

Additional propagation bulletins are a part of the W1AW bulletin service. The bulletin schedules can be found in *QST* every other month (see the Table of Contents) or obtained by writing the ARRL Communications Department. DX-minded amateurs may consider subscribing to one of the excellent DX bulletins available.[8-10] In addition to information about exotic DX locations on the air, these bulletins offer up-to-date propagation forecasts to help you "nab the rare ones."

Happy Hunting

I hope this article has helped to eliminate some of the mysteries of propagation. As you continue to operate and gain experience on the bands, propagation conditions will become "old hat," and practically second nature to you. But beware, for as soon as you take propagation for granted, conditions will change abruptly and surprise you. Call it Murphy's Law or Mother Nature — that's what keeps propagation interesting!

Notes

[1]km = mi × 1.609.
[2]M. S. Wilson, "Midlatitude Intense Sporadic-E Propagation," *QST*, Dec. 1970 and March 1971.
[3]C. L. Hutchinson, "Getting the Most out of Your Antenna," *QST*, July 1983, p. 34.
[4]This explanation of multihop propagation may be a bit oversimplified. For an explanation of several other possible theories, see the *ARRL Antenna Book*, 14th edition, 1982, p. 1-8.
[5]Sunspot numbers are now being recorded and compiled by the Sunspot Index Data Center, 3 avenue Circulaire, B 1180 Brussels, Belgium.
[6]G. Hall, ed., *The ARRL Antenna Book*, 14th ed. (Newington: ARRL, 1982), Chapter 1.
[7]E. Tilton, "The DXer's Crystal Ball," *QST*, June, August and September 1975.
[8]The DX Bulletin, 306 Vernon Ave., Vernon, CT 06066.
[9]The Long Island DX Bulletin, P.O. Box 173, Huntington, NY 11743.
[10]QRZ DX, P.O. Box 4072, Richardson, TX 75080.

Strays

TIS DO'S AND DON'TS

☐ The ARRL Technical Information Service is offered free to members. Although we are eager to help newly licensed amateurs and others with technical problems, in fairness to members we cannot respond to continuing requests for assistance from those who choose not to join the League.

For us to respond promptly to your inquiries we must have:

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2) your amateur call and license class (tell us if you're not licensed)
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When writing, please observe the following guidelines so we may provide the best possible service to the greatest number.

1) Before writing for technical assistance, search your files of *QST* and other ARRL publications. The answer you need may be there, available immediately. Consult the annual index of articles in each December issue.

2) Please do not ask for comparisons among commercial products. Choice of equipment is largely a matter of personal preference. Consult Product Review information in *QST;* compare manufacturers' specifications in their brochures.

Do not ask for information on articles published in other magazines. Write to the editor or author of that article.

Do not request custom designs for amateur gear.

Do not ask advice on nonamateur matters. We cannot respond to questions about CB, marine radio, hi-fi, etc. (unless they concern interference caused by amateur gear).

3) Use a typewriter when possible; otherwise, write or print *clearly*. Please be reasonable in the number of questions you ask; try to limit your questions to three per letter.

4) When writing, please come right to the point, and be sure to share with us whatever experience you have had with the problem in question. This will avoid our reply covering ground you've already been over.

5) Address all technical questions to Technical Information Service, American Radio Relay League, 225 Main St., Newington, CT 06111. — *Bob Schetgen, KU7G, Technical Information Specialist*

QST congratulates...

☐ the following radio amateurs on 50 years as members of the ARRL:

- Charles Winkley, W1EIF, of Plymouth, Massachusetts
- Thomas A. Phillips, W3DOG, of Laurel, Delaware
- Ashod A. Hovsepian, W6EBM, of Sacramento, California
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- Carl C. Drumeller, W5JJ, of Warr Acres, Oklahoma
- William R. Reiss, W1HAX, of Elmwood, Connecticut
- Milton A. George, W1BKG, of Pittsfield, Massachusetts
- Fred L. Whitson, WØGEF, of Jackson, Missouri
- Fergus T. Lea, W1JRM, of Portland, Maine

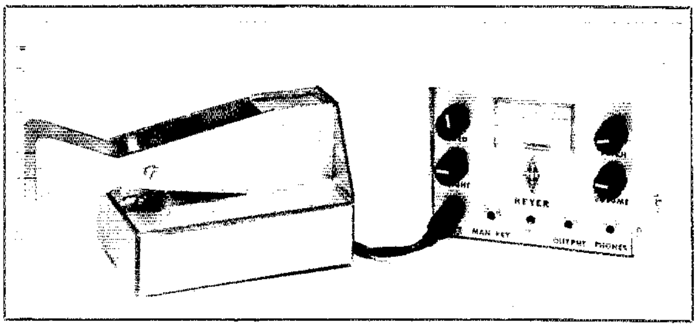
Next Month in QST

To begin the new year, January *QST* will feature the first of a comprehensive series of articles designed expressly for those with little or no electronics background. You may want to pass the word to your pre-Novice friends and neighbors.

Also in January, look for word about the ARRL Antenna-Design Competition, and an article that will tell would-be *QST* authors how to put together a technical article. If you're into kit building, you'll want to take in the January article that provides practical hints to help make your next project one that you'll be proud of for years to come.

CW on a Chip

Craving for a suitable weekend project? Keyers have been a longtime favorite of many! You can even wrangle a *free* paddle out of this offering!



By Bob Shriner,* WAØUZO and Paul K. Pagel,** N1FB

Keyers are fun to build! What makes the assembly of a good-quality, flexible keying system a cinch is the use of the Curtis keyer-on-a-chip. A ready-made pc board and kits of parts for the keyer and a unique, *no-cost* (no foolin'!) paddle make such an evening or weekend project that much more attractive. Does all this appeal to you? Then let's get on with it!

For some years now, the Curtis keyer-on-a-chip has been a regular part of the code transmission chapter of the *Radio Amateur's Handbook.* Although schematic diagrams and pictures of completed units using the Curtis ICs have been provided, a circuit-board template for would-be "homebrewers" has never appeared in the *Handbook.* This lack was the seed for this project. It grew rapidly from the presentation of a template only to that of offering a flexible keyer design.

The Curtis Chip

There are basically four varieties of the 8044 IC. Two of these (8044/8044B) are contained in a 16-pin package and the others (8044M/8044BM) in an 18-pin package.[1] The additional pins are connected to internal circuitry that provides a keyer sending speed monitoring function by means of a meter and a few other external components. The whole family of ICs features contact debouncing, rf immunity and self-completing character generation. A weight control, sidetone output and dot memory are also included. The memory function helps to prevent dot loss if the operator "leads" the keyer. With a quiescent current drain of about 50 μA, an on/off switch is not really required.

The "plain vanilla" (no suffix) and B-suffix ICs offer two slightly different iambic (squeeze) keying methods in addition to single-lever (non-squeeze) keying.[2,3] With the no-suffix IC, a dot or dash being sent when the paddles are released is completed and nothing else is sent. The B-suffix IC completes the dot or dash being sent upon paddle release, and then sends an opposite element; that is, a dot after a dash or a dash after a dot. Many squeeze-key operators prefer the latter method of iambic operation. If you're a single-lever paddle operator, you don't have to concern yourself about these factors; either IC should suit you.

Board Design

To make the keyer as universal in application as possible, the board is patterned so that any of the ICs mentioned earlier can be used with or without some of their inherent capabilities.[4,5] The board is single-sided, and is small enough to fit inside almost any transmitter or transceiver.

Any or all of the variable controls may be mounted on the board or brought out for external adjustment. Two on-board output keying options are provided: reed relay output (with or without arc suppression components across the contacts) or a transistor-keyed output that can be configured to fit your requirements.

Assembling the Keyer

Refer to Figs. 1 and 2 and the accompanying photographs during assembly. The parts overlay is shown in Fig. 3; the pc-board layout is in the Hints and Kinks section of this issue. The IC should be the last item installed. We'd recommend using a socket for it. If you install an 18-pin socket, you'll be able to use the 16- or 18-pin IC, the two unused socket positions simply being left empty when the 16-pin 8044 or 8044B IC is used.

You build the keyer to suit your personal requirements. Simply omit any of the components associated with the features you don't need or want. These may include: sidetone output, level and pitch control, and the speed-meter function. If you want transistor-output keying, you don't need to install the relay and arc suppression network. Should you not want the weight control, you'll still need to install a fixed-value 5.6-kΩ resistor between pins 15 and 16 of U1. If the weighting effect appears to be too heavy, reduce the value of C6 at pin 15 or remove it entirely from the circuit. The MANUAL KEY input can be used as a TUNE function; an spst switch that brings the line to ground will create a key-down condition.

If you elect to use the IC with the speed-meter function, any meter with a full-scale

[1]Notes appear on page 19.

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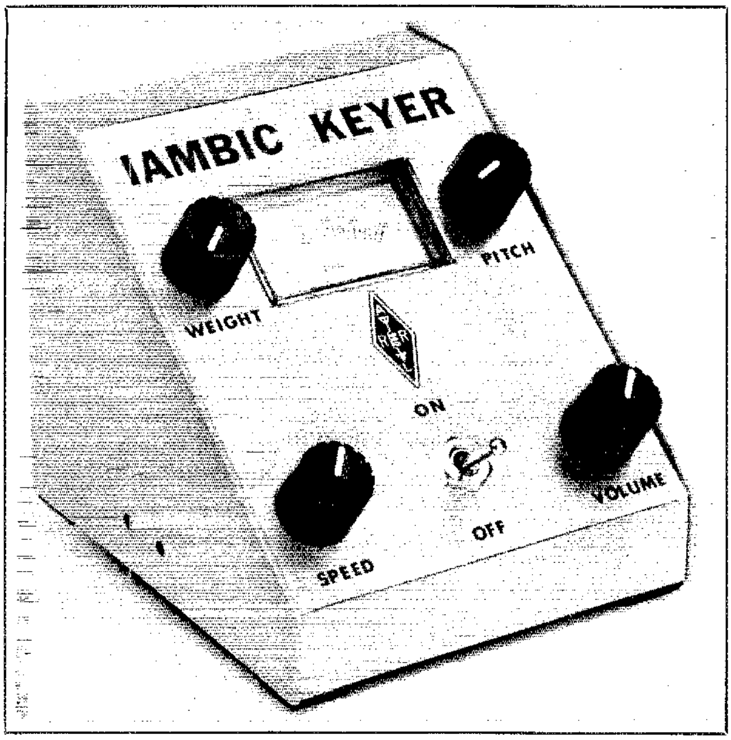
**Assistant Technical Editor, ARRL



deflection of from 50 to 500 μA can be used as long as it has a linear scale. A modified VU meter is used in the prototype shown in the photos. A new meter scale was made; it has 2-word-per-minute increments of from 0 to 100.

With the 100-kΩ resistor shown in series with the SPEED potentiometer in Fig. 1, the maximum speed of the keyer is about 50 wpm. Alter the value of the fixed resistor to modify the speed range. The keyer has a top-end speed of about 80 wpm.

A switch is included to turn the keyer and/or the transistor audio amplifier, Q2, on and off. You can prolong the life of the battery by leaving the audio amplifier off. The sidetone oscillator probably won't be required since most modern transmitters and transceivers have built-in sidetone monitoring circuits. The keyer monitor does serve as a good indicator of battery condition: As the battery becomes depleted, the note will become quite chirpy.

Relay-Contact Arc Suppression

Certain transmitter keying lines may require the inclusion of an arc suppression network across the keying relay contacts. Most modern transmitters and transceivers should not need this network (C10, R19), as they are usually operated at low voltage levels. But keying some transmitters and transceivers using tubes in the final amplifier may require the relay contact pro-

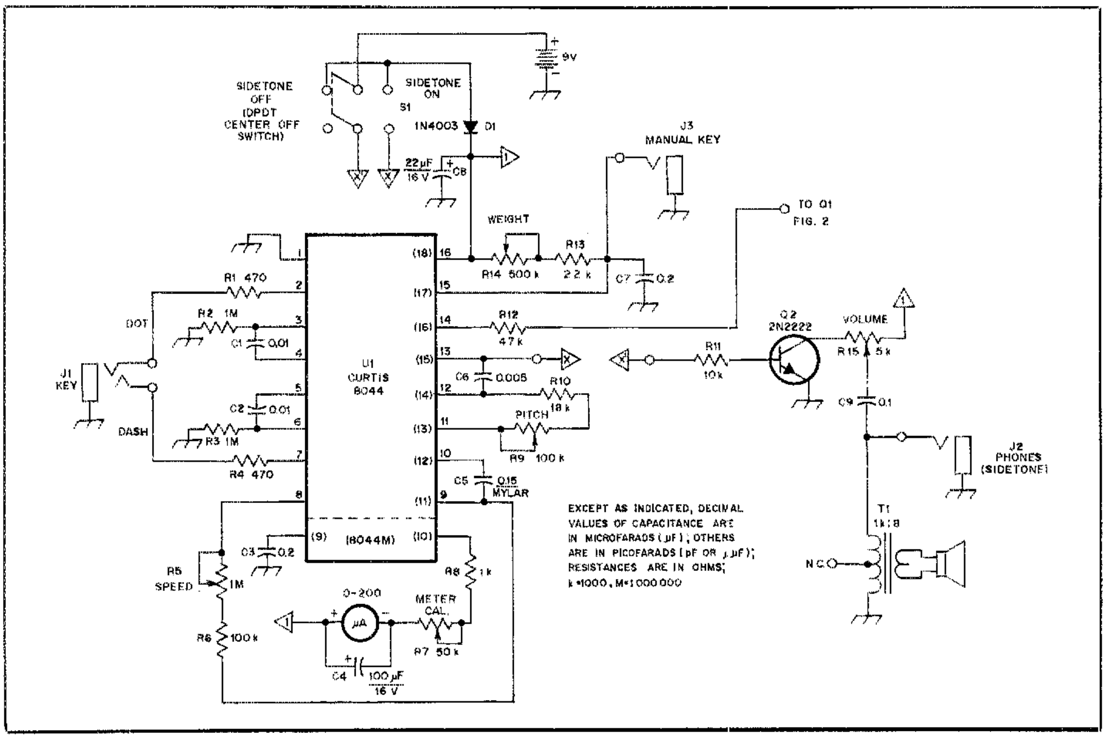


Fig. 1 — Schematic diagram of the keyer. All resistors shown are ¼-W, 5% types.

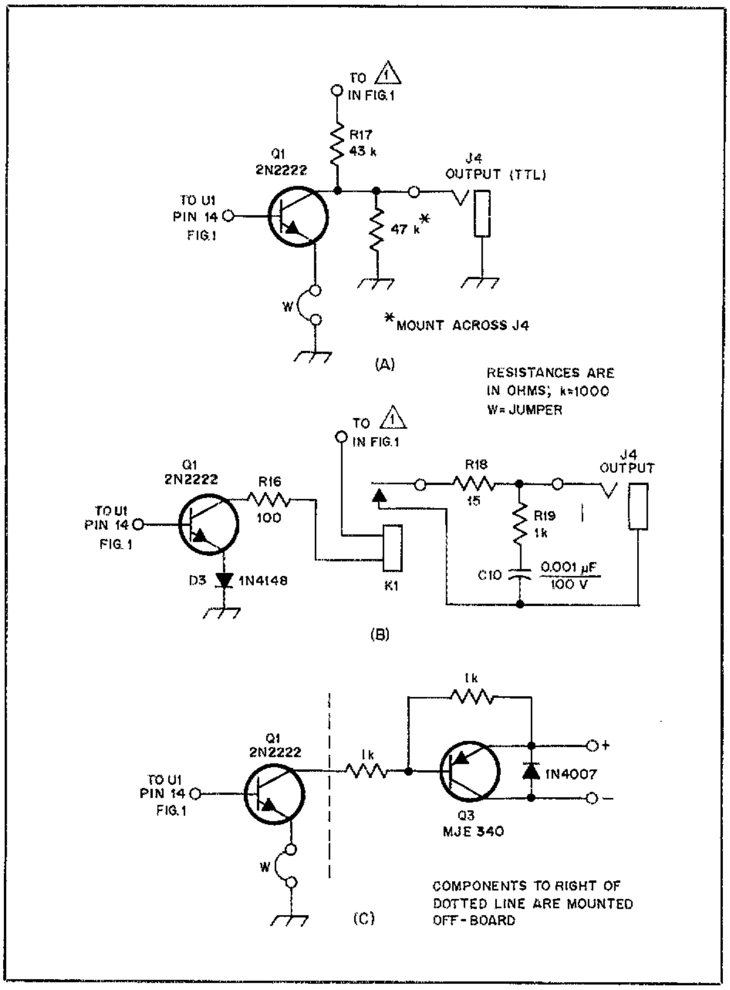


Fig. 2 — Some possible output circuit arrangements. The jumper (W) shown at (A) and (C) is inserted in place of D3. D3 is included to ensure rapid energization of K1; it may not be needed. Q3 and the components to the right of the dashed line in (C) are not mounted on the keyer circuit board.

tection. N1FB has successfully used the prototype keyer with a Kenwood TS-820S transceiver without the arc suppression network; no problems resulted. The key line voltage in that rig is −65-V dc. The appendix information should help you determine the network values required.

Calibrating the Speed Control

Sending speed can be determined by counting the number of dashes sent in a five-second period. That number is the code speed in wpm. A more refined measurement method uses an oscilloscope or a frequency counter connected to pin 12 of U1. Use the formula

$$\text{speed} = \frac{f}{1.2} \qquad \text{(Eq. 1)}$$

where speed is in wpm and f is the dot frequency measured at pin 12 of U1. Thus, for speeds of 10, 25 and 50 wpm, the frequency counter should display dot frequencies of 8.33, 20.8 and 41.6 Hz.

Keyer Case Construction

Two enclosure styles were tried. The first — the rectangular configuration — is easy to assemble. The control labels may be difficult to read at some angles, however. The sloping-panel version is more attractive, but a bit more difficult to put together. Either case style is available from Circuit Board Specialists; specify which you prefer.

Double-sided pc-board material is used for the box parts. Once the box parts have been cut to size, burnish them with fine steel wool. The parts' edges must be beveled to provide a good fit. Use a sharp file to produce this beveled edge. Work slowly and check the parts periodically for a snug fit. Remember: You can always file off more material, but it's impossible to replace material that's been filed away!

Lay the speaker panel on a flat surface and place the speaker grill in position. Use a toothpick to apply quick-drying epoxy cement to the grill-and-panel joint.

When the box parts are ready, tack solder them together and check for alignment and correct fit. If all is well, lay a bead of solder around each seam. A 25- to 45-W soldering iron should be sufficient. Another pair of hands can help to hold the parts in position.

Finish the case to suit your personal tastes. The original models have a combination of clear polyurethane varnish over most of the box. Light-blue epoxy spray paint accents some panels.

Install the panel mounted controls, jacks, meter and speaker. When mounting the keyer board, orient it with the meter-calibration potentiometer on the bottom.

A Free Paddle

Feast your eyes on that dandy paddle in the title photo! If anything has possibilities, that's it! To fishing enthusiasts, it might appear as a side view of headless, parallel-swimming fish — perhaps deserving the name "Tuna Twin Paddle." One might modify the design slightly to provide heads for the fish and have the output line exiting as a fishing line. Boaters might choose a different form and evolve a "Canoe Paddle." The possibilities are endless! Use your imagination and come up with something entirely "you."

To top all this excitement off, you can get the paddle parts *free!* Yes, indeed! Your request for the paddle parts will be honored by the delivery to your domicile of the pc-board material you'll need to construct one of these divine digital dexterity determinators.[4]

For someone wishing to experiment with iambic keying, this is an inexpensive route to follow. Sure, it's not the best paddle in the world, but it will give a good account of itself (more on that later).

Outside of the pc board material required, you'll need a couple of contacts from a junk-box relay and some machine screws and nuts to fashion the contact points. Some lead shot or other material can be used to fill the base. Scroungers can pick up lead shot from the shooting area of a local skeet-shooting club, about 150 yards from the firing line. (Please wait until the shooting is over.) Or, you might try the local garage; perhaps they can supply you with some lead tire-balancing weights.

Constructing the Paddle

Cut the paddle parts to shape. Dress up the parts' edges using a sharp file. Be careful not to get them out of square. With some steel wool, buff the parts to a sheen and spray them with a coat or two of clear acrylic lacquer to retain the finish. Paint the paddle if you wish.

First assemble the base and fill it with the lead shot or other weighting material. Pour in some epoxy cement to hold the material in place. To prevent marring the surface of your operating desk, cover the bottom of the base with a piece of felt or install some rubber feet. Be sure to drill the holes in the paddle arms for the contact screws and drill a hole in the rear panel to pass the key wires. Also, remove a strip of copper from each side of the contact-mounting block to isolate the contacts from the key frame and one another.

You'll need an extra pair of hands to hold the paddle parts in place while you tack solder them together. Once they are aligned properly, solder them along the entire seam.

Break off a couple of contact points from a discarded relay. Solder them to the upright contact piece between the two paddles and attach the output line. Two no. 4-40 machine screws are passed through holes drilled in the dot and dash paddles and secured with one nut on each side of each paddle. These are then adjusted to provide the contact spacing you desire. If the paddle is too stiff to suit your keying style, file the paddle arms to achieve a lighter touch.

How's It Work?

Admittedly, there were some chuckles, grins and outright guffaws when some members of the Hq. staff eyeballed the paddle. Not to laugh! Cw is serious business! Undaunted, N1FB toted the Dual Dolphin home. On the way, he found that the paddle rested rock solidly on the console of his car, the keyer occupying the passenger's seat. With this arrangement, he could easily send 20 wpm while zipping along at 55 mph. (Oh, for a 40-meter mobile rig and antenna!) The paddle/keyer combo became the main means of cw generation at N1FB. Keying must have been decent, since at least one station, during a long cw chat on 10-MHz, asked if he was using a "pokeboard" (keyboard)!

Lest you think this is a fish story, give the combination a try yourself. Perhaps you may even think of a minor modification or two. What the heck, the paddle shouldn't cost you anything and the keyer's a worthwhile addition to any shack.

Appendix

Most modern Amateur Radio transmitters, particularly solid-state designs, do not require additional relay-contact protection. Should your transmitter key line be a low-voltage type and not tolerate or require R18, it can be removed or jumpered.

Some grid-block-keyed transmitters will require additional relay-contact protection in the form of an added series-connected resistor if the key-line voltage exceeds 70 and a large-value bypass capacitor is tied between the key line and ground. Select the appropriate resistor value according to the accompanying table. Determine the resistor power rating by multiplying the resistor value by the square of the key-down circuit current in amperes. The added resistor can often be placed within the body of the key-line plug. For keying inductive loads (such as another relay), a silicon diode should be connected across the contacts of K1 (in place of C10, R19) to absorb the inductive kick.

Voltage	*Series Resistor Value (ohms)*
70	100
100	200
150	450
200	800

Notes

[1]Curtis Electro Devices, Inc., generously supplied the ICs used in this project.
[2]L. Fay, "The Iambic Gambit," *QST*, July 1981, p. 52.
[3]Curtis Electro Devices Lil' Bugger, Product Review, *QST*, March 1982, p. 47.
[4]A complete kit of parts is available from Circuit Board Specialists, P.O. Box 969, Pueblo, CO 81002.
[5]The keyer ICs are available from Curtis Electro Devices, Inc., Box 4090, Mountain View, CA 94040. Be sure to specify which IC you prefer.
[6]Templates for the sloping-panel keyer cabinet and paddle, and free paddle pc-board material are available from ARRL Hq. Send a business-size envelope and $2 to cover template and postage costs.

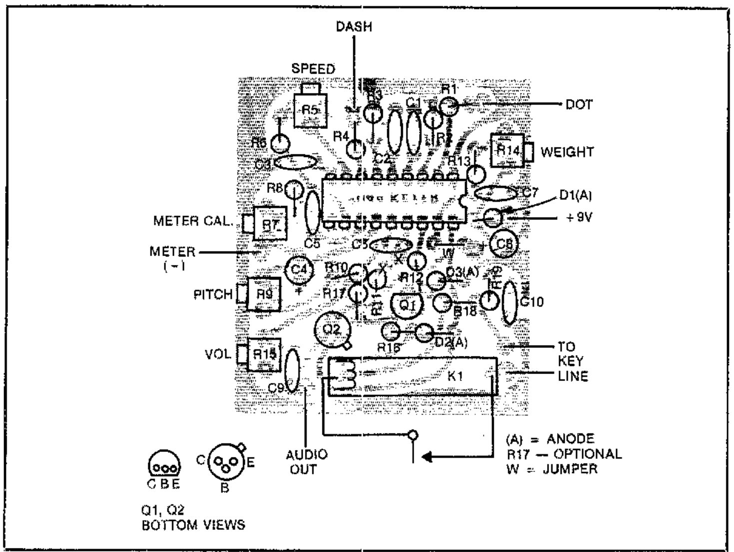


Fig. 3 — Parts-placement guide for the Curtis-IC keyer. Parts are placed on the nonfoil side of the board; the shaded area represents an X-ray view of the copper pattern. The etching pattern appears in the Hints and Kinks section of this issue.

Strays

QEX: THE ARRL EXPERIMENTERS' EXCHANGE

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- "MINIMUF for the Ham and the IBM Personal Computer," by John E. Anderson, WD4MUO
- More on AMTOR Protocol Change

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BEGINNER'S BENCH

Feeding Your Station

Fat coaxial line, skinny cable, open-wire feeders or 300-ohm ribbon line — the correct choice can save the beginner a few dollars while helping to ensure maximum performance.

