Index to Volume LXXVII—1993

Compiled By Brian Battles, WS1O
QST Features Editor

How to Use This Index

Items are listed according to the regular column under which they appeared and/or under any category (or categories) that generally describes them. The name of the author(s) appears in parentheses, followed by a colon, then the page number and issue. Examples: "Honey, I Shrunk the Antenna! (Newkirk): Jul, 34" appears under **Feature Articles—Technical** and **Antennas and Feed Lines**; "Packet for Lunch (Larsen): Mar, 38" appears under **Feature Articles—General Interest** and **Packet Radio.** Some items have additional subject-description information or the name of a department in which it appeared, in brackets after the title.

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Picture this: Color SSTV!

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OUR COVER

A picture is worth a thousand words—especially on HF! Hams who have an IBM PC-compatible computer can now operate on all the popular SSTV modes, using up to 32,000 colors. A single plug-in board and companion software can transmit images like these. See the article on page 20. *(photos by Kirk Kleinschmidt, NT0Z)*

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January 1993
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TECHNICAL

NEWS AND FEATURES

OPERATING

DEPARTMENTS

SLOW-SCAN TV—

Ham stations equipped with an IBM PC-compatible computer now have another use for it — *SSTV!* With this plug-in board, you've got most of the makings of a superb SSTV station!

By John Langner, WB2OSZ 115 Stedman St #Q Chelmsford, MA 01824-1823

Most hams who have an interest in operating slow-scan television (SSTV) are often scared by the enduring myth that it's expensive. As outmoded as this tale is, it does have basis in fact. A few years ago, the most popular piece of SSTV equipment was the Robot 1200C scan converter.[1] It's a great product, but its $1300 price tag and the lack of any viable less-expensive alternative severely restricted the widespread use of SSTV.

Now, for the price of a '1200C, you can buy a blazingly fast home computer complete with floppy disk drives, a good-sized hard disk, VGA color monitor and video card, at least a megabyte of RAM, and still have enough left over for an SSTV interface and software. For most hams, this is the only *sensible* approach!

A couple of years ago, comprehensive color SSTV systems became available for use with Amiga and Atari ST computers.[2,3] Both of these systems support a wide variety of transmission modes: Robot, AVT,[4] Martin, Scottie, Wraase[5] (you'll find a Glossary of Terms in Table 1), graphical user interfaces, on-screen tuning indicators, the ability to read and write popular image-file formats and modify received images. Surprisingly, nothing comparable was available for the ubiquitous IBM PC! SSTV home-brewers are nearly extinct, so I figured if *I* didn't do it, it wouldn't get done! As a result, a comprehensive, easy-to-use, low-cost color SSTV system is now available[6] for use with a VGA-equipped IBM PC/AT-compatible.

In this article, I'll provide some background on SSTV, a few tips for home-brewers and the details of my latest project called *Pasokon TV*. (*Pasokon*—pronounced "pah-so-cone"—is the Japanese abbreviation for *personal computer.*) See Table 2 for its specifications.

Some SSTV History

Amateur slow-scan TV began in 1958, when a group of experimenters, organized by Copthorne MacDonald (then WA2BCW, now VY2CM) became interested in sending images on the MF/HF bands.[7] Trouble was, a standard commercial TV signal requires a bandwidth of a few *megahertz*, but signals in the MF/HF amateur bands are restricted to

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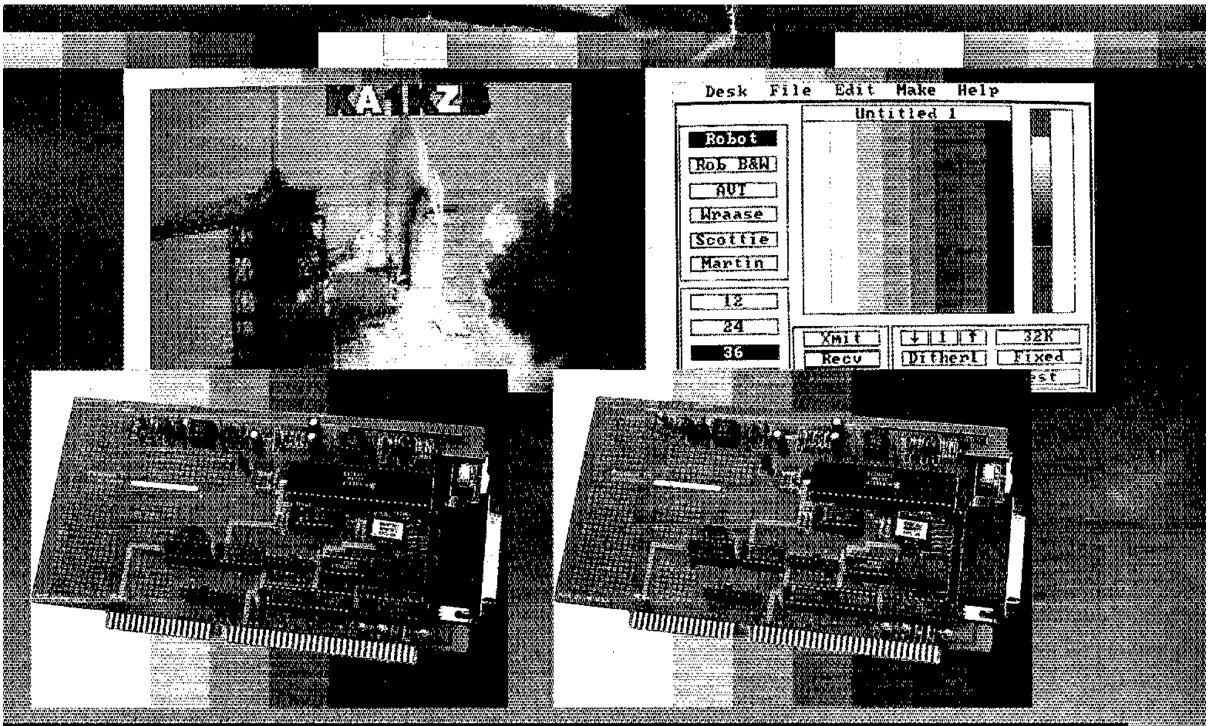
[1] Notes appear on page 30.



IT ISN'T EXPENSIVE ANYMORE!

bandwidths of a few *kilohertz*. Clearly, it wasn't possible to squeeze thirty 525-line images per second into such a narrow bandwidth. Another standard was needed.

The early SSTV 8-second transmission standard is illustrated in Fig 1. Audio tones in the 1500- to 2300-Hz range represent black, white and shades of gray. A short, 1200-Hz burst separates the scan lines, and a longer, 1200-Hz tone signals the beginning of a new frame.

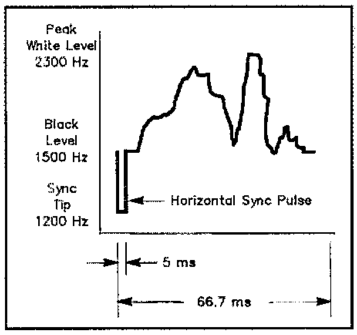


Fig 1—Original 8-second black-and-white transmission format showing the video-signal levels for peak white, black and sync (commonly referred to as "blacker than black"). This transmission mode uses a total of 120 lines transmitted at 15 lines per second for a total transmission time of 8 seconds. The durations of the horizontal and vertical sync pulses are 5 and 30 ms, respectively.

Hams have always been very creative about taking junk and turning it into useful radio equipment. In this case, the early SSTVers located some surplus radar display tubes. These are similar to the CRTs used in oscilloscopes or TV sets, with one important difference: They have a very-long-persistence (P7) phosphor. This means that a spot on the screen continues to glow for *several seconds* after the electron beam has passed that point. By using these long-persistence tubes, the beginning of the image was still visible, though somewhat faded, at the end of the 8-second transmission—but the complete picture could be seen!

Early Color SSTV

Sending color pictures with black-and-white (B&W) equipment was a real challenge! The sender would transmit the same picture three times, each time with a red, green, or blue filter in front of the TV camera lens (see Fig 2). The receiving operator took three long-exposure photographs of the screen, placing red, green and blue filters in front of the film camera's lens at the appropriate times. This is known as *frame-sequential* color SSTV. Fig 3 illustrates the signals sent for a color-bar test pattern.

In the mid 1970s, it became feasible to save these three images in solid-state memory and simultaneously display them on a color TV. But, the frame-sequential method had some problems. As the first frame was received, you'd see a red and black image. During the second frame, green and yellow would appear. Blue, white, and other colors didn't show up until the final frame. Any noise (QRM or QRN) could ruin the image registration (the overlay of the frames) and spoil the picture.

The next step forward was the *line-sequential* method. As shown in Fig 4, each line is scanned three times: once each for the red, green and blue picture components. Pictures could be seen in full color as they were received and registration problems were reduced. The Martin, Scottie, and Wraase modes are all minor variations of the line-sequential theme.

Rather than sending color images with the usual RGB components, Robot Research used luminance and chrominance signals for the 1200C modes.[3] The first half or two-thirds of each scan line contained the luminance information, which is a weighted average of the R, G and B components. The last half or one-third of each line contained the chrominance signal with the color infor-

Table 1
Glossary

ATV—Amateur Television. Sending pictures by Amateur Radio. You'd expect this abbreviation to apply equally to fast-scan television (FSTV), slow-scan television (SSTV) and facsimile (fax), but it's generally applied only to FSTV.

AVT—Amiga Video Transceiver.
1) Interface and software for use with an Amiga computer, developed by Ben Blish-Williams, AA7AS, and manufactured by Advanced Electronic Applications (AEA);
2) a family of transmission modes first introduced with the AVT product.

Back porch—The blank part of a scan line immediately following the horizontal sync pulse.

Chrominance—The color component of a video signal. NTSC and PAL transmit color images as a black-and-white compatible luminance signal along with a color subcarrier. The subcarrier phase represents the hue and the subcarrier's amplitude is the saturation. Robot color modes transmit pixel values as luminance (Y) and chrominance (R-Y [red minus luminance] and B-Y [blue minus luminance]) rather than RGB [red, green, blue]).

Demodulator—For SSTV, a device that extracts image and sync information from an audio signal.

Field—Collection of top to bottom scan lines. When interlaced, a field does not contain adjacent scan lines and there is more than one field per frame.

Frame—One complete scanned image. The Robot 36-second color mode has 240 lines per frame. NTSC has 525 lines per frame with about 483 usable after subtracting vertical sync and a few lines at the top containing various information.

Frame Sequential—A method of color SSTV transmission which sent complete, sequential frames of red, then green and blue. Now obsolete.

Front porch—The blank part of a scan line just before the horizontal sync.

FSTV—Fast-Scan TV. Same as common, full-color, motion commercial broadcast TV.

Interlace—Scan line ordering other than the usual sequential top to bottom. For example, NTSC sends a field with just the even lines in 1/60 second, then a field with just the odd lines in 1/60 second. This results in a complete frame 30 times a second. AVT "QRM" mode is the only SSTV mode that uses interlacing.

Line Sequential—A method of color SSTV transmission that sends red, green, and blue information for *each sequential scan line*. This approach allows full-color images to be viewed during reception.

Luminance—The brightness component of a video signal. Usually computed as Y (the luminance signal) = 0.59 G (green) + 0.30 R (red) + 0.11 B (blue).

Martin—A family of amateur SSTV transmission modes developed by Martin Emmerson, G3OQD, in England.

NTSC—National Television System Committee. Television standard used in North America and Japan.

PAL—Phase alteration line. Television standard used in Germany and many other parts of Europe.

Pixel—Picture element. The dots that make up images on a computer's monitor.

P7 monitor—SSTV display using a CRT having a very-long-persistence phosphor.

RGB—Red, Green, Blue. One of the models used to represent colors. Due to the characteristics of the human eye, most colors can be simulated by various blends of red, green, and blue light.

Robot—(1) Abbreviation for Robot 1200C scan converter; (2) a family of SSTV transmission modes introduced with the 1200C.

Scan converter—A device that converts one TV standard to another. For example, the Robot 1200C converts SSTV to and from FSTV.

Scottie—A family of amateur SSTV transmission modes developed by Eddie Murphy, GM3BSC, in Scotland.

SECAM—Sequential color and memory. Television standard used in France and the Commonwealth of Independent States.

SSTV—Slow Scan Television. Sending still images by means of audio tones on the MF/HF bands using transmission times of a few seconds to a few minutes.

Sync—That part of a TV signal that indicates the beginning of a frame (vertical sync) or the beginning of a scan line (horizontal sync).

VIS—Vertical Interval Signaling. Digital encoding of the transmission mode in the vertical sync portion of an SSTV image. This allows the receiver of a picture to automatically select the proper mode. This was introduced as part of the Robot modes and is now used by all SSTV software designers.

Wraase—A family of amateur SSTV transmission modes first introduced with the Wraase SC-1 scan converter developed by Volker Wraase, DL2RZ, of Wraase Electronik, Germany.

mation. Existing B&W equipment could display the B&W image on the first part of each scan line and the rest would go off the edge of the screen. This compatibility was very beneficial when most people still had only B&W equipment.

The luminance-chrominance encoding also made more efficient use of the transmission time. A 120-line color image could be sent in 12 seconds instead of the usual 24. Our eyes are more sensitive to details in changes of brightness than color, so the time could be used more efficiently by devoting more time to luminance than chrominance. The NTSC and PAL broadcast standards also take advantage of this characteristic of our vision and have narrower bandwidth for the color information.

The 1200C introduced another innovation of encoding the transmission mode in the vertical sync signal. By using a narrow frequency-shifted (FSK) encoding around the sync frequency, compatibility was again maintained. This new signal just looked like an extra long vertical sync to the older

Table 2
Pasokon TV Features

- Send and receive all popular modes:

Robot (color):	12, 24, 36, 72 seconds
Robot (B&W):	8, 12, 24, 36 seconds
AVT (color):	24, 90, 94, 188 seconds
Martin (color):	M1, M2, M3, M4
Scottie (color):	S1, S2, S3, S4
Wraase SC-1 (color):	24, 48, 96 seconds

- Displays full color, real-time images during receive.
- The interface PC board fits into an IBM PC-compatible computer's expansion slot.
- Automatic switching between the microphone and SSTV audio for transmit.
- Does not require a separate power supply.
- Does not interfere with printer or serial port use.
- Graphical user interface—use a mouse and/or keyboard.
- Histogram-style on-screen tuning indicator.
- Images are stored as 320 × 240 pixels with 32,768 colors.
- Displays full-screen, 256-color images on standard VGA with dithering or custom palette options.
- Displays full-screen, 32,768-color images on Super VGA cards equipped with the HiColor feature.
- Reads and writes popular image-file formats (GIF, PCX and Targa).
- Test-pattern generation.
- Image manipulation, such as zooming.
- Requires a minimum IBM PC-compatible computer configuration of an 80286 (or later) processor, 640 kbytes of RAM and a VGA display adapter.

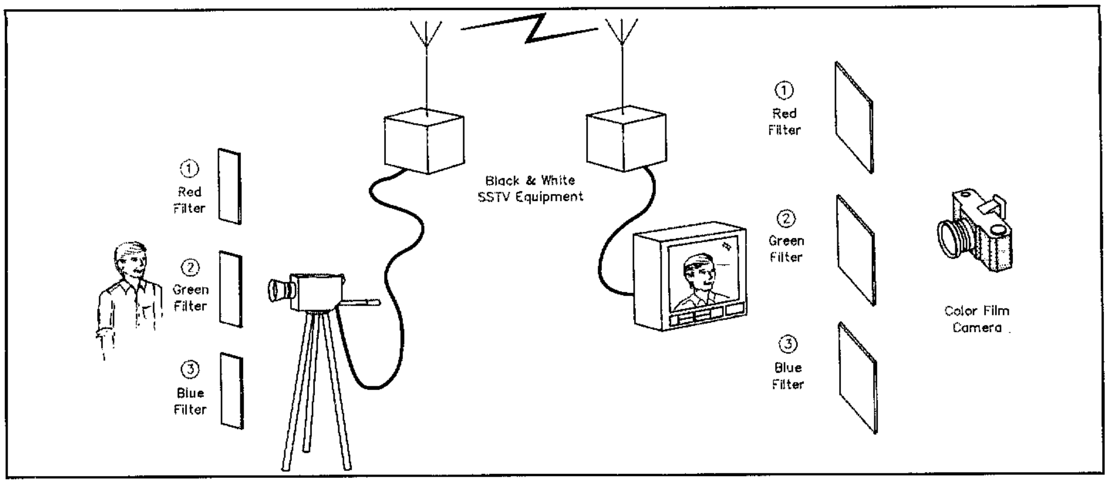


Fig 2—Early color transmission. At the transmitting end, red, green and blue filters were sequentially placed in front of the TV camera to send frame sequential signals. The receiver followed a similar procedure, taking multiple exposures of each color frame.

equipment. (See the sidebar "Examining Robot's Vertical-Interval-Signaling (VIS) Code" for more details.)

Although the Martin and Scottie modes have horizontal sync, some implementations ignore them on receive. Instead, they rely on very accurate time bases at the receiving and transmitting stations to keep in step. The advantage of this synchronous strategy is that missing or corrupted horizontal sync pulses won't disturb the received image. The disadvantage is that even slight timing inaccuracies will produce slanted pictures.

A couple of years ago, yet another incompatible mode was introduced. The *AVT* mode is different from all of the preceding in that it has *no horizontal sync!* To maintain synchronization, it relies on very accurate oscillators at the sending and receiving stations. If the beginning-of-frame signal is missed, it's all over. However, it's *much* harder to miss the 5-second AVT header than the usual 300-ms vertical sync pulse. Redundant information is encoded 32 times, and a more-powerful error-detection scheme is used. It's necessary to receive only a small part of the AVT header to achieve synchronization. After this, noise can wipe out part of the image itself, but the image alignment remains correct and the image retains the correct colors. These SSTV modes are described in more detail in several references identified in the end Notes and Bibliography. Also, see the sidebar "Popular SSTV Modes."

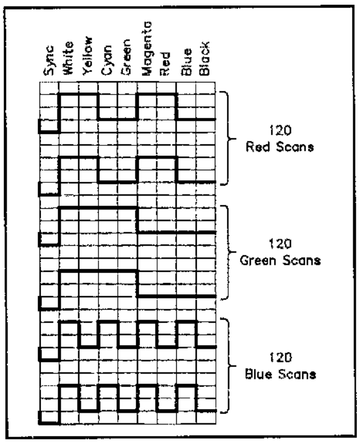


Fig 3—Waveforms of the frame-sequential color transmission employed in Fig 2, here using a color-bar test pattern.

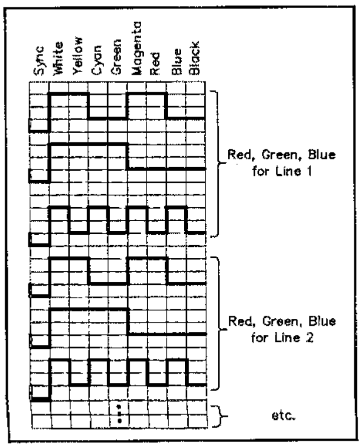


Fig 4—Line-sequential color transmission.

Examining Robot's Vertical-Interval-Signaling (VIS) Code

The original 8-second black-and-white SSTV-image standard used a 30-millisecond, 1200-Hz pulse to signal the beginning of a new frame. In the Robot 1200C, Robot Research increased the vertical sync period by a factor of 10, encoded 8 bits of digital data into it and called it *vertical-interval signaling* (VIS). VIS is composed of a start bit, 7 data bits, an even parity bit, and a stop bit, each 30 milliseconds long. (See Fig A).

Since then, inventors of new SSTV modes (Martin, Scottie, AVT, etc) have adopted Robot's scheme and assigned codes to their particular mode that are unused by the Robot modes. So, each of the SSTV transmission modes has a unique VIS code. This allows new equipment to automatically select any of the new SSTV modes while maintaining compatibility with the older equipment.—*WB2OSZ*

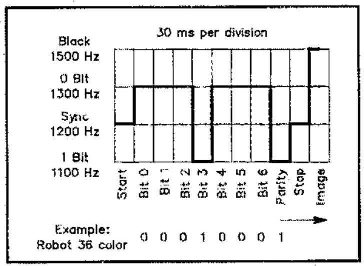


Fig A—Composition of the vertical interval signaling (VIS) code.

SSTV Transmission and Reception with a Computer

Transmitting SSTV images is easy. You simply generate fairly accurate tones and change them at the proper pixel rate. Tones in the range of 1500 to 2300 Hz correspond to the pixel intensities, and most modes use a 1200-Hz tone for sync pulses. A very low-cost SSTV system can even use the computer's built-in sound generator (such as of the low-priced Ataris and Color Computers [CoCos]). *What* you transmit depends on your resources; see the sidebar "Video Digitizers."

SSTV reception is trickier. First, you must measure the frequency of the incoming audio tone. You can't simply measure the number of cycles in a second, or even 1/100 second, because the frequency is changing a couple of thousand times a second. Fig 5 illustrates one method of rapidly measuring the incoming-tone's frequency. Band-pass filters are tuned a little beyond the ends of the frequency range of interest. The output of one filter is rectified to become a positive voltage; the output of the other filter is rectified to become a negative voltage; then the voltages are summed. A low-pass filter with a cutoff frequency of about 1 kHz removes the audio-carrier ripple, while allowing the slower video signal to get through. With careful filter design, the result is a voltage that's fairly proportional to the input frequency. Finally, an analog-to-digital (A/D) converter is used to allow the computer to recognize this voltage.

Another frequency-measuring approach uses digital circuitry to measure the period of each audio cycle (see Fig 6). At a zero crossing of the signal, a counter is reset. It then proceeds to count pulses from a crystal-controlled oscillator. At the end of the audio cycle, the counter content is snatched, the

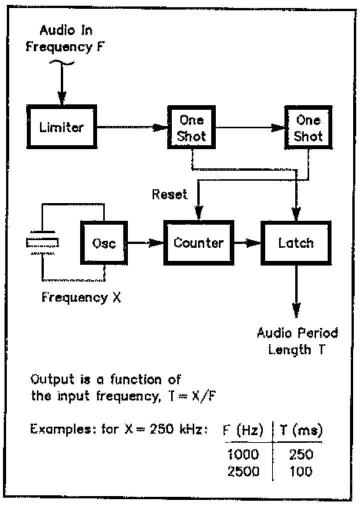


Fig 6—Block diagram of a digital SSTV demodulator.

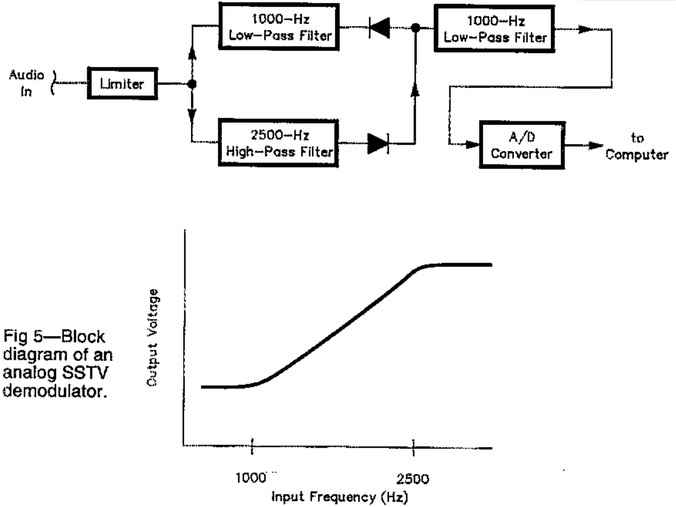


Fig 5—Block diagram of an analog SSTV demodulator.

Popular SSTV Modes

Family	*Name*	*Type*	*Time (Sec)*	*Lines*	*Notes*
Robot	8	B&W	8	120	A,E
	12	B&W	12	120	E
	24	B&W	24	240	E
	36	B&W	36	240	E
	12	Y&C	12	120	
	24	Y&C	24	120	
	36	Y&C	36	240	
	72	Y&C	72	240	
Wraase SC-1	24	RGB	24	128	C
	48	RGB	48	256	B
	96	RGB	96	256	B
Martin	M1	RGB	114	256	B
	M2	RGB	58	256	B
	M3	RGB	57	128	C
	M4	RGB	29	128	C
Scottie	S1	RGB	110	256	B
	S2	RGB	71	256	B
	S3	RGB	55	128	C
	S4	RGB	36	128	C
	DX	RGB	269	256	C
AVT	24	RGB	24	120	D
	90	RGB	90	240	D
	94	RGB	94	200	D
	188	RGB	188	400	D
	125	B&W	125	400	D

Types: B&W = Black and white.
Y&C = Color as luminance and chrominance.
RGB = Color as red, green and blue components.

Notes:
A—Very similar to the original 8-second SSTV standard.
B—The top 16 lines are always sent as gray scale, so 240 lines are usable.
C—The top 8 lines are always sent as gray scale, so 120 lines are usable.
D—AVT modes have a 5-second digital header and don't use horizontal sync.
E—The Robot 1200C doesn't actually have monochrome modes to send the luminance (Y) component of an RGB image in memory. Instead, it sends any one of the R, G, or B components.

You'll note that the number of pixels per scan line is not listed. That's because the demodulated slow-scan video is a continuously varying signal *that can be sampled at different pixel rates* depending on the implementation. Hardware-based equipment (eg, the Robot 400) used powers of two (such as 256) because they're easy to implement with hard-wired logic. Using software, it's just as easy to use any other number. Pasokon TV uses 320 pixels per line to get the proper aspect ratio with a VGA display.—*WB2OSZ* counter is reset and the process starts all over again. The digital approach offers a few advantages over the analog filter approach. A single chip can contain the counter and handle several other functions as well. The analog approach requires a handful of op amps, resistors, capacitors, diodes and finally, an A/D converter. The digital approach has crystal-controlled accuracy—*and no adjustments are required.* The frequency-to-voltage transfer function of the analog version isn't exactly linear and can change with temperature, power-supply variations and component aging.

Digital signal processing (DSP) is another exciting possibility for making SSTV demodulators. With DSP, a high-speed A/D converter is used to sample the audio input: All filtering is done in *software.* This approach can be used to construct filters that are more flexible, accurate, stable and reproducible than their analog counterparts.

Once you have the tone-frequency information, the real work begins. Next, you must separate the composite signal into its synchronizing (sync) and video components. To reduce the effects of noise, the sync pulses are cleaned up by a low-pass filter and Schmitt trigger. Then, the sync is used to control the pixel sampling. Fig 7 contains a high-level outline of a program used to receive an 8-second B&W image. Receiving color isn't that much more difficult. For nonRobot modes, gather the red (R), green (G) and blue (B) scans for each line, combine them, and display a line in full color. Robot modes require considerably more calculation to undo the encoding.

Pasokon TV Interface

Fig 8 contains the Pasokon TV interface schematic. Received audio appears at pin 11 of J1. C1, R2, C2 and R3 act as a high-pass filter to remove low-frequency noise. U1A and its associated components form a low-pass filter that reduces high-frequency noise. Limiter U1B reduces the effect of signal-amplitude variations. Finally, a Schmitt trigger (part of U6) is used to shape the signal in preparation for the digital circuits.

U3 is a member of the Motorola 68000 microprocessor family. It contains four counter/timers, parallel I/O, a universal asynchronous receiver/transmitter (UART)

```
Set line number, L, to 1
Repeat:
  Wait for sync
  Wait for end of sync
  If it was vertical sync, set L = 1

  Gather 128 pixels
  Display pixels on line L
  Increment L
  If L > 120, set L = 1
```

Fig 7—Outline of typical software written to receive an SSTV picture.

Video Digitizers

To get started transmitting on SSTV, you can use image files found on bulletin boards or prepared with a computer painting program. But to send live pictures of yourself, or objects inside or outside of your shack, you'll obviously need a TV camera (eg, a camcorder). You'll also require a *video digitizer* to get the video signal from the camera into your computer.* Many people mistakenly use the term *frame grabber* to describe all video digitizers. A frame grabber employs a lot of expensive, high-speed memory, an amount sufficient to hold *the entire image,* which it grabs in 1/30 second. Other video digitizers can be made at much lower cost by using less memory.

At the extremely low-cost end of the scale is one called Frugalvision. This clever little circuit uses only *six chips* and enough memory for *a single pixel!* It operates by grabbing *only one pixel per scan line.* During the first frame, it takes a pixel from the first column of each scan line. During the second frame, it snatches a pixel from the second column of each scan line, and so on. The trade-off is cost versus speed. For good results, you need a very stable video source—and the subject must remain *perfectly still* for *many seconds.* The amount of time depends on the number of columns, but it's about 8 seconds for 256 pixels and 21 seconds for 640 pixels.

Line grabbers are a reasonable compromise between these extremes. A line grabber has enough memory to hold a scan line. It grabs *all the pixels in the first scan line,* then transfers these to the computer. Then, it determines the next available line, grabs it, and transfers it to the computer. For example, a line grabber might take lines 1, 6, 11, 16, etc, from one frame, lines 2, 7, 12, 17, etc, from the next frame, and so on. The limiting factor is the *transfer time:* the time it takes to transfer the pixels from the line grabber memory to the computer. Total times are approximately 1/5 to 1/2 a second to capture the entire image, depending on the computer's speed. A stable video source and fairly stationary subject are required: But sitting still for a half second is a lot easier than remaining motionless for almost a half minute!

Here is a small sampling of typical, readily available video digitizers. All of them accept NTSC composite video. Some units also have S-video and RGB inputs; most have PAL versions available. This information was derived from the manufacturers' literature.—*WB2OSZ*

Product	*Resolution*	*Colors*	*Capture time (Seconds)*	*Price ($US)*
VIP-8800	(Not mentioned)	16 million	1/30	1995
CV-1010 + FG-1100	640 X 480	262,144	2	299 & 395
Computer Eyes/RT	512 X 512	16 million	1/30	599.95
Computer Eyes/Pro	640 X 480	16 million	24	399.95
DoubleTake AV+	640 X 480	16 million	1/2	345
VIP 640C	640 X 480	16 million	1/5	299.95
Frugalvision (kit)	256 X 244	64 gray scale	8	89.95 (kit)

Frugalvision is available from Idec, Inc, 1195 Doylestown Pike, Quakertown, PA 18951, tel 215-538-2600. The construction of this project is described in "Video Capture on the Cheap," *Radio Electronics*, Dec 1991, pp 37-46.

Computer Eyes/RT and Pro, Digital Vision, Inc, 270 Bridge St, Dedham, MA 02026, tel 617-329-5400, fax 617-329-6286.

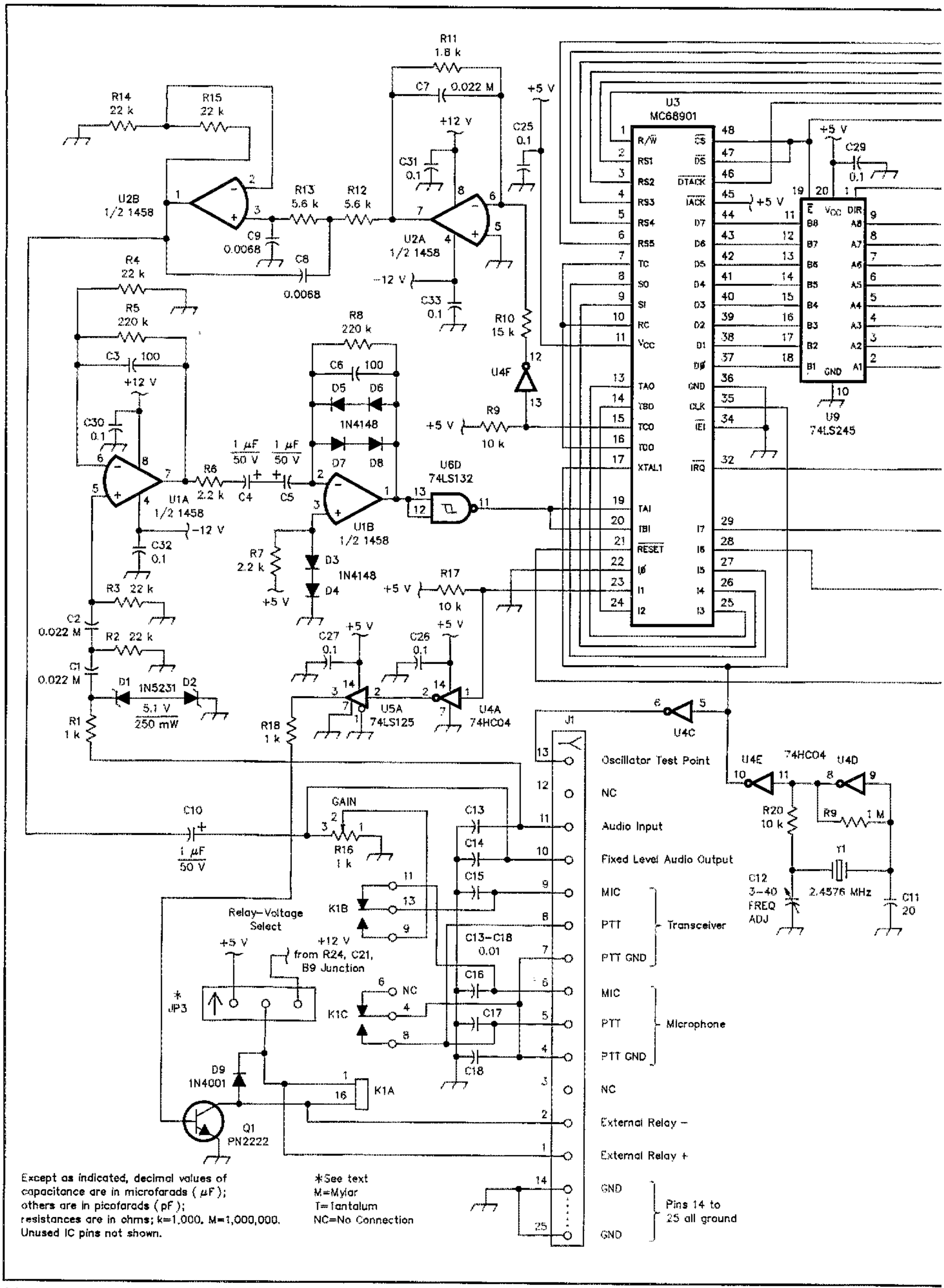
CV-1010 NTSC-to-RGB converter, FG-1100 Video digitizer with 4 inputs, Imaging Automation Inc, 7 Henry Clay Dr, Merrimack, NH 03054, tel 603-598-3422, fax 603-598-3422.

DoubleTake AV+, Logos Systems Int'l, 100 Royal Oak Ct, Scotts Valley, CA 95066, tel 408-438-5012, fax 408-439-9440. (This unit also contains an *audio* digitizer.)

VIP 640C, Ventek Corporation, 31336 Via Colinas, Suite 102, Westlake Village, CA 91362-9897, tel 818-991-3868, fax 818-991-4097.

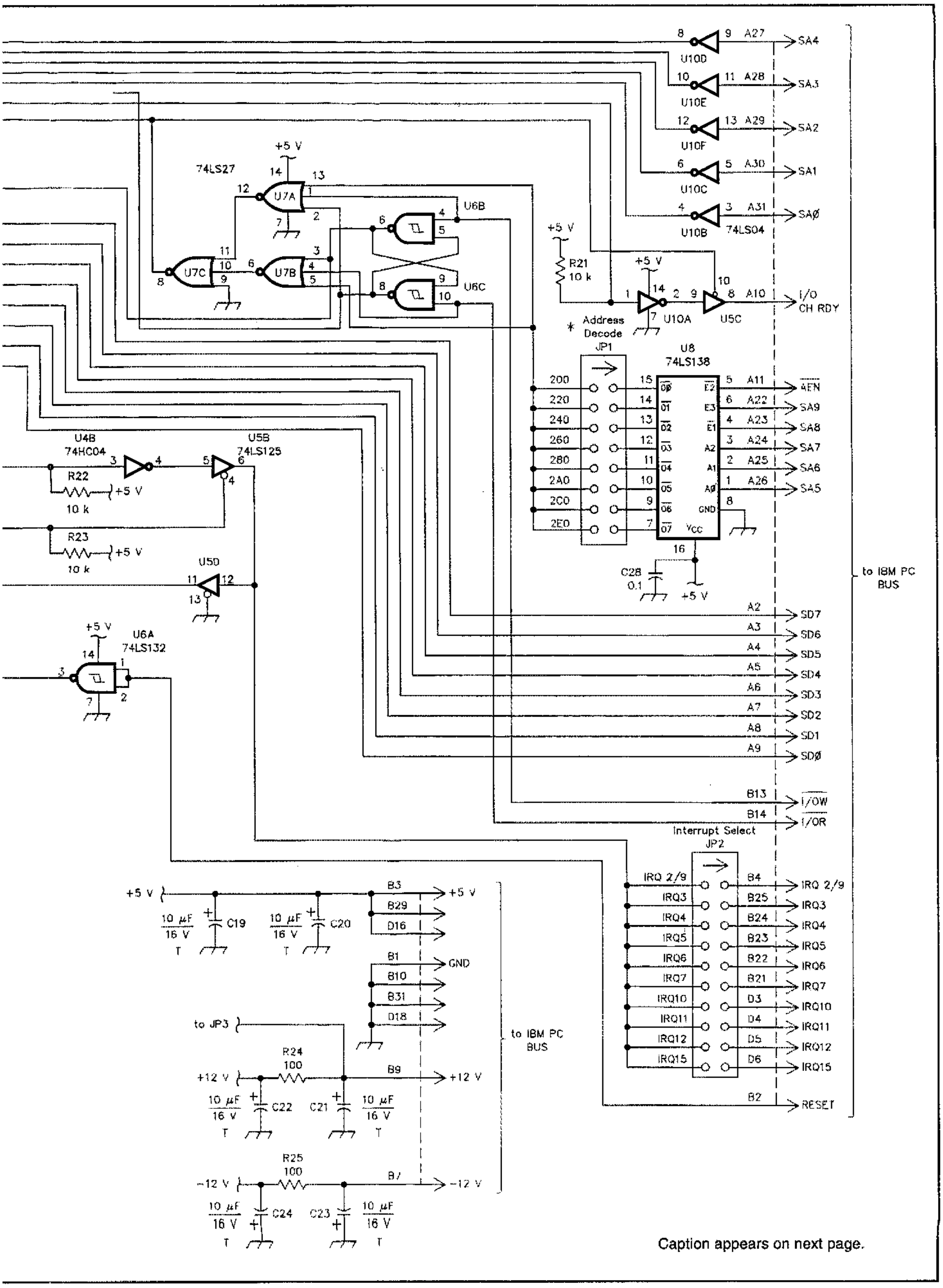
VIP-8800, IEV Corporation, 3030 S Main St, Salt Lake City, UT 84115, tel 801-466-9093 or 800-IEV-6161, fax 801-466-5921.

*By the time you read this, Pasokon TV should be able to take images directly from a VIP 640C digitizer. This means it won't be necessary to run a separate program to digitize images, save them to a file, then run Pasokon TV again and read the image files.



Except as indicated, decimal values of capacitance are in microfarads (µF); others are in picofarads (pF); resistances are in ohms; k=1,000, M=1,000,000. Unused IC pins not shown.

*See text
M=Mylar
T=Tantalum
NC=No Connection



Caption appears on next page.

Fig 8— (Figure appears on previous two pages.) Schematic of the Pasokon TV SSTV interface circuit. Unless otherwise specified, resistors are ¼-W, 5%-tolerance carbon-composition or film units. JP1, JP2 and JP3, are jumper positions; see text. (*QST* style ordinarily identifies jumpers as W numbers.)

C12—3- to 40-pF right-angle-mount trimmer capacitor.
D1, D2—1N5231 5.1-V, 250-mW, or 1N751 5.1-V, 250-mW Zener diode.
D3-D8—1N4148 silicon switching diode.
D9—1N4001 100-V, 1-A rectifier.
J1—DB25S right-angle PC-mount **D** connector.
JP1—8-position DIP male header (I/O address selection).
JP2—10-position DIP male header (IRQ selection).
JP3—3-position SIP male (relay voltage selection).
K1—5- or 12-V, DPDT, 16-pin DIP relay. (OMRON G5V-2 or Radio Shack #275-215.)
R16—1-kΩ, 15-turn, trimmer potentiometer.
U1, U2—1458 dual op amp.
U3—MC68901 multifunction peripheral.
U4—74HC04 CMOS hex inverter.
U5—74LS125 quad three-state buffer.
U6—74LS132 quad **NAND** Schmitt trigger.
U7—74LS27 triple 3-input **NOR** gate.
U8—74LS138 1-of-8 decoder.
U9—74LS245 octal transceiver.
U10—74LS04 hex inverter.
Y1—2.4576-MHz crystal.

Misc: PC board, rear-panel bracket with holes for a 25-pin **D** connector and access to C12 and R16 for adjustment; three shorting blocks for JP1, JP2, JP3; two 8-pin IC sockets (for U1 and U2); five 14-pin IC sockets (for U4, U5, U6, U7, U10); two 16-pin IC sockets (for U8, K1); one 20-pin IC socket (for U9); one 48-pin IC socket (for U3).

and an interrupt controller. Two of the counter/timers are used to measure the period of the incoming audio tone. Details of the timing are given in Fig 9. One counter measures the length of the positive half cycles. Another counter measures the negative half cycles. The samples are stored in variables Ta and Tb. Then, the audio frequency is computed as:

f = 1 ÷ (Ta + Tb) (Eq 1)

Another timer generates the tones used for transmission of the SSTV picture. A low-pass filter (U2) removes the high-frequency components before the signal is applied to the transmitter's microphone input. One of the parallel I/O bits is used to activate the relay (K1) that connects the transceiver's audio input to the SSTV tone generator and activates the PTT control. With a good old-fashioned electromechanical relay, you can always keep the interface connected between your microphone and transceiver. Less-expensive electronic switching isn't such a good idea here. With a chip instead of a relay, the SSTV interface must always be turned on, or you must swap cables to use the microphone.

The final counter/timer in U3 is the pixel-rate clock. If you're thinking about home-brewing your own SSTV system, be sure to use an *external* crystal-controlled timing source! With cache memories and interrupt handling occurring in the PC, any attempt to use software timing loops is doomed to failure. You should also resist the temptation to use the PC's programmable timer. This normally generates interrupts at about 18.2 Hz. Messing with this timing confuses the time-of-day clock, floppy-disk motor time-out, mouse double-click detection, screen savers and other time-dependent functions of the computer. The synchronous modes, such as AVT, require very accurate timing to avoid slanted pictures (see the sidebar, "*Slanted Pictures?*").

The rest of the schematic contains the interface to the computer's bus. Most of it's straightforward. U8 and jumper block JP1 are used to decode a group of 32 I/O addresses in the 2xx (hexadecimal) range. JP2 is used to pick an interrupt level not used by any other device in the system. The only "weird" part is the conglomeration of U6 and

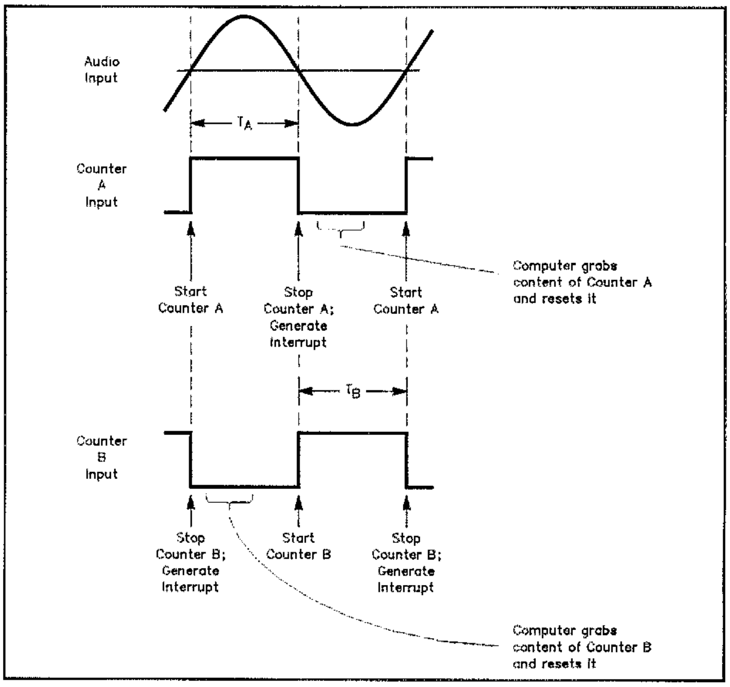


Fig 9—Details of input-period measurement.

Slanted Pictures?

As mentioned in the text, the synchronous modes, such as AVT, require very accurate timing to avoid slanted pictures. Let's quickly calculate the accuracy required.

One of the most popular SSTV modes is the AVT 94-second, which has 320 X 200 pixels using three color components per pixel. Multiply all that together and the result is 192,000 samples. (192,000 ÷ 2048 per second = about 94 seconds.) Suppose the maximum acceptable slant was an offset of 4 pixels at the bottom of the screen. This requires an accuracy of 4 ÷ 192,000 = 0.002% = 20 parts per million. Actually, if the transmitting station was off this much in one direction, and the receiving station was off the same amount in the opposite direction, the bottom of the picture would shift 8 pixels. Rather shabby, but it's the worst case. There's no need for a microcomputer time-of-day clock to have this accuracy, so many PCs probably won't even be this close. Be suspicious of *any* SSTV interface *that doesn't have its own crystal-controlled oscillator* that you can tweak to avoid slanted pictures!—*WB2OSZ*

U7 to process the I/O read and write signals. It might be more complicated than necessary, but it *absolutely guarantees* that the read/write control signal for U3 is stable slightly before, during and after the data strobe for reliable operation on a wide variety of systems.

Pasokon TV Operation

Fig 10 shows the screen layout for the software's main menu. It uses the 320- × 200-pixel, 256-color mode available on all VGA cards. The most frequent operations, such as mode selection, transmitting, receiving and full-screen image display, are activated by clicking a mouse button on a control-panel button on the screen. Less-common operations—such as reading files, generating test patterns, or zooming in on part of an image—are selected from pull-down menus. Some of them cause dialog boxes to appear for further interaction. Most features can also be accessed with keyboard equivalents, so a mouse isn't absolutely necessary to run the program.

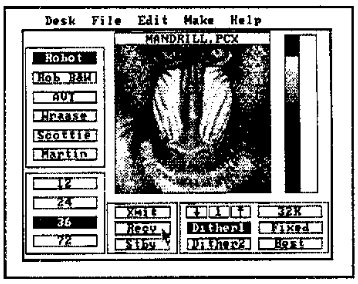


Fig 10—The Pasokon TV software main screen display. Although the software is mouse-oriented, keyboard commands can also be used to call up its various menus and help screens.

A Typical (?) SSTV Contact

Amanda Amateur fires up her Pasokon TV system and tunes to one of the SSTV calling frequencies: 3.845, 7.171, 14.230, 14.233, 21.340, 28.680 or 144.5 MHz. There, she hears Hambleton Ham calling CQ:

HH: "CQ CQ CQ. CQ Slow Scan. This is Kilowatt One Hotel Oscar Hotel."

AA: "K1HOH, this is Amanda, W7AA. Hello, Ham."

HH: "Hi, Amanda! Say, I just finished home-brewing a slow-scan TV system and need some help in testing it."

AA: "OK, let's see some pictures." Amanda uses a mouse to click on the **Recv** menu button, or presses **Enter** on the keyboard to begin receiving.

HH: "Here's a picture of my cat, Figaro, in Scottie S1." He sends the picture. (Note that *he announces the mode before transmission* because not all systems automatically select modes from the VIS code.) "How was that?"

AA: Amanda presses any key or mouse button to end receive mode. "That's a beautiful picture of a very talented animal!"

HH: "Talented? What do you mean talented?"

AA: "He's holding onto the ceiling! I'll send the picture back to you. Which mode would you like?" (Dear reader: The cat is actually *standing on the floor,* but Amanda's decided to play a joke on HH and send the picture *upside down* to HH. [It's certainly not a way to win friends with experimenters! HH better have a sense of humor, or AA is going to wind up in first position on HH's (expletive deleted) list!—*Ed.*])

HH: "How about Robot 72 second?"

AA: Amanda clicks on the **Robot** menu button and the color Robot transmission speeds appear below. She clicks on the **72** button. (Alternatively, she could have used the keyboard function keys.) Next, she picks **Invert** from the **Edit** menu to turn the picture upside down. There's no need to convert an image from one format to another—it's fully automatic.

"Here's your talented cat in Robot 72." (She clicks on the **Xmit** menu button, or presses the **X** key.)

HH: "H-m-m-m-m-m. The picture *is* upside down! Very interesting! How about sending me something in AVT 94?"

AA: With the mouse, Amanda selects **Open** from the **File** menu and gets a list of subdirectories and files. She clicks on the **DOGS** subdirectory and scrolls through the list until **BLACKLAB.GIF** is visible. After clicking on the file name and the **OK** button, the file selector disappears. Next, she selects **Rotate** from the **Edit** menu to rotate the image by 90 degrees and **Negative** to reverse the colors. "Here's a picture of my black Labrador Retriever in AVT 94." She clicks on **AVT**, **94,** and **Xmit.**

HH: "The picture appears to be rotated by 90 degrees and the colors are all wrong. The dog is white and the grass is magenta! Dang! This seemed like such an easy program to write, but I've clearly made some serious mistakes! I'll have to go do some debugging. Thanks for your help, Amanda. 73! K1HOH."

AA: "Glad to help! I'll be here for a while. Give me a call when you're ready to try a new version. 73! W7AA clear and listening."—*WB2OSZ*

Although all images are stored as 320 × 240 pixels with 32,768 colors, the image portion of the main screen is only 160 × 120 pixels with 256 colors to be compatible with even the oldest and cheapest VGA card. The control-panel buttons in the lower-right screen corner select a full-screen image display. The **Fixed** button is the fastest, with a predetermined palette of 256 colors. The two **Dither** buttons produce the illusion of thousands of colors when you stand back 10 or 15 feet from the display. (Up close, it doesn't look so good.) The **Best** button calculates a custom palette of 256 colors for the particular image. This calculation can take a minute or more on a 33-MHz 80386 machine (math coprocessors aren't used by the software), so think before using the **Best** button on a slower machine. To get a feel for what an SSTV QSO might be like, see the sidebar "A Typical (?) SSTV Contact."

Many super VGA cards have a "HiColor" feature that allows the simultaneous display of 32,768 colors. The **32K** button selects this display mode to create beautiful pictures. If you're planning to buy a VGA card, consider one with the HiColor feature—they're not overly expensive.

Construction

Construction is fairly simple, especially with the available PC board (see Note 6). Wire wrapping the circuit takes a *lot* of patience because you have to make hundreds of connections. If you do use the wire-wrap approach, keep the interconnecting wires short (especially those connected to the PC bus!) and use plenty of bypass capacitors on the power-supply lines. Placing the variable capacitor (C12) and potentiometer (R16) near the rear edge of the board allows them to be adjusted from the back of the computer without removing its cover.

To help preserve the almost-lost art of home-brewing, I'll send a free copy of the software and additional documentation to anyone who sends me a photograph of their home-brewed Pasokon TV interface.

Adjustments

The Pasokon TV construction/instruction manual provides detailed information for the whole system. Fundamentally, you make two adjustments and install three SIP jumpers:

- Adjust C12 to put the 2.4576-MHz oscillator on frequency.
- Adjust R16 for the desired audio-output level.
- Install jumper JP1 to select the I/O address.
- Install jumper JP2 to select the desired IRQ level.
- Install jumper JP3 to select the relay operating voltage (5 or 12 V dc).

That's it.

Summary

After I gave several SSTV presentations and taking quick audience polls, a few trends became clear: (1) There is great interest in SSTV; (2) most people never get started because they think it must be very expensive; (3) most hams have an IBM PC-compatible computer. For them, using an SSTV interface with an IBM PC-compatible computer is the only *sensible* approach. (Table 3 contains additional information on how you can find out more about SSTV.)

I'm not the only one to come to these conclusions. Since the time this article was written, two other PC-based SSTV systems have become available; at least one more is rumored to be under development. There will probably be more new SSTV users in 1993 than any other year since it was invented!

Table 3

For More Information About SSTV . . .

Weekly nets:

Saturdays, 1500 and 1800 UTC
14.230 MHz

Magazines specializing in ATV:
The SPEC-COM Journal
PO Box 1002
Dubuque, IA 52004-1002
Tel 319-557-8791, fax 319-583-6462.
This journal features the column "Satellites, Facsimile & Slow-Scan TV Imaging" conducted by Fred Sharp, W8ASF.

*CQ-TV**
British Amateur Television Club
Dave Lawton, G0ANO
Grenehurst, Pinewood Road
High Wycombe, Bucks HP12 4DD
England
Tel 0494 28899
This publication features a frequent column "SSTV Revisited" conducted by Roland Humphreys, G4UKL.

Amateur Television Quarterly
1545 Lee St
Suite 73
Des Plaines, IL 60018
Tel 319-266-7040

Old A5 Magazine *reprints:*
ESF Copy Service
4011 Clearview Dr
Cedar Falls, IA 50613
Tel 319-266-7040

Amateur Television Today!
QCD Publications
PO Box 1677
Weatherford, TX 76086
Tel 817-599-7509

Handbooks:
SLOW SCAN TELEVISION HANDBOOK
Don Miller, W9NTP, and Ralph Taggart, WB8DQT
2nd ed (Peterborough, NH: 73 Inc.)
2nd printing March 1975; 1st ed printed Dec 1972.

*Subscriptions to *CQ TV* can be arranged through Wyman Research or *ATV Quarterly* so US subscribers don't have to deal with foreign currency exchanges.

Notes

[1] Robot Research was the dominant supplier of SSTV equipment for many years, first with the Robot 400, then the 400C and finally, the 1200C, which arrived around 1985.

[2] AVT Master available from AEA. See ad in any ham magazine.

[3] SSTV (kit #168) available from A & A Engineering, 2521 W LaPalma, Unit K, Anaheim, CA 92801, tel 714-952-2114, fax 714-952-3280. Charge cards accepted.

[4] J. Langner, "SSTV—The AVT System Secrets Revealed," *CQ-TV 149*, Feb 1990, pp 79-80.

[5] M. Emmerson, "SSTV Compatibility! Robot 1200C and Wraase SC-1," *Spec-Com Journal*, p 39, no date.

[6] Complete kits are available for $199.95 from: John Langner, WB2OSZ, 115 Stedman St #Q, Chelmsford, MA 01824-1823, tel: 508-256-6907. Kits include a *four-layer* PC board (with plated through-holes and gold-plated edge-connector, solder mask and silk-screen), all board-mounted components, software and printed documentation. Kits *do not* include the connectors required for attaching the interface to a transceiver because the connectors used differ among transceiver manufacturers. For assembled and tested units, add $30. Specify your preference of 5.25-inch (1.2-Mbyte) or 3.5-inch (1.44-MB) diskette. Free shipping in the USA; foreign orders add $15 for shipping and handling. Massachusetts residents must add 5% sales tax.

To use the Pasokon TV system, you need an IBM PC AT or compatible computer operating under PC/MS-DOS with a 80286 (or faster) CPU, a minimum of 640 kbytes of RAM and a VGA display. A mouse is strongly recommended, but not required.

Because this is a *four-layer* PC-board, the ARRL *does not* offer a PC-board layout or part-overlay template.

[7] C. MacDonald, "A Narrow-Band Image Transmission System—*Part I*," *QST*, Aug 1958, pp 11-15, 140, and 142; —*Part II*, Sep 1958, pp 31-36, 146 and 148;

[8] Section 9 of Robot model 1200C color scan converter instruction book. Also in ESF reprint booklet, number 110, p 43.

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J. Montalbano, "The ViewPort VGA Color SSTV System," *73 Amateur Radio Today*, Aug 1992, pp 8-16.

B. Brown, ATV, *73 Amateur Radio Today*, Aug 1992, pp 50-53. (This column contains a discussion of several video digitizers.)

D. Goodman, "SSTV with the Robot 1200C Scan Converter and the Martin Emmerson EPROM Version 4.0," *73 Amateur Radio Today*, Jul 1991, pp 46-48.

J. Langner, "SSTV—The AVT System Secrets Revealed," *CQ-TV 149*, Feb 1990, pp 79-80.

J. Langner, "Color SSTV for the Atari ST—Part I," *73 Amateur Radio*, Dec 1989, pp 38-42; Part II, Jan 1990, pp 41-43.

G. Cameroni and G. Morellato, "get on SSTV—with the C64," *ham radio magazine*, Oct 1986, pp 43-47, 51 (translated by Jim Grubbs, K9EI).

R. Taggart, "The Romscanner," *QST*, Mar 1986, pp 21-27. [Editor's note: By the time this article appears in print, V 2.0 of the Romscanner will be available. Circuit changes obsolete the older schematic. PC boards and a new schematic copy are available from FAR Circuits, 18N640 Field Ct, Dundee, IL 60118-9269; price not available at press time. A PC-board template and part-overlay drawing are available free from ARRL HQ. Please send a business-size SASE and address your request for the TAGGART ROMSCANNER V 2.0 PC-BOARD TEMPLATE to: Technical Department Secretary, ARRL, 225 Main St, Newington, CT 06111.

C. Abrams, "In Search of the Perfect Picture," *QST*, Dec 1985, pp 14-17.

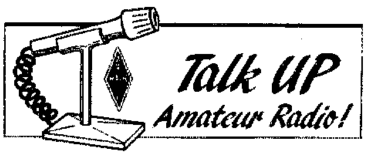
M. Schick, "Color SSTV and the Atari Computer," *QST*, Aug 1985, pp 13-16.

C. Abrams and R. Taggart, "Color Computer SSTV: Part I," *73 Magazine*, Nov 1984, pp 10-13, 16, 18, 20 and 21; Part II, Dec 1984, pp 18-22, 24-26 and 28-32.

John Langner, WB2OSZ, obtained his ham license around 1970 at the Tech Hams ARC (thanks, Don, WA2NRE!). John has a BS in computer science from the Rochester (NY) Institute of Technology. John is an incurable home-brewer and has had a number of construction articles published in BYTE, 73 Magazine, ham radio *and* QST.

For the last several years, John has dedicated himself to making low-cost SSTV available for everyone. When not busy tinkering with ham radio gadgets, spending time with his wife and three sons (and getting some occasional sleep!), he's a Member of the Technical Staff at Advanced Visual Systems, Inc, a leading supplier of scientific visualization software.

Phil Calciano, KA1KZB, provided all the SSTV pictures appearing in this article. (other photos by Kirk Kleinschmidt, NT0Z.)



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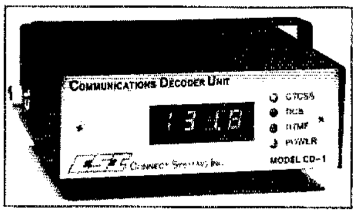

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The Beat of a Different Drum: The Cop MacDonald Story

Care to speculate on the life and times of the father of SSTV? Unless you're "in the know," you'll probably be way off base....

By Rick Booth, KM1G
232 Washington Street
Norwood, MA 02062

Whatever happened to W9OLS? If you recognize the call sign, chances are you're among the faithful. It dates to the birth of slow-scan television (SSTV), a mode many acknowledge that its holder, Cop MacDonald, founded. The man who was W9OLS then, and who is today VY2CM, clearly remembers SSTV's seminal moment. It was 1953:

"I was at engineering school at the University of Kentucky, in the library reading a paper by a guy from Bell Labs. Bell was working on a way to send signatures by telephone. As I thought about it, it suddenly dawned on me that TV didn't have to go at its usual rate." Scan rate, that is.

If slow-scan television interested Ma Bell, it thoroughly captivated Cop. He was already a ham, so the next step was almost a foregone conclusion: Let's get SSTV on the air! But this was 40 years ago, and the technology didn't just jump out and grab him. Then as now, the lure was in the challenge.

"It took seven or eight seconds for the glow of the P7 phosphor display tube to die," Cop says, "so one parameter for our early SSTV experiments was a six-to-eight second frame time. That was all you could get." The signal had the bandwidth of a voice signal—an AM voice signal, from 50 Hz to 3 kHz.

Cop's interest attracted more and more converts, and gradually the infant SSTV movement gained ground. Progress must have seemed agonizingly slow to those early devotees; not until mid-1958, Cop said, did *QST* offer the first comprehensive article on the new mode. Predictably, *QST* lit a fuse and the responses came from around the globe. Slow-scan was taking off.

Cop remembered when things were a little less refined:

"The first system had an AM subcarrier—the one the first transatlantic tests were done with." It had a six-second scan rate and a 20-Hz horizontal rate. "It took a year or so to switch to a frequency-shifted tone on the subcarrier." By then, SSTV systems had progressed to an eight-second scan, with a 15-Hz horizontal rate.

Between 1958 and 1968, SSTV didn't have a formal place in FCC regulations, living from a suitcase of temporary approvals. But its devotees—"a core group of super enthusiasts"—carried the flame until 1968, when the commission formally recognized SSTV as an amateur mode.

Cop MacDonald's star was ascending at the same time, and not just in amateur circles. His light burned increasingly bright at a series of technology-based companies: ITT, Westinghouse and Ball Brothers Research. In the New York City of 1970, contemporary social ferment found ample food in Cop's formidable intellect:

"I was already director of research and development for a small company, and I figured I could become vice president of a mid-sized company—*if* I was willing to work 60 hours a week. I wasn't."

Unsure whether to work in the corporate world at all, Cop turned his back on what was then a new buzz word, "The Establishment," and walked. He kept walking, too—right around the world.

With his wife and 11-year-old daughter at his side, Cop set out out on the trip of a lifetime: a backpacking tour of the globe. The trio went first to South America, then to the islands of the Pacific, and then on to Asia, Europe, Africa and home. Or at least back to the USA. It no longer seemed like home. The States, to his thinking, had profoundly changed: "Nixon was doing his Watergate thing."

The man who once questioned his corporate place now questioned his native place, and the echoes didn't please him. Cop looked elsewhere to live, but his travels warned of a life in certain quarters abroad. Canada seemed the best compromise, with a language and culture (then) linked in what Cop still calls "this overtone of caring." The nomad sank roots in Canada's Maritime provinces.

The son of a food salesman and a homemaker, Cop MacDonald grew up in the suburbs of several major cities, especially Chicago. He got his ham license at age 15, which opened the way for what he now calls, "my first career, as your standard sort of corporate engineer."

After transplanting to Canada, Cop's first vocational ventures veered radically from his technical past: "I explored human services, first as a nursing orderly and later in geriatric counseling. Although relatively short-lived, the episode did have one lasting effect.

"The prevailing theory on people has always seemed to be either someone is good at intellectual pursuits, but a lifeless nerd; or artistic but flaky." Cop found he could train his other right-brained self. "It wasn't an either/or thing at all." He had already discovered a skill for explaining technical subjects to nontechnical audiences, a knack that led him into writing. Since 1972, three years before he moved to Canada, Cop had been writing for magazines.

"I started doing a column on SSTV for *CQ Magazine*. Then a year later, I began a column for *Mother Earth News*, called New Horizons Radio." The *CQ* column was fairly straightforward technical material. The other teetered on radicalism.

Actually, it was the beginning of what we now know to be the worldwide ecological movement. But in the early 1970s, it was *avante garde*, just the thing for a bright light like Cop MacDonald. *Mother Earth News* gave him a chance to "promote ham radio as a tool for the back-to-landers, plus talking about something besides the rig and the weather." His column, New Dimensions Radio, ran until 1983.

In 1980, Cop began four years as the head of a government-sponsored energy conservation program in the province. Since then, he's made his living writing for public- and private-sector clients, meeting the ceaseless need to translate technical material into something understandable.

That writing may pay the rent, but it's not where Cop's heart is at these days. His life began to change 15 years ago, at his first meditation retreat. "My high-speed left brain got quiet for the first time since I was about two years old. And you know, when that happens, you can really see how things are." Cop's been meditating since, meditating his way into a book: *Toward Wisdom: Finding Our Way to Inner Peace, Love, and Happiness*, only just published in Ontario.

"It's (about) a way of looking at reality which hams are familiar with," Cop said, "because reality is an enduring medium, modulated by information."

If that sounds like new intellectual ground, why not? Cop MacDonald is still tilling the soil of the mind. Only the field is different.

High-Performance, Single-Signal Direct-Conversion Receivers

The direct-conversion receiver described in August 1992 *QST* featured high dynamic range, low-distortion audio and a super SSB filter shape factor. What more could you ask for? How about adding an image-reject mixer to that basic design for great opposite-sideband rejection?

By Rick Campbell, KK7B
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Michigan Technological University
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Direct-conversion receivers are capable of outstanding performance. The The high-performance receiver described in August 1992 *QST*[1] has now been used from 25 kHz to 6 GHz with excellent results. Nicknamed "R1," that receiver works well and sounds good because it combines several desirable traits:

- high third-order, two-tone dynamic range;
- moderate noise figure;
- low distortion from antenna to speaker leads; and
- 60 dB of output signal-to-noise ratio.

The significant flaw in R1 is that it has no opposite-sideband rejection. This flaw can cause real problems when you try to use the receiver for serious listening. On the crowded HF bands, the opposite sideband is almost always occupied by an interfering signal or two. On VHF and microwaves, the noise in the opposite sideband reduces the signal-to-noise ratio by up to 3 dB.[2]

There are two ways to get rid of the opposite sideband: (1) a narrow filter before the downconverter; or (2) an image-reject mixer. A fixed-frequency direct-conversion receiver preceded by a narrow filter and tunable converter is a conventional superhet. The image-reject mixer is less familiar.

Fig 1 is the block diagram of an image-reject mixer, along with the mathematical relationships between the local oscillator (LO) and signals above and below the LO frequency. Readers comfortable with trigonometric identities may verify that signals above the LO frequency cancel at the output, while signals below the LO frequency add. Less ambitious readers may simply recognize the block diagram of a phasing SSB exciter (from any radio handbook published in the last 30 years), with the arrows turned around.

Old timers will remember the phasing-type SSB rigs of the 1950s and '60s. Those memories may be a bit less than fond. Maintaining amplitude balance to less than 0.1 dB and phase error of less than 1° in a band-switched, vacuum-tube-and-paper-capacitor transmitter was truly a nightmare. The rigs often sounded bad on the air. Many operators just gave up trying to adjust their rigs after a while and transmitted poor signals until they could afford a filter-type SSB radio.

The situation is reversed today. The required phase and amplitude tolerances are easy to obtain with modern components, and a properly designed phasing exciter has fewer adjustments (and fewer spurs) than a filter rig!

Although image-reject mixers and phasing SSB receivers are scarce in the North American Amateur Radio literature, they are very common in professional circles and in other parts of the world. *SPRAT*, the journal of the G-QRP club, published in England, has presented *three* HF phasing receivers in the past year. Commercially, the Drake R8 shortwave receiver uses an image-reject mixer to downconvert from a VHF first IF to a 50-kHz second IF, and the Kenwood TS-950SDX uses a digital signal processing (DSP) phasing exciter and detector.

R1 Becomes R2

With a wealth of excellent technical material at hand, in particular the outstanding papers by Oppelt,[3,4] I did not have to start

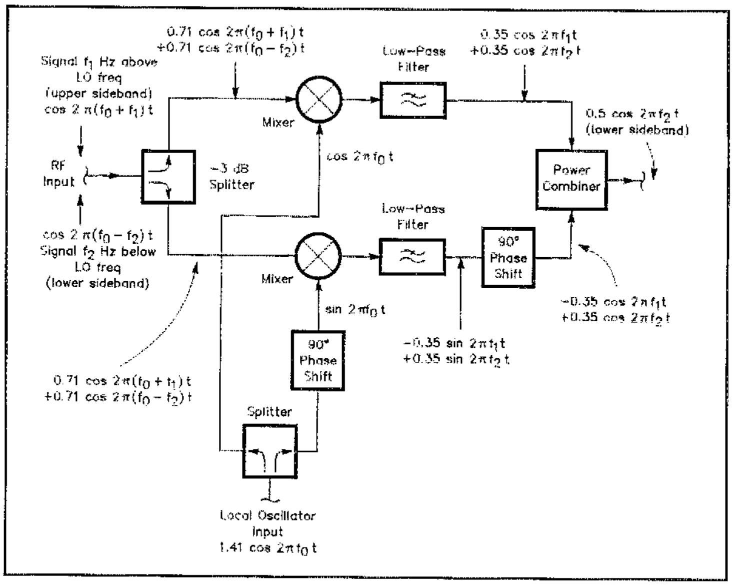

Fig 1—Block diagram of an image-reject mixer. The sines and cosines represent the local oscillator signal and signals above and below the LO frequency. Signals above the LO frequency cancel at the output, while signals below the LO frequency add.

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[1] Notes appear on page 40.

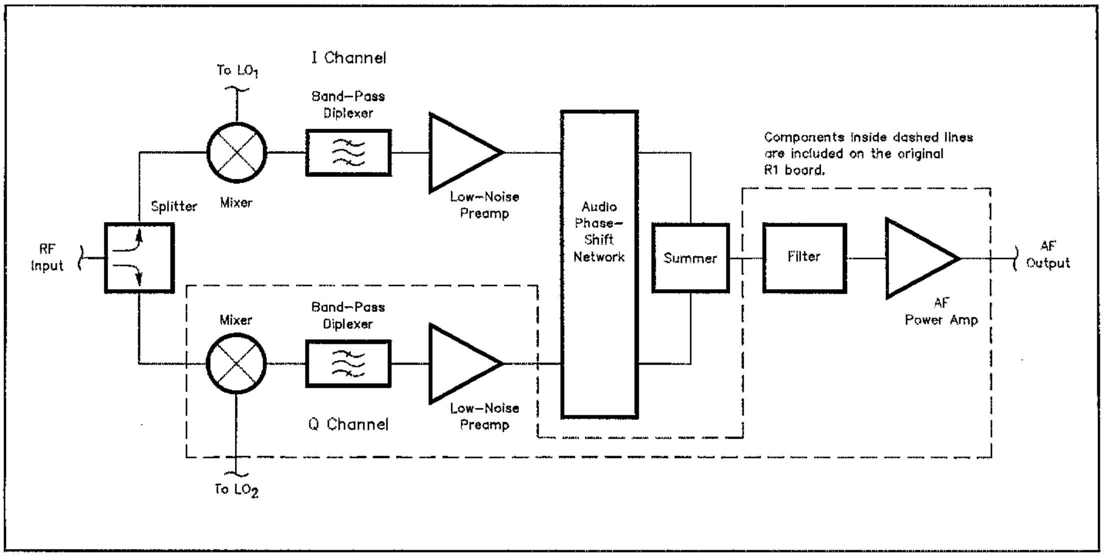


Fig 2—Block diagram of the R2 single-signal direct-conversion receiver. The sections inside the dashed lines are included on the original R1 board. Added are an RF splitter, second mixer diplexer and low-noise preamp, an audio phase-shift network and a summer.

from scratch to design a single-signal direct-conversion receiver. My goal was to add single-signal capability to the R1 board without degrading its otherwise outstanding performance.

Fig 2 is the block diagram of the R1 receiver board, with the necessary blocks added to suppress one sideband. The resulting receiver is called "R2." The additional parts on the R2 board are an RF splitter, a second mixer, diplexer and audio preamp, and an audio phase-shift network and summer using a pair of quad op amps. For maximum flexibility, the 90° phase-shift network for the LO is off the board. The complete schematic is shown in Fig 3. All of the parts fit comfortably on a 3½ × 5-inch double-sided PC board.[5]

Circuit Details

The first component the input signal sees is the RF input splitter. I used a Toko TK2518 because it's cheap ($2), small, and available from Digi-Key.[6] It is rated from 20 MHz to 600 MHz, but I tried one on 40 meters and it works. A Mini-Circuits PSC2-1 splitter worked fine as well, but it is a lot more expensive ($12) and will have to be mounted off the board. For operation below 20 MHz, a home-brew alternative to commercial splitters is shown in Fig 4. I am using the one in Fig 4 on 40 meters.

The RF level at each mixer is 3 dB below the RF-input signal level. That translates to an improvement in the input third-order intercept point. Since there is also a 3-dB (maximum) improvement in signal-to-noise ratio, the R2 board has intrinsically better dynamic range than the R1 board. I haven't built an R2 board with high-level mixers, but based on measurements of the high-level R1 board and the R2 board with SBL-1 mixers, it should have a two-tone, third-order IMD dynamic range of greater than 100 dB.

After the splitter, the input signals are multiplied with the 90°-out-of-phase LOs, filtered and amplified in a pair of identical channels. They are referred to as the I channel (for "*I*n phase") and the Q channel (for "*Q*uadrature," a fancy word for "90° out of phase"). Amplitude and phase-shift errors in the I and Q channels must be carefully controlled. If there is a difference in the conversion loss of the two mixers or the gain of an audio preamp, it can be compensated by adjusting the **AMPLITUDE BALANCE** potentiometer (R30).

If there is an amplitude difference that is a function of frequency, however, R30 can only correct for it at a single audio frequency. Since the diplexer network between the mixer IF port and the audio preamp in each channel has an amplitude and phase characteristic that varies rapidly with frequency over the audio range, it is necessary to carefully select the components marked with asterisks in the schematic. Each component needs to be within 1% of the value of its counterpart in the other channel. The matched pairs may vary from the schematic value by up to 10% with no significant change in performance, *but they must be within 1% of each other.* I bought 10 of each of the marked parts and matched them with an RLC bridge. I obtained four sets matched to within 1%, and built R1 boards with the extras.

In an ideal world with ideal components,

Fig 3 (see next page)—The R1 receiver board circuit board holds all of the non-frequency-sensitive components. The local oscillator, LO phase-shift network and front-end filter are located off the board. The resistors in the audio phase-shift networks (R20-R28 and R35-R43) are 1%-tolerance metal-film units. Other resistors are 5%-tolerance carbon-film or composition units. Inductors are Toko 10RB series fixed inductors. Polarized capacitors are aluminum electrolytics rated at 16 V dc. The capacitors in the audio phase-shift networks (C21-C23 and C26-C28) are 1%-tolerance XICON polyester-film units available from Mouser (see text and Note 5). The capacitors in the diplexers (C1, C3-C5, C11, C13-C15) and 300-Hz high-pass filter (C43, C44) are Panasonic type ECQ-E(F) 100-V, 10%-tolerance miniature metalized polyester film units. C32-C37 in the low-pass filter are Panasonic V-series 50-V, 5%-tolerance metallized film capacitors; C31 is a 50-V Panasonic P-series polypropylene capacitor. C38, C46 and C48 are disc-ceramic capacitors. The other capacitors can be metalized polyester or ceramic-disc capacitors. See Note 5 for part sources.

C2, C12—0.001-µF chip capacitor. These parts are necessary only if the board is used at VHF or UHF.

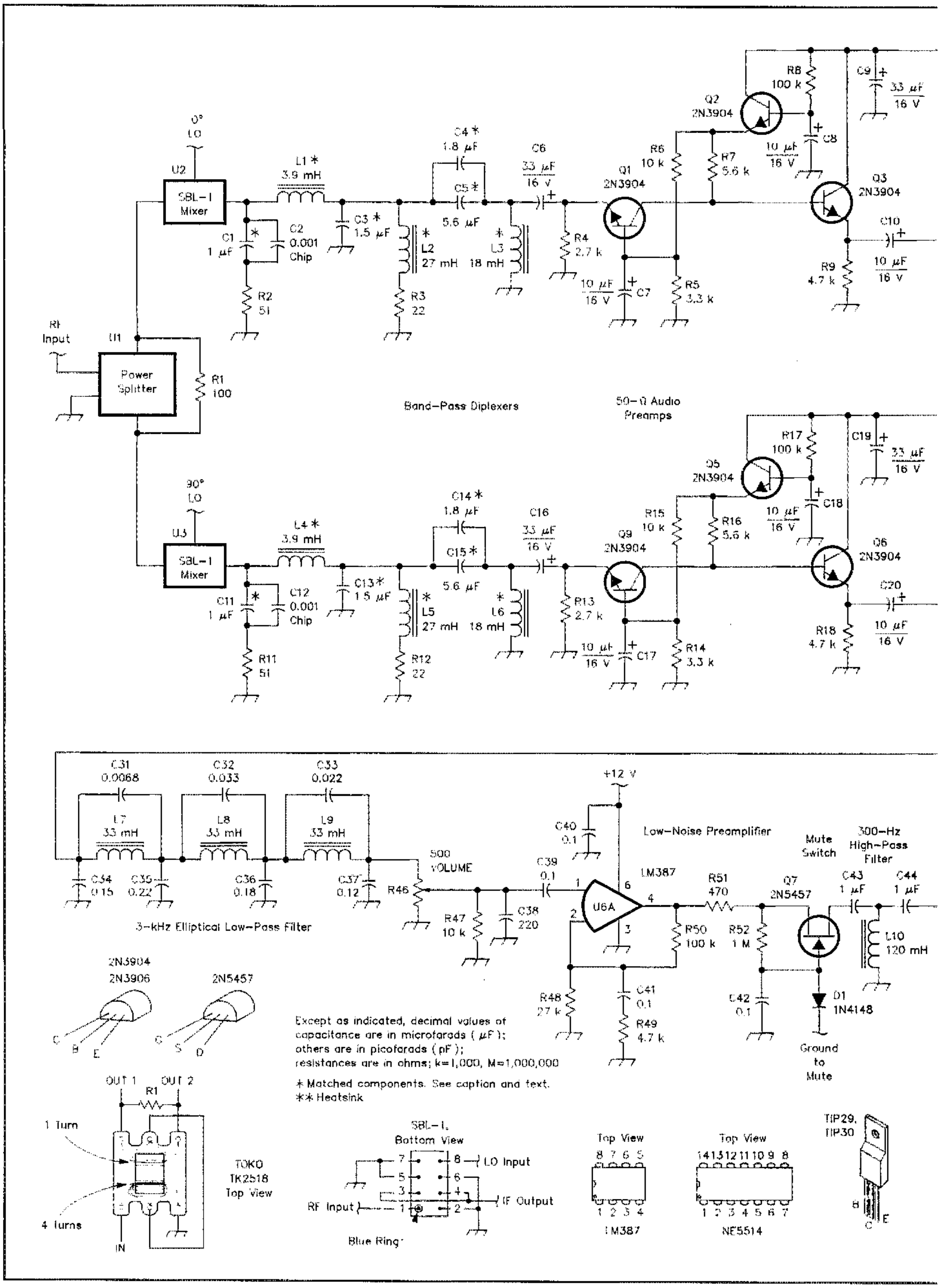
R30—10-kΩ single-turn PC-board-mount control (Panasonic MAG14 or equivalent).

R53—500-Ω audio-taper control.

R57—10-kΩ single-turn PC-board-mount control (Panasonic MAG14 or equivalent).