By Doug DeMaw,* W1FB

There are a host of important factors to consider when planning the feed system for any antenna. The cost of the line should *not* be the first item to ponder. Rather, we need to be aware of the *loss* (measured in decibels per foot) at our chosen operating frequency and the quality of the cable — insulating material, extent of shield-braid covering and inner conductor size. Not all similar feed lines provide the same quality and performance. The choice between coaxial lines and balanced open-wire or ribbon lines is still another matter to contemplate. Finally, we should determine the longevity of the line. Pollutants in the soil and air, plus the effects of sun and temperature, have a significant deteriorating effect on feed lines; certainly, we don't want to replace the transmission line more often than necessary. After all, these are tough economic times, and nothing we buy seems to be priced as we feel it should be! Let's examine the high points of selecting the best feeder cable for several types of applications.

The Matter of Losses

How much loss can we tolerate in a transmission line? Our objective should be to reduce the losses as much as possible, but there must always be some loss in the system, however slight it may be. Most manufacturers of transmission line rate the product in decibels of loss per 100 feet for a specified range of operating frequencies.[1] As the frequency is raised, the losses increase. For a given length of coaxial cable the losses at 1.8 MHz may be so minor that we don't pay much heed to them. But at, say, 28 MHz and higher, they may be so great that half or more of our available transmitter power is lost en route to the antenna! This can happen even when the feeder is matched properly to the antenna at one end, and to the transmitter at the other end.

Fig. 1 contains a chart that will aid us in learning the various characteristics of feed lines, including the losses per 100 feet at frequencies from 1 to 1000 MHz. We can observe from Fig. 1 that the least effective line we can select is miniature RG-174/U. The loss per 100 feet exceeds 5 dB at 29 MHz. Thus, if our transmitter puts out 100 W of rf power, only 31 W will reach the antenna, assuming we are operating at 29 MHz with 100 feet of RG-174/U cable! Imagine what this could mean if we were using a QRP (low power) station with only 1 or 2 W of transmitter output power. Ouch!

Now, let's envision the same general situation while using the least lossy line — open-wire feeders. The loss per 100 feet at

[1]Notes appear on page 23.

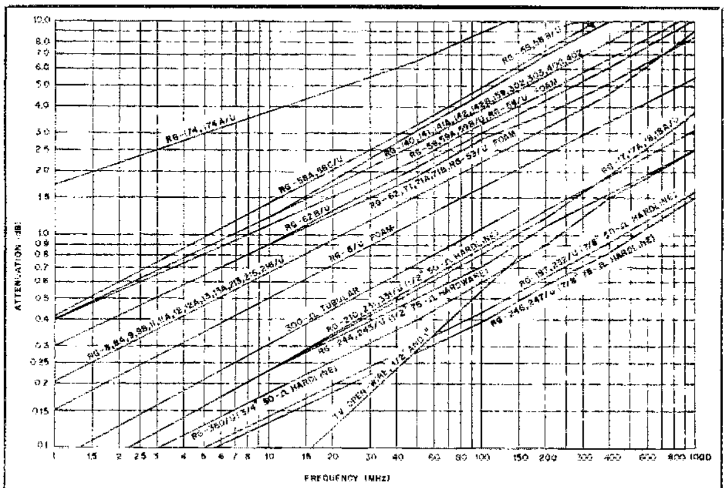

Fig. 1 — Method for determining transmission-line loss.

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29 MHz will be roughly 0.175 dB. Hence, for 100 W of output power we will be able to deliver 96 W to the antenna feed point, assuming the impedances are matched correctly at both ends of the line. So, we can see that even with the least lossy of feed line we will still be forced to sacrifice some of our power in the system. Additional losses will occur, however minor they may be, if we include Transmatches (antenna tuners) and baluns (balanced-to-unbalanced transformers).

There is no such thing in this business as 100% efficiency. So, somewhere along the course of our planning we must strike an acceptable compromise between losses and the cost of our feed line. RG-8/U coaxial line that contains foam dielectric is the frequent choice of amateurs for use from 1.8 to 30 MHz. The loss is 0.9 dB at 29 MHz and only 0.2 dB at 1.8 MHz. So, at 29 MHz our 100-W transmitter will deliver approximately 81 W of rf energy to the antenna feed point per 100 feet of foam RG-8/U line.

Semiflexible, aluminum-jacketed, foam-dielectric cable (Hardline) is the style of transmission line preferred by those amateurs who are willing to spend extra money for a low-loss feed system that should last for many years. This type of coaxial cable is well-suited to vhf and uhf work because of reduced losses and relative immunity to damage from the environment. Although the 0.5-inch-diameter Hardline (RG-210/U, etc.) is quite good in this respect, the 0.75-inch-diameter type (RG-360/U) is better with respect to losses. But, it costs substantially more than the smaller Hardline. The "flea on the dog's back" in either situation is obtaining the connectors for Hardline. Not only are these special fittings hard to locate — they cost a bundle! A method for adapting standard coaxial connectors to Hardline was treated in *QST* a few years ago.[3]

Using Open-Wire Feeders

You may hear old-timers extolling the virtues of open-wire feeders or "ladder line." In bygone days, the standard ham antenna was the end-fed or center-fed Zepp. It got its name from a style of antenna that was used on Zeppelin airships. Open-wire feed line was used with a tuner to permit operation over a wide range of frequencies. The tuner was used to provide an impedance match between the feed line and the station transmitter and receiver. The arrangement did not ensure a match at the antenna feed point; because the feeder losses were so low, however, the mismatch was inconsequential.

Many amateurs still use Zepp-style antennas because they are convenient for multiband operation when there is room for only one antenna. Such a system is shown in Fig. 2. The purpose of the Transmatch (transmitter to transmission-line matcher) is to convert the balanced feed-line format to an unbalanced 50-ohm condition for interface to the transmitter and receiver. The Transmatch is adjusted to provide an SWR (standing-wave ratio) of 1:1, thereby assuring maximum power transfer from the transmitter to the antenna system.

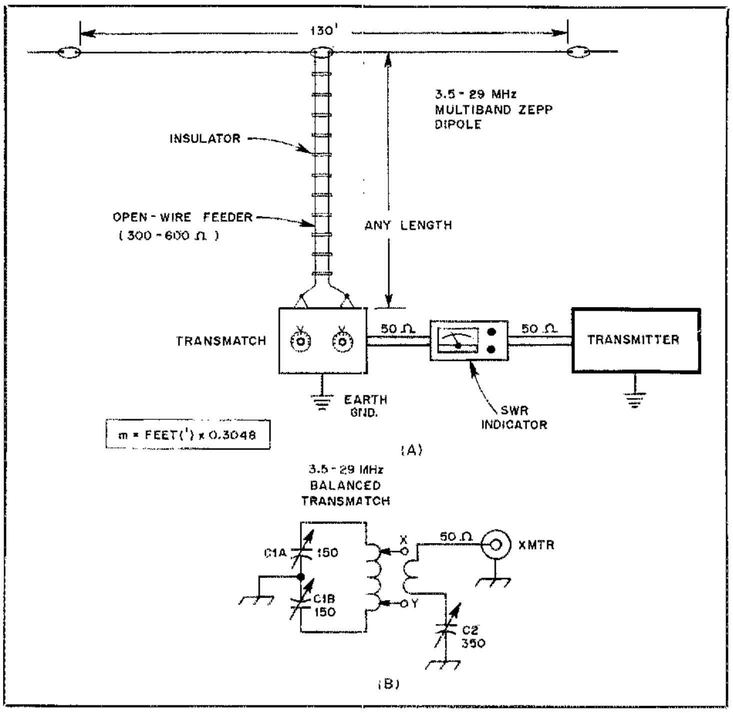


Fig. 2 — Example of a center-fed Zepp antenna for multiband use with open-wire feeders. A Transmatch is used to convert the system from a balanced to unbalanced condition while matching the line impedance to that of the transmitter (see text). Transmatch circuit is shown at B.

Some of you are no doubt wondering, "What should the open-wire line impedance be?" Actually, its value is not critical. A multiband antenna of the type seen in Fig. 2 may not be matched to the antenna feed point on any band, irrespective of the typical balanced feeder impedance. Generally, we will use open-wire line that ranges in impedance from 300 to 600 ohms. This depends on the wire diameter and center-to-center spacing between the two wires. Most commercially made open-wire feed line is 300 or 450 ohms with respect to the characteristic impedance. Homemade ladder line may be as great as 600 ohms in value. We can easily calculate the line impedance by applying Eq. 1.

$$Z_o = 276 \log \frac{2S}{d} \text{ ohms} \qquad \text{(Eq. 1)}$$

where Z_o is the feeder impedance, S is the distance (center to center) between the two wires and d is the diameter of the conductors. The dimensions are in inches. Hence, if we made an open-wire line from two lengths of wire that had an outer diameter of 0.125 inch, and the center-to-center spacing of the wires was 4 inches, the impedance would be

$$Z_o = 276 \log \frac{8}{0.125} = 498.5 \text{ ohms} \qquad \text{(Eq. 2)}$$

Spacers of good insulating quality should be used at 12-inch intervals, approximately, to maintain the desired spacing between the wires. Steatite, ceramic, Plexiglas,® fiberglass or phenolic is suitable. Some amateurs, in an effort to keep the cost low, have used pieces of wooden dowel rod after boiling the pieces in paraffin (canning wax).

What About TV Ribbon Line?

Can we use 300-ohm TV ribbon as balanced feeders? Sure! It suffers from a significant limitation, however: The insulating material — usually polyethylene — increases the line loss, as is the case with coaxial cable. Therefore, we will not enjoy the benefits of the almost air-dielectric open-wire line. But, we will have a balanced system.

An annoying problem with the use of TV ribbon is a change in the SWR when it rains, when the line becomes dirty, or when ice and snow are present. This can happen to a lesser extent when using open-wire line, but it seldom presents the annoyance that we will experience with ribbon types of 300-ohm line.

If TV line must be used, try to obtain the uhf style of feeder. It will be foam filled, or may have sections of the insulating material cut out at short intervals along the inner portion of the line. Also, try to select 300-ohm line with large conductors. Some of the "cheap" ribbon line has inferior insulation and very tiny conductors. Avoid this variety! It will be very lossy.

Back to Coaxial Cables

Commercial antenna designers need to consider a number of factors when choosing a proper feed line: power rating, operating voltage (rms or peak), shielding integrity, velocity factor (V_f) and environmental damage, for example. Not all of these points will be of concern to us amateurs, since we are dealing with low levels of rf power, comparatively speaking, and will not be operating our stations at altitudes greater than 10,000 feet or in ambient temperatures exceeding 100° F. Those two conditions must be taken into account by many commercial operators, however, since the greater the heat and altitude the lower the power capability of the transmission line. An article on this subject, along with derating charts and other feed-line matters not discussed here, was published in *RF Design.*[3] We will focus instead on the velocity factor and potential environmental damage. The operating voltage will also be treated.

Velocity Factor

Each type of coaxial or balanced feed line has a specific velocity factor. The V_f is dictated by the dielectric material used as insulation between the conductors. We can learn the V_f from

$$V_f = \frac{100}{\sqrt{\epsilon}}\% \qquad \text{(Eq. 3)}$$

were V_f is the velocity factor as a percentage of the speed of light and ϵ is the dielectric constant of the insulation. Solid polyethylene (for feed lines) has a dielectric constant of 2.31, and Teflon is rated at 2.07. Air is specified as 1.0, which indicates no need for concern about the V_f of air-insulated lines. Foam polyethylene has an ϵ constant of 1.56, so we must be mindful of whether our coaxial line contains solid or foam insulation. In accordance with these constants, we may apply the V_f of 66% for solid polyethylene cables and 80% for foam types of feed lines. If we should be fortunate (or wealthy) enough to obtain Teflon-insulated feeder cable, we can use 69% for the V_f.

Since each style of cable has a V_f, we can think of this as a time delay factor along the line. This time delay (T) can be obtained from

$$T = 1.016\sqrt{\epsilon} \qquad \text{(Eq. 4)}$$

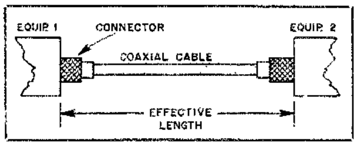


Fig. 3 — When you are cutting a length of coaxial cable to a specific part of a wavelength, it is essential to include the coaxial connectors in the measurement. The velocity factor of the cable is also included in the calculations (see text).

Table 1

Characteristics of Some Popular Coaxial Feed Lines

Cable Type	*Nom. Impedance*	*pF/ft*	*Max. rms Voltage*
RG-174/U	50 ohms	30.4	1500
RG-58A/U	50 ohms	28.5	1900
RG-59A/U	75 ohms	21.0	2300
RG-8A/U	50 ohms	29.0	5000
RG-11A/U	75 ohms	0.5	5000
Alum. jacketed Hardline (foam dielectric)			
1/2 inch	50 ohms	25.0	2500
3/4 inch	50 ohms	25.0	4000
7/8 inch	50 ohms	25.0	4500
1/2 inch	75 ohms	16.7	2500
3/4 inch	75 ohms	16.7	3500
7/8 inch	75 ohms	16.7	4000

Additional types and characteristics can be found in Chapter 3 of the *ARRL Antenna Book*, Fourteenth Edition.

where T is in nanoseconds/ft and ϵ is the dielectric constant of the cable insulation.

What does all this mean? Is it worth bothering with? By all means! Suppose you needed to cut a piece of RG-8/U coaxial line (solid polyethylene insulation) to one quarter wavelength at 3.9 MHz. Perhaps this line section is needed as a matching transformer. The free-space length of such a line would be found from L(ft) = 246/f(MHz). Therefore, the line for that condition would be 63 feet long. But, we must take into account the V_f of our feedline section. Hence, we will multiply 63 by 0.66 to secure the correction factor. The total length (including connectors) of the quarter-wave section then becomes 41 feet 7 inches.

Indeed, the coaxial connectors are a part of the electrical line length, as illustrated in Fig. 3. In an ordinary antenna system that is matched properly to the terminations at each end, we need not be concerned about the precise length of the feed line, except for the losses resulting from long runs of such cable.

Operating Voltage

Feed lines are rated for maximum safe rms (root mean square) operating voltages. The rf voltage from our transmitters is the rms type, assuming the wave form is pure. We can learn what the line input voltage is if we first measure the transmitter output power into a 50-ohm dummy load. E(rms) is obtained from

$$E = \sqrt{PR}\ \text{volts} \qquad \text{(Eq. 5)}$$

where P is in watts and R is in ohms. Therefore, if we delivered 100 W to a 50-ohm load, our rms voltage would be 70.7. This assumes an SWR of 1:1. If the line to the antenna exhibits an SWR other than this, however, we need to include this factor when determining the *effective* line-input voltage. This is accomplished by multiplying the actual input voltage by the square root of the SWR. So, if we had 100 W of power (70.7 volts rms) and an SWR of 1.7:1, the effective input voltage would be 92. We can learn from this that the higher the SWR the greater the possibility of damage to the feed line at the higher power levels. Table 1 lists the maximum safe rms voltage for a number of common feed lines.

Environmental Considerations

It's no secret that air contains all manner of man-caused pollution. These acid and alkali materials can, with time, contaminate the plastics used as coverings on coaxial cables. The poison migrates into the cable and causes it to be lossy. How bad the contamination, and the length of time required to spoil a feed line, depends on the level of contamination for a given region. Those living in large cities and industrial regions will fare the worst in this respect. The ultraviolet (UV) radiation from the sun will also affect the plastics in our antenna systems.

Cables that contain foam dielectric are generally more resistant to pollution damage than are the solid-dielectric types; the air bubbles in the insulation help to restrict the migration process. Aluminum-jacketed Hardline is more likely to resist the effects of pollution than is RG-style flexible line with its braided-copper outer conductor. Moisture is the first-order enemy of coaxial cables, especially when the moisture carries damaging chemicals into the feed line. Acid rain, a common threat in some areas nowadays, can cause oxidation of the feeder conductors. Because of UV radiation in areas of high sun activity, polyethylene jacketing is preferable to PVC outer-jacketed lines. In areas of high sun intensity, PVC lasts about half as long as polyethylene.

Special Impregnated Cables

If we chance to live where the salt content in the air is high, or if a substantial level of air pollution is prevalent, we may be wise to invest in one of the impregnated RG-8/U types of coaxial line. I use this style of line for all underground installations, and tend to prefer it for above-ground use as well. The center conductor is solid rather than stranded. Otherwise, the cable is very similar to RG-8/U. The transmission line is pressure-pumped

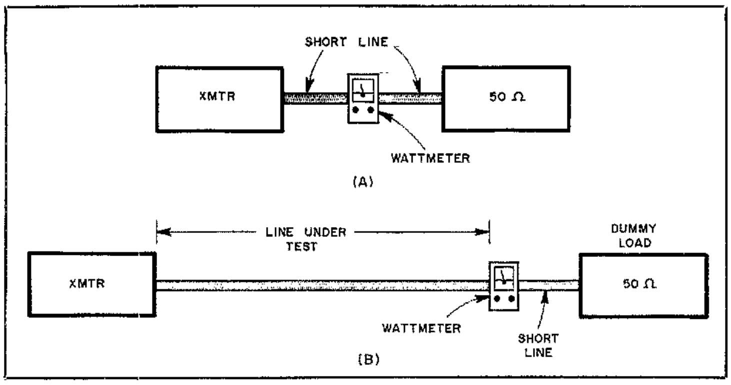


Fig. 4 — An old piece of coaxial feed line can be tested for losses by establishing a given transmitter output power (A) and maintaining it. Next, as shown at B, the wattmeter and dummy load are moved to the far end of the line to be tested and a second power reading is taken. The loss in decibels is then found from dB = 10 log P1/P2, where P1 is the highest power observed and P2 is the lowest reading.

with a sticky, low-loss compound to prevent dirt, air and moisture from penetrating it. It costs more than standard RG-8A/U, but it will last a lifetime under normal circumstances. It is available from at least two manufacturers.[4] It is also immune to damage from rodents, in that it is self-healing (some rodents delight in gnawing on exposed wires).

If you are suspicious of your feed line because it has been out of doors for a long time, or if it has been buried in the soil, it would be prudent for you to do a power-loss measurement of the line. This is a simple procedure. The essentials are shown in Fig. 4. The transmitter power is set to a specified amount after we connect a power meter and dummy load to the transmitter output. Next, the suspected coaxial line is attached to the transmitter, while the power meter and dummy load are moved to the far end of the feed line. A second power reading is taken. The loss in decibels is then compared to the loss of a new feed line, at the operating frequency. (See Fig. 1 and calculate the loss per foot.) If it is greater than 10% of the rated value, the feed line should be replaced.

We can slow down the deterioration process measurably by applying a quality sealant to the open portions of the cable when the antenna system is first commissioned. This will prevent unwanted moisture from entering the line and flowing along the inner and outer conductors. I like to use noncorrosive Silastic® compound or five-minute epoxy cement. I also apply sealant around the back side of my coaxial connectors. The putty used for dry floral arrangements works nicely as sealant for coaxial fittings.

Shielding Integrity

Bargain-price coaxial cable often has very poor shield-braid construction. The manufacturers of this material skimp in an effort to speed production and increase their profits. Well designed single copper braid provides 85 to 90 percent shielding efficiency. Two layers of this braid are even better, and triple-layer braid is best.

But these are quite expensive, and are not necessary for most of us. The extra shielding is essential for duplexer harnessing in repeater stations, and where incidental radiation must be held to a minimum. By comparison, aluminum-jacketed Hardline has an isolation rating (per foot at 100 MHz) of 700 dB. Single-braid line, such as RG-8A/U, provides only 48 dB of isolation. These isolation amounts are referenced to the desired transmitter output power.

If you are tempted to buy low-cost feeder cable, sneak a peek at the shield braid before paying for it. If there are gaps between strips of stranded braid, don't buy it. Some imported cables fit this description all too well!

What Have We Learned?

The object of this discussion has been to caution the newcomer about the matters of feed-line loss, potential deterioration of the cable and the velocity factor of a length of transmission line. A single lost decibel in a transmission or reception system may seem rather inconsequential. After all, few of us could discern an increase or loss of 1 dB while listening to a signal. But decibel losses are cumulative, so it is to our advantage to salvage every decibel possible. When we add up the potential losses in a complete amateur system — losses from feed-line attenuation, baluns, Transmatches and poor matching — it can become startling, especially at the higher frequencies! It isn't a casual matter to have even a 3-dB system loss, for that means we have cut our available rf power by 50%. We must accept some losses, but being miserly about the situation is wise. Three dB, for example, could mean the difference between being copied or lost in the QRN (noise) or QRM (interference).

We might summarize by saying that we should try to buy quality feed line, avoid the small-diameter coaxial cables for runs greater than 50 feet up to 30 MHz, and seal the points where moisture can enter the line. If we need to bury the feeder cable, we should consider using impregnated feed line. For vhf and uhf antenna systems, we should attempt justification of Hardline in preference to flexible feeders to minimize losses. RG-174/U line should be avoided at all costs, even for use in the lower part of the high-frequency spectrum.

Generally speaking, we will fare pretty well for all but the most demanding applications if we use foam-insulated RG-8A/U line in our 50-ohm systems. It represents a reasonable compromise between cost, longevity and minimum losses. Open-wire line will always be best for very long feed lines, however.

Notes

[1]mm = ft × 0.3048; mm = in. × 25.4.
[2]DeMaw, "Connectors for CATV 'Hardline' and Heliax," *QST*, Sept. 1980, p. 43.
[3]DeMaw, "RF Coaxial Cables — Choosing the Right One," *RF Design*, May 1980.
[4]VB-8 coaxial cable from Decibel Products Corp., 3184 Quebec St., Dallas, TX 75247 (W5KF, VP). Imperveon 50-ohm cable from Times Fiber Communications, 358 Hall Ave., Wallingford, CT 06492.

Strays

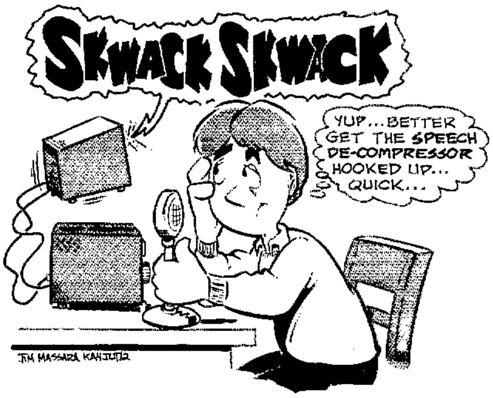
STRAY HINTS

☐ "Strays" are those interesting fillers used when space allows in *QST*. Think you have an item with Stray potential? Here are some hints to help your submission become one. (1) Be sure the information will be of interest to most readers of *QST*. (2) Submit your material before deadline — the 8th of the second month preceding desired publication (i.e., arrive at Hq. before December 8 for February *QST*). (3) Any photographs you send should be good-quality, black-and-white glossy prints. Color prints, slides and instant photos do not usually reproduce well.

Items submitted are normally acknowledged, but that doesn't necessarily mean that your Stray will be appearing in *QST*. We receive far more material than we can find room for. If you want your material returned, please include a statement to that effect and an s.a.s.e.

Follow the above hints and maybe your Stray will find a home in *QST*. — *Andrew Tripp, KA1JGG*

Try This Speech "Decompressor"



Tired of listening to unpleasant or monotonous speech-processed signals? Restoration, in part, can be accomplished at your receiver output with this interesting gadget.

By Eric Nichols,* KL7AJ

Speech processing for a transmitter is probably as good as the circuit design or the operator's ability to adjust the processor correctly. Unfortunately, neither condition is met most of the time. But, even when the overall system is functioning as prescribed, our ears or minds tend to tire of the restricted-speech sound. It's unlikely that we would be so brash as to ask the other station, "Could you please turn off your processor for awhile?" A more practical alternative is to "decompress" the speech at the output of our receiver or transceiver.

It appears that speech processing is here to stay. Practically all new transmitters and transceivers contain processors of some form. Isn't it surprising that no manufacturer incorporates the natural complement to speech compression — audio expansion at the receiver end? Audio expansion is no stranger to the hi-fi enthusiasts among us. These dynamic-range enhancers have been around for years. They compensate for the lack of "head room" on audio recordings. These devices are expensive when designed for music systems, but a simpler and less expensive expander can be devised for speech-only reception.

Here we will consider a simple decompressor which, when connected between a low-level audio output of a receiver or transceiver and an external power amplifier, can effectively restore to the original product even the most "squashed" audio. This project can be assembled in an evening or during a weekend.

Circuit Information

Fig. 1 shows the circuit of the decompressor. We should be aware that all processors — rf or audio — achieve the same end results: to reduce the level of difference between audio peaks and to bring the average power to a higher plateau. This is done with some type of nonlinear amplification.

D1 and D2 of Fig. 1 cause U1 to operate nonlinearly in the opposite sense. As the level into the diodes increases, the amplifier gain also increases. This effect accentuates the difference between the peak voltages. The values for R1 and R2 can be changed to any resistance from 0 to 10 kΩ to tailor the expansion to the amount desired. The values specified in Fig. 1 provide approximately 20 dB of expansion.

Laboratory Tests

I recently demonstrated this compressor at a club meeting. The test system consisted of a closed-circuit audio chain with speech that was prerecorded. The audio was first compressed with 20 dB of instantaneous peak clipping (the most "trashy" method of speech processing). Next, the clipped audio was expanded and fed to a quality audio amplifier; the difference between the original and the expanded audio (after clip-

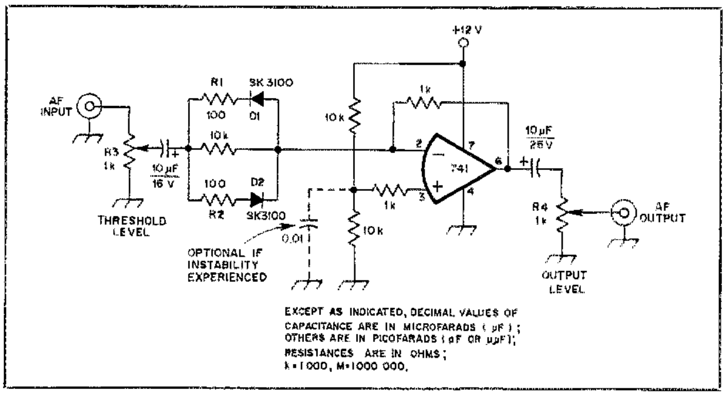


Fig. 1 — Schematic diagram of the speech decompressor. Resistors are ¼- or ½-W carbon composition. V_{CC} can be from 9 to 13.5, positive. R3 and R4 are linear-taper carbon controls. D1 and D2 are small-signal silicon diodes (1N914 or equiv.).

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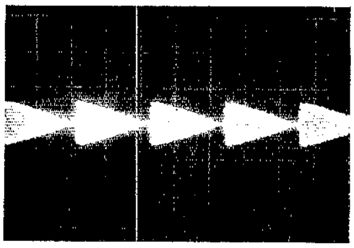


Fig. 2 — Original audio wave form without processing or expansion, as viewed on a scope.

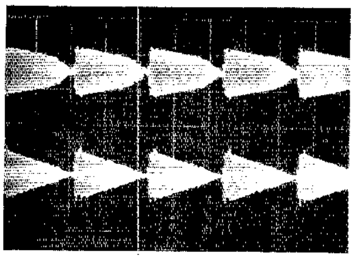


Fig. 3 — The upper trace shows the audio energy after being processed. Processed audio, recovered then expanded by the circuit of Fig. 1, is shown in the lower trace. The ratio between the peaks and ebbs of the audio signal is more pronounced in the lower waveform example.

ping had taken place) was indistinguishable to anyone in the room. Figs. 2 and 3 contain oscillographs of the unprocessed audio, compressed audio and expanded audio energy.

The circuit of Fig. 1 has an unexpected bonus. It reduces the background noise, particularly the high-frequency hiss, considerably. The speech information seems to jump right out of a noiseless background. This greatly enhances the listening quality.

Adjustments

The expander input level should be low. It can be taken from a low-level output point in the receiver. The 600-ohm phone-patch output on most transceivers is ideal for the take-off terminal.

First, we need to set the output level for a fairly high amount. Next, we advance the threshold-control setting until background noise can just be heard. Then the control is backed off just below this point. Now we can listen with pleasure to those hams who don't know that their processors have an "off switch" at our end of the line!

The decompressor has what might be considered a disadvantage: It requires an outboard audio amplifier. If you don't care for a collection of outboard equipment, you may want to try my "Mickey Mouse" way of achieving the necessary result. This method works only with solid-state rigs that contain complementary-symmetry output stages. It will not work with rigs that use conventional Class A output amplifiers.

Most of today's rigs have an af-output stage that is similar to the circuit of Fig. 4. D1 and D2 provide a slight forward bias for Q1 and Q2 to reduce crossover distortion. Sometimes a resistor is used in place of D1. All we need to do in order to introduce crossover distortion intentionally is to change the bias on Q1 and Q2 of Fig. 4. The crossover distortion provides the same results as is created by D1 and D2 of Fig. 1, surprisingly. If there are two such diodes in your rig, you may jumper one of them (as in Fig. 4), and that will do the

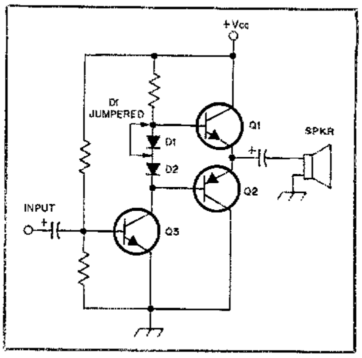


Fig. 4 — Typical circuit of a complementary-symmetry audio channel found in a modern receiver or transceiver (see text).

trick. As an alternative, you may replace the silicon diodes with germanium units, which have roughly half the forward-conduction voltage characteristic of silicon diodes. If a resistor is presently in use at D1 of Fig. 4, you may need to reduce the value of the resistor to approximately half the present value. Do not replace both diodes with resistors, because at least one diode is necessary to ensure thermal stability.

Final Comments

The human voice is a great thing to hear. The state of the art seems to require that the voice quality be "butchered." So, let's use the available technology to increase the effectiveness of our amateur communications. But let's also try to recover the pleasant character of the human voice.

Strays

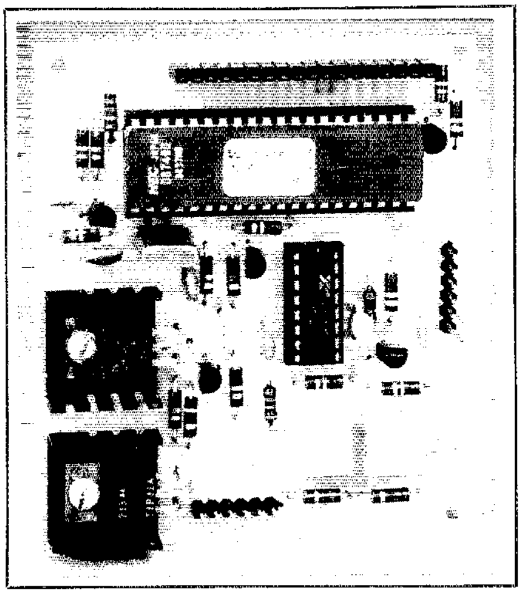
Gary Firtick, K1EB, is one of about 25 members of the Hen House Gang in Bethlehem, Connecticut, who will be operating W1FHP until January 7, 1984, looking to contact hams in as many towns of Bethlehem across the country as they can. This will be their 28th Christmas season operation. See Special Events, Nov. 1983 *QST*, for more details. *(photo by KA1YP)*

CALL FOR PAPERS

☐ Papers are invited for the Experimental Packet Radio Networks International Symposium to be held in Linköping, Sweden, May 26-27, 1984. Areas of interest are experimental packet radio, networking routing and control, distributed processing, channel-access protocols and proposed standards. Write to SOFTNET User Group, Dept. of Electrical Engineering, Linköping University, S-581 83 Linköping, Sweden, or call 46 13 28 10 00 and ask for Per Lundgren or Gunilla Svahn, Dept. of EE.

Deadlines are January 31, 1984, for the notice of intent, and March 31, for camera-ready manuscripts. Author's kits will be mailed by February 29. — *Jens Zander, SM5HEV*

The Microcomputer Repeater Controller



Planning to upgrade your repeater control system? Step into the world of microcomputers with this simple, inexpensive controller. It is easily interfaced to most existing repeater circuits.

By Scott M. Freeberg,* WA9WFA

Many repeater owners and users would like to have a computer-controlled repeater, but are reluctant to build such a system. This may be caused by unfamiliarity, computer complexity, potential high cost, or the nightmare of interfacing the system to a new or existing repeater. The Microcomputer Repeater Controller should quiet all of these complaints. The system is simple, low in cost and easy to interface. Two ICs on a 3- × 3-inch pc board provide complete microcomputer-controlled operation for your repeater.[1]

You are probably wondering what features a controller that only uses two ICs can possibly provide. The answer is *plenty*, if the ICs are a completely self-contained computer and a sine-wave tone generator! Just consider some of the features: two cw identification (i-d) messages, an i-d to i-d timer, a hang timer, a time-out timer, courtesy beep, pre- and post-time-out cw messages, transistor drivers for direct or indirect push-to-talk (PTT) operation and a flexible carrier-operated relay (COR) input interface. A single 12-V dc power supply is required to operate the unit. An optional User Module allows all timer functions to be reprogrammed instantly. Auxiliary control inputs are also provided, for use with an autopatch or a voice-message generator.

System Concepts

My goal at the start of this project was to design a completely self-contained microcomputer repeater controller. I wanted to include all functions and system-interface hardware on a single pc board. At the heart of the controller is a single-chip computer that does not require any support ICs. Computer software performs the functions normally done by two or more pc boards full of digital circuitry. This software can be personalized with i-d messages and timer features that meet the needs of each repeater owner.[2]

I decided to use a sine-wave tone-generator IC to produce a clean, crisp cw note for the i-d messages and courtesy beep. This seems to be an often-overlooked function, judging by the chirpy, harsh tones that I have heard from repeaters.

Problems normally associated with interfacing a controller with a repeater have been reduced by including a variety of interface circuits right on the controller pc board. For example, a two-level PTT interface , which will allow high- or low-current, direct PTT or indirect relay drive, is included. The receiver COR-to-controller interface is a transistor input circuit that can be configured to accept a variety of COR signals. I included several auxiliary control inputs in my design, which can be used by other repeater devices. The computer can recognize the presence of these devices, and aid in interfacing them to the repeater system.

Most repeaters are able to supply +12-V dc for the controller. The tone generator can operate from this voltage, but the microcomputer requires a +5-V dc supply. A 5-V regulator is included in the circuit, to meet this requirement. The current drawn from the 12-V supply will be under 100 mA.

Before starting to design the hardware, I made a list of all the features that should be included in the controller. This is an important part of any project, because your design can take into account the most efficient method of meeting your goals.

CW Features

Most users are keenly aware of the repeater i-d features. I gave this area special attention to help our club machine have a good reputation. An EXAR 2206 tone generator produces a clean sine-wave note for the cw functions of the repeater controller. Two cw i-d messages are provided by this controller. I call them the "first" cw i-d and the "normal" cw i-d. After the repeater has not been used for several minutes the microcomputer stops all control timing. When a signal is received, the first message is sent. Our repeater transmits VIA WA9WFA/R FORT ATKINSON. The normal cw i-d is sent for each succeeding i-d

[1]Notes appear on page 31.

*809 West Cramer St., Fort Atkinson, WI 53538

during *that activity* period. We use a shortened version of the first message, WA9WFA/R. These i-d messages are programmed into the microcomputer, and can be almost anything you want.

Another feature that is included with the cw i-d is a timing window designed to avoid double i-ds. On many repeaters, when the activity stops, the controller is reset immediately. If the machine is accessed again after a pause of a few seconds, it will send the i-d message. The microcomputer-controller timing window looks for activity after the last normal i-d. If someone accesses the repeater before the window closes, the controller is not reset, and the next i-d will not be sent until the regular i-d time has elapsed. After the window is closed, the controller is reset. The next user to access the repeater will cause the first i-d message to be sent.

The cw speed can be programmed as 10, 15 or 20 wpm if you order your microcomputer chip from Processor Concepts.[5] These speeds can be changed at any time by using the optional User Module that is available from the same supplier.

A short tone, called a courtesy beep, is sent a few seconds after a user stops transmitting. The next users should wait for this beep before starting their transmission so others can join the conversation. The time-out timer is reset just before the beep.

The delay time before sending the courtesy beep and the beep length are programmed into the microcomputer. Common values for these times are 0.5, 1, 1.5 and 2 seconds for the delay, and 0, 0.04, 0.08 and 0.12 seconds for the beep length. These parameters can also be changed by means of the User Module.

Timer Features

All of the timing functions for the repeater controller are programmed into your microcomputer. They are set to your specifications when you buy the chip, and can also be changed later with the User Module.

When a user stops transmitting, the controller is programmable to hold the repeater transmitter on for up to five seconds (the maximum hang time allowed by FCC rules). During this time the courtesy beep will be sent. There are arguments for and against using a hang timer. A long hang time results in fewer closures and extended working life for relays. This is one good reason for using a hang timer.

FCC rules require a remotely controlled transmitter to be turned off within three minutes if a malfunction should occur. Many control operators use a time-out timer to comply with this rule. A time-out timer will also prevent one user from monopolizing the repeater for long periods. When the carrier is dropped and the courtesy beep sounds, the timer is reset to zero. The time-out timer can be programmed for from no time-out to 9 minutes.

Rather than just abruptly turning off the repeater when a time-out occurs, the computer is programmed to send a pre-time-out warning message. This message consists of a series of 10 beeps sent just before the transmitter is shut off. Listeners will be alerted to what is about to happen. After the carrier has been dropped the computer will turn on the transmitter and send a post-time-out message. This cw message can be just about any length you want. Our club message is ?? TIME. These two messages provide a nice feature to keep the users informed about the repeater status.

The time interval between cw i-ds is called the normal time. This timer function is programmable from 30 seconds to 10

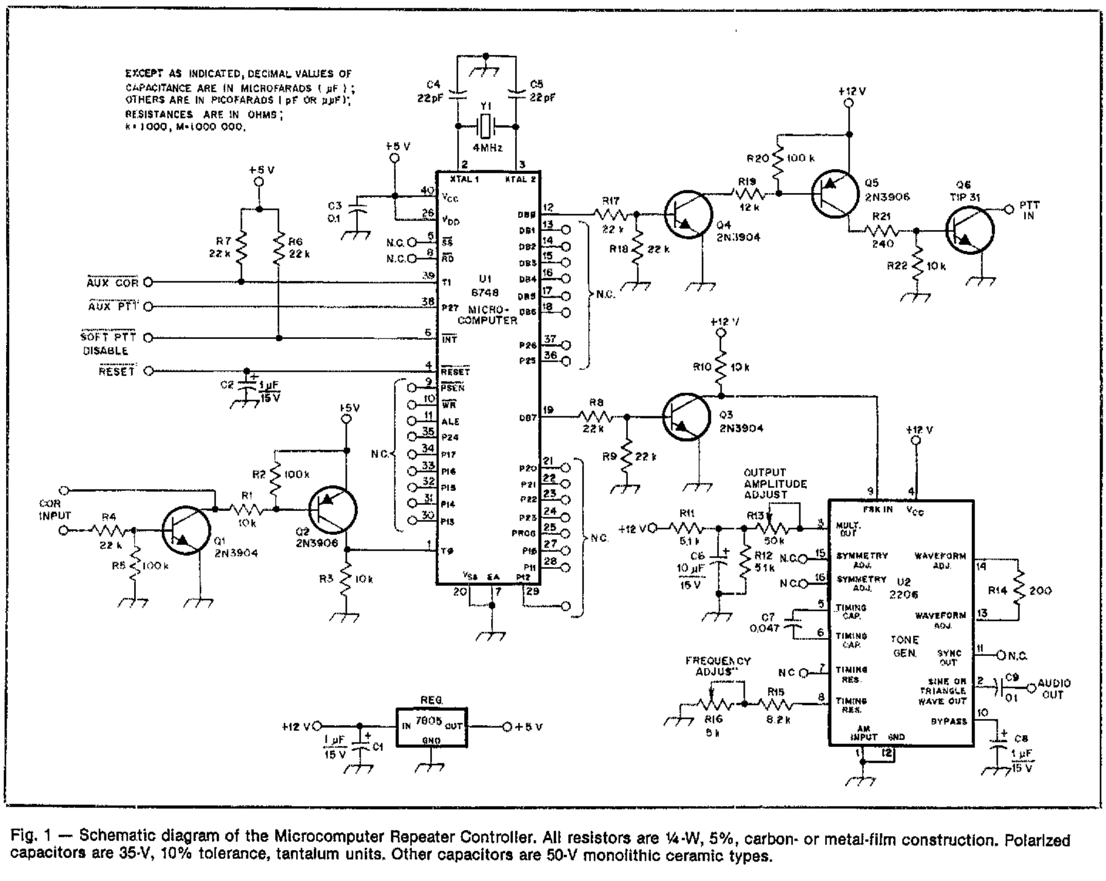


Fig. 1 — Schematic diagram of the Microcomputer Repeater Controller. All resistors are ¼-W, 5%, carbon- or metal-film construction. Polarized capacitors are 35-V, 10% tolerance, tantalum units. Other capacitors are 50-V monolithic ceramic types.

minutes. The normal timer resets each time an i-d is sent. Also, the timer-window feature uses the normal timer to reset the microcomputer controller.

Hardware

A complete schematic diagram for the Microcomputer Repeater Controller is given in Fig. 1. The heart of this project is an Intel 8748 single-chip computer. One unique aspect of this chip is that all of the microprocessor support functions normally supplied by peripheral chips are contained in this one 40-pin IC. The 8748 features 1 K of Programmable Read Only Memory (PROM), which stores the control program permanently. There are 64 bytes of random-access memory (RAM) available for data and system memory. The computer clock oscillator circuit is self contained and requires only an external crystal and two 22-pF capacitors. For timing applications, an internal 8-bit counter is included. This can be used to generate time delays or internal clocks for real-time processing. It is interesting to note that the computer chip provides 27 I/O lines, divided between 24 bidirectional I/O lines and three input lines.

For the 8748 microcomputer to input and output control signals, transistor interfaces are required to isolate and drive other devices. To simplify the interface problems of connecting the controller to existing repeaters, I designed the inputs to the board so they are pulled to a known state when disconnected. In the case of the COR signal, this means that the transmitter will be held off rather than on if the control line is broken. All of the inputs are designed to use negative logic, which means the input will be high when it is inactive, and pulled to ground to be activated.

I use a 4-MHz crystal for the clock. Other frequencies can be used, but any change will affect the internal counter, and that will affect the controller timing features. The higher the clock frequency, the faster the internal timer will count.

Tone Generator

An Exar 2206 function generator provides a pure sine wave output for use with the cw messages. I selected this chip because it is capable of producing a high-quality sine, square or triangular waveform that has excellent stability and accuracy over a wide range of operating temperatures and voltages. The frequency and amplitude of the generator are established by means of the components connected to IC2 of Fig. 1. R13 controls the output amplitude. The 50-kΩ variable resistor shown will provide an output of up to 3 V. C7 and the series combination of R15 and R16 set the tone frequency. R15 presets the upper frequency limit. Exar recommends that this value be not less than 4 kΩ. R16 varies the frequency over the desired range. For the values given, this

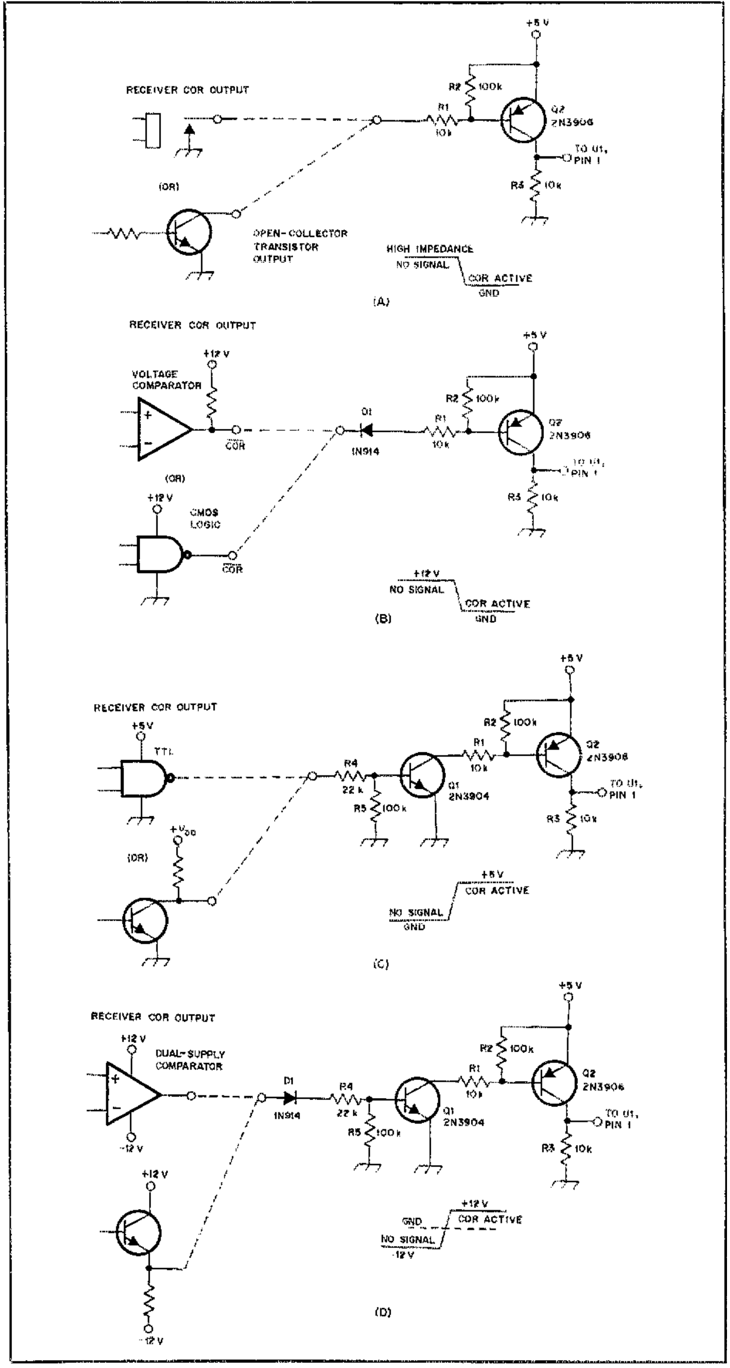


Fig. 2 — Four methods of interfacing the receiver COR signal to the controller board are shown. If the input is pulled to ground when the COR is active, but no voltage is present when it is inactive, use the configuration shown at A. Use the circuit given at B if the COR line supplies a positive voltage when inactive and is pulled to ground when active. Part C shows an interface circuit that can be employed for a receiver COR line that is at ground potential with no signal present, and supplies a positive voltage when active. Part D shows a configuration that will work with a COR that is at ground or a negative potential with no signal and a positive voltage when active.

range is from 1650 to 2550 Hz. For a different range of available tone frequencies, you will have to calculate the new resistor and capacitor values, using Eq. 1.

$$f = \frac{1}{(R15 + R16)(C7)} \qquad \text{(Eq. 1)}$$

where R15 + R16 is used to determine the lowest frequency. To calculate the highest frequency, the denominator in Eq. 1 is simply (R15)(C7). The frequency stability of the tone will depend on the temperature characteristics of these components. I recommend that you use a 0.047-μF ceramic monolithic capacitor with an X7R temperature characteristic ($\pm 15\%$ capacitance variation over a temperature range of -55 to $+125°$ C). Use a good-quality variable resistor for R16.

The tone generator is controlled by the microcomputer-bus I/O port on pin 19. When this pin is low, the tone is off. If pin 19 goes high, Q3 turns on and enables the tone. R9 is required because of the high impedance state of the I/O port on reset or power up. The output from the tone generator is capacitively coupled, and can be connected directly to the mixer in most repeater transmitters. Use shielded cable for this connection to help eliminate interference.

COR Interface

The COR interface is designed to handle just about any receiver COR output. This includes TTL, CMOS, HTL and open collector transistor circuits. The interface will accept 5-V logic levels, 12-V levels and the output from split-supply voltage comparators.

Fig. 2 shows a variety of ways that the COR input can be interfaced to the microcomputer. By reviewing each configuration, you should be able to determine which circuit is compatible with your receiver COR. The circuit of Fig. 2A is for use with a receiver COR output that is pulled to ground when active. It does not supply a voltage when no signal is present. Examples of this type of COR are open-collector-transistor types and relays. The circuit of Fig. 2B can be used with a COR output that is pulled to ground when active, but which has a positive voltage on its terminals when it is inactive. This type of output could be from 12-V CMOS logic, 5-V comparator circuitry or from some transistors.

A positive-logic COR output requires an interface circuit such as is shown in Fig. 2C. With this arrangement, the output is at ground potential when inactive and at a positive voltage when active. An npn switching transistor keys Q1 in this case. R5 forms a voltage divider for the COR input. It also serves to maintain a stable off condition even if the receiver COR is disconnected. This will prevent the transmitter from turning on in the event of a COR cable failure.

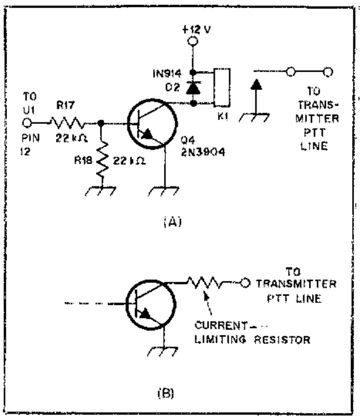


Fig. 3 — PTT driver transistor Q4 can interface directly to some transmitter circuits. Normally open contacts on a low-current relay can be used to close the transmit PTT line, as shown at A. The diode, used to protect the transistor from the inductive kickback that occurs when the relay is turned off, can be any small-signal switching diode, such as a 1N914. B shows Q4 used to drive a solid-state transmitter directly. The series current-limiting resistor value should be determined for your particular transmitter.

Some receiver COR outputs switch between negative and positive voltage levels. Fig. 2D illustrates how this type provides reverse-current protection for Q1.

One auxiliary COR input is available on the Microcomputer Repeater Controller. This input has an effect similar to the regular receiver COR, except that the time-out function and courtesy beep are not included.

The auxiliary input can be used with an autopatch system. Most autopatch circuits are self contained and provide a digital transmitter-enable signal. This signal is often wired in parallel with the COR or PTT line. When the autopatch has received the proper access code it will turn on the transmitter. By connecting the autopatch control line to the auxiliary COR input, the transmitter will be turned on, the courtesy beep will be disabled and the time-out timer on the controller will be disabled. The autopatch control circuit will handle the autopatch time limit.