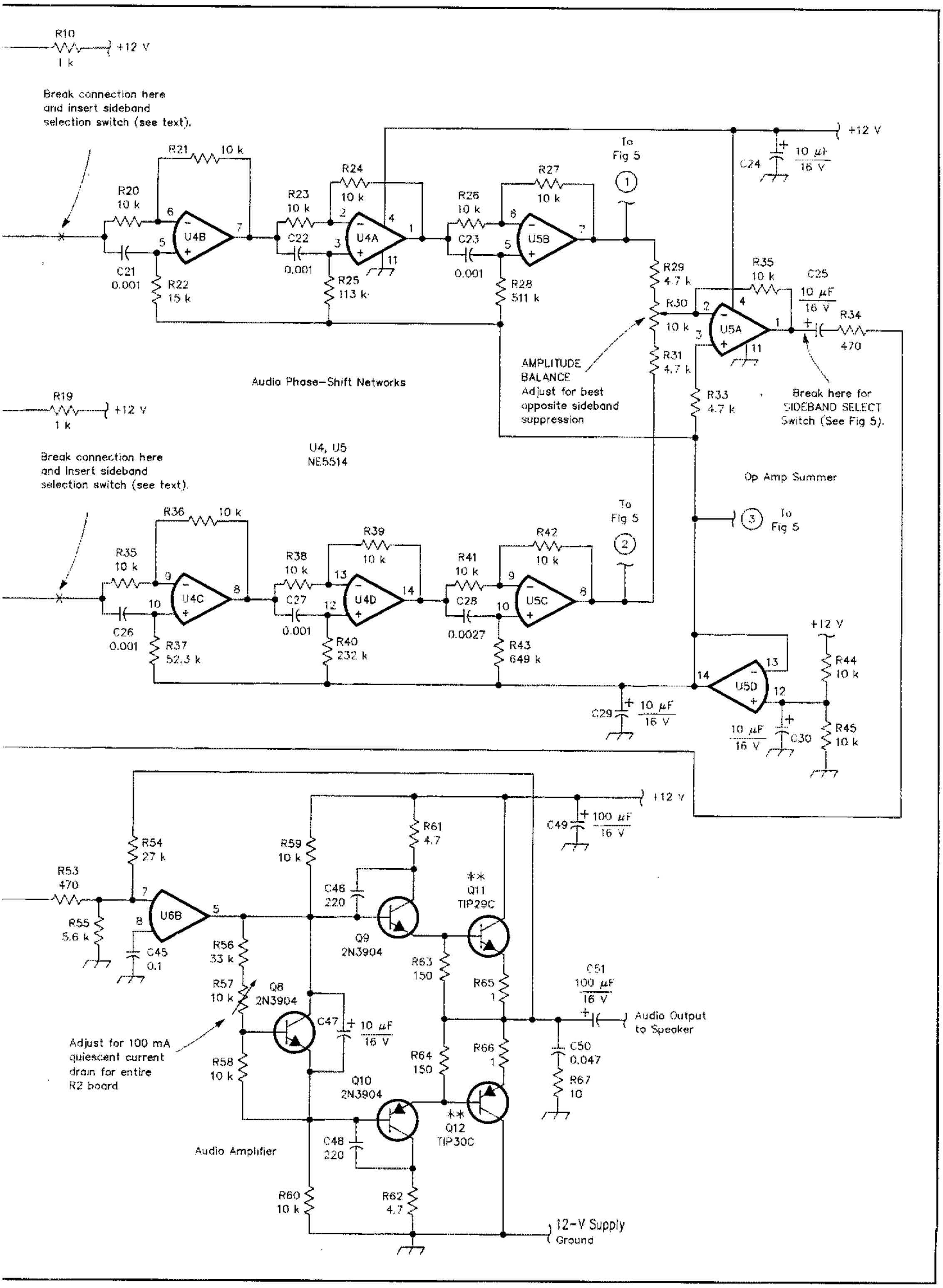
U1—Toko TK-2518 power splitter (this part is called a balun transformer in the Digi-Key catalog). See text.

U2, U3—Mini-Circuits SBL-1 double balanced mixer. Other mixers may be substituted if higher dynamic range is needed or other frequency ranges are desired.



Except as indicated, decimal values of capacitance are in microfarads (µF); others are in picofarads (pF); resistances are in ohms; k=1,000, M=1,000,000

* Matched components. See caption and text.
** Heatsink


R10
1 k
+12 V
Break connection here and insert sideband selection switch (see text).
R21 10 k
R20 10 k
U4B
C21 0.001
R22 15 k
R23 10 k
R24 10 k
U4A
C22 0.001
R25 113 k
R26 10 k
R27 10 k
U5B
C23 0.001
R28 511 k
To Fig 5
①
C24 10 µF 16 V
+12 V
R29 4.7 k
R30 10 k
R31 4.7 k
R35 10 k
U5A
C25 10 µF 16 V
R34 470
AMPLITUDE BALANCE Adjust for best opposite sideband suppression
R33 4.7 k
Break here for SIDEBAND SELECT Switch (See Fig 5).
Audio Phase-Shift Networks
R19
1 k
+12 V
U4, U5 NE5514
Break connection here and insert sideband selection switch (see text).
Op Amp Summer
③ To Fig 5
R36 10 k
R35 10 k
U4C
C26 0.001
R37 52.3 k
R38 10 k
R39 10 k
U4D
C27 0.001
R40 232 k
R41 10 k
R42 10 k
U5C
C28 0.0027
R43 649 k
To Fig 5
②
+12 V
R44 10 k
U5D
C29 10 µF 16 V
C30 10 µF 16 V
R45 10 k
+12 V
C49 100 µF 16 V
R54 27 k
R53 470
R55 5.6 k
U6B
C45 0.1
R59 10 k
R61 4.7
C46 220
Q9 2N3904
**
Q11 TIP29C
R56 33 k
R57 10 k
Q8 2N3904
R63 150
R65 1
C51 100 µF 16 V
Audio Output to Speaker
C47 10 µF 16 V
Adjust for 100 mA quiescent current drain for entire R2 board
R58 10 k
R64 150
R66 1
C50 0.047
R67 10
Q10 2N3904
**
Q12 TIP30C
Audio Amplifier
C48 220
R60 10 k
R62 4.7
12-V Supply Ground

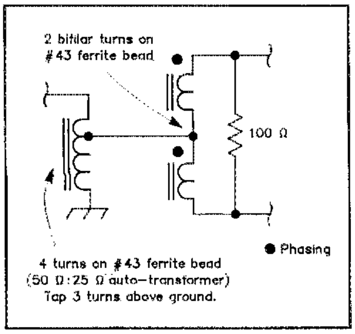


Fig 4—This home-brew in-phase splitter is an alternative to the commercial RF power splitter (U1) of Fig 3.

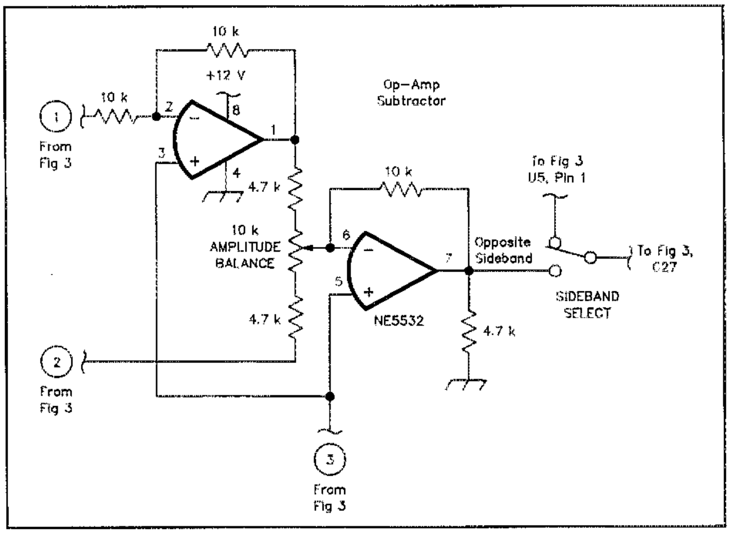


Fig 5—This subtractor circuit may be used to switch between sidebands or for simultaneous reception of both sidebands (see text).

it would be possible to have identical I and Q channels, a perfect 90° audio-phase-shift network, and thereby obtain infinite opposite-sideband suppression. Oppelt has done an excellent study of the errors introduced using real components (see Note 3). To paraphrase his many pages of analysis, it is trivial to build a receiver with 20 dB of opposite-sideband suppression, easy to get 30 dB, not too tough to obtain 40 dB, a real stretch to reach 50 dB, and quite likely impossible to hit 60 dB.

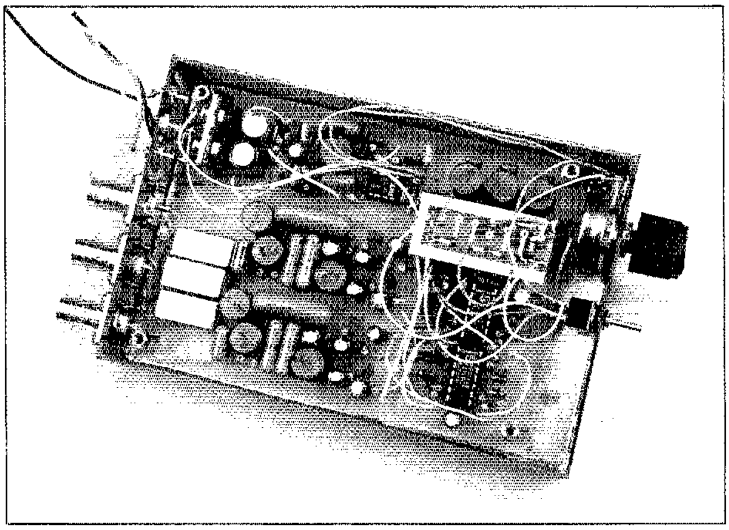
Since I wanted easy reproducibility and decent performance with off-the-shelf components, I settled on the "not-too-tough" 40 dB and designed my audio-phase-shift network accordingly. My phase-shift network is similar to the one used by Breed, but it uses the third-order coefficients calculated by Oppelt (see Note 4). In the audio phase-shift network, R20-R28, R35-R43, C21-C23 and C26-C28 must be within 1% of the schematic value. For the first prototype, I measured and selected six capacitors from a large bag obtained at a flea market, and then made up series combinations of 5% metal film resistors using a digital ohmmeter. For all the subsequent versions I have used off-the-shelf 1% resistors and capacitors (see Note 6). There is no performance difference.

The phase-shift network and combiner are capable of handling fairly large signals without distortion. The design goal was to have the preamplifier distort before the phase-shift network, since the performance of the preamplifier on the R1 board is well characterized. If distortion occurs between the RF splitter and op-amp summer, the distortion products will not have the proper phase relationship, and they will not be suppressed in the opposite sideband. The result will be that strong signals have worse opposite-sideband suppression than weak signals! This would be an unhappy circumstance indeed, but I have not encountered it at input signal levels well up into the millivolt range.

A DPDT sideband selection switch may be inserted at the points marked in Fig 3. I built one receiver that way, but I do not recommend that method of switching sidebands because the optimum setting of R30 is different for suppression of the upper or lower sideband. A better approach is to switch the 90°-shifted LO between the mixer LO ports.

The best approach is to add the op-amp subtractor circuit shown in Fig 5, with its own **AMPLITUDE BALANCE** pot. Then sideband selection is simply a matter of connecting the volume control to the appropriate sideband output with an SPDT **SIDEBAND SELECT** switch. An independent filter and audio output stage could be added if ISB reception is desired.

Many other audio-phase-shift networks may be used in place of the op-amp version



The 3.5 x 5-inch board R2 PC board layout follows the schematic of Fig 3 fairly closely. This version uses a Mini-Circuits PSC2-1 power splitter and a pair of SBL-1 mixers at the lower left-hand edge of the board. The mixers are followed by the diplexers and 50-ohm preamplifiers of the I and Q channels, followed by the audio phase-shift network (a pair of NE5514 quad op-amps) in the lower right corner. The three inductors and seven capacitors of the 1-kHz elliptical low-pass filter are arranged just above the op-amps, near the **VOLUME** control. A second filter, this one a 4-kHz Butterworth, is built on a separate PC board and glued (upside down) to the main R2 board. Filter bandwidth is switched from the front panel. The intermediate audio amplifier, 300-Hz high-pass filter, mute switch and audio power amplifier are on the top half of the board. All that is needed to make a complete receiver is a front-end filter, a VFO and 90° phase-shift network, and a speaker.

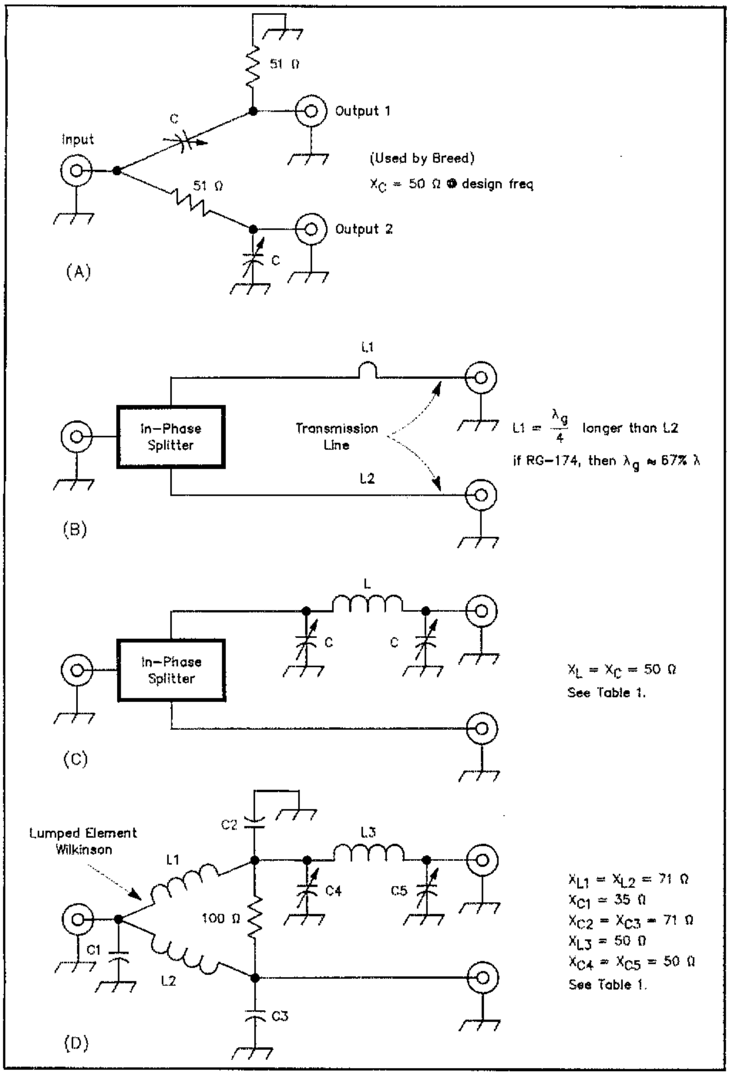


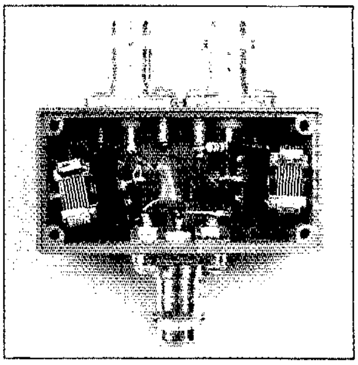
Fig 6—Four possibilities for a 90° LO phase-shift network. The simple RC network shown at A is fine for many MF/HF applications and has been used by the author on 40, 20 and 2 meters. The actual value of the capacitors depends on terminating impedances. Start with 450 pF on 7 MHz, 225 pF on 14 MHz and 22 pF on 144 MHz. See text. The circuit at B uses transmission lines and is practical at VHF and UHF. Examples: If L1 and L2 are made from RG-174, L1 is 69.1 inches longer than L2 at 28.5 MHz; 13.7 inches longer at 144.2 MHz and 8.86 inches longer at 222.1 MHz. The circuits at C and D will work better for more serious applications (see text). Component values for a variety of frequencies are shown in Table 1.

shown here. The I and Q channel signals may be digitized at the points marked X in the schematic, and all of the phase shifting, combining and filtering operations performed in DSP hardware and software. That will probably be the preferred receiver architecture for any band and any mode in the 21st century.

All of the circuitry after the op-amp summer is identical to the R1 receiver described in August '92 *QST*. Fig 3 shows component values for a 3-kHz elliptical low-pass filter. Information on other filters is given in the August '92 article. As with the R1 board, audio hiss may be objectionable during headphone listening. If you're bothered by hiss, add the RC networks described in the August '92 article to U6.

Receiver Construction

I built the first prototype using cut-up pieces of R1 boards interconnected over an unetched PC board ground plane. I followed good RF grounding techniques and hoped for the best with my audio grounds. It worked fine. On the single-board layout, I was much more careful with the audio grounds, and still didn't have any problems. The only instability I have found is audio feedback via magnetic coupling between a speaker lead (either one!) and the inductors in the diplexer stages. This is easy to cure by simply moving the speaker wires a few inches away from the inductors, but it can be a real puzzle if you don't expect it.



Close-up of the LO phase-shift network from Fig 6A. This version is built for 40 meters.

I built two more prototypes on single-sided PC boards to work out the bugs, and then produced a double-sided R2 board with plated-through holes (see Note 1). The R2 board has a number of jumpers to make it easier to try different filters and sideband-selection methods.

The R2 board is almost as easy to use as the R1 board. As explained in the August 1992 R1 article, a band-pass filter on the RF input is good practice, although a simple low-pass filter to suppress signals near the odd harmonics of the LO will suffice. On 40 meters, a narrow band-pass filter to attenuate broadcast stations may be necessary. Above 10 MHz, a preamp will be needed in quiet locations. The preamp will also help isolate the LO from the antenna. The R2 board may be used on any frequency from 1 to 500 MHz with the SBL-1 mixers shown in the schematic. Other mixers can be substituted for higher or lower frequencies.

If you do all the trig identities, follow the phase shifts through the schematic, and carefully connect everything, Murphy guarantees that you will end up with the wrong sideband. It is much easier to simply hook it up and then switch the LO connections if it's wrong. If you can convince Murphy that you really don't care, you have a 50% chance of getting it right the first time!

Local Oscillator

The R2 board with SBL-1 mixers needs two +7 dBm (nominal) LO signals, 90° out of phase. I have used all of the LO phase-shift techniques shown in Fig 6, and there are

Table 1
LO Phase-Shift Network Component Values

For the network in Fig 6C

Frequency (MHz)	*C (pF)*	*L (µH)*
3.525	903	2.26
7.05	452	1.13
14.05	227	0.566
21.05	151	0.378
28.5	112	0.279
50.1	63.5	0.159
144.2	22.1	0.0552
222.1	14.3	0.0358

For the network in Fig 6D

Frequency (MHz)	*L1, L2 (µH)*	*C1 (pF)*	*C2,C3 (pF)*	*C4,C5 (pF)*	*L3 (µH)*
3.525	3.21	1290	645	903	2.26
7.05	1.62	645	323	452	1.13
14.05	0.809	325	162	227	0.566
21.05	0.541	216	108	151	0.378
28.5	0.399	160	80	112	0.279
50.1	0.227	91	45	63.5	0.159
144.2	0.0789	31.6	15.8	22.1	0.0552
222.1	0.0512	20.5	10.2	14.3	0.0358

many others I have not tried. Since there may be a few degrees of phase error in the splitters and mixers, it is necessary to tweak the LO phase-shift network for best opposite-sideband suppression. Some phase-shift networks are easier to tweak than others—it is easier to adjust a trimmer capacitor than cut ⅛-inch pieces off RG-174 phasing lines!

The simple LO phase-shift network in Fig 6A has been used on 40, 20 and 2 meters by tacking in different capacitors and adjusting them for best opposite-sideband suppression. It works fine, but has two peculiarities:

- When adjusted for a 90° phase difference with loads connected to outputs 1 and 2, the output levels will be different.
- The phase at each output depends on the load impedance.

The LO-port impedance and conversion loss of the SBL-1 mixers are functions of drive level, so there is a complicated interaction between the adjustments of the network in Fig 6A and the **AMPLITUDE BALANCE** potentiometer (R30). In practice, you just tweak them both and forget the theory, but strange things can happen with the network in Fig 6A in the system.

Crunch Time: We Test R2 in the ARRL November Sweepstakes

"That good, huh? Maybe I'll hook it up to my 40-meter beam and give it the acid test this weekend in CW Sweepstakes." Rick Campbell had just finished telling me how well his 40-meter R2 prototype (the one in the photos) stacks up against commercial radios. Playing the skeptic, I told Rick I'd connect the transceiver to a good antenna and try to use it on a weekend when the band was sure to be full of rock-crushing signals.

Saturday afternoon rolls around. Hmm...better hook up Rick's radio and listen to a few signals. Maybe I can even work a few people with that dinky (2-inch-square) 1-W transmitter. Out come the clip leads...let's see, we need 12 V, a keyer, headphones and an antenna...wow! this thing really does put out a watt...there's Mike, W9RE, warming up for Sweepstakes...he's *loud* but there's just a trace of his signal on the other side of zero-beat...this receiver *works*.

The contest starts. I call W9RE. He comes right back and gets my entire exchange without missing a beat. Must be a fluke. Next, W2RQ...then KW8N..and so it goes—42 contacts logged in the first hour. This might be fun! During the weekend I manage to squeeze in nine hours of operation, an hour here and an hour there amid raking about a billion leaves and other domestic activities. On Sunday night, the log shows 313 QSOs in 60 ARRL sections—all on 40 meters with Rick's 1-W transceiver.

The R2 receiver is impressive. I have a very low tolerance for poor receiver performance: I'd rather spend a sunny November afternoon picking up leaves one by one than working Sweepstakes with a lousy receiver. Fortunately, R2 sounds great—no IMD or blocking problems show up. CW signals please the ear. Compared to R2, signals on a commercial transceiver with 500-Hz filters sound harsh. Some audio hiss is noticeable during headphone listening, but that can be tamed with a couple of RC networks as explained in Rick's August 1992 *QST* article.

This receiver has no AGC, which I thought would be a problem after years of listening to radios that do a great job of flattening out signals on the band. After a while, my ears and brain figured out how to deal with a receiver without AGC, and I found myself adjusting the **VOLUME** control for only the weakest and strongest signals.

Although the CW filter in this version has a rather broad 700-Hz bandwidth, I had no problem tuning the band and separating stations from each other. Even with a band full of big signals, I only heard the opposite sideband on the strongest stations. Even then, it was just "oh, there's the other sideband." It was not a bother except when I cranked up the power supply voltage from 12 to 14 (this phenomenon is explained in the text).

Bottom line: This home-brew receiver holds its own, even under demanding conditions. Some radios are fun to build but less fun to use ("it doesn't work quite right, but after all it's only a home-brew rig," or "what do you expect for the money?"). I like projects you can build and enjoy without apologies. This is one of those projects.—*Mark Wilson, AA2Z*

Anything that changes the LO drive level at the input or the impedances connected to the outputs of the network will require readjustment of both the phase-shift network capacitors and the **AMPLITUDE BALANCE** control. One operator noted greatly reduced opposite-sideband suppression when he increased the supply voltage from 12 to 14!

The network in Fig 6A is easy to build and works well within the limitations described above. For more serious applications (especially if sideband selection or operation from a variety of supply voltages is anticipated), I recommend using one of the LO phase-shift networks shown in Fig 6C, Fig 6D and Table 1.

It is difficult to recommend an LO for a receiver board that may be used anywhere from VLF to microwaves! For HF use, *The ARRL Handbook*, *QRP Classics* and *Solid State Design* contain many suitable examples. For VLF and LF, old signal generators work fine, and they have incredible bandspread. At VHF and up, the literature is a bit sparse, but an HF VFO followed by a simple transverter works well. At higher frequencies, a VXO and multiplier chain may be used.

The cost of direct digital synthesis (DDS) is falling rapidly, and many of the DDS chips offer two outputs 90° out of phase. The phase-noise performance of low-cost DDS chips is not good enough for a high-performance direct-conversion receiver, but that will change in the future. Fig 7 is the block diagram of an all-mode receiver using a DDS chip for the LO and a DSP chip for the signal processing.

Tuning

There are two trimmer potentiometers on the R2 board and a single LO phase-shift adjustment off the board. It's not quite "no tune," but it's easier than even a simple superhet! R57 sets the quiescent current of the audio-output stage. It can be set for a total R2 board current of about 100 mA—it's not critical.

The **AMPLITUDE BALANCE** pot (R30) and the LO phase-shift adjustment can be set by ear by tuning in a strong carrier on the wrong sideband and then alternately tweaking the two controls for a null.

I tune for best opposite-sideband suppression by setting a signal generator on the audio image ("wrong side of zero beat") and then alternately adjusting the phase shift and amplitude balance (R30) while watching the output on an audio-frequency level meter. I optimize the circuit at a frequency near the middle of the audio range, and then tune around to make sure the opposite-sideband suppression is good from 300 to 3000 Hz. I then adjust a step attenuator for the same signal level on the desired and undesired sideband and record the respective attenuator settings. The difference in step attenuator settings is the opposite-sideband suppression. Once the phase shift and amplitude balance are adjusted, they may be locked in place with nail polish. After a year and thousands of miles on the road, the first prototype shows no signs of needing alignment.

R2 Performance and Impressions

The R2 prototype with a 3-kHz elliptical filter was assembled and tested on the bench. There were no surprises—it was an R1 board with slightly better dynamic range, slightly better noise figure, and 41 dB of opposite-sideband suppression. The real shock came when I connected a 40-meter antenna and VFO. CW signals simply disappeared as I tuned through zero beat. Murphy had it connected for USB, and there was no way I could copy any LSB up in the phone band. It sounded too good—something was wrong with the picture.

I connected my old transceiver with its 2.4-kHz IF filter, and sure enough, CW signals were audible on the wrong sideband. I switched the antenna over to a Collins 75S-3C with its F455FB21 mechanical filter, and they were still there. After listening to the other radios, I returned to the R2. It sounded even better than before. Why?

I have been pondering this for about a year, and I've thought of several explanations. The first is that the commercial radios both have AGC, so the receiver gain is reduced on the desired sideband and increased on the undesired sideband—in other words, the AGC is trying to remove the IF selectivity from the system. I defeated the AGC on both radios and listened again. I could still hear low-frequency CW signals on the wrong sideband that were not audible on the R2. I made a few measurements and discovered that low-pitch CW signals (up to several hundred hertz) on the wrong side of zero beat are suppressed considerably less than 40 dB on both radios. The old Collins dipped below 40 dB at about 400 Hz, and the other radio went below 40 dB a little higher in frequency. The bottom line is that the R2 sounds better than I expected, because it has better selectivity than the radios I'm used to.

I demonstrated a 40-meter R2 at the QRP Hospitality Suite in Dayton this year. Several experienced HF CW operators were amazed when we could not find any CW signals that were strong enough to detect on the wrong sideband.

Comparison with a Superhet

Receiver fans will point out that the schematic in Fig 3 is as complicated as a superhet. If the image-reject direct-conversion receiver has so many parts, why not just "do it right" and build a superhet?

First, this is a high-performance direct-conversion design, with no attempt to reduce the parts count. It could be greatly simplified with only a small reduction in performance. Even in simplified form, it could still outperform most of the simple superhets I've encountered—especially those based on the ubiquitous NE602. The NE602 is not a bad part, but its limited dynamic range should confine its use to low-cost, low-current-drain, minimum-parts-count applications.

Direct-conversion receivers have a number of significant advantages over superhets. Because there is only one LO, there are no internally generated birdies. There are no image frequencies to filter out, and no spurious receiver tuning ranges at strange combinations of LO harmonics and the intermediate frequency. A low-pass front-end filter to reject signals near odd multiples of the LO frequency will ensure a spurious-free receiver.

Since direct-conversion receivers do all of their signal processing at audio, the input frequency is unrestricted. An R1 board with an SRA-3 mixer and an old signal generator for an LO works well below 25 kHz. At the other end of the spectrum, I have an R1 board

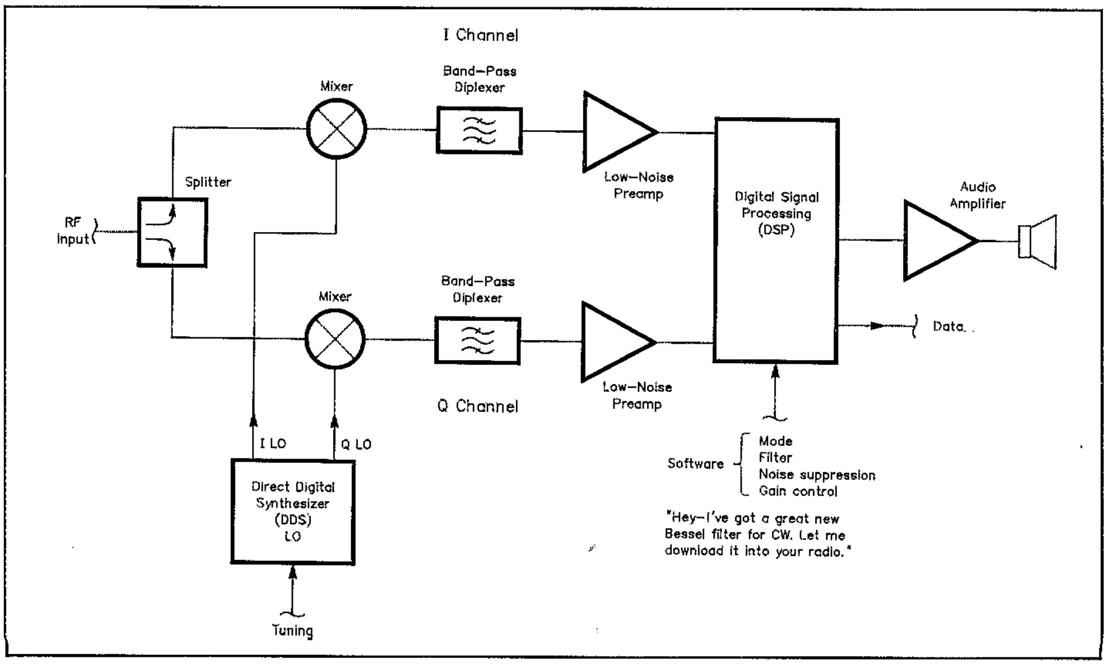


Fig 7—The R2 board can serve as a basic building block for a receiver of the future. Just add a direct-digital synthesizer for the LO and digital signal processing for the audio processing.

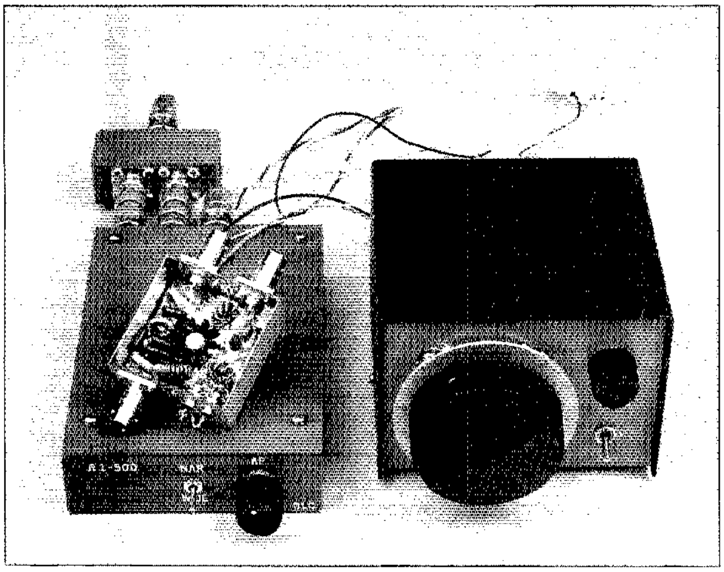


Fig 8—A simple modular 40-meter CW QRP transceiver built around an R2 board. R2 is built in the box on the lower left; the LO phase-shift network from Fig 6A is in the small box connected to rear of R2 with BNC connectors. A 7-MHz VFO is in the larger box on the right. Resting atop R2 is a 1-W CW transmitter with low-pass filter and break-in circuitry. The transmitter and VFO were borrowed from other published projects (see text). This transceiver acquitted itself well in the 1992 ARRL Sweepstakes—see the sidebar. *(photos by Kirk Kleinschmidt, NT0Z)*

with a printed rat-race mixer working at 5760 MHz.

A direct-conversion receiver can be combined with a VFO-controlled CW transmitter to build a transceiver that is simpler than a superhet transceiver. The only spurs are harmonics, which are reduced by the transmitter low-pass filter.

Direct-conversion receivers also have some disadvantages. The need for isolation between the antenna and LO makes direct conversion a poor choice for hand-held radios with integral whip antennas. At HF, a full-size outdoor antenna works best.

I do not recommend the direct-conversion, image-reject approach for a band-switched radio. It is easy to obtain good performance over a single amateur band, but the RF phase-shift network and amplitude balance must be tweaked when changing bands.

Finally, 41 dB is about the practical limit for opposite-sideband suppression in an easily reproduced analog design. While this is as good as some HF and most VHF SSB rigs, the fact remains that the basic superhet design is *capable* of better performance.

R2 Applications

The modular 40-meter QRP CW transceiver shown in Fig 8 features full break-in keying, 1 W output, a slow tuning rate, switched SSB and CW bandwidths, 93 dB two-tone, third-order IMD dynamic range, low-distortion audio, and true single-signal reception. The transmitter, VFO and break-in circuitry were borrowed from "The Ugly Weekender" and the "Optimized QRP Transceiver" in *The 1992 ARRL Handbook*.[8] The receiver is an R2 board with the RC LO phase-shift network from Fig 6A. It is a joy to operate.

The R2 board is small enough to build a high-performance portable HF CW transceiver in a smaller package than the popular commercial versions. For portable use, the receiver's audio power transistors can be removed to reduce the current drain, as discussed in August *QST*.

At Microwave Update '92, I demonstrated another interesting R2 application: a no-tune microwave transceiver, using a Down East Microwave no-tune 1296-MHz transverter with an R2 board and premixed VFO as a tunable 2-meter IF. The transmit IF uses an SSB/CW exciter board, which I'll describe in a future issue of *QST*. The use of image-reject techniques makes it possible to build single conversion microwave radios with any desired IF. I'll explore this concept more fully in a future *QST* article.

Conclusions

The image-reject, direct-conversion receiver is a viable approach to SSB and CW reception for many applications. When combined with high-performance, direct-conversion receiver techniques, it can provide basic receiver performance that surpasses many superhets. Image-reject techniques are in wide commercial use, and both Drake and Kenwood have recently introduced high-performance products that combine image-reject and superhet techniques. In the future, high-performance receivers will be built by combining a high dynamic range I and Q channel front end with a digital signal processor. Future Old Timers will sit around reminiscing about multiple-conversion superhets!

Acknowledgments

Dr Ward Helms, W7SXM, at the University of Washington, and Dr Ben Logan, WB2NBD, at Bell Labs taught me to look for unconventional solutions to problems. Without their influence I would be less inclined to pursue strange ideas like microwave transverters with no tuning and SSB receivers with no IFs.

Notes

[1] R. Campbell, "High-Performance Direct-Conversion Receivers," *QST*, Aug 1992, pp 19-28.

[2] R. Campbell, "Low Noise Receiver Analysis," in *Proceedings of Microwave Update '91*, published by the ARRL. Available from your local dealer or the ARRL Publications Catalog elsewhere in this issue (order no. 3703).

[3] R. Oppelt, "The Generation and Demodulation of SSB Signals Using the Phasing Method Part 1: Basic Theory," *VHF Communications*, vol 19, ed 2, summer 1987, pp 66-72.

[4] R. Oppelt, "The Generation and Demodulation of SSB Signals Using the Phasing Method Part 2: Signal Processing for a SSB/DSB/AM Transceiver Without Using Crystal Filters," *VHF Communications*, vol 19, ed 3, fall 1987, pp 130-140.

[5] Etched, plated and drilled PC boards (double-sided, with plated-through holes) for the R2 board are available from Applied Radio Science, PO Box 225, Houghton, MI 49931 for $20 postpaid (send an SASE for a catalog with current kit information). For individuals wishing to make their own PC boards, an etching template/part-overlay package for a single-sided version of the R2 PC board is available from the ARRL for an SASE. Address your request for the CAMPBELL R2 BOARD TEMPLATE to Technical Department Secretary, ARRL, 225 Main St, Newington, CT 06111.

[6] One source for SBL-1 mixers is Oak Hills Research, 20879 Madison St, Big Rapids, MI 49307, tel 616-796-0920). XICON 1% polyester film capacitors used in the audio phase-shift network are available from Mouser Electronics, 2401 Hwy 287 N, Mansfield, TX 76063, tel 800-346-6873, 817-483-4422, fax 817-483-0931. All other parts are available from Digi-Key, PO Box 677, Thief River Falls, MN 56701-0677, tel 800-344-4539, 218-681-6674, fax 218-681-3880.

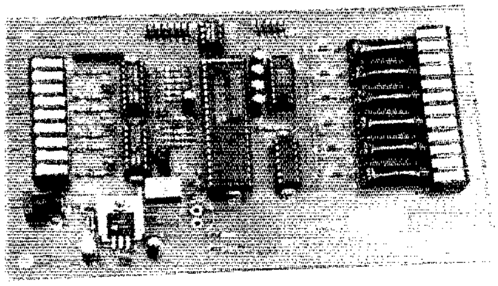
[7] G. Breed, "A New Breed of Receiver," *QST*, Jan 1988, pp 16-23.

[8] R. Lewallen, "An Optimized QRP Transceiver," *QST*, Aug 1980, pp 14-19; also see Feedback, *QST*, Nov 1980, p 53. The Optimized QRP Transceiver also appears on pages 30-37 to 30-40 of *The 1993 ARRL Handbook*, and in the second printing of *QRP Classics* (available from the ARRL Bookshelf as #3169). R. Hayward and W. Hayward, "The 'Ugly Weekender,'" *QST*, Aug 1981, pp 18-21. The Ugly Weekender also appears on pages 30-33 to 30-36 of *The 1993 ARRL Handbook*. Etched, plated and drilled PC boards for the VFO/buffer are available from FAR Circuits, 18N640 Field Ct, Dundee, IL 60118-9269. Price is $4; add $1.50 for shipping and handling to each order. Use the board's **AUX OUT** output for the correct drive level for the SBL-1 mixer.

Introducing METCON, a New Remote Control and Telemetry System

Sure you can do all sorts of telemetry and remote control via Amateur Radio—but how? METCON, Tucson Amateur Packet Radio's new standardized system for remote event measurement and control, may be your answer.

By Paul Newland, AD7I
PO Box 205
Holmdel, NJ 07733-0205



Amateur packet radio has been around for more than 10 years. Hams first used packet for keyboard-to-keyboard communications. Later, computer BBSs became common and allowed hams to exchange electronic mail and bulletins via packet. This article outlines another use for amateur packet radio: telemetry and control of remote environments. Specifically, it discusses a new telemetry and control system for amateur packet radio developed by Tucson Amateur Packet Radio (TAPR). The system is called *METCON-1*,[1] or simply *METCON*.

METCON is a simple device you can use for monitoring and controlling electronic devices at remote locations. It derives its name from the words *teleMETry* and *CONtrol*. Although TAPR designed METCON to be "talked to" via packet radio, you can also use AMTOR or a dial-up telephone line and modems for the communication link.

METCON Basics

Fig 1 shows a basic METCON system configuration. In this example, a computer terminal communicates with METCON via a packet-radio link. You, the user, enter command lines at the terminal, and METCON responds to each command. With METCON configured as a remote device controller, for instance, you might tell METCON to open or close a particular relay. Alternatively, you might ask METCON to report the voltage, temperature, frequency or switch position at one of its inputs. METCON doesn't have to wait for your command; you can program it to report data automatically—at 15-minute intervals, for example—or whenever it senses a change in any of the values you've configured it to sense or measure.

Uses for METCON are limited only by your imagination. Repeater control is a likely candidate. Because Amateur Radio's loss of spectrum from 220 to 222 MHz included popular repeater control-link frequencies, alternative spectrum for primary repeater control is scarce. Using METCON for primary control can allow many repeaters to share the same control channel. Additionally, thanks to METCON's password and authentication features, it's difficult for intruders to break electronically into the system and set things awry.

METCON can let you monitor a remote animal burrow or den in the wild, or keep watch on the temperature and furnace status of a vacation home. The sky's the limit when it comes to applications. You may even dream up applications that aren't permitted on Amateur Radio frequencies. (For example, your application may be commercial, or require security or reliability a radio link cannot afford.) This needn't stop you cold, however. If Amateur Radio isn't an appropriate communication link, consider alternatives. Hayes-compatible modems and a telephone line might be an easy answer. If you need to control something on the other side of the world, consider using the international TELEX system to link you with your remote METCON system.[2] If you must use radio, but your application is incompatible with the Amateur Radio rules, consider a system that makes use of business-band frequencies with appropriate FCC licenses and type-accepted equipment.

METCON itself doesn't know or care how you communicate with it; getting that link up and running properly is *your* job. Now, let's look at *METCON*'s job. Specifically, we'll examine METCON's output and input circuits to understand what they can and cannot do.

METCON Output Circuits

Fig 2A shows one of METCON's eight relay-contact output circuits (the title photo shows a prototype with only six outputs). You can hook up almost anything you want to these relay contacts because they're isolated from other METCON circuitry. The voltage applied to the contacts should not be more than 24 volts above or below METCON's ground,[3] and the current flowing through the contacts should not exceed 250 milliamperes.

METCON Input Circuits

Fig 2B shows a standard METCON input, which is configured as a current loop.

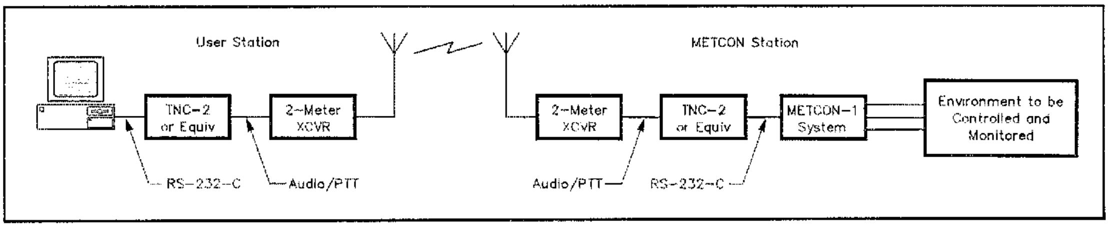


Fig 1—Tucson Amateur Packet Radio's METCON-1 system allows you to measure and control distant events via a half- or full-duplex communication channel—in this example, radio. METCON requires 12-volt dc power.

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[1] Notes appear on page 48.

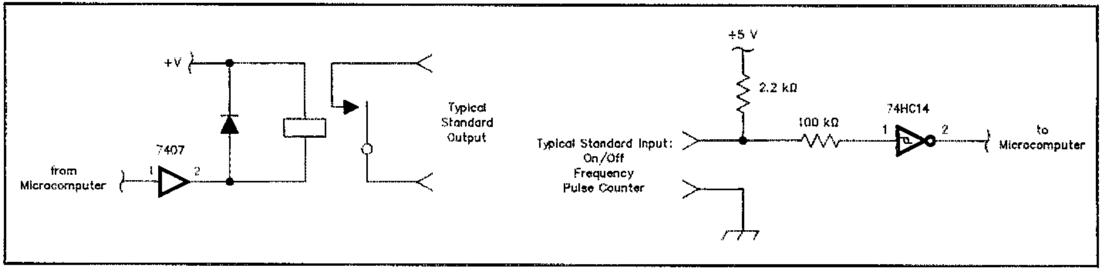


Fig 2—A typical METCON standard output consists of normally open relay contacts capable of handling 24 volts at 250 milliamperes. A typical METCON standard input exhibits an open-circuit voltage of +5 and sources 2.3 milliamperes when short-circuited. METCON's standard inputs can be configured to sense status (open or short circuit), count frequency or totalize pulses.

METCON sources current through a 2.2-kΩ resistor connected to a 5-volt (relative to METCON ground) source. This current returns via an "input common" connection. METCON input devices should be connected to return the source current via this common path. (*Common* doesn't mean the same thing as *ground* in this case. For best safety and operating performance, you should isolate any circuit that you connect to METCON's input from ground and any other circuits.) When a high-impedance circuit (more than 22 kΩ) is connected to a METCON input, METCON considers that input to be active (logic 1).

You'll probably use one or more switches as METCON input devices. A magnetic reed switch, for example, can be attached to a repeater-site door and connected to METCON, with a magnet installed on the door frame to keep the switch closed as long as the door is closed. Current flows; the input is inactive. When the door opens, the magnet moves away from the reed and the switch opens (current does *not* flow; the input is active). It sounds simple because it *is* simple.

But METCON can do much more than report whether any of its inputs are active or inactive. It can also report the number of current on-to-off transitions (up to 10,000) that occur per second, or the *total* number of transitions that have occurred at one of its standard inputs since the system was last powered up. This feature allows any METCON standard input to act as a frequency or totalizing pulse counter. (We'll examine this feature more closely in a bit.)

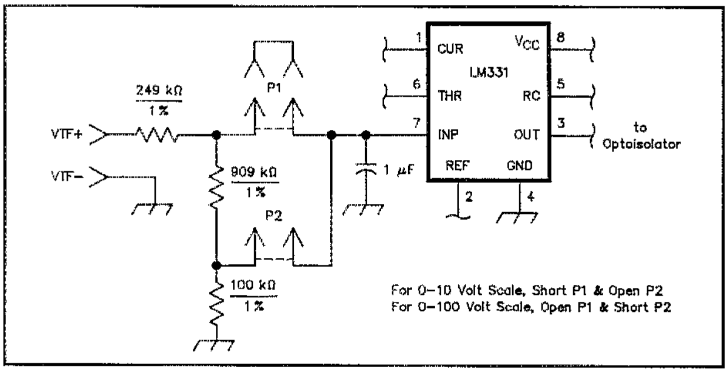
Special Input Modules

As I've described it so far, METCON is designed for basic input (switch open or closed) and basic output (relay open or closed). Thanks to optional input modules and "hooks" built into the METCON system, however, more sophisticated inputs are also possible.

Voltage- or Temperature-to-Frequency Module

METCON's ability to measure frequency may not seem useful with a switch installed as an input device. (A repeater-site door isn't likely to be opened several tens or hundreds of times per second, for instance!) But METCON's frequency-measuring ability affords many possibilities when combined with an input device that converts voltage to frequency. METCON's optional voltage-to-frequency (VTF) converter module does just that.

The VTF module uses National Semiconductor's LM331 VTF chip. Fig 3 shows the module's input circuit. You can configure it for full-scale ranges of 0-1, 0-10 or 0-100 volts. At full scale, the module produces a 10-kHz pulse train; lower input voltages produce proportionally lower output frequencies.



For 0–10 Volt Scale, Short P1 & Open P2
For 0–100 Volt Scale, Open P1 & Short P2

Fig 3—METCON's voltage-to-frequency (VTF) module uses a National Semiconductor LM331 IC for VTF conversion and jumper-programmable resistors for range selection. Ranges of 0-1, 0-10 and 0-100 volts are possible with the standard resistor values shown for this typical VTF-module input. Other ranges can be set by installing resistors of different values.

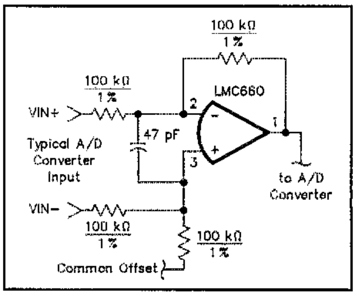


Fig 4—The METCON system can also use an optional eight-bit, successive-approximation analog-to-digital converter (ADC) module capable of sampling up to eight inputs. This drawing typifies the circuitry of one of the ADC's eight inputs.

The voltage to be measured must be positive with respect to the VTF module's ground. For most applications, this isn't a problem.

You can place a METCON VTF module physically close to the voltage source you want to measure. This helps avoid the false readings that noise may cause. Although the module can (of course) accept voltage input from an external source, it can also accept input from an on-board temperature-to-voltage transducer. Jumpers in the VTF module allow you to connect this sensor to the module's voltage-to-frequency converter IC, with appropriate range scaling. This configures the module to perform temperature-to-frequency (TTF) conversion. Depending on the type of sensor you use, you can measure temperature in degrees Fahrenheit or Celsius.

The LM331 VTF IC drives an optoisolator. Connected to one of METCON's standard inputs, the optoisolator's output transistor opens and closes the input circuit at a frequency determined by the LM331. Optoisolation protects the METCON input involved from the effects of external grounds and possible ground loops (provided that the VTF power source is isolated from the METCON power source).

Analog to Digital Converter Module

If you need to measure only one or two analog signals, using a couple of VTF modules is probably the best way to go. But what if you want to measure *eight* analog inputs? You'd

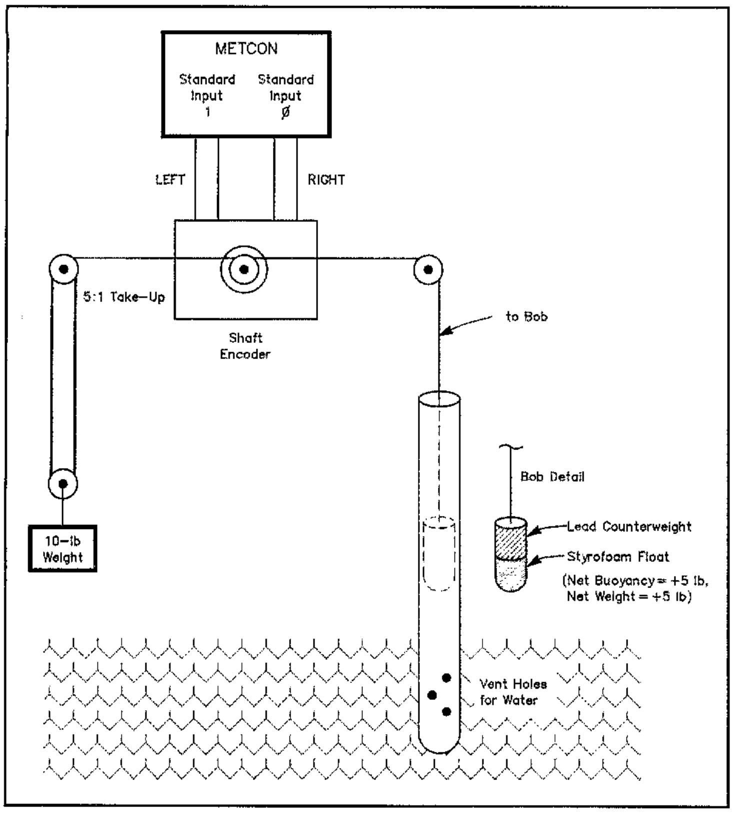


Fig 5—How METCON might be used to remotely measure water level by means of a bob floating in a perforated tube. The shaft encoder generates a uniform number of pulses per unit of bob-cord travel. METCON detects and reports the number of **RIGHT** and **LEFT** pulses the encoder produces; the METCON user crunches these numbers and compares them to a predetermined low-water pulse value to determine the actual water level.

need eight VTF modules, and that can run into some bucks. There *is* an alternative.

METCON includes software and hardware hooks for an optional analog-to-digital converter (ADC) module. The ADC module is built around a National Semiconductor ADC0838 successive-approximation ADC IC. METCON includes a five-pin connector that the ADC module plugs into. At least once each second, METCON samples all of the ADC's inputs and stores the results in memory. You can ask METCON to report the value of any of the module's inputs by reading the corresponding memory location.

As Fig 4 shows, each of the ADC module's inputs operates differentially. The input differential voltage to the ADC module can vary from 0 to 2.55 (the default configuration) or from −1.28 to +1.27 volts. Jumpers on the ADC module select the input range. You can change the scale of any ADC input by changing the values of resistors associated with that input's op amp.

Although the ADC module provides an economic way to measure many different voltage sources, it has limitations. First, fixed resistors determine each input's voltage scale. If you need a voltage scale other than the standard two, you'll need to change resistors. Second, the ADC measures to a precision of 8 bits. If you want to measure small changes in voltage over a period of time (and are not too interested in absolute accuracy), a VTF module is probably a better way to go. Finally, it's sometimes difficult to keep hum and other noise from getting into your ADC measurements. Lack of attention to ground loops usually causes this problem.

Elapsed Time Pulser Module

METCON's optional Elapsed Time Pulser (ETP) module serves as another means of expanding the use of the system's standard inputs. The ETP module generates a pulse for every unit of time its input is active—anywhere from one pulse per second to one pulse every 2 minutes. Connected to a METCON standard input configured as a totalizing pulse counter, the ETP module functions like an engine-hour meter. You can use an ETP module to determine how much time something has spent in a particular state.

Putting METCON to Work

Now that we've discussed METCON basics, let's examine some possible METCON applications. We'll look at three that span the range from simple to complex.

Water-Level Indicator

One possible application for METCON is monitoring the level of some body of water. This might be a bucket, barrel, lake or river. Fig 5 shows an example of a water-level sensor. A weighted float (bob) is placed inside a plastic tube. Vent holes perforate the tube below the minimum expected water level to equalize the water levels inside and outside the tube. The bob drives a counterweight through a reduction mechanism that moves the counterweight one inch for every five inches of water movement.

The key to measuring the bob's relative vertical height is a *shaft encoder*—a device that generates a fixed number of pulses per revolution of a shaft. The bob's vertical movement must be converted into rotary motion to drive the shaft encoder. One way of doing this would be to wind the bob suspension cord (perhaps several times) around a pulley. A disk, connected to the pulley and perforated with an appropriate hole pattern, allows LEDs, shining through the holes onto photodiodes or phototransistors, to signal the disk's direction and number of rotations. As the bob cord moves left or right through the shaft encoder, the disk rotates counterclockwise or clockwise, respectively. For every 10 degrees the disk rotates counterclockwise, the **LEFT** output provides one contact closure (the equivalent of an open switch closing for 50 milliseconds and then returning to open). For every 10 degrees the disk rotates clockwise, the **RIGHT** output provides one contact closure. We'll also assume that the shaft encoder provides 10 output pulses for every inch the bob cord moves.

With such an arrangement, if the water level (and hence the bob) drops 2 inches, the cord moves through the shaft encoder 2 inches to the left. Over the course of this 2-inch movement, the shaft provides 20 contact closures. For a water-level rise of 3 inches, the shaft encoder provides 30 closures.

Recall that a METCON standard input can determine one of three things: static value, frequency (counts per second) or total pulse count. For this application, we'll use METCON's totalizing feature to determine the bob's position.

Initializing the system comes first. To do this, pull the bob cord to the left to position the counterweight at its minimum-level position. Next, zero the totalizing pulse counters for Standard Inputs 0 and 1. Slowly release the bob cord just enough to rest the bob on the water surface without the cord going slack. As you do so, the shaft encoder creates a series of pulses on the **RIGHT** output. The number of pulses equates to 10 times the number of inches the shaft moved through the encoder. For this example, let's assume that there were 350 pulses (35 inches). If, at this point, we command METCON to read the counter for Standard Input 0, METCON will respond with 00350.

Now that we've initialized the system, we can measure water-level changes. A 4-inch drop in water level causes the shaft encoder to generate 40 pulses at its **LEFT** output. METCON counts these pulses and stores them in the counter for Standard Input 1. At this point, we have a count of 350 on Input 0 and a count of 40 on input 1. To determine the water level, we subtract the value in Counter 1 from the value in Counter 0. In this example, the result is 310 (310 = 350 − 40), or 31.0 inches. METCON reports only reports the value in the counters; you (or your computer) must do the subtraction.

If the water now rises 6 inches, the shaft encoder generates 60 pulses at its **RIGHT** output, increasing the count in the Channel 0 counter from 350 to 410. Again, the water level should be equal to the value in Counter 0, less the value in Counter 1, divided by 10. The result is 370 (370 = 410 − 40), or 37.0 inches.

Measuring water level with a system like this involves several potential problems. First, this system uses moving parts. Moving parts wear, so a system *without* moving parts would be more reliable over time. Second, if METCON suffers a power failure, the information corresponding to the shaft encoder's 0 position is lost. Third, the bob's buoyancy may change with time. Fourth, foreign matter may jam the bob within the tube. (I'm sure that many other faults can be found in this hypothetical system. I offer this particular example only as a means of describing METCON's features.)

The need to subtract the two counter values points out a limitation of METCON itself. For telemetry, METCON just *accumulates* signals; METCON does not *process* signals.

Moored-Boat Monitor

A friend whose hobbies include sailing and Amateur Radio suggested to me that METCON might also be used to monitor a moored sailboat. Fig 6 illustrates some of the ideas we discussed. I present it as an example of METCON's capabilities rather than as a suggestion of the best way to monitor a boat.

My friend's boat is moored on a buoy in a bay. Its bilge pump has float and manual switches. If the bilge water rises high enough, the float turns on the pump to remove the water. Eventually, the pump lowers the water to a level that opens the float switch and turns the pump off. This cycle repeats to keep the bilge water to a safe level.

The problem with this arrangement is that sometimes the pump clogs, overheats and shuts down (it has a thermal cutoff switch). My friend can't tell that this has happened until he checks the boat on a weekend. More than once, he has found the bilge full of water because the pump clogged. He'd like to determine the pump's status, and the bilge water level, from his ham shack at home. Additionally, he'd like to know if the boat has moved "off station"—more than a normal radius from its anchor buoy.

I proposed that we wire METCON to several key points on the boat. First, we'd connect METCON's Standard Input 4 to a relay in parallel with the bilge pump. This would let METCON tell if voltage was applied to the pump (that is, that the pump should be pumping). Additionally, we'd configure Input 4 as a totalizing pulse counter so METCON could report *how many times* the pump had been energized.

We'd also connect an ETP module, connected to METCON's Standard Input 3, in parallel with the pump. Configured to provide an output pulse for every 60 seconds its input is high (that is, when power is applied to the pump), this module would allow us to determine how many minutes the pump had been running.

Just because the bilge pump is energized doesn't mean it's working. An additional float switch, placed about 3 inches above the bilge pump's regular float switch, would provide safety backup in case the bilge pump failed or couldn't keep up with water inflow. If this switch closes, the boat may be in danger of sinking! We'd connect this switch to Standard Input 2, configuring Input 2 so that any change of state would immediately send a message to the remote monitoring station.[4]

The boat's radionavigation system has an off-station alarm that METCON can monitor. The alarm closes a pair of relay contacts if the radionavigation system detects that the boat has moved more than a preset number of meters away from a programmable geographical position. METCON Standard Input 1 would monitor this alarm's status. We'd also configure Input 1 to immediately report any status change.

Finally, we'd attach a magnetic reed switch to the boat's cabin door and connect it to METCON's Standard Input 0. METCON would therefore detect removal or opening of the door. Like Inputs 1 and 2, a change in status on this input would immedi-

Communicating with METCON

So, just exactly how do you communicate with METCON? Here's an example that outlines only the simplest METCON commands.

You can connect with METCON over just about any channel—radio, fiber or wire—that allows half- or full-duplex communication. For this example, we'll use VHF packet radio. We'll log in, check the system clock, read a switch input, read a frequency counter location, set and clear a relay, and then look at METCON's memory display. Finally, we'll log out and disconnect.

To start the process, we need to connect to the packet station attached to METCON. Follow the instructions in your packet TNC for the connect stuff. All METCON commands begin with =*A* and all METCON responses begin with -*A*-. (You can change the letter from *A* to some other value, but we'll assume the default value here.)

METCON's login command is **=ALI*nnnnnn***, where *nnnnnn* is the authentication sequence. (METCON provides security by requiring users to transmit an authentication sequence—in effect, a password—at log-in time.) If the login is successful, METCON responds with **-A- LOGGED IN**. We can then check the clock with the time command **=AT**, and METCON might respond with **-A- THU/920326/2025/51**, which shows day of the week, hours, minutes and seconds, in that order. Now to writing to and reading the I/O ports.

You don't really read or write to the METCON's microcomputer I/O ports directly. Instead, you read or write port "images" that are contained in the microcomputer's memory. At least once a second, the port images are updated: The port inputs are moved to the input memory image and the output memory images are moved to the output ports. The commands we'll use read from or write to specific memory locations; METCON updates its actual I/O ports for us.

METCON stores its main input port image at memory location 09. To read a bit of this input port image, we can use the bit display command. If we want to display the value of bit 2 of the input port memory image, this command would be **=ARB092**. METCON responds with the value in bit 2 of memory location 09 (**-A-092-1**). Reading a 16-bit counter image at memory location 1B would require the METCON command **=ARW1B**, to which a possible response might be **-A- 1A-01222 04C6 0000 0100 1100 0110**. When you read a byte or word value, METCON responds with the value in decimal, hex and binary numbers, in that order. Thus, *01222* is the decimal value of the 16-bit word, *04C6* is the equivalent hex value, and the rest is the binary value—with spaces every four bits to improve readability.

METCON stores its main output port image at memory location 05. To activate Relay 3, we can use METCON's bit write command to write a *1* to bit 3 at address 05. That command would be **=AWB0531**, and METCON would respond with "**-A- 053-1**". To deactivate Relay 3, we again use METCON's bit write command to write a *0*. This command would be **=AWB0530**, to which METCON would respond with **-A- 053-0**.

So far, we've been looking at only one bit, byte or word. Another METCON command is of interest to those who want the whole story. If you give METCON a memory display command (**=AM**), it responds with a formatted memory map.

Having completed our tasks, we can now log out. METCON's logout command is **=LO**, to which METCON responds with **-A- LOGGED OUT**. We can then disconnect from the METCON station using the usual packet disconnect routine.

Fig A shows what we just did would look like onscreen, from start to finish.
—*AD7I*

ately send a message to the remote station to indicate the change. This arrangement would serve as an intrusion alarm.

Repeater-Site Monitoring

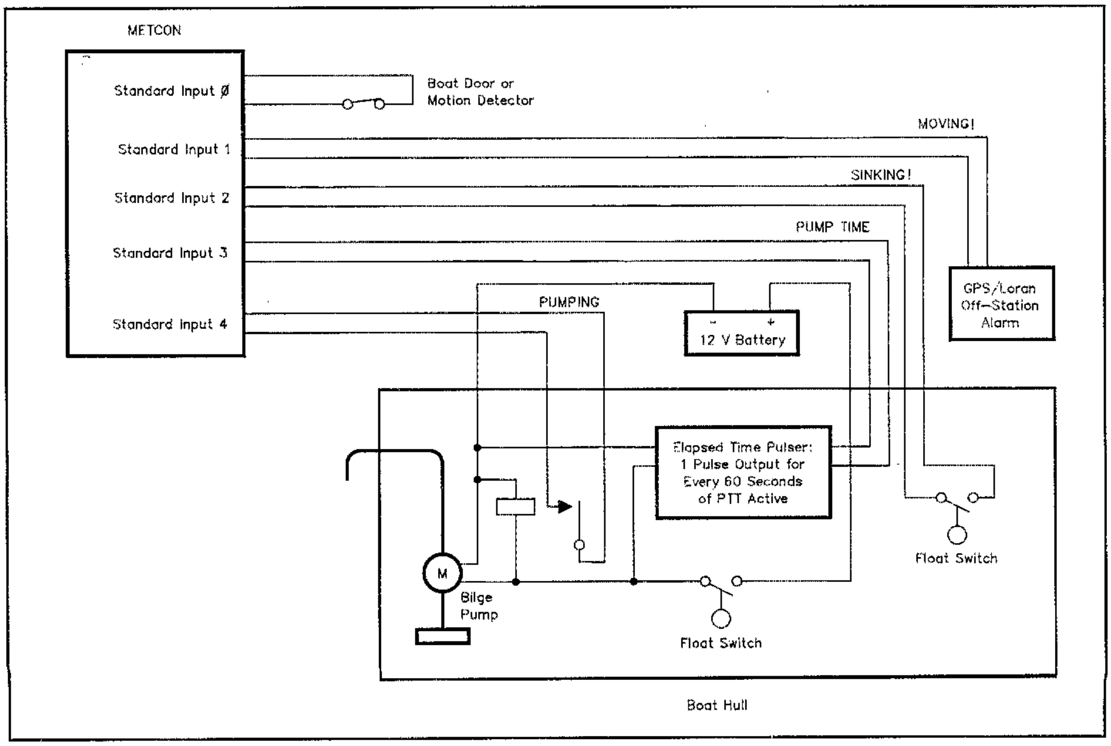
All of the example applications I've presented so far have used METCON only for telemetry. As a final METCON application example, I'll outline a repeater monitoring and control system (Fig 7). This might be a shared site, where a ham repeater shares space with other ham and/or commercial equipment. Again, this example is for illustrative purposes only.

A magnetic reed switch on the building door provides basic intruder detection. (If there's more than one door, each door must have a switch, and all switches must be wired

```
cmd:c ad7i-8
*** CONNECTED to AD7I-8
=ali000000
A- LOGGED IN
=at
A- THU/920326/2025/51
=arb092
A- 092-1
=arw1b
A- 1B-01222  04C6 000 0100 1100 0110
=awb0531
A- 053-1
=awb0530
A- 053-0
=am
A- MEMORY DISPLAY

    P6          P5          P4          P2          P0
0000 0000  0000 0000  0000 0000  0000 0100  0000 0000   OUT  (08 07 06 05 0
0000 0000  0000 0000  0000 0000  0000 0100  1110 1111   IN  (0D 0C 0B 0A 0
0000 0000  0000 0000  0000 0000  0000 1100  1111 1111   CHG (12 11 10 0F 0

  7      6       5      4      3      2      1      0
00000  00000  00000  01260  00000  00000  00000  00000  F/P CTRS (21/22-13

 7     6     5     4     3     2     1     0
255   255   255   255   255   255   255   255   ADC (2A-23)

A- MEMORY DISPLAY COMPLETE
=alo
A- LOGGED OUT
```

Fig A—A sample METCON communication session. The text describes the action line by line.

Fig 6—This hypothetical METCON moored-sailboat monitoring system can detect whether the boat's cabin door is open (switch opening on the **DOOR** line); whether the boat's bilge pump is energized (relay closure on the **PUMPING** line); how many minutes the bilge pump has been energized (number of pulses produced on the **PUMP TIME** line by a METCON Elapsed Time Pulser module); whether the boat's bilge water level is abnormally high (float-switch closure on the **SINKING!** line); and whether the boat has moved abnormally far from its mooring buoy (relay closure on the **MOVING!** line by the boat's radionavigation system off-station alarm).



in series.) This switch drives METCON's Standard Input 0.

If the repeater site is shared, many people may enter and leave the building without our prior knowledge. Only designated people should enter our repeater cabinet, however, so METCON Standard Input 1, configured to immediately report status changes, accepts input from a switch on the repeater cabinet door.

Input 0 might be configured as a totalizing pulse counter so we could determine how many times the building door was opened. If we have exclusive access to our repeater building, we should also configure Input 1 to immediately signal status changes. This would alert us to someone opening our repeater cabinet door as soon as it occurred.

METCON's Standard Output 0 provides the first order of repeater control. It does so by enabling the transmitter's PTT line. If Output 0 is off, the PTT line is open, so the transmitter does not come on even if the receiver hears a signal. An ETP module monitors the transmitter's PTT input and drives METCON's Standard Input 3 so that we can tell how many total minutes the transmitter has been active.

Two other standard METCON outputs control receiver settings. When Output 1 (**PL REQUIRED**) is active, the receiver must detect the proper subaudible tone before the repeater's carrier-operated relay (COR) or signal-detection circuit is activated. With Output 1 active, the repeater retransmits only signals that include a subaudible tone of the correct frequency.

METCON Standard Output 2 controls the COR's sensitivity via the **TIGHT SQUELCH** line. When Output 2 is inactive (open), the COR sensitivity is normal. When Output 2 is active (closed), the COR sensitivity decreases by 10 dB. (This can be useful during periods in which undesired weak signals from adjacent service areas would otherwise bring up the repeater.)

METCON also measures two analog signals in this application. One is the float voltage of the battery back-up system, and the other is the transmitter heat-sink temperature. VTF modules handle both tasks.

A VTF module configured to measure temperature rather than voltage monitors the transmitter temperature. The module's temperature sensor, which looks like a TO-92 transistor, is normally mounted on the VTF module's PC board. In this application, however, we'd remove it from the module and place it in thermal contact with the transmitter heat sink. A short three-wire cable would connect the sensor to the VTF module. METCON Standard Input 2, configured as a frequency counter, accepts input from this module.

A second VTF module, configured to measure dc voltages from 0 to 100, monitors the battery float voltage. METCON's Standard Input 5, configured as a frequency counter, accepts input from this module.

Now It's *Your* Turn

This article has introduced you to Tucson

METCON's Authentication Procedure

METCON makes life hard for would-be system crashers by requiring *authentication* on the part of loggers-in. Logging in using the proper authentication procedure is probably the most complex operating challenge that METCON offers. Here's how it works.

All authorized users of a METCON system must possess the necessary *authentication table* to gain system access. Any user without the table will have extreme difficulty logging in. Because the table itself is never transmitted over the air, would-be intruders listening in on METCON-1 communications can learn little that compromises system security.

Generating the Table

You don't have to know how METCON generates the authentication table to be able to log in, but it's interesting, so I'll outline it here. Again, one important consideration about the authentication table is that it should *NEVER* be sent over the radio channel. The System Administrator (SA) is *physically present* at the METCON site when the system generates the authentication table.

The SA directly connects a computer or terminal to METCON through a null-modem cable and installs an option strap that tells METCON that the SA is connected. With the SA option strap in place, METCON opens itself up and no longer rejects reserved commands. The SA can now issue *any* reserved METCON command and METCON will comply. The SA then loads a six-character "seed" into METCON. METCON uses this information to initiate a random-number generator. The SA then commands to METCON to output an authentication table. The SA writes this table to disk for printing onto paper for distribution to all authorized system users. The SA then removes the hardware option strap, thus disabling all reserved METCON commands. Without this strap installed, the seed cannot be changed or read, and the authentication table cannot be generated. The authentication table itself looks something like this:

```
SEED:  FF 1234
   00    01  02  03  04  05  06  07  08  09  0A  0B  0C  0D  0E  0F
00:F4    C9  78  AA  C8  78  29  29  C6  78  29  F4  58  7F  C8  F4
10:C9    F4  48  78  95  AA  58  7F  9A  95  AA  8A  68  EC  F8  C9
20:88    4A  88  49  DA  DA  F8  AA  4A  AA  1F  14  FD  98  49  98
30:7F    C8  68  14  B8  68  7C  58  58  7F  CB  68  88  4A  88  49
40:2A    68  98  49  58  7F  CB  68  88  4A  88  49  49  98  4A  AA
50:7F    C8  78  98  7C  98  68  58  CB  68  B8  6B  88  68  58  7F
60:74    F1  F8  C6  C8  C8  C8  66  66  81  74  76  4A  AA  00  68
70:6B    98  49  98  7F  CB  68  2A  4A  88  49  58  98  4A  AA  88
80:D9    B5  DA  D9  58  D4  5F  DA  A3  CA  C8  6F  DA  A3  DA  DA
90:DA    D9  B5  DA  E8  58  D4  7C  DA  43  FB  4F  FB  4F  68  CA
A0:98    68  AA  1F  F4  D3  1F  F8  68  AA  88  68  D2  1F  F8  98
B0:D3    1F  F8  F4  AA  88  68  F4  7B  DA  D9  B4  0F  AA  58  D4
C0:CA    F4  D3  AA  D0  AA  34  D2  34  FD  CA  F4  DA  D9  CE  AA
D0:D4    DA  D9  D0  7E  AA  DA  D9  1F  79  DA  DA  14  1F  F8  AA
E0:34    AE  CA  99  E1  CA  99  AA  AA  1F  14  34  1F  50  08  CF
F0:68    AA  6D  AA  18  C4  6D  A4  D7  6D  A4  5D  7D  A4  4B  08
```

Using the Table to Log In

Okay, back to our example. The user would connect to the packet station at the METCON site.

Next, the user commands METCON to respond with a challenge by sending **=ALC**. (As discussed in the "Communicating with METCON" sidebar, all METCON commands begin with **=A**, and all METCON responses begin with **-A-**.) METCON responds with a random challenge (**-A- CHAL: A1 B8 8F**).

The authorized operator applies each of these values to the authentication table as *YX coordinates* to determine the correct response to each challenge value. Both axes' labels run from 0 to F (ignore the trailing *0* in the Y-axis labels and the leading *0* in the X-axis labels). For our example of a challenge of A1, B8 and 8F, the table value for a challenge (*index*) of A1 is 68; for an index of B8, 7B; and for an index of 8F, DA. In this example, then, the operator logs in with values of 68, 7B and DA via the command **=AL687BDA**. A login example is shown below (commands on the left, comments on the right).

```
    =alc                              ; Ask station A to issue LOGIN
                                      ; challenge.
    A- CHAL: DC A3 2A                 ; METCON provides a random
                                      ; challenge. Use each
                                      ; value as index into authen table.
                                      ; Respond
                                      ; with value of each entry for LOGIN
                                      ; command.
    =al141f2a                         ; Log in with value 141f2a (last byte
                                      ; wrong).
    A- !!! ERROR, USER MUST LOGIN!!!  ; METCON error message.
```

```
A- CHAL: A1 B8 8F          ; New challenge. Look up new data
                           ; in table.
=al687BDA                  ; Log in with value 687bda (correct
                           ; lookup).
A- LOGGED IN               ; METCON says user is logged in.
```

Note that the first attempt to log in failed because the last byte of the login command was incorrect. *Every incorrect login command causes METCON to issue a new random challenge*—to which the logging in station must respond with correct *new* table values.

There are three things to keep in mind about METCON authentication. First, the table's contents depends on the seed value the system administrator loads into METCON. Different seed values result in different table values. *Be sure to keep both the seed and table contents private;* anyone knowing the seed value you chose can use another METCON system to generate an authentication table exactly the same as the one *you* generated!

Second, METCON randomly chooses challenge values for each login attempt. If you send an incorrect login command, you'll receive a **USER MUST LOGIN** message and new challenge values. This means that you'll need to look up the corresponding values from your authentication table for every challenge offered. If you're unable to log in after three tries, METCON disconnects the packet link. METCON does this to slow down intruders who might try all combinations of login values.

METCON's authentication routine may not be perfect, but it raises the bar for unauthorized operators who might try to break into your system.—*AD7I*

Amateur Packet Radio's METCON-1, a simple telemetry and control system that allows its user to monitor and control an environment at a remote location. METCON can operate over a shared packet radio channel, a telephone modem channel, or almost any other communication channel that permits half- or full-duplex data transmission. Its possible applications are limited only by your imagination. I invite you to apply your ingenuity to making METCON work for you.

Notes

[1] Information and kits for METCON and METCON modules can be obtained from Tucson Amateur Packet Radio (TAPR), PO Box 12925, Tucson, AZ 85732, tel 602-749-9479, fax 602-749-5636.

[2] If you use the TELEX network, an ASCII/Baudot converter would be needed. An 8751 microcomputer running an appropriate program could easily provide this function.

[3] For best safety, METCON's ground should always be at earth potential.

[4] Note that this only works when the remote station is connected to METCON. METCON does *not* have the ability to establish a connection on its own. The user must connect to METCON; METCON cannot connect to the user.

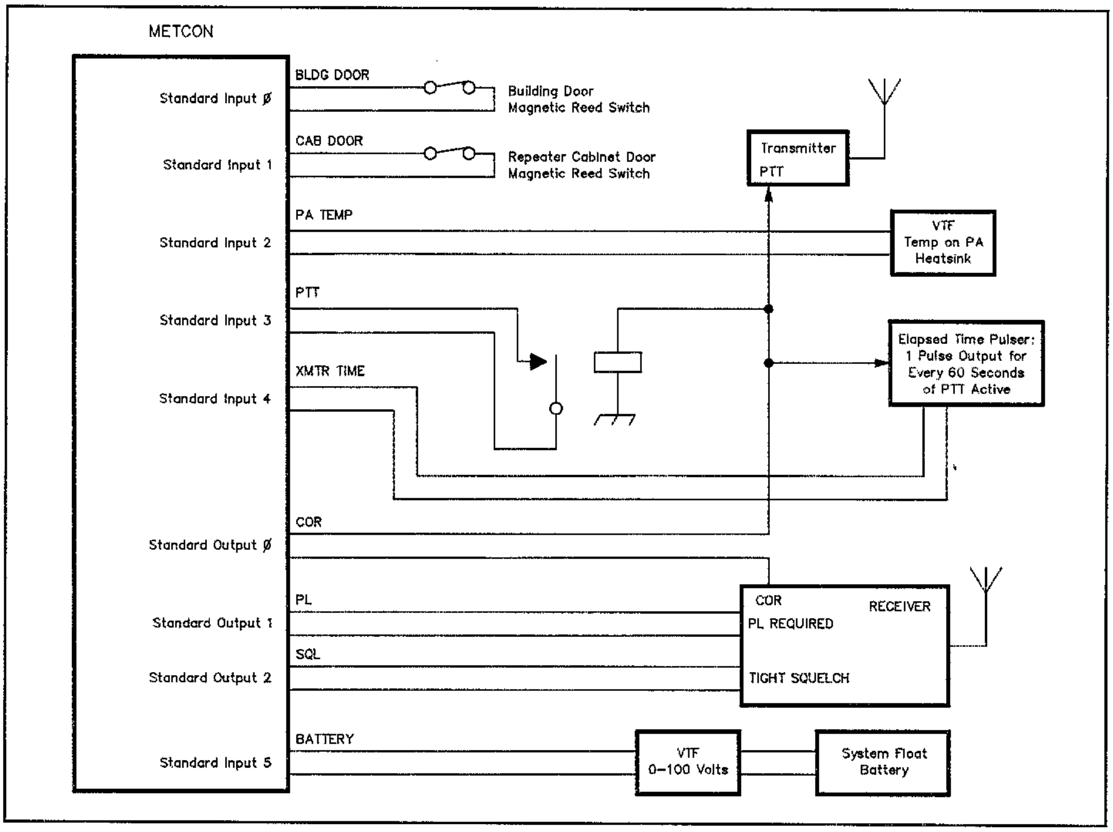


Fig 7—A METCON repeater monitoring and control setup might look something like this. Switches on the **BLDG DOOR** and **CAB DOOR** lines signal that these doors have been opened; the frequency of the pulse train applied to the **PA TEMP** line by a METCON voltage-to-frequency (VTF) module conveys the temperature of the transmitter power amplifier heat sink; relay closure on the **PA TEMP** line indicates transmitter activation; the number of pulses produced on the **XMTR TIME** line by a METCON Elapsed Time Pulser module provides transmitter-minute metering; and the frequency of the pulse train applied to the **BATTERY** line by a METCON VTF module conveys the system's float-battery voltage. METCON standard outputs control the system via **COR** (opening this line disables the repeater transmitter); **PL** (closing this line limits repeater access to input signals containing the proper subaudible tone); and **SQL** (closing this line reduces the repeater's carrier-operated-relay sensitivity by 10 dB).