PTT Interface

Pin 12 of the microcomputer chip is the port used to key the transmitter PTT line. Driver transistor Q4 can be used either as

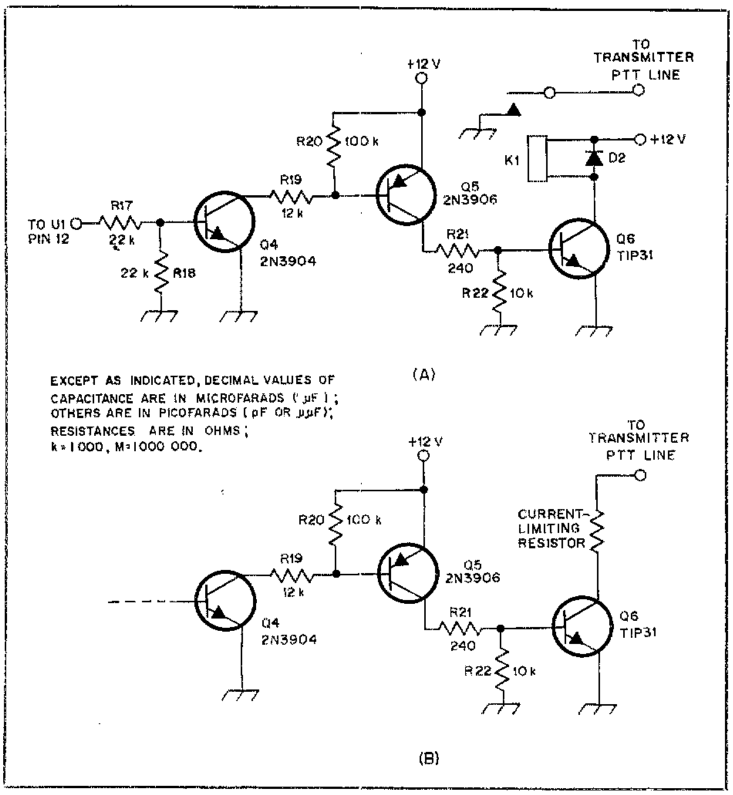


Fig. 4 — High-current PTT demands can be met by using a high-power npn transistor for Q6. A relay can be used, as shown at A, or a current-limiting resistor can be added to allow direct keying, as shown at B. The resistor value will depend on your transmitter.

a direct PTT-line interface or to drive Q5 as a keying transistor for the npn power transistor, Q6. By giving special attention to interfacing problems during the design process, I have provided a circuit that can be adapted to a wide variety of keying requirements. Fig. 3 shows two examples of PTT interfacing using the low-current transistor. If the PTT current requirement is higher than the 2N3904 can handle (200 mA), then Q5 and Q6 can be included to drive a larger relay or key the transmitter directly. The TIP-31 transistor will drive a 12-V PTT line at up to 1 A. Fig. 4 shows the circuit arrangement for two high-current interface configurations. Be sure to use an adequate heat sink with the high-current interface circuit. If the heat sink gets too hot to hold your finger on, increase its size to provide extra cooling.

An auxiliary PTT line is provided in the design. This digital control bypasses all timer functions, and holds the transmitter on for as long as the input line is active. This feature can be used with a voice i-d tape or other devices. When it is time for a voice i-d, the tape machine activates the auxilary PTT input, which turns on the transmitter and holds it on until the tape machine releases the control line. None of the normal repeater functions will be affected.

To bypass the receiver COR control and turn off the transmitter locally or remotely, two PTT-disable levels are provided. The first level, called a "soft" disable, can be used with a DTMF decoder for remote control. The actual PTT disable is performed by the computer software. Fig. 5 shows two examples of interfacing the soft disable to remote decoders. When the control pin is pulled to ground the computer will stop all timing functions and turn the transmitter off.

The second level of PTT disable is a hard-wired control. A switch is used to break the electrical connection between the controller and the transmitter. This is a local control, and should be used to prevent the transmitter from being turned on while someone is working on it. Fig. 6 illustrates the wiring for this control. Two wires connect between the pc board and an spst switch that is located on a control panel or other convenient location.

Resetting the Computer

Problems such as static discharges from lightning, power-line transients or low voltage can cause the microcomputer to stop or to begin executing what seems to be a different set of instructions. To get the computer back to normal, a reset line has been provided on pin 4 of the IC. This line is normally high. Grounding it will cause the computer to be reset.

Fig. 7 shows how I interface the reset line to a 12-V CMOS DTMF decoder. The diode in series with the line provides compatability between the 12-V decoder signal and the 5-V computer control line.

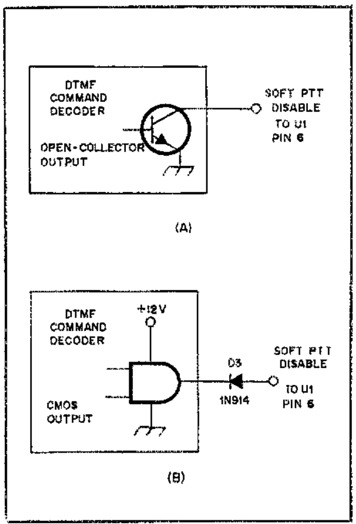


Fig. 5 — A DTMF decoder that uses an open-collector output can be connected directly to the soft PTT-disable line, as shown at A. For a decoder that supplies a positive voltage when inactive, a diode, such as a 1N914, should be added in series, as shown at B.

It is a good idea to provide a local reset control, and Fig. 7 also shows how this can be included. A momentary push-button switch is all that is required. The series diode prevents the decoder output from being damaged when the switch is depressed.

Power Supply

The two IC chips in the controller have different power requirements. The 8748 microcomputer requires 5 V at a maximum current of 165 mA. The tone generator can operate from a supply voltage of from 10 to 26 V, at 25 mA maximum current. Since most repeaters have a source of 12 V, I included a 5-V regulator for the microcomputer supply, and power the tone generator from the repeater supply. You will have to use a heat sink on the 5-V regulator to prevent the operating temperature from going too high. The supply voltage is not critical, and can vary from about 10 up to 16 without causing any problems.

As a final note on construction, I recommend that you use some type of connectors between the controller and your repeater-equipment lines. Then the pc board can be removed quickly, or lines disconnected for troubleshooting purposes. Repair or modification can be difficult if the wires are soldered directly to the pc boards.

Software

The repeater controller program gives "personality" to the microcomputer hardware. This is where you tell the computer what to do and when to do it. An object code listing of the RPT-1 version Ø software is available from the author. (Send an s.a.s.e. for information.) This version does not include the User Module software option, so you will not be able to use this unit to change system parameters. If you purchase the microcomputer chip from Processor Concepts, version 1 software is programmed into it. This is the version designed to work with the User Module.

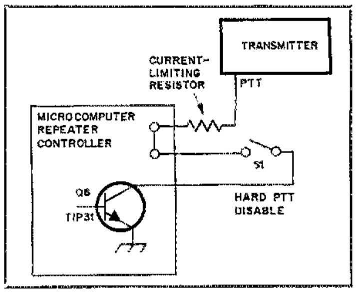


Fig. 6 — A typical PTT wiring scheme includes a hard PTT disable between the output keying transistor and the transmitter line.

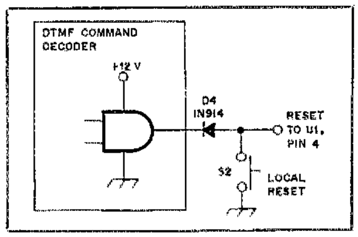


Fig. 7 — The microcomputer reset input can be wired to both a local push-button switch and a DTMF decoder output.

If you decide to generate your own program, I have a few suggestions to make your task a bit easier. First, make a list of all the functions that you want your controller to perform. This list should include requirements such as "Generate a cw i-d" or "Give a courtesy beep." The idea of this list is to reduce the chances that you will complete the program, only to discover that you forgot the cw i-d or some other important feature.

Once you have listed your requirements you must generate a flow chart. This is like a map of the program, showing what activity will be done in which sequence. Spending some time developing the flow chart will make your programming efforts flow much smoother.

The final step is to actually write the program. It is extremely important to use comments in your listing. A short statement

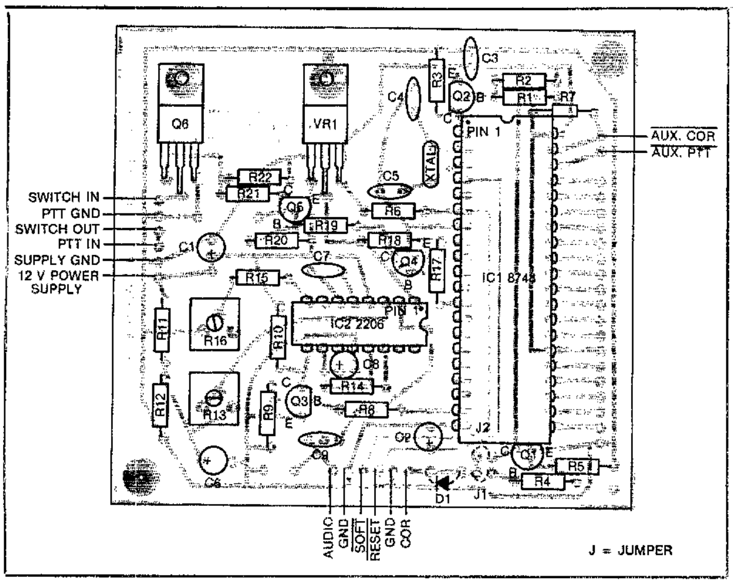


Fig. 8 — A parts-placement diagram for the Microcomputer Repeater Controller. The board is shown from the nonfoil side, and gray areas represent an X-ray view of the unetched copper. The full-size etching pattern is given in the Hints and Kinks column.

next to a set of instructions will tell you at a glance what that section does. If your comments closely follow the flow chart, it will be easier to follow the program and locate a particular activity.

Conclusion

The Microcomputer Repeater Controller is designed to be a simple, low-cost, easy-to-interface repeater controller. It makes a self-contained package that will bring you and your repeater into the computer age.

Notes

[1]mm = in. × 25.4.

[2]Assembled and tested units or complete kits for the Microcomputer Repeater Controller are available. Assembled User Modules can be purchased from Processor Concepts. An etched and drilled circuit board for the controller is available for $17 postpaid, and programmed microcomputer chips cost $49.95. Complete kits cost $89.95. For complete ordering and pricing information send a large s.a.s.e. to: Processor Concepts, P.O. Box 185, Fort Atkinson, WI 53538. The ARRL and *QST* in no way warrant these offers.

[3]See note 2.

Scott Freeberg has been licensed since 1967, at the age of 14. He holds a BSEE from Milwaukee School of Engineering. He is presently employed by Norland Corporation, designing microcomputer heart pacing systems and nuclear-medical analyzers. His Amateur Radio interests include 2-meter fm, 20-meter RTTY and 10-meter ssb.

SEASON'S GREETINGS FROM THE HAMS AT ARRL/IARU HQ.
(Listed in alphabetical order of call sign)

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Naoki Akiyama	N1CIX/JH1VRQ	Edith Holsopple	KA1KRQ	Leo D. Kluger	WB2TRN
Laird Campbell	W1CUT	Brian Downey	WA1KSF	Mark J. Wilson	AA2Z
George Grammer	W1DF	Bernice Dunn	KA1KXQ	Christopher Imlay	N3AKD
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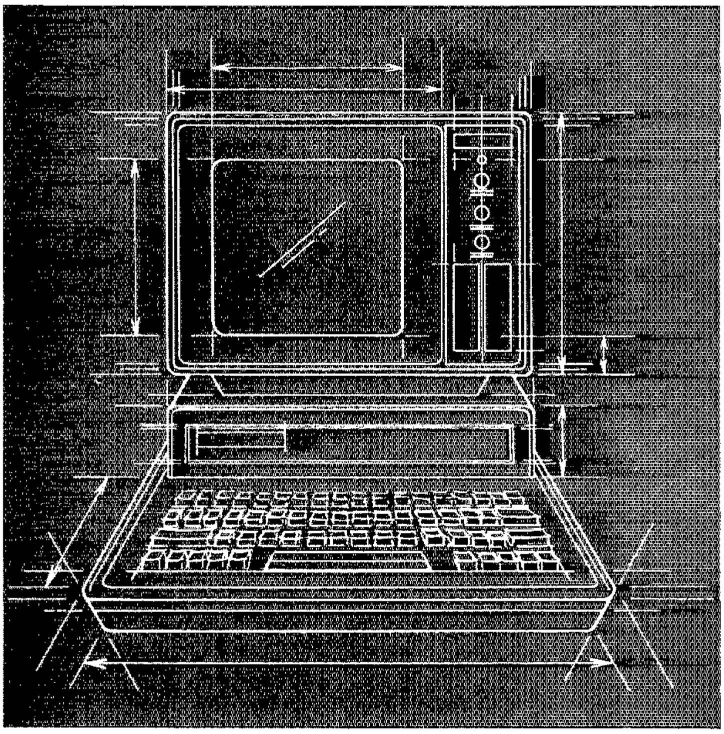
Elizabeth H. Karpiej	KA1DTU	Stan Horzepa	WA1LOU	W. Dale Clift	WA3NLO
Joan Merritt	KA1DTV	Phil Accardi	AJ1N	Larry Wolfgang	WA3VIL
Maureen Thompson	KA1DYZ	Peter R. O'Dell	KB1N	William A. Tynan	W3XO
Hal Steinman	K1ET	Sally H. O'Dell	KB1O	Steve Ewald	WA4CMS
Stephen C. Place	WB1EYI	Mike Kaczynski	W1OD	Paul Rinaldo	W4RI
Paul K. Pagel	N1FB	Bruce Kampe	WA1POI	John Troster	W6ISQ
Doug DeMaw	W1FB	Richard L. Baldwin	W1RU	Wayne Yoshida	KH6WZ
Marian Anderson	WB1FSB	John Huntoon	W1RW	Bob Schetgen	KU7G
Marge Tenney	WB1FSN	Lee Aurick	W1SE	Chuck Chadwick	K8AXL
John Nelson	W1GNC	Gerald L. Hall	K1TD	Chuck Hutchinson	K8CH
Bill Webb	WB1GOO	Perry Williams	W1UED	Jeff Ward	K8KA
Bob Atkins	KA1GT	Arline Bender	WA1VMC	Curt Holsopple	K9CH
Ed Tilton	W1HDQ	Greg Bonaguide	WA1VUG	Jim Clary	WB9IHH
Steffie Nelson	KA1IFB	Bill Jennings	K1WJ	Bernard D. Glassmeyer	W9KDR
Joan Becker	KA1IFO	Chuck Bender	W1WPR	B. Robert Benson	VE2VW
Jean Peacor	K1IJV	Bob Halprin	K1XA	Harry MacLean	VE3GRO
Cheryl Sowers-Clift	KA1IXI	John Lindholm	W1XX	Maxim Memorial Station	W1AW
Andrew Tripp	KA1JGG	Sandy Gerli	AC1Y	ARRL Hq. Station	W1INF
Jodi McMahon	KA1JPA	Ellen White	W1YL/4		
Lori Weinberg	KA1KOW	David Sumner	K1ZZ		

The Personal Computer†

Part 2: Peripheral devices allow a computer system to operate efficiently. Such peripherals as disk drives and printers are worth considering when you put together a personal computer system.



The role of a computer is to obtain data from an input device(s), process the data, and deliver the final results to an output device(s). To perform any useful task, a computer must interact with the outside world. A computer peripheral is any device that is attached to a computer for the purpose of getting useful information into and out of the computer's Central Processing Unit (CPU).

A peripheral is defined as being external to, or away from, the center. The name "peripheral" was selected to designate an input or output (I/O) device external to the CPU. Originally, these devices were physically separated from the CPU. However, since the advent of microprocessor chips, this is no longer the case. Now, complete computer systems, including CPU, memory, and peripheral interfaces are being packaged on single printed-circuit boards and as single ICs. Therefore physical placement can no longer be used to distinguish computer peripherals from the CPU.

This should not cause much of a problem if we think of peripherals in terms of their function rather than their placement. Some devices (printers, disk drives, terminals, etc.) can clearly be defined as computer peripherals, but what about switches, lights, motors, internal RAM and ROM, etc.? These types of devices are being used more and more with microcomputers and personal computers since new and different applications are being thought of and introduced. If you consider the function of these "gray area" devices, they must be categorized, at least for the purposes of this discussion, as peripheral devices, since they are input or output devices.

To attach certain peripheral devices (keyboards, CRTs, printers, tape and disk drives, modems, etc.) to a computer and make them functional, two things are required: interface (hardware and software for device control and communication), and operating system support. Generally, the more complex the peripheral, the more complex and costly the required hardware and software support packages. Some simple peripherals, like discrete I/O devices (switches, lights, etc.), require minimal hardware interfacing and no device-driver routines or operating system support.

What Is Required

All personal computers require at least two peripherals: (1) some amount of memory (RAM and ROM), and (2) a keyboard. Some come with a CRT display and a tape or disk drive, while others provide only the hardware interface and software drivers, and you must furnish the peripheral devices. An area that should be examined closely before making a purchase is whether or not you have definite plans to add peripherals to your system at some future date. Because, as so many personal computer owners have discovered (unfortunately too late), their needs and desires have changed after using the system and becoming familiar with it. Generally, the more peripherals with a system, the more useful the system will be. Some of the more common peripherals you may want to add to your system are:

- Printer
- Additional memory (RAM, EPROM)
- Mass storage device (hard disk, floppy disk or cassette tape)

†Adapted from *Personal Computer*, a Combustion Engineering Corporate Technology Newsletter, published by the Electronics Technology Applications Center, Combustion Engineering, Inc., 44 South 122 East Ave., Tulsa, OK 74128. The individual articles were written by several C-E employees in an internal newsletter for the benefit of the corporation.

• Color video monitor
• Audio response unit (speaker)
• Modem
• Joysticks
• Discrete I/O devices (switches, lights, etc.)
• Analog I/O devices (potentiometers, meters, etc.)

The thing to look for here is how difficult and costly it is to add these devices to your system. Some manufacturers provide spare slots for plugging in hardware-interface boards. These may be optional boards offered by the manufacturer, boards offered by other vendors or, if you really know what you are doing, a special-purpose board that you design and build for interfacing with a peripheral device. Other manufacturers provide only a connector for plugging in an interface board or card rack.

In addition to the hardware-interface circuitry required to connect the peripheral device with the computer, you may also need an I/O device-driver routine to control, synchronize, and format the data to or from a particular hardware interface or device controller. Usually, these I/O device drivers, sometimes called I/O handlers, are provided with the software package that comes with the unit, or they are offered optionally by the manufacturer or other vendors.

I/O Types

There are three basic types of I/O classified according to the method of controlling and synchronizing data transfer to peripheral devices: program-controlled I/O, interrupt-controlled I/O and direct memory access (DMA) I/O.

Program-Controlled I/O

With this type of I/O, the CPU is in complete control of handling the input or output transfer of data. Control and status words are issued and read by the CPU to synchronize and control the transfer of data. Address and data words define what the data is and where it goes (memory location or device).

Interrupt-Controlled I/O

This type is used in cases where an input needs to be serviced immediately or occurs very seldom. When an interrupt input occurs, the CPU stops what it is doing, services the interrupt, then goes back to what it was doing before it was interrupted. Servicing the interrupt usually involves branching to what is called an interrupt service routine, to perform some task related to the input.

Direct Memory Access (DMA) I/O

DMA transfer of data is used with I/O devices, such as disk drives and high speed line printers, which require rapid rates of data transfer. The information is transferred directly between the I/O device and memory without microprocessor intervention. The data transferred is controlled by a dedicated, high-speed logic circuit (DMA controller), capable of operating at higher speeds than the microprocessor.

Hardware Interface

Modularity has become an important consideration with all computer manufacturers and many computer buyers. This concept enables computer systems to be expanded or altered by adding or changing modules or peripheral devices. To make systems modular it was necessary to develop standard "computer bus" structures, which permit other manufacturers to develop compatible products that may be used with any microcomputer. Some of the more common standard bus structures are: S-100, IEEE 488, STD and Multibus. The purpose of interfacing is to match the peripheral device to the computer systems bus in the following characteristics: data speed, control codes, data format and electrical properties.

The format for interfacing data between computer and peripheral may be in serial or parallel form. Parallel data transmission is the transfer of one word of data at a time between the CPU and peripheral device. Although parallel transmission is much faster than serial, it requires more wires, one for each output data word bit.

Serial data transmission is the transfer of one bit of data at a time. It is used in cases in which data is transmitted over great distances and would be impractical to use many wires, for example, with remote terminals or modems. Some of the more common interface standards used for serial data transmission are RS-232-C, RS-422 and current loop. These interfaces establish required control codes and electrical properties.

Serial data may be transmitted either synchronously or asynchronously. Synchronous transmission is much faster than asynchronous, but is more critical as far as synchronizing the CPU with the peripheral device. Synchronous transmission is usually used on larger systems where rapid transfer of data is required.

Synchronous transmission is the sequential transfer of many characters or a block of data at a time. Characters are usually ASCII coded. The entire block of data is synchronized by timing signals. Asynchronous transmission is the transfer of these serial data bits, one character at a time. The characters are usually ASCII coded and bounded by start and stop bits.

Serial data may be transmitted either in a full- or half-duplex mode. In full-duplex mode, data may be transmitted and received simultaneously, whereas in half-duplex it can only do one at a time. Although full-duplex is the most desirable since it is faster, it also requires more wires, one set for transmit and another for receive, so it is not always feasible.

An important factor in the determination of the cost of a personal computer is the method of storing its software (programs and data). No matter how large the memory capacity of a personal computer may be, you will have more information available for use than can be contained within its memory at one time. It is necessary, therefore, for all systems to provide a means of saving programs and data for later use so the computer's memory capacity is made available for other uses. This storage method must allow the program or data to be reloaded into the computer when the program is to be used or when the data is needed.

Most low-cost personal computer systems use an ordinary cassette tape recorder to store and retrieve both programs and data used by the computer; some models use a modified tape recorder. This is an advantage because the tape recorder need not be supplied (and is frequently not included in the system price), or if it is, it is a low-cost peripheral item that adds little to the system cost. Since most systems use audio tones to place the data on tape, ordinary low-noise recording tape may be employed to store programs. This combination of low peripheral cost (the tape recorder), and low media cost (the cassette tapes), makes cassette-tape storage desirable, and almost all personal computer systems will accommodate cassette data storage. Indeed, some of the very low-cost computer systems are available only with cassette storage.

Cassette storage is popular because of its low cost. However, it exhibits some deficiencies. For example, it is slow compared with other methods — it may take several minutes to load a large program into memory each time it is needed. While this is certainly better than reentering these programs by the computer keyboard, it makes applications requiring frequent swapping of programs or data virtually impossible. Also, most cassette-storage systems offer no help in locating a program once it is placed on tape, and it may be necessary to read (play) an entire tape to find a single program. This problem may be solved by placing only one program on each cassette, but this can make the cost of data media (the cassettes) very expensive.

Disk Storage

Floppy disks: An alternative to cassette-tape storage provided as an option with most personal computer systems is flexible diskette storage, commonly known as "floppy-disk" storage. Floppy disk systems consist of one or more disk-drive assemblies, the floppy disk and any necessary circuits and cables to make it function with the personal computer. The disk drive performs the same function as a tape recorder in a cassette-based system — it reads and writes programs and data to the floppy disk.

The disk resembles a small, thin

phonograph record housed in a paper jacket and is inserted into the drive through a slot in the front of the unit. Data is recorded magnetically on the disk (in a fashion similar to audio-tape recording) as the disk spins continuously past the disk-drive head assembly. The head assembly (similar in function to a record-player arm) allows the drive to record or read data anywhere on the disk. The associated electronics allow the computer to transfer data from its memory to the disk, or from the disk to the computer.

The floppy-disk system of data storage has a number of advantages over cassette storage. A most obvious advantage is that data transfer is much faster, with large programs being loaded or saved in seconds instead of minutes. The large storage capacity of the disk allows a number of programs to be stored on a single disk. The average cassette tape can hold large quantities of data, but since it is not practical to search the entire tape for a single program, much of its theoretical capacity is often not used.

Floppy-disk systems have an additional feature that makes storage of several programs on a single disk practical. Most disk systems contain features that catalog the various programs contained on the disk. This permits all files to be given descriptive titles and allows verification of whether or not a particular title exists on the disk. Program loading and saving is also accomplished by similar title reference. With this method, it is no longer necessary to be concerned with where a particular program is stored on the disk, as the operator can refer to it simply by title.

Another advantage of disk storage is that the media is accessed randomly. This means that every portion of the disk is available to the drive at all times, without the need to move the media forward or backward as with a cassette. The disk spins constantly, and the head can move immediately to any place on the disk. When the disk is removed and replaced with another, no attention need be paid to media movement to access data.

Floppy disks can cost more than cassette tapes, but the price is decreasing steadily. Their additional usable storage capacity, as well as their added convenience and flexibility, make them cheaper to use per program stored. They are, however, more fragile than cassettes, and require greater care in handling and storage. The access holes in the jacket expose the media to damage, and care must be taken to ensure that the diskette is not touched in these areas. The disks must be stored and transported in such a manner that they cannot be bent or creased, as this will make the media permanently unusable. If precautions are taken, disks will provide reliable operation for a long time.

Perhaps the biggest disadvantage of using floppy-disk storage systems is the initial cost of the disk drives and interface circuitry. In a typical personal computer system, the initial cost of disk storage is from $300 to $900 per drive. This is a significant cost factor, as many applications require the use of at least two such drives. The cost of a disk-storage system can exceed the cost of the personal computer itself. For this reason, cost comparisons of personal computers should always be made with the desired storage systems (as well as any other peripherals not included in the base price) taken into consideration.

Hard disks: A few applications of personal computers, such as those that involve handling large quantities of data, require storage capacities that exceed those of flexible disk systems. These systems utilize rigid disk media, and are usually referred to as "hard" disks. A hard disk is quite rigid both in composition and in mechanical attachment to the drive. This lack of movement allows these disks to store data in a compressed form, allowing large quantities of data to be contained on a single disk. Because of the mechanical precision required, most of these systems do not have removable disks. This means that systems that use hard disks rely on other storage systems to save (back up) programs against disk failure, and to store programs not currently in use.

Hard-disk storage systems are expensive. A single-drive hard-disk system can cost $1000 to $5000, depending on features and storage capacity. To this cost must be added the expense of the required back-up system. In spite of this, hard disks are advantageous in some systems because they have a much larger storage capacity and are operationally much faster than floppy disk systems. They are currently too expensive to be used for most personal computer applications.

Summary

Peripheral devices are worth considering for any computer system. They expand the system capability, making it more useful and versatile. Before making a purchase of any computer, consider all aspects of adding peripherals to your system: cost, effort, additional space and power requirements, and hardware and software support packages.

New Products

EMERGENCY COMMUNICATIONS: AN ORGANIZATIONAL AND OPERATIONAL HANDBOOK

by F. Dale Williams, K3PUR. Published by FDW Arts, P.O. Box 2540, Falls Church, VA 22042. Soft-bound, 6 × 9 inches, 175 pages. Single copies are $9.95 plus $1.50 postage and handling.

□ According to the history books, Amateur Radio operators have been providing emergency communications since 1913, and since then a magnificent record of accomplishment has been built up, most recently shown in the service provided during Hurricane Alicia. Such volunteerism on the part of radio amateurs is an even more vital asset in this era of limited government resources. Substantial opportunities now exist for contributions by organized volunteers such as the League's Amateur Radio Emergency Service (ARES), as well as the Radio Amateur Civil Emergency Service (RACES). Author (and publisher) Williams's effort is certainly timely.

This publication delivers exactly what the title denotes — a comprehensive guide to organizing an Amateur Radio emergency communications group and efficiently employing that group in times of need. Subjects given in-depth treatment include the kinds of Amateur Radio public service; liaison with government and relief agencies; the art of communicating; how to organize the Amateur Radio public service group; appropriate equipment for both fixed and field operations; proper procedures for emergency and public service activities; and effective drill/test scenarios. Four appendices provide additional information.

The author is well-qualified to add to the emergency communications literature. K3PUR has been a licensed amateur for more than 20 years, and his credentials include serving in a variety of leadership roles in ARES and RACES, most recently as ARRL Section Emergency Coordinator for Colorado. He also has worked extensively in the area of emergency/disaster communications with government and public service agencies, and has served in administrative and advisory positions with the American Red Cross.

As awareness grows that Amateur Radio has a unique ability to survive when commercial communications break down, we, as amateurs, must rise to that challenge. *Emergency Communications* can be an important tool in meeting that challenge — *Robert Halprin, K1XA*

The Extended-Element Beam

Using only three elements, this beam has the gain of a 6-element Yagi array. Extending the elements is the key to such success.

By Richard C. Fenwick,* K5RR, Richard C. Fenwick, Jr.,* N5BXB and Bobby Schroeder*

It is well known that the optimum length of a dipole antenna for maximum gain is approximately 5/4 wavelength. Such an antenna is commonly called an "extended double Zepp."[1] The radiation pattern is as shown in Fig. 1, and the gain is 3.1 dB over a half-wave dipole (3.1 dBd). It does not seem to have been recognized that extended-double-Zepp elements can be utilized in parasitic arrays. Such use results in a compact, high gain antenna, with characteristics generally similar to ordinary Yagi antennas, but with 3-dB greater gain and narrower beamwidth for the same number of elements. This has been verified both analytically and experimentally for 3-element arrays, shown schematically in Fig. 2.

An extensive experimental investigation was conducted on a 3-element beam designed to operate at 280 MHz (Fig. 3). The model was designed to permit exploring a range of element spacings and element tuning. Element tuning was accomplished by incorporating variable capacitors in parallel with the fixed inductors shown in Fig. 2. Element length is not at all critical; all elements were made 51.3 inches, or 1.22 wavelengths, long. The driven element was fed through a coiled coaxial type of balun, ensuring balanced feed. The radiation pattern of the driven element alone was observed to be virtually identical with what was expected theoretically (Fig. 1). E- and H-plane patterns and impedance measurements were made on the model for director and reflector spacings of 0.15 and 0.2 λ respectively, 0.175 and 0.225 λ, and 0.2 and 0.25 λ. The antenna was tuned for two conditions at each spacing — maximum gain and maximum front-to-back ratio. The results generally correspond to those expected for ordinary parasitic arrays.

Effect of Element Spacing

The wider spacings gave reduced gain and front-to-back ratio and increased bandwidth compared to the 0.15- and 0.2-λ director and reflector spacing. Table 1 gives a comparison of these characteristics for the three spacings. Closer spacings were not explored owing to the increasingly narrow bandwidths encountered, although the maximum front-to-back ratio would presumably be improved. Reduced bandwidth is the price paid for the extended-element beam advantages of compactness and mechanical convenience when compared to two ordinary Yagi antennas arrayed side-by-side and spaced at 1 λ. An extended-element beam performs very similarly to such an array in all respects except bandwidth.

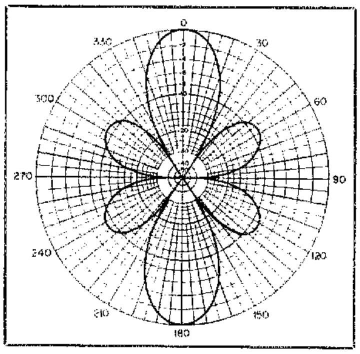


Fig. 1 — Free-space directive diagram for the extended double Zepp. This array has a gain of 3.1 dB over a dipole. (In this plot, the radiator lies along the 90°-270° line.)

Performance

Fig. 4 summarizes the front-to-back ratio, gain and VSWR observations for 0.15- and 0.2-λ director and reflector spacing. Increased gain with increasing frequency is a familiar characteristic of all Yagi antennas. The bandwidth over which the front-to-back ratio is high is less for the extended-element beam than with ordinary Yagi antennas, judging from Lawson's epic work.[2] If the maximum gain of an ordinary

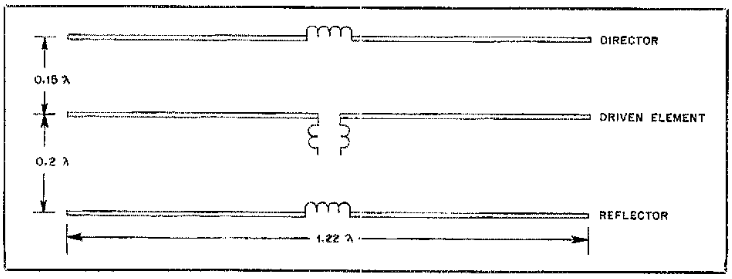


Fig. 2 — Extended-element beam, with suggested dimensions.

[1]Notes appear on p. 37.

*c/o Electrospace Systems, Inc., Richardson, TX 75083-1359.

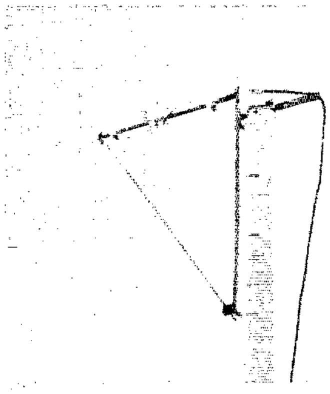


Fig. 3 — Extended-element beam model, designed to operate at 280 MHz.

3-element Yagi is 7 dBd, then we'd expect the maximum gain of an extended-element beam to be 10 dBd. This is very nearly what was observed. A gain penalty of about 0.7 dB is incurred when choosing tune-up for maximum front-to-back ratio rather than maximum gain, but often this will be the preferred choice, especially if bandwidth is a consideration.

Fig. 5 shows measured radiation patterns for the maximum gain and maximum front-to-back conditions. Lack of symmetry in the back lobe is believed to be caused by imprecision in model construction.

Tune-up Procedure

Considerable experimentation was carried out to determine the optimum tune-up procedure. Whether tuning for maximum gain or maximum front-to-back ratio, it was found that the director should be tuned first, with the reflector removed or open circuited. The director is tuned for maximum signal off the front or minimum signal toward the rear, as desired. Then, the reflector is tuned for maximum effect. The process converges very rapidly, such that it is not absolutely necessary to retune the director after tuning the reflector. However, this was sometimes found to improve the results. A convenient method that has been used successfully for tuning an antenna for maximum front-to-back ratio is to lash the antenna onto the side of a tower, with the reflector closest to the ground and at least λ/2 above ground. Then transmit from a dipole near ground at the base of the tower. The antenna under test is connected to a receiver, and the director and reflector tuned sequentially for minimum S-meter reading.

The suitability of the extended-element beam to this tuning approach is one of its advantages over ordinary Yagi antennas; the usual practice is to adjust element lengths to achieve the desired performance. A single ordinary Yagi requires six elements for 10-dBd gain. Optimizing such an antenna experimentally is clearly much more difficult than optimizing a 3-element extended-element array.

The driven element may be tuned and matched by conventional means quite in-

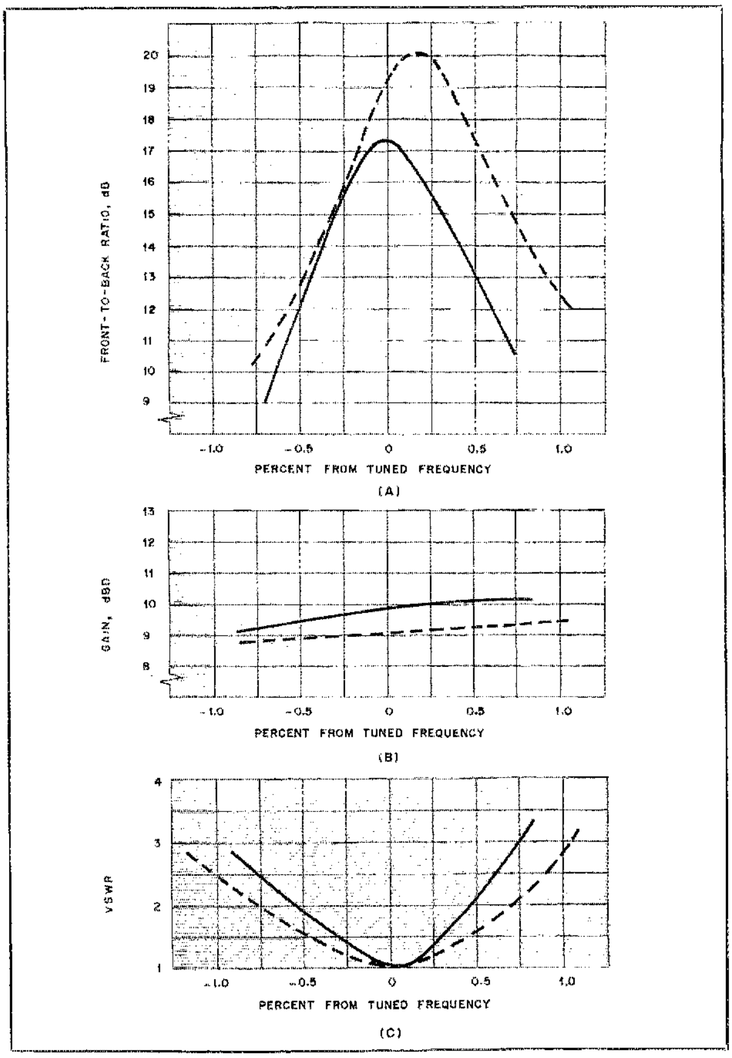


Fig. 4 — Performance of the extended-element beam with dimensions shown in Fig. 2. At A is the front-to-back ratio; at B, the gain over a dipole; and at C, the VSWR. In each plot, the solid line shows the performance with the beam tuned for maximum gain, and the broken line when tuned for maximum front-to-back ratio.

Table 1

Measured Performance of 3-Element Antennas

Dir. and Refl. Spacing (λ)	*Tuning Condition*	*Gain* (dBd)*	*F/B* (dB)*	*Input R* (Ohms)*	*Bandwidth for 2:1 VSWR (%)*
0.15 and 0.20	max. gain	9.8	17.3	25	1.0
0.15 and 0.20	max. F/B	9.1	19.2	34	1.5
0.175 and 0.225	max. gain	9.5	14.9	34	1.3
0.175 and 0.225	max. F/B	9.0	15.9	39	1.8
0.20 and 0.25	max. gain	9.5	9.3	31	1.1
0.20 and 0.25	max. F/B	8.2	13.8	51	2.1

*At tuned frequency (280 MHz)

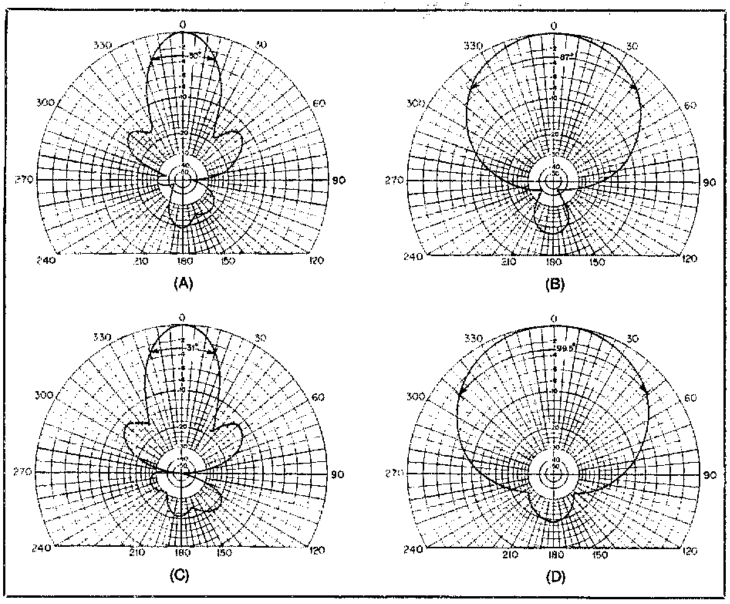


Fig. 5 — Measured patterns of the extended-element beam of Figs. 2 and 3. E-plane patterns are shown at the left, and H-plane patterns at the right. The patterns at A and B were obtained with the beam tuned for maximum gain, while those at C and D were for the beam tuned for maximum front-to-back ratio.

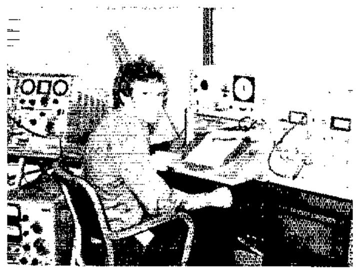


Fig. 6 — Pattern recording equipment, with author Schroeder.

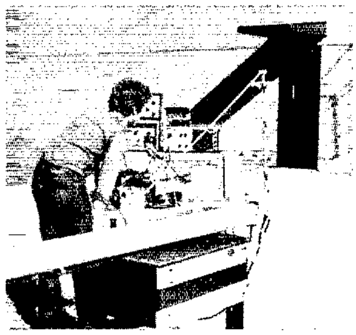


Fig. 7 — Impedance measurement in progress.

dependent of the tuning of the director and reflector. The patterns of the antenna are not a function of the driven-element tuning.

The values of the inductors needed to tune the antenna must be determined experimentally. As a start, it is suggested that the reactance for the director be about 400 ohms, and that for the reflector about 600 ohms. A range of adjustment of approximately ±50% from these values is recommended. Probably the easiest mechanical arrangement for variable inductors is a length of transmission line made of rod or tubing, with a movable shorting strap.

Method of Measurements

All model measurements were made on the Electrospace Systems, Inc., antenna pattern range (Figs. 6-8). The range was set up carefully to minimize ground reflections. Impedances were measured with a Hewlett-Packard network analyzer (Fig. 7). Frequency settings were quite critical, and a frequency counter was used for all measurements.

Gains were calculated from the radiation patterns using the familiar expression

$$\text{Gain} = 10 \log \frac{41{,}253}{\theta_E \theta_H} - 2.14 \text{ dBd}$$

where

θ_E is the E-plane 3 dB beamwidth, and
θ_H is the H-plane 3 dB beamwidth.

(Gain and directivity are assumed to be synonymous here.) A gain-correction factor was developed to account for sidelobes. As a check, gains were measured experimentally by rotating the parasitic elements 90° to the driven element and observing the decrease in signal at the receiver. It was concluded that the gains calculated from measured patterns are probably accurate to within 0.2 dB.

Care was taken to ensure that the presence of the tuning person's body did not affect the tuning of the antenna (Fig. 8). Even so, the tuning for maximum front-to-back ratio was critical, and it is entirely possible that higher ratios than we observed can be achieved. Some scatter in the data (Table 1) suggests that tuning was not always optimized.

Future Work

A modest attempt to evaluate a 4-element array was made; it seems logical that such an array would work. However, no significant improvement in gain over a 3-element array was observed. More investigation of such arrays is needed.

Extended-element beams of three elements each for 10 and 15 meters, mounted on the same boom, are under construction as of this writing. This project may be the subject of a future article.

Acknowledgment

Thanks go to Toney Magnino, W5MVK, who provided considerable assistance in the experimental work, most particularly in the construction of the antenna model.

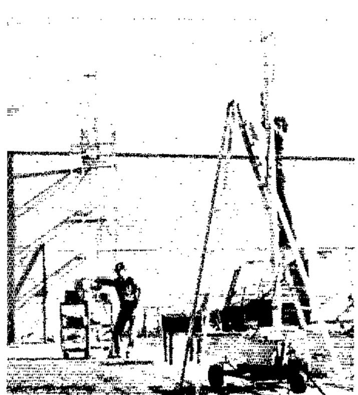


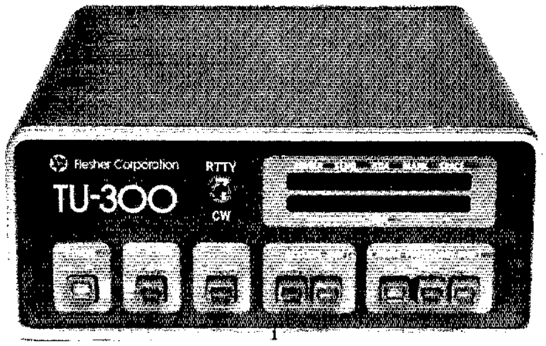
Fig. 8 — Antenna model being tuned on the pattern range.

Notes

[1]G. Hall, ed., *The ARRL Antenna Book* (14th Ed.). Newington: ARRL, 1982, page 6-8.

[2]J. Lawson, "Yagi Antenna Design," *Ham Radio*, May 1980, p. 22.

The TU-300 — Modified



Mating a computer, a TU and a transceiver can often be a stumbling block. This information should provide some ideas and prevent you from stubbing your toe!

By Paul K. Pagel,* N1FB

Many radio amateurs have joined in on the computer-buying spree and are sharing the fun of RTTY and cw operation using their new-found "toy." I recently joined the crowd.

Right from the start, the new Amateur Radio/computer operator discovers it's generally not a matter of "plug and play" (although we're getting there!). Sooner or later (with any luck sooner!) you realize that you should take some precautionary measures to keep the computer and any units matched to it from electrically "fighting" each other. Usually when such a fight occurs, the computer is the loser. As a fighter is prepared for a match, the computer must be prepared to protect itself during these matches.

In my shack, I use an Apple IIe® computer, a Flesher TU-300 modem and a veteran TS-820S transceiver. For the computer, a first line of defense occurs at the ac-line connection. I've installed a Drake FL-6 line filter between the ac line outlet and the computer ac line input.[1] This filter/outlet strip combination provides brute-force line-filtering to keep rf energy from leaving or entering the computer through the ac line. The filter also contains MOVs (metallic-oxide varistors) that suppress potentially harmful line-voltage transients.

When you are mating a computer to a transceiver or other peripheral equipment, accidental application of voltage levels harmful to the computer becomes possible. Isolation is the word to keep in mind, and optoisolators are a main ingredient in filling this requirement. These devices are simple to use, inexpensive and consume little power; they also provide a high degree of circuit isolation.

The optoisolators I use in the circuit consist of an LED that is optically coupled through a translucent insulator to a photosensitive output transistor. The insulating material between the LED and the photosensitive transistor has a voltage breakdown rating of about 7500 volts. Thus, it would require an exceptionally large voltage to puncture the insulating material for the two circuits to interact directly. If proper design procedures are followed, this type of failure should never occur.

At the Transceiver

Before attempting an interface to the computer, I decided to eliminate one concern — the −65 V level present at the TS-820S key jack. That potential has the potential (pardon me!) of ruining a transistor. It would be redundant to adapt every external device to be able to handle the high voltage present at the key jack. A better approach is to modify the transceiver to be compatible with the rest of the low-voltage outside world. The change I made removes the −65 V level from the key jack and places a positive 4.4-V potential there.

The simple circuit shown in Fig. 1 does the job. A single high-voltage pnp transistor is mounted on a three-lug terminal strip secured to the bottom of the 5-V regulator "bathtub" heat-sink by one of the heat-sink mounting screws. R1 is connected between the appropriate terminal lug and the 5-V foil of the regulator board. The end of R12 (220 ohms) attached to the key jack is removed and connected to the transistor collector tie point. Wire the base of Q1 to the key jack and install R2 as shown. The key jack voltage is now essentially TTL compatible and is active low.

On To the TU

Physical Considerations

I initially thought I would mount the interfacing optoisolators on a small pc board supported by a bracket inside the TU-300. This would keep the computer innards free of excess baggage. I'd also be able to derive the +5 V required for the

[1]Notes appear on page 40.

*Assistant Technical Editor, ARRL

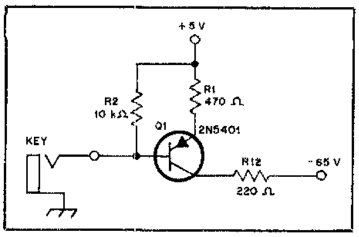


Fig. 1 — Schematic diagram of the TS-820S keying circuit modification. Q1 is a high-voltage pnp transistor. R12 is part of the original transceiver circuit. All resistors are ¼-W, 5% types.

optoisolators from the '300 supply.

It was at this time that I had a chance to see the TU-470, successor to the '300. I noted that the '470 had some added transistor switches that enabled switching of positive and negative polarity PTT or other such lines. This feature is missing in the '300. There, the mechanical SEND/REC switch is used to perform such switching manually. Why not incorporate the '470 circuit modification into my '300?

It then struck me that there were two expansion slots in my '300 that were doing nothing at the time. (One was eventually filled with the cw demodulator board.) I could put the interface and transistor switching circuits on a small plug-in board that would occupy one of the expansion slots.[2] That would certainly be a lot neater than a board secured to a chassis wall with a bracket.

The left-most expansion slot (8) was selected for the new board. Don't be tempted to use an empty filter-board slot. Those slots are connected to the signal buses. The expansion slots have no connections made to them other than ± 12 V and ground. This allows one to wire the other pins to any desired source or destination without the need for cutting pc-board traces.

Circuit Description

Fig. 2 is a schematic diagram of the interface and PTT-line switching circuits. I've retained the individual equipment plug and socket numerical designations for clarity. As shown in Fig. 2, I use two optoisolators to interface my computer with the TU-300. These devices are connected to the computer game I/O socket, J15.

The RTTY/cw software I have delegates J15 pin 15 (ANØ) as the RTTY/cw output port, pin 12 (AN3) as the T-R control line, and pin 2 (PBØ or SWØ) as the RTTY/cw input port. The outputs, once the program has initialized them, are normally high during a no-signal, receive mode condition. When AN3 goes low, the TU-300 SEND N line is grounded, putting the TU into the transmit state. ANØ then toggles the afsk/fsk circuit of the TU or the cw key line, depending upon the position of S1.

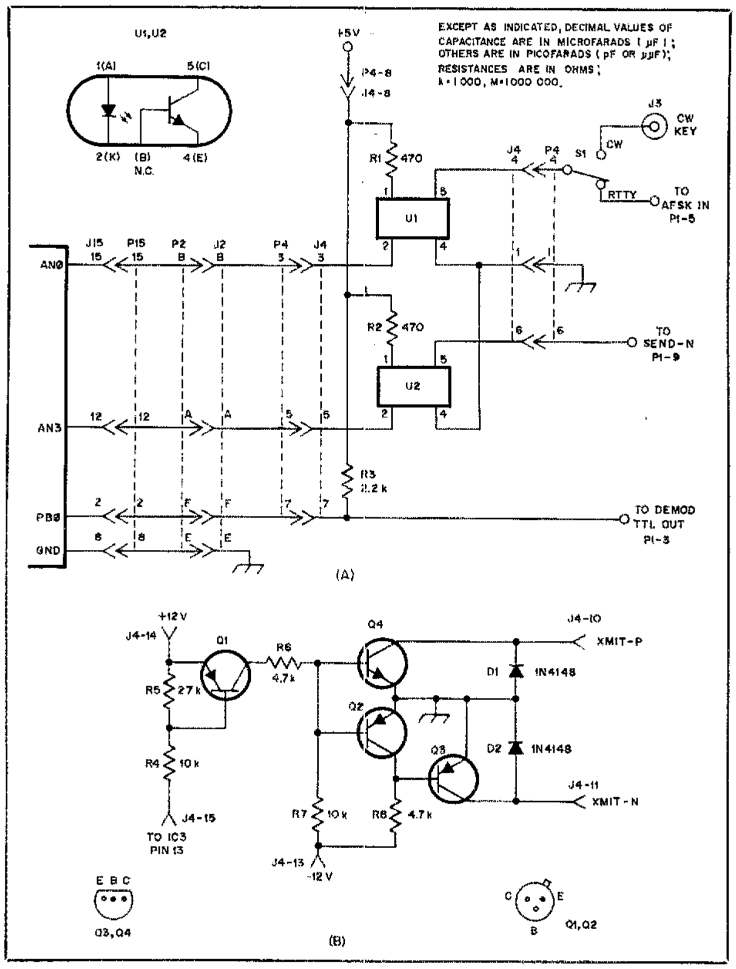


Fig. 2 — Computer/TU interface circuit and TU-300 keying circuit modification schematic diagram. All resistors are ¼-W, 5% types.