A $5 Headset Mike

This easy-to-construct boom microphone clips onto your headphones to make operating more pleasurable. For a couple of dollars and hours, you can make one yourself!

By Doug Brede, W3AS/7
14410 E 45th Ct
Veradale, WA 99037

I'd long wished for a headset mike! But their prices—and the fact that commercial mikes usually come with a standard headset—had long deterred me from buying one. A headset mike keeps the mike element at the correct distance from your mouth for optimum modulation and clarity and frees your hands for writing, typing or flipping switches. A headset mike is especially convenient when you're DXing, contesting or ragchewing for hours at a time.

My option? Build one! And that I did. This mike is inexpensive and easy to construct, even for a novice builder. Parts are easy to obtain: They're available from Radio Shack and several mail-order houses. I bought the mike element on sale; the rest of the parts came from my meager junk box. The entire project cost me about a dollar. Even if you don't have a junk box, it shouldn't cost you much more.

You may think that such an inexpensive mike sounds tinny. That's not true of this one! The electret mike element is a high-fidelity instrument. In side-by-side, on-the-air comparisons with a Kenwood MC-50, an Astatic D-104 and my headset mike, listeners had a hard time telling them apart. Many listeners were surprised by the high-quality signal from the tiny headset mike.

Construction

The mike circuit schematic is in Fig 1. Not very complicated, is it? For this job, use a fine-pointed tip on your soldering iron. Trying to solder these parts using a large soldering gun or iron is bound to be frustrating at best: The mike element is less than ½ inch in diameter!

First, attach a 6-inch piece of slightly rigid copper wire (#18 or equivalent) to a claw test clip (see Fig 2); a claw test clip provides a firmer hold than a small-toothed alligator clip. The wire acts as the mike boom. Snuggle into your headphones and locate the best spot to attach the claw clip to the headset—a mirror may help you find that spot.

Next, trim the support wire to length. Ideally, the mike should be positioned at the corner of your mouth, rather than directly in front of it. This location helps minimize popping—the unwanted noise you get when you speak and breathe directly into the mike.

Slide the support wire and the end of a length of miniature coax (such as RG-174)—or a small-diameter, shielded, single-conductor cable—through a 6-inch piece of heat-shrink tubing. Solder the cable to the two mike-element connectors and carefully solder the support wire to the mike's case. If your mike element has an aluminum case, solder the support wire to the mike's ground connector instead. After you've verified that the mike element is in the correct position, heat the shrink tubing.

The tiny electret mike element contains a small amplifier (see the sidebar "How Does an Electret Microphone Work?"), so it requires a dc power source to operate. My Kenwood TS-440S—and many other popular rigs—has a low dc voltage on one of the transceiver's mike-connector pins. You can derive the mike's operating voltage from

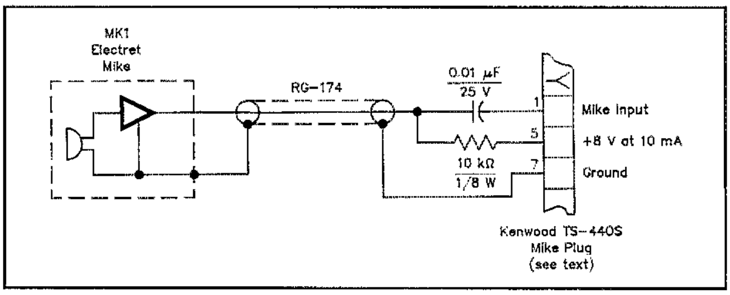


Fig 1—Schematic of the $5 Headset Mike circuit. Part numbers in parentheses are Radio Shack; equivalent parts can be substituted. MK1—Miniature electret microphone element (270-090 or 270-092). Misc: Claw clip (270-345), 4 to 5 feet of RG-174 miniature coaxial cable, 6 inches of heat-shrink tubing, a 6-inch piece of semirigid bare #18 copper wire, an 8-pin microphone connector (274-025A), or other connector required to match your rig's mike jack.

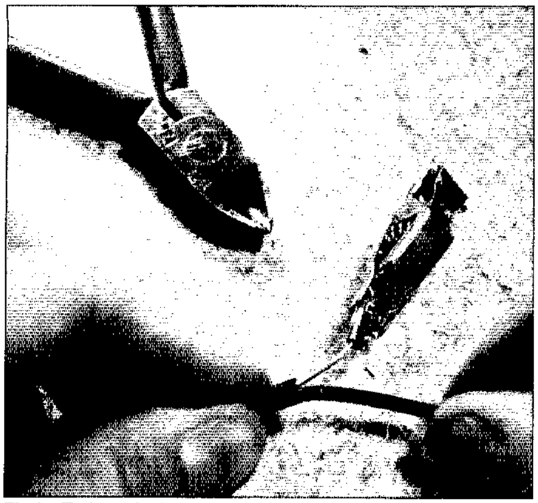


Fig 2—To form the headset-mike boom, slide the support wire and the end of a length of miniature coax through a 6-inch piece of heat-shrink tubing. Solder the coaxial cable to the two connectors on the mike element and carefully solder the support wire onto the mike's case. If your mike element has an aluminum case, solder the support wire to the mike's ground connector.

this pin because the mike element draws only 1 mA. If your rig doesn't have a dc voltage on the mike connector, you can power the mike with a 1.5- to 9-V battery. I installed the resistor and capacitor inside the mike plug. Only miniature components will fit, of course.

Summary

Because of the stony soil at my location, my shack has a notoriously poor RF ground. As a result, I have RF all over the place. In spite of this, the headset mike operates without RF pickup or distortion. Having the circuit components within the shielded mike plug helps, I'm sure.

You'll find this tiny mike has a lot of output. I've used many mikes that barely drive my transceiver; this one's output is a pleasant surprise. You may need to crank back a bit on your rig's mike and VOX gain controls. I'm sure you'll find the couple of bucks and hours you've put into this project worthwhile.

How Does an Electret Microphone Work?

The electret mike is a relatively modern innovation that relies on a very old electrical principle. The sound-receiving mechanism of the electret mike is a variable capacitor. One capacitor plate is fixed, while the other is flexible. The flexible capacitor plate vibrates in response to sounds it picks up. As it moves, the device capacitance changes.

To make the movable capacitor plate sensitive to high-pitched sound, a very thin, movable diaphragm is used. The diaphragm is often constructed of a plastic material, such as polyester film, coated with an extremely thin layer of gold.

A dc potential of 50 to 200 volts is used to form the polarized capacitor element. Early models of this microphone (known as *condenser* mikes) required a hefty bias supply to provide this voltage. Because of the required power supply, condenser mikes were too big and heavy for anything but fixed applications.

The introduction of the electret capacitor eliminated the need for the bulky bias supply. The electret capacitor has an advantage over older capacitor designs: With traditional capacitors, if you apply and remove a voltage to the plates, the charge gradually diminishes. The electret capacitor, on the other hand, holds a charge *indefinitely*. That's because at the heart of an electret is a dielectric piece that is permanently polarized electrically and so possesses a permanent electric field. This means that the electret capacitor charging can be done by the manufacturer, eliminating the need for you to provide a polarizing voltage.

An electret mike, however, still requires a small external supply to operate. That supply is used to power a small, internal FET amplifier which serves as an impedance converter. An electret element itself has an impedance of about 10 megohms—that won't match the mike-input impedance of your rig. The FET converter output impedance is between 50 and 2200 ohms—depending on the design of the particular electret mike—and that's suitable for most ham transmitters.

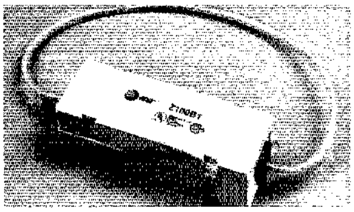
One to 10 volts at 1 mA is sufficient to power the FET amplifier. A small penlight battery can provide the voltage, often for several years. Power can also be drawn from the transmitter's mike connector; that eliminates the need for a battery. The voltage is applied to the electret mike element by the same two wires used to carry the audio signal.—*W3AS/7*

Doug Brede is a research director for a major West Coast seed company and a former Associate Professor of Horticulture at Oklahoma State University. He has written over 100 technical articles for magazines in the lawn and golf-course industry. His previous articles in QST *appeared in 1980 and 1989. Doug holds an Extra Class license and operates on all HF bands. He's been licensed for 24 years and enjoys contests and DXing.*

New Products

TELEPHONE RFI FILTER

☐ When you want to fix a problem with a telephone, check with the obvious source: The engineers at AT&T have developed the Radio Interference Filter Z100B1 (a replacement for the Z100A), designed to reduce interference from AM and FM broadcasters, ham stations and CB radios. The manufacturer also claims that it may help reduce interference to telephone modems. The RFI filter plugs into any standard modular telephone jack, between the wall outlet and the phone or modem. Retail price $17.99 from AT&T Phone Centers and retailers. AT&T, 5 Wood Hollow Rd, Parsippany, NJ 07054-2899; tel 201-581-3000, fax 201-428-8553.



Hints and Kinks

Conducted By David Newkirk, WJ1Z
Senior Assistant Technical Editor

1 WATT IN, 30 WATTS OUT WITH POWER MOSFETS AT 80 METERS

☐ Here (Fig 1) is a stable broadband linear amplifier that uses two inexpensive IRF511 power MOSFETs in push-pull to obtain about 30 watts on 80/75 meters with 1 watt of drive. It was patterned after an amplifier described by Doug DeMaw.[1]

Because I heat sink the MOSFETs directly to the chassis enclosure with insulating wafers (Fig 2), no expensive heat sink need be purchased. An 8 × 8 × 2-inch aluminum enclosure is adequate for shielding and sinking purposes. I originally mounted and heat sinked this amplifier on an old frying pan that my wife gave me. One might say that I had a sizzling signal on 75.

Since the MOSFETs' turn-on voltages may differ slightly from device to device, I decided to bias each one separately by employing a scheme designed by Wes Hayward and Jeff Damm.[2]

To adjust this amplifier for Class B linear operation, remove drive and terminate the output with a 50-ohm dummy antenna. Insert an ammeter in the drain supply line and adjust R1, **BIAS 1**, and R2, **BIAS 2**, to the threshold at which quiescent current just starts to flow. If you prefer class AB operation, tweak R1 to show, say, 10 mA, and then tweak R2 to double this current—20 mA in this example. The quiescent current level is not critical if kept small, but the device-to-device balance is. I suggest that you not exceed 3.5 volts bias on either gate; I prefer a value in the 2.5-volt region.

The amplifier's stability is excellent. Proper power supply decoupling and keeping the input impedance low (200 ohms or less) were the only steps necessary to achieve what I call unconditional stability.

I should mention in closing that the

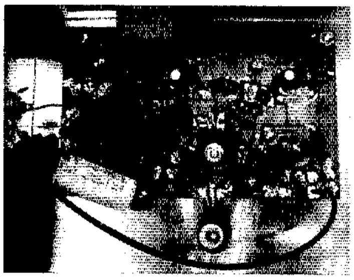


Fig 2—Version 1 of the amplifier used a frying-pan heat sink. This second version is a bit less adventurous: it uses a *chassis* as a heat sink.

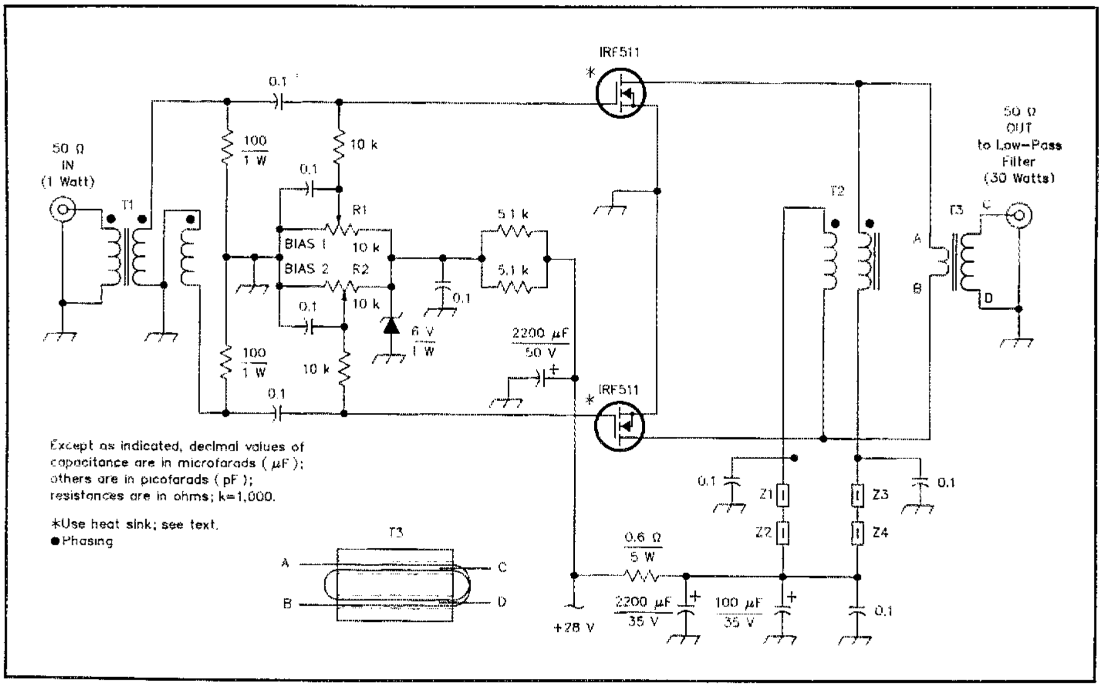


Fig 1—Jim Wyckoff runs two push-pull IRF511s at 28 volts to produce 30 watts at 80/75 meters. Resistors are ½ watt except where noted. Jim used an NTE #TP0006 insulating wafer between each IRF511 and the heat sink and wrapped their mounting bolts with electrician's tape to avoid short circuits. You *must* add low-pass filtering to this circuit to comply with FCC purity-of-emissions regulations. See the low-pass filter tables in Chapter 2 of the 1992 *ARRL Handbook.*

T1—12 trifilar turns of #26 enameled wire on an FT-50-43 (Amidon) or F-50-43 (Palomar) toroidal ferrite core.

T2—12 bifilar turns of #22 enameled wire on two stacked FT-50-43 (Amidon) or F-50-43 (Palomar) cores.

T3—See drawing inset. Wound with #18 plastic-insulated hook-up wire on an Amidon BN-43-7051 jumbo balun core. Primary (AB), 2 turns; secondary (CD), 3 turns.

Z1-Z4—Ferrite bead (Amidon FB-43-801 or Palomar FB-8-43).

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[1] D. DeMaw, "Go Class B or C with Power MOSFETs," *QST*, Mar 1983, pp 25-29. Also see Feedback in these 1983 *QSTs*: May, p 43; Jul, p 41; Dec, p 50.

[2] W. Hayward and J. Damm, "Stable HEXFET RF Power Amplifiers," *QST*, Technical Correspondence, Nov 1989, pp 38-40. Also see Feedback, *QST*, Mar 1990, p 41.

IRF511s' power output diminishes gradually as the operating frequency approaches 18 MHz. You can expect about 30 watts out on 80 meters and about 10 watts out on 20 meters. I get 5 watts out on 17 meters. (I report this last number only as a curiosity. ARRL Lab Engineer Zack Lau, KH6CP/1, suggested in a reply to my query about this phenomenon that the IRF511's high drain-to-ground capacitance tends to shunt more signal energy to ground as the operating frequency is increased.) Considering that these little devices generally cost under $2, they really do play from 3.5 to 14 MHz.—*Jim Wyckoff, AA3X, Severn, Maryland*

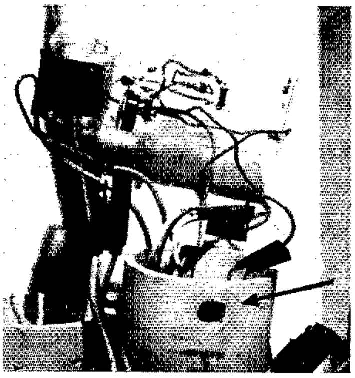

Fig 4—The circuit mounts on a plastic box lid that fits into a piece of PVC pipe. The phototransistor looks out of the pipe through the hole indicated by the arrow. *(photos by the author)*

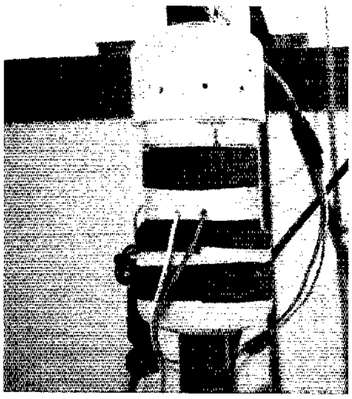

Fig 5—Holes of varying sizes drilled through the top-cap rim allow adjustment of the circuit's light sensitivity in conjunction with the circuit's THRESHOLD control.

A LIGHT-OPERATED SWITCH FOR SOLAR PANELS

☐ When I purchased my HF rig, I began to think about using Arizona sunshine as a power source. I already had a 2000-mA solar panel that I was using for various hobby projects. This was a good start, because a good solar panel costs $200 or more—the same as a line-operated power supply. I added a heavy-duty deep-cycle battery for energy storage.

A battery directly connected to a solar array starts discharging into the array as soon as light fails. A series diode can prevent this at the cost of voltage drop (hence, energy loss) in the diode. Diode loss would eat up the output of one or two of my panel's 33 cells! I wanted *all* of the panel's output voltage to be available for battery charging.

I decided to solve this problem by building a light-operated relay based on a cadmium-sulfide photoresistor (Fig 3). It consists of a voltage regulator (D1) followed by a transistor (Q1) that drives a normally open relay (K1).

To minimize the circuit's power drain, use a relay with as high a coil resistance as possible. My relay has a coil resistance of 500 ohms. Applying 9 volts to this coil results in a power dissipation of only about 0.2 watts.

Note that the system's solar panel, and not the battery, powers the circuit at start-up. As I sketched out the circuit, it seemed to me that once the relay had closed, feedback from the battery might keep it closed. This doesn't happen because the photoresistor does not receive enough light at night to allow Q1 to keep the relay on. When the relay opens at dusk, the battery is disconnected and the device will not operate again until dawn.

I constructed the circuit on a plastic box lid that slips into a short section of PVC pipe (Fig 4) and positioned the photoresistor in an opening near the pipe end. The pipe's top cap plays a role in adjusting the circuit's sensitivity. I drilled a series of holes of varying diameter around its circumference (Fig 5), and these can be centered over the photoresistor by twisting the cap. Selecting the proper aperture and adjusting the circuit's THRESHOLD control sets K1 to open under the desired light conditions. I taped the top cap in place once I'd found the right hole size, and taped a cap with a center drain hole to the pipe bottom.

For best exposure to ambient light at my location, I positioned the pipe on one of my carport's columns. Turning the relay off at will is easy: A piece of tape over the pipe's light aperture does the trick. —*Lawrence Harbison, N7HRN, Thatcher, Arizona*

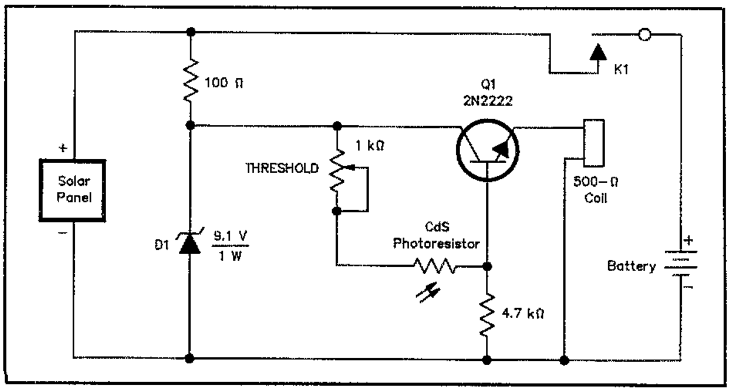

Fig 3—Lawrence Harbison uses this circuit, a light-operated relay, to disconnect his solar panel from his system battery between dusk and dawn.

CURING RFI IN THE HAM IV CONTROL UNIT

☐ When my beam was pointing toward my house and I was transmitting with about 400 watts, the needle of my HAM IV control unit's meter moved with the rhythm of my voice modulation and Morse code. The effect was quite distracting, and I tried to eliminate it using several methods but none of them worked. I grounded the unit, threaded its ac cable through ferrite beads and installed toroidal coils on the ac line. Using aluminum foil, I even created a grounded screen inside the top and bottom covers of the control box. None of these remedies helped.

Finally, I did what I should have done in the first place: I installed two 0.047-μF bypasses between the terminal board and ground—one between the #3 terminal and ground, the other from the #7 to ground. (These terminals connect to the rotator's direction-sensing potentiometer.) This completely solved the problem.

The particular capacitor value is probably not critical; anything from 0.01 to 0.1 μF should work. The capacitors used should be rated for at least 25 WVDC. —*George Pataki, WB2AQC, Queens, New York*

TAKE NOTES WITH A CAMCORDER

☐ We've all had the experience of working on a piece of equipment for the first time and forgetting exactly how it goes back together—things like how certain wires were routed or what order to reassemble things, or, worse yet, which wires went where. Murphy's Law says that this last will *certainly* happen when we don't have the schematic for the device in question.

Something that I've found can help is to set up my camcorder so that it has a clear view of what I'm doing, and let it record every action. Then, when I get confused about whether it was the green wire or the yellow wire that was attached to terminal C, I can review the tape. I'm careful to show the camcorder such details as which connector goes where, the position of DIP switches, color coding of wires, and such, and make verbal comments as I go. The camcorder often picks up things that I don't even notice at the time, but which can be important later on.

Since tapes are so inexpensive these days, I label the tape with the name of the piece of equipment and keep it for future reference. Although I use reasonably good tapes for this purpose, I don't use the highest grade, as I'll probably never want to copy it or even view it more than a time or two. The really cheap tapes, though, are too likely to shed oxide and foul the heads in the camcorder.

Don't restrict your use of this idea to just working on Amateur Radio gear, either. Anytime you have something complex to work on it can come in handy, whether it's working on the answering machine or a lawn-mower engine. If nothing else, you can at least use the tape to prove to a doubting spouse that you didn't spend the whole afternoon watching football!—*Clark Jones, N7RPQ, Phoenix, Arizona*

Club Competition Rules and Contest Disqualification Criteria

The 1993 contest season is upon us. Four ARRL-sponsored contests in 1993 include an ARRL-affiliated club competition: January VHF Sweepstakes, February/March International DX Contest, November Sweepstakes and December 160-Meter Contest. There are a few ground rules to follow to ensure that your club's scores are properly credited (and to ease the log checker's burden). These are detailed below.

From time to time it becomes necessary to consider disqualifying an entry to an ARRL contest. The particulars are listed below. Most often, the reason is simply that the person submitting the entry wasn't accurate in copying call signs or contest exchanges. As long as you're careful only to log QSOs when you're sure of the information, you should have nothing to worry about. (The use of standard ARRL contest forms and computer disk entries will help ensure that your score is figured properly and will help speed up the publication of contest results in *QST*.)

Call or write if you have a question about the rules listed here or for any contest. The time to ask is before the contest, not afterward.

Club Competition

Only ARRL active affiliated clubs may participate in the club competition. A member must be listed in the regular score listings to be counted for a club.

For a club to be listed, two conditions must be met:

1) At least *three different entries* from members of the club must be submitted.

2) All members wishing to be included in the club score *must indicate the club name on their summary sheets and the club secretary must send a list of all club members eligible to compete for the club and which level (unlimited, medium, local) they wish to enter for each competition.* Remember to meet the mailing deadline!

There are three levels of club competition:

1) **Unlimited:** Any club submitting *51 or more entries.* (One station can submit two entries—one on CW and one on phone in the November Sweepstakes and the DX Contest.) All stations and all operators must reside within 175 miles of the club's center. All members must attend at least two club meetings per year to be eligible to submit an entry. If, however, they haven't been a member for a year's time, they must have attended a meeting as a member prior to the contest. Club members who are disabled in such a way that they are unable to travel are exempt from the two meetings per year rule, but they must be regularly active in club affairs. To be considered *bona fide*, a member must be active in club affairs. Members living outside of 175 miles and/or members operating stations outside 175 miles may not compete in the club competition. The club must be actively ARRL affiliated.

2) **Medium:** Any club submitting *50 or fewer entries*, except as noted in local club criteria below. The same mileage and attendance requirements apply as the unlimited class club. The club must be actively ARRL affiliated.

3) **Local:** Any club submitting *10 or fewer entries.* All members must reside and operate within 20 miles of the club's center. There is no attendance requirement. The club must be actively ARRL affiliated.

Single-operator and multioperator station scores may be counted. At a guest-operated single-operator station, *both* the guest operator and the station licensee must be members of the same club to count the score for that club. At multioperator stations, at least 66% of the operators must be members of the same club for the score to count for that club. *A multioperator entry may (optional) use nonmember operators licensed one year or less without including such operators in the above 66% calculation. The intent here is to encourage clubs to recruit contesters from newer amateurs without adversely affecting the club aggregate score.*

In conjunction with the two meetings per year rule, the club must hold *at least four* in-person meetings per year. A club's entry classification may be changed if, in the opinion of the ARRL Awards Committee, the club has manipulated its number of entries to fall into a lower classification (eg, if a club with 100 members submits only the 10 highest scores, even if more than 10 of its members wish to compete).

It is not within the intent of these rules that a club should vote out a member or that a member resign and then be voted back into the club later so the member-attendance rule can be met.

The highest scoring active affiliated club entry will be awarded a gavel in each category (unlimited, medium, local).

The highest single-operator CW score and the highest single-operator phone score (ARRL International DX Contest and ARRL November Sweepstakes) in any active affiliated club entry will be awarded with a club certificate when at least three single-operator CW and/or three single-operator phone scores are submitted.

Disqualification

If the claimed score of a participant is reduced by 2% or more, the entry may be disqualified. Score reduction doesn't include correction of arithmetic errors.

Score reduction may by made for taking credit for unconfirmed QSOs and/or multipliers, duplicate contacts and/or other scoring discrepancies.

An entry with more than 2% duplicate contacts left in the log or an entry in which more than 2% "rubber clocking" (altering the actual time to increase the operating time so that it's greater than the allowable limit) is detected will be automatically disqualified.

If a participant is disqualified, he or she will be barred from submitting an entry in the next annual running of that specific contest, eg, disqualification from the 1993 phone SS prohibits submission of an entry for the 1994 phone SS, but 1994 CW SS participation is okay.

The call signs of all disqualified participants will be listed in the *QST* contest report.

A participant on the borderline of disqualification, but not actually disqualified, may receive a warning letter.

For each duplicate contact or miscopied call sign removed from the log by HQ, three additional contacts will be deleted as a penalty. The penalty will not be considered part of the 2% disqualification criteria.

In all cases of question, the decisions of the ARRL Awards Committee are final.

Technical Correspondence

Conducted By Paul Pagel, N1FB
Associate Technical Editor

TOUCH-CONTROL-LAMP RFI

◊ Many Amateur Radio operators have experienced interference in the form of a broadband, slowly drifting, ac-modulated signal. In most cases, the listener brushes off the problem as "another TV birdie," and grudgingly puts up with the noise, knowing full well the frustration involved in identifying and removing the interfering signal.

Although radiated TV horizontal-oscillator harmonics are still a problem (particularly with older sets), there has been a marked reduction in the number of observed cases of interference from these sources. Improved TV-set design and the advent of cable TV, with the attendant removal of outside antennas, has contributed to a decrease in interference from TV (ITV) in many areas.

Another Monster Rises

Unfortunately, as that problem lessens, a relatively new problem has appeared: touch-controlled lamps and similar devices. These products operate on the capacitance-relay principle: Each generates and detects a weak RF signal forming a tuned control loop. The touch-sensitive device has a metallic surface (sensing contact) that you touch to control the device. The metallic surface is connected to the transmitter or receiver tuned circuit within the touch-controlled device. Your touching the metallic surface adds sufficient capacitance to detune the circuit. This detuning results in a loss of signal to the detector portion of the control system, which is translated into a command to change the state of the lamp (or other device). Depending on the complexity of the control unit, lamps can be turned on or off; some units can be adjusted to various brightness levels.

For the most part, the RF-sensing portion of these devices operates in the 150- to 200-kHz region, using a self-excited oscillator—usually a single transistor, but occasionally a TTL gate IC. Oscillator types vary from vendor to vendor, and include blocking oscillators and LC types. Power required to operate the unit is usually derived from the ac line by a simple half-wave rectifier and filter system. At best, the dc supply is poorly filtered, often containing a considerable ac component. Component aging often worsens these problems. The generated signal is both rich in harmonics (caused in part by the simplicity of the oscillator design) and accompanied by considerable FM, a result of the poorly filtered dc.

Identifying the Interference

From the ham's point of view, harmonics from these devices appear as broadband, ac-modulated signals, often appearing every 150 to 200 kHz across the band, with each appearance covering a bandwidth of 10 to 25 kHz on the lower bands, increasing to 20 to 50 kHz on the higher bands. Because the signals drift slowly, they may be noticeable one day and absent the next, depending on where in the band the station receiver is tuned. If you have access to a spectrum analyzer (or if your receiver is fitted with a band-scanning oscilloscope) you may be able to see that the interfering signals sweep across the band at a 60-Hz rate.

The signals sound much like TV set horizontal-oscillator harmonics, except: They cover a much wider bandwidth; they have a definite edge when tuning across the interference and variations within the window they occupy and they drift slowly up and down the band, sometimes taking several minutes to more than an hour to move into or out of a particular band segment. When tuning across such a signal with the station receiver in the CW/SSB mode, you'll observe conspicuous changes in the signal (both the beat-note character and strength). TV-set harmonics, by comparison, are relatively fixed in frequency and are (typically) only 2 or 3 kHz wide, even on the higher bands.

The strength of the RFI generated by these touch-control devices varies greatly, but in severe instances, the signals can create problems as distant as 1500 feet from the source. Conducted radiation from signals traveling along power wiring significantly extends this range.

A Two-Way Street

An interesting correlation exists between RFI generated by these devices and their susceptibility to RF interference from amateur transmitters. I've found that strong (kilowatt-range) amateur transmissions often cause the devices to respond as if someone had touched the control plate. In fact, this response is what helped me identify the source of a mysterious drifting RFI signal I first observed on 20 meters some years ago.

I was experiencing QRM from a weak, ac-modulated signal that slowly drifted up and down the band—it appeared to come continuously from one direction. My direction-finding gear led me to a neighbor's home about 300 feet from mine. I contacted him, and after I explained that I was looking for a source of interference, he agreed to let me inspect his home to find the source of the problem. Originally, I suspected a TV set or computer monitor, but couldn't find such a culprit. Then, the neighbor mentioned that sometimes his living-room lamps went on and off for no reason! Could the troubles be related? I learned that he had two touch-control lamps. We discovered that whenever I used high power on 40-meter CW, his lamps turned on or off every time I pressed the key! Because my on-air time is usually brief (as when making a DX contact) the trouble, from my neighbor's point of view, always disappeared after a few moments. He didn't connect the flashing lamps with my activities.

Further experimentation showed that his lamps responded to high-power signals on 20 and 10 meters as well (but not to 15-meter signals), and then only when my beam was pointed at his home. On 160 meters, a 100-watt "barefoot" signal was enough to cause his lamps to flash. I suspect that the differences are caused by resonances in his house wiring. I also noticed, after careful listening to the "inward" RFI on my station receiver, that the beat note changed pitch slightly each time I pressed the transmitter key. This indicated at least one way to identify these sources.

Experience shows that by the time the signal from a touch-controlled device is strong enough to cause problems in your receiver (above S3), it's usually vulnerable to your transmitted signals. So, if you suspect touch-control interference, turn your antenna (if you can) until the interference is loudest, then transmit a few test signals (slow dashes) at your highest power level. If the received-signal audio note changes each time you transmit, you probably have lamp-dimmer RFI! By contrast, TV-oscillator harmonics shift abruptly with changes in the video portion of the TV signal (as during scene changes) and won't be affected by your transmissions. You may have to use different frequency bands to get a response. Of course, failure of the dimmer to respond to your signals doesn't eliminate them as a source of QRM. In some cases, I noticed that extended operating (ragchewing) caused the QRM to move quickly out of the band I was using. I suspect that rapid cycling of the device caused it to warm up somewhat. That shifted the oscillator a few kilohertz (at its fundamental), which was enough to move the harmonic away from where I was operating.

Noise-Blanker Action

Modern noise blankers are remarkably effective in eliminating noise interference, and will reduce (but not eliminate) noise from the lamps. What they do, in some cases, is to partially blank the interfering signal, so that your receiver seems to suffer from an increased noise floor, which returns to normal as you tune outside the sweep range of the harmonic. This effect is most noticeable in the CW/SSB mode. Try turning off your noise blanker, and/or using the AM receive mode when searching for this type of interference. Also, I've found that the edges of the swept interference are frequently stronger than the central portion of the signal.

Locating the Interference

Once you suspect a lamp as the source of your problem, you've got to find it. One of the best tools I've used is a portable shortwave receiver (I use a Sony ICF-2010) equipped with a loop for the band(s) of interest. Use your beam to get a general interfering-signal direction, then move along that line listening first with the nondirectional whip on the portable receiver. When the signal gets louder, change to the loop, and it will often lead you directly to the source. If you have no DF gear, use any portable radio (preferably one with shortwave coverage) and attempt to localize the source by signal-strength sniffing.

Simply walking under the service entrance wires with your radio is usually enough to get a usable signal. This type of interference tends to peak sharply near the meter base of the house or apartment involved. If the owner agrees to let you inside to continue your search, you'll find a strong peak as you approach the offending device. Many people own more than one such lamp. Be sure to *ask* if they have more than one, as different units may be causing problems on different bands, or at different spots in the same band. Also, I've found multiple units in the various apartments of a multi-unit complex, each contributing to the overall problem.

Fixing the Problem

When you've located the interfering lamp, you'll probably have to talk the owner into taking it out of service to clear up the problem. Plug-in ac filters or common-mode RF chokes (winding the line cord around a ferrite core) sometimes are effective, but more often the touch plate of the device acts as an antenna and directly radiates the signal. In this case, plug-in filters may actually make the problem *worse!*

Touch-control devices are covered by FCC Part 15 rules, which—among other things—prohibit them from interfering with communications services. Unfortunately, the FCC has better things to do than bust some homeowner, and will probably be uninterested in helping you solve the problem other than sending the homeowner an RFI booklet. It doesn't hurt to ask the Feds to intervene, however. Who knows, they may be having a slow day and be willing to help! In any case, endeavor to obtain the make and model number of the touch-control device, its purchase date and place of purchase. Forward this data and any other pertinent details—such as how the interference impacts your operations—to the FCC and the ARRL. Remember the golden rule of interference: If no one complains, it's not interference!

If you are "fortunate" in that your transmitted signals cause the device to flash on and off, the owner may be willing to take it out of service or return it to the store where it was purchased. If this is not the case, you may be able to talk the owner out of using it on safety grounds ("It isn't supposed to cause interference, but it does, which means it may be defective, and you certainly don't want a piece of defective electrical apparatus in your home"). If none of this works, and if filtering is ineffective, you either put up with it, or offer to purchase the lamp and trash it yourself. Of course, this last solution is fraught with the danger that the owner will just go out and buy another lamp! And yes, write the FCC and request that they send the owner of the interfering device a letter advising them of the fact that their lamp (or whatever) is causing illegal interference to radio communications, and requesting that it be permanently removed from service. Be sure you give the FCC a summary of the facts and what steps you have taken to date. They may not agree to send such a letter, but if they do, it will likely be very effective. For most people, the receipt of an official letter is enough to effect the desired result.

Summary

If a sufficient number of complaints are received by the FCC,[1] perhaps manufacturers can be made to modify the devices to reduce or eliminate the problem. The solution seems simple: All that's required is to use an RF oscillator that produces a sine-wave signal having little or no harmonic content, a reasonably clean source of dc to operate the unit and the problems should disappear. A bit of RF bypassing will simultaneously reduce radiated harmonics and reduce susceptibility to RFI from amateur sources. Economic considerations are probably not a factor either, since it might take as little as 50 cents difference in the manufactured cost to convert to sine-wave oscillators and properly filter dc sources for the control module. It's even possible that at least some of the manufacturers are unaware there's a problem, and education may be in order as to the ramifications of oversimplified design.—*James B. Wiley, KL7CC, 8023 E 11th Ct, Anchorage, AK 99504*

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[1] Send for the ARRL RFI Package. You can get yours by sending a 9- × 12-inch SASE with 98 cents postage to the Technical Department Secretary, ARRL, 225 Main St, Newington, CT 06111. The package contains a list of the telephone numbers and addresses of local FCC Field Offices.

Note: All correspondence addressed to the Technical Correspondence column should bear the name, call sign and complete address of the sender. Please include a daytime telephone number at which you can be reached if necessary.

Keep the author(s) in the communications loop. Whether praising or criticizing a work, copy the author(s) on comments sent to Technical Correspondence.

Feedback

◊ In November's Technical Correspondence column, there's an error on page 68, in the center column, almost directly across from Fig 5's caption. In the sentence fragment ". . . to the feed point by a factor $1 + \sin 2L$, but . . . " the formula should read "by a factor $1 + \sin^2 L$, but . . ."–*tnx Charles Michaels, W7XC*

◊ There is an error in the schematic diagram shown in "A Dual Radio Speaker," *QST*, Nov 1992, p 57. The connections to headphone audio switch S2 (section A and B) should be modified as shown in Fig 1.

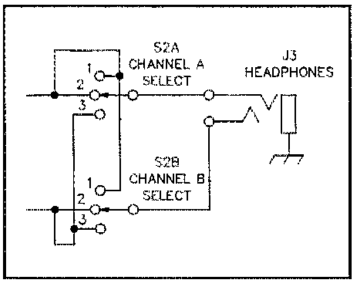

Fig 1.

New Products

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☐ The PackeTwin System speeds up packet radio with a dual-port IBM-compatible card that uses a single expansion slot and feeds two radios. You can select modems for each port that run independently at 1200, 2400, 9600 or 19,200 bauds, and add a compact, specially designed 440-MHz transceiver. The PackeTwin system comes with standard rig interfaces and a free version of KA9Q *NOS* software. The system's UHF transceiver has fast switching time for high-speed packet and two watts of RF power output. You specify two crystal-controlled frequencies from 430-450 MHz, as a dual-channel simplex rig or in split mode for use with packet repeaters, and the factory ships it ready to run—all you add is a 70-cm antenna.

The complete PackeTwin system retails for $499 with the interface card, a 9600-baud modem, transceiver, cable and software. The card, modem and radio are also available separately. Gracilis Inc, 623 Palace St, Aurora, IL 60506; tel 708-897-9346 or 708-801-8800, fax 708-844-0183.

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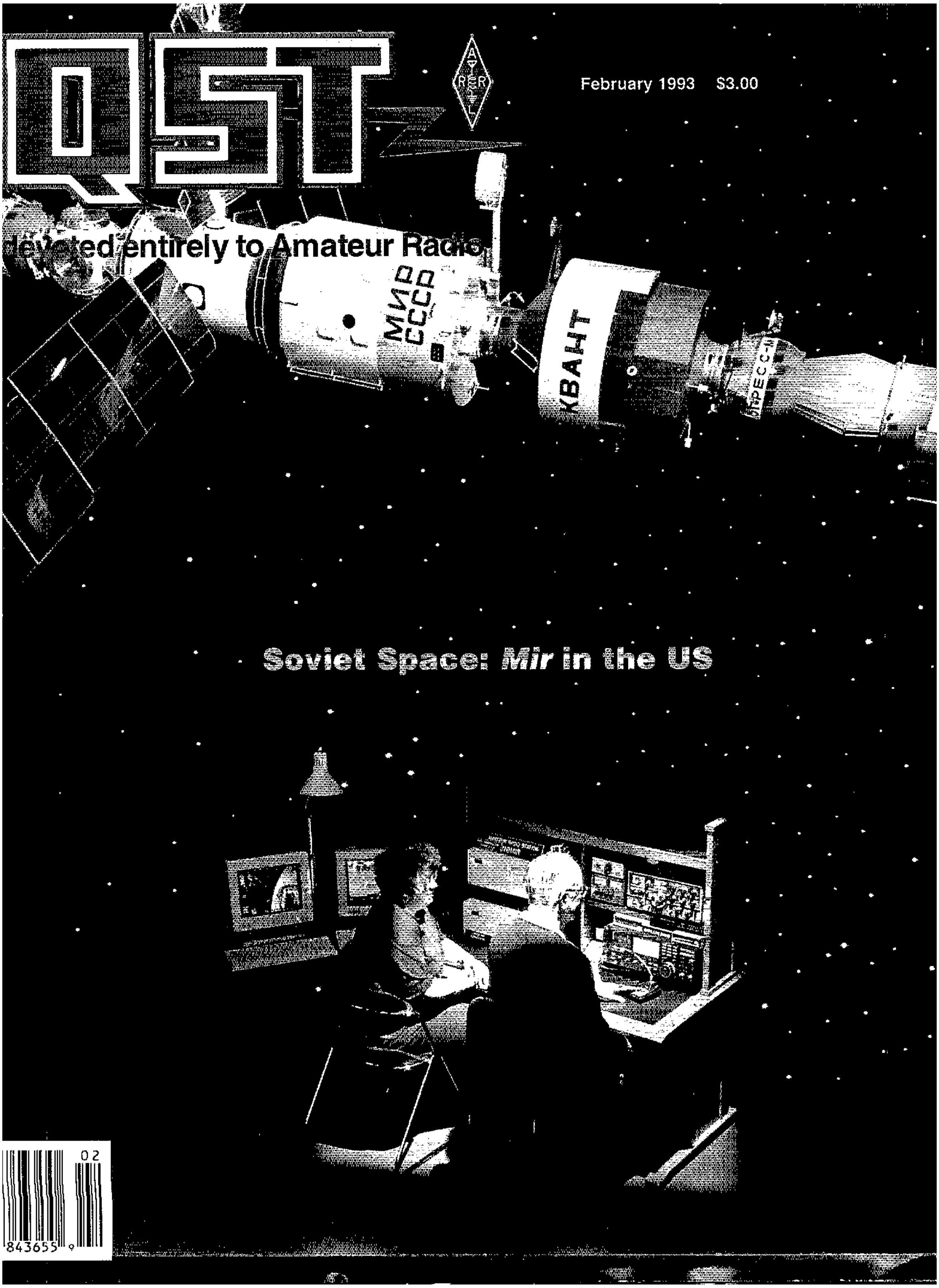

Soviet Space: *Mir* in the US

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OUR COVER

As the Russian *Mir* space station drifts overhead (a ⅛-size scale model), Roy Welch, W0SL, and Anna Tsadka operate the Amateur Radio station at the Soviet Space exhibit at the St Louis Science Center. Anna, who immigrated to the US in 1980, was interpreter for the presentation. See the story of the Soviet Space exhibit's travels to Ft Worth and St Louis, in this issue. *(photo by Sue Vanderbilt)*

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TECHNICAL

NEWS AND FEATURES

NEW HAM HORIZONS

OPERATING

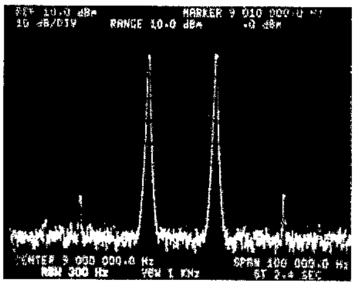
DEPARTMENTS

A High-Dynamic-Range MF/HF Receiver Front End

Read a two-tone IMD dynamic range specification of over 100 dB for an MF/HF receiver and you'll probably also see the qualifications "preamp off," "500-Hz IF bandwidth" and "signal spacing, 100 kHz." *This* front end hits 111 dB with preamp *on*, a 2.4-kHz IF bandwidth and the ARRL standard test-signal spacing of 20 kHz.

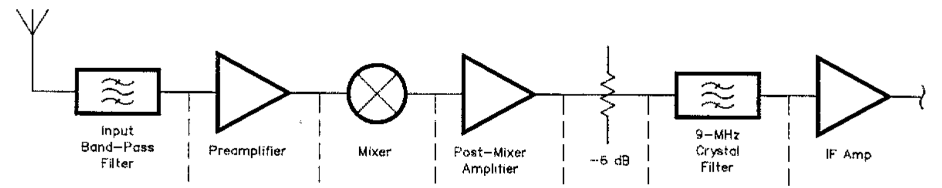
By Jacob Makhinson, N6NWP
1100 N Sunset Canyon Dr
Burbank, CA 91504

Proliferation of high-powered transmitters and electronic noise pollution imposes severe signal-handling requirements on modern receivers. Strong signal-handling capability has received significant attention from the manufacturers and reviewers of MF/HF amateur equipment during the last decade. Despite these technical advances, there is definitely room for some improvement. Yaesu's FT-1000, a state-of-the-art transceiver, sports a claimed dynamic range of 108 dB and an intercept point of +32 dBm (RF amplifier off, 500-Hz IF bandwidth, 50-kHz spacing). For a 2.4-kHz bandwidth, this dynamic range would drop to 103.5 dB.



A spectrum analysis of the front end's performance, preamplifier off, while converting two 0-dBm (1 milliwatt) 14-MHz signals spaced 20 kHz apart to the front end's 9-MHz IF. These appear at 8990 and 9010 kHz, with the system's third-order intermodulation products (8970 and 9030 kHz) down 64 dB relative to either test signal. This equates to a third-order input intercept of +42 dBm. *(spectrum photograph by the author)*

This article shows how to build a front end which achieves a two-tone dynamic range of 111 dB and an intercept point of +33 dBm—with *20-kHz* test-tone spacing, RF amplifier *on*, and using an IF bandwidth of 2.4 kHz. That this performance can be realized with a preamplifier in line is significant. It allows the realization of a receiver that combines strong signal-handling capability with excellent sensitivity—a challenge to any receiver designer. The circuit can be adjusted using test equipment available to most construction-minded amateurs, and its 50-ohm modu-



	Input Band-Pass Filter	Preamplifier	Mixer	Post-Mixer Amplifier	-6 dB	9-MHz Crystal Filter	IF Amp
Gain (A_V) (dB)	-0.3	8	-7	8	-6	-2	--
Noise Figure (NF) (dB)	0.3	2	7	2.5	6	2	3
Third-Order Input Intercept (IP_{3in}) (dBm)	+33	+32.7	+40.7	+34.5	--	--	--
1-dB Input Compression Point (P_C) (dBm)	+15.3	+15	+23	+18	--	--	--
1-dB Input Desense Point (P_D) (dBm)	+14.3	+14	+22	—	--	--	--

Fig 1—Stage-by-stage gain, noise figure, third-order input intercept, 1-dB compression point and 1-dB desensitization point performance of the high-dynamic-range front end.

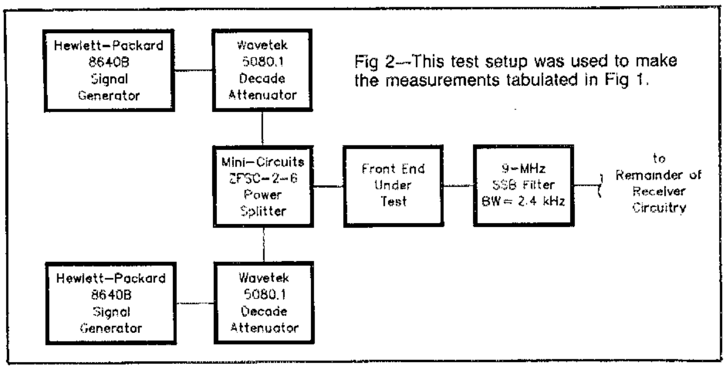


Fig 2—This test setup was used to make the measurements tabulated in Fig 1.

lar construction provides flexibility and encourages experimentation.

Front-End Design Considerations

The primary motivation was to design a front end capable of withstanding an abundance of strong signals several kilohertz away from a desired weak station. A traditional approach to this problem is the use of switchable preamplifiers and attenuators. Professional communications equipment sometimes employs PIN-diode attenuators,[1] correlating attenuation to signal level. Depending on the implementation, these methods may have one drawback: Sensitivity suffers, so the suppression of strong interfering signals is achieved at the cost of losing weak stations. Therefore, early in the design of this front end, it was decided that a preamplifier should precede the mixer to establish a low system noise figure.

Fig 1 shows a partial block diagram of a single-conversion receiver. This article covers only the preamplifier, mixer and the post-mixer amplifier in detail. A low-loss, band-pass input filter serves to protect the front end from strong out-of-band signals. The filter insertion loss is assumed to be 0.3 dB. For the sake of calculations, the IF amplifier noise figure is assumed to be 3 dB. Every block preceding the 9-MHz filter has 50-ohm input and output impedances. This greatly facilitates block replacement, promotes experimentation, and makes possible the removal of the preamplifier to connect the input band-pass filter directly to the mixer.

Measurement and Calibration Results

The measurement results are summarized in Fig 1. The preamplifier was evaluated at 14 MHz, and the mixer and post-mixer amplifier at 9 MHz. The test setup is shown in Fig 2. Table 1 summarizes the system measurement and calculation results.[2] Table 1 indicates that the preamplifier allows the achievement of a receiver sensitivity of 0.15 μV, on par with or better than any commercially produced MF/HF amateur transceiver, while keeping third-order input intercept and dynamic range at the respectable levels of +33 dBm and 111 dB, respectively. Extensive experiments on 14 MHz prove the front end's performance to be quite adequate under the most adverse receiving conditions.

Those wishing more flexibility can implement a preamp on/off option, although the intent of the design was to avoid the need of this switch. Amateurs who find a sensitivity of 0.32 μV adequate for their needs may choose not to use the preamplifier at all. If this is done, the front end becomes truly "uncrunchable" with an impressive third-order input intercept point of +41 dBm and a spurious-free dynamic range of 112 dB. The 1-dB input desensitization point[3,4,5] reaches +22 dBm in this case—a signal level of 2.8 V RMS at the antenna connector!

Overall System Considerations

Although sensitivity and dynamic range are primarily determined by the quality of the front end, some of the remaining receiver stages are of considerable importance. The IF amplifier should have an input signal handling capability of several volts, its distortion characteristics should be adequate, and its noise figure must be low enough not to appreciably degrade the front end's noise figure. A good crystal filter is a must; its shape factor should be 1.8 or better, and its ultimate attenuation should exceed 100 dB. Cascading two filters—one before the IF amplifier, and one after—should be considered as a way of improving the selectivity and also reducing the wideband noise generated by the IF amplifier. The product detector should not degrade the signal coming from the IF strip, and the audio amplifier's frequency response should be tailored with low- and high-pass filtering. Finally, extra care should be taken when designing the local oscillator (LO). A poorly designed oscillator may negate improvements in the front-end's dynamic range. Excessively high close-in phase noise[6,7] may degrade the receiver's ability to separate closely spaced signals. If we assume that the crystal filter provides out-of-band rejection of 100 dB at an offset of 2.0 kHz, the LO's phase noise must be better than −133 dBc/Hz at a 2.0-kHz offset to preserve the filter's performance. High far-out phase noise may raise the receiver noise floor; this can degrade the receiver's dynamic range. The dynamic range limited by phase noise

Table 1
Front-End Performance

Preamplifier	*With*	*Without*
Third-Order Input Intercept (IP_{3in}, dBm)	+33	+41
1-dB Input Compression Point (P_c, dBm)	+15.3	+23.3
1-dB Input Desensitization Point (P_D, dBm)	+14.3	+22.3
System Noise Figure (NF, dB)	6.6	12.9
Minimum Discernible Signal (MDS, dBm)	−133.4	−127.1
Sensitivity for 10 dB (S+N)/N Ratio in 50 Ω (Vs, μV RMS)	0.15	0.32
Spurious-Free Dynamic Range (SFDR, dB)*	111.0	112.1
Blocking Dynamic Range (BDR, dB)	147.7	149.4

*Also known as *two-tone dynamic range*; measured at ARRL Lab standard test-signal spacing of 20 kHz.

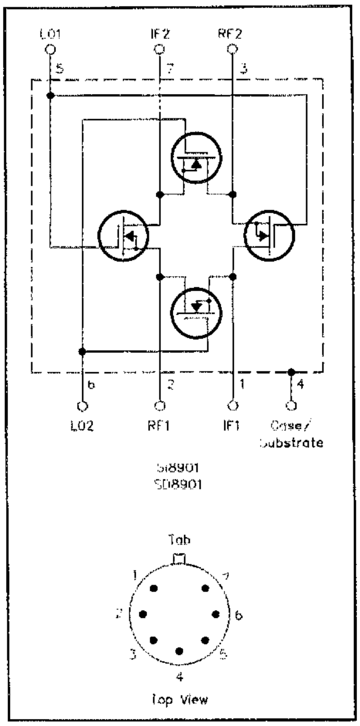


Fig 3—The front end uses a Siliconix Si8901 DMOS FET quad mixer. Siliconix has since discontinued this part, but an exact replacement (the SD8901) is available from Calogic Corporation.

[1] Notes appear on page 28.

should be equal to or better than the dynamic range limited by IMD.[8] Calculations show that the LO's far-out phase noise should be better than − 145 dBc/Hz. Although the LO spectral-purity requirements are quite stringent, they can be met by employing an LC-type oscillator with a high-Q coil.

Circuit Description

Mixer

This stage received special attention since its characteristics have a profound effect on dynamic range. Numerous articles have been written covering the performance of different types of mixers. For those interested in this subject, the works cited at Notes 1, 5, 9, 10, 11 and Additional References 1 through 4 are recommended.

After an extensive survey and experimentation, it was concluded that a passive mixer with active devices can provide the highest input intercept point. The Siliconix Si8901—a monolithic quad DMOS FET designed by Ed Oxner, KB6QJ—was chosen for the mixer. This excellent device was specifically designed for mixer applications and truly belongs at every experimenter's bench. Siliconix Application Note

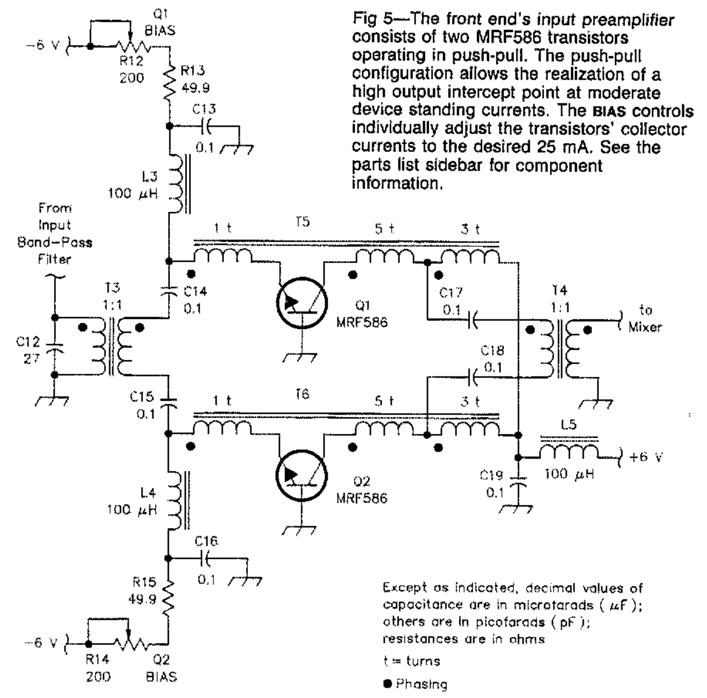


Fig 5—The front end's input preamplifier consists of two MRF586 transistors operating in push-pull. The push-pull configuration allows the realization of a high output intercept point at moderate device standing currents. The **BIAS** controls individually adjust the transistors' collector currents to the desired 25 mA. See the parts list sidebar for component information.

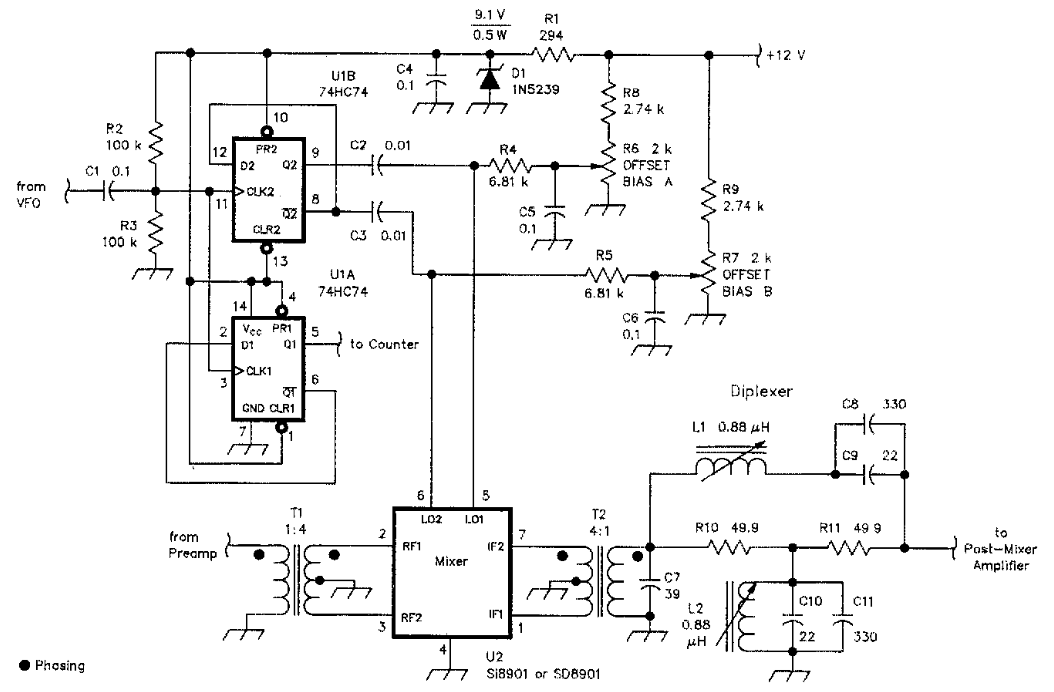


Fig 4—The front end's mixer circuitry applies the Si8901/SD8901 in conjunction with a 74HC74 dual flip-flop, a diplexer network and a VFO operating in the 10.5-MHz region—twice the frequency necessary to convert 14-MHz signals to the system's 9-MHz IF. Both 74HC74 sections divide the LO frequency by two and square it; one section (U1A) provides LO output for a frequency counter, and the other (U1B) drives the mixer's LO port. The **OFFSET BIAS** controls adjust mixer balance. D1 sets the 74HC74's supply voltage at a level considerably higher than the part's specified 6-V maximum; see text. See the parts list sidebar for component information.

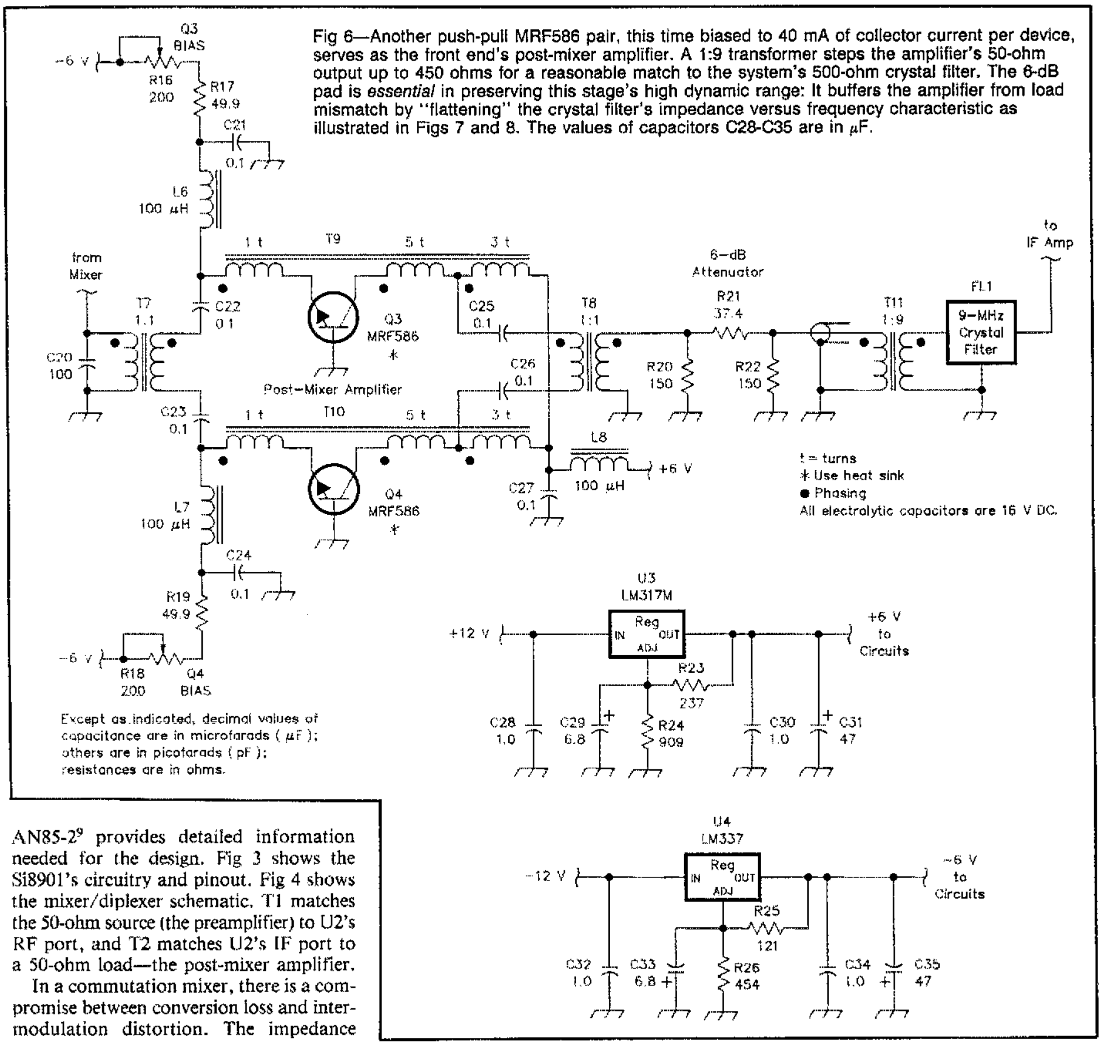


Fig 6—Another push-pull MRF586 pair, this time biased to 40 mA of collector current per device, serves as the front end's post-mixer amplifier. A 1:9 transformer steps the amplifier's 50-ohm output up to 450 ohms for a reasonable match to the system's 500-ohm crystal filter. The 6-dB pad is *essential* in preserving this stage's high dynamic range: It buffers the amplifier from load mismatch by "flattening" the crystal filter's impedance versus frequency characteristic as illustrated in Figs 7 and 8. The values of capacitors C28-C35 are in µF.

Except as indicated, decimal values of capacitance are in microfarads (µF); others are in picofarads (pF); resistances are in ohms.

AN85-2[9] provides detailed information needed for the design. Fig 3 shows the Si8901's circuitry and pinout. Fig 4 shows the mixer/diplexer schematic. T1 matches the 50-ohm source (the preamplifier) to U2's RF port, and T2 matches U2's IF port to a 50-ohm load—the post-mixer amplifier.

In a commutation mixer, there is a compromise between conversion loss and intermodulation distortion. The impedance

Parts List for the High-Dynamic Range Front End

Unless indicated otherwise, all of the circuits' resistors are ¼-W 1% tolerance, and all of its capacitors are 20%-tolerance ceramic. C7-C12 and C20 are dipped mica, polyester film or epoxy-coated ceramic.

C7—39 pF ±10%.
C8, C10—330 pF ±5%.
C9, C11—22 pF ±5%.
C12—27 pF ±10%.
C20—100 pF ±10%.
D1—9.1-V, 0.5-W Zener diode, 1N5239 suitable. See text.
L1, L2—Toko 332PN-T1012Z (Digi-Key TK5130) 1.0-µH variable inductor.
L3-L8—100-µH RF choke.
Q1-Q4—Motorola MRF586 RF transistor. Use heat sink (Thermalloy 2228B or Aavid 3257) on Q3 and Q4.
R6, R7—2-kΩ multiturn potentiometer.
R12, R14, R16, R18—200-Ω multiturn potentiometer.
T1, T2—Mini-Circuits T4-1 4:1 RF transformer.
T3, T4, T7, T8—Mini-Circuits T1-1T 1:1 RF transformer.
T5, T6, T9, T10—homemade trifilar transformer wound with #32 enameled wire on Amidon BN-43-2402 core (Fair-Rite 2843002402) as per Fig 9.
T11—Mini-Circuits T9-1 9:1 RF transformer.
U1—74HC74 CMOS flip-flop.
U2—Calogic Corporation SD8901HD mixer (replaces the Siliconix Si8901HD). Siliconix has discontinued the Si8901; the Calogic SD8901 is an exact replacement.
U3—LM317MP adjustable voltage regulator IC.
U4—LM337MP adjustable voltage regulator IC.

A list of the names and addresses of parts suppliers, keyed to the components in this list, is available as detailed in Note 2.

ratios of transmission-line transformers T1 and T2 are chosen to achieve low intermodulation distortion while keeping conversion loss at an acceptable level.

The LO waveform applied to U2 is of great importance. To achieve a high intercept point, the LO drive must approach an ideal square wave with a 50% duty cycle.[10,11] Flip-flop U1B divides the LO signal frequency by two and provides 50% duty cycle square waves at its complimentary outputs. The offset bias adjustment potentiometers, R6 and R7, allow compensation for mismatch among the Si8901's MOSFETs and ensure that the commutation mixer switches operate in a 50% duty cycle mode. U1A provides a signal that can be applied to a digital frequency counter.

The mixer's intercept point also depends on the LO drive level. Zener diode D1 establishes the power supply voltage for U1 and, therefore, the voltage swing at the mixer LO port. In the interest of obtaining a high intercept point, a 9.1-V diode is used at D1. This voltage exceeds the 74HC74's maximum supply specification. Experiments with samples from different manufacturers show that there is a considerable safety margin, however: The 74HC74s tested failed when the power supply was adjusted to around 11 V. Those who feel uncomfortable in exceeding the part's maximum rating may use a 6.8-V Zener diode at the cost of a few decibels in the intercept point value. Since double-balanced mixers are sensitive to IF port mismatches,[12] and a reactive load can cause an increase in conversion loss and degradation of the third-order intercept point, a band-pass diplexer follows the mixer. The diplexer comprises L1, C8, C9, L2, C10, C11, R10 and R11. The 9-MHz signal passes through this network with minimum attenuation, while out-of-passband signals over a wide frequency range are dissipated. Resistor R10 presents a 50-ohm impedance to the mixer, while R11 presents a 50-ohm impedance to the input of the post-mixer amplifier. C7 cancels the inductive reactance of the load presented to the mixer's IF port.

The measured performance of this circuit is: third-order output intercept, +35 dBm; 1-dB output compression point, +16 dBm; 1-dB output desensitization (blocking) point, +15 dBm; insertion loss, 7 dB.

Preamplifier

To preserve the mixer's dynamic range, the preamplifier's output intercept point should exceed the mixer input intercept point by at least 3 dB, and its compression point should exceed the mixer's by at least 1 dB.[13] The preamplifier's design criteria are as follows: third-order output intercept, ≥45 dBm; 1-dB output compression point, ≥24 dBm; gain between 6 and 9 dB; noise figure <3.0 dB; −1-dB frequency response, 1.8-30 MHz.

If conventional resistive feedback techniques were used to meet the design objectives, the required transistor standing current would become prohibitively high, resulting in a high noise figure. Therefore, a combination of a "noiseless feedback" method[14-18] and a push-pull configuration was tried, and resulted in the circuit shown in Fig 5.

The MRF586 is a high-frequency transistor designed for use in low-noise, ultralinear amplifiers and packaged in a TO-205 case. Used in a common-base configuration, this device provides a low noise figure because its optimum source resistance is close to 50 ohms. Bias controls R12 and R14 allow the collector current of each transistor to be adjusted to the desired value of 25 mA. T3 matches the amplifier input to the 50-ohm band-pass input filter; T4 matches the preamp output to the mixer signal port.

Transformers T5 and T6 are designed to couple part of the collector signal back into the emitter (negative feedback), set the gain to 8 dB, and set the input and output impedances to 50 ohms. C12 cancels the inductive reactance of the source presented to the mixer's RF port at 14 MHz. (In a multiband receiver, this capacitor may be part of switchable preselector filters.) The measured performance of the circuit is: third-order output intercept, +48 dBm; 1-dB output compression point, +25 dBm; gain, 8.0 dB; noise figure, 2.0 dB; −1-dB frequency response 1-40 MHz. This performance fully meets the preamplifier design goals.

Post-Mixer Amplifier

Although the post-mixer amplifier (Fig 6) uses the same basic configuration as the preamplifier, its design criteria are different. The post-mixer amplifier's input intercept point should exceed the mixer output intercept point by at least 3 dB, and its compression point should exceed the mixer's by at least 1 dB. The post-mixer amplifier must therefore exhibit these characteristics: third-order input intercept, ≥38 dBm; 1-dB input compression point, ≥17 dBm; gain between 6 and 9 dB; and noise figure, <3.0 dB.

These design goals could easily be met if the post-mixer amplifier had a purely resistive termination. However, the amplifier must drive a crystal filter, the impedance of which varies considerably with frequency. Connecting the post-mixer amplifier to the crystal filter via a matching transformer would severely degrade the amplifier's third-order intercept point. Inserting resistive attenuation between the amplifier and filter reduces this effect to an acceptable level, as shown by Figs 7 and 8.

BIAS controls R16 and R18 allow adjustment of the MRF586s' collector current to 40 mA per device. This value is 15 mA higher than the preamplifier's device standing currents to overcome the negative effects of the reactive load on the post-mixer amplifier's intercept point. T11, a transmission-line transformer, matches the crystal filter's impedance (500 ohms in this case) to the amplifier's 50-ohm output impedance. C20 cancels the inductive reactance of the source presented to the amplifier's input. The measured performance of the post-mixer amplifier, terminated by a 9-MHz filter via the 6-dB pad, is: third-order input intercept, +34 dBm; 1-dB input compression point, +18 dBm; gain, 8.0 dB; and noise figure, 2.5 dB. In spite of the measures taken, the input intercept point requirement has not been met. Overall system measurements indicate, however, that the mixer's third-order input

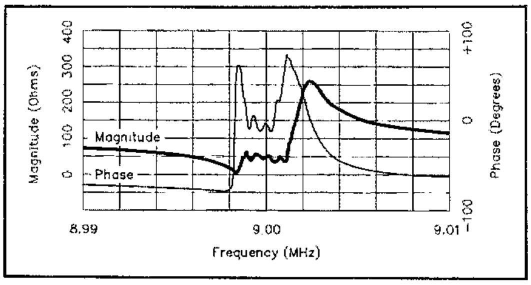


Fig 7—The crystal filter's input impedance as seen by the post-mixer amplifier through a 1:9 matching transformer. The reflections caused by this characteristic seriously degrade the amplifier's strong-signal performance if the 6-dB pad shown in Fig 6 is omitted.

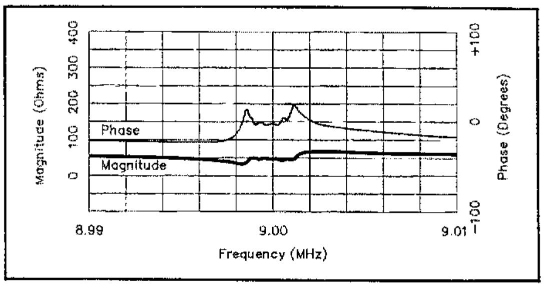


Fig 8—The filter impedance as seen by the amplifier through the 1:9 transformer and a 6-dB pad. The reflections still occur—they must for the filter to act as a filter—but the pad reduces their magnitude sufficiently to preserve the amplifier's high dynamic range.

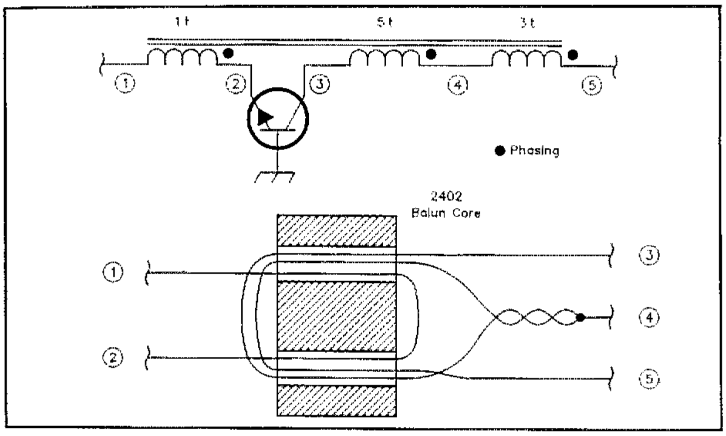


Fig 9—Transformers T5, T6, T9 and T10 are wound with #32 wire on two-hole ferrite cores (see the parts list sidebar for core information). Wind the 1-turn winding (1-2) first, the 3-turn winding (3-4) second and the 5-turn winding (4-5) third.

intercept point has been degraded by less than 1.5 dB (40.7 dB versus 42 dB) and is still very high.

Construction

The three modules are enclosed in separate, unpainted aluminum boxes. Each module's components are mounted on Vector 8007 board (perforated, with a solid copper plane on one side). The circuit components are mounted on the solid copper side, which serves as a ground plane. Vector T68 pins serve as terminal posts as necessary. All modules' signal inputs and outputs are made via BNC connectors; RG-174 coaxial cable interconnects the modules.

Properly built and shielded, the amplifiers should not oscillate. Because they operate in push-pull, strict symmetry should be observed in constructing the upper and lower halves of each amplifier.

The matching transformers specified at T1 and T2 (1:4 ratio) T3, T4, T7 and T8 (1:1 ratio); and T11 (9:1) are commercial units. If home-built matching transformers are used instead, they should be transmission-line types and highly symmetrical. Transformers T5, T6, T9 and T10 are homemade as shown in Fig 9.

Alignment

Amplifiers

Adjust controls R12 and R14, **Q1 BIAS** and **Q2 BIAS** (Fig 5), for a collector current of 25 mA per device. Adjust controls R16 and R18, **Q3 BIAS** and **Q4 BIAS**, for a collector current of 40 mA per device.

Mixer

The diplexer inductors, L1 and L2, must be aligned first. If the mixer is a part of an assembled receiver, inject a signal (100 mV P-P at, say, 14.250 MHz) into the antenna terminal. Tune the receiver to this frequency and adjust L1 and L2 for maximum signal while observing the signal at the post-mixer amplifier's output on an oscilloscope.

As an alternative method, apply a 14.250-MHz, 100-mV P-P signal to the preamplifier input. Using another signal generator, apply 4 V P-P at 10.5 MHz to the input (**from VFO**) of U1 (Fig 4). Peak L1 and L2 as described above.

For optimum performance, adjust R6 and R7 (**OFFSET BIAS A** and **OFFSET BIAS B**) while viewing the mixer's output spectrum with a spectrum analyzer. A two-tone IMD test setup (Fig 2) is required. (Detailed information on performing this test can be obtained from the works cited in Notes 3, 4, 5 and 10, and Additional References 1 and 8.) Connect the spectrum analyzer to the post-mixer amplifier output. Adjust R6 and R7 to minimize third-order-IMD products. Monitor the mixer insertion loss while doing this; it should be around 7 dB.

If a spectrum analyzer is unavailable, another method can be used. A 9-MHz IF strip with defeatable AGC is required. Turn off its AGC, connect it to the post-mixer amplifier, and connect a scope to its output. Tune the front end's VFO to convert one of its third-order IMD products ($2f1 - f2$ or $2f2 - f1$, where $f1$ and $f2$ are the test-signal frequencies) to 9.000 MHz. Adjust R6 and R7 to minimize the level of this signal.

Summary

By properly applying known design principles, radio amateurs can construct a high-performance front end which combines a very high intercept point with excellent sensitivity. Used with a low-noise LO, the front end described in this article achieves a wide dynamic range even with its preamplifier stage in line. A receiver incorporating such a front end can provide strong-signal performance that rivals or exceeds that of most commercial equipment available to the amateur.

Notes

[1] U. Rohde, "Recent developments in circuits and techniques for HF communications receivers," *ham radio*, Apr 1980, pp 20-25.

[2] An information package detailing the calculations on which this data is based, as well as a list of suppliers for the components listed in the "Parts List for the High-Dynamic Range Front End" sidebar, is available free of charge from the Technical Department Secretary, ARRL, 225 Main St, Newington CT 06111. Please mark your request MAKHINSON FRONT-END TEMPLATE and include a business-size SASE.

[3] C. Drentea, *Radio Communication Receivers* (Blue Ridge Summit, PA: TAB Books Inc, 1982), pp 88-135.

[4] R. Watson, "Guidelines for receiver analysis," *Microwaves and RF*, Dec 1986, pp 113-122.

[5] Wideband signal processing component catalog, VARI-L Company, Denver, CO.

[6] J. Grebenkemper, "Phase Noise and Its Effects on Amateur Communications," *Part 1, QST*, Mar 1988, pp 14-20; *Part Two*, Apr 1988, pp 22-25. Also see Feedback, *QST*, May 1988, p 44.

[7] P. Chadwick, "Phase Noise Intermodulation and Dynamic Range," *Frequency Dividers and Synthesisers IC Handbook*, Plessey Semiconductors, 1988.

[8] See Note 7.

[9] E. Oxner, "Designing a Super-High Dynamic Range Double-Balanced Mixer," Application Note AN 85-2 (Santa Clara, CA: Siliconix, 1986).

[10] V. Drozdov, *Lyubitel'skie kv Transiveri* (Amateur HF Transceivers), in Russian, Moscow, USSR, 1988.

[11] See Note 9.

[12] P. Drexler, "Effect of Termination Mismatches on Double-Balanced Mixers," *Microwave Journal*, Jan 1986, pp 187-190.

[13] E. Red, "*Daten, Fakten, HF-Grundschaltungen, 50 ohm-Technik*," in German or Russian, Franzis-Verlag GmbH, Munich, 1986.

[14] See Note 1.

[15] U. Rohde, "High dynamic range receiver input stages," *ham radio*, Oct 1975, pp 26-31.

[16] D. Norton, "High Dynamic Range Transistor Amplifier Using Lossless Feedback," *Microwave Journal*, May 1976, p 53.

[17] J. Reisert, "High dynamic range receivers," *ham radio*, Nov 1984, pp 97-105.

[18] See Note 13.

Additional References

1. D. DeMaw, *Practical RF Design Manual* (Englewood Cliffs, NJ: Prentice-Hall Inc, 1982), pp 23-25.

2. E. Oxner, "Junction FETs in Active Double-Balanced Mixers," Application Note AN 73-4 (Santa Clara: Siliconix, 1973).

3. U. Rohde, "Performance capability of active mixers," *Part 1, ham radio*, Mar 1982, pp 30-35; *Part 2, ham radio*, Apr 1982, pp 38-44.

4. W. Hayward, "Experiments with Primitive FET Mixers," *RF Design*, Nov 1990, pp 39-48.

5. N. Ayrandjian, "Simple Computation of Spurious-Free Dynamic Range," *RF Design*, Jan 1987, pp 65-66.

6. W. Richardson, "IMD and intercept points of cascaded stages," *ham radio*, Nov 1984, pp 28-34.

7. W. Hayward and D. DeMaw, *Solid-State Design for the Radio Amateur* (Newington: ARRL, 1986), Chapter 6: Advanced Receiver Concepts.

8. *RF/IF Signal Processing Handbook*, Vol 1 (Brooklyn, NY: Mini-Circuits, 1985), pp 7-23.

A radio amateur since 1986, Jacob Makhinson holds an Advanced Class license and has 20 years of experience in analog design. He received an MSEE from the Civil Aviation Engineering Institute, Riga, Latvia, in 1972, and since 1984 has been employed at Philips Ultrasound, where he is involved in the design of ultrasound imaging systems. His areas of technical interest in Amateur Radio include low-noise techniques and widening the dynamic range of HF receivers.

You Can Operate HF Mobile!

Interested in HF mobile operating, but strapped for cash? Don't worry about fancy antennas and expensive mounting hardware. With a little effort and imagination, you'll be on the air before you know it!

By Roger Burch, WF4N
Route 3, Box 235
Central City, KY 42330

As I read VE7AQC's letter in the March 1992 *QST* "Correspondence" column, I sympathized with Dr Campbell's frustration in trying to create an HF Amateur Radio mobile installation. In December 1992 *QST*, Ed Karsin, W3BMW, offered his idea for an HF mag-mount antenna. After operating HF mobile CW for nearly 10 years (and more than 150,000 miles), I've found another successful approach to the problem.

A Simple Antenna Mount

Beginning with the I-can't-mount-an-antenna-on-my-car's-rubber-covered-rear-bumper dilemma, Fig 1 shows the antenna mount I've used successfully on three different vehicles. It's simply a strip of aluminum plate of approximately 3/16 × 3 × 18 inches. The aluminum strip is attached to the rear frame rail with two ¼-inch sheet-metal screws (see Fig 2). The strip is bent so as to drop below, and extend just beyond, the rear bumper. This makes a very effective spring mount, allowing approximately 15° of rearward mast deflection at highway speeds, and even more if the mast is struck by low-hanging branches!

Fig 1—An aluminum strip secured to the car frame makes an excellent antenna mount. (*photos by Roger Burch, WF4N*)

I expected the metal to eventually fatigue and break, but that has yet to occur. On the contrary, the mount appears to be as sound as when I installed it in 1982. This is even more remarkable when you consider it was bent 90° during a minor mishap a few years ago. After quickly straightening the mount, I was back on the air!

The length of the mount doesn't significantly affect the impedance characteristics of my Hustler antenna. The coax braid is grounded to the mounting strip using a ¼-inch bolt, nut and lock washer. Using its insulating hardware, the Hustler is installed through a 5/16-inch hole drilled in the end of the mount. When the antenna and mount are assembled properly, obtaining a good impedance match is never a problem.

I originally used a "guy string" with my antenna. It was attached to the mast at the resonator-to-mast joint and anchored to the corner of the car roof. This was a big mistake! The guy formed a fulcrum which allowed the forces of wind and obstruction-induced deflections to be transmitted to the hinge of the mast. Within a year, the hinge was destroyed. The unguyed replacement mast has held up quite well—despite carrying as many as four resonators at once!

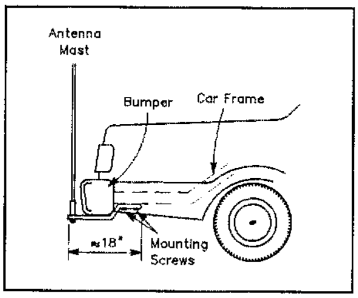


Fig 2—The antenna mount is attached to the frame with ¼-inch sheet metal screws. When bent into the proper shape, it extends below, and just beyond, the rear bumper.

Multiple Resonators

I've fabricated a simple adapter that allows me to install multiple resonators on a single mast (see Fig 3). It's also made of sheet aluminum, approximately 1/8-inch thick and cut in the shape of an *X*. Five 13/32-inch holes are drilled as shown in Fig 4. Each leg of the X is bent downward at a 30° angle. This keeps the outer resonators away from the center resonator. The center hole in the adapter fits the threaded end of the mast. The center resonator is then screwed onto the mast to hold the adapter in place. The outer resonators are attached with a 3/8-24 × 3/4-inch bolt and washer.

Mounting the Rig...and a Key!

It seems as though the available room in modern automobiles is shrinking faster than the sizes of our radios. (Would you believe my list of mobile rigs has included a TR-3 and an FT-101EE?) Nevertheless, with a little effort I've always been able to find a spot for my radio.

Take a look at Fig 5. The rig is mounted

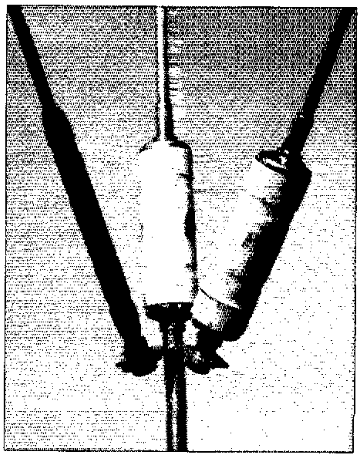


Fig 3—Three resonators on a single mast!

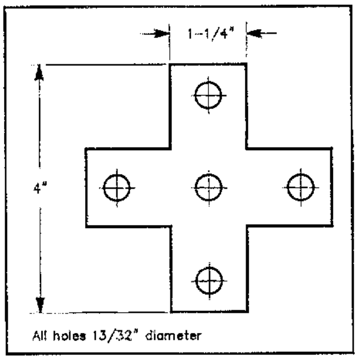


Fig 4—To create the multiple resonator adapter, a ⅛-inch thick piece of sheet aluminum is cut and drilled as shown. Each leg is bent downward at a 30° angle. The center hole is slipped over the top of the mast, followed by the center resonator which holds it in place. Additional resonators can then be attached at the corners.

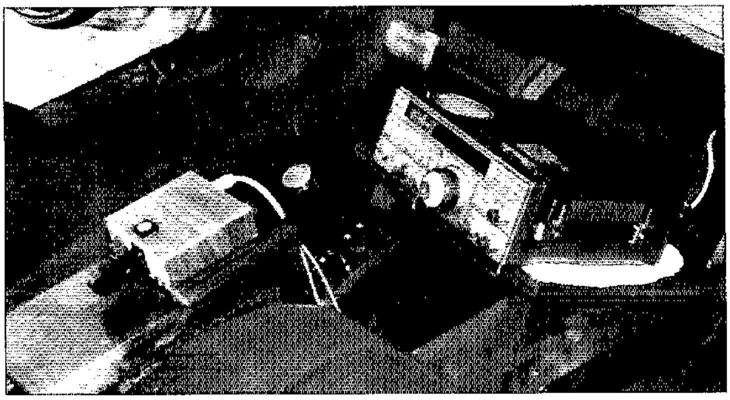


Fig 5—A heavy wood platform holds the HF transceiver while a smaller platform cradles the CW keyer.

on a platform I constructed from ¾-inch plywood and two short pieces of 2- × 8-inch lumber. The platform straddles the center hump and has yet to exhibit a tendency to wander. Gravity and the fold-down foot on the FT-757 keep the rig in place. In the interest of safety, however, the rig should be secured to the platform and the platform attached to the vehicle.

The CW keyer and paddles are mounted on another wood platform made from 2- × 6-inch stock. Not visible in Fig 5 are two 2- × 6-inch legs which extend from the base of the keyer platform to the floor of the car. Since the mount is wider than the gap between the seats, the pressure from the seats holds it in place.

Two pieces of quarter-round stock secure the keyer to the mount. A finish nail is driven through the center of each piece of stock, forming rails to grip the keyer housing. These pieces pivot on their nails as the keyer is installed and are spaced to provide a snug fit. [This setup is fine for casual, infrequent operating. In the event of an accident, however, the keyer and platform are likely to become airborne. Consider a more secure mounting system if you plan to operate CW mobile on a regular basis.—*Ed.*]

I've found the slight inconveniences posed by my HF mobile installation to be more than offset by the satisfaction of enjoying worldwide communication as I travel. As you can see, operating HF mobile isn't nearly as difficult as it looks. Try it yourself!

Roger Burch, WF4N, obtained his Novice license in 1982 and upgraded to Extra the same year. He is employed in the maintenance department of Big Rivers Electric, an electric generating utility. Roger owns and operates 2-meter and 70-cm repeaters in the Central City, Kentucky area. He's also active in QRP, computers and antenna design. Roger holds a General Radiotelephone license, a Journeyman Communication Certification and he recently accepted an appointment as an ARRL Technical Specialist. QST

New Books

THERE ARE NO ELECTRONS: ELECTRONICS FOR EARTHLINGS

By Kenn Amdahl. Published by Clearwater Publishing Co Inc, PO Box 1153, Arvada, CO 80001-1153, 322 pp, illus. Softcover, 8½ × 5½ inches. Second printing, Sep 1991. $12.95.

Reviewed By Brian Battles, WS1O
QST *Features Editor*

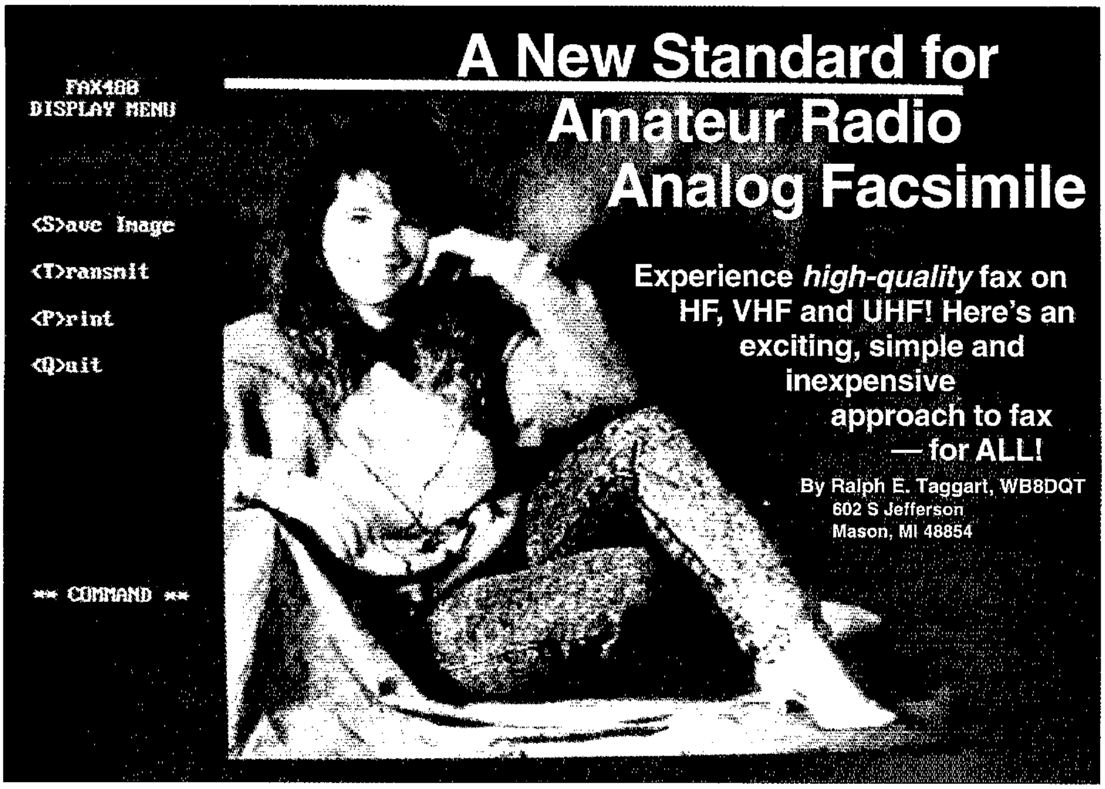
What do a green-skinned hippie, a hoary time-traveling wizard, waterskiing ducks, a beautiful nude Norwegian girl rowing a boat across a fjord and breakdancing chickens have to do with electronics? A lot, according to Kenn Amdahl, who claims to have inherited his knowledge of electronics from his father, Bernhard Amdahl, N0BCD, a telephone company engineer; his uncle John Amdahl, KH2BG, who taught electronics for the US Navy; his uncle Lowell Amdahl, WB6IRF, who repaired computers for IBM; and, of course, Eugene Amdahl (a second or third cousin of his father), founder of the Amdahl Corporation, which builds supercomputers. Kenn claims that his lack of a PhD in electronics—and total ignorance of the subject—is one of his main qualifications for writing about it. He makes a very good point.

If this was 1969 and hallucinogenic drugs were the rage, the reader might well assume that Amdahl has written most of this material while under the influence of certain pharmaceutical compounds.

Eschewing traditional discussions of physics and the usual dull, cranium-boggling miasma of jargon and cliches, Amdahl creates a surreal world of fantasy and adventure, rich with fables and allegories that make the way electricity works come to life. What he purposely lacks in scientific "why" he makes up for in allegorical "how." You won't soon forget that listening to Marvin Gaye singing *Heard it Through the Grapevine* is part of a powerful force that attracts positive "female" particles to negative "male" counterparts to produce the potential we've always been satisfied to simply call "voltage." Describing how series and parallel resistances work by comparing them to vehicles crossing bridges is a brilliant metaphor.

Amdahl takes you back and forth between simple parables and their translations into electronics theory in a roller coaster blur of hallucinogenic fantasy and friendly narrative. He confesses his ignorance of the subject matter even as he elucidates, and he delights in gently bashing those who are enamored of jargon and academic dogma.

The list goes on, but people who have always been hopelessly brainlocked trying to grasp the concepts of how electronics really works will find it clearly laid out via vivid, bizarre images in a loosely woven tale of adventure, sorcery and silliness. Don't let nonham friends or family members find this book in your library, though—if they read it, they'll find out that there's no big mystery to electronics and spoil their image of you as a master of a stupefying technical art.



A New Standard for Amateur Radio Analog Facsimile

Experience *high-quality* fax on HF, VHF and UHF! Here's an exciting, simple and inexpensive approach to fax — for ALL!

By Ralph E. Taggart, WB8DQT
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Mason, MI 48854

Facsimile (fax) is the oldest of the image-transmission technologies. It evolved early in this century as newspapers developed the means to print photographs on a routine basis. With the maturation of radiotelephony, it became possible to transmit and receive high-quality still images anywhere in the world, and press organizations created worldwide communications networks to ensure that photographs of major events appeared in newspapers with minimum delay. Sending a detailed picture over voice-grade cables or wireless links requires that the rate of image transmission be quite slow. Depending on the circuit and the amount of detail required, transmission times might range from 5 to 20 minutes or more. (For more information on fax, see the sidebar, "Some Fax Basics.")

What Have Amateurs Been Doing With Fax?

Amateur fax activity has been primarily limited to a spin-off of fax, slow-scan television (SSTV) and reception of weather-satellite images. Until only a few years ago, weather-satellite enthusiasts typically modified commercial fax recorders, or built their own machines from scratch. The era of mechanical systems has now faded into history. Most current weather-satellite work involves the use of computers and high-resolution graphics display systems. IBM PC, AT, PS/2 and PC-compatible computers equipped with VGA display systems are, by far, the most popular.

SSTV is an amateur innovation, pioneered in 1958 by Copthorne MacDonald (now VY2CM).[1] After a period of experimentation, SSTV activity settled on a standard using 120-line images, transmitted at a rate of 15 lines/second, for a total of 8 seconds per image. This method used an FM-modulated subcarrier, adopting standard fax frequencies of 1500 Hz for black and 2300 Hz for white. Because SSTV is based on cathode-ray-tube (CRT) technology, the fax standard was modified to include the use of a 1200-Hz subcarrier frequency as a sync pulse. The vertical sync pulse, to start each *frame* or image, is a 30-ms interval of the 1200-Hz subcarrier, while each *line* is triggered by a 5-ms horizontal sync pulse.

From Long-Persistence Phosphors . . .

The original SSTV image format was a trade-off between bandwidth considerations and the available analog-display technology. A 15-Hz line-transmission rate (about 1920 picture elements/second [or pixels/second]) is about the maximum data-transmission rate suitable for use with HF SSB equipment. Pictures were displayed on long-persistence (P7 phosphor) CRTs that would hold a useful image for about 8 seconds. If an image is transmitted at a rate of 15 lines/second for 8 seconds, the result is an image consisting of 120 lines. While relatively crude, the results are quite acceptable for close-up portrait views (Fig 1) and the short transmission time means that at least one or two frames would dodge the QRM if you sent each picture several times.

. . . to Digital Memory

The advent of digital scan converters—where the image is stored in digital memory and displayed on a TV monitor—did away with the limitations of the P7 CRT. As memory-chip capacity increased (and prices dropped), amateurs experimented with 120- and 128-line color images and 240- and 256-line higher-resolution monochrome (B&W) formats. By 1985, with the introduction of the Robot Research Model 1200C scan converter, 240-line full-color images became the norm. The 240-line pictures, in B&W or color, are a tremendous improvement over the older 120-line standard and approach the resolution you see when viewing videotapes (Fig 2). As computer graphics standards evolved and improved, various microcomputers such as the Amiga, and PC compatibles with VGA graphics adapters, became more prevalent as SSTV operating platforms. Even stations using dedicated scan converters—such as the Robot

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[1] Notes appear on page 36.

Table 1

Proposed Standards for the Transmission of 512 × 480 × 16 Images

Image aspect ratio: 1:1 (square).

Horizontal line rate: 3.7416 Hz.

Direction of horizontal scan: left to right.

Pixel display/sample rate: 1953.125/s.

Frame scanning lines: 480.

Direction of vertical scan: top to bottom.

Video modulation: audio FM subcarrier; sync, 1200 Hz; black, 1500 Hz; white, 2300 Hz.

Line synchronization: crystal reference and/or line-sync triggered; 5.12 ms of 1200-Hz line sync, followed by 262.144 ms of active video.

Frame start/synchronization: 5 seconds of 244-Hz modulation of the subcarrier between black and white limits (4 clock cycles of 1500 Hz followed by 4 clock cycles of 2300 Hz, repeated 1220 times), followed by 20 phasing lines consisting of 10 clock cycles of sync (1200-Hz subcarrier) followed by 512 clock cycles of white (2300-Hz subcarrier). Frame period: 2 minutes 18.3 seconds (10 seconds for start/phasing plus 480 active image lines).

1200C—employ computers to store and retrieve images, overlay call signs and otherwise manipulate the pictures.

Why Was Fax Sleeping?

Although SSTV has seen steady growth, there has been almost no activity in terms of *high-resolution fax* work. There are two significant reasons for this: First, there's no single, coherent fax standard. Surplus commercial fax machines vary considerably in terms of standards, and it's difficult to obtain spare parts and specialized supplies (such as the multitude of different recording papers) that these machines use. Secondly, there's the length of time required to transmit a fax picture. Depending on the system, anywhere from *5 to 20 minutes* are required to transmit and receive an image! For radio amateurs, because FCC regulations require periodic station identification, transmitting for more than 10 minutes without identifying is illegal. Even transmission times of 4 to 6 minutes are impractical, given the congested nature of our HF bands. With the heightened interest in amateur video communications, it's time we looked at alternatives for the transmission of more-detailed images.

A Computer-Based Fax Standard

Rather than tie amateur fax standards to outmoded 1930s mechanical technology, it seems reasonable to focus on available high-resolution *computer video standards* as a starting point. Now, IBM PCs and compatibles unquestionably have the largest installed user base. VGA graphics adapters and monitors have become the entry-level standard for such computers, and older units can be upgraded to VGA standards for as little as $50 for a VGA adapter card and about $100 for a monochrome monitor. Although VGA systems are backwardly compatible with the older monochrome, CGA and EGA standards, two VGA-specific modes have the greatest potential interest for amateur imaging:

Pixels/ Line	*Image Lines*	*Colors/ Gray Scale*
320	200	256
640	480	16

The 320 × 200 standard is widely used in systems designed for SSTV image display, although the 256-color limit demands creative programming to achieve acceptable color display. The 640 × 480 × 16 color/gray-scale mode provides an excellent option for a significant increase in monochrome image resolution. If the horizontal display is limited to 512 pixels, an area to one side of the image can be reserved for image data and menu options.

A 512 × 480 image (see the lead photograph and Figs 3 and 6) represents the same increase in resolution over the 240-line SSTV standard as the 240-line pictures do in relation to the original 120-line format. The increase in image detail is *striking* and greatly expands the range of options for pictures and subjects.

If you've purchased a computer in the last two years, you probably already have all the computer hardware you need. All that's required to put high-resolution fax pictures on the air is a transmission standard that takes advantage of the many options already developed for SSTV.

Fig 1—Here's my daughter, Jennifer, in a portrait view using the original 120-line SSTV display standard. The relatively low resolution of this format is a result of the trade-off between signal bandwidth and the limitations of the analog equipment used to display pictures until the mid-70s.

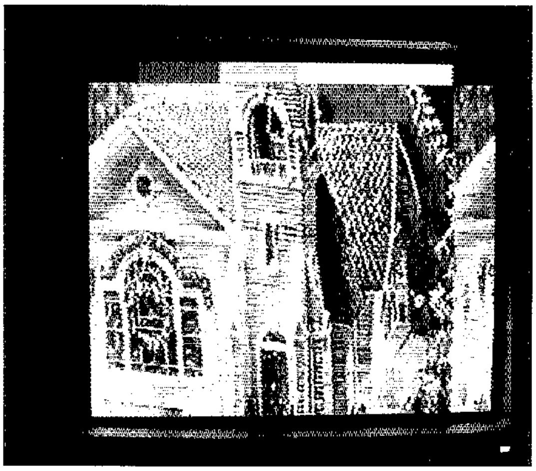


Fig 2—Digital scan converters facilitated the development of 240-line SSTV formats as illustrated here. This sample was displayed from a 20-meter transmission by K4ZET using the Scottie 1 line-sequential color mode. The SSTV routines incorporated into the *FAX480* program were used to display this image. The increase in effective resolution over that achieved with 120-line images is dramatic and expands the possibilities for exchanging detailed pictures.

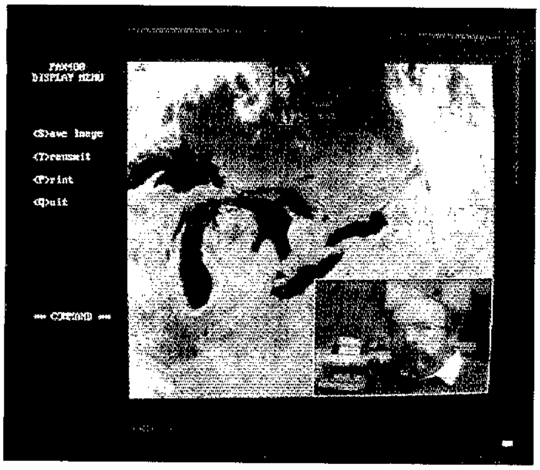


Fig 3—Here's an example of a test transmission of the 480-line fax format on 70-cm FM. The significant increase in image resolution provides even more opportunities for possible picture sources. Image transmission using the proposed fax standard requires 2 minutes and 18 seconds. This image was created from a 480-line VGA weather-satellite image file, with the inset derived from a 240-line TV camera frame-grabber.

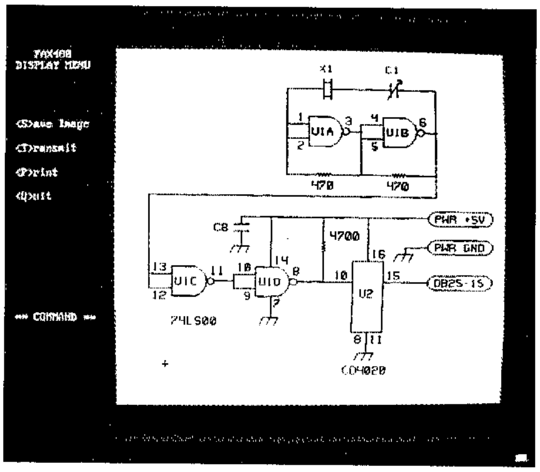


Fig 4—The original schematic of the simple clock circuit. This isn't in *QST* style. That's because it's an author's original that was scanned into a computer file, transmitted and received on 70-cm high-resolution fax, then photographed off the screen! The circuit uses a 4-MHz microprocessor crystal. Resistors are 1/4-W, 5%-tolerance carbon-composition or -film units. U1 is the oscillator/buffer, followed by U2, a CMOS counter, providing a divide-by-2048 function. The GND and +5-V leads are connected to points on the bottom of the ViewPort VGA PC board at the dc power connector; the lead to pin 15 (**ERROR**) of the parallel connector is connected beneath the ViewPort board near the DB25F I/O connector.

Video Modulation Standards

Table 1 lists the specifications for a VGA-based image-transmission format. The standard fax and SSTV audio FM modulation format (1500 Hz black to 2300 Hz white) is retained, enabling use of existing HF fax and SSTV modems.

Line and Frame Rates

The standard specifies a 3.7416 lines/second horizontal rate. This fractional value may seem odd, but is a result of bandwidth and system clock factors. The present HF SSTV standards transmit pixels at a rate of 1920/second. Transmitting pixels at a significantly slower rate results in unnecessary extension of the image-transmission time. A significant increase in the pixel rate requires a wider video bandwidth and degrades resolution because of the audio and RF filtering used in modern HF SSB equipment. A pixel rate roughly comparable to the SSTV rate is the most practical alternative.

In the next section, I'll emphasize the benefits of maintaining very precise rates for image transmission. This can't be achieved by relying entirely on the computer's internal clock and/or software timing loops. A simple crystal clock circuit that can be trimmed to a precise frequency provides the requisite accuracy.

Tick! Tock!

We can take two approaches to the system clock. If we want a nice, even-numbered line rate, we'll need a custom-ground crystal whose odd frequency will make it difficult to adjust without a very accurate frequency counter. Alternatively, we can use a universally available crystal (cheap!) and accept whatever line rate that produces. Fig 4 shows a simple and frugal clock circuit using a 4.0-MHz microprocessor crystal. U1 is the oscillator/buffer, followed by U2, a CMOS counter, providing a divide-by-2048 function. Assuming the 4-MHz oscillator is precisely on frequency, the clock output is 1953.125 Hz. If a well-calibrated frequency counter is available, connect it to TP1 and adjust C1 (use a nonmetallic tuning tool) for a reading of precisely 4 MHz. Alternatively, set the crystal calibrator of your receiver for zero beat with WWV at 10 MHz. Connect a short test lead to TP1, set the receiver to 4 MHz and adjust C1 for zero beat with the calibrator harmonic.

The system software uses this clock to time the pixel transmission and display, so the resulting pixel rate is 1953.125 pixels/second. Each clock cycle requires 0.512 ms; that sets all the other system rates. Each line is preceded by 10 clock cycles of 1200-Hz subcarrier sync (a total of 5.12 ms), followed by 512 image pixels, requiring an additional 262.144 ms (512 × 0.512). Thus, the total time required for line transmission is 267.264 ms (5.12 + 262.144). Because each line requires 267.264 ms, the line rate works out to 3.7416 lines/second (1000 ms ÷ 267.264). The image frame consists of 480 lines, so the time required to transmit the image data is 2 minutes, 8.3 seconds (480 lines ÷ 3.7416 lines/second = 128.3 seconds). We need to add an additional 10 seconds at the start of the picture to ensure reliable starting (to be discussed shortly), so the total time required for transmission of a complete picture is 2 minutes, 18.3 seconds.

Synchronization

SSTV image lines are triggered individually by the presence of a 5-ms, 1200-Hz sync pulse tone at the start of each line. Because the sync pulses maintain line registration at the start of each line, SSTV timing is relatively noncritical to received image display. SSTV images can be recorded on standard monaural audio-tape systems of modest quality and still retain acceptable image registration. The major drawback to line triggering is that interference or noise bursts can disrupt line triggering and thus destroy an entire line of image data.

Fax systems don't use sync pulses; they rely on precision speed control at the transmitting and receiving ends to maintain synchronization. This makes line registration independent of noise or interference; only individual pixels are disrupted on a noisy path. The main problem with such an approach is that it's best implemented live. Recording such a transmission either requires the use of a precision recording system (a VCR locked to a local TV station or a digital audio-tape system), or the use of a stereo audio recorder with crystal-referenced tones and phase-locked loops to track inevitable speed variations in the audio recorder.

The proposed standard takes advantage of the positive aspects of both the line-triggered and crystal-locked approaches. All image transmissions are made using the

1953.125-Hz master clock frequency as a standard. The image is transmitted with the same precision as a fax signal, but conventional 5-ms, 1200-Hz SSTV horizontal sync pulses are inserted at the beginning of each image line. Thus, two modes of operation are possible. Display of the signal, using the crystal clock, begins with the detection of the image-start routine (to be discussed). The system then looks for the first 5.12-ms, 1200-Hz line sync pulse (10 clock cycles). At that point, the system displays the 480 image lines entirely by reference to the crystal-controlled clock. Noise or interference impacts *individual image pixels only!*

Because the system also transmits line sync pulses, the signal can be recorded on a standard *monaural* cassette recorder. A second display routine, which loads lines after receipt of each line sync pulse, can be used to display the image from the tape recorder. Once loaded, the image can then be transmitted by reference to the system clock. The result is a very flexible system with the precision of conventional fax, but with the audio-tape convenience of SSTV!

Image Start

When I first began work on the 480-line fax format, I initiated each frame with a conventional 30-ms, 1200-Hz SSTV vertical sync pulse. This worked well all the way through the early tests on UHF FM, but was less than satisfactory when the typical 20-meter bedlam was mixed with the signal while I was experimenting on the test bench. Commercial HF fax links place a premium on reliable image starts. If their customer's machines are subject to false starts, the customer wastes money on fax paper and loses the opportunity to receive a possibly important image. For an amateur fax standard, if we're going to commit slightly over 2 minutes to the transmission of a picture, we want starting to be as reliable as possible to make optimum use of the frequency.

HF and satellite data links solve the starting problem by transmitting several seconds (typically about 5) of a distinctive *start tone.* The start tone consists of square-wave modulation of the subcarrier between black-and-white limits (often at a 300-Hz rate). This start tone is quite distinctive and can be reliably detected in a number of ways, triggering the recorder at the receiving end. The start tone is then followed by a *phasing interval* of about 5 seconds duration. During this period, black pulses are transmitted to mark the start of image lines, with the rest of each line consisting of white subcarrier. The black-to-white transitions are easily detected and the phase interval permits the recorder at the receiving end to get into step (phase) with the fax transmitter. At that point, actual transmission of image data begins.

My experience with weather satellites had demonstrated how reliable such a system can be, so I incorporated it into the 480-line standard. The start tone for this format involves 5 seconds of a 244-Hz square wave modulating the subcarrier between black and white limits. This is achieved by setting the transmitted subcarrier to white (2300 Hz) for 4 clock cycles (2.05 ms), then setting it to black (1500 Hz) for 4 more clock cycles. If this is repeated 1220 times, the result is 5 seconds of start tone (1220 ÷ 244 = 5). At the receiving end, the start tone is detected by sampling the video on every clock transition for a total of 488 samples (¼ second). The system tallies the number of black-to-white transitions in the ¼-second interval. If the start tone is present, there should be 61 transitions (244 ÷ 4). The system accepts a value between 59 and 62 (allowing for sampling error), assuring stringent requirements for initiating image display. The use of a 5-second start tone interval allows the receiving operator ample time to initiate display when the start tone is heard, further reducing the possibility of false starts.[2]

Some Fax Basics

Until the recent advent of satellite-based digital systems, the basics of fax technology were quite stable for over 50 years. Fax transmitters and receivers (referred to as *recorders*) are based on a marvelous combination of intricate mechanics and basic electronics. Image scanning at the transmitter and reconstruction at the receiving end is controlled by synchronous motors which maintain constant and equal speeds at both ends of the fax circuit. This is achieved by driving the motors from crystal or tuning-fork frequency standards, in conjunction with precision gear assemblies. The operating speed of these units (the *line rate*) is not easily altered and multimode units are rare. Units designed for press Wirephoto service typically operated at 60, 90 and 180 lines/minute (lpm). Weather map and chart systems operate at 120 lpm, while weather-satellite fax recorders use line rates of 120 and/or 240 lpm.

In addition to variations in line rate and other aspects of image format, fax systems also differ in video modulation. All use some form of modulation of an audio tone or *subcarrier.* Weather-satellite systems and some systems designed for phone-line service amplitude-modulate the subcarrier. In weather-satellite systems, for example, black image data is represented by a subcarrier amplitude of 4%, while white is represented by 100% of peak subcarrier amplitude. Amplitude subcarrier modulation does not work well over HF radio links where fading is common. Weather-chart and Wirephoto systems on HF use FM subcarrier modulation in which black is represented by a 1500-Hz tone and white by a 2300-Hz tone. There is also considerable variation in how the various systems signal the start of an image and how they accomplish *phasing*—assuring that the start of each line at the receiving end corresponds to the start of a line at the transmitter.—*WB8DQT*

The start tone is followed by about 5 seconds of phasing interval, representing 20 nonimage lines. Each of these lines begins with 10 clock cycles of 1200-Hz subcarrier (5.12 ms), followed by 512 clock cycles of white (2300-Hz) subcarrier. This makes it very easy for the system to detect that starting time for the forthcoming image lines, at which point display can begin using either the crystal-clock reference, or triggering using the line sync pulses. With the software to be described shortly, about 10 of these white lines are displayed at the top of the picture if the start tone was detected immediately. This provides a buffer of about 3 seconds for start-tone detection under noisy conditions, without loss of image data.

Hardware

One of the problems associated with attempting to introduce a new communications standard is the vision of all the new hardware required to give it a try. This is *not* the case with this computer-based fax standard! Most interfaces designed to receive and transmit SSTV using a PC-compatible computer can be used to transmit and receive images using this standard. One such unit is the ViewPort VGA SSTV interface.[3,4]

A&A Engineering's ViewPort VGA is an excellent general-purpose SSTV interface that requires the simple addition of the clock circuit shown in Fig 4 to be compatible with the 480-line fax mode. The clock circuit, built on a small piece of perf or PC board (see Note 3), can be installed in the ViewPort VGA (see Fig 5), which supplies the needed 5 V dc. The interface connects to the host computer via the parallel printer port.[5] The interface uses almost all of the I/O capacity inherent in the PC's parallel port, with the exception of a single input bit available at pin 15 (the **ERROR** bit) of the DB25 connector on the ViewPort VGA rear apron. A single wire, connecting the output of the added clock circuit to this pin, is all that's required to put the system into fax service. The added clock circuit *has no effect* on the use of the interface for SSTV.

Software

The software required to implement 480-line fax is *hardware-specific.* To encourage experimentation in this mode, I've written a PC-compatible program, *FAX480,* which is being made available free of charge on a number of bulletin board systems. (A&A Engineering also ships the software with ViewPort VGA kits and assembled and tested units; see Note 3.) The software can be found on the ARRL's BBS (203-666-0578), the Dallas Remote Imaging Group (DRIG) BBS (214-394-7325) and on my Weather Satellite Handbook BBS (517-676-0368). Experimenters may copy and distribute the software to others, as long as no fee is charged and commercial use is not made of the program code. The software is distrib-

Table 2

***FAX480* Software Functions**

- Receive live (directly from the computer) images using the internal crystal clock.
- Display audio-tape-recorded images using frame- and line-triggering provided by the 1200-Hz sync pulses.
- Transmit images from the RAM image buffer.
- Display, on demand, the current image in the RAM buffer.
- Save to disk and load from disk any image (128 kbyte total file size).
- Display a 16-step gray scale to optimize adjustment of the VGA monitor's brightness and contrast controls.
- Add a one-line (60-character) label at the bottom of the image. This label can identify the image and/or provide information relevant to a particular contact.
- Print fax images on LaserJet and DeskJet printers.
- An SSTV operating module that lets you receive the most-popular SSTV modes and transmit both low and "high"-resolution SSTV images from within the *FAX480* program.
- Routines to permit precise receiver tuning prior to picture transmission.

uted in an archived file (*FAX480.ZIP*), which contains the following files:

FAX480.EXE—the operating program to control the A&A ViewPort VGA interface.

FAX480.DOC—an ASCII documentation file describing the use of the *FAX480* program.

BINFAX.EXE—a utility file to convert VGA-screen-captured files to the *FAX480* image-file format.

FAXSSTV.EXE—a utility to convert *FAX480*-format images for use in the SSTV portion of the program.

Sample images—typically one 480-line fax image and several SSTV-format images are included in the archived distribution file.

With this software, you can operate two-way 480-line fax without any ancillary equipment other than the modified A&A ViewPort interface. The computer must be equipped with a VGA display adapter, a minimum of 640 kilobytes of RAM and a parallel printer port. There's no installation program: The software *automatically* determines the parallel port to which the interface is connected and alerts you if you've failed to connect the interface or apply power to it! The program can run from a floppy disk of virtually any capacity, or a hard disk drive. See Table 2 for a brief description of the program's features. The *FAX480* program and A&A interface were used for all the on-the-air tests. This 480-line fax capability can be incorporated into other PC-based SSTV systems; I encourage developers to do so. Anyone experienced in programming SSTV routines will have no trouble, as sync and video handling is identical to that of SSTV except for reference to the clock-pulse train for timing.

Image Sources

Untold high-resolution images are available from BBSes and other sources. These images can be easily translated to a *FAX480*-compatible format. Terminate-and-stay-resident (TSR) software is available that grabs a 640 × 480 VGA screen and converts the image to a large binary file (307 kbytes). *VGACAP* (V 6.1) by Lawrence and Marvin Gozum, a shareware program available on the DRIG bulletin board,[6] is an example of such software. The *BINFAX.EXE* utility program supplied in the *FAX480.ZIP* archived file converts such binary files to the disk format used by the *FAX480* program.

Hand-held and flatbed scanners can also be use to digitize photos and other material. Most of these support a 640 × 480 VGA display option. If these programs don't directly generate the needed binary files, the screen-grab approach mentioned earlier can be used. Hand-held scanners represent one of the most cost-effective ways to produce images for fax and SSTV. They're particularly useful in computer-based systems that may lack TV frame-grabbing hardware. Almost all the images in this article were scanned from standard color prints using a Logitech 256 hand scanner. Images scanned at 100 dots per inch (dpi) equal or exceed the resolution obtainable with the 512 × 480 fax format. I saved the scanned images as .PCX files; *VGACAP* created a binary image file and *BINFAX* converted the binary file to the *FAX480* format.

TV frame grabbers can be used if they support the needed resolution. Most inexpensive frame grabbers produce 240- or 256-line images. In that case, you might as well rely on 240-line SSTV, as there's no advantage to using a higher-resolution format. A 240-line frame-grabber can be used to cap-

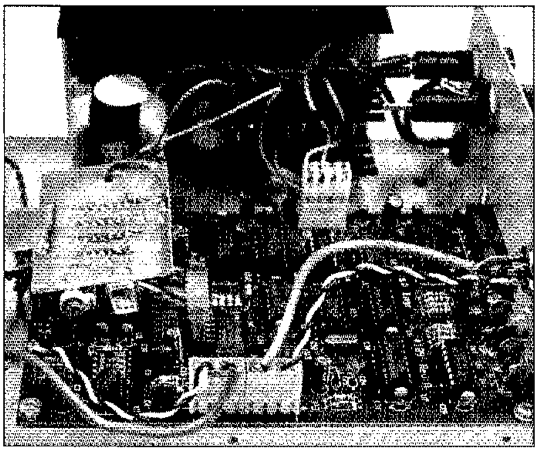


Fig 5—An inside view of the A&A Engineering ViewPort VGA SSTV interface with the two-IC clock circuit installed. The ground foil of the clock's PC board is soldered to a strip of spring brass held in position by the MIKE GAIN pot mounting hardware. Other mounting methods—such as double-stick tape—can be used. (P2, the unused connector near the power supply board accommodates the interface's optional tuning indicator.) Installation takes but a few minutes as only three leads need be connected to the interface's PC board. The circuit doesn't affect the interface's SSTV operation, and provides the precision timing required for fax transmission and reception.

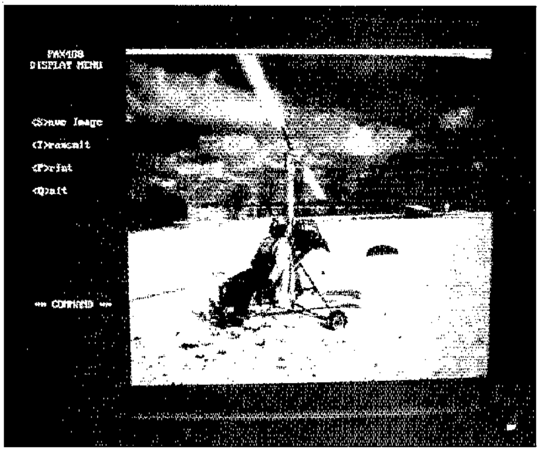


Fig 6—My flying partner, Don Chubb, spins the blades of our ultralight gyroplane in this 512 × 480 image. This picture and the title-page photo, were transmitted to W8SH on 20 meters by W1AW using a transmitter power output of 100 watts fed to a three-element Yagi at 120 feet. At W8SH, I used the live (crystal-locked) display program option. The white bar at the top of the picture is the remnant of the phasing signal and characterizes pictures transmitted by the fax standard proposed in this article. This picture is far too detailed for 120-line SSTV and would be marginal for 240-line transmission. The 480-line fax format provides the resolution required to create a sharp picture.

ture an image and insert it into a larger fax image (see Fig 3), or several 240-line pictures can be combined to make one larger fax image. Better (and more expensive) frame grabbers integrate *both* fields of an interlaced TV picture, capturing a 512-line image. Such systems can be used, along with a TSR screen-capture program, to produce the binary files, which can then be converted by the *BINFAX* program for use by the *FAX480* software.

On the Air

Prior to testing the system on HF, I made a number of contacts on VHF and UHF to get a feel for the operational aspects of the system. Four images were recorded on the audio channel of an 8-mm videotape deck, with the video input locked to a local TV signal. This provided me with a recorded signal that could be transmitted with essentially the same timing precision as a live signal. Jeff Broughton, WB8RJY, one of our local ATV crowd, took custody of the VCR and made provisions for tapping the audio signal into one of his 70-cm FM H-Ts. At the receiving end, I displayed the signal using the crystal-locked mode, simultaneously making audio recordings to test the line-triggered display mode.

The tests went off without a hitch. The image in Fig 3 is an example of pictures transmitted in this sequence. Clearly the system was working well over any voice link that could be maintained on VHF and UHF FM. If nothing else, the tests demonstrated the capability of transmitting very detailed images over paths that would be difficult or impossible using the conventional ATV approach.

HF tests of this fax system were conducted on 20 meters between W1AW in Newington, Connecticut, and W8SH, the Michigan State University Amateur Radio Club station in East Lansing, Michigan. In effect, the tests celebrated a 25-year reunion. (ARRL Executive Vice-President David Sumner, K1ZZ, and I were responsible for putting W8SH on SSTV in 1967, and collaborated on a number of pioneering experiments over the next few years.) There is no comparison between the current 480-line images and the 120-line pictures we worked with so many years ago, but the sense of adventure and experimentation was as fresh as ever!

Both stations were equipped with modified A&A Engineering ViewPort VGA units, standard VGA PC-compatible computers and the *FAX480* software. Last-minute problems with interfacing to the transmitter kept W8SH from transmitting, so W1AW assumed that role, operating at 100 watts. Three *virtually perfect* pictures were received in Michigan (see the lead photo and Fig 6). Only one picture had to be repeated, and that was because *I* missed the start tone! The W1AW crew, who had never used the software prior to the tests, performed without a single slip. After all the endless, early-morning hours, consuming gallons of coffee while debugging computer code and hardware, the transition from concept to reality was almost anticlimactic! The results are excellent—and the system is easy to use. You can't ask much more of a mode that is both simple and inexpensive to implement!

The SSTV Bonus

One of the problems with being a fax pioneer is that you're initially going to run into more SSTV stations than you will those equipped for 480-line fax. As an added inducement for you to give the program a try, I've included a very effective SSTV module within the *FAX480* program to ensure that you don't get lonely while searching for other fax stations! All the SSTV options use 64 gray-scale steps for transmission and reception, and you can copy *all* of the standard monochrome and Robot color SSTV formats as well as the popular Scottie 1 (Fig 2) and 2 modes. If the interest is there, I'll add other modes in the future.

Any images you receive can be retransmitted in either the 120- or 240-line format. In addition, the program allows you to transmit any of your 480-line fax image files in either SSTV format. All SSTV images use compact 64-kbkyte files that save to disk and load from disk almost instantaneously. In addition, the SSTV module incorporates a *contrast-expansion* routine to improve pictures washed out as a result of less-than-optimum setup at the transmitter end, or slight mistuning at the receiving end.

Of course, if you have a color VGA display, you can use the software supplied with the ViewPort VGA interface to transmit and display color SSTV images. You'll find that program is a bit more complex to use and will probably require some setup. If you're using a monochrome VGA monitor, the image quality is better if you use the *FAX480* program. *FAX480* is also considerably easier to use for routine SSTV work if you're content with a high-quality, gray-scale display.

Summary

ATV and SSTV are two healthy modes that provide a useful range of image-communications options. Unfortunately, a number of factors combined to stifle the development of amateur fax capability and (until now) there hasn't been a good option for handling higher-resolution imagery. This proposed 480-line fax mode, based on the nearly universal 640 × 480 VGA display, provides such an option. The standard is completely compatible with existing SSTV hardware for PCs and compatibles. Software, in the form of the *FAX480* program, is available free for a popular SSTV interface unit. Modifying the interface for the fax standard doesn't alter its normal SSTV capabilities in any way: It simply opens new possibilities. My intention is not to attempt to displace any of the existing SSTV imaging modes; each has its place in the continuum of amateur image communications. What has been missing is a coherent standard for handling images of *higher resolution.* Hopefully, this proposed standard, which can be implemented very economically, will encourage experimentation with the exchange of higher-resolution images and provide amateurs with a viable facsimile option.

Notes

[1] For more information on SSTV and Copthorne MacDonald, see J. Langner, "SSTV—It isn't Expensive Anymore!," *QST*, Jan 1993, pp 20-30, and R. Booth, "A Profile of Copthorne MacDonald," *QST*, Jan 1993, p 31.

[2] The start-tone system was tested by initiating the image-display routine while driving the system with a tape recording of about 45 minutes of the weekly SSTV net on 20 meters. This tape represented SSTV image transmissions, chatter, interference and everything else that can possibly occur on 20 meters on a Saturday afternoon. The time between false starts averaged about 5 minutes. Because you normally would initiate display just moments before the other station began transmitting, I consider the result a highly reliable approach to triggering the display.

[3] See Product Review in this issue. The ViewPort VGA system 4-MHz clock and accompanying software is available from A&A Engineering, 2521 W La Palma Ave, Unit K, Anaheim, CA 92801, tel 714-952-2114, fax 714-952-3280.

A PC-board template package is available free from the ARRL. Address your request for the TAGGART CLOCK TEMPLATE to: Technical Department Secretary, ARRL, 225 Main St, Newington, CT 06111. Please be sure to enclose a business-size SASE.

[4] See also pages 15 to 17 of the May 1988 issue of *QST*, in which we published "Pictures by Packet," written by Corby B. Pratt, K4WV, and Virgil L. Yarbrough, K4IEK. This system uses a C64 computer.—*Ed.*

[5] You'll need a DB25 extension cable (25-pins connected straight through) with a DB25M (male) connector on each end to connect the interference to the computer's parallel port. —*Ed.*

[6] Dallas Remote Imaging Group (DRIG) BBS, tel 214-394-7325.

Ralph Taggart was first licensed as WV2EMC in the late 1950s and now holds an Advanced-class license. He earned a BA in Biology from Rutgers University (1963), an MS in Botany from Ohio University (1967), and a PhD in Botany from Michigan State University in 1971. Since that time he has been on the faculty at MSU and is a Professor in the department of Botany and Plant Pathology and the department of Geological Sciences, and is Curator of Fossil Plants.

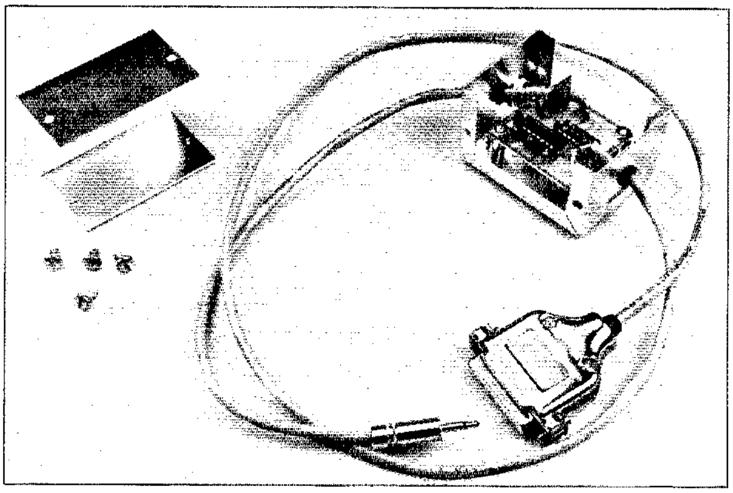
Ralph's been active in amateur television (ATV) since 1963 and slow-scan television since 1967. In 1972 he was coauthor (with Don Miller, W9NTP) of the Slow Scan Television Handbook. *Not long after that, he became interested in weather-satellite image reception display and authored the first edition of his* Weather Satellite Handbook *in 1976. This book, now in its fourth edition, is published by the ARRL.*

In addition to his ongoing work in all aspects of image communications, Ralph is active in ultralight aviation, flying a homemade gyroplane of his own design. Ralph has authored many articles for Amateur Radio and various aspects of ultralight aviation, a book chapter on ultralight ski flying, numerous technical papers in his research area and is coauthor of two general biology textbooks.

Everything You Always Wanted to Know About Hardware for Computer-Controlling Modern Radios

Want to interface your ICOM, Kenwood, Ten-Tec or Yaesu transceiver to your computer? Here are the basics—and easy-to-build interface circuits for each!

By Wallace R. Blackburn, AA8DX
1346 Erickson Ave
Columbus, OH 43227



Almost every modern MF/HF Amateur Radio transceiver has provisions for external control by a computer or terminal. Most hams take advantage of this feature by using software intended exclusively for radio control, or primarily intended for some other purpose (such as contest logging), with rig control as a secondary function.

A number of recent, informative articles have been published describing hardware for interfacing radios to computers.[1,2,3] But none of them have covered interfacing basics and circuits that work with *all* modern computer-controllable radios. That's what I'll do here. You can easily build these interfaces from readily available parts. PC boards, templates and parts kits are available to make things easy.[4]

Why Do I Need an Interface?

You may ask, "Why can't I simply connect the radio to the computer, using the appropriate connectors at each end?" The trouble is, almost all *computers* use RS-232-C levels for controlling external devices, and just about every *radio* requires TTL control signals. The two systems aren't compatible—they use different voltage levels. See the sidebar, "TTL and RS-232-C: What's the Difference?"

The interface simply converts the transistor-transistor logic (TTL) levels used by most radios to RS-232-C levels most computers use, and vice versa. Interfaces of this type are sometimes referred to as *level shifters.*

Interconnecting the Hardware

Commercial interfaces are available from radio manufacturers, at rather high prices. As mentioned earlier, recent articles have described simple and inexpensive interfaces. Some of these fail to fully implement the radio manufacturers' specifications; others don't provide adequate buffering between the radio and computer. Here are the details of interconnecting modern radios to computers, one radio manufacturer at a time.

ICOM Radios and the Ten-Tec Argonaut II, Delta II and Omni VI

The system ICOM uses in its CI-V interface—and that Ten-Tec uses in the Argonaut II, Delta II and Omni VI—is quite interesting. And it's the easiest system for which to build an interface!

These systems use a CSMA/CD (carrier-sense multiple access/collision detect) bus. This long name refers to a bus that a number of stations share to transmit and receive data. In effect, the bus is a single wire and common ground that can be used to interconnect a number of radios and computers. See Fig 1.

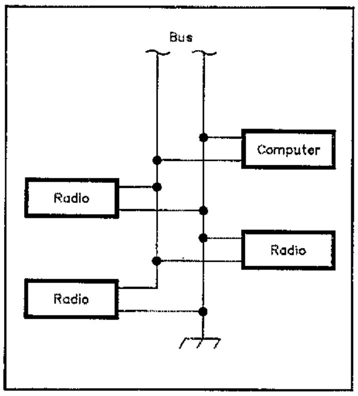


Fig 1—The basic two-wire bus system that ICOM and newer Ten-Tec radios share among several radios and computers. In its simplest form, the bus would include only one radio and one computer.

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[1] Notes appear on page 41.

TTL and RS-232-C: What's the Difference?

The term *TTL* usually refers to a type of integrated circuit. The popular 7400 series of digital logic ICs are TTL. Recent editions of *The ARRL Handbook* provide an excellent description of these devices.

For our purposes, what's important is that the term *TTL* is also used to refer to the *interface attributes* of these devices. In this context, TTL simply means that +5 V represents a binary or logical one, and 0 V indicates a logical zero. More exactly, any voltage equal to or greater than +2.4 V is a one and a voltage of +0.4 V or less is a zero. The region between +0.4 V and +2.4 V is undefined. See Fig A. TTL circuits can handle voltage maxima of 5.25 and -0.5. Most computer-controllable Amateur Radio transceivers and receivers use TTL control signals.

It's important to note that even though +5 V is a logical one, this level does not necessarily indicate the device's *active* level. *Active-low* logic denotes a resting signal of +5 V, which is pulled low (to less than 0.4 V) to represent a logical one. The advantage of active-low logic is better noise immunity. If the line is typically at +5 V and a noise spike of a volt or two occurs, it goes unnoticed. If the line is resting at 0 V and a spike of 2.4 V or more appears on the line, the device will probably see it as an erroneous logical one. A signal line that is active (or *asserted*) when low is represented by a bar over the signal label on schematic diagrams and chip pinouts.

Useful only for short distances (a few feet or so), TTL is typically used only within the confines of PC boards and pieces of equipment. As distance increases, TTL signals are affected by noise, reflections in the line, and the like. Another scheme is necessary to send binary data over longer distances.

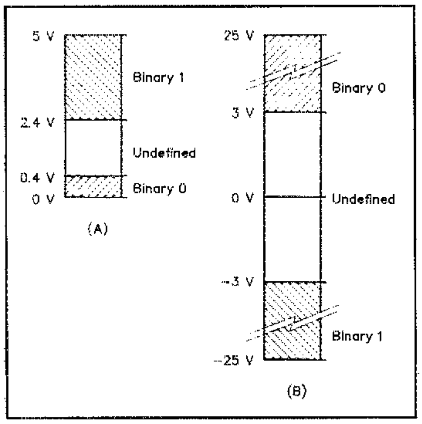


Fig A—At A, logic levels in a TTL system. Most modern transceivers with computer-control capability require TTL control signals. At B, RS-232-C logic levels, used by most computers and data terminals.

RS-232-C

The RS-232-C standard (slightly refined and renamed EIA-232-D in its latest form) is a system used to send binary data over relatively long distances (50 feet or more). This standard also specifies the physical connection and signal lines. The serial ports on most personal computers comply with the RS-232-C standard. These ports are typically used for modems, radios and other accessories.

To allow data signals to travel longer distances and provide better noise immunity, RS-232-C lines use a wider range of signal voltages. A logical one (or *mark*) voltage is negative, and the logical zero (or *space*) voltage is positive. Specifically, -3 V or less is a mark and +3 V or greater is a space. As with TTL, the area between these voltages is undefined. RS-232-C systems nominally use ±12 V or so for their logic signals and have absolute limits of ±25 V.

RS-232-C specifies a 25-pin D connector (DB25) for the physical connection. The individual pins and their purposes are also specified. Some are unused, some are rarely used, and a few are always used. We'll discuss just the most important ones, from a radio-interfacing standpoint.

To understand RS-232-C signals, it is important to know that equipment connected via RS-232-C is usually classified in one of two ways: *data terminal equipment* (DTE) and *data communication equipment* (DCE). Terminals and computers are examples of DTE, while modems and TNCs are DCE. (Radios don't fit comfortably in either category!) The signal names are given as they apply to the DTE.

Serial data is transmitted and received over pins 2 and 3 of a DB25. Pin 2 is known as *TxD* (transmit data) and is used to send data from the DTE to the DCE. Pin 3 is referred to as *RxD* (receive data) and transports data from the DCE to the DTE. Since the signal names apply to the DTE, pin 2 is where the DTE transmits and pin 3 is where it receives. Pin 7 is the *signal* ground and pin 1 is the *shield* ground. These two grounds are often tied together.

Two other pins, 4 and 5, are sometimes used and sometimes not. These are known as *RTS* (request to send) and *CTS* (clear to send), respectively. These are known as *handshaking* signals, which are sometimes used for data *flow control*. They provide a way for one piece of the connected equipment to tell the other to hold off sending data. This might allow one side to "catch its breath" or do some other processing.

When used, the RTS line is used by the DTE to indicate to the DCE that it has data to send. The DCE uses CTS to signal the DTE that it is ready to receive data. Most software simply brings the RTS line active at startup and leaves it there while ignoring CTS. Some more sophisticated software uses it as it's intended to be used, however. Two other handshaking signals, *DSR* (data set ready) and *DTR* (data terminal ready), are assigned to pins 6 and 20, respectively. DSR and DTR are usually used simply to indicate the equipment on the other side is powered up and active. Even though none of the current computer-controllable radios use DSR or DTR, we still need to pay attention to them at the computer end. For the Kenwood interface we tie DSR to CTS—a common practice. For the other interfaces we will simply tie DSR and DTR together at the PC.

On the TxD and RxD lines, a data bit of one, transmitted or received, is represented by a −12-V signal for a specific period of time that depends on the data rate at which the equipment is communicating. A zero is +12 V of the same length. An idle line may rest at binary zero (+12 V) or in the undefined state between the two levels.

The RTS and CTS lines work somewhat differently. These are active when +12 V is on the line. This may seem backwards (and a little confusing) unless you think about it a different way. When CTS is active, it means that the device on the other end is ready to accept data. Most of the time, this is the state of the device. By holding +12 V on the line, the device is saying, "Okay, keep the data coming." When the device needs to hold off accepting data for some reason, the line is *dropped* by putting −12 V on it. This allows the device to catch its breath or do some other processing. When a radio does not implement RTS and CTS, we simply tie them together at the computer end, as with DSR and DTR. As you look over the interface schematics and your radio's documentation, you'll see why each of the interfaces differs slightly in these respects.

—AA8DX

The single wire is used for transmitting *and* receiving data. Each device connected to the bus has its own digital *address*. (Each radio comes programmed for a different default address; you can change this address, usually by resetting internal DIP switches.)

Information is sent on the bus in the form of packets that include the control data and the address of the device (radio or computer) for which they are intended. All computers and radios on the bus receive each packet. Each device decodes the address in the packet and acts on only the data intended for it, ignoring packets intended for other devices.

Each device listens to the bus to make sure it's not in use before transmitting. A problem occurs when two devices transmit on the bus at the same time: They both listen, hear nothing, and start to send. When this happens, the packets garble each other. This is known as a *collision*. That's where the CSMA/CD bus collision-detection feature comes in: The devices detect the collision and each sender waits a random amount of time before sending the data again (listening first, of course). The sender waiting the shorter random time goes first.

Connecting transmit- and receive-data lines together on an active-high bus makes buffering similar to what the Yaesu interface requires (as discussed later). It's important to ground the interface, radio, and computer for maximum safety and noise immunity.

The schematic for the ICOM/Ten-Tec interface is shown in Fig 2. It's also the Yaesu interface. The only difference is that TxD and RxD are tied together for the ICOM/Ten-Tec version, as shown in the inset. The PC board is also the same—a jumper connects the TxD/RxD for ICOM/Ten-Tec use.

The signal lines are active-high TTL. This means that a logical one is represented by a binary one (+5 V). To shift this to RS-232-C, we need to convert the TTL +5 V to the RS-232-C's −12 V, and vice-versa. In the same manner, the TTL 0 V converts to an RS-232-C +12 V and vice-versa. As we'll see, this is quite simple with modern devices.

To meet the radio's interface specifications and to provide some buffering and isolation, the circuit uses a buffer/driver (U1). This provides buffering that some other circuits lack (including some sold by radio manufacturers!).

U2 is a 5-V-powered RS-232-C transmitter/receiver chip that has made interfacing TTL and RS-232-C much easier than in the past. Previously, interface chips required ±12 V supplies. This device, however, uses *charge pumps* to produce approximately ±10 V from a +5-V supply. Eliminating the need for the negative supply makes things much easier. We'll use this device in all three interfaces.[5]

Along with the RS-232-C standard DB25 female (DB25F) connector for the PC, the cable simply uses a ¼-inch phone plug at the radio end. The sleeve is ground and the tip is the bus connection. If you want to connect multiple radios, connect a number of the plugs in parallel with the one shown.

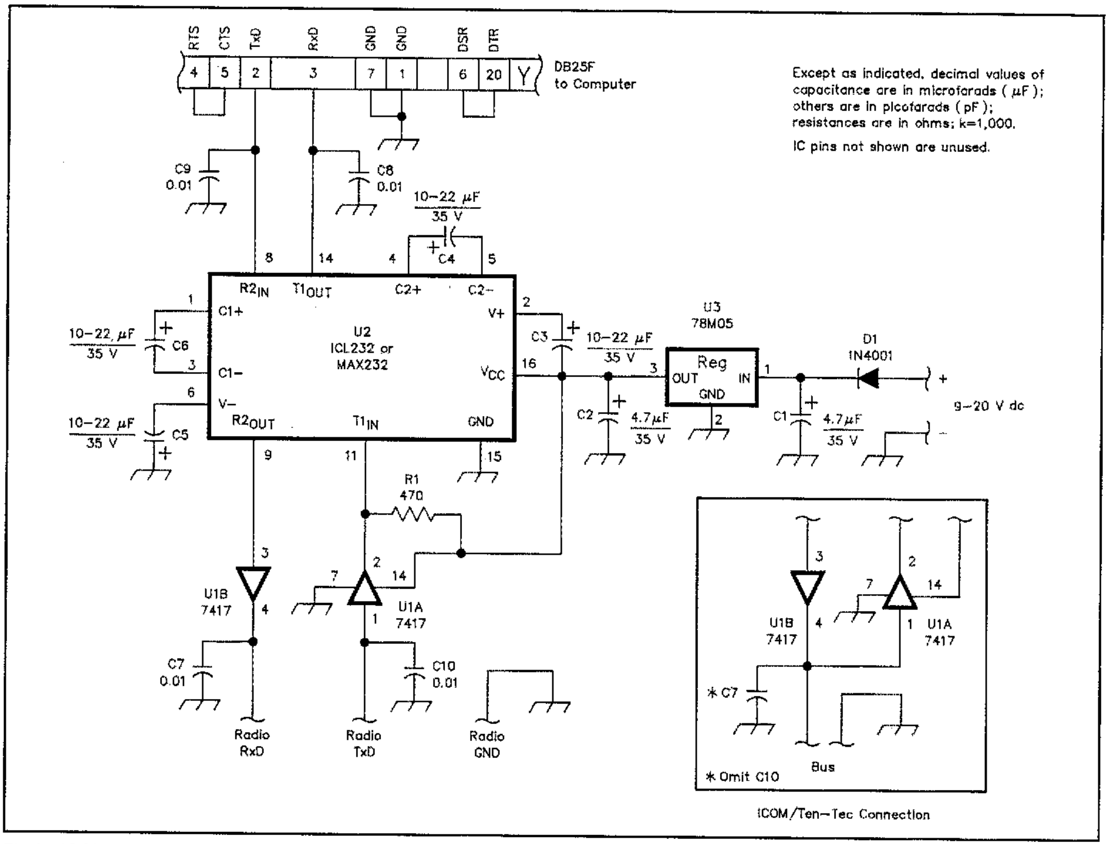


Fig 2—ICOM/Ten-Tec/Yaesu interface schematic. The inset shows the ICOM/Ten-Tec bus connection, which simply involves tying two pins together and eliminating a bypass capacitor.

C7-C10—0.01-μF ceramic disc.

U1—7417 hex buffer/driver.

U2—Harris ICL232 or Maxim MAX232 (see note 5).

It's worth noting that the ICOM and Ten-Tec radios also use identical basic command sets (although the Ten-Tec set includes additional commands). Thus, driver software is compatible. I commend the manufacturers in working toward standardizing these interfaces. This allows the Ten-Tec radios to be used with all popular software that supports ICOM's CI-V interface. When setting up the software, simply indicate that an ICOM radio (such as the IC-735, which has the same default address as the Omni VI) is connected. The Ten-Tec ports operate at rates up to 19.2 kb/s for very fast response.

Kenwood

Kenwood's system is similar to Yaesu's. The difference between the two is that Kenwood uses the RTS/CTS lines, and the received data (RxD) and transmit-data (TxD) lines are active-low. As the sidebar explains, this means that a binary zero (0 V) represents a logic one. This characteristic makes it easy to fully isolate the radio and the PC, since you only have to ground the radio's signal lines to assert them. Doing this with optoisolators and keeping the radio and computer grounds separate provides excellent isolation. To maintain isolation, a separate power supply should be used for the interface. A wall transformer works just fine for this. Fig 3 shows the Kenwood interface.

The interface uses a DB25 connector at the computer end. Kenwood uses a 6-pin DIN connector at the radio. Refer to the radio's manual for pin connections.

Some Kenwood radios made a few years back require additional parts to use the interface. The TS-440S and R-5000 require an 8251A and a 4040 chip.[6] Some other radios, such as the TS-940S, require an interface board available from Kenwood. The newer radios, including the TS-450S, TS-690S, TS-850S, TS-950S/SDX and TS-790A, require no additional parts.

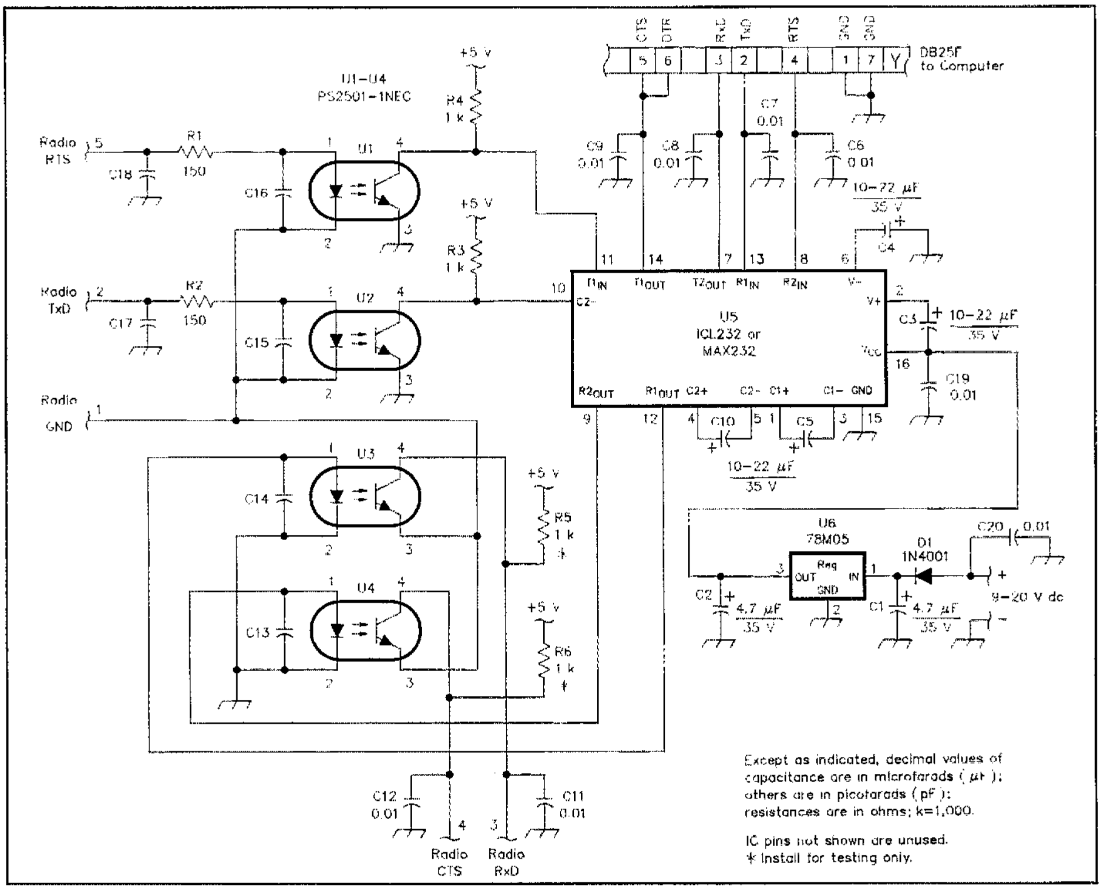
Ten-Tec Paragon, Omni V and Omni VI

If you own a Ten-Tec Omni V or Omni VI, you don't need an interface: These radios have RS-232-C interfaces. (This interface is also a Paragon option.) All you need is a cable to connect the radio to the computer's serial port.

The Omni VI is the most flexible Ten-Tec model. It has two ports: In addition to the one described above, it also has a port for a CSMA/CD bus, as described in the section on controlling ICOM radios.

Yaesu

Yaesu's system is the simplest of those described here. Only the RxD and TxD lines are used; RTS and CTS aren't used to control



Except as indicated, decimal values of capacitance are in microfarads (µF); others are in picofarads (pF); resistances are in ohms; k=1,000.

IC pins not shown are unused.

* Install for testing only.

Fig 3—Kenwood interface schematic.

C6-C9, C11, C12, C17, C18—0.01-µF ceramic disc.

C13-C16, C19-C21—0.001-µF ceramic disc.

U1-U4—PS2501-1NEC (available from Digi-Key—see note 5).

U5—Harris ICL232 or Maxim MAX232 (see note 5).

the data flow. This may seem foolish, but in most cases these signals aren't necessary, and most software ignores them when they are present!

The connector for the computer end is a DB25F. The radio end varies with the model. Some, such as the FT-980, use a 6-pin DIN connector. These radios transmit and receive data. Others, such as the FT-757GXII, use a three-pin connector. These radios only have an RxD connection on the radio—the radio doesn't send information back to the computer. The other two pins are for ground and AGC. Refer to your radio's documentation for details.

Testing

It's relatively easy to test the completed interfaces; remember, all they're doing is shifting signal levels. You'll need a 5-V supply, a 9-V battery and a voltmeter. Simply apply voltages as described in Table 1, 2 or 3 and look for the proper outputs with the voltmeter. When an input of −9 V is called for, simply connect the positive side of the 9-V battery to ground and the negative side to the test point.

For the Kenwood interface, don't forget the separate grounds. Use the right one to ground a signal line for testing. Also, install R5 and R6 only during testing. Because Yaesu uses more than one type of connector on their radios, Table 3 lists signal names instead of pin numbers. Consult the radio documentation to associate the signal names to the correct pins. Also, don't forget that Yaesu radios using the three-pin connector use only the RxD line at the radio—these radios don't transmit data to the computer.

Construction Tips

The interfaces can be built on a PC board or breadboard, or using point-to-point wiring. PC boards, templates, and parts kits are available.[7] The parts are readily available.

It's a good idea to enclose the interface in a metal case and ground it well. Use of a separate power supply is also a good idea. You may be tempted to take 13.8 V from your radio—and this works well in some cases—but you sacrifice isolation and may have noise problems. Since these interfaces draw very little current, a wall transformer fits the bill nicely.

Table 1
Kenwood Interface Testing

Apply	*Result*
GND to Radio-5	−8 to −12 V at PC-5
+5 V to Radio-5	+8 to +12 V at PC-5
+9 V to PC-4	+5 V at Radio-4
−9 V to PC-4	0 V at Radio-4
GND to Radio-2	−8 to −12 V at PC-3
+5 V to Radio-2	+8 to +12 V at PC-3
+9 V to PC-2	+5 V at Radio-3
−9 V to PC-2	0 V at Radio-3

Table 2
ICOM/Ten-Tec Interface Testing

Apply	*Result*
GND to Bus	+8 to +12 V at PC-3
+5 V to Bus	−8 to −12 V at PC-3
−9 V to PC-2	+5 V on Bus
+9 V to PC-2	0 V on Bus

Table 3
Yaesu Interface Testing

Apply	*Result*
GND to Radio TxD	+8 to +12 V at PC-3
+5 V to Radio TxD	−8 to −12 V at PC-3
+9 V to PC-2	0 V at Radio RxD
−9 V to PC-2	+5 V at Radio RxD

Another subject worth touching on is the radio's communication configuration. The computer's serial port must be set to the same data rate, parity and number of start and stop bits as the radio. Check your radio's documentation and configure your software or use the PC-DOS/MS-DOS **MODE** command as described in the DOS manual.

A great deal of software is commercially available for controlling radios via IBM PCs and other popular computers. See the ads in *QST* and the *National Contest Journal* for details.

Summary

Understanding the basic principles of radio-computer interfacing, you can easily and inexpensively build an interface for your equipment. Enjoy the flexibility your radio, computer and software can give you!

Notes

[1] G. Kalata, "An Inexpensive Computer/Radio Interface," *QST*, Sep 1991, pp 24-25. Also see Feedback, *QST*, Dec 1991, p 46.

[2] N. Thompson, "A Low-Cost PC Interface for ICOM Radios," *QST*, Jul 1992, pp 37-38.

[3] Z. Baranski, "Easy Computer Control and Bandwidth Switching for Kenwood Transceivers," *QST*, Jun 1992, pp 23-26.

[4] PC board templates are available from the ARRL Technical Department Secretary for a business-size SASE. Etched and drilled boards are available from CW Enthusiasts, 1346 Erickson Ave, Columbus, OH 43227. Boards cost $8 each, postpaid. COD orders are accepted at 800-547-7479 (24 hours, orders only). Add $5 to COD orders. Complete parts kits with assembly instructions are also available from CW Enthusiasts. Prices: Kenwood, $47 postpaid; ICOM/Ten-Tec and Yaesu, $44 each, postpaid. Kits include an etched and drilled PC board, all board-mounted parts, cabling and connectors. A case is not included.

[5] The Harris ICL232 chip is available from Digi-Key Corp, 701 Brooks Ave S, PO Box 677, Thief River Falls, MN 56701-0677, tel 800-344-4539 or 218-681-6674, fax 218-681-3880. Orders under $25 incur a $5 handling charge. The chip, equivalent to the Maxim MAX232, is also available from CW Enthusiasts (see note 4) for $6 postpaid.

[6] This chip set is available from Kenwood as the IC-10 for approximately $40. Another source is CW Enthusiasts (see note 4) as the CW-10, for $19 postpaid. Follow the installation instructions in your radio's manual.

[7] See note 4.

Wally Blackburn was first licensed in 1990 and holds an Amateur Extra Class license. He earned a BSEET from DeVry Institute of Technology and is currently working on a Master's degree in computer science at Ohio State University.

Wally started his career with AT&T Bell Laboratories and has recently moved to AT&T's Network Services Division, where he works as a Unix system administrator.

Wally enjoys playing with computers and mild contesting, and works CW almost exclusively.

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The DCI-100 is a standard single-slot add-on card that directly addresses the CPU's I/O bus addresses, so it doesn't tie up a serial (COM) port. It interfaces with most amateur transceivers (it's a lot like hooking up a packet TNC). More than one card can be installed in a single PC for use on separate frequencies, bands, etc, with voice messages stored in separate subdirectories.

Retail price $199 per board, $69 software; one complete set of hardware and software $259. Ralph Olds, AA7GT/VE7OV, Digital Communications Inc, 8946 Shook Rd, RR #4, Mission, BC, V2V 5M2, Canada; tel 604-820-1162, fax 604-826-0704.

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QST

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OUR COVER

How much power is enough? FCC rules state that "an amateur station must use the minimum power necessary to carry out the desired communications." You might be surprised at how little that can be! See the article by Diehl Martin, N5AQ, and David Newkirk, WJ1Z, on page 28. *(photos by Kirk Kleinschmidt, NT0Z)*

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March 1993
Volume LXXVII Number 3

TECHNICAL

NEWS AND FEATURES

NEW HAM COMPANION

OPERATING

DEPARTMENTS

The NRY A Simple, Effective Wire Antenna for 80 through 10 Meters

Known as a *broadside collinear curtain array*, this antenna is simple to build, rakes in DX signals and has gain over a dipole on all the bands it covers!

By Rick Olsen, N6NR
ARRL Technical Advisor
PO Box 538
Issaquah, WA 98027

Hams' thoughts ultimately turn to antennas. So it was with me a while ago. I had just moved to the emerald beauty of the Pacific Northwest and began pondering the 160-foot-tall Douglas firs in my backyard. I had considered putting up a crank-up tower, but decided that neighbors, architectural committees and crank-ups don't mix. Then I thought about topping a tree. Nope—I moved here because I love these trees.

I started sniffing around for alternatives. I obtained a copy of fellow ARRL Technical Advisor Roy (W7EL) Lewallen's antenna-modeling program, *ELNEC*, along with a copy of *The ARRL Antenna Book*, and began to incubate the seeds of what would become my most enjoyable antenna project since I built a four-element quad back in 1974.

Hanging ropes from big trees is not a new idea.[1,2] I began by pacing off the distances between the monsters in my yard so I would know how much space I had to work with and how I could optimally direct my signal to places like Europe, Africa and the Caribbean.

Next came the selection of a configuration. Because the trees are approximately 120 feet apart, my first choice was a multielement collinear array. But a good friend, Terry Conboy, N6RY (in *NRY*, I'm the *NR* and he's the *RY*), suggested stacking a pair of *extended double Zepps*. I filed his suggestion away in my memory for the time being. I started my analysis.

The ARRL Antenna Book, on page 8-32, describes the basic function of the collinear array. In sum, it concentrates energy perpendicular to the wire, hence the term *broadside array*. This antenna's physical layout and pattern are shown in Fig 1. Such an antenna's gain can be increased by varying the end-to-end spacing of the elements.[3]

Chapter 8 goes on to explain three- and four-element arrays. Disappointment set in when I sized a four-element in-line array for 20 meters: It wouldn't fit between the trees! What *would* fit, however, was a "two-over-two." In fact, depending on the vertical separation, this array might even have some gain over the four-element collinear arrangement.[4]

My choice was made. I wanted to maximize performance on 20 meters and take what I could get on the other bands. I chose two 1-λ, center-fed wires vertically spaced ⅝ λ apart. According to *ELNEC*, with the array 100 feet high at the top and modeled over lossy ground, this antenna's gain is approximately 12 dBi.[5] Not bad! The only problem is that this antenna's 20-meter feed-point impedance is very high.