J1 — Part of TU-300 (DB-25P), not shown.
J2 — Amphenol 126-192 or equivalent.
J3 — Phono jack.
J4 — 15-pin socket, Molex 22-15-2151 (gold-plated contacts, 22-15-2156 (tinned contacts) or equivalent.
J15 — Part of computer (game I/O socket).
P1 — Part of TU-300 (DB-25S), not shown.
P2 — Amphenol 126-195 or equivalent.
P4 — 15-pin SIP plug, Molex 22-03-2151 or equivalent.
P15 — 16-pin DIP plug.
Q1, Q2 — 2N4125, 2N2907 or equivalent.
Q3 — MPSL51, 2N4888, ECG288 or equivalent (see text).
Q4 — MPSL01, ECG194 or equivalent (see text).
S1 — Part of 3pdt toggle switch (C&K 7301 or equivalent). The two poles not shown are used in the TU-300 cw/RTTY selection circuit.
U1, U2 — ECG3041 optoisolators or equivalent.

During receive, the incoming demodulated cw or RTTY signal toggles PBØ, the computer "does its thing" and displays the information on the screen.

The XMIT-P and XMIT-N outputs shown in Fig. 3 are keyed from two sources: the front panel SEND/REC switch, and the SEND N input. When either of these sources is keyed, IC3 pin 13 goes low. Q3 and Q4 are then driven into conduction and key the respective XMIT-N and XMIT-P lines.

The transistors used at Q3 and Q4 have a high collector-to-emitter voltage rating (about 100 V). If the lines you intend to key are low-voltage ones, you may substitute transistors having correspondingly lower ratings.

TU-300 Modifications

I replaced the front-panel RTTY/CW

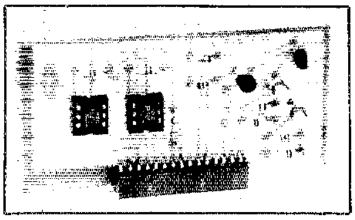


Fig. 3 — A view of the completed interface circuit board.

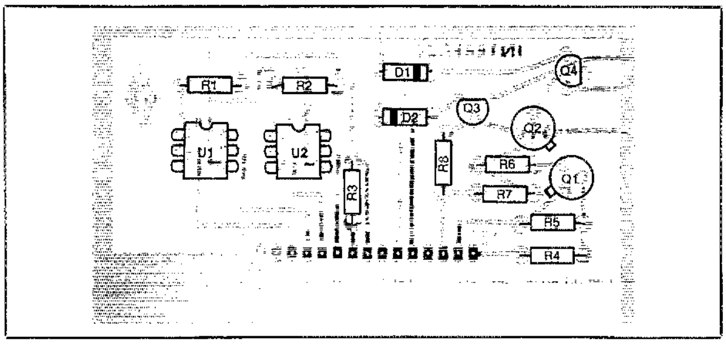


Fig. 4 — Parts-placement guide for the interface board. This view is from the component side of the board. Gray areas represent an X-ray view of the unetched foil. A circuit-board etching pattern appears in the Hints and Kinks section of this issue.

Table 1
Modification Interconnections

Slot 8 Pin No.	*Connected To*
15	IC3 pin 13
14	+12 V
13	−12 V
12	N.C.
11	XMIT-N
10	XMIT-P
9	N.C.
8	+5 V
7	P1 pin 3/J2, F
6	P1 pin 9
5	J2, A
4	S1
3	J2, B
2	N.C.
1	Gnd/J2, E

N.C. = not connected

selection switch (provided with the optional cw demodulator board) with a 3pdt toggle switch. The third pole of the switch is used as S1 of Fig. 2. Next, two foil cuts were made to adapt the XMIT-N/XMIT-P keying modification to the '300. (If this circuit addition is not to be made, these foil cuts are not needed.) The first cut is done on the top side of the board. Locate the trace that connects IC3 pin 13 to the fifth SIP (single in-line package) resistor lead. Cut this trace with a sharp instrument. To separate the TR-N line from the SIP resistor, another cut beneath the board is required. Find the trace that leads from D8 to the fifth SIP resistor connection. Cut this trace at a convenient point. Now connect the cathode of D8 to IC3 pin 13. No. 30 wire-wrapping wire is sufficient for this purpose and will serve to make most of the other modification changes.

I installed a phono jack in the upper-left rear-panel corner of the '300. This jack serves as the cw key line connector. A multi-pin socket is mounted at a convenient point next to the phono jack. This socket and its mating plug are used to interface the computer I/O lines to the TU. If your setup is similar to mine, you'll need at least four contacts on the connectors.

A 15-pin SIP male plug is soldered in place at expansion slot 8. A mating female pc-board socket, similar to those used on the Flesher boards, is required for the added board. After assembling the plug-in board, chassis wiring may begin.

Table 1 provides the information you'll need to wire the expansion slot to the rest of the TU. The slot 8 pin numbers are counted from the rear of the unit forward. (This procedure conforms to that used by Flesher in the cw demodulator modification.) Note that normally unused pins of P1 (actually a socket) are connected to the XMIT-P and XMIT-N lines. This leaves the SW1/SW2 mechanical switching arrangement undisturbed. Although the ±12-V leads are already connected to the expansion slot, you'll have to connect slot 8 pin 8 to +5 V with a jumper wire. When you have all the connections made, check for solder bridges, shorts and wiring errors. Having convinced yourself all is well, you can make the appropriate I/O wiring connections between the units and plug in the board.

Interfacing

I use a length of four-conductor ribbon cable with matching connectors to interface the computer and the TU. A length of multi-conductor wire connects the TU to the transceiver. This keeps most of the wiring in two neat bundles.

Perhaps your software calls for different I/O assignments at the game port. The program may also toggle low to high instead of the high-to-low toggle required for this setup. If that's the case, probably the easiest thing to do is to modify the software. Find the I/O address locations within the program and shift them to suit your requirements. I had to do this with another program I have to make everything compatible. I wasn't about to switch plugs and other hardware around!

Summary

I hope these ideas will be of use to you. The project is one I consider worthwhile. It has made RTTY/cw operating much more sophisticated (at least from a hardware standpoint!) at my station. The computer keyboard is the sole controlling factor during operation. RTTY and cw transmissions are silent except for the click of the keyboard, or sound of the cw monitor tone. What a pleasure!

Notes

[1] R. L. Drake Model LF2 and LF6 Line Filters, "New Products," *QST*, Jan. 1983, p. 27.
[2] A circuit board and complete kit of parts is available from Circuit Board Specialists, P.O. Box 969, Pueblo, CO 81002.

Strays

SSC LOGO CONTEST UPDATE

☐ Thanks for the great response to the Special Service Club (SSC) logo contest. There are only a few weeks left. The deadline for submissions is January 15, 1984. Get your entry in before it's too late. (Complete rules are in October *QST.*) — *Sally O'Dell, KB1O, Club Program Manager, ARRL*

TIME/FREQUENCY MANUAL AVAILABLE

☐ Haven't completed your Christmas shopping yet? How about a copy of *Time and Frequency User's Manual No. 559* for the ham who has everything? Compiled by the National Bureau of Standards, this 256-page publication will be of interest to the casual reader as well as the laboratory engineer. Send $6 to Superintendent of Documents, U.S. Government Printing Office, Washington, DC 20402. Refer to Library of Congress No. 79-600169.

Hints and Kinks

Conducted By Larry D. Wolfgang,* WA3VIL

AN AMPLIFIER FOR HIGH-IMPEDANCE VOLTMETERS

☐ When I read the Beginner's Bench article in the Jan. 1982 issue of *QST*, I became interested in the high-impedance voltmeter project described there.[1] I already own a Heath VTVM, so I did not need another meter, but the project offered a convenient starting point for gaining some experience with op amps. I decided to build an amplifier to use with my VTVM instead. Now I can measure very small voltages with the meter.

The circuit I developed is shown at Fig. 1. I used a 741 op amp to provide an artificial ground for the signal voltage, which permits the use of a single 9-V battery to power the amplifier. My goal was to provide two amplification ranges: × 10 and × 100. By using 5%-tolerance resistors, and allowing for the voltage-divider effect of the 1-MΩ resistor in the probe and the 12-MΩ unit at the amplifier input, the actual multiplication factors for my amplifier are 10.1 and 93. I decided that is close enough for my purpose.

The 2.2-kΩ resistor on the output of U1A provides a load for the op amp, as described in *The Radio Amateur's Handbook*.[2] I used the second half of the LF353N op amp to provide an adjustable zero function for my meter, as suggested in Collins' article. Ferrite beads were added as shown in Fig. 1 to cure a problem of self oscillation in my amplifier. The four beads that I used came from a "mystery" assortment that I had. Radio Shack beads (no. 273-098) should work. [Micrometals beads made from no. 73 or 75 material should also be effective. These are available from Amidon Associates, 12033 Otsego St., North Hollywood, CA 91607. — Ed.]

To use the amplifier, I plug the test probe from my Heath VTVM into the amplifier, plug its output into the VTVM, short the probe to ground and zero the meter using the amplifier zero control. The circuit is very sensitive in the × 100 position. Just twisting around in my chair with the probe in my hand is enough to send the meter movement into fits from static buildup.

I have discovered some interesting applications for a meter with an amplifier like this one. For example, a signal as small as 0.01 V will deflect the movement in my VTVM. When the amplifier is connected and set to the × 100 position, this becomes 0.0001 V, or 100 μV! Now it is possible to measure small currents flowing through a circuit — 0.1 μA through a 1-kΩ resistor can be detected with no trouble.

*Assistant Technical Editor, ARRL

[1] G. Collins, "Some Basics for Equipment Servicing," *QST*, Jan. 1982, pp. 38-41.

[2] C. Hutchinson, *The Radio Amateur's Handbook*, 61st ed. (Newington: ARRL, 1983), p. 4-44.

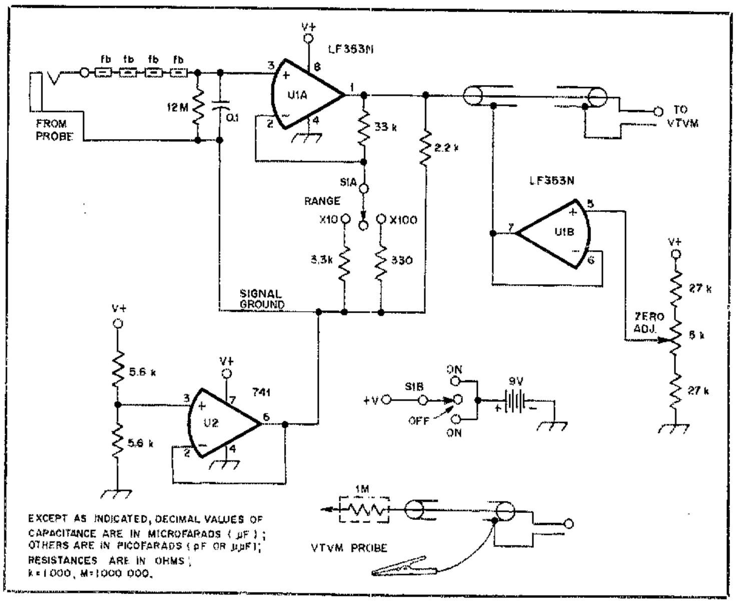


Fig. 1 — Schematic diagram of a high-impedance meter amplifier built by KR8L.

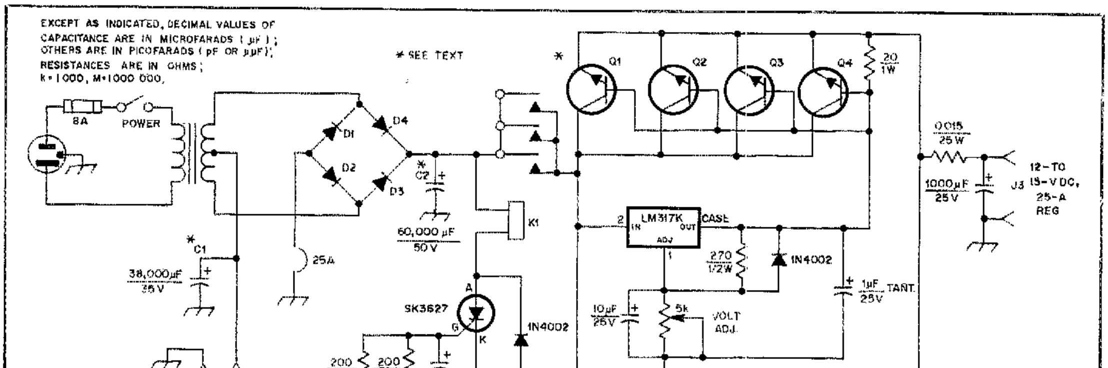


Fig. 2 — Schematic diagram of a 25-A power supply built from an automotive-type battery charger by KØUDZ.

D1-D4 — 25-A bridge rectifier (Radio Shack no. 276-1185) or individual diodes with 25-A (or greater) rating (1N2155 or equiv.).

J1-J3 — Power connector, builder's choice.

Q1-Q4 — 2N3055 or equiv.

Perhaps the most useful application I have found is for troubleshooting. I have an rf probe to use with my meter and amplifier. This combination provides a noticeable deflection of the meter with any signal much above S9 at the antenna terminals of my receiver. I can follow a signal like that all the way through the receiver now.

My amplifier is built into a small plastic box. It has a ¼-in. phone jack for the probe and an 18-in. shielded wire with a ¼-in. phone plug to connect to the VTVM.[3] The on/off and range switch is a single dpdt, center-off switch. A Radio Shack IC experimenter's board (no. 276-159) serves as the pc board. — *Bill Parmley, KR8L, Leslie, Michigan*

POWER SUPPLIES FROM OLD BATTERY CHARGERS

□ I have seen old automotive-style battery chargers for sale at many flea markets. Most of these high-current devices sell for a few dollars, and they usually have a burned-out rectifier, a broken wire or some other minor fault. The 15- or 30-A ratings make them ideal candidates for power-supply parts. Fig. 2 is the schematic diagram that I used to build a high-current power supply. You can change the number of 2N3055 transistors and other component ratings to suit the supply you are building.

A 25-A bridge rectifier is available from Radio Shack (no. 276-1185). C1 and C2 are surplus units that I had on hand. The unregulated voltages at J1 and J2 will vary depending on the actual capacitance you use for these two components. Automotive parts stores have circuit breakers that are direct replacements for the automotive fuses. These are available in a variety of current ratings, and the price is only a few dollars each. I used a surplus relay for K1. It has three sets of contacts, rated at 10 A each, so I wired them in parallel to obtain the required current rating. — *Gene Wasson, KØUDZ, Rapid City, South Dakota*

TUNABLE FILTER FOR THE CODE*STAR CODE READER

□ After using a Microcraft CODE*STAR code reader for a short time I realized the audio filter stage in the unit limits its usefulness.[4] The audio output from a receiver passes through an agc circuit in the reader, then goes through a narrow band-pass filter and finally on to the digitizing stages that produce the required alphanumeric display. The center frequency of the band-pass filter is fixed at approximately 750 Hz, and this filter has a bandwidth of less than 100 Hz. From an operator's point of view, there are times when a received signal produces comfortable audio copy, but with a tone that does not fall within the passband of the filter. When this happens, the code reader does not decode the signal.

Solving this problem proved to be quite simple. The filter must be modified to allow the center frequency to be tunable, while maintaining a fixed bandwidth and gain. After studying the schematic diagram for the CODE*STAR reader and reviewing some basic circuit theory, I realized that only a minor change would be needed.[5,6] By replacing each of two 6.8-kΩ resistors (R6 and R22) with a series combination of a 2.2-kΩ fixed resistor and a 10-kΩ potentiometer (see Fig. 3) I now have the desired filter action. Since both potentiometers must be adjusted simultaneously, I used a dual control, such as Radio Shack no. 271-1732.

I mounted this new control on the front panel where the TUNE LED was located previously. The hole had to be enlarged slightly. I drilled a new hole for the LED in the upper-right corner of the front panel. I removed all of the ICs from the circuit board before unsoldering the resistors and installing the new components.

This modification allows the center frequency to be varied from about 600 Hz up to about 1200 Hz. I believe this will be adequate for almost all situations. To operate the code reader, a station is tuned in at a comfortable listening frequency, and the new TUNE control is rotated until the LED lights and the copied signal appears on the alphanumeric display. — *Allan Mense, W4PRB, St. Louis, Missouri*

[5] D. Lancaster, *Active Filter Cookbook* (Indianapolis: Howard W. Sams & Co., 1975), Ch. 7.
[6] D. DeMaw, "Understanding and Using Audio Filters," *QST*, April 1983, pp. 45-48.

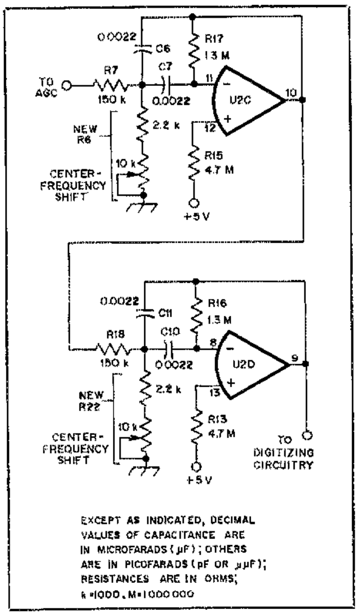


Fig. 3 — Schematic diagram of the band-pass filter section of the CODE*STAR reader showing changes made by W4PRB to provide a variable center frequency. Part numbers refer to the manufacturer's schematic diagram.

SELECTABLE CW FILTERS FOR THE YAESU FT-901DM

□ I have made two modifications to my Yaesu FT-901DM that make it more enjoyable for me to use on cw. My first complaint with the rig was that after installing a cw filter I could not operate with the wide (ssb) filter on cw. I wanted to be able to select either filter while operating cw. The automatic mike gain control (AMGC) is not functional on cw, and this is a dpdt switch, with both sections connected in parallel. Fig. 4 shows how I used one side of this switch, and connected three wires to the filter board. Now I can select either filter in the cw mode, but the ssb filter is selected automatically in the ssb mode. The AMGC operates normally on ssb. [Be sure to use shielded wire, and keep the leads to the filter board as short as possible to keep increased leakage around the filter to a minimum. — Ed.]

My second complaint about the '901 was that

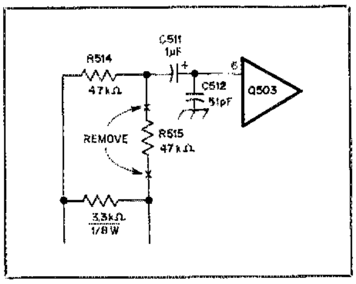


Fig. 5 — A simple modification for the FT-901DM to allow the af gain control to adjust the sidetone-monitor level. Part numbers refer to the schematic diagram in the owner's manual.

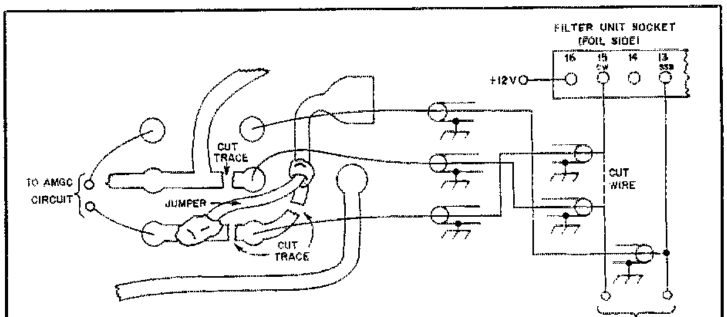


Fig. 4 — Diagram of the modification done by SV2IL to allow the selection of either the cw or ssb filters in the cw mode on the Yaesu FT-901DM.

[3] mm = in. × 25.4.
[4] P. O'Dell, "Microcraft CODE*STAR Reader Kit," Product Review, *QST*, July 1983, p. 45.

I prefer to be able to adjust the level of the sidetone monitor from outside the rig after the internal trimmer is set to an acceptable level. Fig. 5 shows a simple modification that allows the sidetone level to be adjusted by turning the af gain control. Remove R515 and replace it with a new resistor of 3.3 kΩ, but connect the top of this new component to the opposite side of R514, as shown. — *D. Vassiliades, SV2IL, Thessaloniki, Greece*

RETUNING TRAPS FOR THE WARC BANDS

☐ Now is the time to start looking for antennas to use on the 18- and 24-MHz WARC bands. You can learn about propagation on these bands by listening, even before they are opened for amateur use. The traps in a Hustler 4BTV or 5BTV are tunable, so it is easy to convert one of these antennas for coverage of the new bands.

To retune a trap, remove it from the antenna. If you remove more than one trap at a time, be sure you can read the label or have some other way to tell them apart. Couple a dip meter to the trap and check its resonant frequency. The method you use to couple to the trap may vary with the type of meter you are using. I simply connected a 5-pF capacitor between the inner end of the trap and the hot side of my dip-meter coil. Each trap should give an indication of resonance in the band it was intended for. If not, then you should take the trap apart and look for a damaged coil. The outer shell is one plate of the resonating capacitor for these traps. If your trap checks out okay, readjust the resonant frequency by sliding the outer shell up or down the coil. The tuning range seems to be as much as ±20% of the marked frequency. I moved the 10-meter trap to 24.8 MHz and the 15-meter trap to 18.1 MHz.

When you reassemble your antenna, you will have to adjust the section lengths for the new bands. The lengths I used are given in Table 1. Each length includes the length of the trap, to the bottom of the next trap or the top of the antenna. You may have to use some new tubing sections.

Joseph Boyer, W6UYH, wrote a series of articles in *CQ* about calculating the lengths of elements for multiband trap vertical antennas.[7] I recommend that you read this series if you want to calculate the new lengths for other frequencies.

Mount the antenna and connect at least four radials, each one about ¼ λ at the lowest operating frequency. Check the SWR and readjust the section lengths if necessary. Start with the highest frequency band. If you run out of adjustment range, you can use a capacitance hat to lower the resonant frequency. The Hustler design should give some ideas on this.

Some trap antennas do not use this type of trap. Some use sealed capacitors, and in others the entire trap is sealed. If your antenna uses traps with sealed capacitors, you should be able to retune them by rewinding the coils. Sometimes this requires using smaller wire to allow more turns to fit into the space. With care you should be able to retune most types of trap antennas. — *R. P. Haviland, W4MB, Daytona Beach, Florida*

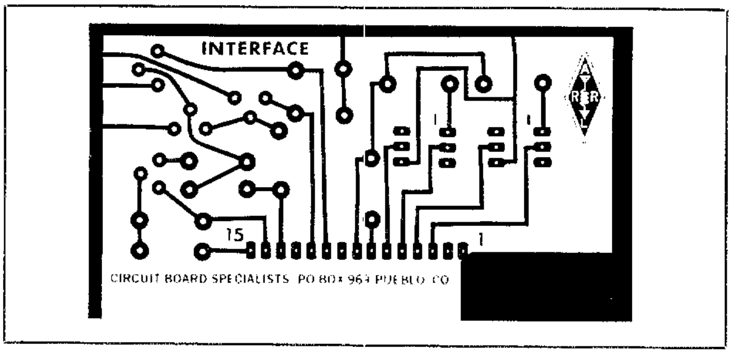
[7] J. Boyer, "The Multi-Band Trap Antenna," *CQ*, Feb.-May 1977.

Table 1

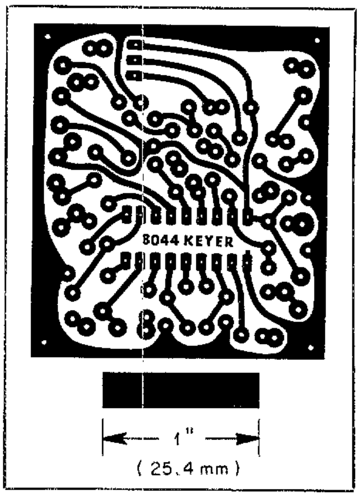
Trap Vertical Element Lengths

Band (MHz)	*Section Length (in.)*
24	108
18	24
10	49

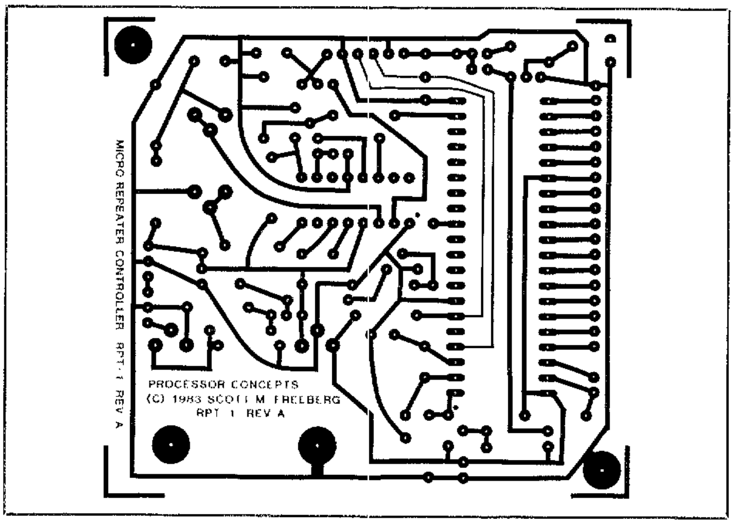
mm = in. × 25.4.