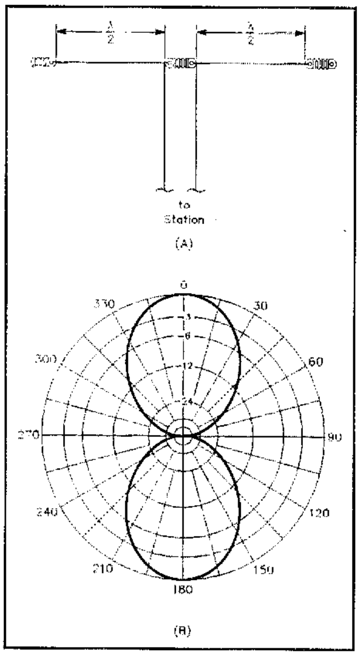


Fig 1—A two-element collinear array (A), also known as *two half-waves in phase*, is merely two half-wave wire elements placed end to end. This antenna has about 1.5 dB gain over a dipole in the same surroundings, and B shows its free-space pattern.

Reenter N6RY and the extended double Zepp. Take a look at Fig 2A. The extended double Zepp is simply a 1.25-λ, center-fed wire (two end-to-end ⅝-λ elements). There's a hidden message here! Have you found it?

Well, at first, I didn't either. Terry explained to me that the extended double Zepp (EDZ) is essentially two half-wave

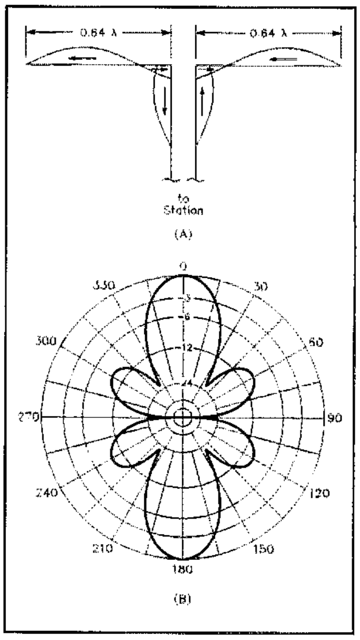


Fig 2—At A, the physical layout of and current distribution on an *extended double Zepp* (EDZ), a special case of the two-element collinear array shown in Fig 1A. This antenna exhibits more gain than the Fig 1A antenna because lengthening the elements to 5/8 λ has the same effect as spacing the element ends of two collinear half-wave elements 0.28 λ apart. (See Note 3.) At B, the antenna's azimuth-plane pattern in free space. This antenna has about 3 dB gain over a dipole in the same surroundings.

---

[1] Notes appear on page 24.

collinear elements spaced just over ¼ λ apart. The extended double Zepp's added gain over two half waves in phase comes *not* from the extra 0.28 λ of wire between them, but from the *increased end-to-end spacing of the outermost pair of half-wave elements.* (Radiation from the inner 0.28-λ actually works *against* that from the outer half-waves, but the gain increase from the new element spacing more than overcomes this disadvantage.) Figs 1B and 2B are similar. The EDZ has broader lobes and useful smaller lobes, both of which are attractive for a fixed array.

Sure enough, it was right in front of me all the time in the *Antenna Book* (page 8-34, fifth paragraph). In fact, the *whole darned array* was right in front of me in Chapter 8! I just needed some help piecing it all together. I modeled this array every way I could imagine, looking for the best possible performance. My wife didn't see me in the evenings for two weeks! Fig 3 shows the final configuration.

Figs 4, 5 and 6 show some of the results of my modeling. I used a design frequency of 14.2 MHz, a height of 100 feet above ground (for the top wire), elements made of #14 copper wire, and ground coefficients typical of the Pacific Northwest. All the elements in this array are fed in phase, as described in a bit.

As the figures show, the antenna's gain and patterns are impressive. On the 80- and 40-meter bands, the antenna has remarkably dipole-like patterns and gains and seems to work at least as well as a dipole in the same height range. On the higher bands, it works *much* better than a dipole. Its predicted gains in my installation are 9, 11 and 14 dBi at 7, 10.1 and 14 MHz, respectively. At 18.1 MHz (Fig 6) and above, the array begins to take on long-wire properties. It has several major lobes in directions closer to the wire axis, which the lower-frequency patterns don't provide.

To help you evaluate the NRY's patterns, Figs 4-6 superimpose the patterns of half-wave dipoles cut for 10.1, 14 and 18.1 MHz, respectively, on the NRY's patterns. These comparisons assume that the dipole is the same height as the NRY's feed point (halfway between the elements, or about 22 feet lower than the top wire), and over the same ground.

You may wonder about this antenna's effectiveness as a function of height. According to the *Antenna Book*, arrays like this have the best gain when the lower element is at least ½ λ above the ground (33 feet at 14 MHz). And, of course, as with all horizontally polarized antennas, this array's radiation angle decreases as you raise the antenna, which makes for better DX performance. The bottom line is that for DX work, you don't have to get the array up 100 feet for it to work well—you just need to put it up as high as you can. For closer-in coverage, it will work well at lower heights.

Another advantage of this array is that its feed-point impedances are manageable. To maintain my goal of broadband use, I decided to feed the antenna as shown in Fig 3. This is the "lazy-H" configuration described starting on page 8-37 of *The ARRL Antenna Book.* The attractiveness of this configuration is that it doesn't rely on specific phasing-line lengths to work properly. To the extent made possible by the different mutual impedances between each element and the ground, the phase relationship between elements remains constant regardless of frequency.

Final Dimensions

The antenna's physical dimensions (86 feet, 10 inches long and 43 feet, 5 inches high, not counting insulators) are easy to remember because the height is half the length. Because the antenna is basically 1.25 (5/4) λ long, center-fed and vertically spaced 5/8 λ apart, you can easily scale it to whatever band you like. For instance, to scale it to 10 meters, divide all the dimensions by two (28 MHz/14 MHz = 2).

Construction

Now for the really fun stuff: putting this thing up in the air! Building the antenna is a breeze. It took me all of about an hour to cut and solder the wires and attach the tethers. The harder part was preparing the trees.

If you have a friend like Bernie Olshausen, N6RUX, you have it made. He's a crack shot with a slingshot, similar to that described by Wade Calvert, WA9EZY, in *QST* a couple of years back.[6]

Once Bernie landed the fishing lines over the right tree limbs, the next job was to pull up some lightweight twine (we used seiner twine–the stuff fish nets are made of) with the fishing line. This is important; rope sometimes snags as it goes over limbs, and you don't want to try to pull it over with 6-pound-test fishing line! With the twine, we pulled up some 800-pound-test nylon rope with a pulley on one end, using another piece of nylon rope through the pulley to act as a catenary for the array.

It's important to use a strong, weather-resistant catenary rope to hang the array, because the antenna loses its desirable properties quickly when it takes on the shape of a **V**. You can expect a 2-dB gain decrease at 20 meters with a 20° variance from the flat-top configuration. The rope, therefore, must be strong enough to minimize droop and maintain integrity in the sun and wind. We'll get back to the rope in a minute.

For the phasing harness I used low-loss, 300-Ω twinlead. Any parallel-wire feeder should work fine. Don't use coaxial cable in the feed system. It's too heavy, unbalances the array and has very high loss when operated at a high mismatch.[7]

To assemble the feed system, I used porcelain insulators and equal lengths of 300-Ω line. I then cut the #14 antenna wire to the proper length and soldered it to the feed system in a standard dipole configuration.

Next came the tethers. For an antenna almost 87 feet wide and 43 feet tall, you need to provide some means of maintaining the proper element spacings. I used equal lengths of seiner twine (nylon or dacron rope would also work fine) to vertically space the element ends. I also attached plenty of line to the lower end insulators so I could tie them to the appropriate fence posts once the antenna was in the air. See Fig 3.

The catenary rope is the trickiest part. The antenna won't hang right if you don't attach the top three insulators in the right spot. I know—I did it wrong the first time!

I cut a piece of catenary rope 95 to 100 feet long and carefully marked the center. Next, I measured the distances from the center to each end of the array. (The array itself is a handy measuring line.) Then I added about 10 more inches and marked the line. I'll tell you why in a minute.

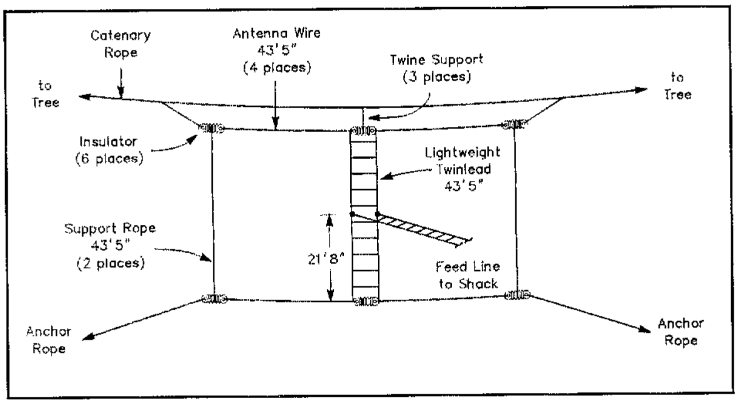


Fig 3—The broadside collinear curtain array—the NRY—in its final configuration. This antenna is cut for 20 meters, but works well on all the HF ham bands. The array hangs from a catenary rope via three short lengths of twine. The lower end insulators are supported by ropes to those of the top wire, and the center insulators are joined by the phasing line, which is 43 feet, 5 inches of parallel-wire feed line. The feeder to the radio is made of the same type of line connected at the halfway point on the phasing line.

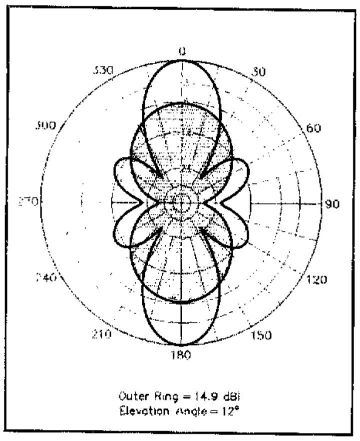


Fig 4—The NRY's 14-MHz azimuth-plane pattern with the bottom wire at 56.5 feet and top wire at 100 feet. For comparison, the shaded part shows the pattern of a half-wavelength, 14-MHz dipole at 78 feet (the same height as the NRY's feed point).

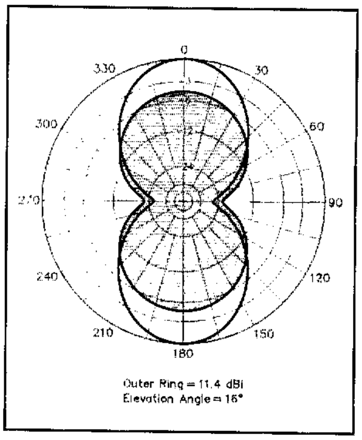


Fig 5—The NRY's 10.1-MHz azimuth-plane pattern with the bottom wire at 56.5 feet and top wire at 100 feet. For comparison, the shaded part shows the pattern of a half-wavelength, 10.1-MHz dipole at 78 feet (the same height as the NRY's feed point).

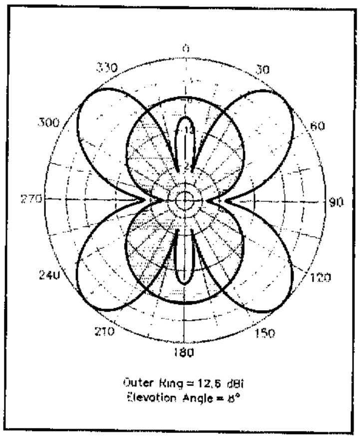


Fig 6—The NRY's 18.1-MHz azimuth-plane pattern with the bottom wire at 56.5 feet and top wire at 100 feet. For comparison, the shaded part shows the pattern of a half-wavelength, 18.1-MHz dipole at 78 feet (the same height as the NRY's feed point).

Then I cut three pieces of twine, each about 24 inches long. At the center of the catenary rope, I carefully opened the rope weave and passed about 6 inches of twine through it, then let the catenary rope retake its natural shape. Wrapping the twine around the rope several times on either side of the joint relieved the tension placed on the rope introduced by passing the twine through it.[8] I then tied the other end of the twine to the top center insulator so that the spacing between the rope and the insulator is 12 inches.

After doing that, I attached the twine at the end marks on the catenary rope in the same manner. The distance to the end insulators is 18 to 20 inches. This causes the ends to hang off the rope at an angle and provides strain relief.

If you use a braided rope (with a weave you can't open to pass twine through it), one good alternative is to use insulators in the catenary rope (at the points where I attached the twine) as tie points for the array.

Now you're ready to haul the antenna up on the lanyards. Consult your US Navy *Marlinspike Seamanship* manual (just kidding, of course) and join the ends of the catenary to the lanyards that are hopefully by now through the pulleys, up in the trees. Use a strong knot that won't slip.

Before hoisting the array into the sky, *make sure* that the connections to the feed harness at the top and bottom are the same (ie, that the left elements both attach to the same wire in the twinlead). This thing does not work right if the top and bottom are fed 180° out of phase!

Now you're ready to haul it up and connect the feed line to the antenna tuner. Check to make sure that the symmetry and top-to-bottom wire spacings are within reasonable limits, then tie the bottom tethers off so that the bottom wire is as flat as you can make it.

You're Done!

This is quite an effective antenna. I've installed two of them: One is boresighted on the Middle East and works like gangbusters into Europe as well; the other favors the Caribbean and most of the US. In the first three months after putting them up, I worked 105 countries on CW, including some rare ones, with a 100-watt transceiver. I'm surprised at how easily I can bust pileups—especially on 30 and 20 meters.

I hope you have an opportunity to try this array on for size. Don't limit yourself to just the backyard, either: It's a great Field Day antenna, to say the least. It also comes in handy for emergency work where you need a gain antenna.

You don't have to be a rocket scientist to design effective antennas! Your old friend *The ARRL Antenna Book*, and other ARRL publications, like Walt (W2DU) Maxwell's *Reflections*, the *Handbook*, the *Antenna Compendium* series and Wilfred Caron's *Antenna Impedance Matching*, are serious reference guides and sources of numerous ideas. It doesn't hurt to have your favorite antenna-modeling software loaded on your PC, either!

Notes

[1] D. Brede, "The Care and Feeding of an Amateur's Favorite Antenna Support—the Tree," *QST*, Sep 1989, pp 26-28, 40.

[2] J. Hall, ed, *The ARRL Antenna Book*, Chapter 22 and pp 4-3 and 15-3.

[3] *The ARRL Antenna Book*, Fig 8-38, p 8-32, illustrates how gain varies with the spacing of the two adjacent element ends.

[4] The *Antenna Book*'s Fig 8-45, p 8-35, suggests that a spacing of 5/8 to 2/3 λ yields good broadside gain. For the four-element array, upwards of 8.5 dBi gain is obtainable.

[5] *dBi* means decibels relative to an *isotropic radiator* (point source) in free space. It *does not* denote gain relative to a point source above real ground, or a dipole in any surroundings.

[6] W. Calvert, "The EZY-Launcher," *QST*, Jun 1991, pp 34-35.

[7] *The ARRL Antenna Book* shows how coaxial-cable loss varies with matched loss and SWR in Fig 18, p 24-18. (The *Antenna Book* also contains a table of matched loss per 100 feet as functions of frequency and cable type in Fig 26, p 24-18).

[8] If you use braided rope for the catenary, you'll need to use another method to secure the seiner twine to the catenary rope.

Rick received his Novice license in 1964 (WN7CNP). His enjoyment of the hobby extends from DX to moonbounce. He has been active in public service and disaster communications since his Novice days and has held numerous ARRL field appointments. Rick has been an ARRL Technical Advisor since the beginning of the TA program in 1976.

Rick enjoys CW and holds a 40-WPM code-copying certificate. He is an active low-power DXer, favoring the challenge of sneaking around the pileups instead of boring holes through them. Rick is an instrument-rated private pilot and a first tenor in the Seattle Opera Chorus.

Professionally, Rick is the Director of Engineering for US WEST's NewVector Group, which is the parent company for their cellular and paging operations. He did his undergraduate work in electrical engineering at Arizona State University. He also holds an MBA from La Jolla University.

Rick has also authored numerous technical articles for magazines and The ARRL Handbook.

Automatic RF Power Control for AMTOR Operation

How do you comply with the FCC Part 97 rule that requires us to use the minimum necessary power? Most of us do it manually—but now we can do it *automatically* in AMTOR operation.

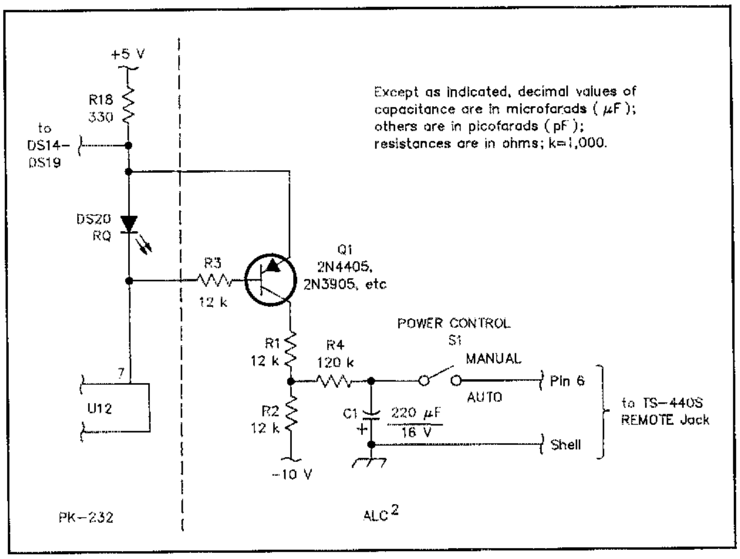
By Carl D. Gregory, K8CG
2002 Kenny Ave
Champaign, IL 61821

Part 97 of the FCC rules requires hams to use the "minimum transmitter power necessary to carry out the desired communications." When was the last time you operated in compliance with this rule? If you operate AMTOR Mode A (ARQ), this simple project enables you to comply with it *automatically*!

In a recent *QEX* editorial,[1] Paul Rinaldo, W4RI, challenged hams to develop automatic power-control systems similar to those used by commercial networks, such as cellular telephone systems. As Paul pointed out, most amateur communication is simplex, with typical transmission times of a minute or more, so real-time power control is difficult.

I was intrigued by Paul's challenge. AMTOR Mode A is an exception to the normal simplex operation mode, and *does* provide a near-real-time feedback signal: About twice a second, the remote station gives a one-bit signal report. (Cellular phone systems use a three-bit report.) The report is in the form of either an acknowledgment (signal okay) or request for retransmission (*RQ*—signal too weak to copy due to QSB, QRM, QRN or whatever).

I decided that this information, present as an electrical signal in my AEA PK-232 multimode communications processor (MCP), could be used to adjust the transmitter power. The radio's external ALC input, intended to prevent destructively overdriving a linear amplifier, would provide the connection. A simple interface was all I'd need. The result, which I call the Automatic AMTOR Link Level Control Circuit (ALC²), is described here with the hope of stimulating more development in this important area. It's effective, simple and compatible with the existing Mode A protocol. Unlike other current proposals, which aim to increase data rate (often to levels that casual keyboard operators can't use) while maintaining constant power, this technique proposes to optimize channel use by adapting *power* while maintaining a constant data rate.



Except as indicated, decimal values of capacitance are in microfarads (µF); others are in picofarads (pF); resistances are in ohms; k=1,000.

Fig 1—Schematic of the simplest version of ALC², showing connections to PK-232. All resistors are ¼ W, 5% tolerance.

Planning the Project

My first step was to determine the characteristics of the transceiver and data controller. I use a Kenwood TS-440S, which has a rear-panel ALC input (pin 6 of the **REMOTE** DIN connector). I determined that this input responds to voltages in the range of −4 to −9, with most of the action taking place between −6 and −8 V. Output power can be controlled over a range of more than 36 dB (from 100 watts down to 25 milliwatts—as little as I can measure). The impedance of this input is a few hundred kilohms.

My PK-232 has an LED labeled **RQ**, which flashes every time the other station requests a repeat of a three-character group. Don't confuse this with the **ERROR** LED, which flashes when the local station (mine) can't copy the remote station's transmission. There's no way that increasing power at the local station will help this situation, so we ignore it, except to send an RQ to the remote station.

The **RQ** LED (the PK-232's DS20) is controlled by the microprocessor in the PK-232 via an encoded signal on Port B of the 8536 parallel I/O chip (U6). The signal is decoded by U12, a 7445 BCD-to-decimal decoder/driver. Thus, the only place the "signal report" is present as a discrete electrical signal is at the output of U12. Unfortunately, the open-collector nature of U12, and the sharing of the current-limiting resistor R18 between six LEDs, means that there is usually no standard TTL level signal at that point. However, a pull-up resistor of 10 kΩ or so can be connected from the cathode of DS20 to +5 V to put a TTL level here.

To generate a negative voltage for the rig's ALC circuit, we need a negative power supply. The PK-232 has a low-current negative supply available at the anode of D9 (−10D on the schematic). Signal ground and +5 V are available lots of places in the PK-232 (the latter, for example, appears at

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[1] P. Rinaldo, "Power Control in System Design," *QEX*, Oct 1991, page 2.

the cathode of D12). Similar signals and voltages should be available in other AMTOR controllers.

Now that we know what the interface has to interconnect, we can design it. We'll need at least two functional blocks. One is a buffer/amplifier/level shifter to convert the signal at U12/DS20 to positive-going pulses in the range of −10 to −5 V, compatible with the transceiver's ALC circuit. The other is a low-pass filter, or integrator, to average this signal and thereby produce a smoothly varying feedback signal for the transceiver.

A critical feature of this project is the very slow feedback of signal-strength information from the other station. Only one bit of information is available each half second. If we want to cover the 36-dB range in 3-dB steps, we would have to wait at least 6 seconds for this information to be returned by the remote station. On the other hand, an ALC switch based on the RQ signal could instantly change the power over the entire range.

In the language of control systems, the potential gain is high, and the phase shift (delay in getting information back) is great, as a result of the low-pass filter imposed on the signal reporting data by the slow link speed. This combination creates the potential for oscillation. To damp the oscillation, I chose a low-pass filter with a very long time constant in the feedback circuit. (This is probably not the best way to do this, but it works.) The result is that the power level changes slowly, and the feedback from the remote station can catch up with the power changes.

The choice of a time constant is a compromise. If it's too long, the circuit will not track QSB; if it's too short, the power will oscillate wildly between 100 watts and a few milliwatts. I didn't optimize this value—perhaps someone with a theoretical bent will do so. The components shown result in an oscillation period of more than 10 seconds (ie, hard to tell from QSB), and the oscillation amplitude is reasonable (a few decibels). Someone skilled in feedback-loop analysis could undoubtedly design a better integration scheme than this simple RC filter.

The First Circuit

The simplest circuit I came up with, as a feasibility study, is shown in Fig 1. It uses only seven components. PNP transistor Q1 detects the voltage drop across DS20 when the LED is turned on. R3 limits the base current so that Q1 doesn't steal all the current (which would keep the LED from lighting). This is necessary because the junction voltage of a GaAs or InP LED is larger than that of a silicon transistor.

As long as DS20 is off (indicating that the remote station is correctly receiving the transmitted data), C1 gradually charges to a negative voltage through R2 and R4 (time constant R4 × C1 = 26 seconds). This gradually reduces the transmitter output power (via S1 and the ALC line) to the point at which the signal finally becomes inaudible at the remote station. This is a level just below the FCC 97.313(a) specification.

At this point, the remote station sends an RQ—request for retransmission—and the LED comes on. Q1 turns on, partly discharging C1 through R1 and R4 (at roughly the same rate). The ALC voltage becomes less negative, and the power increases for the next transmission. This continues until the remote station signals that the link is solid. Q1 now turns off and the process repeats. If interference or fading intervene, the RQ signal may come earlier. In any case, the transmitter power varies around the optimum value, and automatically compensates for changing band conditions.

I included S1 because you might want to transmit using modes other than ARQ, for which manual power control must be restored. Also, during the establishment of a link (**PHASE** mode) and during changeover (**OVER**) the **RQ** light is off. A fade during these portions of the QSO will result in eventual link loss, unless S1 is opened to restore full power momentarily. Thus, the circuit is not quite fully automatic.

R1 and R2 form a voltage divider between about +1 V and −10 V when Q1 is on. As a result, the ALC output range is limited to −4.5 to −10 V if R1 and R2 are equal. This works fine with my TS-440S, but other R1:R2 ratios may be needed for other rigs.

Construction and Testing

I built the circuit on a scrap of perfboard. Layout is noncritical. Almost any PNP transistor should work at Q1. I used a 2N4403, because it was at the top of my junk box. A 2N3905, or one of the SK equivalents (SK3466 or SK3114A), or a grab-bag PNP transistor, should work fine.

Before connecting the circuit to your MCP, record all your parameter settings, turn off the power and remove the battery-backup connection (if your MCP has one). In the PK-232, you'll also need to remove the mailbox daughter board (if you have it) to reach the anode of D9.

Connect the circuit to the MCP with short lengths of hookup wire. I soldered connections directly to the component side of the main board of my PK-232 for the +5 V, −10 V and ground connections, and to the back side of the display board for the LED connections. This saved me from having to further disassemble the PK-232. A length of small-diameter coax (RG-174) can be threaded through one of the holes in the rear panel for the connection to the transceiver's ALC jack.

After putting everything back together and verifying that the PK-232 still works like it originally did, you are ready to test the ALC^2. Use a high-impedance voltmeter to monitor the ALC output signal. Put the data controller in packet mode (the **RQ** LED doubles as the **CMD** indicator in this mode). After a few seconds, the voltmeter should read about −4. Now go into Converse mode. The **CMD/RQ** LED should go out, the **ERROR/CONV** LED come on, and the meter reading should slowly drop to −10 V. Return to Command mode, and the voltage should climb back to −4. Connect the transceiver and get on the air!

A Better Circuit

As I mentioned, the circuit of Fig 1 isn't fully automatic. I built it only as a feasibility demonstration. A more practical circuit could use additional inputs from other LEDs and some logic to make a better decision about when to reduce power. A future data controller could be designed to provide a single output directly from the CPU for this purpose—possibly using a smarter algorithm, such as counting the number of retries, and so on.

My attempt at a fully automatic circuit is shown in Fig 2. The complexity has about doubled (to 15 components!). The basic control circuit is quite similar to that of Fig 1. I changed R1/R2 to a potentiometer (R1) to permit experimenting with the ALC threshold. I set the pot to midrange before installing it and have not changed it since. Returning the emitter of Q1 to +5 V gives a slightly larger range for the ALC output.

The big change in this circuit is the addition of logic—three **NAND** gates (U1A-U1C) and a diode-**OR** circuit (D1-D3). The circuit now responds to the PK-232's **PHASE** (from DS15), or **OVER** (from DS16) signals as well as **RQ**. If any of these conditions are detected, transmitter power is increased slowly. This guarantees at critical points in the life of the link that the circuit will not shut the transmitter down. Also, it now detects whether the controller is in Mode A (**ARQ** LED, DS8), and inhibits the power control if it is not. As a result, the switch is no longer required, although it may be retained for manual override in Mode A (or for the time-honored on-the-air A/B test) if desired.

I also built this circuit on a scrap of perfboard. I used a 74C00 to minimize current drain, in keeping with the QRP philosophy. Any ordinary signal diodes (1N914, 1N4148, etc) can be used for D1-D4.

Testing, as described earlier, requires overriding the Mode A logic by grounding the anode of D4 with a clip lead. (In principle, it is possible to address the LEDs directly using the *ADD*ress and *IO* commands in the July 1990 and later PK-232 firmware; however, I could not determine how to use the 8536 to do this in my PK-232.)

This circuit performs just as well as the first one, and requires less attention.

Results

It is really amazing to see how little power is sometimes required to maintain solid communications. I have a simple field-strength meter sitting on the shelf above the rig. On 20 meters, the needle is pegged during transmit at 100 watts. At times, with the ALC^2 in line, the meter does not move *at all* (perhaps a full 36 dB of power reduction), yet link data flow is unimpeded, even over distances of thousands of miles. (My first QSO with this gadget was with a station in

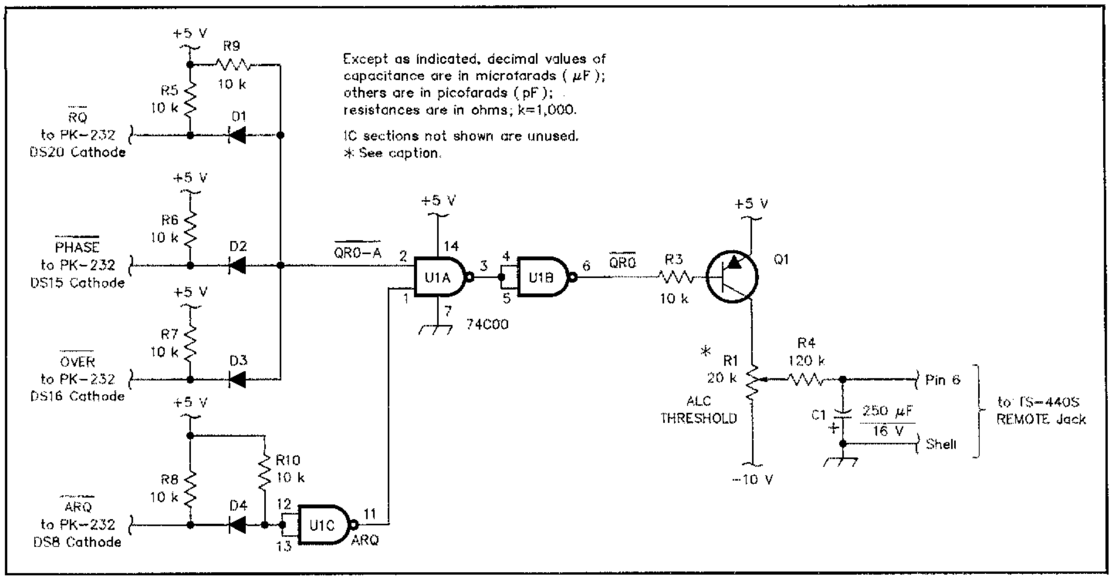


Fig 2—Schematic of a fully automatic version of ALC², showing connections to the PK-232 and logic-signal names. R1 replaces the R1/R2 combination of Fig 1.

D1-D4—1N914, 1N4148 or similar silicon diode.

Q1—2N4405, 2N3905 or similar PNP transistor.

R1—20-kΩ linear potentiometer.

U1—74C00 quad NAND gate.

Chile.) The error rate (ratio of chirps with **RQ** lit to chirps with **TFC** or **IDLE** lit) hovers around 10-15%, regardless of band conditions. This is fine, since with Mode A, there's no need for perfect copy. This suggests the following (hopefully provocative) statement: The power required for effective communication with AMTOR Mode A *actually may be less than that required for CW under the same conditions!* In other words, the signal-to-noise ratio for Mode A can be much lower than for CW for the same error-free copy. If this is true, then AMTOR Mode A is currently the ultimate mode for QRP operation.

Summary

I hope this concept (if not the circuit) will find wide use in the Amateur Radio community. Obviously, if a lot of us use this sort of system, the AMTOR frequencies can be simultaneously used by more stations, to the benefit of all. Stations forwarding HF traffic via APLink on emergency power, where reduced battery drain during good copy can be translated directly into longer up-time, should be real beneficiaries.

I leave it for other experimenters to improve the feedback loop, or to design a software scheme that optimizes the power control, yet maintains the basic AMTOR Mode A protocol for compatibility with existing systems. And, of course, the challenge to develop compatible systems for power reduction on voice and other modes remains open!

Carl Gregory earned his Novice ticket in 1971 at 20. He upgraded to Advanced, and finally Extra Class, in 1976. A Life Member of ARRL and AMSAT, Carl occasionally gets on the air to work teletype, but his primary interest has always been tinkering. He believes strongly in technical training as a major long-term justification for the existence and frequency allocations for the Amateur Radio service.

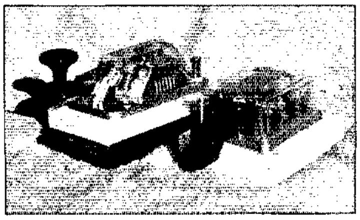
When he's not tinkering, Carl, with his wife, Faye (a confirmed nonham, but supportive of the hobby), enjoys traveling, hiking in the woods, and animals.

Carl attended the University of California at Riverside and Northwestern University. He earned a PhD in physical chemistry, and points out that training in electronics was notably absent from his formal education. Amateur Radio filled the gap. He's a laboratory manager at the University of Illinois Biomedical Magnetic Resonance Laboratory in Urbana, Illinois.

New Products

HI-MOUND CODE KEYS

◊ Impressed by the extraordinary keys manufactured by the Hi-Mound Company that he purchased on a trip to Japan, shipboard radio officer Mike Zbrozek, K8XF, has arranged to import the firm's products for US hams. CW aficionados and key collectors can obtain a unique Morse code key with a smooth touch, excellent workmanship and solid feel. The white marble MK-705 paddle is $129.95; the HK-702 hand key is $149.95 and the bakelite-based BK-100 bug is $154.95. Add $8 s/h. Mike Zbrozek, K8XF, 9929 Fox Squirrel Dr, New Port Richey, FL 34654; tel 813-862-6328.



Strays

I would like to get in touch with...

◊ Ten-Tec enthusiasts, collectors, fans and people with Ten-Tec circuit modifications. Mark Murray, WX3H, 2425 Jackson Pkwy, Vienna, VA 22180.

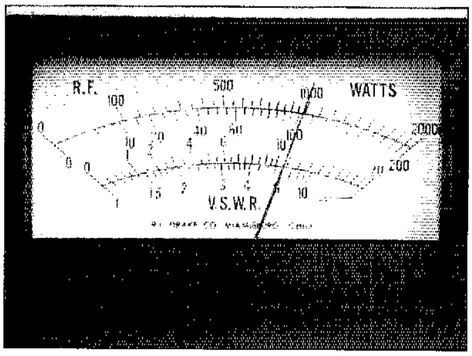
◊ anyone interested in collecting Amateur Radio postage stamps as a hobby. We'll send you a listing of others who reply and their current ham postage stamp lists. Fried (WA6WZO) and Sandy (WA6WZN) Heyn, 962 Cheyenne, Costa Mesa, CA 92626.

◊ amateurs who are career TV and radio broadcasters. Randy Berry, N3LRX, 2703 Keystone Ln, Bowie, MD 20715-2617.

Transmitter Power:

What It Is, What It Does and How to Use It

Sure we're required to use no more transmitter power than necessary—but how much do we *need*? We all know that using the minimum necessary transmitter power reduces interference—but *how*? Some of the answers may surprise you.



By Diehl Martin, N5AQ
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and

David Newkirk, WJ1Z
Senior Assistant Technical Editor

From Novice and Technician right on up to Extra, *all* FCC-licensed radio amateurs must obey the first rule of ham transmitter power: "An amateur station must use the minimum power necessary to carry out the desired communications." This article will help you understand why the amount of power you use matters, and what happens when you use too much or too little. Once you understand the effects of using too much and too little power, you'll be better equipped to run the *right* power more often.

Your Transmitter as Power Converter

Like miles per hour, power is a *rate*—how fast (or slow) something occurs over time. If you drive 60 miles in a hour, your driving rate for that hour—your *speed*—equals 60 miles per hour.

Your transmitter is, in effect, a power converter: It converts power-supply energy into radio energy. The more power your transmitter runs, the more power-supply energy it converts to radio energy every second. The unit of energy is the *joule* (pronounced *jewel*); the unit of power is the *watt*. One watt equals an energy-use rate of one joule per second.

The faster your transmitter uses energy, the louder your signal sounds at the receiving end. Thus, *more* power generally means *stronger* signals. The slower your radio converts power-supply energy to radio energy, the weaker your signal sounds at the receiving end. Thus, *less* power generally means *weaker* signals.

Power and Distance

How far your signal travels has nothing to do with how much power you run. As long as you operate your transmitter at a power greater than zero, your signal will travel everywhere propagation allows it to. Your signal will be vastly stronger in some places than others, and far too weak to be useful in many places. Increasing power doesn't make your signal travel farther, it just makes your signal stronger. The trick is to use just enough transmitter power to make your signal *usefully copiable at the place you want to contact*. How much power you *need* to do this depends on more than propagation conditions and what kind of antenna you use. It also depends on what kind of signal you send and the receiving situation at the station(s) you contact.

Transmitter Power from the Receiving End

Considered from the standpoint of good radio *reception*, the minimum power rule might go something like this: Use only enough transmitter power to make your signal usefully distinct from *noise* at the receiver.

Don't Get Caught in a Power Race

Among the worst types of interference to overcome is another signal of the same mode you're using. The most common way to overcome this is to increase your power level. But because the interfering station can probably hear you as well, its operator will overcome *your* interference by increasing *his* power output also! The result is that both of you are now using higher power, and neither of you is communicating any better. The additional power just lets you be heard farther beyond the station you're trying to reach. Result: a higher interference level on the band.—*N5AQ*

Noise can take many forms. If you turn up the volume on an operating radio receiver with its antenna disconnected, you'll hear hiss. Incoming signals must be strong enough to be distinguishable from receiver hiss or you can't communicate.

Connecting an antenna to your receiver may bring in "band noise" static, galactic noise and other natural sources that cause a continuous dull roar or hiss—that's strong enough to override the receiver's internal hiss.[1] Whichever is stronger—band or receiver noise—incoming signals must sufficiently override it or you can't communicate.

Whether it arrives in the form of other hams' signals or from man-made electrical pops, roars and buzzes, *interference* is a major source of noise. If incoming signals don't override it, you can't communicate.

The kind of signal you send is part of the equation, too. Different types of emission—CW, FM, SSB, RTTY and so on—take up different amounts of radio spectrum. For example, much of a CW transmitter's energy is confined to a relatively narrow frequency band—on the order of 200 Hz or so. An SSB transmitter, on the other hand, spreads its energy out over a bandwidth of 2.5 kHz or so.

Even when it's exactly tuned to a transmitter, a receiver picks up noise *and* transmitted signal. The trick is to maximize transmitted-signal pickup and minimize noise pickup. To do this, the receiver's band-

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[1] This statement applies mainly to AM, CW and SSB receivers. An unsquelched FM receiver produces about as much hiss with its antenna on as with its antenna off.

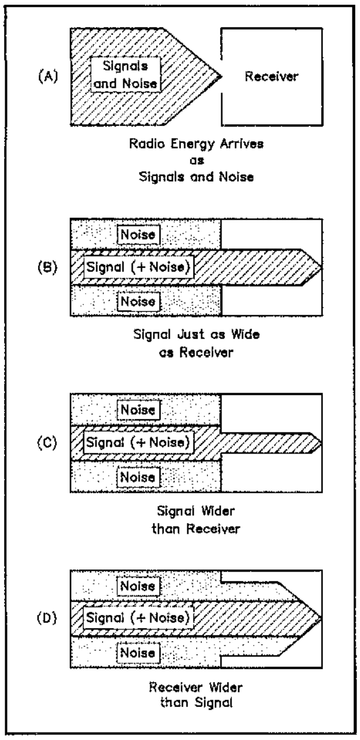

Fig 1—Your receiver (the box at A) doesn't just let *all* radio energy—signals and noise—enter. Its *selectivity*—the hole in the box wall—lets only some of the incoming radio energy enter. Ideally, your receiver's selectivity should only be wide enough to let in all of the transmitted signal (B). (*Some* noise always enters the receiver along with the signal.) Making the receiver *narrower* than the transmitted signal (C) wastes whatever signal power the receiver doesn't pick up. Making the receiver *wider* than the transmitted signal (D) lets in extra noise that makes the signal harder to understand.

width—its *selectivity*—should match the transmitter's bandwidth. Too-narrow receiver selectivity wastes transmitter power because the receiver doesn't receive it. Too-wide selectivity picks up all of the transmitted energy but adds additional noise.

Fig 1 illustrates this idea. The receiving station(s) in a contact help the transmitting station make the best use of its transmitter power by optimizing their selectivity to maximize the signal and reject noise. This can be as simple as switching from "CW wide" to "CW narrow," but it can make a big difference.

VHF and UHF FM radios generally already have optimized selectivity. You can't adjust their receiving bandwidth. But if you operate between 160 and 10 meters, your transceiver probably *does* allow some selectivity choice. By all means, try narrowing your radio's selectivity when noise and interference get tough.

Why Excessive Power is Worse Than Waste

Excessive power is actually worse than waste: *It's potential interference.* Fig A is one way of visualizing how this works. Operating your transmitter at a power significantly higher than what you need for comfortable audibility in your target area can cause interference to stations beyond your target. If you want to talk only as far as Philadelphia, running enough power for comfortable copy in Los Angeles is poor operating practice.

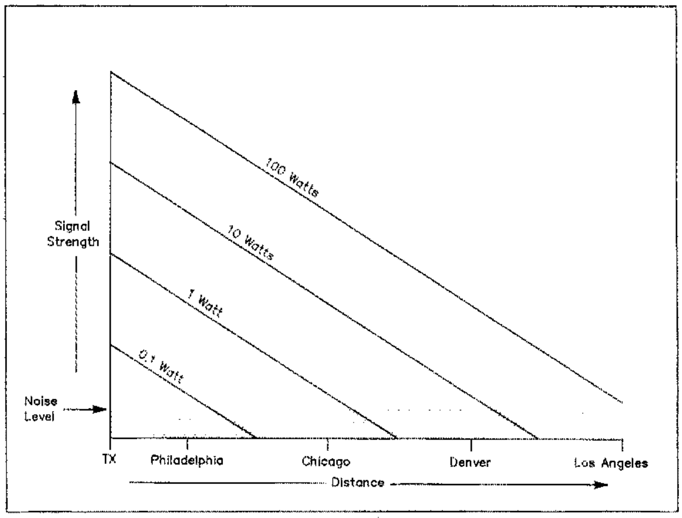

Fig A—To anyone not wanting to copy it, your signal is a form of noise called *interference.* This graph shows the communication distance possible at a fictional frequency and time with transmitter powers of 0.1, 1, 10 and 100 watts. In this example, operating your transmitter at 100 watts to contact Philadelphia when a tenth of a watt would do the job can cause significant interference far beyond Philly. On the other hand, trying to contact Denver with 1 watt *also* wastes power because your signal uselessly falls into the noise near Chicago. Ideally, you should adjust your power for comfortable copy at your target. Doing so ensures solid copy at your target station while allowing other stations in areas far beyond your target to simultaneously operate on—*reuse*—the same frequency.

Don't be misled by this graph's big-city labels. Redone with the names of world capitals or towns in your region, it communicates a concept that applies just as much to worldwide and local/regional communication. That concept is this: *Power that routinely lets stations comfortably hear you far beyond your target is* too much *power.*

Excessive power isn't just extra power. It can disallow successful frequency *reuse* where it might otherwise be possible. *Reuse* is when stations sufficiently remote from your target area operate on the same frequency as you, simultaneously. Ideally, they can't hear you and you can't hear them. If you run more power than you need, your signal may needlessly make it into the receivers of others attempting to reuse the frequency. When this happens, your signal stops being useful and starts being *interference*!—*WJ1Z*

Thinking about power from a receiving standpoint can help you see why increasing transmitter power isn't always the best answer to combating noise. If, for instance, a dying component makes your receiver excessively hissy, it's a much better solution to fix the receiver instead of having everyone you contact turn up their power. If a defective aquarium thermostat blankets your favorite band with buzz, finding it and fixing it is a much better idea than asking everyone you contact to increase power.

So it turns out that using the minimum power necessary isn't just the responsibility of transmitting stations. *All* stations using the radio spectrum must work together to use the lowest transmitting powers necessary.

Putting Your Power Where You Want It

Transmitting *and receiving* stations can improve their use of transmitting power in another, very important way: They can use *directional* antennas—antennas that direct energy *to* the right place and respond best to signals *from* the right place. Just as a flashlight bulb is a much poorer illuminator without a reflector behind it, a transmitter or re-

Minimum, Maximum or What?

The absolute maximum power output allowed for the Amateur Radio service over most of its bands is 1500 watts peak envelope power. Depending on your license class, operating band and the type of signal you transmit, your legal maximum power may be much less—200, 25 or 5 watts, for instance. Operating with your legal maximum power may sometimes be appropriate—in some emergency, contest and DX situations, or when working VHF meteor scatter and moonbounce communication, for instance.

So how much power do you need to talk from Point A to Point B? It's simply not possible to say that a given power level (100 watts output, for instance) is *always* enough, not enough or too much. Signal *fading*—signal-strength change over time—is the reason for this. Changing propagation can cause the minimum power you need to vary by a factor of more than 1000 to 1 in a matter of minutes. You may be able to communicate effectively over a 2000-mile path using less than 1 watt output at some times, and not be able to communicate there at all an hour later using 1500 watts. Especially on the shortwave ham bands, the *exactly* appropriate amount of power you need may vary continuously. But you needn't slavishly ride your transmitter's RF output control. the FCC's power rules let you play it by ear.—*N5AQ*

Too Much Power, Too Much Light

This page can communicate meaning to you only because you can sense meaning in the symbols it conveys through the arrangement of ink on its paper. You can't determine how the ink is arranged unless your eyes can tell where blank paper ends and ink begins. The ink must be sufficiently dark and the paper must be sufficiently light. Paper doesn't emit light of its own, so you need a light source to illuminate it. If the light isn't bright enough, though, your eyes won't be able to sense enough contrast between ink and paper for you to understand what this page's printing means.

With effort, you can successfully read even when there's too little or too much light for comfort. When poor lighting causes poor contrast, you can make up for it by concentrating harder or rescanning until you're sure of what you're reading. If there's too much light, you can reduce contrast by squinting or putting on sunglasses. But if you illuminate this page to the point of pain, you may actually be unable to read it!

Receiving radio signals is very much like reading printed text. The signal you want to copy is the ink. Noise (band sounds, hiss, static and interference) is the paper. Unless the signal is strong enough to contrast with the noise, your ears may not be able to read the signal.

Just as you can work harder to read an underlit printed page, under noisy radio conditions you can concentrate harder and have a sending station repeat until you copy its message. (Some forms of data communication do exactly this, automatically.) Just as you turn on a lamp to improve the contrast of ink on paper, you can ask a sending station to increase its transmitter power to make its signal stand out more clearly above the noise. Like using a reading lamp that's too bright, increasing transmitter power beyond a certain point doesn't improve copy. You can't improve copy to more than 100%!

This is what using "the minimum power necessary to achieve the desired communications" is all about. It's about using only enough transmitter power to ensure sufficient contrast between signal and noise. Reading light that you end up throwing away by squinting or putting on sunglasses is wasted light. Transmitter power that raises your signal farther above the noise than necessary for 100% copy is *wasted* power.—*WJ1Z*

ceiver with a nondirectional antenna may work, as the saying goes, "equally poorly in all directions."

An antenna that strengthens transmitted and received signals compared to another antenna is said to exhibit *gain* over the poorer antenna. Most of the gain antennas hams use are also *directional*. They achieve gain in one or more directions by reducing their performance in other directions. A directional gain antenna—many hams call them *beam* antennas—transmits more radio energy *to* its target and intercepts more radio energy *from* its target than an antenna with less or no gain. On transmit, this means a better signal-to-noise ratio at the target receiver. On receive, antenna gain improves signal-to-noise ratio by rejecting noise from nonfavored directions.

A flashlight reflector makes a flashlight brighter by concentrating lamplight in a beam. The lamp itself doesn't get brighter. Likewise, antenna gain increases your station's *effective radiated power* without your having to touch your transmitter's power control. On receive, antenna gain makes *other* stations stronger because the antenna intercepts more of their transmitted energy.

Directional gain antennas, then, are valuable tools in making better use of transmitter power *and* spectrum space. Antenna gain helps you use less transmitter power to achieve acceptable signal-to-noise ratios at other receivers. Antenna directivity helps keep your radiated power away from where you don't want it to go. This makes frequency reuse more possible because stations located where your antenna *doesn't* point may be able to operate on the same frequency without interference to or from you.

If directional gain antennas are so wonderful, why don't all hams use them all the time? One reason is that gain antennas are not always practical. Especially on the 160- through 30-meter bands, gain antennas can be pretty large. Even for bands where directional antennas are compact, installing a beam antenna, mast and rotator equipment can cost more than you may be willing to spend.

Antenna directivity may work *against* the kind of communicating you want to do. You may *want* omnidirectional coverage—for local and regional work, mobiling, or when you just want to get on the air and chat with whoever comes back to your calls.

Whatever type of antenna you use, you can still make plenty of contacts *and* operate in compliance with FCC's minimum power rule. Now's the time to talk about how.

Making the Most of Your Transmitter Power

Adjust your transmitter exactly as its manual describes. This is especially important for voice and data modes, in which misadjustment of transmit-audio levels can make your signal simultaneously strong and uncopiable.

Locate your radio's power control. Many modern radios can be conveniently adjusted to more than one power level. Some VHF and UHF radios provide a high/low power switch that allows the user to choose between two fixed power levels—for instance, 2 watts and 100 milliwatts. Many shortwave transceivers can be continuously adjusted between their maximum power output and some lower level—for example, between 100 watts and 10 watts. One or more controls marked **RF POWER OUTPUT**, **CARRIER** or **DRIVE** usually perform this function. With SSB transmitters that do not directly allow power output reduction, you can usually reduce output power by turning down the mike gain control. If you own an external power amplifier, this would be a good time to reacquaint yourself with its **OFF** or **STANDBY** switch!

Experiment with speech processing if your rig allows it. Adjusted *with care*, speech processing can help your voice cut through noise better by making weak speech components stronger. Adjusted carelessly, speech processing can make you sound like you're shouting from the cockpit of a jet taxiing through a bathroom. Don't consider your adjustments ready for prime time until you've monitored your own signal or made on-the-air tests with friends who'll tell you the truth.

Use the appropriate band for the distance involved. Is the band you're about to use open to the area you would like to contact? Scan the band for awhile before you transmit to see what propagation is doing. VHF or UHF bands—or, during daylight hours, 160 or 80 meters—are best for strictly local contacts. For contacts outside your immediate area, 160 meters or an HF band is probably appropriate. and deciding which band to use is a matter of knowing the usual propagation conditions for the bands at that time. Avoid making local contacts on a band that's open for DX—you may cause interference to other amateurs over much of the world and reduce chances for frequency reuse. Likewise, don't keep using a repeater when simplex or "direct" communication will work. Repeaters are a great place to make initial contact. Then, move to simplex if possible.

Ask the other operator how well you're being received. Hams traditionally give numeric "signal reports" that only *may* tell you something useful about how understandable your signal is. If in doubt, ask for a *verbal* report. What you really want to know is whether your signal is usefully distinct from noise and interference. As long as you can be easily understood, you're running as much power as you need.

Switch to a mode that makes better use of your transmitter power. This suggestion is only so useful, because many of us have preferred modes that we use through thick and thin (or until the contest ends or we crack the DX pileup!). But if just making contacts is your goal, some transmission modes may "get out" better than others, especially if you have a poor to modest antenna. Watt for watt, AMTOR and Morse code generally get through better than SSB, and SSB generally gets through better than AM.

The digital modes, such as RTTY, AMTOR, ASCII, and packet, show a similar relationship between bandwidth and power required. Generally, the faster you send information, the more transmitter power is required for the information to be successfully received. This is why you can sometimes get through with the old 45.45-baud (60 WPM) RTTY when 75-baud RTTY or 110-baud ASCII *cannot* get through. Fig 1 can help you understand the reason for this. The faster you send information, the wider your signal spreads across the spectrum. The receiving station must then widen its selectivity for best reception of your signal, and this lets in more noise. So you usually won't need as much transmitter power for successful operation at lower data rates.

Experiment with reducing your transmitter power. If your signal is well above the noise, cutting your power by half should be barely noticeable to the other station. Cutting your power to one fourth of its original level will reduce your signal about one signal-strength (S) unit. You'll probably find that your results differ very little between running 100 watts and 25 watts output. If your radio has a **HIGH/LOW** power switch, switch to **LOW** to see if high power is necessary. You may *still* hit that repeater "full quieting"! If the other operator has trouble hearing you, you can always just turn the power back up.

Improve your antenna. As mentioned earlier, changing to a directional gain antenna can make all the difference in the world. Which antenna should you try? Ask around! Experimenting with antennas is something many hams love, and earfuls of advice—not to mention armloads of books, including *The ARRL Antenna Book*—are available on the subject.

Use Your Knowledge on the Air

In this article, we've considered appropriate transmitter power mainly from the standpoint of achieving useful signal readability at the receiving end of your contacts. *That's* the purpose of the FCC's minimum power rule. This article presents basic transmitter power issues and helps you take your first steps toward compliance. What's the bottom line? Don't let your **RF POWER** control be a stranger!

Turning Up the Wick on Smart Power

You and the other hams you contact do much more than just control how many watts your radios run. You also work together to overcome varying conditions by changing *how* you use your radios' power. Listening during pauses in your own transmission, you can stop sending when you hear interference—or when you hear the other operator tell you to wait—and start sending again when the interference goes away. If you miss the other operator's name, you can ask for a repeat—as many times as necessary.You can speak faster or more slowly, change how fast you send, or change how you space words or letters. You can change mode or frequency, or switch to another antenna. *And* you can adjust your transmitter power.

All of these possible actions result from one main *perception*: that information has been received incorrectly or lost. All of these actions result from one main *decision*: whether the information received incorrectly or lost must be recovered.

For many years, people—radio operators—handled all aspects of this job. We perceived when information was missing, decided whether lost information had to be recovered, decided how to recover it, and did whatever was necessary to do so. By 1980, radio amateurs had added mechanization to this process by adopting AMTOR and packet radio—modes that include automatic error detection and/or correction. These modes improve accuracy mainly by sending duplicate data. Their error-reduction protocols do not involve automatic transmitter power control.

Now, in 1993, smart power—transmitter power automatically and adaptively adjusted to the optimum necessary for solid communication—is alive, well and growing in Amateur Radio. As you read this, hams are using at least one digital communication protocol—CLOVER—that includes automatic transmitter power control as one means of maximizing throughput. If you're already running AMTOR, you can get your feet wet in automatic power control without rolling in CLOVER: Beginning on page 25 of this issue, Carl Gregory, K8CG, shows how you can add simple, experimental automatic power control to AMTOR modems.

Operating with the minimum necessary power isn't just about using the *right* power. It's also about using the right power *right*. Automatic power control stands to help us do even better at achieving that goal.—*WJ1Z*

Strays

THE PLACE FOR SPACE

◊ If you've come in contact with a news item concerning Amateur Radio in Space and education, the students at Chaminade College Preparatory School would like to hear from you. Rich Ensign, N8IWJ, has turned over responsibilities of the Radio Amateur Satellite Corporation (AMSAT) Education News to this active physics group, headed by Dave Reeves, KF6PJ. Please send Amateur Radio in Space news items of educational interest to Reeves at 7500 Chaminade Ave, West Hills, CA 91304.

CABLE COMPANY HELPS HAMS

◊ When the Edisto ARS needed a backup VHF repeater, it requested $1400 in financial assistance from Jones Intercable of Orangeburg, South Carolina. When Hurricane Hugo struck the area in September of 1989, the Orangeburg repeater was the only one operating in that part of the state. Jones provided the needed funds for the 146.805-MHz RACES machine, which went into service on November 16, 1992. It provides vital coverage from Charleston to Columbia and throughout Orangeburg, Calhoun and Bamberg Counties.—*Jim Brown, WM3O, Orangeburg, South Carolina*

Predicting Transatlantic 50-MHz F-Layer Propagation

When does 6 meters open to Europe? A simple forecasting method, based on analysis of three years' transatlantic openings, provides a reliable guide to 6-meter propagation.

By Emil Pocock, W3EP
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Six-meter operators, perhaps more than those on bands below 30 MHz, are anxious to know when F_2 propagation is likely to open the band to other parts of the world. This happens so infrequently at 50 MHz, even during the peak of the 11-year solar cycles, that accurate predictions of when and where openings might occur are invaluable for planning when to devote precious time to 6-meter operating. Unfortunately, both long-term forecasts and predictions based on current solar data are so poor at 50 MHz that they are nearly useless for day-to-day planning. That's because the required 50-MHz data based on past experiences apparently have never been systematically collected and analyzed. Such an analysis is presented in this article with a simple, new graph for forecasting openings between New England and Europe—probably the most common 50-MHz DX path in recent years.

Limitations of Available Forecasts

Long-term forecasts geared for HF (3-30 MHz), such as those published in *QST* and other Amateur Radio journals, provide only the most general indications of when F_2 propagation may reach 50 MHz.[1] These forecasts suggest that openings between the Eastern US and Western Europe would occur during morning hours (1200-1700 UTC) between late October and early March, when the 2800-MHz solar flux is very high. How high is not certain. Even with a predicted average solar flux of 234 (the highest forecast for any November-February period during Cycle 22), the chart published in *QST* made it appear that the chances for 50-MHz transatlantic propagation were very small. See Fig 1. In fact, European openings were common from New England when the daily solar flux exceeded 200, but the probabilities for any particular flux value could not be determined from these charts.

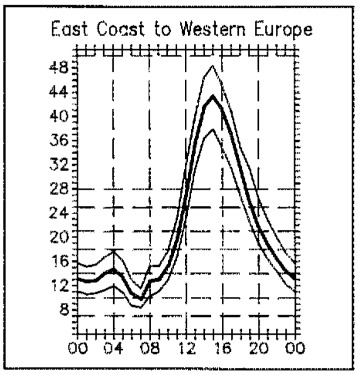


Fig 1—Propagation predictions for East Coast to Europe for the mid-November to mid-December period for a smoothed 2800-MHz solar flux of 234. The upper line indicates the highest possible frequency; the center bold line is the maximum usable frequency; and the lower line represents the optimal traffic frequency. This chart appeared in November 1989 *QST*, page 60.

Amateurs have generally assumed that increasingly disturbed geomagnetic conditions, as measured by the A and K indexes, progressively degrade propagation. The effects of geomagnetic conditions are not accounted for in forecast charts and are largely undocumented as high as 50 MHz.

Personal computer programs such as *Miniprop*, *IONCAP* and *MINIMUF*, generate more precise forecasts for HF propagation using current solar flux values, but they, too, are quite misleading at 50 MHz. Two examples based on the Boston-to-London path during late December demonstrate their limitations. With a solar flux of 210, *IONCAP* predicts that the highest possible frequency (HPF) will be just 43 MHz.[2] The actual HPF *must* be much higher than that, because the 50-MHz path from New England to Europe was open on about 40% of the days when the flux was 210! Similarly, *Miniprop* indicates that the maximum usable frequency (MUF) between Boston and London reaches 45 MHz only when the flux is at least 300,[3] but this, too, vastly underestimates amateur experiences. Even daily predictions made with popular computer programs clearly fall short at 50 MHz.

An Empirical Approach

One way around the misleading results of these programs is to base predictions on actual observations of transatlantic paths at 50 MHz. By comparing past 50-MHz propagation conditions with solar and geomagnetic indicators, it is possible to construct a new set of empirically based forecasts. Such an analysis depends on having a complete and reliable set of observations for the path in question—in this case New England to Western Europe. The operators of five of the most active 6-meter stations in New England during the peak of Cycle 22 kindly provided their logs for this study: K1TOL, WA1OUB, W2CAP, K1JRW and W3EP, all of whom were located within a 150-km radius of Boston (see Table 1). They were similarly equipped with medium-sized (5- to 8-element) Yagis, favorable locations, used 500- to 1000-watt transmitters, and were among the East Coast stations heard most often in Europe.[4]

All five operators listened toward Europe

Table 1
Stations Contributing Log Information for the Study

Station	*Location*	*Grid Square*	*Distance to London (km)*
K1TOL	Southern Maine	FN53	5085
W2CAP/1	Cape Cod, Massachusetts	FN41	5260
WA1OUB	Southern New Hampshire	FN43	5270
K1JRW	Central Massachusetts	FN32	5355
W3EP/1	Eastern Connecticut	FN31	5390

[1]Notes appear on page 34.

at 50 MHz most mornings of three 120-day study periods, November 1 through February 28, 1988-89, 1989-90 and 1990-91.[5] Few European openings are likely to have escaped the attention of all of them. Operating practices typically included long periods of listening, but also of calling CQ when the band seemed closed. The latter technique was often effective in initiating contacts, because the better-equipped American stations were likely to be heard in Europe before Europeans could be heard in New England.[6]

Once the band opened, the New Englanders tended to make lengthy runs of European contacts for as long as conditions remained favorable. The logs included notes on beacons and stations heard, but not necessarily worked, which were also extracted as data. Log entries are thus reliable propagation indicators, especially in determining the onset of openings.

The compiled data base incorporated the date with the earliest and latest times Europeans were recorded in any of the five logs. Only stations in Ireland, the countries of the British Isles, France and the Netherlands were included in order to limit the geographical area on the European end of the path. Signal strengths and other subjective evaluations were not considered. These data were then combined with corresponding daily solar flux and geomagnetic A-index values.[7]

Openings occurred on 133 of the 360 days (36.9%), during which the solar flux averaged 203.6 and ranged from 135 to 359. More than two-thirds of the openings were recorded in at least two logs and nearly half appeared in at least three logs. The three stations closest to Europe—K1TOL, W2CAP and WA1OUB—made nearly all the unique reports, most of which were brief. They also recorded more European openings, generally heard openings earlier than other stations in the region, and reported them for longer periods of time. Logs from VE1YX in Nova Scotia (4705 km from London) showed that he had an even greater advantage, often working Europeans an hour or more before they were heard in New England.

Relation to Solar Flux

Many different kinds of analyses were run on these data. The simplest and most revealing can be reduced to a single graph—Fig 2. It shows the portion of days that European stations were reported in any of the five logs as a function of solar flux. A curved regression line is superimposed on the data for ease of interpretation, but it's unknown if this is the "true" line in spite of its reasonable fit.[8] In any case, the graph clearly indicates a strong positive relationship between solar flux and 50-MHz propagation between New England and Western Europe.[9]

The significant feature of this graph is its application for predicting future conditions. The probability that the 50-MHz path between New England and Western Europe will be open between November and February can be read directly from the figure. When the flux is 220, for example, the graph suggests there is a 50% chance that the band will be open.

Probabilities of future conditions derived from this graph must be used with some caveats because of the way it was derived. It represents past performance based on a reasonably reliable set of data collected over three years. Nevertheless, there is no guarantee that future conditions will duplicate the past.

Predictions are likely to be high for the beginning and end of the November-February period and low for the optimal time of late December and early January. The predictions may also be somewhat high for any single location in New England, because openings can be geographically selective even within a 150-km radius of Boston. In addition, the graph may underestimate propagation to Europe generally, because not all continental European stations that appeared in the logs were included in the data base. Even given these cautions, the graph probably represents the best available aid for predicting 6-meter European openings from New England.[10]

Effects of Geomagnetic Conditions

It is widely assumed that disturbed geomagnetic conditions adversely affect F_2 propagation, especially over polar and near-polar paths. Contrary to these expectations, analysis of log data and corresponding daily A indexes, a common measure of the condition of the earth's magnetic field, indicates that geomagnetic conditions made very little difference to propagation when the A index was less than 30. The band opened on about 37% of the days whether the magnetic field was quiet, unsettled or active. See the summary in Table 2. Propagation was adversely affected only during geomagnetic storms (A index of 30 or higher), but the band still opened on 25% of those days. Average solar flux rose slightly with geomagnetic activity, but had only a slight tendency to moderate the effects of higher A indexes.

Single-Hop or Double-Hop?

One of the reasons that popular computer programs fail to predict 50-MHz propagation conditions between New England and Western Europe is that they assume that the 5300-km path must be spanned by a minimum of two hops of about 2650 km each.

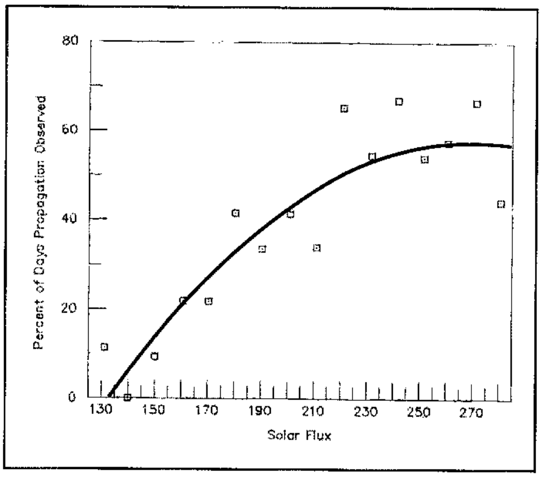


Fig 2—Portion of days 50-MHz propagation was observed from New England to Western Europe during three November-February periods, 1988-89, 1989-90 and 1990-91. Solar-flux values are grouped in *deciles* (130 = 130-139), with the exception of 280, which represents 16 data points between 280 and 359.

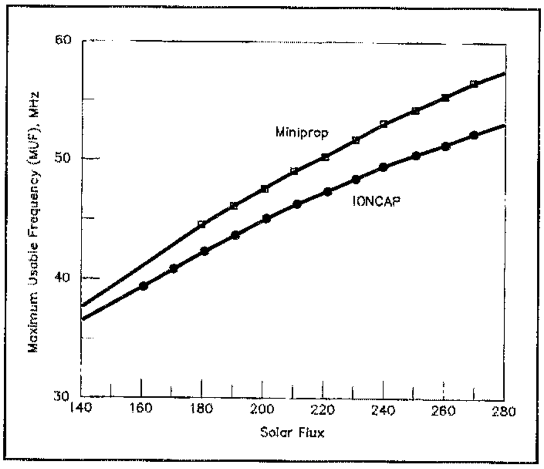


Fig 3—Maximum usable frequency predictions for a 3850-km path across the North Atlantic in late December. The distance is nearly the maximum *Miniprop* and *IONCAP* accept for single-hop analysis. It constitutes the center segment of the complete Boston-to-London great circle path of approximately 5300 km. See text for discussion.

Table 2

Geomagnetic Conditions and 50-MHz Transatlantic Openings, November through February 1988-1991

Geomagnetic Conditions (K index)	*Average Flux*	*Total Days Open*	*Percent of Days*	*Average Minutes Per Day*	*Average Minutes Per Opening*
Quiet (0-7)	198.0	147	36.7	48.2	131
Unsettled (8-15)	206.9	136	36.7	43.0	117
Active (16-29)	207.3	65	38.5	42.7	111
Storm (30+)	214.7	12	25.0	31.7	95

This is entirely consistent with ordinary assumptions that the F_2 layer can support hops no longer than 4000 km. The problem is that single 2650-km hops capable of supporting 50-MHz propagation require solar flux values much higher than required for a single hop of 4000 km, and certainly higher than those observed during the study period. Clearly this approach is flawed.

It is more likely that the New England-to-Europe path is covered in a single hop, despite the 4000-km maximum rule that is applied to HF propagation. Nevertheless, specialized studies have demonstrated that single-hop paths as long as 7000 km are possible at frequencies higher than 40 MHz with observed solar-flux values.[11] This solution could explain why 6 meters opened between New England and Europe with flux values much lower than those predicted by *Miniprop* and *IONCAP*.

In this regard, the results of *Miniprop* and *IONCAP* analyses of a single 3850-km hop, aligned with the midpoint of the 5300-km Boston-to-London path, are quite interesting. *Miniprop* predicts an MUF of 50 MHz with a solar flux of 220; *IONCAP* places the solar flux at 250 for the same MUF. See Fig 3. These flux values are well within the observed range during European openings. Indeed, the MUF can also be inferred from the observations presented in Fig 2, because the MUF indicates the operating frequency that F_2 propagation will reach on at least half the days. The band was actually open on 50% of the days when the flux was between 220 and 229! This coincidence with *Miniprop* and *IONCAP* calculations based on a 3850-km hop may provide some clues to the characteristics of longer single-hop paths, but its significance is unclear without further investigation.

Conclusions and Further Work

Analysis of 50-MHz F_2 propagation between New England and Western Europe over a three-year period has demonstrated that propagation forecasts designed for HF work do not provide suitable accuracy at 50 MHz. In the absence of better propagation models, short-term forecasts can be improved by basing predictions on past observations. Data extracted from amateur logs were used to derive a graph of solar flux versus the portion of days on which 50-MHz propagation existed between New England and Western Europe. This graph can be used directly to forecast the probability the band would be open on future dates based solely on daily solar-flux values. Additional analysis showed that daily A-index values of less than 30 had little effect on these probabilities. Techniques used in this study can be adapted easily to other paths where sufficiently reliable and complete long-term log data are available.

Acknowledgments

Many thanks to those who provided copies of their logs for this study. They were: Bob Billings, VE1YX; D. J. Butler, G4ASR; Lefty Clement, K1TOL; G. D. Eddowes, G3NOH; Richard Lax, G4AHN; Bob Mobile, WA1OUB; Rick Robinson, K1JRW; Henk Schanssema, PA2HJS; J. E. Tindle, G3JXN/GW3JXN; and Steve Wilson, W2CAP. I appreciated correspondence with Henk Schanssema, who has been using a similar approach to the problems of worldwide 50-MHz F_2 propagation. I benefited as well from discussions with Jerry Hall, K1TD, *QST* Associate Technical Editor [now retired], and appreciated the lengthy data he generated for me using *IONCAP*. Dr Michael Adams, Eastern Connecticut State University, helped create the graphics.

Notes

[1] See the prediction charts published in *QST*'s "How's DX?" column and E. Pocock, "Propagation Forecasting During Solar Cycle 22" *QST*, Jun 1989, pp 18-20, for a short discussion of their uses and limitations. General introductions to propagation are C. L. Bixby, "The Art and Science of DXing," *QST*, Jan 1979, pp 11-14; R. Healy, "Propagation Broadcasts and Forecasts Demystified," *QST*, Nov 1991, pp 20-24; and "The Role of the Sun," J. Hall, ed, *The ARRL Antenna Book* (Newington: ARRL, 1988), pp 23-22 through 23-27.

[2] In effect, the highest possible frequency indicates that there is only a 10% chance that the actual propagating frequency will reach or exceed 43 MHz. The likelihood of propagation decreases rapidly with frequency above 43 MHz.

[3] There is a 50% chance that the actual propagating frequency will reach or exceed the maximum usable frequency. In forecasting, the MUF does not refer to the maximum propagating frequency actually observed.

[4] See, for example, the many times these five calls appear among European reports in the *UK Six Metre Group Newsletter*, *Six News*, the Radio Society of Great Britain's *VHF/UHF Newsletter*, and its successor, *The Six Meter and Up DXer*.

[5] Very few 50-MHz transatlantic contacts were made via F_2 propagation outside these four months. Late December is the theoretical peak for such paths and this was generally born out by the logs.

[6] British stations were limited to 100 watts effective radiated power, or the equivalent of about 20 watts and a five-element Yagi. Other European countries, including France, had restrictions that were much more limiting than those in the US.

[7] Daily solar flux and A indexes were supplied by John McKinnon, National Geophysical Data Center, Boulder, Colorado.

[8] The regression line used is a second-order equation in the form $y = k + ax + bx^2$. A straight line also makes a pretty good fit, but it creates the mistaken impression that propagation improves indefinitely with flux. The data clearly cannot support this interpretation, as there appears to be some tendency for leveling out at flux values above 230.

[9] Analysis of the logs of five European stations on the eastern end of the study path, G3NOH, G3JXN/GW3JXN, G4AHN, G4ASR and PA2HJS, corroborated these general results. The European logs recorded fewer openings to New England and contained fewer entries, probably because the British stations were not as well equipped, there were many fewer Americans than Europeans on the air, and log entries indicated they were not as active as their New England counterparts. The resulting graph of European data indicated a lower portion of days open for given flux values, but the slope and shape of the curve was very similar to that based on the New England logs.

[10] The observatory at Penticton, British Columbia, source of 2800-MHz solar flux used by WWV and other reporting services, makes its measurements daily at 2100 UTC. Because the daily solar flux is available only several hours *after* European openings are likely, it might appear that the graph has limited value as a predictor. This is not quite the case. Relatively small changes in flux commonly observed from day to day do not significantly alter the predicted outcome on 50-MHz propagation. Thus the previous day's solar flux, or an average of the previous five days' solar flux, yield results that are nearly identical with same-day flux readings. The usefulness of previous days' solar flux as a predictor probably has little to do with cumulative effects of solar radiation on F-layer ionization, because the F layer is essentially recreated each day by ultraviolet radiation. Rather, it may be attributed to the imperfect relation between actual ultraviolet radiation and measured 2800-MHz solar flux, the uncertain effects of geomagnetic activity, and a number of other factors that are as yet imperfectly understood.

[11] F. Kift, "Single-Hop Propagation of Radio Waves to a Distance of 5,300 km," *Nature*, May 24, 1958, pp 1459-1460; and T. S. Rappaport, R. L. Campbell and E. Pocock, "A Single-Hop F_2 Model for Frequencies Above 30 MHz and Path Distances Greater than 4000 km," *IEEE Transactions on Antennas and Propagation*, Dec 1990, pp 1967-1968.

An Easy, On-Glass Antenna with Multiband Capability

Are you mobile on one VHF/UHF band, or two? Either way, this on-glass antenna design is for you!

By Robin Rumbolt, WA4TEM
1202 Wilkinson Rd
Knoxville, TN 37923

With every new car purchase comes the agonizing decision of where to punch the hole for my 2-meter antenna. Recently I purchased a dual-band transceiver, and the problem became where to punch *two* holes. I'd rather punch *no* holes at all!

An on-glass antenna seemed like the ideal solution. Such antennas couple RF through the windshield without the need to drill holes for cables and mounting hardware. Being a builder at heart, I designed an on-glass antenna to suit my needs. Not only does it feature the ability to disconnect the radiating element quickly (for car washes, etc), it has multiband capability, too!

Construction

I built the base of my antenna out of heat-sink material (see Fig 1). I happened to find a piece of bare aluminum heat-sink stock with long, straight fins. Each fin was spaced about ¼ inch apart. You can find similar heat-sink material at your local hamfest flea market. It's cheap and relatively easy to machine. You can also use aluminum channel stock, which is available from a variety of sources.

The first step was to cut out a piece roughly 1⅝ inches square and remove all but the two middle fins. The fins were a bit too tall, so I carefully trimmed them to ½ inch in height. I used a grinding wheel to round the corners and drilled 9⁄64-inch holes in the centers of both fins.

The antenna coupling plate is cut from a piece of sheet steel. Its dimensions equal those of the antenna mount. (Avoid using aluminum for the coupling plate, since it's very difficult to solder.)

The quick-disconnect assembly is made from two hexagonal brass standoffs just wide enough to fit snugly between the fins. One standoff has a hole threaded through its entire length. The other standoff has a threaded stub on one end and a threaded hole in the other. I carefully drilled a 9⁄64-inch hole through the open end of the second standoff. Using a #6-32 × ½-inch screw, I assembled the standoff to the base as shown (see Fig 2).

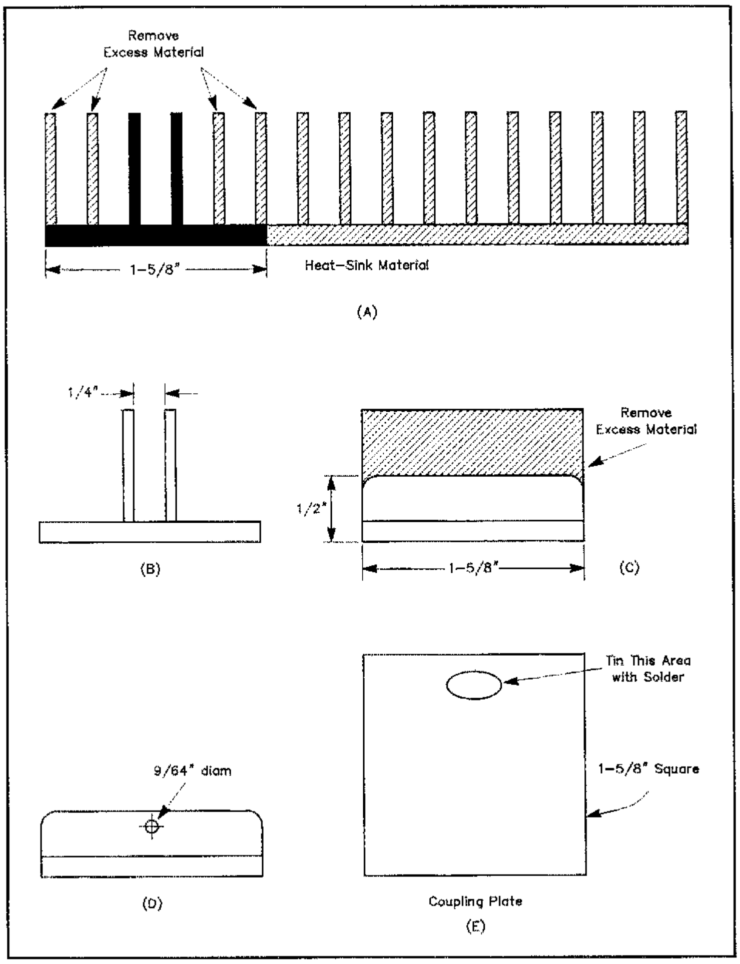


Fig 1—The antenna base is fabricated from a piece of heat-sink stock. Cut out a 1⅝-inch section and remove all but the two center fins (A and B). Round off the sharp corners of the fins and trim for a ½-inch height (C). Drill a 9⁄64-inch hole through the centers of both fins (D). The coupling plate is cut from a 1⅝-inch section of sheet steel. Tin a small area as shown at E.

The radiating element is made of 3⁄32-inch brass welding rod. I cut a #6-32 thread about ¼ inch up one end. This end is screwed tightly onto the first standoff. If you lack the tools to thread the rod yourself, use 1⁄16-inch welding rod and solder it to the standoff.

The total antenna length depends on the band you wish to use. See Table 1 for approximate lengths for various bands. As you can see in Fig 2, the finished section

Fig 2—The on-glass antenna is made of brass welding rod attached to two threaded standoffs (see text). Drill a #6-32 hole in the bottom standoff as shown. Using a #6-32 × ½-inch screw, nut and lock washer, secure the antenna to the base. The total length of the antenna (L) is measured from the tip of the welding rod to the mounting screw. Use the lengths listed in Table 1 and then trim as necessary to obtain a low SWR.

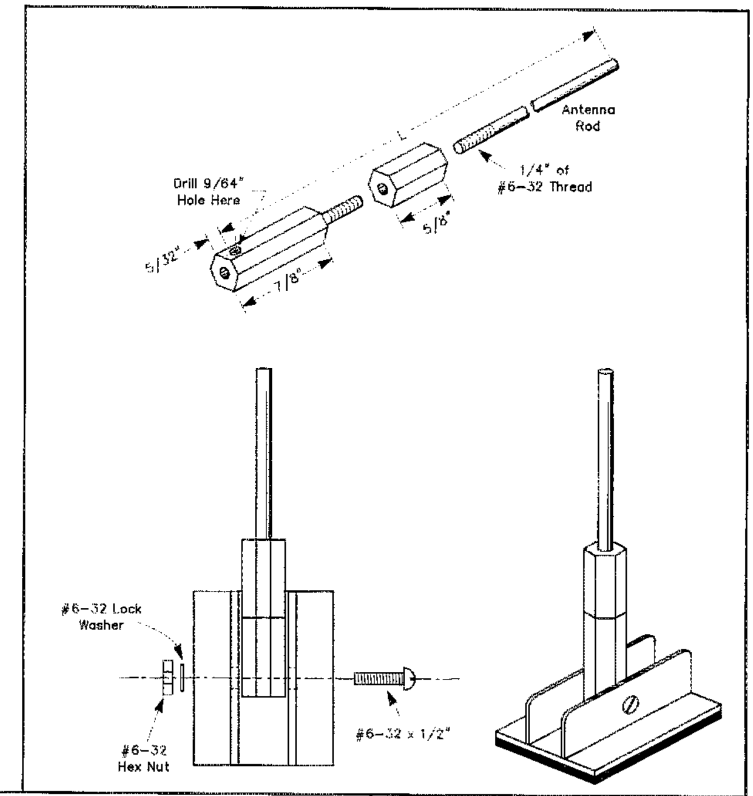


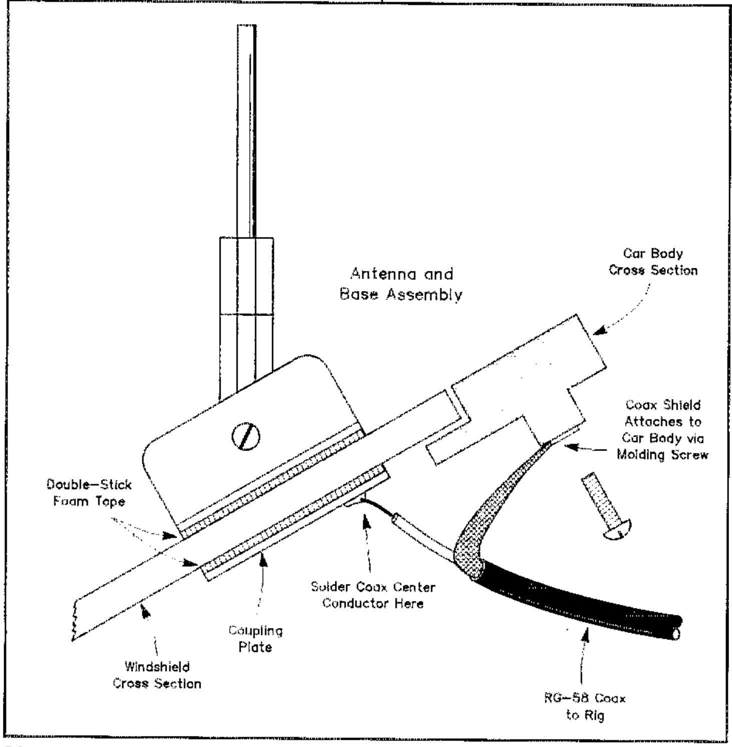
Table 1

Antenna Lengths for Various Bands

Frequency (MHz)	*Length (inches)*
145	23¼
223	15¹⁄₁₆
444	7¹⁷⁄₃₂
880 (cellular telephone)	11⅜ (¾ λ)
920	11 (¾ λ)
1296	7¾ (¾ λ)

screws onto the stub of the base-mounted standoff. Whenever I need to remove it, a few twists is all it takes!

The coupling plate and the antenna base are attached to the windshield with double-sided foam tape (Radio Shack 64-2361). One tape strip isn't wide enough to cover the base and the plate, so I applied two strips side-by-side. It was a simple matter to cut the strips, peel off the backing and apply the tape to each piece. Any excess is easily trimmed away. The important thing to remember is *not* to peel the paper backing from the tape until just before you're ready to install the antenna.



← Fig 3—Use strong foam tape to hold the base and the coupling plate to the windshield. The coaxial cable center conductor is soldered to the coupling plate. The braid is grounded to the car body via a nearby molding screw. The braid *must* be grounded at the antenna for proper performance.

Mounting

As you search for just the right spot to mount your antenna, bear in mind that you *must* ground the coaxial cable shield to the car body *near* the mount. In most cars, the top center of the front or rear windshields is best. Older cars usually have screws to attach the molding in these areas. These screws can often be used for grounding. If you own a newer car without strategically located screws, you'll have to install one yourself. In my Dodge Caravan, I drilled a small hole in the roof support (not the roof itself!) and used a small sheet-metal screw to fasten a solder lug in place. Whichever approach you use, check the screw with your VOM and ensure that it really makes contact

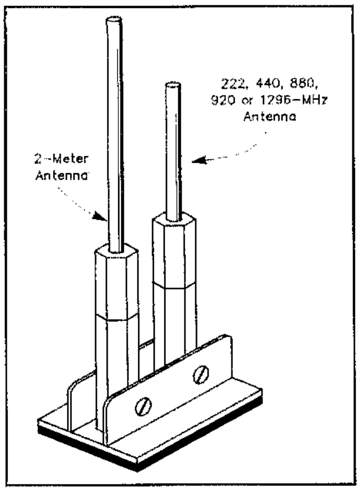


Fig 4—The multiband option. You can mount two antennas in the same base. This is ideal for today's dual-band, VHF/UHF transceivers.

with the chassis of your car. Many screws anchor in metal; but the metal isn't always grounded!

Hold the base to your windshield in the area where you intend to install it. Adjust the antenna until it is vertical, then tighten the nut. Remove the mount and spray paint the entire assembly black—or whatever color looks best with your car!

When the paint is dry, clean the glass thoroughly (inside and outside). Check your chosen antenna location one more time. Is it in the path of windshield wipers? If you open the trunk or hatch, will the antenna be crushed?

If everything looks safe, peel the paper from the foam tape and attach the base to the outside glass. Press firmly to ensure that the tape sticks to the surface. Attach the coupling plate to the inside glass directly opposite the base. Solder the center coax conductor to the coupling plate and connect the coax shield to the ground screw or lug.

Tuning

With an accurate SWR/power meter, make SWR measurements and begin pruning the antenna for the lowest SWR. In my 2-meter installation, with 50 watts forward power, the needle doesn't even wiggle in the reverse-power position (a 1:1 SWR). If you can't get the SWR below 3:1, check your coax ground at the antenna. This is often the culprit.

Multibanding

I saved the best for last! You can create a dual-band antenna by simply mounting a second antenna and quick-disconnect assembly on the same mounting base (see Fig 4). For example, here's a fancy system for hams who own cellular phones: Install 2-meter and 880-MHz antennas in the same mount. The 2-meter whip will do double duty on 2 meters and 70 cm, while the 880-MHz antenna is perfect for your cellular telephone. An antenna farm on glass! (This configuration must be fed with a single feed line and an appropriate diplexer must be purchased or homebrewed.[1])

Conclusion

I am extremely happy with the antennas I have made using this on-glass method. No external holes were necessary and the antennas disconnect easily. SWR is low on every band and the antenna's radiation efficiency seems to rival any hole-mounted antenna I've used in the past!

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[1] D. Jenkins, "A Simple VHF/UHF Diplexer," *QST*, October 1991, pp 18-25.

Robin Rumbolt, WA4TEM, has been active in many different phases of Amateur Radio for over 29 years. In addition to his Advanced class and General Radiotelephone licenses, Robin earned BS and MS degrees in engineering. He is currently employed as a staff engineer at Philips Consumer Electronics.

Strays

SMOKE TEST

◊ A friend of mine was trying to tune in the Jackalope Net on 40 meters, but there was a strong heterodyne right on the net frequency. His brand-new transceiver wouldn't notch it out, no matter how hard he tried. He tweaked the knobs and cussed the radio until his wife—also a ham—came out of the kitchen and asked, "Don't you know the difference between a heterodyne and the smoke alarm?"—*Lynn Bilyeu, K0ODF, Chadron, Nebraska*

FAMILY OF HAMS

◊ In late October 1990, our family decided to take up Amateur Radio as a "family project." I was all for it, but my older brother, Jay, needed some extra encouragement. By listening to the local Quakertown, Pennsylvania, repeater, we were able to find a hamfest. We stopped by and looked around with great anticipation. We picked up a copy of the ARRL's *Tune in the World with Ham Radio* and asked around for information on Novice classes. We were directed to Warren Erdman, K3ZXQ. Warren took our names and "signed us up" in his soon-to-start Novice class.

The class didn't last long for the family. My dad passed his Novice theory and code on November 15, 1990, after an extended study session while waiting (more than four hours) for his boss at the Orlando Airport. Upon returning home, Mr Erdman came over to our house to administer the tests. It was great to have our family "Elmer," Mr Dee Hester, N6BQR, there to sign the paperwork for submission to the FCC.

"Black Friday," November 23, 1990, wasn't so black for our family. It seemed as if it were "Good Friday" for us. Mom, Jay and I passed our Novice tests that day. Dad even passed his Technician theory test. Mr Erdman accidentally played a 13-WPM tape and I thought I was a goner. Dad had promised Jay and me he'd get us a radio if we passed our tests. Dad came through with two brand-new hand-held 222-MHz transceivers for the holiday. Our new licenses showed up in the mail the last week of 1990. Mom was issued KA3YFL, Jay was issued KA3YFJ, Dad received N3IVB and I got KA3YFK. We were on our way.

Although Dad wanted all of us to at least try to reach Technician class, Mom was the only one to do so, in January 1991. Mom was issued the new call sign N3IWL in early February. I think Dad had it planned that way. He had gotten Mom a dual-band (2-meter/440-MHz) radio, which she couldn't use until she was a Technician.

Instead of watching TV in November of 1991, I listened to the Gordon West, WB6NOA, theory tapes. Dad had gotten them for Mom. Each evening I'd go to sleep listening to "Gordo." Then, after reviewing the question pool with Dad, I took my test. I aced it! Wow, what a feeling! Dad said it would be easy to go on to General theory with the Technician so fresh in my mind. Again, I listened to "Gordo" as I went to sleep, reviewed the question pool, and again I aced the test!

Then I was on to the 13-WPM code test. Dad has a copy of the *Super Morse* IBM-compatible code training software by M. Lee Murrah, WD5CID, and that helped me get back in the swing of things. I hadn't used the code for almost a year. It took about two weeks (10 minutes in the morning and 10 minutes in the evening) to get up to speed to pass the test. In late 1991, Dad offered Jay and me an incentive to upgrade. We could add something to the shack if Jay upgraded to Technician and I upgraded to General by January 6, 1992. When Jay and I passed, we suggested adding packet to our station. Dad got us a multimode TNC, so now we log onto the WB3JOE PBBS, not to mention the other modes.

I was looking forward to getting on 40 meters and the Fish Net, now I can. Some of my friends are Tom, KA1UGH; Rick, KD1BR; Brian, WS3F; Joe Sr, WA1JKS; Jack, N2JTO; Ken, N1GPL; Alan, KF8PM; Derrick, WM1U; Marty, KA1IRA; Dave, N4FFG; and Fran, KA3WTF.

I enjoyed working the W3OK Christmas City special-event station on December 21, 1991. I made 40 contacts. Some of the contacts were my Fish Net friends.

Dad kept ahead of us: He took his Advanced and Extra Class elements in November and December of 1991. He was issued his KE3AH call sign on December 12, 1991.

February 11, 1992, brought three new licenses to our house. Jay was issued N3LQV, Dad received WV3M and I got KA3YFK (Young Fine Kid), only this time with "General" printed on it.

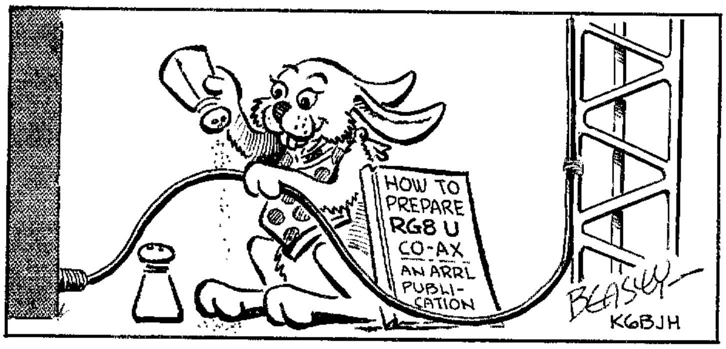
Mom may try to go on to General. I hope she does, so when I go away to school I can talk to her. I was eight years old (almost nine) when I passed my Novice tests. I was nine when I passed the Technician and General elements. I hope to upgrade to Extra Class within the next few months.

I want to thank all my friends who helped me and encouraged me along the way. Steve, N3ANW, took time to come and see me take my General code test; Bob, WF3H, saw to it I could work W3OK, all my Fish Net buddies (even though they are "a little older than me"), and of course my "Family of Hams."—*Christopher Breish, KA3YFK, Coopersburg, Pennsylvania*

Packet for Lunch

Many hams like to chew the rag, but how many get chewed right off the air?

By Carol Larsen, KA9HFA
129 N Ninth Pl
Sturgeon Bay, WI 54235



In the summer of 1989, I decided I'd finally enter the world of packet radio. With my busy schedule, I had taken the attitude that, although I was definitely interested, I didn't have the time to decide on equipment, figure out how to connect it, and learn to use a new mode. What a mistaken idea—packet is the perfect mode for busy hams to contact their ham friends who are also busy or keep unusual work hours, enabling us to communicate at our own convenience. Usually faster than the US Postal Service (and often more reliable), packet radio doesn't interfere with your schedule—it lets hams take *better* control of their time. I already had a computer and an extra 2-meter hand-held transceiver, so it was just a matter of selecting a TNC. Connecting the equipment was no problem, either. The same afternoon the UPS truck brought the TNC, I was making contacts. Since then, I've made new friends in the packet ranks and kept in better contact with my local friends. I'm also able to handle traffic that I was previously unavailable to take.

While attending several hamfests, I persuaded a close friend to join the packet ranks, which would greatly facilitate our planning trips to hamfests. (For good reason, my local phonetic nickname is "Ham Fest Addict.") I've noticed that I've become so dependent on packet communications that I'd feel cut off from a major part of my ham world if anything went wrong with my station. I'm at work during the 8 AM 2-meter Good Morning Net and the weekday morning 20-meter skeds with the Door County "snowbirds." Without packet, I'd miss all the local happenings, except on weekends when I can join the groups on the air and attend the weekly HamSat breakfast.

Cut for Maximum SWR

One afternoon, I rushed home from my job and didn't check my mailbox as I normally do. Typical of my busy schedule, I was in a hurry to get to an out-of-town basketball game. By the time I returned that night, it was late when I checked my mailbox—a perfect example of why packet is important to me. Oddly enough, there was no mail in my mailbox and no stations listed on my TNC's Heard list since about noon that day. I tried to connect to some stations on 145.01 MHz, but to no avail. I switched to 145.05 and connected to a station about two miles east of me, so I was satisfied that my equipment was working. Maybe it was the Door County ARC mailbox and node on 145.01 that had a problem. It was almost midnight, so I didn't call anyone to find out.

The next day, I hurried home from work to attend a church supper, so again I didn't have time to check my mailbox in the afternoon. When I returned, there was a message on my telephone answering machine from one of my packet friends who was wondering if I was all right because my packet station had been off the air for the second day in a row and I hadn't picked up the message he'd left for me in the club mailbox. I called him on our local repeater to assure him that I was fine, but apparently my packet station wasn't.

Armed with two other 2-meter rigs, I headed for the shack. Because I had been able to access my own mailbox and make a contact on 145.05, my software, cables, TNC and antenna must be functioning properly, so it seemed that the problem had to be in the ability of my packet-dedicated hand-held transceiver to function on 145.01. After getting the same results with all the 2-meter rigs, I was scratching my head.

After considering the proximity of the only station I was able to contact and the distance to the station that was unable to contact me, I decided that I was transmitting a weak signal and receiving only strong, close signals. I checked the SWR on my antenna and feed line, and found the needle pinning the end of the meter. What a time of year to have antenna problems! It wasn't so bad last summer when the birds sat on the radiator of my old 2-meter ground plane and bent it over. In good weather it's not that hard to find someone willing to climb the tower and install a different antenna. My options in January, other than the unthinkable—being off packet for a while—seemed to be that I'd have to run coax down two levels of stairs in my trilevel house from my only other 2-meter antenna, a ¼-wave 2-meter ground plane made from white house wiring and fed with white coax. It hangs in my bedroom on the third level, while my shack is on the garage level.

Pondering the options, I went back upstairs, stepped out on the patio in the chilly, windy darkness, and shined a flashlight beam at the 35-foot level of the tower where the new bird-proof WB9YUY coaxial/PVC pipe 2-meter antenna is side mounted. It appeared intact. I then turned to the bottom of the tower where the coax comes from the basement and discovered it had been cut completely in two! I must have made my one connect with only a part of my coax for a "long wire" antenna.

Feeding on the Feed Line

Who would cut my coax and why? I've never had any TVI complaints from my neighbors and vandalism isn't prevalent in my neighborhood. Then, upon closer examination of the coax, I discovered which neighbor was to blame. It had to be the rabbit that lives under the patio adjacent to the tower. The coax for the packet station, added most recently, passes through the basement wall in a different hole than does the bundle of other coax and cables, so fortunately it was the only cable attacked. He had first nibbled away the outer insulation and then, after several serious attacks, had made a clean cut in that familiar angle pattern he uses to prune the shrubs I don't want pruned. How appropriate that it must have happened at lunchtime when the TNC's Heard list quit adding stations heard!

I love nature's creatures and have put up with the shrub pruning, but was this excusable? Pulling the loose coax ends into the basement, I headed for my parts cabinet and soldering iron. A couple of PL-259s and a barrel connector put my packet station back on the air. As for the rabbit's taste for coax, he'll have a hard time penetrating the split-loom tubing that now covers the cables between the basement wall and the tower! QST

The Lure of Classic Radio

If you think old tube rigs are doorstops or boat anchors, think again! Vintage radios are in vogue more than ever before. So get with the program!

By Marty Drift, WB2FOU
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and Jim Musgrove, K5BZH
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The increasing popularity of vintage radio gear isn't all that surprising. An interest in things from the past is only natural, and Amateur Radio, with its rich tradition, is no exception. Remember, most of today's classic cars were once second-hand vehicles. And so it is with radio gear.

The popularity of vintage radio has been increasing over the past few years, and the signs are becoming more evident: Note *QST*'s Ham Ads. For some time there has been an "Antique-Vintage-Classic" section. Classic radio enthusiast Barry Wiseman, N6CSW, launched *Electric Radio* in May of 1989, a magazine dedicated to Amateur Radio of the vacuum tube era. Collins collectors have had *The Collins Collector's Magazine* since June of 1991. (See the sidebar, "Vintage Radio Resources" for info about the magazines, and see "Classic Rigs and Amplitude Modulation: Perfect Ham Radio Partners," by Paul Courson, WA3VJB, and Steve Ickes, WB3HUZ, in the February issue of *QST*).

Classic Radio Beginnings

Interest in vintage radio comes about in several ways. Some hams purchase vacuum tube gear to duplicate their first ham setups. Others buy rigs they could only dream of owning when they were youngsters. A few hams acquire older gear because they can't afford the new equipment they desire. Some develop an interest while repairing older equipment they pick up at swapfests to resell at a profit to help support their hobby. The paths to vintage radio enjoyment are numerous and varied.

A few vintage operators would like to permanently move technology back three or four decades, but most collectors appreciate modern technology and its advantages: Accuracy, operating ease and reliability are only a few examples.

Modern VHF equipment makes mobile operation a pleasure. Working another station 150 miles away with one watt on 1296 MHz wasn't even a dream in the '50s! We will continue to look forward to technological advances.

There is a charm to vintage radio, though. The older equipment possesses a certain character that is missing in modern radios. The same is true of a steam-powered locomotive compared to a modern diesel. Glowing fire bottles (vacuum tubes) add an almost spiritual dimension—sort of like the whistle on a steam engine. Ever seen the bumper sticker, "Real Radios Glow in the Dark"? Applying filament voltage is akin to stepping into a time machine and temporarily stepping back to another era.

For old-timers, antique radios bring back fond memories, and newer hams can experience the characteristics of vacuum tube gear and get a feel for a bygone era. Knowing how to tweak the phasing control on an old receiver to obtain single-signal selectivity, or how to tune a vacuum tube transmitter, requires skills no longer needed for modern equipment.

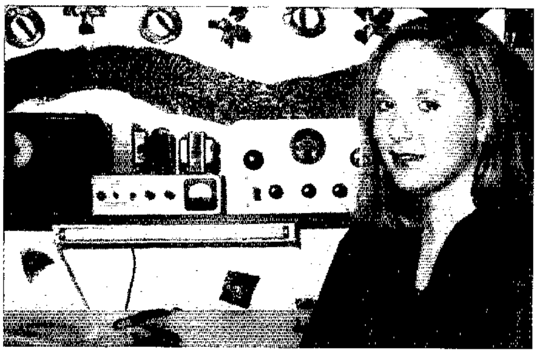
If you have an interest in this facet of our hobby, now is the time to get involved. Vintage equipment is in demand and availability is down. Prices are likely to increase significantly from now on.

What's Available?

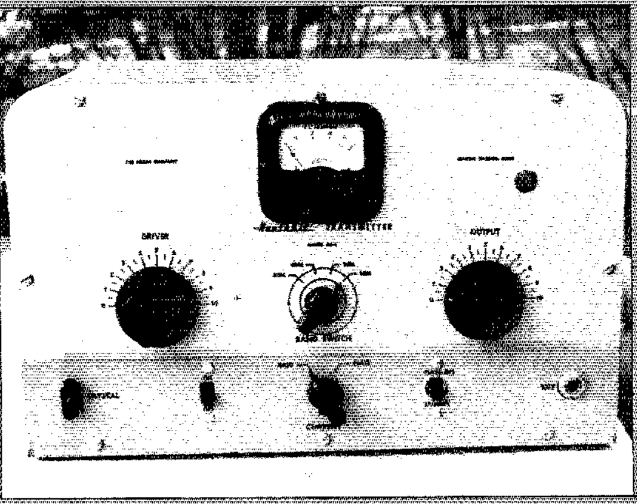
Some will balk at the thought of not having room to house the old "boat anchors," but the perception is not entirely founded. Not all vintage equipment is physically large. A Hallicrafters SX-96 receiver, for example, is about 18 inches wide, and a Johnson Ranger transmitter about 15 inches. An operating desk of 36 by 28 inches will easily accommodate the pair. Some Novice stations, such as the Hallicrafters S-38C receiver and Heath AT-1 transmitter, are even smaller.

The truly grand rigs were few. The Collins KW-1, a refrigerator-sized monolith, is an awesome sight. Only 150 or so were made. A few more than 1600 Collins KWS-1s were produced, and records show that only 402 Johnson Desk Kilowatts were manufactured. These big rigs have a suitably large following, however, and they're in hot demand.

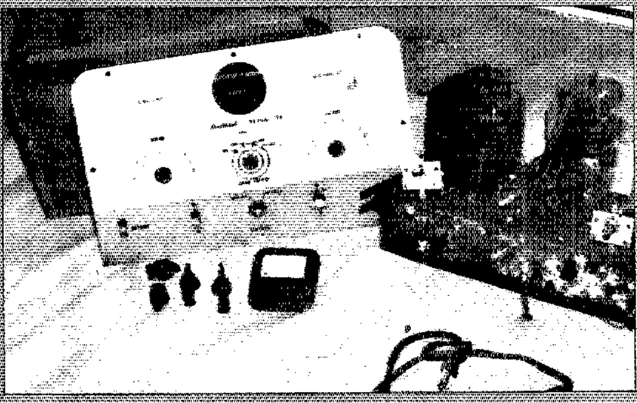
David Bertman, AB7B, has quite a tidy classic station. The goodies include a Heathkit DX-100, a Collins 75A-4, a Hallicrafters HT-44, a Drake 2A, and a Johnson Matchbox antenna tuner. *(photo courtesy of AB7B)*



The coauthor's daughter, April Musgrove, with a homemade 6AG7-6L6 transmitter and a Hallicrafters S-38B receiver. This is a typical mid-'50s Novice station. *(photo courtesy of Jim Musgrove, K5BZH)*



Coauthor Marty Drift's restored AT-1 Novice rig shows the amazing transformations that can take place in the vintage radio biz. The little Heath transmitter isn't perfect, but it's light years ahead of the way it looked originally. You can see why it only cost a dollar! (*photos by Marty Drift, WB2FOU*)



Some serious collectors have literally created museums in their houses. Some collections center on one manufacturer, such as Collins, Hallicrafters, Heath or National, and a few of these folks have rescued prototypes of equipment that, for various reasons, never went into production. A few examples are the Collins KWM-3, Elmac ATR-4, the Johnson Avenger sideband transceivers and the Hallicrafters HT-36 sideband transmitter.

There is a growing interest in AM gear, particularly from the '50s. Johnson Viking IIs, Heath DX-100s, Heath Apaches and Johnson Rangers fit nicely into this category.

Interest has also been generated in vintage sideband gear. The Central Electronics 10A was the most popular of the early sideband exciters. The 10B and 20A are later revisions. Thousands of these transmitters were sold and many still exist. They also do an excellent job on CW. A Central Electronics 20A is perhaps the best buy in the vintage transmitter market.

Many Novice transmitters are prized possessions. The Heath AT-1 may sell for well over $100. Philmore's NT-200 commands a similar price. Other Novice rigs such as the Johnson Adventurer, Heath DX-35 and Globe Scout can be had for a lot less.

One category overlooked by many is home-brew equipment. These units can make excellent display items. Low-power transmitters with one or two stages were common in the '40s and '50s. Homespun units were typically built on aluminum chassis, but some were assembled in wooden boxes, bread pans or even coffee cans. They occasionally appear at swapfests, but few collectors seem interested. How unfortunate—the home-brew rigs reveal a lot about the culture of Amateur Radio.

Early solid-state equipment is typically ignored, too. Don't forget, the transistor is well over 40 years old, and the integrated circuit is not far behind. Early equipment in this category includes the Hallicrafters FPM-300 and the Sideband Engineers SBE-33. Even better is the Hallicrafters FPM-200, which is a rare find.

Many of the vintage receivers still do a good job on the ham bands. Some that come to mind are the National HROs (definitely classics), the Collins 75A series, the Hallicrafters SX-101, Heath Mohawk, Drake 2B and National NC-300.

The Hallicrafters S-76 is a sleeper among early '50s receivers. Rather than the typical single-pole crystal filter at 455 kilohertz, this receiver relied on a 50-kHz IF amplifier that has one of the best passbands found in receivers of its time. The skirt selectivity is very good. In the 500-Hz or 1-kHz position, it does a nice job as a CW receiver. The SX-96 and SX-100 are upgrades.

If a vintage station doesn't interest you, then a display of keys, microphones, tubes, or crystals might be your cup of tea. These displays don't have to be large to be interesting. A collection of 10 or 20 tubes on a bookshelf adds class to your ham shack. A couple of vintage telegraph keys will further enhance it.

Where to Find Vintage Gear

So where do you find this stuff? Hamfests are probably the best places to search. Magazine ads, antique stores and garage sales are other sources.

If you aren't accustomed to playing technician, it would be wise to have a friend who is familiar with vintage sets accompany you. Sniff the transformers. If they have a pungent odor, they are probably defective. Check for missing parts. It's usually wise to avoid sets with obvious unauthorized modifications. You might want to keep it in mind as a parts set, though.

When purchasing a vintage rig, try to get the original owner's manual. Larger public libraries often have *Sam's Photofacts* and *Rider* manuals which document many popular receivers. Vintage collectors can sometimes provide you with copies of manuals, and some dealers sell original manuals or photocopies (these can be expensive, so be sure to shop around).

A Tune-Up is Probably in Order

After purchasing a vintage set, you are likely to have some work ahead of you to put it back into good operating condition. Blow the dust out of the set. Clean the panel smudges. Check the tubes. Filter capacitors are prone to be defective, and paper coupling capacitors are usually leaky.

Most vacuum tubes are readily available on the used market and are reasonably priced ($1-$7), but a few have become difficult to find and command rather steep prices. The 7360, used as a balanced modulator in a lot of early '60s sideband equipment, typically sells for $25 to $40 on the used market. Similarly, the 4D32, used as an RF amplifier in several 100-watt transmitters such as the Collins 32V series, the Hallicrafters HT-20 and the Johnson Viking I, goes for $25 to $50.

Various surplus houses sell many of the components, including tubes, needed to keep the vintage radios alive, and finding the stuff isn't too difficult (see the "Resources" sidebar).

If you aren't familiar with vacuum tube theory, don't assume that you won't be able to gain the ability to understand the circuits. Ask one of the old-timers in your radio club or purchase any older *ARRL Handbook* and do a little studying.

A few cautionary notes are in order:

- Mercury vapor rectifiers require special attention. If the equipment hasn't been used in a long while, filament voltage should be applied for a few hours prior to applying plate voltage. Refer to an older *ARRL Handbook* for details.
- Some older Novice equipment is "ac-dc" (transformerless). Be sure the cabinet and any exposed metal parts are isolated from the ac line before using or working with them. For an extra margin of safety, use an isolation transformer with these sets. The Hallicrafters S-38, a typical ac-dc receiver, requires only 30 watts of power, so large isolation transformers aren't necessarily required.
- The operating voltages found in tube-type gear should be respected. Exercise care in troubleshooting live sets. The old rule was not to work on any live circuit that exceeded 300 volts. The high-voltage supply of a 50-watt Novice transmitter is more than enough to electrocute you. Don't be paranoid, but be conscious of the dangers. Don't work on energized vacuum-tube equipment when you are tired.

Operating Your Vintage Gear

If your interest is operating vintage equipment on today's ham bands, there are some things to consider. You might start with a middle-grade-or-better receiver and an AM/CW or CW transmitter in the 50-watt range. If your interest is sideband, think about a 10- to 20-watt exciter (and perhaps plan to build a small linear amplifier).

AM operation is often found in the 75- meter band on 3880 and 3885 kHz. The 10-meter AM watering hole is from 29.0-29.2 MHz. You will note that the 75-meter group tends to filibuster and that 10 meters features a faster push-to-talk type of operation.

Vintage CW activity takes place throughout the CW subbands, and vintage sideband transmitters can be heard anywhere in the phone subbands. A vintage SSB net meets Sunday afternoons at 2000 UTC on 20 meters near 14.295 MHz.

Vintage sideband transmitters may require more bandwidth than their modern counterparts. If you are sensitive to this, use

A Vintage Set of My Own

By Kirk Kleinschmidt, NT0Z
QST *Assistant Managing Editor*

The flood of vintage radio articles coming through the *QST* editorial office finally overcame my resistance—I had to "restore" a set of my own. Having dealt with "vintage" radios as a kid (crusty, musty and mostly junky ones), however, I knew I couldn't settle for anything less than a "real" vintage radio—one that could be used on the air and enjoyed. (Working for *QST* and having access to all of the latest and greatest rigs had spoiled me. I didn't want some drifty, buzzy and scuzzy rust-bucket, but I didn't want to spend a fortune, either....)

Well, as luck would have it, one of the guys at HQ wanted to part with his National HRO-7—a real classic, and one with a lot of potential—for only $50! He even said it worked (three years earlier), which was true, after a fashion. A set of plug-in coils, a spare power supply, an owner's manual, a matching speaker and a handful of extra dial calibration strips rounded out the deal. As they say, the price was right.

I took my battleship-gray beauty home, hoofed it up to my third-floor (penthouse?) apartment and plopped it (gently) onto my smallish workbench (which shares half of my home office and ham shack).

Before initiating the "smoke test," I grabbed a can of "air" from my darkroom and "blew off" all of the controls, the inside (under the handy flip-top lid, where the tubes live) and underside of the chassis (where all of the melted wax capacitors live).

So far, so good. The inside was very clean—always a good sign—and because the 1946-vintage HRO-7 had plug in coil sets, there were no crusty, multi-wafer bandswitches to worry about. That was a bonus I hadn't considered!

After the dust settled, I gently pulled out each tube and applied a drop of Stabilant 22 (the electrical connection miracle juice) to each pin before plugging it back in.

After giving the power supply and speaker connectors the Stabilant 22 treatment, I plugged the power supply into the ac power strip, stood back a ways (a handy thing to do when smoke testing old tube gear), and flipped on the filament power switch.

The meter lamp popped on (a good sign), the tubes lit up (an even better sign) and nothing hissed, snapped or exploded (the best sign of all). After a minute or so with no pyrotechnics, I carefully flipped the high-voltage toggle.

No prob! Other than the brain-numbing roar of 6V6-powered band noise (the audio gain control was maxed), everything looked and sounded fine! As I tuned the quintessential HRO tuning knob (these are really cool), frequencies changed, but everybody on 75 meters sounded like Donald Duck! No matter what I did, the BFO (called a CWO—CW Oscillator in those days) wouldn't work. A couple of shortwave AM broadcast stations near 4 MHz sounded okay, so I knew the radio was at least functional.

After disconnecting the power and letting the filter caps settle down for a few minutes, I started poking around "underneath."

Sure enough, the 220 kΩ screen resistor in the BFO circuit had "gone infinite." It literally tested as an open circuit on a meter that can measure 20 million ohms of resistance. Old carbon resistors do that sometimes. It's part of the fun....

After replacing the resistor with a similar one (so I was off 30 kΩ—all of the other resistors in the set were, too. No big deal. Tubes can be forgiving), the BFO fired right up and SSB ops now sounded as expected. CW notes, too.

Because everything was working okay, I decided to hedge my bet and replace all of the rather melted, drippy wax and paper capacitors—electrolytics, too. If one of the melted caps shorts the high-voltage bus (a common-enough occurrence), bad things can happen.

The shiny new 0.1 and 0.01 bypass caps look like Barbie Doll parts compared to the bloated, leaking Havana cigar-size originals. Now, there's a *lot* of extra room under the hood! The neon blue mini-caps look a bit out of place, but they're tucked away where nobody sees them, and I know the receiver won't suddenly "go south" because of a pooped capacitor.

I was really into the process now, so I called the friendly folks at Antique Electronics Supply (see the other sidebar), and for only $35 I had a new set of tubes. (One of the original tubes was "microphonic." If I tapped the side of the tube (with a plastic stick) or the receiver cabinet, the speaker reverberated with the sound of an iron-booted giant walking down a marble-floored hallway. This sometimes happens to older tubes as internal elements become loose and vibrate while amplifying.)

To make a long story short, the HRO-7 works like a champ. The dial calibration is accurate, the VFO (HFO in those days) is stable, the IF bandwidths are fine, CW notes sound smooth and pure—even SSB signals sound good. With proper fiddling with the crystal phasing control, proverbial single-signal reception is possible.

It's not as crunch-proof as an Omni VI or FT-1000D, but few tube radios were, are, or likely will be. But it's more than functional, and I get a lot of satisfaction from using it, all the while knowing that I dusted it off, fixed it up and gave it new life.

Yep, that nifty HRO is all mine!

Neil Wiegand, WA5VLZ, with his shack full of goodies. To the right of HRO "Senior" is a Stancor 20P. Next to Neil's hand is a home-brew regenerative receiver in a wooden cabinet. *(photo courtesy of Jim Musgrove, K5BZH)*

Electric Radio publisher Barry Wiseman, N6CSW, recently acquired this one-tube transmitter built in an old-style coffee can. It's an excellent example of vintage "homebrew." *(photo courtesy of N6CSW)*

Vintage Radio Resources

Heathkit Repair and Parts
RTO Electronics
4166 Maple St
Berrien Springs, MI 49103
616-473-3201

Transformers—Repair & Custom Made
Peter Dahl
5869 Waycross Ave
El Paso, TX 79924
915-751-2300

Tubes, Parts, Books
Antique Electronics Supply
6221 S Maple Ave
Tempe, AZ 85283
602-820-5411

Collins, Paint, Parts, Rigs
Surplus Sales of Nebraska
1315 Jones St
Omaha, NE 68102
402-346-4750

Ron Foffmar—Collins Junk Yard
1409 West Willis
Alvin, TX 77511
713-331-1074

Manuals for All Rigs
HI Manuals
PO Box Q-802
Council Bluffs, IA 51502

The Collins Collector's Magazine
2645 W Chicago St
Rapid City, SD 57702

Parts, Equipment
Fair Radio Sales
1016 East Eureka St
PO Box 1105
Lima, OH 45802

Hallicrafters Manuals
Ardco Electronics
PO Box 95
Berwyn, IL 60402

Mosley Manuals & Parts
Mosley Antennas
1344 Baur Blvd
St. Louis, MO 63132

Electronic Parts
Newark Electronics
4801 N Ravenswood Ave
Chicago, IL 60640

Metal Inserts for Knobs (Heath & Collins)
Charles Tabbott
13192 Pinnalle Ln
Leesburg, VA 22075

Electric Radio Magazine
PO Box 57
Hesperus, CO 81326
303-247-4935

an audio bandpass filter prior to the transmitter's audio (microphone) input.

Many of the vintage transmitters do an excellent job on CW. Make sure you have a clean note with no chirps and you're all set. If one isn't included, an external VFO can be added for greater flexibility.

In the old days, because many rigs were crystal-controlled, operators typically tuned 10 to 15 kHz up and down the band after calling CQ. The resulting QSO would be "split frequency." These days, answering someone's CQ 10 kHz away isn't likely to result in a contact! If your transmitter is crystal-controlled, you may have to initiate most of your contacts by calling CQ.

Conclusion

We hope that you've enjoyed this introduction to vintage radio collecting and operating, and that you have an interest in the heritage of Amateur Radio. Our intention was mainly to help "set the hook" on those who may already be interested in classic radio, but if you're looking for something to spark your interest and enjoyment of our diverse hobby, vintage radio may be just what the doctor ordered.

We'll be listening for you with our HROs!

Marty Drift is employed by Texas Instruments and Jim Musgrove is employed by Motorola. Both have been involved with classic radio for several years and have helped with vintage radio presentations at conventions and hamfests. QST

Strays

CQ CHEMISTS

◊ Any amateurs who attend the Pittsburgh Conference on Analytical Chemistry, PITTCON '93, in Atlanta, Georgia, March 7-13, please put your QSL card on the message board under "Q." Also, any suggestions for using a particular repeater there would be appreciated. Dave Sher, W9LYA, 9614 Kostner, Skokie, IL 60076.

QST congratulates...

◊ ARRL *Repeater Directory* Editor Jay Mabey, NU0X, of Newington, Connecticut, on attaining the rank of Master Instructor through the Professional Association of Diving Instructors (PADI). Achieving the coveted Master Instructor rating also places Jay on the PADI Educational Review Committee.

◊ Bruce Cohen, NS3R, of Gaithersburg, Maryland, on being promoted to lead systems accountant at the US Information Agency in Washington, DC. Cohen, a CPA and CFE, has more than 13 years experience as an auditor, systems consultant and supervisory operating accountant with the federal government.

Crossband-Repeater Operation

New dual-band VHF/UHF rigs offer flexible operating capabilities, as pointed out in the September 1992 FM/RPT column. Here's how to take advantage of this feature and avoid potential problems.

By Bob Witmer, W3RW
146 Forest Trail Dr
Lansdale, PA 19446

Would you like to be able to check into that distant 2-meter repeater swap net while taking a walk in your neighborhood? How about monitoring 52.525-MHz FM simplex activity with your 440-MHz hand-held transceiver while tinkering with your car? Do you need quick repeat/relay capability for public service or emergency communications?

If you have a 440-MHz hand-held transceiver and one of the dual-band FM transceivers that offers crossband-repeater (CBR) operation, all of the above is possible with, in most cases, only a minor modification to the dual-band transceiver to activate the CBR mode. In some cases no modifications are necessary. See your rig's operating manual or contact the manufacturer for details on modifications and activation procedures.

What is Crossband-Repeater Operation?

Unlike normal repeater operation which takes place using two frequencies on one band (Fig 1A), a CBR uses one frequency on each of two bands (Fig 1B).

When CBR operation is activated, the transceiver's receiver switches between the two bands. When a signal is received on one band, such as 2 meters, it connects the audio output of the 2-meter receiver to the audio input of the 440-MHz transmitter and keys the 440-MHz transmitter. (On some equipment, such as the Yaesu FT-470 hand-held transceiver, there's no direct audio connection for CBR—the audio is acoustically coupled.) When the input signal disappears, the transmitter is unkeyed and the transceiver goes back to alternating receive on the two bands until a signal is again received and the transmit cycle is repeated (no pun intended).

Crossband-Repeater Operating Modes

Depending on the radio, there are three types of CBR operation:

Simplex-to-Simplex CBR

In simplex-to-simplex CBR operation (Fig 2), the transceiver is set for simplex frequency operation on both bands. This is a good way to keep track of activity on a local frequency when you're away from the base rig.

Simplex-to-Repeat CBR

In simplex-to-repeat CBR operation (Fig 3) the transceiver is set for simplex operation on one band and repeater frequency offset operation on the other. On some transceivers, such as the Kenwood TW-4100A, simplex-to-repeat CBR is only possible with standard frequency offset repeaters—though this shouldn't be a problem for most users. In this mode, you're dependent on the characteristics of the repeater you'll be operating with.

One-Way CBR

An additional mode, one-way CBR, is shown in Fig 4. This mode, available on some late-model dual-band rigs, such as the Kenwood TM-741, provides a way to remotely monitor activity. Because this could be considered one-way transmitting, I recommend that the transmit side (band) of the rig be connected to a dummy load during operation. This greatly limits the effective range of the transmitter, while still providing receive coverage close to the dual-band rig and dummy load.

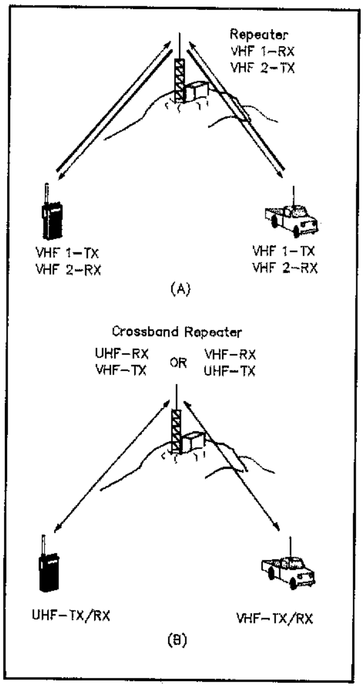


Fig 1—**Normal Repeater Operation** (A) **Crossband repeater operation** (B)

Using It

Read the sidebar "The Law Says. . ." before you begin operation.

Base/Mobile CBR with Hand-Held Transceivers

Repeaters you can hit with a more powerful base or mobile rig may not be accessible from a hand-held radio. With CBR, you can extend the range of your portable rig. If you have specific areas or directions of operation in mind, choose antennas that favor those directions. For example, if there's a distant 2-meter repeater due west that you'd like to be able to reach via CBR and your primary hand-held transceiver operation is located due north of your CBR's location, you'd equip the CBR with a 440-MHz beam pointed north at your portable location and a 2-meter beam pointed west at the distant repeater to provide optimum operation with minimum operating power. If your operation is limited to your immediate vicinity, consider feeding the local—usually 440-MHz—transmit side of your dual-bander into a dummy load to limit the transmit range and limit the ability of undesired signals to access your system.

Hand-Held CBR with a Hand-Held Transceiver

You may be able to put a hand-held radio at a temporary location to provide CBR coverage where it wouldn't be practical to operate a higher-powered base or mobile transceiver.

The coverage advantage provided by a hand-held at a choice location may be superior to that provided by the location obtainable by a higher-powered mobile. As with any hand-held operation, battery consumption is an important consideration. If a particularly good location is available, CBR operation with a low power setting may be sufficient.

Application Examples

Operation from Inside a Partially Shielded Building

Fig 5 shows an example that, unfortunately, most of us are familiar with. Computers have brought an increasing level of RFI that's

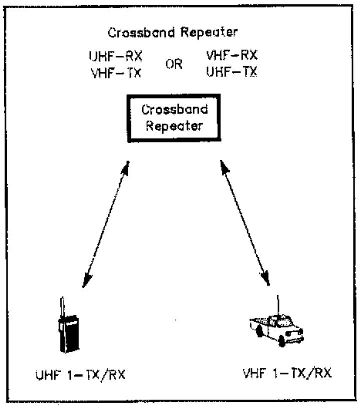


Fig 2—**Simplex-to-Simplex Operation**
In this mode, each band has the same frequency for input and output.

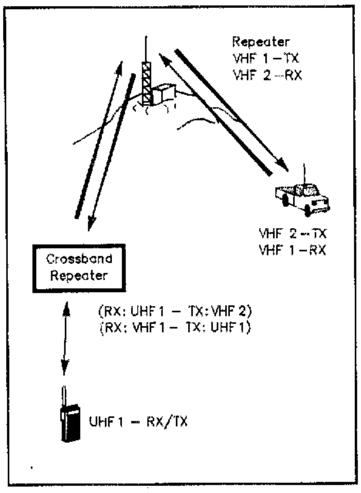


Fig 3—**Simplex-to-Repeater Operation**
In this mode, UHF has the same input and output frequency, and VHF has different input and output frequencies (differs by repeater offset).

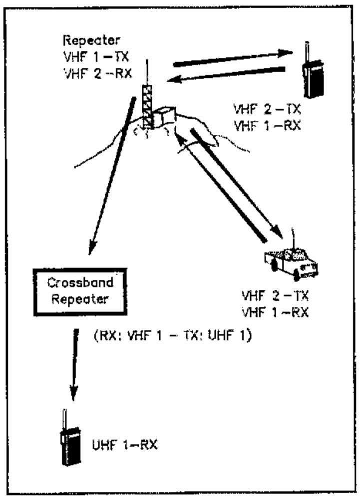


Fig 4—**One-Way Repeat Operation**
In this CBR mode, VHF is designed as receive and UHF as transmit. Reverse operation is not activated.

made operation near them difficult—especially inside a building that may already be partially shielded—providing difficult repeater operation. With a mobile rig operating as a CBR just outside in the parking lot, hand-held access to your favorite 2-meter repeater from inside the building is easy.

Access Assist

With the high sensitivity of modern hand-held transceivers and the increasingly tough RF environment provided at repeater sites, balanced repeater operation is often impractical, even with mobile equipment. It's often possible to hear a repeater easily but not be able to access it with a hand-held transceiver. Fig 6 shows how CBR operation with a dual-band hand-held can provide access-assist capability. Operating this way eliminates several problems associated with normal repeater-carrier tails. For example, CBR autopatch operation is possible when operating this way, provided your dual-band rig in CBR mode passes the DTMF tones without too much distortion. The disadvantage of this approach is that there may be audio feedback. You can minimize this by using an earphone or by operating CBR without simultaneous-receive/full-duplex operation.

Remote Monitor

One-way monitoring is possible. Fig 7 shows how, with the addition of a receive converter (in this case, a 6-meter converter), CBR can be used to provide remote-receive capability that lets you listen while walking around with a hand-held radio.

Fig 8 shows the concept of Fig 7 taken a step further. Instead of the receive converter, a transverter is used. In Fig 8, a 6- to 2-meter transverter used with a dual-band transceiver in CBR mode provides 6-meter remote-base capability.

Operating Practice: Avoiding Problems

Long Repeater-Carrier Tails

A major potential problem with simplex-to-repeat CBR is operation with repeaters that have long transmitter carrier tails. The long repeater carrier tail (originally called a "squelch tail") came about when the reset function for repeater control system time-out timers switched from repeater carrier-transmitter drop to received signal-carrier drop, which spares users the squelch crashes associated with the repeater carrier dropping between input transmissions. The "courtesy beep" replaces this characteristic on many repeaters and the repeater carrier remains on the air. Because of the way CBR works, you can't respond to someone on a repeater until its transmitter carrier drops. It's frustrating to have to wait 10-15 seconds before being able to respond, and it sure limits rapid exchanges. A short (one-second) carrier-tail repeater facilitates rapid CBR mode exchanges.

Two-Way CBR and Repeater Autopatches

CBR makes it essentially impractical to use an autopatch through a remote repeater. Initiating an autopatch call in this mode typically times out the patch. Once the patch is activated, it looks for signals from your transmitter for control. Unfortunately, your CBR's receiver still hears the repeater's transmitted carrier, preventing the transceiver's transmitter from functioning. Without a dial or disconnect signal, the repeater's autopatch will eventually (hopefully) self-disconnect.

Don't Link Two Repeaters

Because of the way CBR works, it isn't possible to link two normal repeaters. A "normal" repeater means a machine that has a carrier tail. If two such repeaters are CBR linked, once the first outside signal keys one of the repeaters, the repeaters will be ping-ponged on and off by each other's carrier tails—not a desirable condition.

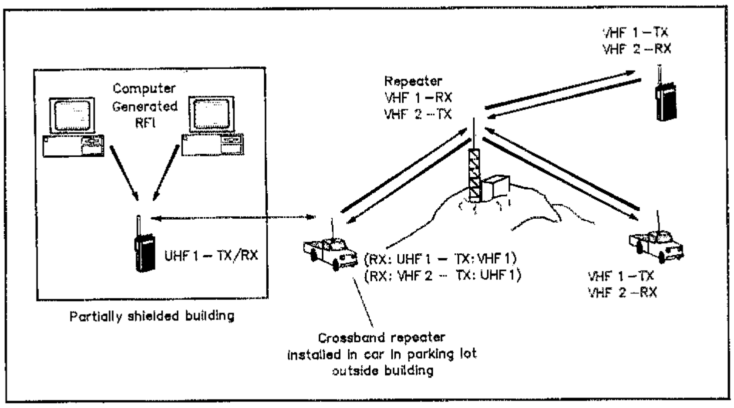


Fig 5—**CBR Assists Operation From Inside Building**

"Open" Transceiver Mike in CBR Mode

Check your dual-band transceiver's operating manual to determine the method for CBR audio connection and the effect of CBR operation on front-panel controls. Some radios require special consideration:

When operating the Kenwood TW-4100A in CBR mode, you'll probably want to disconnect the microphone because it's "live," though the push-to-talk switch isn't functional, when the transmitter is activated. If you don't disconnect it, local audio will mix with the repeat audio and possibly cause feedback, depending on factors such as your speaker volume setting. You could use this to mix in ID audio.

The same open-mike precaution applies to the Yaesu FT-470 hand-held transceiver. In the CBR mode, repeated audio is acoustically coupled, with no direct audio connection, so the microphone is open to all local audio, not just that picked up from the speaker.

Transmit Power and Duty Cycles

If a repeater you operate with in CBR mode has a lot of activity, lower your transceiver's transmit-power level. Most FM transceivers aren't designed for continuous high-power transmission. Consider the heat-dissipation environment of your rig's installation, too: The typical mobile installation on a hot summer day isn't an ideal environment! Check the rating of the power source you use to avoid overheating it or running down your car's battery in a stationary mobile situation with the engine off.

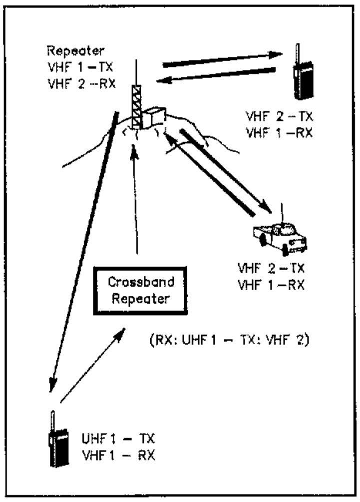


Fig 6—**Access-Assist Operation**
In this CBR mode, the crossband repeater permits access to a repeater not normally accessible by hand-held power levels, but within hand-held receive range.

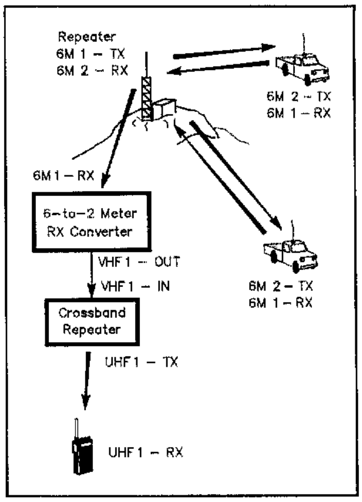


Fig 7—**Remote-Monitor Operation**
In this CBR mode, one band is designated as receive, with an external converter and the other band as transmit. Two-way transmit is not activated.

Check Your Arithmetic!

Harmonically related frequencies won't work in CBR mode. For example, if you monitor the output of a repeater on 442.5 MHz using a simplex frequency of 147.5 MHz, you'll run into trouble because 442.5 MHz is the third harmonic of 147.5 MHz (147.5 × 3 = 442.5), which will interfere with reception of the 442.5-MHz signal.

Finding a Clear Frequency

Observe the band plans when selecting your simplex frequency for CBR operation. Carefully select a CBR frequency to avoid interference problems. You're responsible for what's retransmitted by your station! The old

The Law Says...

A full discussion of the Part 97 rules covering repeater operation is beyond the scope of this piece, but the following review should help you avoid problems:

Part 97.3 (a)(35) defines a repeater as "An amateur station that automatically retransmits the signals of other stations." CBR operation is repeater operation. Because the CBR is a repeater, it's essentially under automatic control. Transmitters under automatic control must be able to be shut off if anything goes wrong. Wherever you set up your CBR, be sure you're in a position to turn it off in a timely manner if a problem arises. Review the following sections of Part 97 to avoid violating FCC rules:

Part 97.205 Repeater Station Requirements:

- Permitted control operator license classes (Novice not permitted).
- Permitted operating frequencies.
- Automatic control operation permitted.
- It's okay to limit repeater access to certain users and CTCSS access is allowed.

Part 97.119 Identification Requirements:

- You should identify the CBR by voice when operating, eg, "W3RW clear via W3RW Repeater."

Part 97.109 Station Control Requirements:

- Automatic control.

Although this article is targeted at CBR operation using 146/440-MHz dual-band rigs, the same type of operation is possible with 146/222-MHz dual-band transceivers. A Novice operator, however, may *not* be the licensee or control operator of any repeater.

Aside from the basic FCC rules, there are other important considerations to look at before you fire up your shiny new dual-bander as a crossband repeater. Established, recognized band plans set aside specific frequencies and band segments for repeater inputs and outputs. On top of this, your local repeater-coordination or spectrum-management group is responsible for keeping track of what repeaters are operating on which frequencies. In some areas, there are *no* available repeater frequencies on a given band.

Flipping a couple of switches on your mobile or hand-held rig may not seem like a major undertaking, but it *is* essentially the same thing as installing a regular repeater. If you suddenly appear on a well-known repeater input or output frequency, or one used for established simplex, packet, RTTY, weak-signal work, satellite contacts or other purposes, you aren't going to make many friends! Also, the FCC says that in matters of interference between repeaters, the repeater recommended by a recognized coordinator is given preference and the licensee of the noncoordinated repeater has primary responsibility to resolve the interference (ie, shut down, move or face possible charges of malicious interference). See the FCC rules, Section 97.205(c). Contact your local repeater coordinator for details; addresses are published in the latest ARRL *Repeater Directory*. The *Directory* also lists band plans and specific repeater frequency pairs.

Use your head—after carefully checking for a clear frequency (and one authorized for repeater operation), operating a low-power, restricted-range crossband repeater temporarily in a remote area may be okay. Randomly kicking your dual-band mobile into CBR mode while cruising through a congested metropolitan area is asking for trouble. Enjoy CBR operation when it's an advantage, not an annoyance.

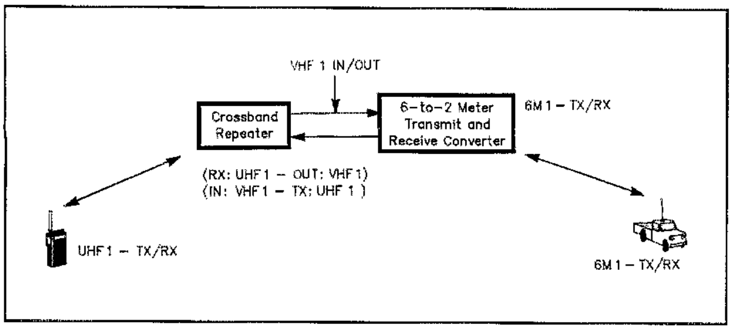


Fig 8—**6-Meter CBR Remote Base Operation** In this mode—a variation of simplex-to-simplex operation—a 2- to 6-meter transverter provides 6-meter operation.

adage applies: "Listen before transmitting!" Monitor your selected frequency for as long as possible—at least a few weeks or so, preferably a couple of months—to determine how clear it is of activity. Once you've picked a relatively clear frequency, it may be your best bet to operate CBR using continuous-tone-coded squelch system (CTCSS) on your receive frequency to eliminate undesired access.

Do it Right and Have Fun

These are just a few uses for CBR. Many of these could be used to expand repeater access in emergencies and in normal operating situations.

The CBR capability available in many dual-band mobile and hand-held FM transceivers can be used to improve your communications capability. By following simple guidelines, potential operating, equipment and FCC problems can be avoided. Experiment and let me know if you come up with any new applications. This is one part of ham radio I find interesting, as do many operators of "regular" repeaters.

References

FCC Rule Book, 8th Ed (Newington: ARRL, 1991).

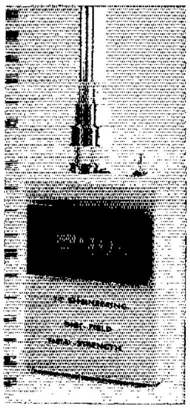
Kenwood USA Corp *TW-4100 Repeater Operation Supplement.*

Witmer, R., "Kenwood 4100A Crossband Repeater Operation," *73 Amateur Radio*, November 1988, p 39.

New Products

WIDE-RANGE SIGNAL-STRENGTH METER

◊ Designed for use in a wide range of applications, the Digi-Field instrument can be used as a "sniffer" for dc-1.3 GHz emissions, such as 60-Hz noise sources, RF and microwave oven leakage. Its clear visual readout lets the user acquire calibration data for plotting antenna patterns, checking polarization, measuring antenna radiation gain or loss, comparisons and other measurements. The hand-held device comes with its own telescoping antenna and a PL-259 connector for use with an external antenna. It has a 3½-inch LCD readout and low-battery indicator (requires one standard 9-V battery). Retail price $79.95. IC Engineering, 16350 Ventura Blvd, Suite 125, Encino, CA 91436; tel 800-FIELD-58 (343-5358) or 818-345-1692, fax 818-905-3374.



Nominations Open for Philip J. McGan Memorial Silver Antenna Award

Know someone who stands out in Amateur Radio public relations? Get a nomination in *now*—the deadline is almost here!

By Steve Mansfield, N1MZA
ARRL Public Information Manager

The deadline for the second annual Philip J. McGan Memorial Silver Antenna Award is April 30, 1993.

This award is given jointly by the ARRL and the New Hampshire Amateur Radio Association to the ham who best exemplifies the volunteer public relations efforts of Phil McGan, WA2MBQ (SK), on behalf of Amateur Radio.

Phil McGan served as the first chairman of the ARRL's Public Relations Committee, which helped reinvigorate ARRL's commitment to public relations. He was also the League's New Hampshire Section Public Information Coordinator. The first award was presented to James Heil, KB5AWM, of Houston, Texas.

CALL FOR 1993 ENTRIES:

1) Eligibility: The award is given to an individual (not a group), who is a full ARRL Member in good standing at the time of nomination. The nominee must not be compensated for any public relations work involving Amateur Radio (including payment for articles) and may not be a current officer, director, Vice Director, staff member or member of the current selection committee.

2) Selection criteria: The winner of the Philip J. McGan Memorial Silver Antenna award will demonstrate volunteer public relations success on behalf of Amateur Radio at the local, state or national level, and will live up to the high standard of achievement exemplified by Philip J. McGan.

3) How to nominate someone: Anyone may make a nomination. Nominations must be on an official entry form, available from ARRL headquarters. The nomination will include a written summary of activities with supporting documentation whenever possible.

4) Deadline: Nominations must be received at ARRL Headquarters in Newington by 5 PM April 30, 1993. Nominations arriving after the deadline or without an entry form will not be considered.

5) How the process works: Nominations will be screened by the ARRL Public Relations Committee, which will forward its recommendation to the Volunteer Resources Committee of the ARRL Board of Directors. The Board will make a final determination at its July meeting and the winner will be notified shortly thereafter.

6) To obtain an entry form, call ARRL HQ at 203-666-1541, ext 245; e-mail smansfield@arrl.org; MCI Mail 519-3279. Ask for an official Philip McGan Memorial Silver Antenna Award entry form.

7) Return completed entry form and supporting materials to Philip J. McGan Memorial Silver Antenna Award, c/o Steve Mansfield, N1MZA, Public Information Manager, ARRL, 225 Main St, Newington, CT 06111.

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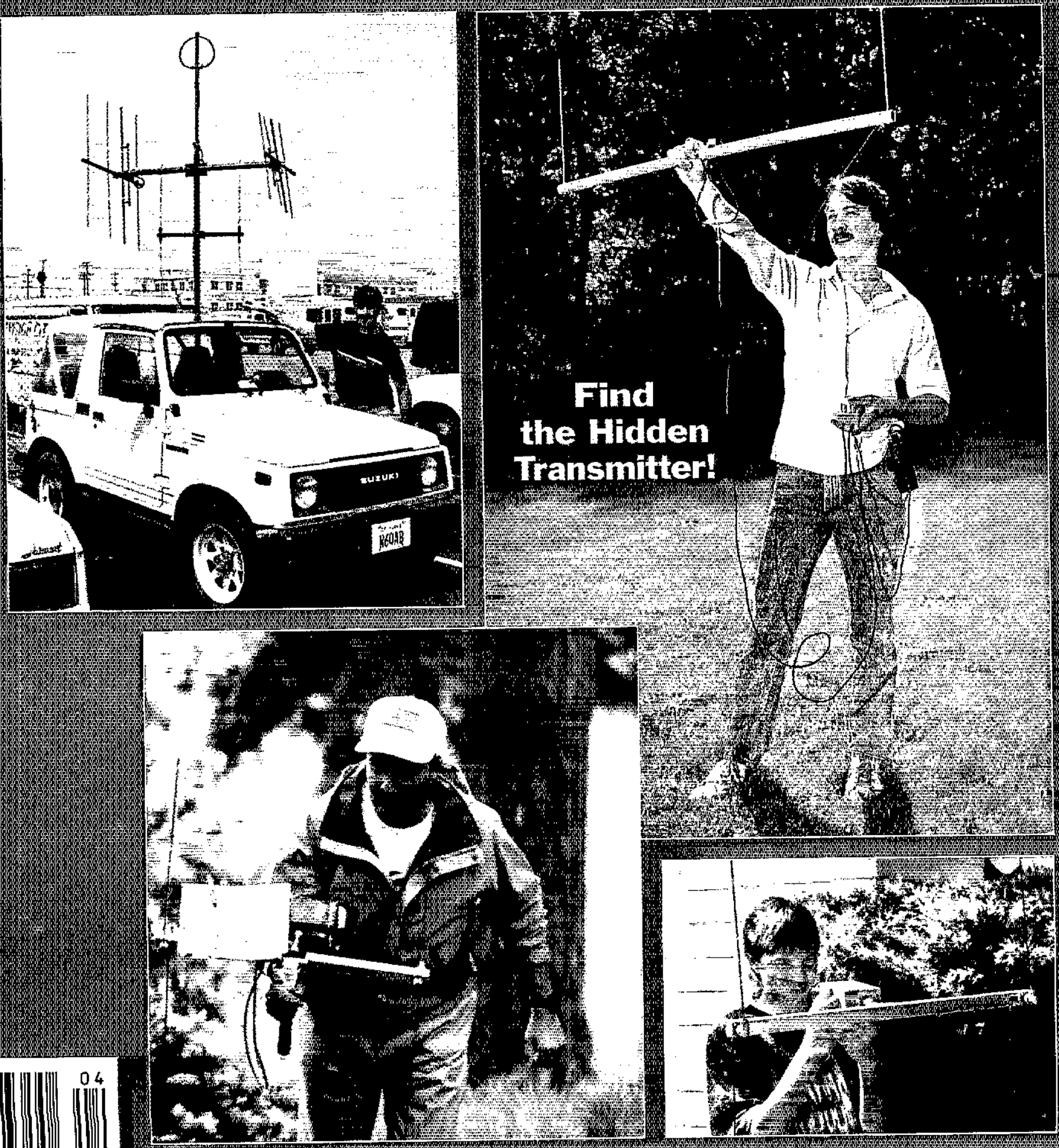

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Membership in the ARRL, including a subscription to *QST*, is available to individuals at the following rates: $30 per year in the US and possessions, $42 elsewhere, payable in US funds. Age 65 and over, with proof of age, $24 US, $36 elsewhere. Licensed radio amateurs age 17 and under may qualify for special rates; write for application. Membership and *QST* cannot be separated. Fifty percent of dues is allocated to *QST*, the balance for membership. Subscription rate for libraries and institutions: $30 per year postpaid in the US and possessions, $42 elsewhere. Single copies $3 in the US.

Foreign remittances should be by international postal or express money order or bank draft negotiable in the US and for an equivalent amount in US funds. Canadians apply to CRRL address; see page 9.

Second-class postage paid at Hartford, CT, and at additional mailing offices. Postmaster: Form 3579 requested. Send change of address to: American Radio Relay League, 225 Main St, Newington, CT 06111.

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OUR COVER

Amateur Radio direction finding is a fine art that's fun to learn. Clockwise from upper right: ARRL Senior Technical Illustrator Dave Pingree, N1NAS, sniffs 2-meter RF in a Newington park *(photo by Kirk Kleinschmidt, NT0Z)*; Nebraskan Eric Jensen, 13, learns about one facet of ham radio by joining monthly T-hunts with the Lincoln ARC; Kevin Kelly, N6QAB, wins the 5-transmitter on-foot foxhunt at the 1991 Friendship Radio Games in Oregon; Kevin tracks hidden transmitters on three bands at a hunt in Los Angeles. *(photos by Joe Moell, K0OV).* Read about the fine points of transmitter hunting on page 48 of this issue!

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NEWS AND FEATURES

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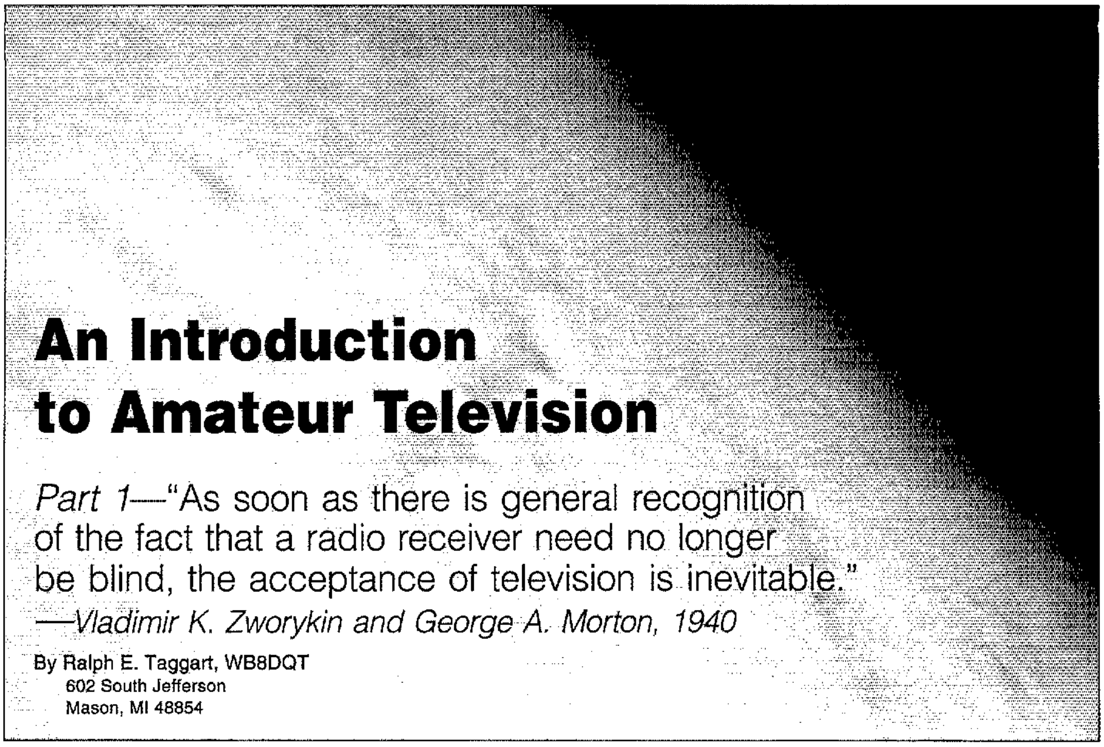


An ATVer's view of earth and space from an altitude of 125,000 feet. This picture was transmitted to earth from a balloon-launched, 1-watt ATV transmitter. *(photo by WB8ELK)*

An Introduction to Amateur Television

Part 1—"As soon as there is general recognition of the fact that a radio receiver need no longer be blind, the acceptance of television is inevitable."

—Vladimir K. Zworykin and George A. Morton, 1940

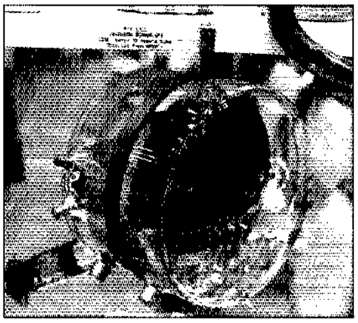
By Ralph E. Taggart, WB8DQT
602 South Jefferson
Mason, MI 48854

When Zworykin and Morton penned these words at the end of their classic text, *Television*,[1] NBC had concluded a major test of the RCA TV system, and tens of thousands of people had had their first glimpse of television at the 1939 World's Fair. World War II would delay further TV experiments, but would also provide technical advances that set the stage for the widespread introduction of broadcast TV service. Fifty years later, TV has impacted our lives in ways Zworykin could never imagine in 1940. Full color and stereo sound, cable distribution, TVRO satellite dishes, laser video disc, the ubiquitous VCR and the camcorders that have eliminated the movie camera from family events—all of these are the bits and pieces of the mosaic we grandly refer to as the "Video Age."

Almost anyone's vision of the future seems to incorporate TV as a medium of personal communication. From the video phones in *2001—A Space Odyssey*, to Captain James T. Kirk "opening a channel" from the bridge of the starship *Enterprise*, communications with a future slant seem incomplete without a live-action picture to accompany the interplay of voice communication. If, indeed, this is the future, we ought to be able to see some faint prophesy of things to come among the ranks of the penultimate communicators—radio amateurs. Strangely, although video technology has invaded the home, it has largely stopped short of the ham shack. With all of its high-tech sophistication, most of our equipment is just as blind as it was when Zworykin uttered his prophesy!

Changing Times

Amateurs have been involved with TV since the days of mechanical scanning disk systems of the 1920s, yet only a few thousand amateurs are active on this mode. This is all the more amazing when you realize that practically everyone has one or more TV sets, and each year that passes makes it more likely that we'll also have a camcorder gathering dust between family milestones!



In the beginning...a 1935 Zworykin iconoscope TV camera tube. *(photo by KB9FO)*

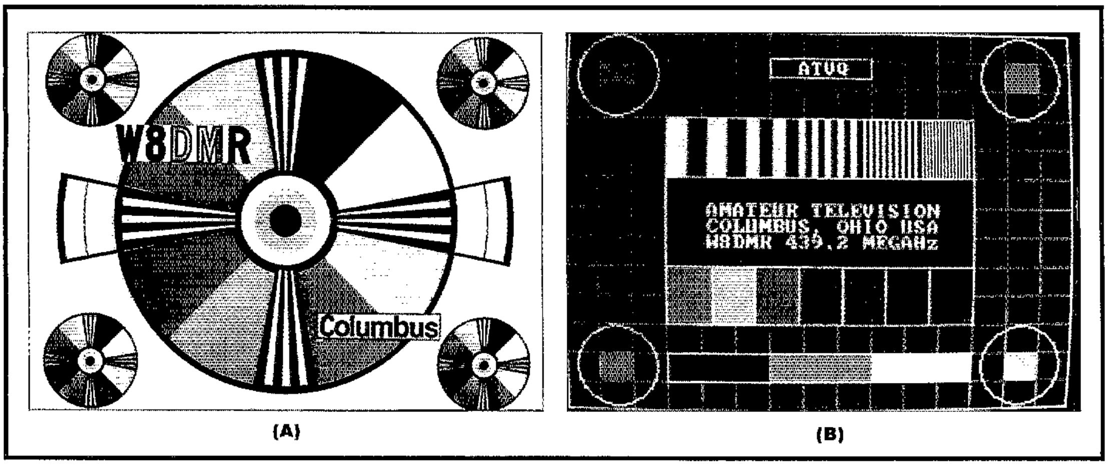
Many amateurs think: "It's too expensive!"; "You have to be a technical wizard!"; "Assembling a station is a major building project!" and "Why go through all that trouble to work across town?"

Like many misunderstandings, these have their origins in a kernel of historic truth. When I assembled my first ATV station in 1963, the effort took more than two years—including the construction of a camera—and I ended up with a rack full of equipment that could transmit a signal for all of two miles! Although I'm proud of the accomplishment, those days have about as much relevance to modern ATV as spark does to packet!

There's Lots to Tell...

There's a lot going on in ATV in the '90s, and it simply isn't possible to pack even an introduction to ATV into a single article. If you missed the earlier *QST* article "Smile—You're on Ham Radio,"[2] go back and read it.[3] It contains an ATV resources guide (where to find out more about ATV and the equipment available) and lots of other useful information.

[1] Notes appear on page 23.



Examples of homemade—and *handmade* ATV test patterns. The "wagon wheel" or "propeller" pattern at A is a product of handiwork and photography by Bill Parker, W8DMR. Bill used a thumbtack, string and marking pen to make the original (middle) circle. The white and colored wedges were cut from construction paper. Bill scissored his call sign from a QSL card, and the *Columbus Dispatch* provided the word *Columbus* (with a bit of help from a black marking pen). All this, combined with photocopying and photography, resulted in the attractive test pattern you see here.

At B, an IBM BASICA program written by Bill creates the computer-generated test pattern, which includes linearity bars (the gridwork), multiburst (the alternating white and black vertical bars used to check frequency response), and gray scale and color bars. The wedges are used to check horizontal and vertical resolution. Creativity at work here! *(photos by W8DMR)*

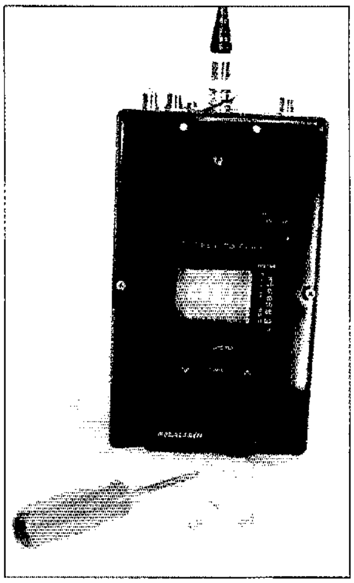
I'll break my presentation into three parts. In this first installment, we'll look at some of the things that are being done on ATV, take a quick survey of the bands being used and provide a few guidelines on how you can eavesdrop on existing ATV activity in your area. Part 2 covers the basics of putting an ATV station on the air: the antennas, transmission lines and the receiving and transmitting equipment you'll need to get started on ATV. In Part 3, we'll delve into slightly more technical areas: adding transmit power amplifiers, the how and why of ATV repeaters and a discussion of some of the experimental work with modes such as FM TV on our microwave bands.

As in most areas of Amateur Radio, there is more than enough technical detail to keep you fully occupied if you are so inclined, but technical expertise is *not required* to get a station on the air. The most complex pieces of the station—the TV cameras and receivers—you may already own and know how to use. Getting this equipment up and running on ATV is no more demanding (or expensive!) than adding a new FM rig to the shack, so let's start with *why* you might want to do just that!

Bands and Formats

Worldwide, there are three major TV formats in use: NTSC, PAL, and SECAM (see the Glossary of Terms). Each format differs in the number of lines that make up a TV picture, the rate at which individual pictures or *frames* are transmitted to give the illusion of motion, the details of the modulation formats and how they handle color. The technical details need not concern us since you'll be using equipment compatible with standard TV broadcasting in your area. For those of us in North America, that's NTSC; PAL and SECAM are used throughout Europe. Which format is used in other parts of the world is largely a matter of political history and who got the contracts for installing the first national TV systems!

Wherever you live, the ATV activity uses the *same standards* as that of the TV broadcast stations in your area—that allows the use of readily available cameras, VCRs and TV receivers. My discussions will center on equipment suitable for NTSC standards employed in the United States, Canada, and Mexico, but most of the principles apply to areas where other formats may be in use.



This *hand-held* ATV transceiver and camera (a "walkie-lookie") was built by Earl Campbell, KS8J. *(photo by KS8J)*

Where Do We Operate?

The vast majority of ATV operation is in the 70-cm band, the lowest amateur band authorized for ATV operation. There is some simplex operation on the 33- and 23-cm bands (particularly in urban areas), but these frequencies present problems in designing and building equipment to handle higher transmitter power levels. Experience shows that path losses caused by obstructions (buildings, trees, etc) are significantly higher there than on 70 cm.

ATV Repeaters?

Yup. There are now one or more ATV repeaters operating in over 40 states and around the world. Many of these repeaters are "in-band," using 70-cm frequencies for input and output. Where the 70-cm band is particularly congested—or where proximity to line "A"[4] limits the frequencies available—repeaters typically use a 70-cm input or output frequency, cross-banding to 33 or 23 cm. Unless you have a local ATV repeater, 70 cm is the band of choice, combining excellent receiver capabilities, easy power generation, compact antennas, moderate path loss, and the opportunities for DX contacts when the band is open.

What Do You Do On ATV?

Why spend the time and money to watch

the same people night after night? Good question! Why do so many amateurs talk to the same folks again and again on 75 or 2 meters? With ATV, you get to *see* friends rather than just staring at the LEDs and meters on the front of your transceiver. With ATV, you also get the bonus of seeing spouses, kids (and their birthday parties), parades, the new car, antennas, towers, the latest gear, and—of course—anybody who happens to stop by for a visit!