Etching pattern for the interface circuit board. Black represents unetched copper, viewed from the foil side of the board. The pattern is shown full size. A parts-placement diagram is shown in Fig. 5, page 40.



Circuit-board etching pattern for the Curtis-IC keyer. The pattern is shown full size from the foil side of the board. Black areas represent copper foil. The parts-placement guide appears on page 19.



A full-size etching pattern for the Microcomputer Repeater Controller. The board is shown from the foil side, and black represents unetched copper. A parts-placement diagram appears on page 31.

Product Review

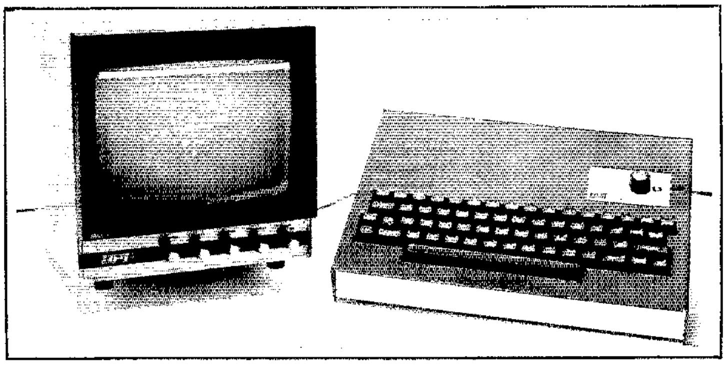
Conducted By Paul K. Pagel,* N1FB

The SRT 3000 Send/Receive Terminal

A versatile, self-contained unit that transmits and receives Baudot, ASCII and Morse code, the SRT 3000 is lightweight, has a 63-key keyboard and a built-in modem (terminal unit). Just three connections to your transceiver and one to a video monitor, and you're on the air in style!

The '3000 gives no external evidence of its tremendous capabilities. To access different functions, various keys are pressed in conjunction with the CTRL or CTRL and SHIFT keys. There are no codings or guides on the keyboard as to which key controls which function (except for a few standard functions such as REPEAT, CQ, AS, DE, KN, SK and ID). It will take time to develop proficiency in handling this unit.

The SRT is best hooked up to a video monitor, (not included) although 5-V dc at 70 mA is provided for powering an rf modulator. Back-panel connections — all phono jacks except for the Morse keyed output, which is ¼-in. phone — include cassette control and audio lines, RS-232-C lines for computer and printer, oscilloscope vertical and horizontal connections for tuning purposes, sidetone output and the standard lines such as audio in, afsk out, video out, and a PTT line for transceiver control.[1] A built-in 0.5-W audio monitor and a small speaker enhance the independent status of this unit. An internal battery is used to retain some selective programmable memories when the power is turned off.



Innards

When the DGM and video monitor are turned on, the first thing you see is a status line displaying the information you need to know. Since the DGM contains a built-in modem, variables for different modes of operation are given. A tuning bar, which works like the "tuning eye" that was used on old receivors shows when the received signal is in optimal tuning range. During Morse reception, when the desired signal is correctly centered in the passband, a double asterisk (⁑) will appear to the right of the tuning bar. The status line also shows the mode, speed, RTTY shift and T-R status, and a counter for the 1000-character buffer. More specialized functions are displayed at the right half of the status line. These include the 24-hour-clock readout and other functions, too many to be explained in this review. They are covered in the manual.

A series of selectable features makes versatile RTTY operation a reality: If for some reason you want to read the screen from across the room, you can select one of three different screen sizes. Page scrolling allows you to review a word or a sentence that has passed from view. A handy feature is the split-screen mode. It displays transmitted and received text on different halves of the video monitor, with reversed video to differentiate between the two.

One of the interesting aspects of the SRT is the ease of function selection. The instruction manual calls this "toggling." Word wraparound, carriage return, unshift-on-space, normal or reverse video — all are keyboard switchable with just the simple press of the CTRL and/or SHIFT key(s) and the correct key, mnemonically coded (i.e., CTRL Q = QBF message; CTRL SHIFT T = time display, etc.). Other keyboard controls adjust the built-in TU variables: Shift settings, tone pairs and normal or reverse shift are all changed easily.

Using a programmable 8-character code, the SELCAL functions of the SRT are activated. This allows two stations to automatically send messages to and receive messages from each other. [Note: Under FCC regulations, control operators are still required to be present at both ends of the circuit.] The SELCAL functions include printer and tape-recorder control so you can have a copy of the message sent to you. If you have a message for the other operator, the "Who Are You" (WRU) function will, upon reception of the correct code, activate your transmitter, transmit whatever is stored in MEMORY ONE and turn on the SELCAL for any possible response.

Other Fine Points

- A 24-hour clock on the extreme right of the status line, and a time memory that sends not only the displayed time, but any other 36 characters of information you want (such as the date).
- Full-duplex operation, which makes use of the split-screen function to give you a truly interactive conversation, especially if you have a full-break-in transceiver.
- A break-in mode that allows you to transmit new information without disturbing what is already entered into the 1000-character buffer.
- A number of preprogrammed and programmable memories: Ten 80-character memories may be loaded. The capabilities of the SRT allow you to set up programming loops that cause one memory to include another, or to repeat by calling *itself* again (think of a flow chart).

A few "goodies" enhance Morse code operation with the SRT. Variable weighting and character spacing, an 800-Hz sidetone, and, for those of you with little to say during a QSO, an automatic idle that can be turned on to provide B̅T̅ between characters.

Impressions

The SRT is certainly a versatile machine. The complexity of the controls necessitated a "crippled" approach to operation for a few weeks until I memorized the various command combinations. For the duration of learning time, DGM provides a handy one-half page (!) synopsis of operating commands that may be taped in a convenient place. I used the terminal at my home QTH, W1AW and N3KZ — it's lightweight and portable. The SRT can't generate large character symbols, but with quick manipulation of the memories and tape-recorder controls (also from the keyboard), I was able to insert fancy labels and brag tapes whenever and wherever I wanted. Tuning in signals with the status line bar is not difficult. The combination of a receiver RTTY filter helped when available, but even with a 2.7-kHz (ssb) passband I had little trouble copying signals. The internal demodulator filters did a good job of keeping things in the clear. When a scope was available, I made use of the crossed ellipse tuning method, but it really was not necessary.

The Morse functions are superlative. When venturing down to the low end of 40 cw, I was able to "make like the big boys" — those ops who coast at 50 wpm. By adjusting the weighting and intercharacter spacing control for high-speed intelligibility (and by preloading the text buffer to compensate for my 45-wpm maximum typing speed), I could keep up with the best of 'em.

With a maximum speed of 99 wpm, meteor-scatter work on vhf was made a bit easier by loading up a few memories with the correct information and then increasing the code speed for transmission in the proper sequence. The same function was used for moonbounce work, but with the message slowed to the appropriate rate for that mode. Even with 600 W of 432-MHz

[1]mm = in. × 25.4

*Assistant Technical Editor

rf five inches above the keyboard, no difficulties were encountered, and the SRT itself gave no RFI problems in any application — it's a well-shielded machine.

The manual is, for the most part, understood easily. Twenty-nine of its 35 pages are devoted to descriptions of the unit's capabilities. There's no explanation of the theory behind the SRT, but several pages consist of schematic and chip layout diagrams. A description of how to hook the SRT into a Microline 82A printer is also given. I never had problems with station equipment organization, for I could move the keyboard out of the way when it was not in use. The clean, sleek appearance and coloring (black, gray, silver) of the keyboard was much appreciated. When turning the unit on, I always made a few keystrokes to toggle and take advantage of those operating functions I like to use.

A Few Drawbacks

Despite its versatility and practicality, there are a few problems with the SRT-3000. The internal speaker faces out of the rear of the unit and gives a rather tinny sound. When used as a monitor for received signals, I found this limitation especially annoying, and feel there is no reason to use it as a substitute for your regular station speaker.

When switching into the split-screen mode, the monitor dims substantially, a result of the reverse video that sets off the transmit buffer display. The programmable clock must be reset every time the unit is turned on, a minor inconvenience. A feature included in many modern RTTY keyboards (but not the SRT) is an incremental QSO counter that would be perfect for contest operation.

When you are changing page-size formats, whatever is displayed gets erased. This is not a problem if you know what you want from the beginning of your operation. But if you decide to switch page sizes for any reason in the middle of a QSO, you'll lose whatever was there.

When you are selecting a speed for Morse code operation, the status indicator doesn't display the newly entered speed until you enter the transmit mode. This was a bit disconcerting at first.

The instruction manual organization is a bit strange, requiring one to read through the entire booklet before getting a truly composite idea of the SRT-3000 capabilities or even instructions as to how to turn the thing on. There is one typo in the digested command listing in the back of the book: The control commands used to turn on the random code generator for A-Z and for A-Z plus numbers and punctuation are reversed.

On the review unit, the detachable power cord unplugged a bit too easily from the rear of the keyboard, causing loss of the most recent commands and page memory. This happens, however, only when the unit is moved around. I quickly learned to hold onto the plug to keep it inserted properly when moving the keyboard.

The ac-line fuse is mounted internally, making access difficult; removing the cover is a bit of a chore. All in all, though, the advantages and pluses far outweigh these minor inconveniences.

The SRT-3000 is available from DGM Electronics, Inc., 787 Briar Ln., Beloit, WI 53511. Price class: $795. — *Leo D. Kluger, WB2TRN*

AEA KT-2

☐ After the thrill of making the first few 5-wpm cw QSOs, most Novices develop a growing urge to raise their code speeds to 13 wpm, 20 wpm and beyond. As any QRQ operator will tell you, the true joy of cw lies where syllables, words and phrases are recognized at higher speeds with little effort.

How do you get from QRS to QRQ? Practice. Advanced Electronic Applications KT-2 Keyer/Trainer can help you along the journey.

The KT-2 is a high-quality, key-pad-programmable keyer and Morse code trainer. An attractive 2-1/2 × 4-3/4 × 4-3/8-inch black and charcoal-gray metal case encloses a single glass-epoxy etched circuit board on which all components are mounted. On the top, forward-sloping surface is an on-off/volume control knob and a 12-button key pad. The back panel has two phono jacks for positive or negative transmitter-line keying, a 1/4-inch stereo phone jack for paddle connection, a 3.5-mm mini-phone jack for headphones and a power jack for connecting 12-V dc. The right side panel is perforated over the internal speaker, and the case sits atop four rubber feet to prevent marring desk tops.

Packaged with the KT-2 are a power cord, three-wire paddle cord and transmitter-keying patch cord, all with appropriate, high-quality plugs. An optional wall-mounted power supply, external rechargeable NiCd battery pack and an automobile cigarette-lighter cord for mobile operation are available at extra cost.

The Keyer

One of the KT-2's two modes of operation is as a straightforward electronic keyer with programmable sidetone frequency and speed, and other features. At turn-on, the sidetone is set at 833.3 Hz, though the frequency can be increased or decreased with two keystrokes. The code speed, initialized at 20 wpm, can be set precisely at speeds from 1 to 99 wpm. Regardless of dot or dash ratios (also programmable), speed calibration is exact. AEA uses the FCC standard of 50 code elements per word, using the word "PARIS" as the reference (a code element is equal to the time of one space between within-character dots or dashes). In other words, the word "PARIS" is comprised of 50 code elements, and the number of times it is sent per minute is the code speed in words per minute.

When using the KT-2 as a keyer, you have several mode options. At turn-on, the KT-2 functions as an iambic keyer: When using a two-lever paddle, holding the dot and dash paddles closed at the same time will cause alternating dots and dashes to be sent, a useful function when sending such characters as the letter C or a period.

Also, both dot and dash memories are enabled at turn-on; either one or both may be switched on or off from the key pad. Dot memory permits inserting a dot in a string of dashes by momentarily closing the dot paddle; conversely, dash memory permits inserting a dash in a string of dots.

For those who prefer semi-auto or "bug" operation, the KT-2 can be set up to form dots automatically, but generate a continuous tone for as long as the dash paddle is closed, mimicking a bug. Using this feature, you can also wire a straight key to the dash contact and use the Keyer/Trainer as a sidetone generator while keying a transmitter.

As mentioned earlier, dot and dash ratios (initialized at 1:1 and 3:1, respectively) can be programmed, each referenced to the length of an intra-character space. The dot-to-space ratio can be set from 0.5 to 1.5, and the dash-to-space ratio from 2.0 to 4.0.

The KT-2 will key just about any transmitter. Two diode-protected output jacks permit keying cathode-keyed and most transistor-keyed transmitters (+ jack) and grid-block-keyed and some other transistor-keyed transmitters (− jack).

Though certainly not a fault of the KT-2 in the literal sense, the jack wiring for the paddle connection is the reverse of that for some other common keyers. For the one-keyer user there is *no* problem; everything performs exactly as specified. The contester who uses the KT-2 and another keyer (one backing up the other) should note that several other commercial keyers have the dot and dash contacts wired the opposite way — to swap the paddle between keyers would require reversing the dot and dash wires. (See Hints and Kinks, *QST*, Jan. 1982, and "The CHIP (Cheap, Homemade Iambic Paddle)," *QST*, Oct. 1982, for ideas on how to solve this problem).

The Trainer

The KT-2 is a code trainer intended to be used by those who already know the code and who

want to increase their code speed. It is not intended to be a code teacher; you cannot drill on particular letters or subsets of Morse code characters. Once you are familiar with the code at a slow speed, the KT-2 offers an effective and intriguing method to improve speed quickly.

You set a comfortable starting speed, the desired ending speed and the length of time you want to spend on that session. Then, the KT-2 gradually increases the code speed from the starting speed to the ending speed over the programmed time limit. You can program any start/stop combination between 1 and 99 wpm as long as the speed is greater at the end than at the start. The duration can be programmed from 0.1 to 99.9 minutes. Alternatively, drill at a constant speed (no increase over time) is possible. At initialization, the starting speed is set at 5 wpm, ending speed at 20 wpm and duration at 10 minutes; this is simply the default condition and is easily changed from the key pad to suit a student's needs.

Within the trainer mode two code formats are possible. In the Farnsworth method, at the beginning of your programmed session characters are sent at the *ending* speed with inter-character spacing increased to yield the desired start speed. This is a proven method used at slow code speeds over W1AW and in ARRL code tapes. To increase the code speed over the duration of the session, the KT-2 gradually reduces the inter-character spacing. It is intended to prevent students from counting individual dots and dashes and to help them recognize entire characters as identifiable patterns, eliminating some of the plateaus that occasionally emerge.

Alternatively, you may select "slow code" method in which code is sent with the proper intra-character and inter-character spacings for the current speed. Again, you are cautioned that at speeds slower than 13 to 15 wpm, the slow code method is not recommended.

Code generation in the KT-2 is dubbed pseudo-random; it is not truly random in that the software reads characters from a 24,000-character table in ROM. Though finite in length, the table is large enough (!) that few, if any, users could commit it to memory. A student can select any one of 10 fixed starting points from the key pad or choose a random starting point. AEA will provide (for an additional $2) a written listing of the entire character table with the 10 starting points clearly marked; thus, you can check the accuracy of your copy. Though a few of the starting point character sequences became familiar (i.e., predictable) over time, after 30 seconds or so into any of the sequences the characters were again unpredictable; the random starting points were never identifiable.

Another choice offered by the KT-2 is fixed five-character groupings or random-length groups. With the fixed groupings you know each group will consist of only five characters; with the random-length groups, "word" length is unpredictable. You may also choose between common code characters (26 letters, 10 numerals, the period, comma, fraction bar and question mark) or both common and less common code characters (adding more esoteric punctuation and prosigns).

The only problem evident with the KT-2 (with some start/stop/duration combinations) is an inherent delay between "telling" the training to start and its actually starting. For example, though the delay is unnoticeable under the initialization conditions, when the values are changed to start equals 15 wpm, stop 25 wpm and duration 10 minutes, a 7.5-second delay ensues until the Trainer starts sending Morse. With the same start/stop speeds, a duration of 99.9 minutes (admittedly longer than most users would choose) causes a delay of 75 seconds. Other combinations yield a variety of delays. As the manufacturer explained:

> The condition ... is indeed endemic to the KT-2. The pause is actually "thinking" time for the microcomputer to determine how fast to start sending the code and what rate of acceleration to use. The amount of delay is dependent upon the parameters entered by the user ... The reason lies in the algorithm ...

The delay is little more than a slight irritant that will gladly be tolerated. Once the practice begins, the short delay is of no consequence.

One option that would make the KT-2 an even more effective trainer were it to be incorporated would be *decreasing* code-speed practice. Each approach, increasing and decreasing code speed, has its advantages and disadvantages. An increasing-speed algorithm (as used in the KT-2) starts you at a level where you're comparatively comfortable and takes you to your target level. As you progress through a session, you're constantly fighting complacency or overcoming the "inertia" of comfort; as the speed increases, performance (characters copied successfully) decreases and the target speed *feels* fast. Nonetheless, this method does work and is far more effective than most other approaches.

A decreasing-speed algorithm, however, starts you at a speed well beyond your reach and brings you down to your target speed. It is similar to the fast-code practice transmitted over W1AW. As speed decreases, your performance in that session improves and the target speed *feels* slower than it would if you had approached it from slower speeds.

The disadvantage of this method is that you must initially have a higher tolerance for frustration because you'll probably copy few characters at the start speed. This approach would help many students reach their targets more quickly. A decreasing-speed feature is merely a subjective preference, but one I would have liked to have had as an option.

Using the KT-2, you'll quickly find yourself engrossed in an effective and rather enjoyable quest for QRQ skill. The KT-2 is available from Advanced Electronic Applications, Inc., P.O. Box 2160, Lynnwood, WA 98036. Price class: $130. — *Steve Place, WB1EYI*

KLM 21.0-21.5-6A "BIG STICKER"

☐ When the opportunity arose to test KLM's 15-meter "Big Sticker," I jumped at it. The antenna definitely fulfills the "big" requirement. Its six elements are arranged along a 3-inch-OD, 36-foot boom, and it weighs in at 60 pounds.[1]

[1]m = ft × 0.3048; kg = lb × 2.2

Like most KLM antennas, this beam employs a dual driven element to achieve a low SWR across the entire band.

The boom comes in three 12-foot sections. Two of the boom sections are swaged at one end to fit inside the center section. A couple of 1/4-20 bolts secure each joint.

Each antenna element is made from two halves joined at the center by a Lexan® insulator. Each element half consists of a 6-foot section of 3/4-inch-OD tubing swaged at one end to accept a 1/2-inch tip of the appropriate length. A 6-inch piece of 7/8-inch-OD tubing reinforces each element half where it joins the insulator. Compression clamps hold the tubing sections together.

The reflector and director element halves are connected electrically by short aluminum straps. The two driven elements are interconnected by aluminum phasing straps. Because of the log-cell driven element, the feed-point impedance is 200 ohms. The review antenna came complete with KLM's model 3-60 4:1 balun, but making a balun from coaxial cable as explained in the instruction manual is easy.

Each element is secured to the boom by a single worm-gear hose clamp attached to the underside of the insulator with two 1/4-20 bolts. Sheet metal screws keep the clamps from twisting on the boom. The boom-to-mast plate and associated U bolts accept a 2-inch-OD mast. A steel truss cable running from a riser on the boom-to-mast plate to aluminum clamps near the ends of the boom provides support.

KLM supplies stainless-steel hardware to bolt the antenna together. They also include a generous supply of Penetrox®, a conductive paste used to ensure good electrical contact at the element joints and balun connections.

This antenna took about four hours to assemble. KLM's excellent instruction manual included a step-by-step assembly outline and detailed drawings with dimensions. Some of the assembly steps were more time-consuming than necessary because of ill-fitting parts. For example, some of the boom-to-element hose clamps didn't quite fit into their spaces on the insulators. Also, on some of the element halves, the machine screws that secure the elements to the insulators didn't quite fit through all of the holes. Eventually, with some filing and a lot of patience, I got the beam together.

My only serious complaint is that I found a wax-like film coating the inside of the swaged end of many of the 3/4-inch element sections. A VOM confirmed that this waxy substance does *not* conduct electricity, so I carefully scraped it off with a sharp knife and then used steel wool to clean up the inside of the element. Upon completion, I was mystified by a small handful of leftover hardware. A careful rereading of the instructions confirmed that I had put on every screw, nut and washer as specified.

Clarke Greene, K1JX, and I installed the beam atop 70 feet of Rohn 25. Clarke pulled while I followed the antenna up, and it was bolted in

KLM 21-21.5-6A 15-Meter "Big Sticker"

Manufacturer's Claimed Specifications	*ARRL Evaluation*
Frequency of operation: 21-21.5 MHz.	As specified.
Longest element: 25 ft.	24 ft, 3 in.
VSWR: Better than 1.5:1.	See Fig. 1.
Wind area: 8.5 sq. ft.	Not measured.
Turning radius: 20 ft.	22.7 ft.
Material: 6063-T832 seamless tubing.	

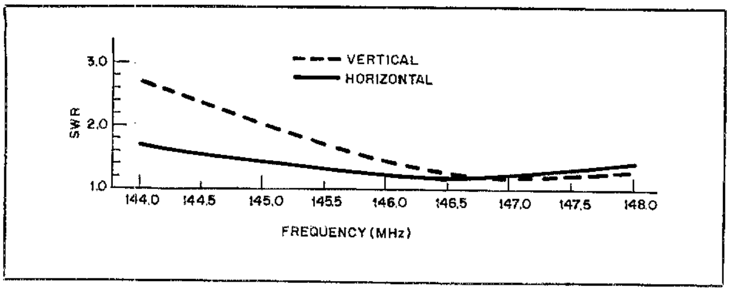


Fig. 1 — SWR curves for the Austin OMNI 2-m mobile antenna. In this installation, the horizontal elements of the test antenna were less than 1/4-λ from the car body (a compact station wagon).

place within 15 minutes of leaving the ground. Most of the weight is in the boom, the elements are short and the antenna is balanced, so installation is relatively easy. The beam is fed with 125 feet of RG-8/U.

On-the-air evaluation confirmed that this big antenna is a big performer. Some casual operation during the cw WPX contest, Field Day and the IARU Radiosport brought many unsolicited "tremendous signal" reports. Received signals were stronger on the KLM than on a 4-element beam at the same height on another tower 60 feet away. This antenna exhibits deep nulls off the back and sides, and I find myself making more small adjustments to the rotor to peak signals than with other 15-meter beams I've used.

Although it remains to be seen if this antenna can survive a New England winter, it has weathered several summer storms with wind gusts in excess of 50 mi/h that felled trees throughout the state. This Big Sticker certainly meets my idea of what a big antenna should do. If you're looking for a big antenna for 15, this one is worth considering. The 21-21.5-6A lists for $500. Manufacturer: KLM, P.O. Box 816, Morgan Hill, CA 95037. — *Mark Wilson, AA2Z*

AUSTIN OMNI 2-M ANTENNA

☐ The question of antenna polarization in vhf mobile work has been with us almost since the first vhf gear appeared in cars. Vertical antenna polarization is favored for obvious aesthetic and mechanical reasons, but horizontal polarization has documented advantages of noise reduction and higher average signal levels in irregular terrain. A 1950s trend to horizontal polarization was reversed with the fm-and-repeaters boom of the 1960s. Since the late 1970s, the effectiveness of ssb communications has rekindled interest in the art of 2-m mobile communication over considerable distances, without the aid of repeaters, mainly through the use of horizontal polarization.

Why not use an antenna that can be changed readily from horizontal to vertical polarization without sacrificing the advantages of both modes? This question was responsible for the development of the OMNI 2-m mobile antenna by Dick Austin, K1QIZ.

How It Works

The OMNI combines an old principle with some new mechanical ideas. It uses two half-wave elements fed in phase, matched to 50-ohm line by means of a corrective stub and a balun. Only the mechanical construction is new.

At the bottom of the antenna is an unobtrusive, chrome-plated cylinder. It is fitted with a threaded stud that can be screwed into a standard Hustler bumper mount or used with a matching 3/8-inch nut to attach the base to any kind of wraparound bumper mount.

The main vertical support is a tapered, white fiberglass tube with diameters of 7/8 inch at the bottom and 1/2 inch at the top. The matching network is entirely within the fiberglass tube. The 50-ohm coaxial feed line, also white, exits through a hole in a chrome-plated sleeve at the bottom of the whip. This sleeve makes a firm slip-on fit to the base assembly. The bottom edge is notched to fit around a steel pin in the base, preventing antenna rotation when the car is in motion. Note that the OMNI has its own transmission line. If you have an hf rig in your car, you do not have to disconnect it when the OMNI is installed.

At the top of the vertical support is a thick plastic sleeve that serves as a mount for the two half-wave elements. These are tapered slightly to give strength at the lower ends while reducing wind resistance. The fed ends of the elements are set into short lengths of solid brass rod that are held in place by means of wing nuts. Slots cut into the plastic mount in the vertical and horizontal planes permit the elements to be fixed in place in a vertical or horizontal position, their alignment being kept at the proper angles by the slots.

For vertical polarization, the elements are aligned with the vertical slots (on opposite sides of the insulating mount), one pointing upward and the other downward. To change to horizontal polarization, loosen the wing nuts and turn the elements to the horizontal plane. The base support is oriented so one element is parallel to the long dimension of the car, pointing forward. The other element is parallel to the rear of the car. The preferred mounting position is at the left rear of the vehicle, as this keeps the elements on the side away from pedestrian traffic. Mounted in this way, the elements are approximately 52 inches above the rear bumper.