What the Local Folks Do

Here in the Lansing, Michigan, area, the dozen or so members of the Central Michigan Amateur Television Society have a wide range of interests in addition to radio, all of which can be shared, courtesy of modern video technology. If a space shuttle mission is in progress, Jeff, WB8RJY, will be relaying NASA Select video from his TVRO installation. Bruce, KA8ZXX, owns a marvelous collection of pictures covering the entire manned space program. Andy, W8AHY, has a collection of videotape shot at steam-engine shows, and Steve, N8LWX, is our roving video reporter at hamfests, antenna raisings and club meetings. I don't require any real persuasion to air the latest weather satellite pictures, videotape shot at ultralight aircraft fly-ins, or footage shot from my ultralight gyroplane (which he built himself.—*Ed.*)

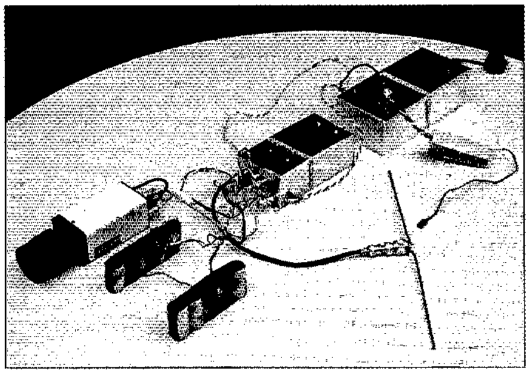
Other ATV groups are similarly diverse. Many have a long history of public service involvement, providing video coverage of parades and other events, volunteering for search and rescue work, undertaking video storm watches and a host of other activities that benefit the community and provide a high and positive profile for Amateur Radio.

Other ATV Activities

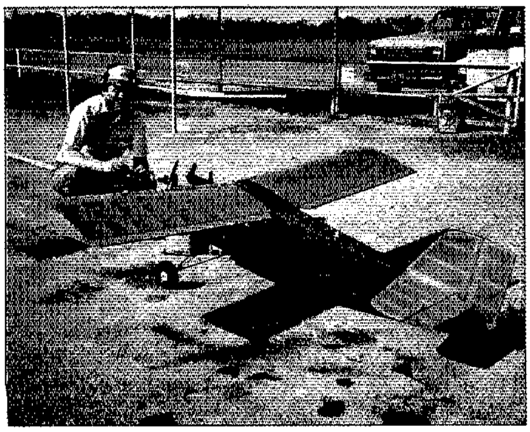
Elsewhere, amateurs are launching model rockets equipped with miniature cameras and ATV transmitters, flying ATV gear in radio-controlled aircraft, and lofting balloon packages to the edge of space to provide spectacular pictures for hundreds of miles! Not very long ago, those of us here in Michigan had the opportunity to watch live pictures from an altitude of 138,000 feet, courtesy of a balloon-born package constructed, launched and tracked by the Dayton, Ohio, ATV group! Lofting repeaters to the edge of space in similar balloon packages is definitely on tap for the future. For those who would reach out a bit farther, there is the potential of amateur satellite TV activity, encouraged by successful ATV operations from the space shuttle (STS-37) as part of the SAREX program.[5]

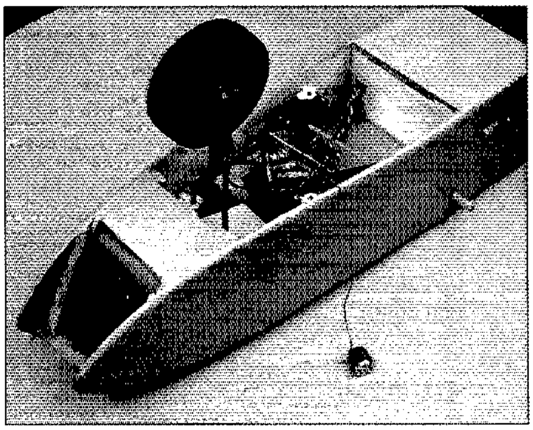
How Far Can I See?

My predicting *your* reliable working range is difficult, because it involves a number of factors: the nature of the local terrain, where you and the folks you want to work are situated on that landscape, and items such as transmitter power, feed-line losses and antenna gain. One of the reasons ATV is limited to the 70-cm band and higher frequencies is the signal's wide bandwidth. A broadcast-standard color signal with sound has a bandwidth of about 4.5 MHz, and the transmitted RF bandwidth ranges from about



Combining two hobbies makes for some excitement! Here's how Larry, K5MWN, loaded his ATV gear into the nose of his radio-controlled model airplane.





The screen shot below was taken while the aircraft was aloft, sending live ATV pictures to the ground. *(photos by K5MWN)*

Off-the-air video on 910.75 MHz using FM video. *(photo by WBØZJP)*

6 to 10 MHz, depending upon factors we'll discuss a bit later. At the receiving end, bandwidth is effectively limited to about 5 MHz by the TV-receiver design.

For Example ...

Most amateurs have used H-Ts to operate VHF or UHF FM in their area, so let's use that as a reference. *Based on bandwidth,* an ATV signal is degraded by about 25 dB compared to a 15-kHz FM signal of similar power. Although this sounds astronomical, let's look at a situation where you're able to maintain solid communication with another station using a VHF or UHF H-T or mobile rig running low power. Your mobile antenna gain is negligible, but a typical ATV system can easily contribute *at least 15 dB* of combined antenna and height gain, less reasonable line losses. Thus, although the ATV signal is 25 dB down (based on bandwidth), it's only about 10 dB down when the ATV antenna and height gain are factored in. So, if you can work a station with your H-T or mobile rig (running somewhere between 1 and 10 watts), you should be able to span the same path with good signals on ATV running 10 to 100 watts output.

In general, if you can work a certain distance from the VHF mobile at relatively low power, you should be able to do the job on ATV with a moderate power output. Using the 1-watt transceivers I'll discuss in Part 2, you should easily achieve a 20-mile working radius with average terrain. Your reliable coverage area can be extended easily to 40 or 50 miles with the addition of a modest "brick" amplifier.

When Unusual Circumstances Prevail

All of the preceding comments apply to normal band conditions, in terms of what can be expected on a reliable, day-by-day basis. Nature has some pleasant surprises in the form of *tropospheric duct* openings that are particularly common in the summer months in the East and Midwest. The current ATV DX record stands at over 1000 miles, and a moderately well-equipped station can expect to work stations on 200- to 400-mile paths. Any ATV station here in central Michigan that has been operated regularly through a summer tropo season has QSL cards confirming two-way contacts with stations as far

Glossary of Terms

ATV repeater—A station designed to automatically retransmit amateur video transmissions. Repeaters serve much the same role in ATV that they do in the case of FM voice communications: They extend the operating range of relatively low-powered ATV stations. In addition, given the highly directive nature of ATV antennas, a roundtable is far more practical with a repeater because all stations in the group have their antennas aimed at the repeater site.

Aural—The audio component of a composite TV signal.

Cross-banding—Operation in which an ATV repeater receives signals on one amateur band while retransmitting the signal on another band. Cross-band repeaters are typically easier to get up and running because isolating the receiver and transmitter is relatively easy. *In-band* repeaters, where the signal is received and retransmitted on the *same* band, present a real challenge in terms of obtaining sufficient isolation between input and output.

NASA Select video—The National Aeronautics and Space Administration leases channel space on commercial communications satellites and uses this capability to provide live video coverage and related material for significant space-related functions. These include shuttle missions and planetary encounters. These signals are *not scrambled*—NASA encourages dissemination of the information for educational purposes.

NTSC—The National Television System Committee, responsible for defining the video and audio standards used in commercial broadcast television in the US and elsewhere. It's also the designation of the television standard used in North America, Japan and some other places.

PAL—Phase alteration line. Television standard used in Germany and many other parts of Europe.

SAREX—Shuttle Amateur Radio EXperiment. Operation of Amateur Radio equipment aboard a space shuttle, with the approval of NASA and the FCC. NASA's intent in making astronauts available for SAREX operations is to involve the largest possible numbers of people, particularly youngsters, in technology and the US space program with the help of Amateur Radio.

SECAM—Sequential color and memory. Television standard used in France and the Commonwealth of Independent States.

Subcarrier—A signal component of the modulated signal that itself carries information. In the case of a TV signal, a frequency modulated audio signal is mixed with the video in the video modulator, resulting in an FM aural subcarrier carrying the voice signal along with the video modulation of the primary carrier.

Subcarrier-sound transmission—A system for simultaneously transmitting video and audio data (see *Subcarrier* above).

Tropospheric ("tropo") ducting—When air masses in the lower atmosphere differ significantly in temperature, conditions can exist where radio waves can be refracted. Temperature inversions, where warm air overlies cold air close to the ground, can cause some of the radiated signal to follow the curve of the earth (*ducting*) for many hundreds of miles.

TVRO—Television receive-only. An earth station installation for reception of television broadcasts from geostationary communications satellites.

Visual—The video or image component of the TV signal.

away as Iowa, New York, and Pennsylvania! Stations all over Illinois, Indiana, Ohio and parts of Ontario, Canada, are worked frequently enough to keep everyone on a first-name basis!

Looking In

If there *is* ATV activity within working range of your location, eavesdropping on the action need not involve a major monetary commitment on your part. Obviously, the first step is to determine *if there is* any activity in your area. Bring up the subject on your local repeaters and at club meetings; you'll soon find out. Chances are, if there is an ATV group, they'll be using a 2-meter FM simplex channel as an intercom, making it easy to monitor who is active. Audio-only check-ins are welcomed with open arms—and you would be hard-pressed to find a more missionary crowd!

Many active groups have one or more ATV converters that they pass around to let newcomers take a peek at what's going on. These simply install ahead of your TV set and output a signal to Channels 2, 3, or 4. Local ATV operators will give you plenty of help in using the converter—*they want you to succeed!*

If a loaner converter is unavailable, there are other options. *Cable-ready TV sets* can tune 70 cm (cable Channel 60 = the popular 439.25-MHz simplex frequency) and, although the converter's sensitivity leaves much to be desired, they can often be surprisingly effective when used in conjunction with an outside UHF TV antenna. Another option is to use one of the many models of 4- and 5-inch monochrome and color TV sets that are equipped with a slide-rule dial. These sets have an internal potentiometer that sets the range of the UHF varactor tuner. You adjust this control to shift a low UHF station up by about 8 channels. That done, you'll then be able to tune in 70-cm ATV at the low end of the UHF tuning range. Such a modified TV receiver can later serve as your station's transmitted-signal monitor.

Till Next Time...

I've only scratched the surface of the breadth of amateur TV operations. Adding the video dimension to your station operations has never been easier. In Part 2, we'll look at what it takes to actually get an ATV station on the air!

A note of special thanks to Henry Ruh, KB9FO,[6] *and everyone else who contributed* the photos used in this article.

Notes

[1] V. K. Zworykin and G. A. Morton, *Television* (New York: John Wiley and Sons, 1940), 646 pages.

[2] See B. Battles and S. Ford, "Smile—You're on Ham Radio," *QST*, Oct 1992, pp 42-44.

[3] Back issues are available from HQ for $3, postpaid. Contact Publication Sales.

[4] Covered by FCC regulations, §97.3(a)(26). See Chapter 5 of *The FCC Rule Book*, published by the League. You can find a copy at your local dealer, or order one from ARRL HQ. (See the ARRL Publications Catalog elsewhere in this issue; order number is 3304, price $9.)

[5] To find out more about SAREX, contact the ARRL Educational Activities Department.

[6] Henry Ruh, KB9FO, is Director of Engineering at WSNS-TV in Chicago, Illinois. To say that Henry is an avid ATVer is an understatement. Henry and Bill Brown, WB8ELK, jointly edit and publish the *Amateur Television Quarterly*, 1545 Lee St, Suite 73, Des Plaines, IL 60018, tel 708-298-2269, fax 708-803-8994. *ATVQ* offers two introductory books for would-be ATVers: *ATV Secrets for Aspiring ATVers*, a 96-page nontechnical introduction; price $9.95 plus $2.90 shipping. The 292-page *TV Secrets, Vol II for Novice and Expert*, has 90 projects to keep you busy; price $24.95 plus $2.90 shipping. QST

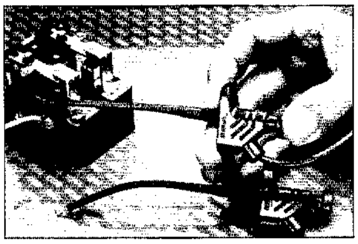
New Products

TINY ANTENNA

◊ For short-range communications without cumbersome whips or awkward "rubber duckies," the CH-32 Miracle Baby, shown here approximately actual size, is a handy dual-band antenna. Today's pocket-size hand-held transceivers will really fit in your pocket at hamfests, Field Day or while you do tower work, and this little 1.75-inch mighty mite works on 2 meters and 440 MHz. Suggested retail price $33.95. Comet Antenna, NCG Company, 1275 N Grove St, Anaheim, CA 92806; tel 800-962-2611 or 714-630-4541, fax 714-630-7024.

HANDY TEST CLIP

◊ Put an end to toasted fingertips and hours of frustration when you need to get a test clip into a hard-to-reach or high-temperature area. The Model 5830 Maxigrabber test clip features a double-gripping pincer that can be rotated for easy installation, and an extended shaft. It's insulated against shorting and can be used in high-temperature environments. The clips come in red or black and are five inches long with a three-inch flexible-shaft. Retail $6.30 each. ITT Pomona Electronics, 1500 E Ninth St, PO Box 2767, Pomona, CA 91769; tel 714-469-2900, fax 714-629-3317.

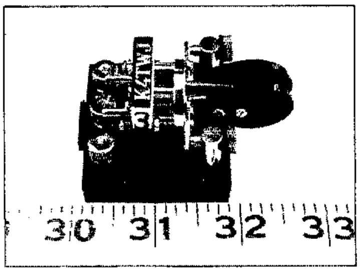


5 BANDS IN ONE

◊ The Pentenna five-band antenna covers 144, 222, 440 and 1280 MHz, and 6 or 10 meters. Quarter-wave radiators on each band use your vehicle's body as a ground plane, allowing the use of a short, 31-inch whip that won't fold back at highway speeds. The Pentenna is tuned and needs no adjusting after installation—it mounts on a standard Motorola/NMO connector. The Pentenna can be used on any of the above bands individually, and an Austin Pentaplexer provides the isolation needed for simultaneous multiband operation. Retail prices: Pentenna $95, Pentaplexer $200. Austin Antenna, 10 Main St, Gonic, NH 03839; 603-335-6339, fax 603-335-1756.

PEEWEE PADDLE

◊ Well suited for QRP, mobile, backpacking or in the shack, one of the tiniest Morse code keys ever developed is the 3-in-1 miniature twin-lever paddle. It measures a scant 1½ × 2 × 2 inches and weighs just 5 oz.



Retail price £70 (about $100 US). SASE/IRCs to Gordon Crowhurst, G4ZPY Paddle Keys International, 41 Mill Dam Ln, Burscough, Ormskirk, L40 7TG England, or in the US, c/o Dave Ingram, K4TWJ, 4941 Scenic Dr, Birmingham, AL 35210. QST

Strays

ATTENTION SCOUTERS

◊ I'm compiling a list of Scouters and Scouts who are licensed hams active in ham radio Scout Troups or Explorer Posts, or regularly give ham demonstrations to Scouts. I'm also composing a list of radio clubs which sponsor Scout Troups or Explorer Posts. Send me a note describing your involvement. Walt Garrett, NØMAL, 8 Willow Ct, Florrissant, MO 63031.

Fighting Antenna Corrosion

Yesterday your antenna was working fine—today it's on the fritz. The problem could be quite obvious, or caused by something you may not suspect.

By Scott Roleson, KC7CJ
14007 Donart Dr
Poway, CA 92064

Here's an assortment of readily available corrosion-fighting compounds. Most of the brand names can be read; the item second from the left is ALUMA-SHIELD.

At Christmastime, *Santa Claus* is supposed to be on rooftops—*I'm not!* But there I was, dismantling my antenna. Don't misunderstand—I enjoy working on antennas, and the winter weather in Southern California makes rooftop work far less arduous than in the New England or more northerly climes. But I had installed my antenna only six months earlier and was hoping to spend my holiday vacation time *operating!* Unfortunately, my otherwise trusty Butternut vertical had ceased to work—*right in the middle of the ARRL 10-Meter Contest a few weeks earlier!* One day it was fine; the next day, it simply wouldn't load properly.

At first, I suspected something simple: a broken connector, maybe. Substituting a dummy antenna for the vertical quickly showed the problem was at the antenna. As I dismantled the vertical, I noticed a fine, white powder at each joint. Apparently, over the six months since I'd installed the antenna, corrosion had crept into every connection. I hadn't anticipated this! I live in an inland part of San Diego County, where the climate is typically dry. I had tightened each clamp and bolt securely. But, my reliance on clean joints and joint pressure alone to ensure good and lasting electrical connections was obviously lacking. Murphy set the timing; Mother Nature had done the work. I was off the air until I solved my corrosion problem.

Simplified Corrosion Physics

In geologic terms, we only borrow metal for a while. After all, metal is usually found as ore, locked up in rocks in compounds of one kind or another. The extraction of metal from ore and the subsequent processing can be viewed as adding energy. This energy is just waiting to be released under the right circumstances.[1] In the long run, metal returns to its natural, corroded state.

My antenna, like most antennas, is made almost entirely of aluminum. Relative to other metals, aluminum needs *lots* of energy to extract it from ore. In the atmosphere, aluminum oxidizes readily, forming a surface oxide with the chemical name Al_2O_3—aluminum oxide: This is the fine, white powder that I found on my antenna.

Bimetallic Corrosion

Bimetallic corrosion also causes a great deal of trouble. It occurs when two metals with the right properties are brought into contact, and an electrolyte is present. It's the same chemical process that occurs in batteries. Specifically, electrons from one metal (called the *anodic* metal) flow across the junction to the other metal (called the *cathodic* metal), as shown in Fig 1. Gaseous hydrogen forms at the surface of the cathodic metal near the junction. The positive ions left in the anodic metal oxidize, either flowing into the solution, or forming an oxide layer at the junction. In bimetallic junctions, the more anodic metal is always the one that corrodes.

The electrolyte is typically some kind of salt or other compound dissolved in water, making the solution conductive. The metals need not be immersed in the electrolyte. Morning dew, or dissolving salts from fingerprints are sufficient for bimetallic corrosion to start. Acid rain is an electrolyte formed when sulfur compounds in air pollution mix with atmospheric moisture.

Galvanic Incompatibility

Metals that readily corrode when in contact with other metals are said to be *galvanically incompatible* because of their relative positions on the galvanic chart (see Table 1). This table relates the *electropotential* of metals, or metal alloys, in sea water, because sea water is a common electrolyte. The

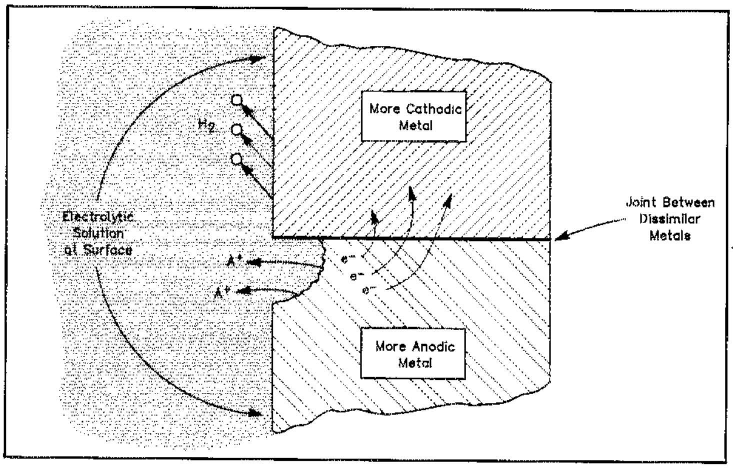


Fig 1—Bimetallic corrosion. When two dissimilar metals are joined in the presence of an electrolyte, the more anodic metal will corrode by releasing positive ions (A^+). This frees electrons (e^-) that move across the junction, forming hydrogen bubbles (H_2) at the cathode. This is the same reaction that batteries use, except the electrons flow through a circuit.

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[1] Notes appear on page 26.

A recently disassembled, partially oxidized interior joint. The short, formerly telescoped tubing section looks much better than the adjoining element section that has been continually exposed to the elements. Nevertheless, it has oxidized.

metals in Table 1 are listed in order of decreasing potential, from most anodic to most cathodic. In practice, when mating metals, it's best to choose those that are close together on the galvanic chart.

Tin and gold are metals that illustrate how troublesome bimetallic corrosion can be. Both metals are commonly used to coat electrical connectors. As you can see in Table 1, these metals are galvanically remote. Sometimes, connectors with pins coated with these metals are inadvertently attached to each other. If the contact pressure is insufficient to keep out moisture, or if the metals are used in an environment where electrolyte forms easily, the tin surface oxidizes. I've seen this happen in personal computers, where plug-in cards with gold-plated edge connectors are plugged into tin-plated motherboard connectors. The resulting problems are usually intermittent and difficult to localize. Simply removing a card and reinserting it may remove enough oxide that the problem disappears—temporarily, at least. After awhile, the oxide reforms and the problem returns.

Ensuring Good Electrical Joints

Making and keeping good electrical connections in antennas is really simple—as long as you pay attention to the basics. The best way to make good electrical connections is to start with galvanically compatible metals, then clean all connections well before assembly (refer to the accompanying photos). To make sure these connections *stay* good, seal all contact points so moisture can't enter the joint to form an electrolyte and start corrosion. In my case, I had done the first, but not the second.

Electrical contact at mating surfaces occurs between microscopic bumps and points where the metals meet. Joint impedance is proportional to the number of points in contact. A smooth, clean surface ensures that there will be lots of these points and little between them to get in the way. For antennas, I've found it's best to first buff all joining parts with steel wool, emery cloth or a wire brush, then with a nylon scouring pad (ScotchBright or equivalent). For metal tubing, it's important not to forget to clean the *inside surfaces* of telescoping parts. I wrap steel wool around a pencil or form it into a pencil-like shape so I can get to the tubing's inside surface. Finally, I use a clean rag to wipe off any powdered metal and oxide. I do my best to refrain from touching the mating surfaces and contaminating them with body oils. (Using cotton gloves during antenna assembly is a good idea.)

Mating-Surface Pressure

Pressure between mating surfaces is important. Oxides start to form immediately, so there must be enough pressure to break through the oxide layer. Furthermore, contact pressures must be high enough to ensure joints are stable and tight during normal flexing. A phenomenon called *fretting corrosion*[2] occurs when a connection is repeatedly opened and closed. (See the sidebar, "Fretting Corrosion.") The closure breaks through a fresh surface oxide, so fresh metal is in contact with fresh metal. However, when the contact opens, this fresh metal is again subject to oxidation. With time, the oxidation builds up. Eventually, the contact pressure isn't enough to break through the oxide layer and the connection fails.

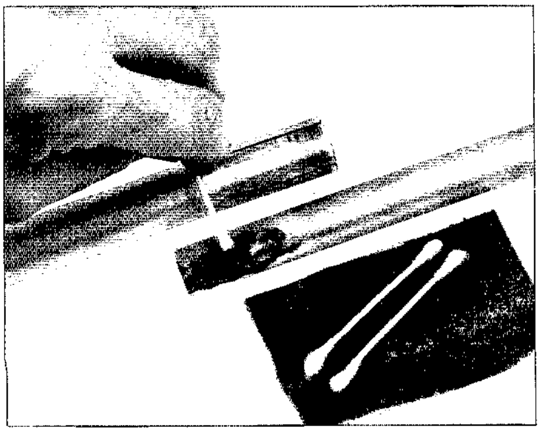
Table 1
Relative Galvanic Series in Sea Water

*** Anodic End ***
Magnesium
Zinc
Aluminum
Mild steel
Iron
50-50 lead/tin solder
Stainless steel (type 304 & 316)
Tin
Nickel (active)
Brass
Aluminum-bronze
Copper
Nickel (passive)
Silver
Gold
*** Cathodic End ***

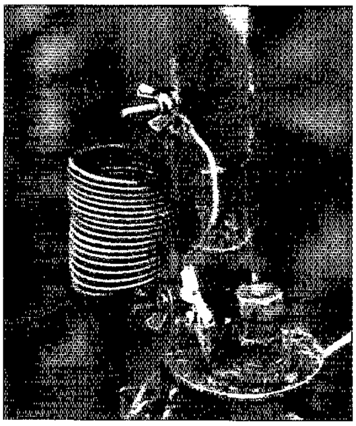
Minimizing Corrosion

Galvanic Similarity

To minimize bimetallic corrosion at contact points, use parts that are galvanically similar. Then, too, assemble them for minimum galvanic differences. For example, aluminum and copper are not very compatible because they are far apart on the galvanic table (see Table 1). Yet, the connection of a copper wire to an aluminum antenna part is often needed. So, tin or solder-plate the copper wire, forming a gas-tight seal between the copper and plating. Then, use stainless-steel hardware[3] to secure the connection. Stainless steel is galvanically close to both tin and aluminum, and won't rust like other steel hardware. Use a stainless-steel washer between an aluminum surface and a tinned wire, or lug, connected to it. This is because the galvanic difference between tin and stainless steel—or aluminum and stainless steel—is less than tin and aluminum. This should help the joint last even longer if (or when) an electrolyte gets into the joint.



After scrubbing the aluminum to remove the oxide, an antioxidant (Ox Gard, in this case) is applied to both sections of the telescoping tubing. *(photos by Kirk Kleinschmidt, NT0Z)*



After assembly, this section of an antenna was painted. I used a brightly colored paint; in this case, red. Painting is especially useful where plastic tape is difficult to apply, or where an impracticably large amount of tape would be needed. When you need to disassemble a joint for cleaning, the bright color makes it easier to see if you've removed all the paint. *(photo by the author)*

Fretting Corrosion

Fretting corrosion occurs whenever there is a small repetitive or cyclical motion at a metal-to-metal joint or seam. This motion can be either a make-break or sliding action. Basically, joint flexing or sliding exposes clean, fresh metal to the air, and an oxide begins to form. This oxide is broken or scraped away with each reconnection or flex, but begins to reform when it is exposed again. Over time, the oxide builds up at the joint, and eventually the reconnection or scraping won't remove all the oxide at the point of contact. Joint resistance gradually increases, and the joint eventually becomes intermittent. How long a joint or contact lasts before it becomes a problem depends on joint pressure. It also depends on how oxidation-prone the metals at the joint are, and other conditions such as temperature, humidity and the existence of surface contaminants.

For example, automotive engineers are concerned about tin-plated connector pins used in automobiles because of fretting corrosion.[1] The trend in automotive connectors is to use more pins (more things to go wrong), make them smaller (with smaller contact surfaces), and to use lower insertion force to make assembly easier. These connectors are also subject to vibration, elevated temperatures and rain or salty mud in some locations. All these things aggravate the fretting corrosion problem.

Over time, normal antenna flexing in the wind can cause fretting corrosion at joints if they're not assembled correctly and well protected. To minimize fretting corrosion at antenna joints:

- Design the joints to have large contact-surface areas.
- Thoroughly clean mating surfaces before assembly.
- Securely fasten joints to minimize flexing.
- Use the highest practical contact force.
- Use oxide-inhibiting lubricants or pastes.
- Protect finished joints from the atmosphere with paint, tape, or other overcoatings.

[1] M. Lee, A. Mao, and M. S. Mamrick, "Fretting Corrosion of Tin at Elevated Temperatures," *Proc. 34th Meeting of IEEE Holm Conference on Electrical Contacts*, San Francisco, Sep 26-29, 1988, pp 87-91.

Use Star Washers

Another way to form a gas-tight seal with hardware is to use star washers. Star washers break through oxides and cut into the mating surfaces under the pressure of the screw and nut used to hold them in place, so that fresh metal is in contact with fresh metal. In some cases, you can get the same effect using sheet-metal screws (again, I prefer stainless-steel screws). This technique works best for hardware that is to go on once, or at most, only a few times. Repeated assembly can damage a surface, possibly providing a path for moisture to enter the joint.

Joint Compounds

Special pastes, fluids, and other joint compounds help seal joints and inhibit electrolyte action.[4] These compounds are squeezed away from the microscopic points of contact by mating pressure, but fill in around the points of contact to inhibit air and moisture passage. For aluminum antennas, I use pastes available for aluminum electrical wiring. These pastes are available at electrical-supply houses, hardware stores and building-supply outlets. A *few* antenna manufacturers provide an antioxidant compound with their antennas.[5] The joint compound can be spread with a cotton-tipped applicator. Wire-brushing the surface after compound application helps ensure all quickly formed oxide is broken down and prevents a new layer from forming. Toothbrush-size wire brushes are available from most hardware stores.

Joint compounds, unfortunately, don't last forever. They harden and crack, or with time and temperature, simply flow away from joints. For this reason, some sort of finishing barrier or overcoat is needed, such as plastic tape, paint, or silicone rubber sealant (bathtub caulking). Choose a material that is flexible and resistant to ultraviolet light. Many paints and plastic tapes eventually harden and become brittle from exposure to ultraviolet light. Normal antenna flexing in the wind cracks them, and moisture seeps in through these cracks. On the other hand, this also argues for regular maintenance. Plastic tape is easy to remove and replace. If you use paint, use a bright color: red or yellow. If you need to disassemble a joint for cleaning, the bright color is easier to see so you can tell if you've removed it all.

Summary

All this probably sounds like a great deal of trouble to go through to keep an antenna working. It is—especially when the antenna is new and you want to get it on the air in a hurry! The key to antenna longevity is to do *all*, not just some, of what I've told you about. The effort ensures reliable and long-lasting antenna-element joints and connections. My vertical antenna is still working fine after two years.

Antenna work is fun, but I'd much rather build new ones than rebuild an old one again and again, wouldn't you?

Notes

[1] L. Van Delinder, ed., *Corrosion Basics*, from the National Association of Corrosion Engineers, 1440 South Creek Dr, Houston, TX 77084, pp 23-32.

[2] J. Whitley, "Investigation of Fretting Corrosion Phenomena in Electric Contacts," *Proceedings 8th International Conference on Electric Contact Phenomena*, Tokyo, Japan, Aug 22-26, 1976. (Copies of this paper can be obtained from Amp, Inc, PO Box 3608, Harrisburg, PA 17105, tel 717-564-0100, as their publication #82418.)

[3] Stainless-steel hardware is available from Small Parts, 13980 NW 58th Ct, PO Box 4650, Miami Lakes, FL 33014-0650, tel 305-557-8222, fax 800-423-9009. (If you buy stainless-steel hardware locally—especially hose clamps—ensure the *entire* clamp is made of stainless steel! There are "stainless-steel" hose clamps of which only the *band*—not the screw mechanism—is made of stainless steel.—*Ed.*)

[4] S. Leibson, "Fluids Vanquish Intermittent Contacts," *EDN*, Mar 14, 1991, pp 59-64.

[5] An extensive list of corrosion-fighting compounds is available free from the ARRL. Address your request for the ARRL LAB's ANTIOXIDANTS TEMPLATE to: Technical Department Secretary, ARRL, 225 Main St, Newington, CT 06111. Please be sure to enclose a business-size SASE.

Scott Roleson was first licensed as a Novice (WA2NYR) in 1965. He earned his Advanced class license while living in Arizona in the early 1980s. Scott's ham radio interests include antennas and the occasional—and all too infrequent—casual ragchew. Scott credits his becoming an engineer to having his ham radio hobby get out of control!

Before getting his BSEE from Arizona State University and his MSEE from the University of Arizona, Scott worked for Motorola as an electronics technician on RF communications and radar equipment, and in radiation-effects testing at Bell Telephone Laboratories. He has over 10 years experience in Electromagnetic Compatibility (EMC) engineering and works at Hewlett-Packard's San Diego facility. Scott has published many technical papers and articles, mostly on EMC subjects. He is a member of the IEEE and the IEEE EMC and Antennas & Propagation societies.

A Multimode Phasing Exciter for 1 to 500 MHz

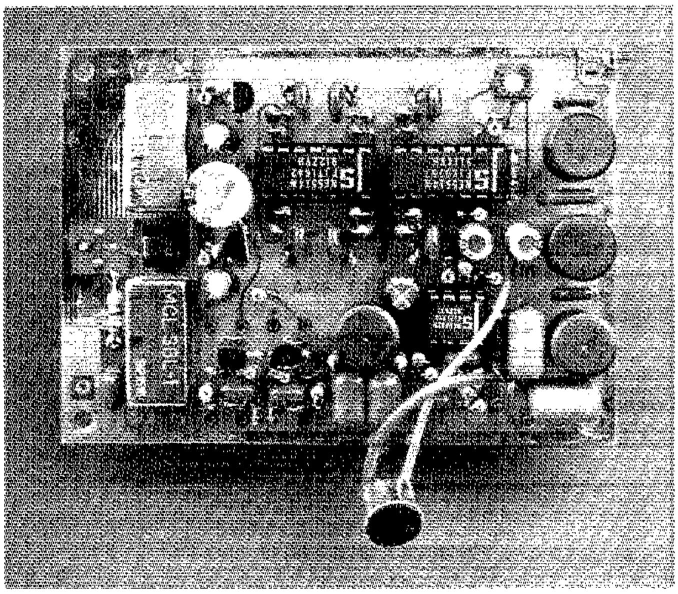
In January, *QST* introduced you to R2, a single-signal direct-conversion receiver for the '90s. This month meet T2, a matching transmitter that generates SSB, CW and more.

By Rick Campbell, KK7B
Department of Electrical Engineering
Michigan Technological University
Houghton, MI 49931

The high-performance direct-conversion receivers described in the references[1,2] may be easily paired with a CW transmitter to make a simple transceiver of very respectable performance. For QRP operation and VHF through microwave weak-signal work, CW is the preferred mode. Whether you enjoy CW operation, or simply use it as a tool to make otherwise impossible contacts, it's nice to be able to switch to a voice mode when the signals are strong and the technical ideas are flowing hot and heavy. Here is a little multimode exciter that makes a perfect companion to a direct-conversion receiver.

Circuit Description

Fig 1 is the block diagram of a direct RF phasing-type SSB exciter. This block diagram has appeared in every *ARRL Handbook* for 40 years, and readers are encouraged to review the basics in those pages.[3] In the 1950s, the blocks contained vacuum tubes, paper capacitors and possibly a Barker and Williamson 2Q4 audio phase-shift network plugged into an octal tube socket. The schematic in Fig 2 is an implementation of Fig 1 using modern components.



Close-up of the top of the 2.5 × 3.5-inch T2 PC board. The power combiner (U6) and mixers (U4, U5) are at the left-hand edge of the board. The audio phase-shift networks (U2, U3) are arranged along the upper half of the board. The audio high-pass (C11, L1, C12) and band-pass (C13-C15, L2, L3) filters are arranged along the right edge. The mike amplifier and sidetone generator (U1) are in the lower right portion, and the control and power-supply circuitry is at the lower left. The MMIC amplifier (U7) and low-pass filter components are located on the back side of the board.

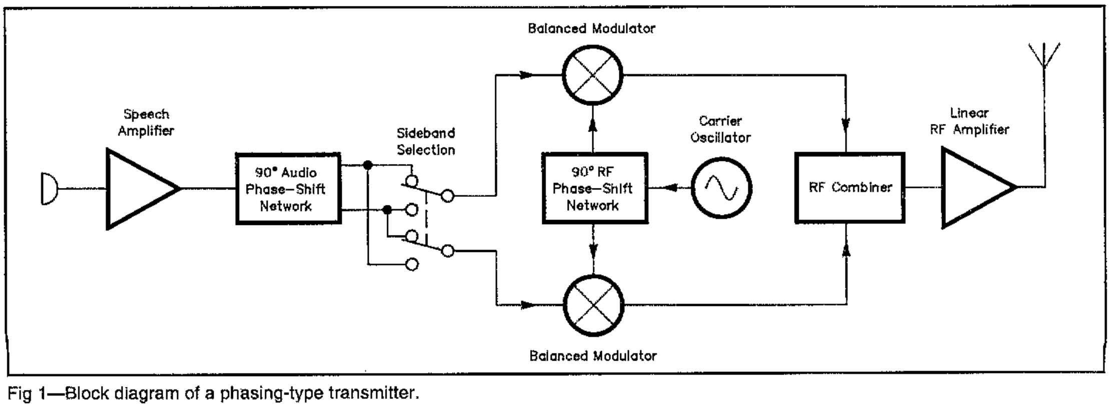


Fig 1—Block diagram of a phasing-type transmitter.

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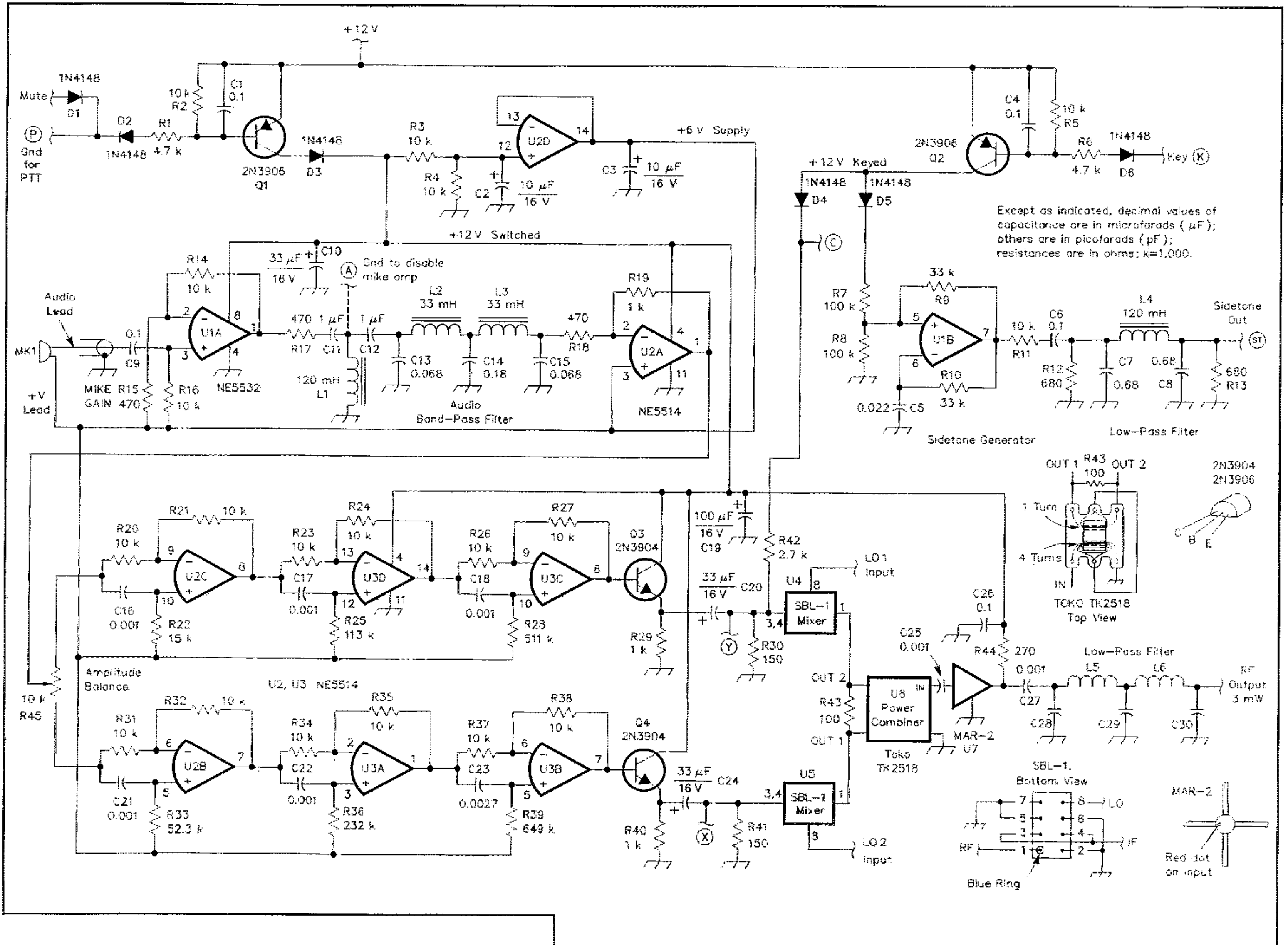
[1] Notes appear on page 31.



Except as indicated, decimal values of capacitance are in microfarads (µF); others are in picofarads (pF); resistances are in ohms; k=1,000.

Fig 2—The T2 transmitter circuit diagram. The local oscillator and LO phase-shift network are located off the board. The resistors in the audio phase-shift networks (R20-R28 and R31-R39) are 1%-tolerance metal-film units. Other resistors are 5%-tolerance carbon-film or -composition units. Inductors are Toko 10RB series fixed inductors. Polarized capacitors are aluminum electrolytics rated at 16 V dc. The capacitors in the audio phase-shift networks (C16-C18 and C21-C23) are 1%-tolerance XICON polyester-film units available from Mouser. C7, C8 and C13-15 in the audio band-pass and sidetone low-pass filters are Panasonic V-series 50-V, 5%-tolerance metalized film capacitors. C28-C30 in the RF low-pass filter can be small 5%-tolerance leaded or chip components (see text). The other capacitors can be metalized polyester or ceramic-disc capacitors. See Note 7 for part sources.

C28-C30—See text and Table 1.
L5, L6—See text and Table 1.
MK1—Electret mike (Radio Shack 270-092 or Mouser 25LM045).
R45—10-kΩ, single-turn, PC-board-mount control (Panasonic MAG14 or equivalent).
U4, U5—Mini-Circuits SBL-1 double balanced mixer. Other mixers may be substituted if other frequency ranges are desired.
U6—Toko TK-2518 power combiner/splitter (this part is called a balun transformer in the Digi-Key catalog). See text.
U7—Mini-Circuits MAR-2 MMIC. See text.

Table 1
Fifth-Order Butterworth Low-Pass Filter Values

Frequency (MHz)	*C28, C30 (pF)*	*L5, L6 (µH)*	*C29 (pF)*
7	273	1.8	884
28	68	0.447	220
50	39	0.257	127
144	13.6	0.0893	44.1
222	8.9	0.0579	28.7
432	4.5	0.0298	14.7

Historians may enjoy comparing Fig 2 with Fig 11-5 on page 307 of the 1959 *ARRL Handbook*. The functions are identical, but virtually all the parts have changed! Following the op-amp microphone amplifier (U1A) is an LC audio band-pass filter, just like the old days, but designed using modern network theory and a computer circuit analysis program (*PSPICE*).[4]

The audio phase-shift networks (U2B, U3A, U3B and U2C, U3C, U3D) are the same as in the R2 receiver (see Note 2). Second-order (like the old 2Q4), third-order and fourth-order networks were explored for the audio phase-shift networks. Fourth-order networks work very well with ideal components, providing more than 45 dB of opposite-sideband suppression in my computer model. The third-order network with ideal components can provide more than 40 dB of opposite-sideband suppression. When 1% component variations were included in the computer models, the fourth-order network model degraded to within a few decibels of the third-order model. For this design, third-order networks with 1% off-the-shelf components are specified. This provides better opposite-sideband suppression than the phasing exciters of the 1950s and '60s, without requiring hand-selection of any components.

One critical area in a phasing SSB generator is the pair of audio amplifiers between the audio phase-shift network and the balanced modulators. These amplifiers have three requirements:

1) They must have identical phase shifts. This is most easily achieved by making them very broadband, by using large-value coupling capacitors.

2) They must have stable gain. If the gain of one of the amplifiers changes, the opposite-sideband suppression is reduced. Stable gain is most easily obtained by using feedback to set the gain, or using emitter followers with a gain slightly less than one.

3) They must have exceptionally low distortion. Any distortion introduced at this point has two equally undesirable effects: The distortion sends high-frequency audio harmonic energy into the balanced modulators; and the distortion products no longer have the desired 90-degree phase relationship.

The successful implementation of point 3 is what separates an acceptable phasing-type SSB generator from an unacceptable one. A phasing-type generator with 10% total harmonic distortion (THD) sounds pretty good on the air, because most of the 10% distortion falls outside the desired signal bandwidth and is rejected by the *receiver*. On either side of the transmitted signal, however, is a wide spectrum of garbage, kindly referred to as "splatter." To eliminate splatter, a phasing-type SSB generator must have audio amplifiers with THD well below 1%.

In contrast, a filter-type SSB generator may have large amounts of distortion in any of the stages preceding the filter. Distortion ("clipping") is often introduced intentionally to increase the average-to-peak-power ratio, or unintentionally by operators who turn up the mike gain to squeeze out the last milliwatt. A phasing-type SSB generator requires low-distortion audio and RF design from the input of the audio phase-shift network to the output of the RF combiner. This is why phasing-type SSB transmitters have a reputation for exceptional audio quality.

The emitter-follower drivers used here (Q3, Q4) were modeled on *PSPICE*, and have less than 0.3% THD when driving the IF ports of the SBL-1 double-balanced mixers (U4, U5). The onset of distortion in the emitter followers occurs above the desired drive level for the SBL-1s. The 150-ohm resistors (R30, R41) were added from the mixer IF ports to ground after I burned out a pair of SBL-1s by dumping the charge on the 33 µF coupling capacitors (C20, C24) through the mixer diodes.

The two double sideband (DSB) signals from the SBL-1 balanced modulators are combined in a Toko TK2518 splitter-combiner. This part is rated from 20 to 600 MHz, but it seems to work as low as 7 MHz in this application. A better choice for lower frequencies is a Mini-Circuits PSC2-1, which must be mounted off the board, or a homebrew replacement as shown in Fig 4 of the January *QST* receiver article (see Note 2).

After the combiner, the signals are amplified by U7, a Mini-Circuits MAR-2 MMIC (monolithic microwave integrated circuit). Readers unfamiliar with MMICs are in for a treat—this part has 50 ohm input and output impedances, boosts the signal up to a few milliwatts with low distortion, is unconditionally stable, and costs 99 cents. It also provides a constant load impedance for the combiner, so that opposite-sideband suppression is not a function of the following circuitry. You can read more about these remarkable devices in the 1987 *QST* series by Al Ward, WB5LUA.[5]

Following the MMIC amplifier is a fifth-order low-pass filter. I debated whether to include the low-pass filter on the PC board, since it is the *only* part of the exciter that must be changed for different bands. I finally included it because it allows the exciter to directly drive an antenna for milliwatt QRP work, or a transverter for the microwaves, or the IF port of a mixer in a heterodyne transmitter. For multiband applications, or for directly driving a linear amplifier, the low-pass filter components can be omitted, with jumpers replacing the inductors. Design equations for low-pass filter elements are given in Chapter 2 of any recent *ARRL Handbook*, and Table 1 gives typical values for a few amateur bands.

The PC-board low-pass filter artwork will accommodate chip capacitors or small leaded components. I prefer chip caps for frequencies above 30 MHz. The inductors can be air core at VHF and small toroids at HF. For lower frequencies, the low-pass components may be too large to fit on the PC artwork. A separate low-pass filter may be constructed off the board in that case.

Several different mixers were tried at U4 and U5 in the prototype exciters. Mini-Circuits SRA-2CM mixers will work up to 1000 MHz, and SBL-3s will work down to 25 kHz, with no other changes in the exciter. Options for LO phase-shift networks are discussed in January *QST*'s receiver article (see Note 2) and *The ARRL Handbook*.

More Power for T2

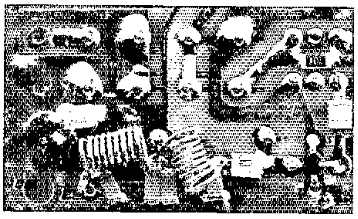
Once you have built a copy of T2 for your favorite band, the next logical question is, "How can I get more power?" Even the most dedicated QRP enthusiasts require more than 3 mW on occasion. Fortunately, League publications offer a number of good designs for a variety of frequencies and power levels.

If you're simply interested in CW operation, the amplifier sections of the QRP transceivers shown in Chapter 31 of the 1993 *ARRL Handbook* are a great place to start. Another place to look is in the projects described in *QRP Classics.*

You probably didn't build T2 just for CW operation, though. SSB operation requires *linear* amplifier stages. The following references describe linear amplifiers that operate over a wide range of frequencies and power levels.

- W. Hayward and D. DeMaw, *Solid State Design for the Radio Amateur,* published by the ARRL. Chapter 8 contains a variety of linear circuits.
- A. Ward, "Monolithic Microwave Integrated Circuits, Part 2" *QST*, Mar 1987, pp 22-28, 33. Describes low-power, broadband, no-tune amplifiers that work from 160 meters through UHF. Also covers combining MMICs for additional power. Later development: Mini-Circuits MAV-11 and Avantek MSA-1104 devices are good for 50 mW (linear) through 432 MHz.
- Z. Lau, "A 1.8 to 54 MHz 5-Watt Amplifier," *QEX*, May 1992, pp 7-8. Uses an MRF137 and a 28-V supply to generate a clean 5-W signal. A gain compensation stage allows it to be driven with 5-10 mW.
- W. Hayward, "Stable HEXFET RF Power Amplifiers," Technical Correspondence, *QST*, Nov 1989, pp 38-40. Shows several amplifiers using the IRF511. Gain and power output varies. Describes an 8-W, 3.5-14 MHz amplifier that runs on 13.5 V.
- Z. Lau, "PIN Diode-Switched Amplifier," Technical Correspondence, *QST*, Feb 1988, p 42. Describes a 1.4-mW input, 1-W output 40-meter amplifier using an MRF237.
- W. Hayward, "A QRP SSB/CW Transceiver for 14 MHz, Part 2" *QST*, Jan 1990, pp 28-31. Describes a 1-W, 20-m HEXFET amplifier and 100-mW 2N5859 driver.

For VHF or UHF work, Chapters 31 and 32 of the 1993 *ARRL Handbook* show how to get from mixer levels to useful power levels on 50, 144, 222 and 1296 MHz. For 432 MHz, see E. Krome, "A High-Performance, Easy-to-Build 432-MHz Transverter, Part 2," *QST*, Sep 1991, pp 18-21. For 903 MHz, see D. Mascaro, "903 MHz Linear Amplifiers," *QST*, Part 1—June 1990, pp 18-25; Part 2—July 1990, pp 28-31. A good general overview of the use of VHF and UHF hybrid power modules is B. Olson, "Hybrid Modules: Building with Bricks," *QEX*, Jul 1988, pp 13-14.—*Zack Lau, KH6CP*



Close-up of the back side of the T2 PC board opposite the mixers. The MMIC amplifier (U7), its coupling capacitors (C25, C27) and the low-pass filter components (C28-C30, L5, L6) are soldered to this side of the board. R43 is also mounted on this side of the board, underneath the power combiner (U6). Although I used chip components for C25, C27-C30 and R43, small leaded components will work as well.

Frosting the Cake

I like to build portable VHF and microwave rigs, and as soon as I work somebody on a 10-mile path I always go farther away and try again. That means at some point I have to switch from SSB to CW. Then we get our antennas aligned and switch back to SSB. The first prototype used just a resistor to unbalance one of the mixers for CW. I hand-wired all the TR switching and didn't put in a sidetone generator because it was too much trouble. I then had a really cute little microwave transceiver and a great big Morsematic keyer that I have to haul up the mountain because I was too lazy to put sidetone in the little microwave rig.

After the second prototype, I decided to put some convenience circuitry on the exciter circuit board to make it easy to build a transceiver.[6] The extra stuff includes Q2, for CW keying; Q1, the TR-switch transistor; and U1B, a sidetone generator. Since I have low tolerance for bad audio, I added a low-pass filter on the sidetone generator to convert the square wave output of the relaxation oscillator to a fairly decent sine wave. Point P is grounded to transmit, Point K is connected to the key for CW, and Point ST is a high-impedance sidetone output that can be routed to an appropriate spot in the receiver audio amplifier. Point A is grounded to disable the mike amp for CW. Interconnections will be discussed in more detail next month.

Other Modes

Old timers may remember that some of the early phasing transmitters offered CW, USB, LSB, AM, DSB suppressed carrier, and narrow-band FM (or PM) at the flip of a front-panel switch. It is possible to generate other voice modes besides USB and LSB with the T2 exciter.

- To generate AM, apply 12 V to point C (to insert carrier) and break the circuit at point X to disable the quadrature modulator.
- To generate narrowband PM, apply 12 V to point C and break the circuit at point Y to disable the modulator that is in phase with the carrier. Mathematicians may want to verify that a DSB suppressed-carrier signal plus a 90° phase-shifted carrier is in fact narrowband phase modulation. The more experimentally inclined can simply listen to the various signals on a multimode receiver (I did both!).

All kinds of advanced digital modes can be generated by injecting the appropriate signals at X and Y, but that's another story.

Performance and Applications

A 2-meter version of this exciter has 40 dB of carrier and opposite-sideband suppression. Distortion products are more than 30 dB down at +3 dBm (2 mW) output. The distortion spectrum falls off very rapidly outside the desired passband. The audio quality is about average for an AM broadcast station—noticeably superior to the average SSB signal on the ham bands. The +3 dBm output level is ideal for upconverting to the microwave bands, heterodyning with a VFO to the HF bands, working across town on 222.1 MHz or driving a linear-amplifier module.

For some applications (running the legal limit on a wide open HF band comes to mind), the casual use of a direct phasing exciter is inappropriate. The potential for interference is too high. A carrier suppressed 40 dB below 1500 watts is 150 milliwatts, and that is enough to work the world on a wide-open HF band. For rock-crushing signals on crowded bands, I recommend operating the phasing exciter at a fixed frequency, following it with a crystal filter, and heterodyning to the desired output frequency, as is done in the Kenwood TS-950SDX. Personally, I'd rather operate a really big bulldozer at a landfill than a legal limit amplifier on a crowded band (the two have a lot in common), but I can find dozens of applications for a multimode exciter that's smaller than most microphones and works on any ham band below 1000 MHz!

A phasing "QRP Gallon" (five watt) transmitter will only have 500 microwatts of carrier and opposite-sideband signal, and conditions have to be really good for 500 microwatts to interfere with anyone. See the sidebar "More Power for T2" for references to some suitable previously published power amplifiers.

On VHF through microwave frequencies, it is routine to make SSB and CW contacts out to a hundred miles or so with very low power from hilltops. A crystal-controlled phasing exciter makes a nice companion to a multimode scanner at a scenic overlook. For the higher bands, the phasing exciter can serve as the IF for a transverter—details next month.

Tuning

The LO phase-shift network (not included on the board; see the January *QST* article referenced in Note 2) and **AMPLITUDE-BALANCE** pot (R45) can be set for best opposite-sideband suppression while listening to the wrong sideband on a receiver with good selectivity. I tack-soldered a 100-kΩ resistor from the output of the sidetone generator to the mike input to provide a tune-up tone. By alternately adjusting the LO phase-shift network and **AMPLITUDE BALANCE** pot, it should be possible to reduce the wrong-sideband tone by about 43 dB.

It is difficult to judge the carrier suppression in a direct phasing transmitter by listening to the signal in the station receiver unless the carrier oscillator and receiver are very well shielded. The carrier oscillator puts out at least 10 milliwatts, and the carrier signal will probably be loud in the receiver even with the antenna disconnected. I measure carrier suppression by heterodyning the exciter to a different frequency using a separate crystal oscillator and high-level double-balanced mixer.

I chose the audio amplifier gain for low-distortion SSB with the Radio Shack electret microphone shown in the schematic. For different microphones (or different operating habits), the value of R15 should be changed while listening to the audio and opposite sideband in the station receiver. I recognize that the absence of a front-panel mike gain control is an inconvenience, but the consequences of audio distortion in a phasing rig are far more serious than in a filter rig. R15 should be selected and then left alone. There is just too much temptation to turn up the gain a little to get that last half decibel out when working a weak signal!

The opposite sideband may be selected by reversing the LO connections at the mixers. It may be necessary to readjust the LO phase shift network and amplitude balance pot when changing sidebands.

Conclusions

The photos show how the parts are arranged on the PC board. This little SSB exciter board looks nice and works well from VLF to the low microwaves. The first prototype worked the first time I applied 12 V and an audio input signal. The problem with a fine-looking PC board layout is that it discourages further experimentation.

One of the requirements I impose on my projects is that they must be reproducible. In most cases that implies an engineering cycle that starts with a concept, explores lots of options, creates a block diagram and schematic, tests all of the circuitry on the computer and at the bench, lays out and etches a PC board, builds a prototype, debugs and tests the prototype and then lays out a second, third and often a fourth PC board. When I am finished, I am confident that anyone can put the parts on the PC board and build a working copy of my project.

That is not the only way to build a radio, and it is not necessarily the best way. Most homebrewers are familiar with a technique called "ugly" construction, in which the circuit is built, designed, tested, modified, redesigned, optimized, used on the air, and reoptimized in a continuous creative process. "Ugly" equipment is never finished, and always makes the best use of available parts and the homebrewer's talent. Ugly circuits often work better than nice-looking PC boards for two reasons:

- The continuous ground plane and short ground connections reduce or eliminate ground-current problems and improve shielding.
- The fact that the circuit is already ugly encourages further experimentation and improvement.

Building and improving ugly constructed RF projects is the best way to learn about RF design. As soon as you read about a new technique you can try it out on the bench. Ugly construction allows the homebrewer to be creative all the way through a project, and for me, that is the fun of ham radio.

Next month, I'll give more details about integrating the R2 and T2 boards into a station and show some interesting VHF and microwave applications.

Notes

[1] R. Campbell, "High-Performance Direct-Conversion Receivers," *QST*, Aug 1992, pp 19-28

[2] R. Campbell, "High-Performance Single-Signal Direct-Conversion Receivers," *QST*, Jan 1992, pp 32-40.

[3] This discussion appears in Chapter 18 of the 1985 through 1993 editions of *The ARRL Handbook*.

[4] *PSPICE1* Circuit Analysis Software is available from the Microsim Corp, 20 Fairbanks, Irvine, CA 92718, tel 714-770-3022.

[5] A. Ward, "Monolithic Microwave Integrated Circuits," *QST*, Part 1—Feb 1987, pp 23-29, 32; Part 2—Mar 1987, pp 22-28, 33.

[6] Etched, plated and drilled PC boards (double-sided, with plated through-holes) for the T2 board are available from Applied Radio Science, PO Box 225, Houghton, MI 49931 for $10 postpaid (send an SASE for a catalog with current kit information). For individuals who want to make their own PC boards, an etching template/part-overlay package for the T2 PC board is available from the ARRL for an SASE. Address your request for the CAMPBELL T2 BOARD TEMPLATE to Technical Department Secretary, ARRL, 225 Main St, Newington, CT 06111.

[7] One source for SBL-1 mixers and MAR-2 MMICs is Oak Hills Research, 20879 Madison St, Big Rapids, MI 49307, tel 616-796-0920. XICON 1%-tolerance polyester-film capacitors used in the audio phase-shift network are available from Mouser Electronics, 2401 Hwy 287 N. Mansfield, TX 76063, tel 800-346-6873, 817-483-4422, fax 817-483-0931. All other parts are available from Digi-Key, PO Box 677, Thief River Falls, MN 56701-0677, tel 800-344-4539, 218-681-6674, fax 218-681-3880.

Strays

QST congratulates...

◊ Seven FCC employees on being awarded the the Commission's highest honors, the Distinguished Service and the Meritorious Service awards. They were presented to FCC employees who have advanced the mission and objectives of the Commission through their sustained extraordinary or exceptional accomplishments. FCC Chairman Alfred Sikes (who resigned his post in January) presented the awards at an agency-wide ceremony December 16, 1992.

Distinguished Service Award gold medals were presented to Managing Director Andrew S. Fishel and Robert M. Pepper, chief of the Office of Plans and Policy.

Meritorious Service silver medals were presented to Janet S. Amaya, assistant chief of management and personnel in the Mass Media Bureau; James L. Ball, associate director of the Office of International Communications; Sheldon M. Guttman, associate general counsel of the Office of General Counsel; Michael B. Hayden, AK3F, chief of the Private Radio Bureau's Microwave Branch; and Alexander J. Zimny, N2EDP, engineer-in-charge of the FCC's New York City Field Operations Bureau.

◊ The following amateurs on 70 years of ARRL membership:

- Frank Gunther, W2ALS, Staten Island, NY
- Robert Morris, W2LV, Sparta, NJ
- Lewis Elicker Jr, W3ADE, Harrisburg, PA
- Bartow Huff, W5AHD, Lawton, OK
- Everett Gibbs, W8AQ, Canal Fulton, OH

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- Sidney Bear, W1ASP, New Haven, CT
- Laurence Stein Jr. W1BIY, Hingham, MA
- James Eckersley, K2IXE, Forked River, NJ
- George Montgomery, W3KVS, Washington, PA
- William Tucker, W4FXE, Hallandale, FL
- John Wallace, W6EP, Westlake Village, CA
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- Edward Van Peenen, W8DM, Kalamazoo, MI
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- Frank Finger, NU1A, New London, NH
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- Joseph Hoffmann, W2DIJ, Daytona Beach, FL
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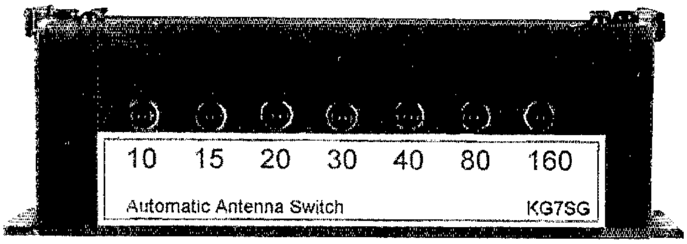
- Howard Bullock, W4LBM, Alexandria, VA
- William Vandal, W4LPD, Palm Bay, FL
- Vance Gildersleeve, K5CF, Poteau, OK
- Louis Lechenger, W5IHL, Houston, TX
- Norman Wehrli, W5JPC, Athens, TX
- Carson Nohrnberg, W6MKO, Berkeley, CA
- Roy Tucker, N6TK, La Mirada, CA
- Richard McNutt, W8CAT, White Lake, MI
- Carroll De Groff, W9OLY, South Bend, IN
- J. Harold Gibson, W9PQO, Fair Grove, MO
- G. Theodore Anderson, W9UHD, Tuscola, IL
- Jack Miller, W9WYN, Brookfield, IL
- Thornton Lyford, W0QIK, Arden Hills, MN

A Remotely Controlled Antenna Switch

Got a feed-line rat's nest? Here's a solution: an inexpensive and easy-to-build remotely controlled antenna switch that works automatically or manually with ICOM MF/HF rigs—and manually with others!



(photos by the author)

By Nigel Thompson, KG7SG
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Carnation, WA 98014

Until recently, my ICOM IC-735 fed an assortment of dipole antennas through several coaxial feed lines. The feed lines connected to a large rotary switch next to the transceiver. In planning to add a multiband beam to my collection, I wondered: "Why do I need so many feed lines when I can use only one antenna at a time?" What I really wanted—and needed—was a *single* feed line and a remotely controlled antenna switch!

The remotely controlled antenna switch I built and describe here works with ICOM MF/HF rigs to *automatically* select the correct antenna for the band you're working. It's a simple design, the switch requires no set up or alignment and it can be adapted in various ways to produce custom systems. Low-cost parts are used throughout and they're items available from a number of component suppliers, including Radio Shack. If you don't have an ICOM rig, you can use a manually controlled version of the switch.

Description

Fig 1 is a block diagram of the system. The decoder sits in the shack next to the rig and connects to it via an accessory connector on the back of the transceiver. A multiconductor cable with 7-pin male DIN plugs interconnects the control box and transceiver. A single length of RG-8 coax and a multiconductor cable run from the rig and the decoder box to the remotely located switch assembly. The remotely located relay box is equipped with seven SO-239 connectors: one for the feed line (**COMMON**) and one for each of the antennas I use.

The Decoder

ICOM MF/HF rigs have a band signal available on a rear-panel accessory connector (**ACC2**). It's designed to be used for automatically selecting the band of operation of a power amplifier connected to the rig. This connector also provides an 8.0-V reference and a 13.8-V dc power-supply tap. Table 1 shows the output voltage at the accessory socket when the radio is switched to the various bands. Note that the 18-MHz-band signal *is the same* as that of the 21-MHz band, and the 24-MHz-band signal *is the same* as that of the 28-MHz band. This means that the ICOM band-control scheme *does not support the use of separate antennas for 18 and 21 MHz, or 24 and 28 MHz.*

These are the **ACC2** connector-pin assignments for my ICOM 735:

Pin 1 +8 V reference
Pin 2 Ground
Pin 4 Band-output voltage
Pin 7 +13.8-V supply

My First Try

In designing the decoder circuit, I tried two different approaches. The first was based on an idea I got from Mark McKibbin, WR7V, and used a bar graph decoder/driver circuit. The idea is to have the decoder IC sense the rig's band signal and use the LED output drivers to control the relays, giving a one-IC solution to the problem.

Unfortunately, the band-output voltage *does not change in linear steps*, and has one voltage of zero which results in no output from the decoder IC. I added a couple of op amps to the design and got it to work, but I found that the circuit was rather temperamental and required some set up. I wanted something more robust, so I opted for a more simplistic design.

Second Try

Fig 2 shows the final circuit diagram of the controller. R1 through R7 divide the rig-provided reference voltage to define midpoints between the band-signal levels. Three LM339 quad comparators do the work of

Table 1
ICOM Accessory Connector Output Voltages by Band

Band (MHz)	*Output Voltage*
1.8	7 - 8.0
3.5	6 - 6.5
7	5 - 5.5
14	4 - 4.5
18, 21	3 - 3.5
24, 28	2 - 2.5
10	0 - 1.2

Note: The voltage step between bands *is not constant*, but close to 1.0 V, and the 10-MHz band is *not in sequence with the others.*

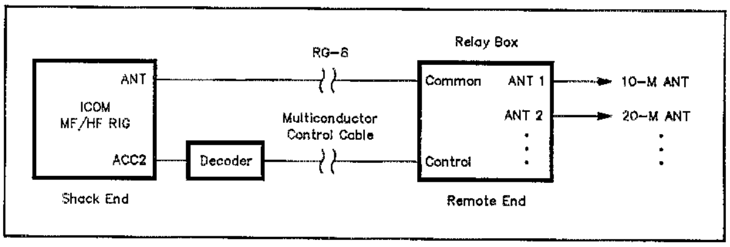

Fig 1—Block diagram of the remotely controlled antenna switch.

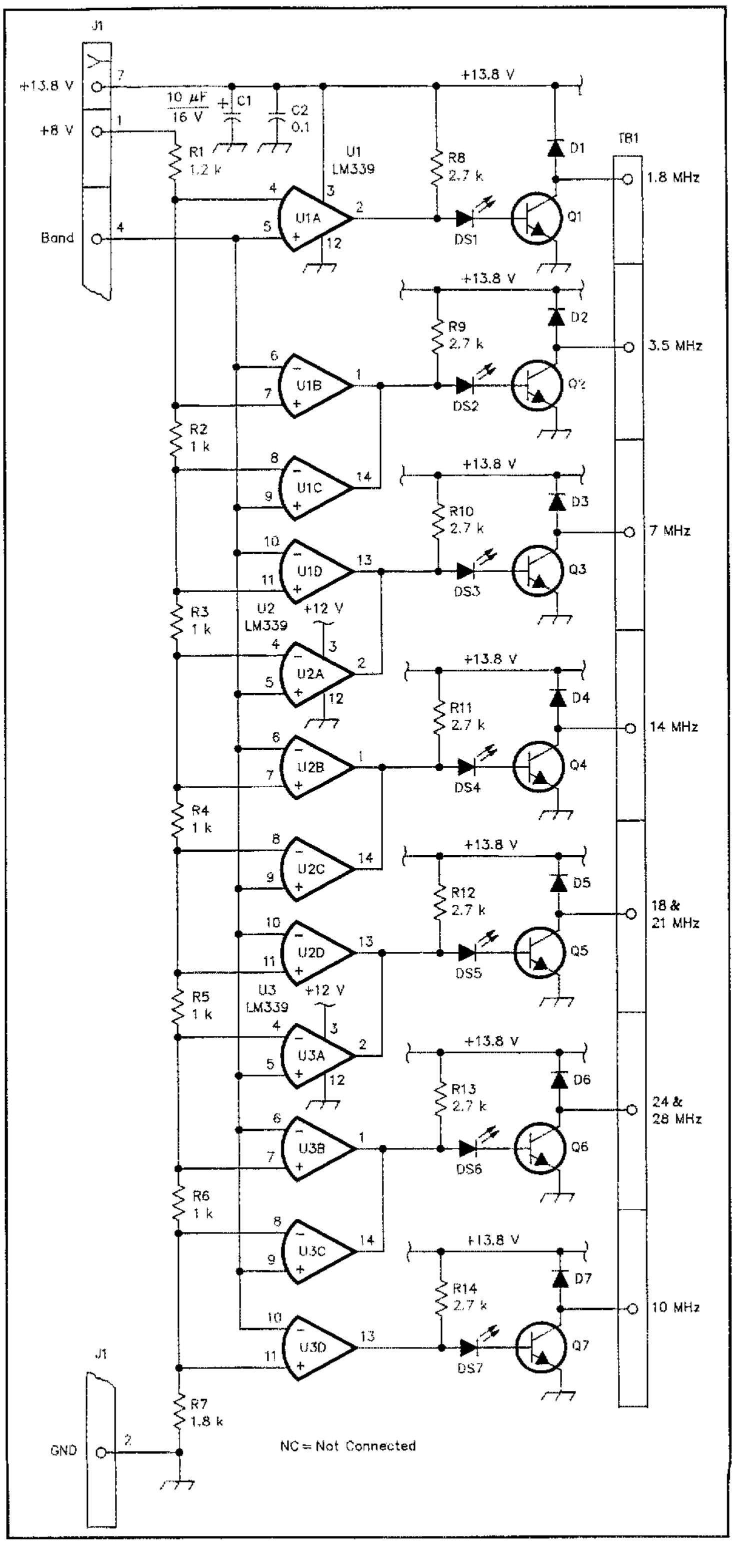

Fig 2—Schematic of the remotely controlled antenna switch control-box circuit. Part numbers in parentheses are Radio Shack; equivalent parts can be substituted. Unless otherwise specified, resistors are ¼-W, 5%-tolerance carbon-composition or film units.

D1-D7—1N4003 (276-1102).
DS1-DS7—LED.
J1—7-pin female DIN chassis-mount connector.
Q1-Q7—Almost any NPN transistor capable of sinking 150 mA (for the Radio Shack relays used in this project). I used ZTX657s. Other types include: MPSA05, MPSA06, 2N4401 (276-2058).
TB1—9-position screw-terminal strip.
U1-U3—LM339N (276-1712).
K1-K7—12-V, DPDT relay with 10-A/125-V ac contacts (275-218). (Presently, local Radio Shack stores stock the 275-218c [note suffix] relay, with a contact rating of 15 A.—*Ed.*)
Misc: decoder enclosure (approximately 2½ × 4 × 3 inches HWD), relay enclosure (approximately 3 × 6 × 8 inches HWD), 7 single-hole-mount SO-239 connectors.

deciding which band the rig is on.

Two bands (1.8 MHz and 10 MHz) need only a single comparator because they are at the ends of the voltage range. All the other bands use two comparators. One comparator of the pair detects if the band signal is *above* a specific level. The second comparator of the pair determines if the signal is *below* a certain level. If the signal lies between the two levels, the output-relay driver transistor is turned on.

Because the LM339 has open-collector outputs, we can conveniently connect them together to a pull-up resistor so that the output signal goes high only if *both* LM339 sections are off. This provides a simple **AND** function that uses few components.

I included LEDs to tell me that the unit is functioning correctly. The LEDs are in the base circuit of each relay-driver transistor. Such positioning ensures the LEDs *always* correctly indicate the band selected—even if more than one driver transistor is connected to a single antenna relay. Such a configuration would be used with a tribander: The outputs for 14, 21 and 28 MHz are simply wired together to a single relay. An inside view of the control box is shown in Fig 3.

The Relay Box

My original relay box design was quite simple. When a band was selected, the antenna was connected to the appropriate feed-line connector. I improved on this by having the antennas shorted to ground when not selected. This helps prevent static buildup on the antenna which otherwise might damage the rig's front end when the rig is first connected to the feed line.

The circuit of Fig 4 has an additional relay that is energized whenever the rig is turned on. Its job is to ground the feed line when the rig is turned off. I thought this might also help prevent static buildup on the open-

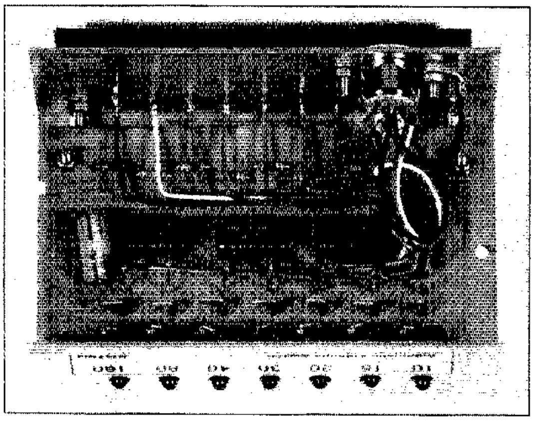

Fig 3—Inside view of the remote-control box. This early vision of the Remotely Controlled Antenna Switch doesn't incorporate the multiposition switch of Fig 7. (The Far Circuits PC board has provisions for the resistive ladder of Fig 7.)

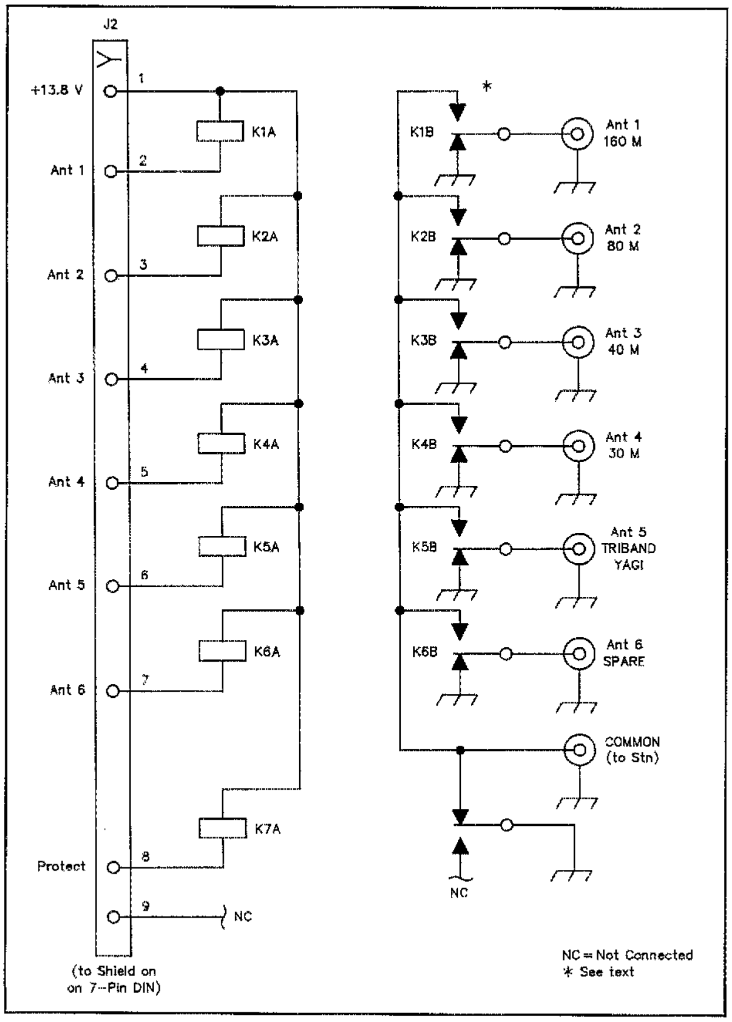

Fig 4—Relay box schematic. The normally open and normally closed contact pairs of the DPDT relays are individually connected in parallel to increase current-handling capability. J2 is a DB9M chassis-mount connector (276-1537).

circuited line when the rig is not in use.

The relays employed handle 100 watts comfortably. If you plan on building a high-power version of this switch, you'll need relays with much higher voltage and current ratings.

For my 100-W prototype, I used Radio Shack 12-V, DPDT relays. I wired both contact sets in parallel for increased current-handling capability. When the switch was connected at the end of 50 feet of RG-8 to a dummy antenna, I measured no SWR change with or without the relay box. The relay box layout I used is shown in Figs 5 and 6. The relays are mounted as close as possible to the connectors to keep the lead lengths short. The diecast aluminum box keeps the RF in and the weather out.

Alternative Designs

If you own a tribander, you might consider modifying the design to have only one output for those three bands. You can do this by changing the resistor ladder to have two fewer nodes. Note that the 14, 18, 21, 24 and 28-MHz signal levels are all in one block. Such a design requires only two LM339s.

If you'd like the ability to manually select a band or automatic operation, add a second resistive divider to deliver the signal levels that the transceiver would normally provide. See Fig 7. This resistive divider and switch are connected *between* the **ACC2** jack outputs and the existing resistive divider of the control box. Use a multiposition switch on the decoder's input to select either the radio's band-output signal (automatic mode), or one of the resistive-divider nodes.

By connecting the multiposition switch of Fig 7 to appropriate voltage sources, you can operate the remotely located relay box

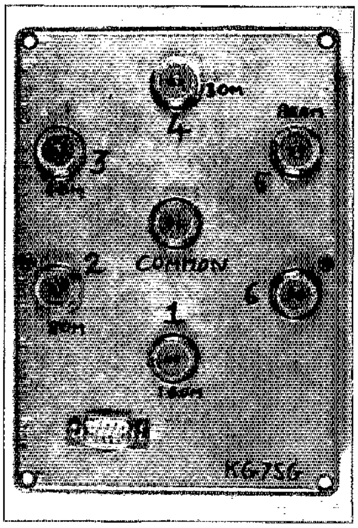

Fig 5—External view of the relay box. The diecast aluminum enclosure, with the liberal application of sealant, provides a weathertight assembly.

with rigs other than ICOMs.

Probably the ultimate version of this antenna switch would send the band signal, the dc power and the rig's RF up the feed line. I didn't bother trying this as separating a few volts of control signal and 100 W of RF seemed to me to be doomed to unreliability. If you want to try this "multiplexing," note that the band-signal output from the rig has a source impedance of around 100 kΩ, which means you'll need to buffer it before you can use it to drive a low-impedance load. The LM339 comparators have a very high input impedance, so buffering is unnecessary.

Construction

The Control Box

My prototype was developed on an experimenter's circuit board.[1,2] The first finished version of the decoder is built on perforated strip board and housed in a small metal cabinet. I use a 7-pin DIN connector for the connection to the transceiver and a 9-position screw-terminal strip for attachment of the control cable. With the screw-terminal strip, it's easy to connect multiple outputs to a single relay drive wire by simply linking them together.

Because the controller requires no power supply of its own, it fits into quite a small space providing a very unobtrusive addition to your station.

The Relay Box

The relay box uses SO-239 connectors for the main feed-line and antenna connections and a 9-pin D connector for the control cable. To protect the relay box from the elements, it's housed in a small wooden enclosure at the base of my roof-mounted tower. My relay box has inputs for six antennas, but any configuration can be built using the same basic design.