An incidental dividend of this antenna and mount design is that the antenna assembly can be removed in an instant for safe keeping. For this purpose, there is a soft plastic clip mounted part way down the fiberglass support and slotted to hold the collapsed elements in place, parallel with the supporting mast.

Performance

Any mobile antenna is bound to be far less effective than even a small directive array mounted in the clear. Still, a vhf mobile enthusiast accustomed to the quarter-wave whip will be pleasantly surprised by results obtained with the OMNI in the vertical or horizontal mode. My first ssb contact while using the OMNI was initiated from my driveway on a small hill in Canton, Connecticut. My CQ was answered by a station on Long Island, about 65 miles away.[3] He was surprised to learn that I was using only 10-W output. Equally good reports have been received out to 75 miles or so over hilly terrain and under essentially normal conditions.

[3]km = mi. × 1.6

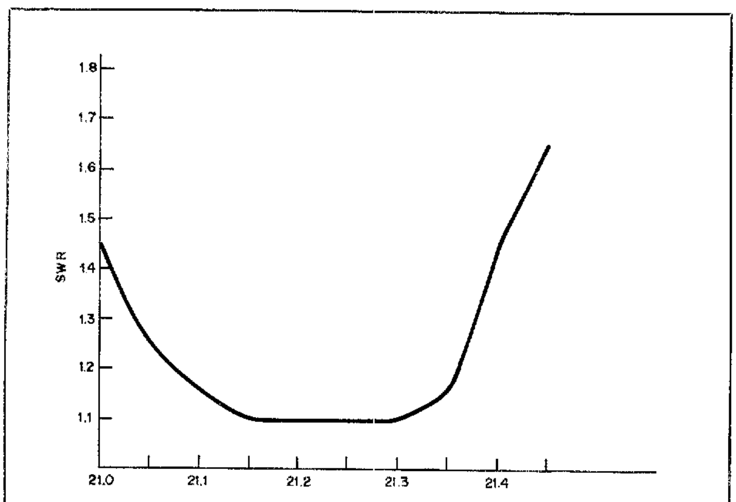


Fig. 2 — SWR curve for the KLM 21-21.5-6A.

Some of the review period was spent in central Florida where the reliable communications range with a 10-W rig, when working with well-equipped stations is contained with a circle roughly 80 miles in diameter. Experience in Florida and on the trip home showed that reliable communications ranges could be extended appreciably with the use of an amplifier; mine delivers 65 watts of output. "Grid" Gridley, W4GJO, also using an OMNI, had just over 100 watts of output available. We both found it possible to often work the Jacksonville and Miami areas about 150 and 230 miles distant, respectively. Gulf Coast tropo contacts out to 400 miles were made before the real season opened for this sort of thing.

Probably the most impressive demonstration of the capabilities of 2-m ssb operation with horizontal antennas was a four-hour continuous contact with W4GJO from his home in northwestern Georgia. This contact continued for 157 road miles (120 air miles) in the hill country of eastern Tennessee. Many other QSOs were held at distances of from 75 to 100 miles in the Shenandoah Valley of Virginia and the Pocono Mountains of Pennsylvania. Almost without exception, the person contacted volunteered the information that he had never worked a 2-m mobile station at that distance before.

The performance of the OMNI as a vertically polarized antenna, working direct and through several repeaters, compares with that of a through-the-glass-fed vertical whip antenna; there is little practical difference in their working range. There may be a difference in the timing of fades, but the signals average out to about the same level. Thus, I am getting good results with the OMNI in fm simplex and repeater work with the dividend of being able to have solid QSOs on ssb over better-than-average distances. The OMNI is a real plus compared with trying to work ssb and fm with a single antenna that favors one mode or the other.

The OMNI can be obtained from Austin Custom Antenna, 38 Terminal Rd., Providence, RI 02905. Price class: $65. — *Ed Tilton, W1HDQ*

SPEEDCALL 312-K TOUCH-TONE® DECODER KIT

□ The only hams who should be interested in a tone decoder are those building an autopatch, right? Wrong. Why? Because the Speedcall 312-K is actually a selective signaling device. Here are a few examples of its possible use: You want to be available for instant contact on the local repeater, but do not want to be distracted by the chatter of other stations in QSO; you're a DXer who wants instant notification when the rare one comes on; you may be in a RACES or ARES group that must respond in an instant, or you may be working in an extremely noisy environment and wish to receive a visual indication that someone is calling you. Many, if not most, amateurs active on the fm mode could make some use of a selective signaling device.

The Circuit

Tone-decoder ICs have improved dramatically in the last few years. Until recently, an array of seven or eight 567s was typical for tone decoding in most amateur applications. Now we have available stable, false-free decoder ICs that decode all 16 digits. Speedcall's 312-K Decoder is an innovative circuit designed around the ITT 3201 Decoder/Filter IC.

The ITT 3201 is a complete two-of-eight DTMF (Dual-Tone, Multi-Frequency) decoder/filter in a single 22-pin package, requiring a single +12-V supply. Audio for the ITT 3201 is taken from the receiver fm detector. Speedcall has made provisions for taking the audio off before or after de-emphasis.

After processing the received tones, the ITT 3201 produces a data valid signal and a hexadecimal character corresponding to any valid DTMF tone received. The hex output becomes the input for an NC 5200 logic array that functions as a field-programmable sequence decoder. Actually, it is only partially field programmable, and the programming done at the factory is what makes the Speedcall 312-K so versatile and useful. The field-programming portion is accomplished with wire jumpers and allows the user to determine the series of digits required to activate the output (address code). Any sequence consisting of three to eight digits (numbers and/or symbols) may be selected and programmed by the user.

When a valid address code has been detected, the logic array output is fed to an inverter, a latch and a 3-Hz oscillator. The oscillator output drives a transistor, causing it to turn an LED on and off at the 3-Hz rate. This transistor can also be used to drive any low-current 12-V device (such as a relay), to connect to additional signaling instruments, or the oscillator section can be bypassed to create a continuous output. In either case, the output will remain activated until the 312-K is reset manually with the reset switch or until it receives the COMMAND RESET signal, if it has been programmed for this function.

Field Programming

Address code field programming is straightforward. The programmer simply inserts wire jumpers between sockets representing digit output and position (in the sequence) registers. The same digit may be used for more than one position in the sequence, e.g. the 312-K could be programmed to respond to 13356, with the digit "3" being used in the second and third registers (positions).

With the COMMAND RESET option programmed in, the calling operator may reset the 312-K from his station by entering the address code followed by the reset digit — whatever the user has programmed into the 312-K for this function. Thus, if an operator does not respond in a specified period of time, the calling operator may cancel the call and deactivate the signaling device.

The MASTERCALL option permits the sending station to substitute the MASTERCALL digit for any digit or digits in the address code. Suppose a receiving unit is coded for 1598 with # as the MASTERCALL digit. The sending station may send 1598 to address the unit, or he may send #598, ##98, ###8, ####, 1###, #5##, etc. This has uses in situations in which receiving stations can be fitted to some grouping. It could be used to alert selectively a group of DXers all needing the same country, or some subgroup of ARES. The COMMAND RESET option may be used in conjunction with the MASTERCALL option.

Construction

The Speedcall 312-K is reasonably simple to build. If any difficulty is to be encountered, it would most likely be in finding the proper connection point in the receiver. Speedcall suggests that it be connected as near the fm detector as possible, which is not particularly difficult to do if you are familiar with your fm receiver. If not, you would probably be wise to enlist the aid of a local radio technician.

The 3-3/4 × 2-inch circuit board might fit inside some fm transceivers. Connections required are +12 V, common, audio and audio shield (common). Once the device is connected to the receiver, the user is required to tune to a station sending a DTMF signal and adjust a potentiometer on the circuit board to provide a 500-mV signal to the input of the decoder IC. No other adjustments are required.

The unit I built and tested has performed flawlessly over the last few months. I added an external piezo buzzer purchased at Radio Shack. The basic kit is in the $90 price class, while a deluxe version with case, switch and buzzer is in the $105 price class. More information is available from Speedcall Corporation, 2020 National Ave., Hayward, CA 94545, tel. 415-783-5611. — *Peter R. O'Dell, KB1N*

LOGBOOK

□ Hams do a lot of paperwork. We keep logs, we send, receive and record the sending and receiving of QSL cards, and, for various operating awards, we record states and countries worked and confirmed. Some of us spend as much time shuffling paper as we do rag chewing!

Today, many hams own Radio Shack TRS-80® personal computers. Jerry Crosby, WB8YUO, has written a program for the Models I and III that lessens the paperwork that the active ham encounters. Called "Logbook (Amateur Radio Computer Logbook Program)," the program allows you to log QSOs, search the log, extract data from the log for QSLing, and keep a record of your DXCC, WAS and 6-band WAS progress.

The log includes the QSO date, call sign of the station worked, operator's name, city, state and country of the station worked, frequency, mode and QSO time, exchanged signal reports, and other notes (such as QSL status). Once you have logged a QSO, a program-editing module allows you to make changes and/or corrections at any time.

Searching the log is a two-step process. Let's say, for example, I want to search my log for all contacts made on March 8, 1983. First, I use the indexing module to index the log according to date (the category I wish to search). Next, I use the search module to find all March 8 QSOs. (The search function works very quickly.)

The program provides 12 ways of outputting (to the CRT or a printer) the information found by the searching process. If you desire, you can output the whole log.

The QSL module of "Logbook" allows you to pull pertinent QSO data out of the log and print that data on tractor-fed, self-adhesive labels. The information printed on the label may be personalized by changing some program lines (the program is written in BASIC) to include your station equipment list.

"Logbook" works as advertised. A lot of thought went into the program. Since the program is written in BASIC, the adventurous user may modify the program to meet his or her requirements.

The program requires a minimum of 32K of RAM and one disk drive. It operates with most of the popular disk operating systems available for the TRS-80 Models I and III. "Logbook" is available from Jerry Crosby, WB8YUO, 6333 Willowdale Ct., Columbus, OH 43229. Full documentation and a program diskette are included for $39. — *Stan Horzepa, WA1LOU*

Technical Correspondence

Conducted By
Bob Schetgen,* KU7G

The publishers of QST assume no responsibility for statements made herein by correspondents.

AMTOR CONTROVERSY

☐ I would like to correct an error in an otherwise excellent article by Paul Newland, AD7I ("An Introduction to AMTOR," July 1983 *QST*). Paul referred to four modes of operation in the AMTOR system: A, B, L and S. Mode S is not included in the definition of AMTOR, however, and I shall explain why.

Mode S, better known in maritime radio as SELFEC, is part of the maritime SITOR system from which AMTOR is derived. This mode enables a shore station to transmit a message to a specific shipboard station that cannot use Mode A because international maritime regulations prohibit hf transmitting while in a foreign port.

Paul describes this mode as, "like Mode B but with selective calling." Not only does it allow the specified receiving station to print only directed messages, it is so designed as to actively prevent reception of the message by *any* other station.

This feature is clearly outside the spirit of Amateur Radio, at least in the view of the licensing authorities in those countries where permission was first sought. The FCC has followed this line, as the recently approved rule change mentions specifically Modes A and B, with no mention of SELFEC. As far as I know, no presently active AMTOR stations have SELFEC mode, and the AMTOR units currently available do not provide this mode. Perhaps it is included in units designed for both amateur and maritime use.

I should point out that the absence of SELFEC in AMTOR does not prevent the use of Mode B with a selective call feature to prevent the reception of unwanted messages. The user can add a suitable hardware or software device to the AMTOR code converter in the same way as is already done with conventional RTTY. — *Peter Martinez, G3PLX, Hants, England*

The author replies: I think the major problem here is the definition of AMTOR. Once a definition of AMTOR is made, several additional problems fall out of the definition.

To the best of my knowledge, Peter was the first to coin the term AMTOR, in his article "Amtor, An Improved Radioteleprinter System, Using a Microprocessor," in August 1979 *Radio Communication*. In my mind, that article defined the term AMTOR for use in the U.K., unless the local regulating administration later redefined it. In the USA, the Federal Communications Commission has defined AMTOR to be the protocol specified in CCIR Rec. 476-2 (1978). Allow me to quote from Sections 97.69[b] and 97.69[b](3) of the Commission's Rules and Regulations: " . . . the following digital codes, as specified, may be used: . . . The International Radio Consultative Committee (CCIR) Recommendation 476-2 (commonly known as AMTOR), provided that the code, baud rate and emission timing shall conform to the specifications of CCIR 476-2 (1978) Mode A or Mode B." Let me point out that the FCC is referring to the CCIR's definition of Mode B, not my definition of Mode B or Mode S. The CCIR's definition of Mode B, as stated in CCIR Rec. 472-2 (1978), is listed in Section 3.2 of that document as: "A synchronous system, transmitting an uninterrupted stream of characters from a station sending in the collective B-mode (CBSS) [I call this Mode B — PBN] to a number of stations receiving in the collective B-mode (CBRS), or from a station sending in the selective B-mode (SBSS) [I call this Mode S, PBN] to one selected station receiving in the selective B-Mode (SBRS)." What I describe as Mode S is referred to by the CCIR in the Forward Error Correction (FEC) section as "Selective Broadcast." It is Mode B, as defined by the CCIR, as described in the protocol specification that the FCC has approved. Use of Mode S (FEC selective broadcast) is permitted by the FCC.

To avoid this confusion in the future, I recommend that administrations approve the use of "CCIR Rec. 476" (as the FCC has done), rather than the term AMTOR. If an administration wishes to restrict the use of some protocol features (although I don't know why they would), they should approve, "CCIR Rec. 476 less Sections 5 and 6," or whatever is appropriate. Without such wording, rules and regulations are subject to operator interpretation.

Peter's statement, " . . . not only does [Mode S] allow the specified receiving station to print only directed messages, it is so designed as to actively prevent reception of the message by *any* other station . . . ," is incomplete. I would finish his sentence with, ". . . using Mode B." What he is describing is like trying to monitor Mode A traffic without a Mode L code converter (recall that Mode L allows a third party to monitor Mode A traffic). To monitor Mode S, you need a Mode S code converter.

My station, and other stations using my firmware, can monitor Mode S traffic as well as send and receive it. There is the option of printing *all* Mode S traffic (on a channel) *regardless* of intended receiving station. It works well! There is no reason that amateur stations can't monitor Mode S traffic. The only reason existing equipment cannot monitor Mode S is that manufacturers haven't provided this capability. It should be a simple matter to update current AMTOR programs for Mode S monitor capability. (My firmware is available to radio amateurs for their own use at no charge. This firmware object code is available by calling N2DSY's computer bulletin board, tel. 201-943-7754.)

I view Mode S not as a facility to prevent others from reading my traffic, but as a facility to prevent other stations from wasting my teleprinter paper. (With my AMTOR converter, I can select whether or not I want to receive Mode B traffic.)

I should have been clearer in my original article,[1] and added the following to my definition of Mode S that appears on page 12: "This doesn't mean that other stations *cannot* print traffic intended for your station. It means they have the *option* of not printing traffic directed to your station, and you have the *option* of not printing traffic directed to their station."

I hope all manufacturers of AMTOR equipment will provide Mode S with monitor capability. However, as I pointed out in note 3 of my article: "Not all (AMTOR) controllers include all modes. . . . make certain the unit . . . (you get includes) . . . the modes you want to use."

Finally, Peter is correct when he points out that selective-call capability can be added, external to the AMTOR code converter, to those stations that require this capability (without resorting to Mode S). However, that usually requires an additional microprocessor system to implement the feature. Rather than add an additional processor, I chose to implement this feature in my converter.

Microprocessor chips themselves are inexpensive (less than $5). But when you add the support chips, pc board, sockets, power supply, cabinet, and so on, that second micro can cost the same amount of money as a homebuilt AMTOR code converter! It is simply more economical to build SELFEC into an AMTOR converter in the first place.

I hope my comments help to clarify these issues. I should add that I, with other AMTOR users, will forever be in Peter's debt for his pioneering work with AMTOR systems. Thank you, Peter! — *Paul Newland, AD7I, Holmdel, New Jersey*

SILENT VHF STATION CAUSES TVI

☐ A recent TVI problem prompted me to write and share my experience. TVI was occurring, only on Channel 6, during 20-meter operation. It was equally strong using either 100 W or 1000 W of power. My station consists of an FT-102 with an Alpha 76 PA and a 204 BA antenna at 72 ft.[2]

A low-pass filter is used between the exciter and PA with another at the PA output. All cable is double-shielded RG-214/U. My 2-meter antenna is about 3 ft above the 20-meter antenna. As a professional communications engineer, I should have known what was happening.

Using a Motorola R2000 spectrum analyzer and a through-line sampler, I found my sixth harmonic to be approximately 0.7 μV at 85 MHz. (TV Channel 6 is approximately 700-μV field strength at my home.) Next, I measured signal strength with the analyzer connected to the TV antenna. Wow! The sixth harmonic was equal to the channel 6 signal.

The answer is obvious: Intermodulation distortion was being developed in the 2-meter radio and re-transmitted through the 2-meter antenna. I might add that the situation was worse when the 2-meter rig was off. A circulator at the output of the 2-meter radio cured the problem. However, moving the vhf and hf antennas apart, or disconnecting the vhf antenna, would be a less expensive cure.

Several 2-meter radios that use solid-state switching were tried, and all retransmit spurious

*TIS Specialist, ARRL

[1] P. Newland, "An Introduction to AMTOR," July 1983 *QST*, pp. 11-13.

[2] m = ft × 0.3048.

products. Only a commercial vhf radio, which uses an antenna relay for switching, causes no trouble.

This may save a lot of hair pulling when TVI suddenly develops. Chances are that a new 2-meter 220-MHz or 440-MHz rig at your station is causing the problem. — *Robert Findlay, W6NZX, Vista, California*

QUAD ANTENNA

☐ When a new 3-element 20-meter quad was put into operation at W5DJ, a couple of problems were encountered. I believe that my experience will save future builders some headaches and needless tower-climbing exercise.

First, the SWR changed drastically with beam heading. When the antenna was pointing north, the SWR was about 1.15:1; turned to the west, it would rise to about 1.8:1. The new quad is supported on a guyed 70-ft tower that previously held a 3-element Yagi. The Yagi did not show any significant SWR change with beam heading. This fact helped hide the cause of the problem.

The tower is supported by two sets of guy wires. One set of three guys at the 25-ft level, and another set at 60 ft. All guy wires are broken into 12-ft sections by insulators.

The problem was eventually traced to the top guy wires, which were attached directly to the tower's anti-twist bars. Considering any pair of top guys, there was a 12-ft section connected to the anti-twist bar and (through the tower and anti-twist frame) to another 12-ft section. Once it was recognized that there was a continuous conducting length of 28-ft (roughly a half wavelength at 14 MHz), the solution was obvious. An insulator installed between each guy wire and the anti-twist frame cured problem number one completely.

This problem did not exist with the Yagi because the Yagi's driven element was above the top of the tower, some 15 ft from the top guys. With the quad, the bottom of the driven loop is within about 2 ft of the anti-twist frame's vertical position. (In the near field, field intensity varies inversely with the cube of distance from the radiator.) Thus, coupling into the top guy wires is several hundred times greater with the quad than it was with the Yagi.

SWR variation resulted primarily from the fact that the plane of the quad loop is perpendicular to the closest guy wires at some headings, and parallel at others. It is not necessary to install insulators at the lower anti-twist bars because their attachment point is 37 ft below the bottom of the quad; coupling is insignificant.

Problem number two proved to be as interesting to diagnose as the first, and more difficult to solve. The antenna was designed for 14,150 kHz, but SWR and noise-bridge measurements indicated resonance at 13,800 kHz. Wire lengths for the reflector, driven element and director had been correctly calculated from equations and accurately cut to those values.[3] Why, then, was the antenna resonant 350 kHz lower than expected?

Design equations and measurements were both known to be good. The only remaining cause was the type of wire used. I had used some 12-gauge, stranded, vinyl-insulated wire to improve mechanical fatigue resistance. (It is well-known that broken wires from continuous wind-induced flexing is a problem with quads.) My rationale was that the stranded wire is better able to withstand mechanical fatigue than solid wire — that much is correct. Further, I reasoned that the vinyl insulation would protect the tinned copper strands from corrosion and add some mechanical strength — this is also true.

However, I overlooked the electrical effect of the vinyl insulation. It finally occurred to me that dielectric loading of the insulation was probably the cause of my woe. To verify this, before lowering the 20-meter quad, I assembled a single, 10-meter quad loop using the same type of wire. A noise bridge indicated resonance at about 28,600 kHz. After the vinyl insulation was stripped off, resonance occurred at about 29,200 kHz. Clearly, dielectric loading of the vinyl insulation causes a drop in resonant frequency of about 2% — similar to the 20-meter quad behavior.

The quad was lowered and the insulated wires replaced with 14-gauge enameled copper-clad wires, cut to the same lengths. With the quad back on the tower, resonance is now at 14,050 kHz. Changing from vinyl-insulated to enameled wire resulted in an upward frequency shift of about 250 kHz. Although not exactly consistent with the design equations yet, a small adjustment will put resonance at 14,150 kHz.

I am not aware of any mention of this loading effect (caused by insulation on wire elements) anywhere in Amateur Radio antenna publications or engineering texts. This education compensated somewhat for the considerable effort required to lower, adjust and raise a quad antenna in near-100° F (38° C) weather! — *John Kennedy, W5DJ, Dallas, Texas*

[3] J. Lindsay, Jr., "Quads and Yagis," *QST*, May 1968, p. 11.

DIAL LINEARITY ERRORS

☐ In appearance, the engine-engraved dial has no rival. Because amateur bands are narrow on a percentage basis, it is often possible to select components such that a linear dial indicates frequency exactly at the ends and shows intermediate values with good accuracy.[4] If a parallel L-C circuit is tuned by a semi-circular-plate variable capacitor, the actual frequency will be lower than the dial indication, except at the tuning-range ends.[5] It is difficult to find the position and value of the maximum error, but the mid-scale error is found easily from the expression

$$f_e = f_{min}\left[\frac{1+K}{2} - \sqrt{\frac{2}{1+\frac{1}{K^2}}}\right] \quad \text{(Eq. 2)}$$

where

f_e = frequency error
f_{min} = low edge of tuning range
f_{max} = high edge of tuning range

$$K = \frac{f_{max}}{f_{min}}$$

Using the 80-meter band as an extreme case, the dial will read a maximum of 24.99 kHz high at 47% rotation, whereas Eq. 2 yields 24.94 kHz. Most other tuning ranges have smaller percentage bandwidths so that Eq. 2 will more closely approximate the maximum error. Note that the mid-scale error decreases rapidly with K.

To minimize end-effect errors, a gear reduction of 2:1 may be arranged. The capacitor then operates in the middle half of its range, where the linearity of capacitance versus shaft angle is nearly perfect. — *Frank Noble, W3MT, Bethesda, Maryland*

[4] F. Noble, "Finding a Pad," *QST*, Oct. 1981, p. 51.
[5] For a derivation of Eq. 2, send an s.a.s.e. to Frank Noble, W3MT, 10004 Belhaven Rd., Bethesda, MD 20817.

Feedback

☐ In "New Ideas for the VHF Wattmeter" (Oct. 1983 *QST*) author Lamb has pointed out a few errors. Under the subheading "Calibration," all references to R3 actually refer to R1. The last sentence of the first full paragraph on page 13 should read: "... to that of the terminating resistor." Additionally, the unmarked capacitor in Fig. 4 should be marked C1.

☐ Several readers have noticed an error in the wavemeter schematic diagram that appears in "The Ever Useful Wavemeter" (Sept. 1983 *QST*). Fig. 4 shows the connection of D1 and D2 at the cathode of D1. This junction should be moved to the anode of D1, so that the cathode of D2 is connected to the anode of D1 and the 5-pF capacitor. Connect the cathode of D1 to the 0.01-μF capacitor. A similar circuit and the theory of operation, are in *The Radio Amateur's Handbook* (Fig. 13, page 5-8).

☐ OK1ABB found an error in, "Go Class B or C with Power MOSFETs" (March 1983 *QST*). The detail drawing of T1 in Fig. 4 has leads E and C transposed. From left to right, the correct lead order is A, C, E/D, B, F.

☐ Author Wayne Cooper points out an error in Fig. 1 of his article, "A New Mixer for the ASTRO 103 Receiver," which appeared in Hints and Kinks, Oct. 1983 *QST*, page 41. A connecting dot should be shown where the input lines to both voltage regulators cross, near the LM 317T.

☐ In "Overvoltage Protection for 13.8-V Power Supplies" (Oct. 1983 *QST*), an error appears on page 40. The 7.5-V, 5-W Zener diode referred to in column one, 10 lines from the bottom of the page should be a 1N5343B.

☐ There are several corrections to "The Pizza Clock: An Exercise in Wire Wrapping," (June 1983 *QST*). On the circuit-board patterns (available from ARRL for a large s.a.s.e.), U8 and U9 are transposed; U9 belongs in the 14-pin socket. Also, the 1-kΩ resistor at pin 6 of U1 has been eliminated; it may be necessary to break the trace and install the resistor if there are problems with the function of U1 (characteristics of some crystals will require the addition of the resistor for proper operation). The IC labeled U4 (on page 31) is actually U10. There are two pin 12s shown on U2. The real pin 12 is shown connected to pin 14 of U3; the other should be labeled pin 13, and is grounded, along with pins 8 and 15.

WAØPBQ informs us that the minimum PIV specification of most LEDs falls below the 8-V reverse bias used in the Pizza Clock. LEDs should be tested before use in the clock; those that fail can still be used if a 100-kΩ resistor is connected across their leads.