Summary

With this remotely controlled antenna switch installed, gone is the rat's nest of feed lines! (I now have 200 feet of coax sitting in the basement ready for another use.) The antenna installation is neater, and best of all, antenna selection is totally automatic!

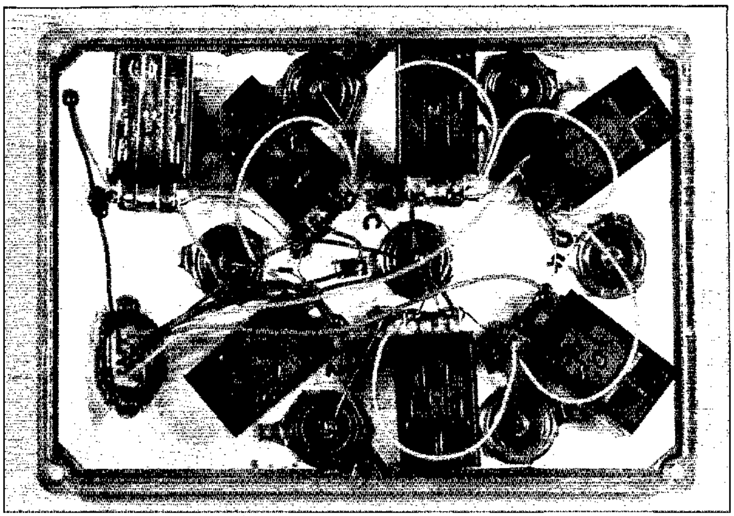

Fig 6—An inside view of the relay box. A piece of PC board fit to the box top serves as a means of easily attaching wires to a single ground plane. The relays are oriented to provide the shortest possible lead lengths to the coaxial jacks. Each relay case is cemented to the PC-board ground plane.

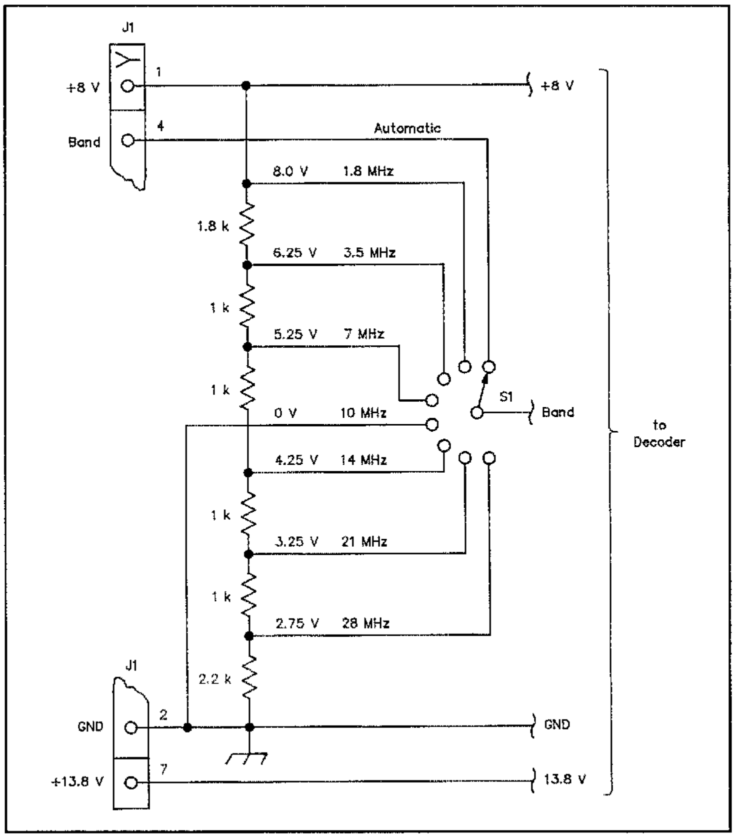

Fig 7—The addition of a multiposition switch provides a means of selecting automatic or manual control of relay switching with ICOM rigs. Rather neatly, this circuit also furnishes a remotely controlled antenna switch for non-ICOM rigs.

Notes

[1] I *love* these things and find them impossible to be without! I've even built a VHF transmitter on one—much to the horror of a friend!

[2] A PC board for the control box is available from FAR Circuits, 18N640 Field Ct, Dundee, IL 60118-9269; price $6, plus $1.50 shipping. The board provides for the addition of the resistive ladder of Fig 7. A multiposition switch is also available from FAR Circuits; price $3. No additional shipping charge is incurred if the switch is ordered with the PC board.

A PC-board template package is available from the ARRL for a business-size SASE. Please address your request for the THOMPSON REMOTE ANTENNA SWITCH TEMPLATE to the Technical Department Secretary, ARRL, 225 Main St, Newington, CT 06111.

A Two-Element Duoband Beam

Explore the 12- and 17-meter bands with this small, lightweight Yagi.

By John V. Mansheim, W0JCS
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Soon after the 12-meter band opened to US amateurs, I built a two-element beam using dimensions taken from Bill (W6SAI) Orr's *Beam Antenna Handbook*. The antenna worked much better than the dipole I had been using and I enjoyed many DX contacts.

The opening of the 17-meter band presented me with a dilemma. The 12-meter beam and tribander on my tower left no room for another beam antenna. I didn't want to give up my 12-meter beam, so the answer was a good duoband design.

I remembered my past success using traps made by Unadilla[1] on the 160, 80 and 40-meter bands. It occurred to me that I could use Unadilla's KW-12 coils on my 12-meter beam to get 17-meter coverage as well.

I took the 12-meter beam down and installed the KW-12 coils at the element ends. I added adjustable stubs for 17 meters to the outer ends of the coils. Then I temporarily mounted the antenna 5 feet above the ground—about as far from the ground as it would be mounted above the triband beam on my tower—and tuned it to resonance (more on this later).

Construction

Study Figs 1 through 4 as you read the construction details. Use electrician's conductive paste to prevent corrosion where the element sections overlap, and stainless-steel and brass (or nickel-plated brass) hardware in construction. A coating of clear plastic spray (such as Krylon) protects the finished antenna.

I bought the aluminum tubing at a hardware store. *QST* advertisers also stock it. The original 12-meter beam was built on a 6-foot length of 1¼ × 1¼-inch square stock with 1/16-inch wall thickness. I cut this piece from an 8-foot section and attached the leftover 2-foot piece at the bottom of the boom at the antenna's center of gravity for additional strength and greater boom-to-mast clamping area. A boom made of round tubing with a traditional boom-to-mast clamp would work fine too. Use what's most readily available to you and don't be afraid to substitute materials (within reason) and modify the design accordingly.

To support the elements, I mounted two 3-foot lengths of ⅛-inch-thick 1- × 1-inch aluminum angle stock to each end of the boom and installed ceramic standoff insulators for the element supports (see Fig 2). The inner sections of the driven element consist of two 3-foot lengths of ⅞-inch tubing mounted to the ceramic standoffs, with a 3-inch gap between the adjacent ends. Before mounting them, slot the outer ends of the tubing to permit the ¾-inch tubing to slide freely, and for the compression clamps to work properly.

The center section of the reflector is a 7-foot, 4-inch length of ⅞-inch tubing, centered on the boom. For additional support and good grounding, drill and bolt the driven-element support and reflector to

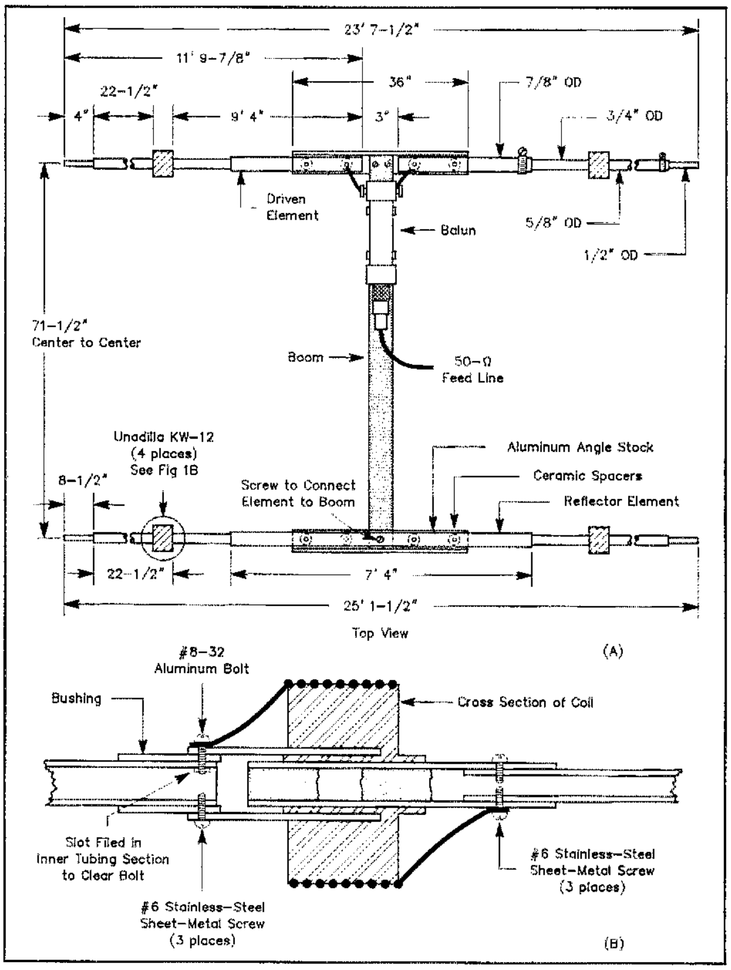

Fig 1—Physical layout of the 12- and 17-meter Yagi. Table 1 lists the necessary parts. Note that the clamps are shown only on half of the driven element for simplicity. Reflector tubing diameters are the same as those of the driven element. At B, a detail of the loading-coil attachment to the element tubing.

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[1] Notes appear on page 37.

the boom where they cross. Also slot the outer ends of the reflector tubing.

Mounting the Coils

Cut a section of ⅞-inch tubing into four 2-inch lengths, and cut a slot lengthwise through each piece with a hacksaw. These will form bushings at one end of each coil. On the larger ends of the coils, locate and temporarily remove the #8-32 aluminum bolt. Push each of the slotted 2-inch pieces of ⅞-inch tubing into the larger end of the coils. Using the coil-bolt holes as a guide, drill through the tubing and loosely replace the aluminum bolt.

File a ½- × ⅛-inch slot into each end of the four 8-foot, ¾-inch-OD tubing sections. Push the slotted end of each piece of the ¾-inch tubing into the slotted sleeves on the coils. Align the ½- × ⅛-inch slot so it passes under the aluminum bolt in the coil, then tighten the bolt. Drill three small holes through the existing holes on the coil tubing, through the aluminum sleeve, and into the ¾-inch element. Place three #6, ⅜-inch stainless-steel, pan-head sheet-metal screws in these holes and tighten them.

Mount the four 2-foot lengths of ⅝-inch tubing to the outsides of the coils as follows. First file a ½- × ⅛-inch slot into one end of each piece of tubing. Loosen the bolt on the small end of the coils and push the tubing inside it until the slot is aligned under the screw. Tighten the screw. Drill three small holes through each element to allow the stainless-steel pan-head screws to secure the element-to-coil connection. I used heat-shrink tubing over the coil ends to protect the hardware and connections (see Fig 3).

Final Assembly

Mount the coil assemblies to the ⅞-inch tubing. Slide the ¾-inch tubing into the ⅞-inch tubing. Use aircraft-grade stainless-steel hose clamps at these junctions.

I used a commercial 1:1 balun at the feed point, as shown in Fig 4. [A coaxial choke balun would work as well.—*Ed.*] Connect the balun terminals to the driven-element halves with short pieces of wire.

Table 1
Duoband Yagi Parts List

- 8-foot length of 1¼ × 1¼ × 1/16-inch square aluminum tubing (boom).
- 8-foot length of ⅞-inch-OD aluminum tubing.
- Two 3-foot lengths of ⅞-inch-OD aluminum tubing.
- Four 8-foot lengths of ¾-inch-OD aluminum tubing.
- Four 2-foot lengths of ⅝-inch-OD aluminum tubing.
- Four 1-foot lengths of ½-inch OD aluminum tubing.
- Two 3-foot lengths of 1 × 1 × ⅛-inch aluminum angle stock.
- Six 1¼ × 1-inch ceramic standoff insulators (preferably threaded).
- Eight stainless-steel hose clamps.
- Small container of electrician's conductive paste.
- 1:1 balun.
- Four Unadilla KW-12 antenna coils.
- Two boom-to-mast **U** clamps.
- Four ⅝-inch caps for element ends.
- Twelve #6 × ⅜-inch pan-head, stainless-steel sheet-metal screws.
- Twelve #6 × ¼-inch pan-head, stainless-steel sheet-metal screws.
- About three feet of heat-shrinkable tubing.
- Assortment of stainless-steel and brass bolts, nuts and washers.

Alignment

Adjustment for 12 meters comes first. Leave off the 17-meter stubs and the balun during this step. Check the resonant frequency of the driven element using a dip meter coupled to it with a one-turn loop of wire across the feed point.[2] I set my antenna to 24.960 MHz by sliding the ¾-inch tubing to the appropriate positions in the ⅞-inch elements. Install the coils and 17-meter, ½-inch-diameter stubs and adjust the driven element to 18.120 MHz by sliding the stubs into or out of the ⅝-inch tubing.

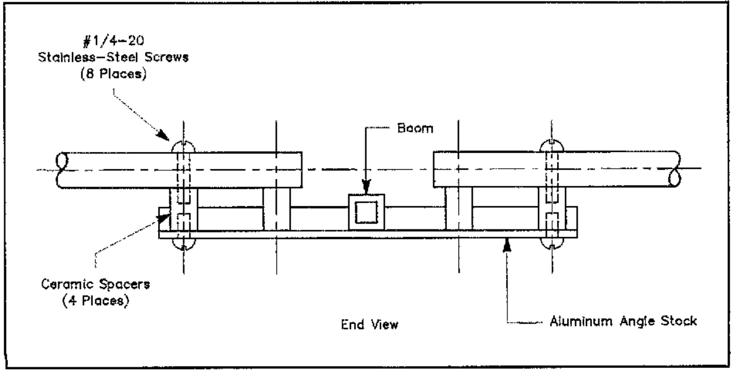

Fig 2—Element-to-boom mounting detail showing how ceramic insulators support the driven element on the angle stock. See Fig 4.

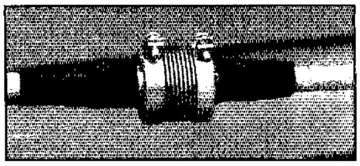

Fig 3—After assembly, the loading coils are exposed, but their connections are protected from the weather by heat-shrinkable tubing. This antenna has served for more than four years with consistently good performance and minimal maintenance.

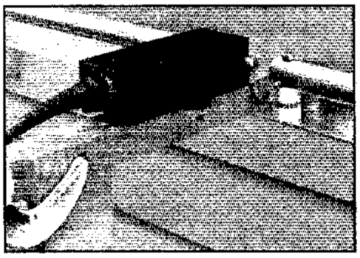

Fig 4—Feed-point detail showing one of the ceramic insulator, angle-stock support and inner driven-element ends.

Set the 12-meter portion of the reflector element (between the inside edges of the coils) to 19 feet, 8½ inches (see Fig 1). The Unadilla coils are equipped with adjustable straps to allow resonant-frequency adjustment. Adjust these straps until your dip meter shows resonance at 18.120 MHz for each coil. The reflector is 18 inches longer than the driven element, but its coils are resonant at the same frequency as the driven element.

On-the-Air Results

The beam typically exhibits a 5 to 6 dB signal-strength advantage over my dipole antennas. The front-to-back ratio is more than 10 dB and front-to-side attenuation is about 25 dB. The antenna doesn't seem to interact with my tribander, as its SWR and patterns haven't changed since I put up the new Yagi.

For only around $100 in parts and a few hours of time, you can make an antenna that covers two great HF ham bands that conventional tribanders don't. It's also small and light enough to put on the same mast. Try building one yourself—you'll be pleasantly surprised at the variety of propagation these bands offer and how many people use them!

Notes

[1] Unadilla Antenna Manufacturing Company, PO Box 4215, Andover, MA 01810.

[2] When you use a dip meter for antenna work, factors like the instrument's age and calibration and the degree of coupling can affect your readings. It's best to verify the dip-meter frequency with a counter or receiver.

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QST

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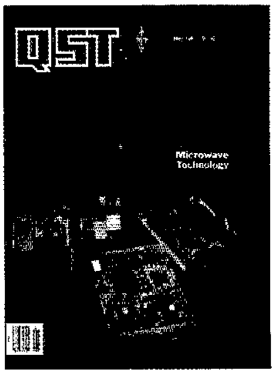

OUR COVER

Fascinating, exotic and fun:
Microwave operating may sound esoteric, but it's not as formidable as it seems. See how easy it is to home-brew a 10-GHz transverter in the article on page 21 by ARRL Lab Engineer Zack Lau, KH6CP, and build yourself a microwave transceiver with the help of Rick Campbell, KK7B, on page 29.

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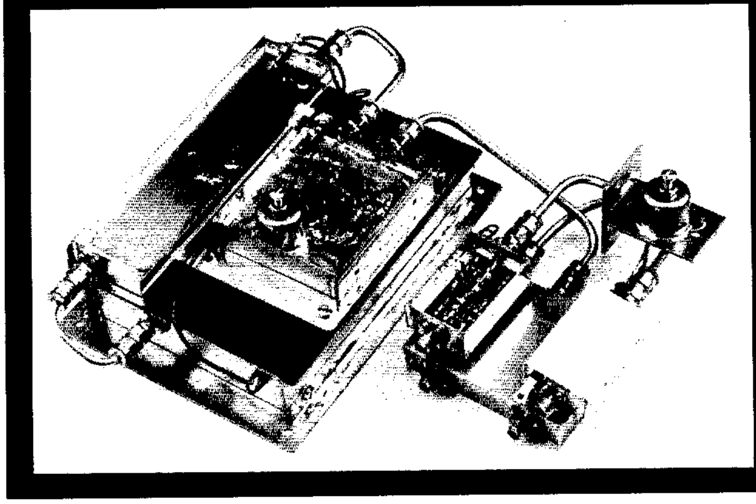

Part 1— Narrowband 10-GHz operation—without exotic or surplus parts—has finally arrived for the microwave builder!

By Zack Lau, KH6CP
ARRL Laboratory Engineer

Looking for some challenging microwave equipment to build? How about a complete 10-GHz transverter with stability good enough for weak-signal CW work? You don't have to find any exotic pieces to build this project—all of the parts are fairly common. In fact, everything has been available for years.

Despite the transverter's compact package, however, it consists of several modules that you must build. And although the VHF/UHF circuitry follows the no-tune concept developed by Jim Davey, WA8NLC; Rick Campbell, KK7B; and others, the X-band (10-GHz) parts need to be tuned up—preferably with a spectrum analyzer that works through 10.4 GHz.[1] To give you an idea what it takes to build this transverter, see the sidebar, "Should I Attempt This?"

Design Philosophy

Unlike the no-tune transverters,[2] I decided to develop the transverter as a set of building blocks with stainless-steel or gold-plated SMA connectors. Although this construction method is more expensive and time-consuming than a more integrated approach, it offers several advantages. Most importantly, it allows you to check small portions of the transverter for proper performance. If something doesn't work, troubleshooting is fairly straightforward. And, if you just can't get one of the modules to work, you can simply build another one. Another advantage of this construction method is the shielding that results from packaging circuits in separate boxes. This helps greatly to keep the transverter spectrally clean, with a minimum of spurious outputs and responses. Finally, the transverter is easily updated or expanded to take advantage of improving technology. Making its receiver section state-of-the-art is simply a matter of adding the 1-dB noise figure preamplifier described in December 1992 *QEX*.[3]

A Brief Overview

Fig 1 and Table 1 show the transverter's configuration and measured performance. A local oscillator (LO) feeds a power splitter that drives a pair of mixers. One mixer is used on transmit and the other on receive. The transmit mixer is followed by a filter and amplifiers. A filter following the final stage is optional. Low-noise amplifiers and an image-stripping filter precede the receive mixer. Without adequate image rejection, the receiver sensitivity can degrade by as much as 3 dB.

Table 1
10-GHz Transverter Performance*

Transmit Converter

144-MHz Drive (dBm)	*10-GHz Output† (dBm)*
–10.0	3.8
–3.0	8.5
0.0	10.3
1.0	10.8
3.0	11.6
5.0	12.2
10.0	12.8

Power Output versus Supply Voltage†
(Drive signal: –0.8 dBm at 144.06 MHz)

Supply (V)	*Output Power (dBm)*
10.34	8.5
10.51	10.0
10.75	10.0
12.34	10.0
14.02	10.0

Receive Converter

IF (MHz)	*Gain (dB)*	*Noise Figure (dB)*
144	8.82	2.73
146	8.79	2.70
148	8.79	2.71

Noise Figure and Insertion Gain versus Supply Voltage
(IF = 144 MHz)

Supply (V)	*Gain (dB)*	*Noise Figure (dB)*
10.3	6.59	3.03
10.7	8.81	2.92
10.9	8.64	2.89
11.3	8.59	2.79
12.4	8.68	2.75
13.4	8.70	2.74
13.5	8.73	2.76
14.9	8.75	2.74

*The data in this table comes from the most recently completed prototype, which consists of the modules described in Part 1 and Part 2 of this article.
†Power output was measured with an uncalibrated HP 435B/8481A.

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[1]Notes appear on page 28.

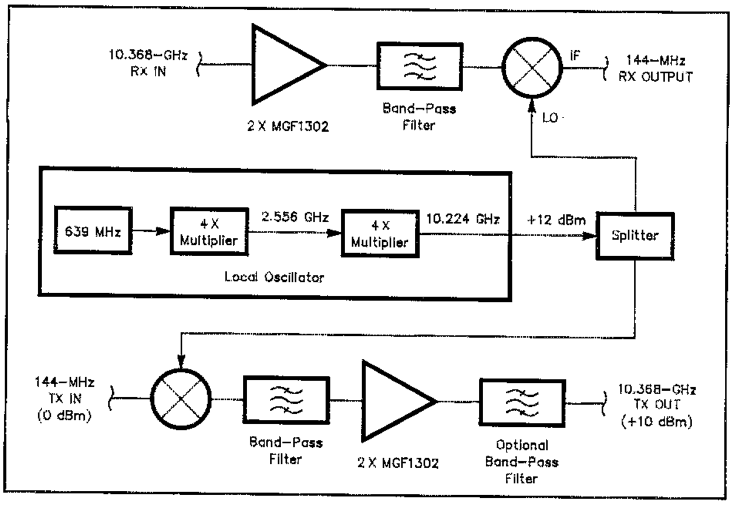


Fig 1—Block diagram of the 10-GHz transverter.

The Local Oscillator

The most critical part of microwave narrowband work, the LO, starts off with the circuitry developed by WA8NLC and KK7B. The 106.5-MHz oscillator (Fig 2A) is multiplied by six (Fig 2B) to produce a 10-dBm (10-mW) signal at 639 MHz. This signal is then multiplied by four and amplified to 7 dBm at 2.556 GHz (Fig 3).

This is essentially the same scheme used in KK7B's 2.16-GHz LO in July 1989 *QST*,[4] except that I modified the filters for 639 and 2556 MHz. I also added a 0.47-µF capacitor to provide a low-impedance input for the 78L05 regulator (it can oscillate if not properly bypassed). These circuits are built on fiberglass-epoxy G10 or FR4 PC-board material; the remaining circuits are built on 5880 RT/duroid.

Choosing the Circuit-Board Material

This part was actually pretty easy: I looked around for something with low enough loss to work well, but that's also readily available to amateurs. The only stock item that meets this description is 0.015-inch-thick (15-mil) 5880 RT/duroid sold by Microwave Components of Michigan (see the sidebar, "Where to Get the Pieces"). I believe that Microwave Components of Michigan began stocking this board material in response to demand generated by the Tuesday Night Transverter published in the *Proceedings of Microwave Update '88*.[5] The thicker 30-mil 5880 RT/duroid is definitely unacceptable, as its radiation loss is rather high.

If availability wasn't an issue, I might have chosen a board thickness that helps to optimize stability via source inductance.[6] Another criteria for choosing board thickness is the interface with the transistors and connectors. Often, it is desirable to minimize the discontinuity between these interfaces by selecting trace widths comparable to the connector diameters and transistor-lead widths. The 15-mil board works pretty well in this area—the 46-mil trace widths fairly closely match the widths of the specified 50-mil chip capacitors.

Crystal Frequency

When choosing an LO crystal, the most important consideration is the crystal's calibration. The tolerance of the International Crystal Manufacturing high-accuracy crystal (#473590) I recommend is 10 parts per million. This means that the crystal can be as much as 1.06 kHz off the marked frequency without deviating from the specified accuracy. Because the LO is multiplied by 96, the transverter's conversion frequency could be as far as *102 kHz* from the expected frequency, even without taking temperature variations into account. Although the oscillator circuit allows some adjustment to compensate for frequency error, attempting to shift the frequency seems to degrade stability.

To make sure that the conversion frequency falls inside the 2-meter band, I specify a 106.499-MHz crystal. Selecting a 106.500-MHz crystal might prove to be unwise if it was cut 10 ppm high—the usual calling frequency of 10.368100 GHz would be just below the 2-meter IF radio's 144.0-MHz band edge—a problem with some radios. You may want to choose another frequency, perhaps even lower, to move the IF to 145 or 146 MHz. If you do this, you'd be wise to investigate possible sources of interference. Keep in mind that hilltops are often pretty bad in terms of interference problems.

The stage following the 639-MHz to 2.556-GHz multiplier is a GaAsFET multiplier, filter and amplifier (Figs 4 and 5) that takes the 2.556-GHz input and provides at least 12 dBm at 10.224 GHz to the LO splitter/mixer board (to be described in Part 2).

Bias Supplies

I know it's not the cheapest way to go, but I decided to build a negative bias supply into each module that requires one (all the stages that use MGF1302s). This reduces the chance

Should I Attempt This?

As someone who's been interested in 10-GHz weak-signal work since I first experienced it last June, I found myself asking this question while editing this article. Although I've built several no-tune transverters and amplifiers and worked with a bit of surplus microwave equipment, I wasn't sure that I was up to this.

I was pleasantly surprised. Though somewhat labor-intensive, this project doesn't pose the challenge that building earlier narrowband 10-GHz equipment has. If you're not familiar with the mechanical construction techniques used in building this transverter, you'll find Chapter 7 of the RSGB's *Microwave Handbook, Volume 2: Construction and Testing*,* and Chapter 8 of *The ARRL UHF/Microwave Experimenter's Manual*† to be very useful.

The most complicated part of this project is *not* making the boards, getting the parts or assembling the pieces, but tuning the mixer, filters and amplifiers once you've built them. (Only the filters *require* tuning to work acceptably, however.) Most of the building blocks are relatively simple and can be functionally tested using a multimeter to check bias voltages, an FM broadcast receiver to make sure that the local oscillator works, and a 10-GHz relative power meter or detector.†† Optimizing the circuits is easiest with the aid of a spectrum analyzer, calibrated power meter and other precision instruments, however.

If you've successfully built any of the no-tune transverters but don't have access to the test equipment necessary for this project (or lack experience using it), you'll find this to be the biggest challenge. If you haven't built a lower-frequency transverter, it's a good idea to get that experience under your belt before going at this one.

Zack has sweated the details—now we can all reap the benefits of an easily duplicated, high-performance transverter suitable for long-distance communications.—*Rus Healy, NJ2L, Senior Assistant Technical Editor*

*This book, as well as Volumes 1 and 3, are available from the ARRL Publication Sales Department, tel 203-666-1541.

†Also available from the ARRL Publication Sales Department.

††To be discussed in Part 2.

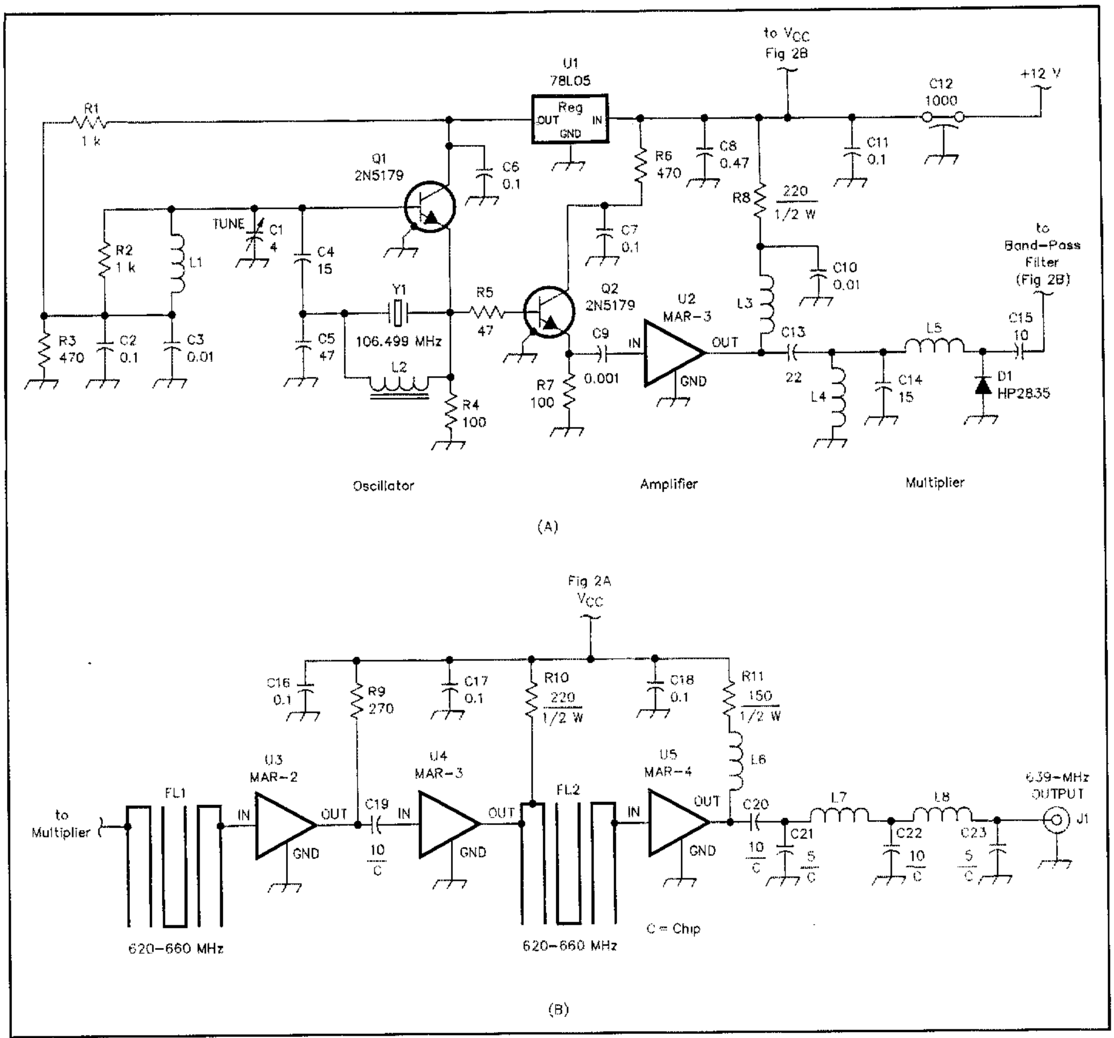


Fig 2—Schematic of the crystal oscillator and times-6 multiplier. Resistors are ¼-W carbon-film or carbon-composition types unless otherwise indicated.

C1—Air-dielectric trimmer capacitor that can be set to approximately 4 pF. Low temperature coefficient is more important than exact value, as L1 can be adjusted to compensate.
C8—Minimum value required to stabilize U1 is 0.33 µF. An electrolytic capacitor can be substituted if proper polarity is observed.
C12—1000-pF feedthrough capacitor. Exact value not critical (100 pF to 0.1 µF should work well).
D1—Schottky diode. Hewlett-Packard 5082-2835, -2811 and -2800 work well.
FL1, FL2—Band-pass filters printed on PC board.
J1—SMA female chassis-mount connector.
L1, L3, L4, L6—8 turns #28 enameled wire, 0.1-inch ID, closewound.
L2—12 turns #30 enameled wire on T-30-6 toroid core.
L5—5 turns #28 enameled wire, 0.1-inch ID, closewound.
L7, L8—2 turns #28 enameled wire, 0.062 inch ID, turns spaced one wire diameter.
Q1, Q2—2N5179 or BFR91.
U1—78L05 5-V, 100-mA, three-terminal regulator.
U2, U4—MAR-3 or MSA-0385 MMIC.
U3—MAR-2 or MSA-0285 MMIC.
U5—MAR-4 or MSA-0485 MMIC.
Y1—106.499 MHz, fifth-overtone, series-resonant crystal (International Crystal Manufacturing #473590).

of misconnecting the positive and negative supplies. I also opted for active bias supplies, as shown in Fig 5. This figure shows the two equations for calculating components for different bias conditions. For instance, to bias an FET at 3 volts and 30 mA, you first calculate the effect of any resistors used for stability. Often, a 51-Ω resistor is used to stabilize the circuit; if present, it increases the circuit bias voltage to 4.53 volts. One set of standard values that comes close to the bias conditions given above, and accounts for the 51-Ω resistors, is: R_{dn} = 16 Ω, R_{an} = 3.6 kΩ, and R_{bn} = 1.1 kΩ.

I used Intersil ICL7660s to generate the negative bias supplies because they require few external parts. A cheaper alternative is to use NE555 timer chips as oscillators driving rectifiers.[7] I published such a circuit, with a PC-board pattern, in March 1991 *QEX*.[8]

Filter Construction

The transverter's band-pass filters are made from half-inch copper pipe caps, as shown in Figs 6 and 7. These were developed by Roman Wesolowski, DJ6EP; and Kent Britain, WA5VJB.[9] They're affordable, too:

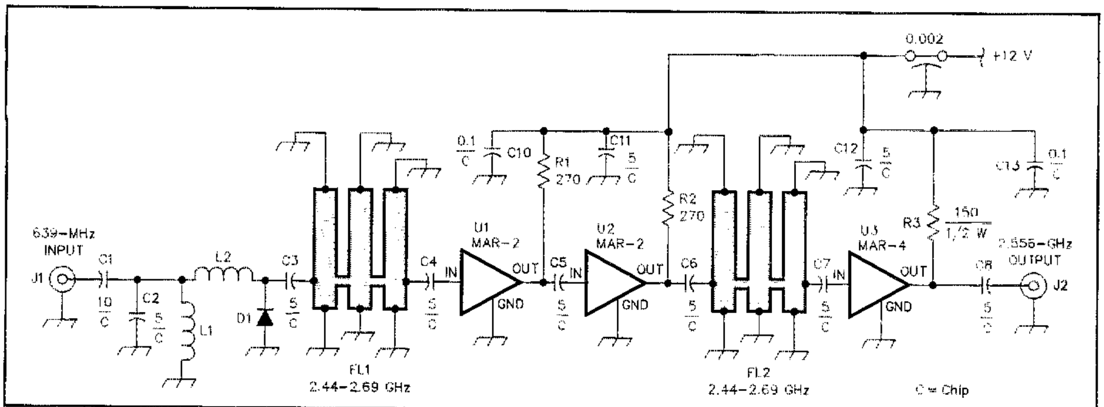


Fig 3—Schematic of the 639-MHz to 2.556-GHz multiplier. Resistors are ¼-watt carbon-film or carbon-composition types unless otherwise indicated.

D1—Schottky diode. Hewlett-Packard 5082-2835 or equivalent.
J1, J2—SMA female chassis-mount connector.
FL1, FL2—Printed band-pass filters.
L1—3 turns #28 enameled wire, 0.062 inch ID, turns spaced one wire diameter.
L2—Printed inductor.
U1, U2—MAR-2 or MSA-0285 MMIC.
U3—MAR-4 or MSA-0485 MMIC.

You can buy half-inch plumbing caps at home-supply stores for as little as 12 cents each. (Designed to cap pipes that are 0.5 inch ID, these caps actually measure 0.62 inch ID and about 5/16 inch long.) I drill and tap the caps (at top center) with #4-40 threads and use nickel-plated brass screws; unplated brass screws should work as well. Kent Britain has forced steel screws through the caps to thread them. Don't use these screws for tuning, though, as steel is unacceptably lossy. I often polish my plumbing caps so that they look nice and solder easily.

A pipe-cap filter ahead of the mixer is adequate in terms of system noise figure, giving an image rejection around 24 dB with a 144-MHz IF. For critical applications, a waveguide filter, such as the one published by Glenn Elmore, N6GN, in July 1987 *QEX*,[10] is recommended. With such a filter, 50 dB of image rejection is easily obtained with a 144-MHz IF. However, for lightweight portable transceivers, plumbing-cap filters seem to be the best compromise. For a clean transmitted signal, you should use one at the final transmit amplifier's output as well.

The filters are built on unetched, double-sided, 1/16-inch G10 or FR4 PC-board material. I recommend that you use 0.141-inch semirigid coaxial cable (UT-141) to make the probes. A probe length of about 75 mils is optimum. If you cut them too short—say, 50 mils—the insertion loss climbs from an acceptable 1 to 2 dB to as much as 5 or 8 dB. If the probes are cut too long—say, 100 mils—the image rejection drops to a measly 10 to 14 dB, though the insertion loss also drops (to 0.5 dB). The probes are spaced 5/16 inch center to center and the pipe cap is soldered to the ground plane so that the probes are centered within it.

How do you determine the best probe lengths and spacing for pipe cap filters? I developed the filters in this transverter using a spectrum analyzer and trial and error. The signal source was an X-band mixer and the 10.224-GHz local oscillator. I could have done a lot better with a network analyzer or a scalar sweep setup, but I used what was available to me.

Filter construction can be fairly critical for optimum performance. In particular, the probes must be accurately cut to length. I estimate my error margin in measuring and cutting probe lengths to be about 10 mils. The ends of the probes are filed flat, not chamfered or rounded. Filter loss seems to be a few tenths of a decibel lower with the dielectric left on the probes, although it is easier to trim the probes exactly with the dielectric removed.

You may be tempted to use 0.085-inch semirigid cable because it's easier to handle than UT-141. A similar filter I made using this material gives 24 dB of image rejection, but has 3.4 dB of loss. The probe length for this cable is 70 mils. A filter using 100-mil probes of 0.085-inch semirigid cable has only 2 dB of loss, but the image rejection drops to a barely acceptable 17 dB. UT-141 is better for this application.

I recommend that you assemble the cable and solder it to the ground plane *before* measuring and cutting the probe length. Otherwise, the length may change as you work on the cable. With these filters, a potential problem is caused by the center conductor moving around slightly, particularly when the cable is straight and the center conductor forms the center contact at the

Where to Get the Pieces

Here's a partial list of vendors that supply the parts used in the 10-GHz transverter. Catalogs and/or price lists are available from each listed supplier.

Down East Microwave, RR1 Box 2310, Troy, ME 04987, tel 207-948-3741, fax 207-948-5157, and Steve Kostro, N2CEI, RFD 1 Box 341A, Frenchtown, NJ 08825, tel 908-834-1304 (days) or 908-996-3584: MMICs, MGF1302, chip capacitors, SMA connectors, mixer diodes (only DEM carries the HSMS 8202; both stock the HSMS 2822).

Mainline Electronics, PO Box 235, Leicester LE2 9SH, England: MGF1302, MMICs, transistors, voltage regulators.

Microwave Components of Michigan, PO Box 1697, Taylor, MI 48180, evening tel 313-753-4581: MMICs, 15-mil 5880 board material, MGF1302, HSMS 2822, chip capacitors, UT-141, SMA connectors, feedthrough capacitors.

Nemal Electronics, 12240 NE 14 Ave, N Miami, FL 33161, tel 305-893-3924: SMA connectors, semirigid cable.

Ocean State Electronics, PO Box 1458, Westerly, RI 02891, tel 401-596-3080, fax 401-596-3590: LM555, 78L05, 2N3906, 2N2907A, ICL7660, ¼-watt carbon-film resistors, electrolytic capacitors, tantalum capacitors.

SHF Microwave Parts Co, 7102 W 500 St, La Porte, IN 46350: HSMS 2822, MMICs. Also sells Gunn transceivers.

Small Parts, PO Box 4560, Miami Lakes, FL 33014, tel 305-557-8222: brass sheet stock, stainless-steel screws, #2-56 taps, many other items of interest to microwave-equipment builders.

RF Parts, 1320 Grand Ave, San Marcos, CA 92069, tel 619-744-0700, fax 619-744-1943: MGF1302.

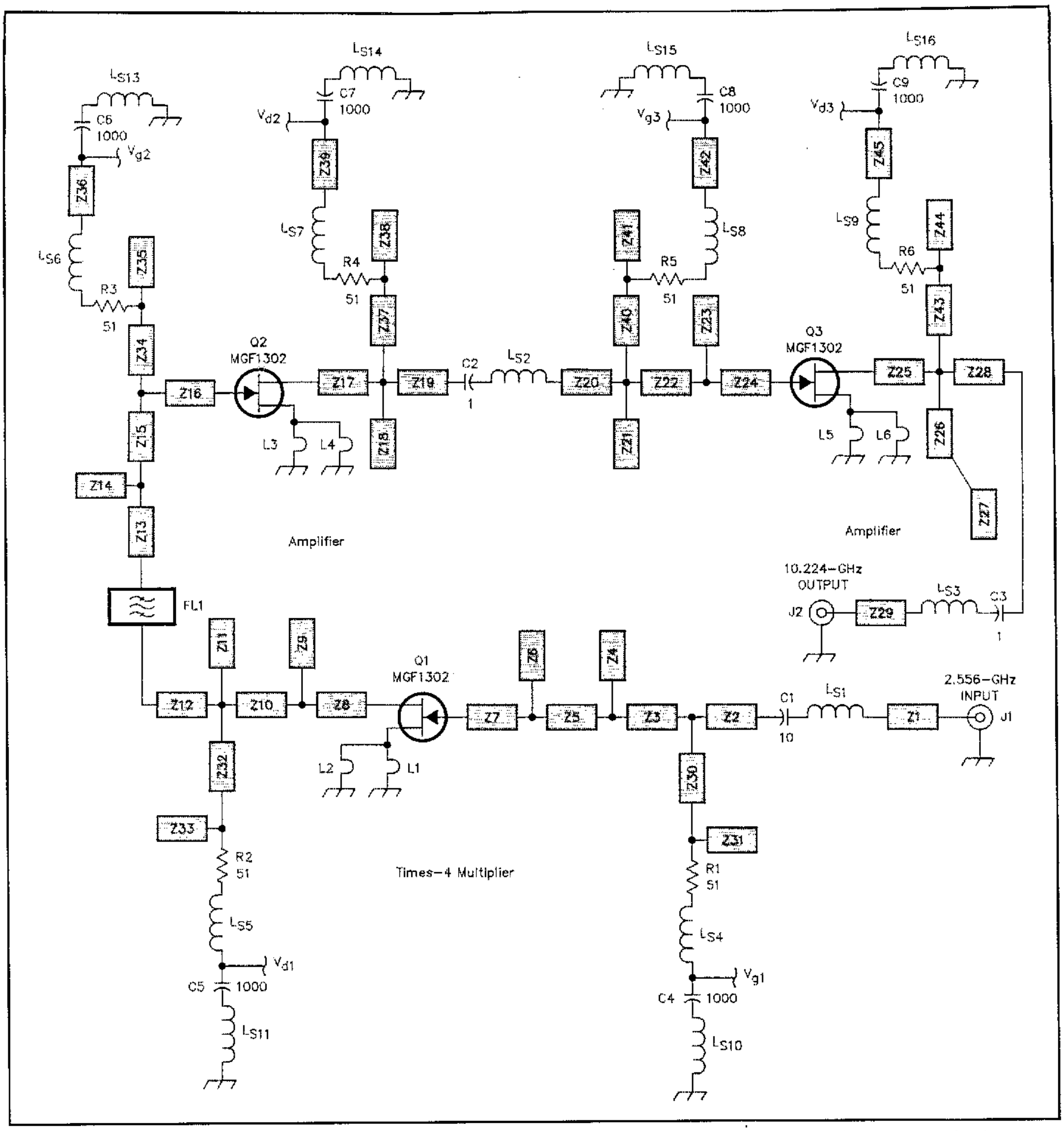


Fig 4—Schematic of the 2.556- to 10.224-GHz multiplier. Resistors and capacitors are chip components. L1-L6 are source-lead inductances. L_{S1}-L_{S16} are stray inductances. Z1-Z45 are etched on the circuit board.

FL1—Pipe-cap filter. See Fig 6. Countersink the ground-plane side of the circuit board hole (by hand) to keep the ⅛-inch UT-141 probe center conductors from shorting to it.

Q1-Q3—Mitsubishi MGF1302 GaAsFET. Substitution not recommended.

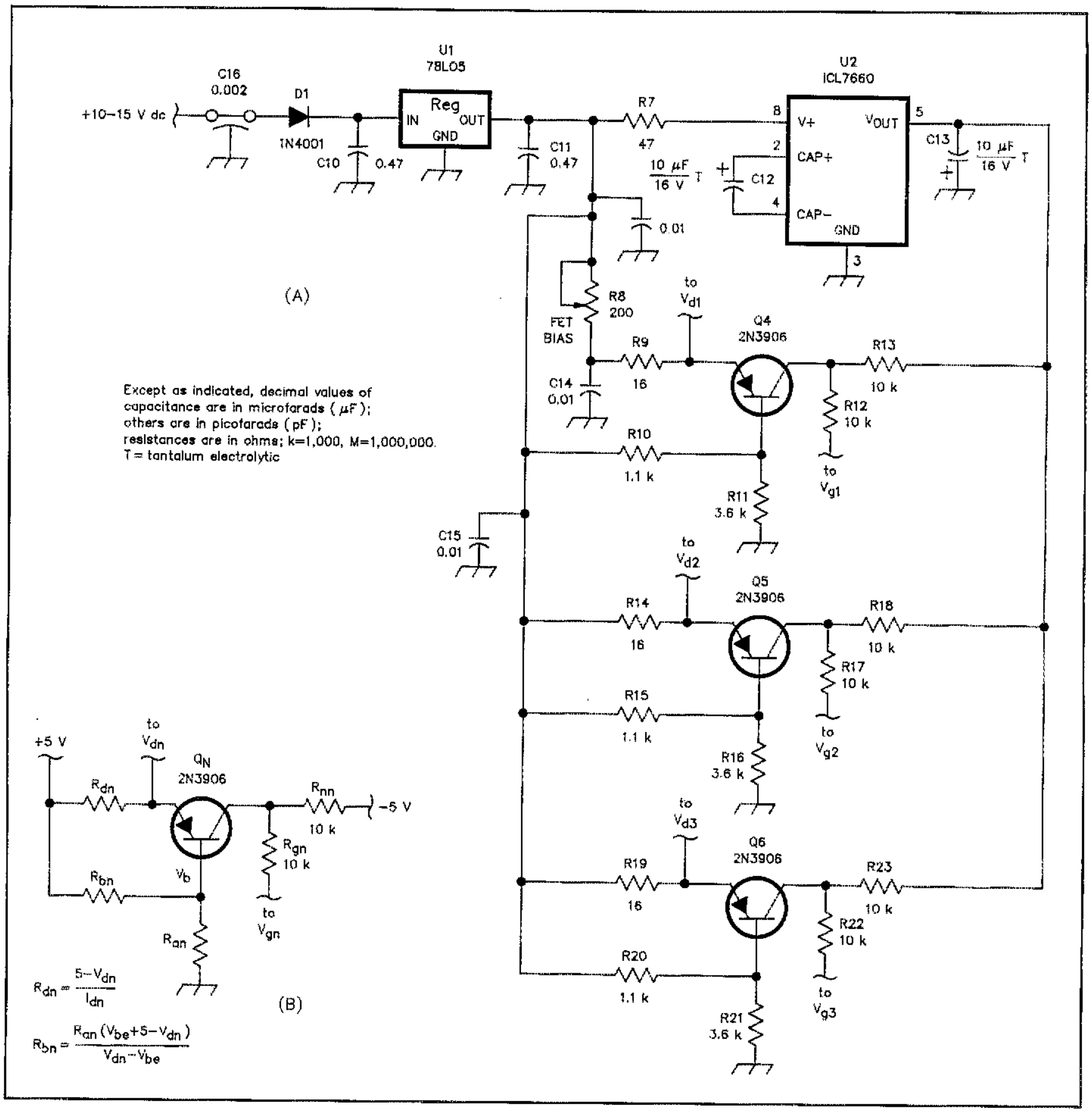
connector end. Bending the cable helps to prevent this problem, but the best solution is to use connectors that captivate the center conductor, keeping it from being pushed inward.

Enclosures

As shown in the title photo, I use 0.025-inch-thick, half-inch-wide brass sheet stock to make the enclosure walls. Instead of soldering SMA connectors to the walls, I attach them with #2-56 screws; either method is acceptable. The 25-mil brass stock is ideal for tapping small screw holes. Other commonly available thicknesses can also be used, although 20-mil stock is a bit flimsy and 32-mil stock is more difficult to solder.

Duplicating the Circuit Boards

Using PC-board layout software, I've developed artwork for each of the transverter's circuit boards. To make it as easy as possible for *QST* readers to build this transverter, ARRL HQ is making the circuit-board artwork available in three forms: as PostScript files downloadable from the ARRL HQ telephone BBS; as negative film for those with access to photographic methods of circuit-board production; and as laser-printed positive images that can be transferred directly to



Except as indicated, decimal values of capacitance are in microfarads (µF); others are in picofarads (pF); resistances are in ohms; k=1,000, M=1,000,000. T = tantalum electrolytic

Fig 5—At A, schematic of the multiplier bias supply. At B, generalized FET bias circuit and equations.

C12, C13—Tantalum electrolytics preferred, but aluminum electrolytics should work.
C16—Feedthrough capacitor. Value not critical.
D1—Reverse-polarity protection diode.
Q4-Q6—General-purpose PNP transistor; 2N3906 and MPS2907 suitable. Plastic-cased devices are easiest to use.
R8—200 Ω used in the prototypes. 100 or even 50 Ω may be suitable.
U1—78L05 5-V regulator.
U2—ICL7660 negative-voltage generator.

the PC-board material.[11]

Several methods are available for transferring toner from the laser engine to the circuit board. Plain paper is my favorite.[12] Start with a clean circuit board (roughed with 400-grit sandpaper) and a laser-printed reversed positive image of the board on plain paper. Then use an ordinary household iron at its linen setting to iron the image onto the board. Buffer the iron from the sheet of paper with the pattern on it with a second, clean sheet of paper. Run the iron over the board in a pattern that uniformly heats the material for 30 seconds or so for the 15-mil Teflon boards and at least a minute for the G10/FR4 boards. The iron's heat liquefies the plasticized toner and fuses it to the circuit board.

After ironing, place the board and paper (now fused to the board) into plain water for a few minutes, then remove it from the water and carefully rub away as much of the paper as you can. If the transfer process leaves incomplete traces, clean the board again with sandpaper and start over with a new copy of the artwork. You can correct minor imperfections with an etch-resist pen and carefully cut pieces of Scotch tape. Cover the bottom (ground-plane) side of each board with Scotch tape, then etch the boards. Peel off the tape and remove the toner with plain steel wool.

Oscillator Construction

I didn't develop circuit-board artwork for the 106.5-MHz local oscillator. If you want stability adequate for a 10-GHz SSB/CW system, a quartz-crystal-controlled system is *marginal*—you really can't throw away any stability to make construction easier. Remember: *The LO is multiplied by 96 before being mixed with the 144-MHz IF signal!* You could use a double-sided circuit-board layout, except that stability is ten times worse than that of a ground-plane version. So, I opted for the ground-plane version (Fig 8A). I also used a high-stability, air-dielectric trimmer at C1, as some ceramic trimmers have a high temperature coefficient. The trimmer value isn't critical, as L1 can be adjusted to compensate.

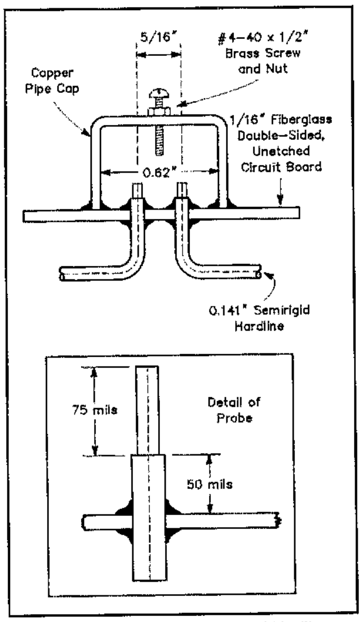


Fig 6—Simple and effective 10-GHz filters can be made using copper pipe caps and probes made from 0.141-inch semirigid coax. The total probe length above the ground plane is 125 mils (⅛ inch). Solder the pipe cap to the ground plane so that the probes are centered within it. A #4-40 × ½-inch brass (or nickel-plated brass) screw with a lock nut tunes the filter.

I recommend that you build and align the oscillator as follows. Build the oscillator with a 47-Ω, ¼-watt resistor in place of the inductor/crystal combination (L2 and Y1). When you power up the circuit, tune C1 so that the oscillator operates at 106.5 MHz. After replacing the 47-Ω resistor with the crystal and its resonating inductor, verify that the oscillator starts reliably as power is applied. A minor adjustment of C1 may be necessary for reliable starting. I don't recommend trying to adjust C1 for a given oscillation frequency.

The 639-MHz to 2.556-GHz multiplier (Fig 9) has no tuning adjustments. You simply verify that its power output is between 5 and 10 dBm.

Amplifier Design

I chose to use MGF1302 GaAsFETs for all the 10-GHz circuits. These seem to be the most readily available, low-cost parts that work well at this frequency. The transverter uses seven of them, and they cost less than $7 each from several sources—see the sidebar, "Where to Get the Pieces." Ideally, a transverter like this would use 10-GHz MMICs for gain blocks, but these weren't available during project development. Not only were the available packaged GaAs MMICs too pricey (around $40 each), but they weren't designed to work at 10 GHz. The second choice was the Avantek ATF13735, but commercial purchasers have made the standard part the short-leaded ATF13736, which is more difficult to use than the long-leaded version. I prefer to use devices with long leads since they're easier to install in circuits that use lead inductance as a circuit component.

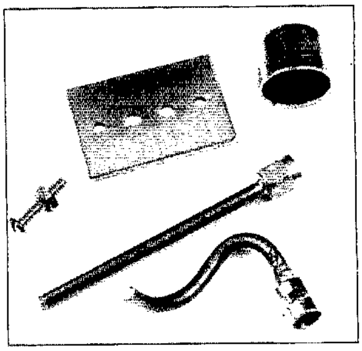


Fig 7—The pieces that make up a 10-GHz band-pass filter, before assembly. *(photos by Kirk Kleinschmidt, NT0Z)*

Multiplier and Amplifier Construction

In each of the transverter's building blocks, I build the RF circuitry on one side of the ground plane and the biasing circuitry on the other. After etching the boards, I drill and countersink holes for the power leads. Also countersink the ground-plane foil around the multiplier board's filter-cable holes so that the UT-141 center conductor doesn't short to the ground plane. Countersink the holes *by hand* with a relatively large drill (3/16 to ¼ inch). It's important to do this manually—you'll be surprised how easy it is to drill through such thin, soft material! Cut slots for the FET source leads as discussed in the next section. Then add the brass walls

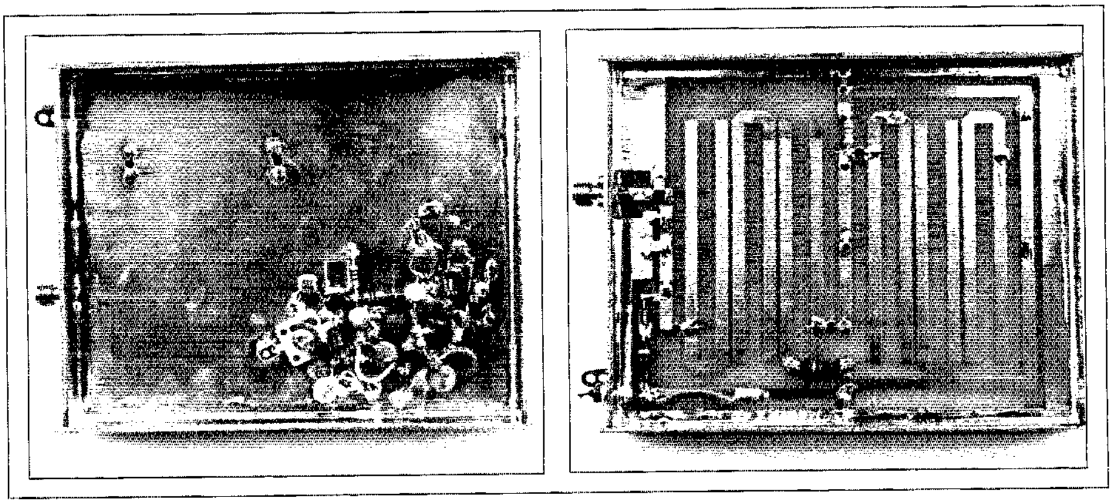


Fig 8—At left, a photo of the 106.5-MHz oscillator circuit, built ground-plane style on the back of the 639-MHz multiplier board. Oscillators built this way exhibit stability an order of magnitude better than etched PC-board versions. This is especially important for minimizing drift at the 96th harmonic of the oscillator frequency—10.224 GHz. At right, a top-side view of the 3¾- × 4¾-inch assembly shows the 639-MHz etched band-pass filters.

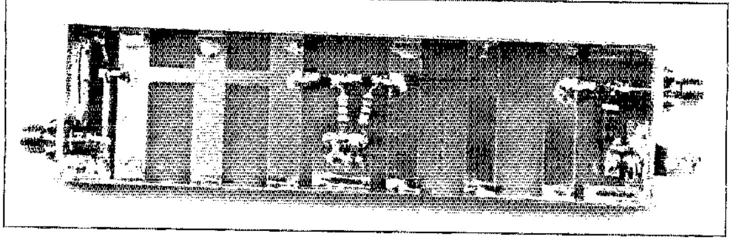


Fig 9—The 1 ¼- × 5-inch 639-MHz to 2.556-GHz multiplier board uses MMICs to provide a 5-mW filtered intermediate LO signal that drives the 10.224-GHz multiplier.

and install the connectors. Build the biasing circuitry after the transistors are installed.

Circuit performance may be improved slightly, as discussed in the next section, though the design is relatively broadband and should operate adequately despite minor construction variations. Computer simulations predict gain flatness within a decibel across the 10-GHz band.

I glue RF-absorptive rubber or foam to the insides of the enclosure lids. This reduces the chance of waveguide effects disrupting circuit operation.[13]

GaAsFET Installation Tips

Beware of soldering irons with significant ac leakage. People blow up lots of devices because their soldering iron tips aren't at ground potential. Measure your soldering iron's tip-to-ground potential if you have any doubts.

The circuits in this transverter use the GaAsFET source-lead inductance as a circuit component. Use the photos as guides when installing them. Bend the source leads down at the ceramic device body, then insert them into holes carefully cut in the circuit boards using a #1 X-ACTO blade or similar weapon, as is done in *The ARRL Handbook's* GaAsFET preamplifiers.[14] Be sure to cut the holes so that the device is centered on the board traces. Once the device is installed, bend the source leads up flush with the bottom of the board and solder them to it.

Of course, take the usual precautions when handling GaAsFETs, which are static-sensitive. Chapter 24 of *The ARRL Handbook* discusses these practices.

Adjusting the 2.556- to 10.224-GHz Multiplier

First, adjust the filter-tuning screw for maximum output. Next, set the bias trimmer for maximum power output. You may then want to tune the amplifiers. Do this using a tuning tool made out of a 1/16-inch-square piece of thin copper sheet or foil stuck into the end of a piece of Teflon tubing. Slide the tool along the input and output lines, looking for hot spots—places where the presence of the foil makes the power output increase. After finding them, turn off the power. Next, solder a piece of foil at each hot spot and adjust its position with high-quality tweezers.

Coming in Part 2

When you finish building the blocks described this month, you'll have a clean 10.224-GHz local oscillator. Next month, I'll describe the mixer/splitter board and the preamplifier/power amplifier circuit, and some 10-GHz antenna ideas.

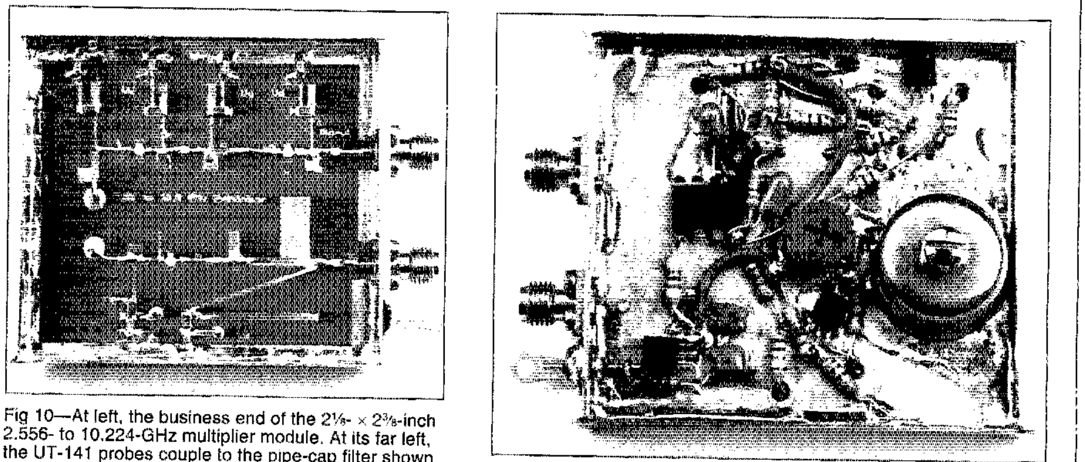


Fig 10—At left, the business end of the 2⅛- × 2⅞-inch 2.556- to 10.224-GHz multiplier module. At its far left, the UT-141 probes couple to the pipe-cap filter shown at right. The FET biasing circuits are also shown at right.

Notes

[1] It *may* be possible to tune up the system using a Gunnplexer unit with an S meter, but I haven't attempted it and can't guarantee that it will work. A Gunnplexer should be able to pick up a properly functioning LO even with lots of attenuation between them.

[2] *QST* has published a series of no-tune transverters developed by KK7B and WA8NLC, including versions for 903, 2304, 3456 and 5760 MHz. The most recent of these is J. Davey, "A No-Tune Transverter for the 2304-MHz Band," *QST*, Dec 1992, pp 33-39. See the notes at the end of D. Mascaro, "A High-Performance UHF and Microwave System Primer," *QST*, May 1991, pp 30-33, for details on the others.

[3] Z. Lau, "The Quest for 1 dB NF on 10 GHz," RF, *QEX*, Dec 1992, pp 16-17.

[4] R. Campbell, "A Clean, Low-Cost Microwave Local Oscillator," *QST*, Jul 1989, pp 15-21.

[5] K. Bailey, R. Larkin and G. Oliver, "TNT for 10 GHz," *Proceedings of Microwave Update '88* (Newington: ARRL, 1988), pp 80-95. This book (and other conference proceedings referenced here) is available from the ARRL Publication Sales Department for $12 plus shipping and handling.

[6] Amplifier stability is affected by the inductance of the FET source leads.

[7] I've also used little surplus boards with surface-mount NE555s, although some of these are poorly constructed and had to be resoldered for reliable operation.

[8] Z. Lau, "Power Supply for GaAsFET Amplifier," *QEX*, Mar 1991, pp 10-11.

[9] K. Britain, "Cheap Microwave Filters," *Proceedings of Microwave Update '88*, pp 158-162.

[10] G. Elmore, "A Simple and Effective Filter for the 10-GHz Band," *QEX*, Jul 1987, pp 3-5, 15.

[11] The ARRL BBS can be reached at 203-666-0578 (1200/2400, N, 8, 1); one 250-kbyte file, KH6CP10G.ZIP, contains all the PC-board artwork. Send paper and film artwork requests to the Technical Department Secretary, ARRL, 225 Main St, Newington, CT 06111. Request the MAY 1993 *QST* KH6CP 10-GHz TEMPLATE and be sure to indicate whether you need paper or film artwork. The template package also includes part-placement diagrams for the transverter's circuit boards.

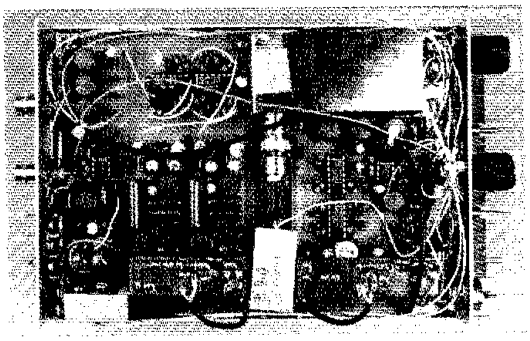
[12] This method will be more fully described by John Grebenkemper, KI6WX, in an upcoming *QST* article.

[13] K. Britain, "Works Great! Until You put it in the Box?," *Proceedings of the 25th Conference of the Central States VHF Society* (Newington: ARRL, 1991), pp 33-34. See Note 5 for ordering information.

[14] L. Wolfgang, ed, *The ARRL Handbook for Radio Amateurs*, 1993 ed (Newington: ARRL, 1992), pp 32-22 through 32-38.

Single-Conversion Microwave SSB/CW Transceivers

No-tune transverters have made it easy to get on the UHF/SHF bands—as long as you have a commercial 2-meter transceiver you can dedicate to that application. Here's how to put two other recent *QST* projects, the R2 direct-conversion receiver and T2 transmitter, to work with a transverter as an inexpensive and lightweight, high-performance, all-home-brew station.



(photos by Kirk Kleinschmidt, NT0Z)

By Rick Campbell, KK7B
Department of Electrical Engineering
Michigan Technological University
Houghton, MI 49931

In January, I described a high performance SSB/CW receiver called R2,[1] and last month I presented a companion multimode transmitter, T2.[2] These two modules can be used to build high-performance, direct-conversion (D-C) transceivers for *any band* below 500 MHz.

One of my favorite activities is weak-signal microwave work on mountaintops using small, portable SSB/CW systems. Using the R2 and T2 boards as a tunable IF for a no-tune microwave transverter, I can build high-performance microwave SSB/CW transceivers that are smaller and use less battery power than systems using commercial multimode 2-meter transceivers and transverters. This article describes some of the design philosophy behind single-conversion microwave systems, and presents a complete 1296-MHz transceiver and an IF system using a choice of two VFO circuits and the R2 and T2 boards.

Basic Design Considerations

Fig 1 is the block diagram of the SSB/CW transceiver. The transceiver has an SSB/CW generator and IF receiver, a pair of mixers and a local oscillator (LO) to perform the IF-to-RF frequency conversion, and RF amplifiers to set the receiver noise figure and increase the transmitter power. The same arrangement can be used for signal frequencies from a few kilohertz to hundreds of gigahertz.

The choice of intermediate frequency (IF) is determined by the selectivity of the RF filtering and the available IF-filtering components. An IF of about 10% of the signal frequency permits good image rejection with a reasonable number of noncritical RF tuned circuits. Conventional SSB/CW superhets for microwaves need to use several frequency conversions to reach a narrow-bandwidth crystal filter operating near 10 MHz. Each frequency conversion adds system complexity, spurious responses, and internally generated receiver signals ("birdies").

We can eliminate the need for crystal filters by using the R2 and T2 boards to build a tunable IF. Since both R2 and T2 work fine at 1/10 the frequency of any microwave band below 6 GHz, we can build single-conversion microwave transceivers that cost less and have fewer spurious responses and outputs than a traditional scheme.

Block Diagram

A complete microwave transceiver using a no-tune transverter board[3] and the R2 and T2 boards is shown in Fig 2. Fig 3 shows the first prototype of this system, and Fig 4 shows IF2, a 144-MHz transceiver intended for direct connection to a no-tune transverter. It's relatively easy to build a transceiver with these boards, because the two share no functions other than the LO. In fact, the interconnections between the boards are almost identical to the cables that interconnect a Collins S line and 62S-1 transverter for VHF-transceive operation!

Each board needs a +12-volt supply at the points marked **+**. Since the T2 board includes transmit-receive switching and R2 has a muting circuit, both boards are normally powered continuously. R2 requires an external speaker, volume control and filter, options for which were discussed in my January 1993 and August 1992 *QST* articles.[4]

T2 has microphone connections, a push-to-talk (PTT) solder pad marked **P** on the circuit board, and a solder pad marked **K** for key. **P** is grounded for transmitting SSB or CW. To disable audio modulation when transmitting CW, the audio preamp output marked **A** is shorted to ground. A grounded solder pad is etched next to **A** for conve-

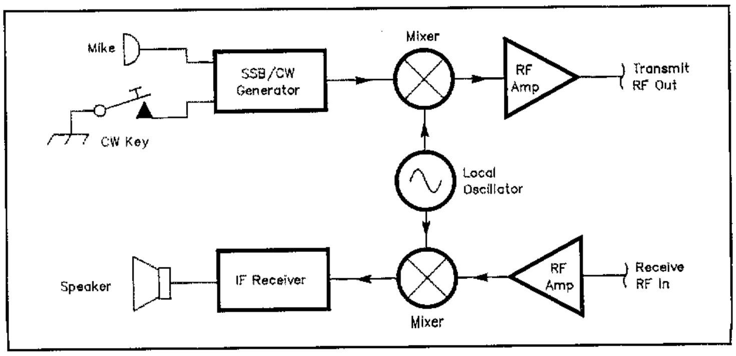


Fig 1—SSB/CW transceiver block diagram.

[1] Notes appear on page 34.

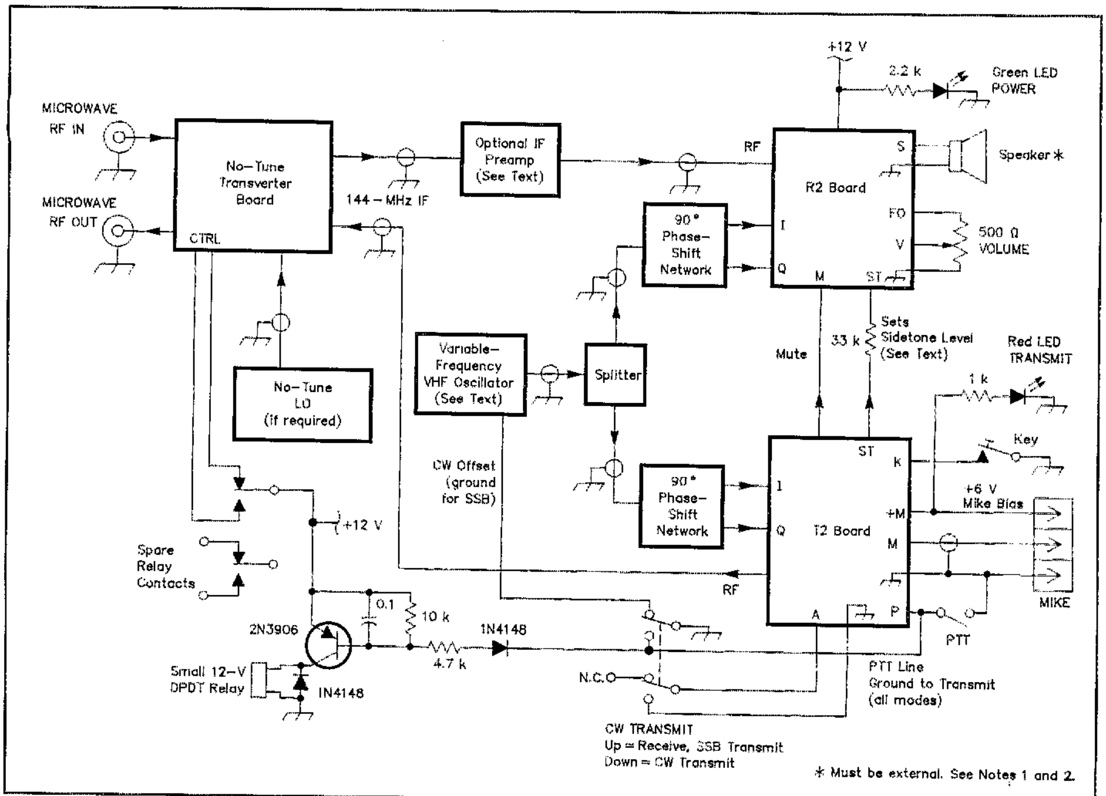


Fig 2—Interconnecting the T2 transmitter board, R2 receiver board, VFO and a no-tune transverter creates a direct-conversion, no-tune microwave transceiver. All of the no-tune transverters except the 903-MHz version use external local oscillators. Set the resistance value in the **ST** line to an appropriate level for your listening environment; 33 kΩ works well for relatively quiet locations. The volume connection (**FO** on the R2 board) *must* be terminated in 500 Ω (see Note 1).

nience in doing this. I use a DPDT **CW TRANSMIT** switch on my radios that simultaneously grounds the audio-preamp output, grounds the PTT line and offsets the local oscillator from zero beat.

Only two wires connect the R2 and T2 boards: mute and sidetone. To mute the receiver during transmit, connect a wire from the unmarked solder pad to the right of **P** on the T2 board to the solder pad marked **M** on the R2 board. The sidetone generator is on the T2 board, with an output marked **ST**. The sidetone input on the receiver board is also marked **ST**. Directly connecting these two points results in a very loud sidetone! For a more comfortable sidetone level, experiment with series resistor values between about 10 kΩ and 100 kΩ. I settled on 33 kΩ for portable operation in quiet locations. Of course, a 100-kΩ sidetone-level pot is an acceptable alternative.

Each board requires a pair of quadrature LO signals, as discussed in the R2 and T2 articles. The first transceiver I built used a

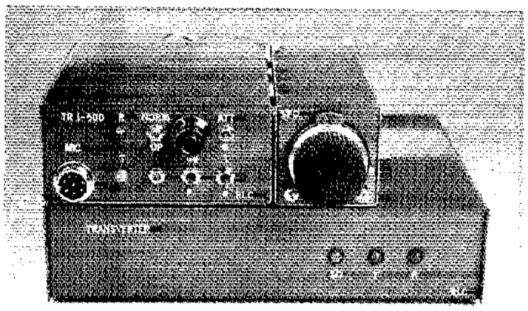


Fig 3—The first no-tune, single-conversion 1296-MHz transceiver that follows the block diagram of Fig 2.

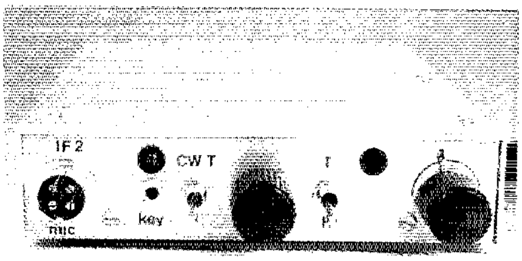
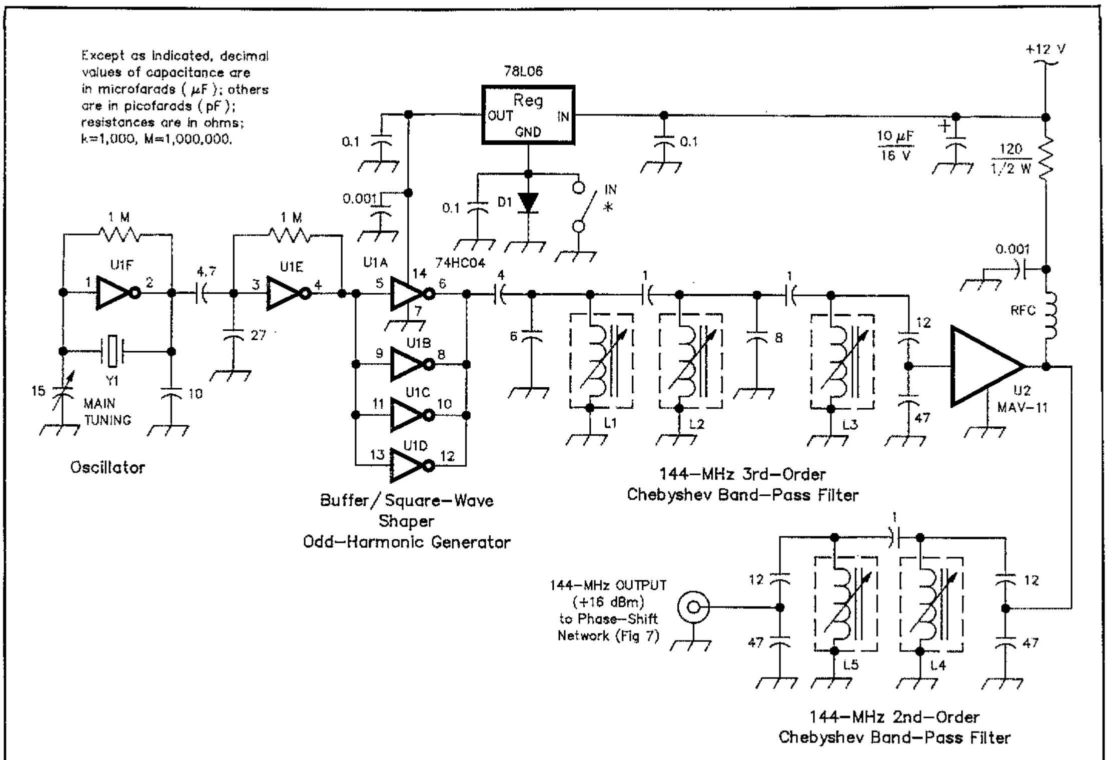


Fig 4—IF2, a D-C transceiver with an integral VXO (Fig 5). This 1¼ x 5¼ x 7½-inch transceiver can be directly connected to a no-tune microwave transverter to create a single-conversion microwave transceiver.



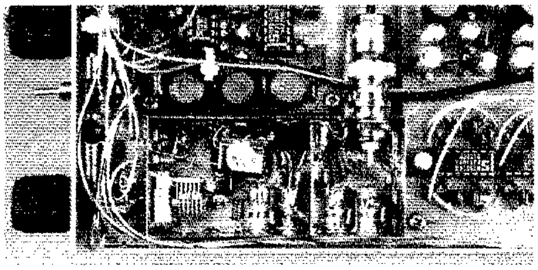


Fig 5—This simple VXO circuit has a tuning range of ±25 kHz. All capacitors less than 50 pF are NP0. When D1's anode is not grounded via switch or relay contacts, a 1-kHz frequency offset is imposed for CW operation. The photo shows the prototype VXO circuit built into the IF2 transceiver of Fig 4. The output network in this prototype differs slightly from the schematic (it uses glass trimmer capacitors for tuning; fixed, hand-wound inductors; and gimmick coupling capacitors, for example).

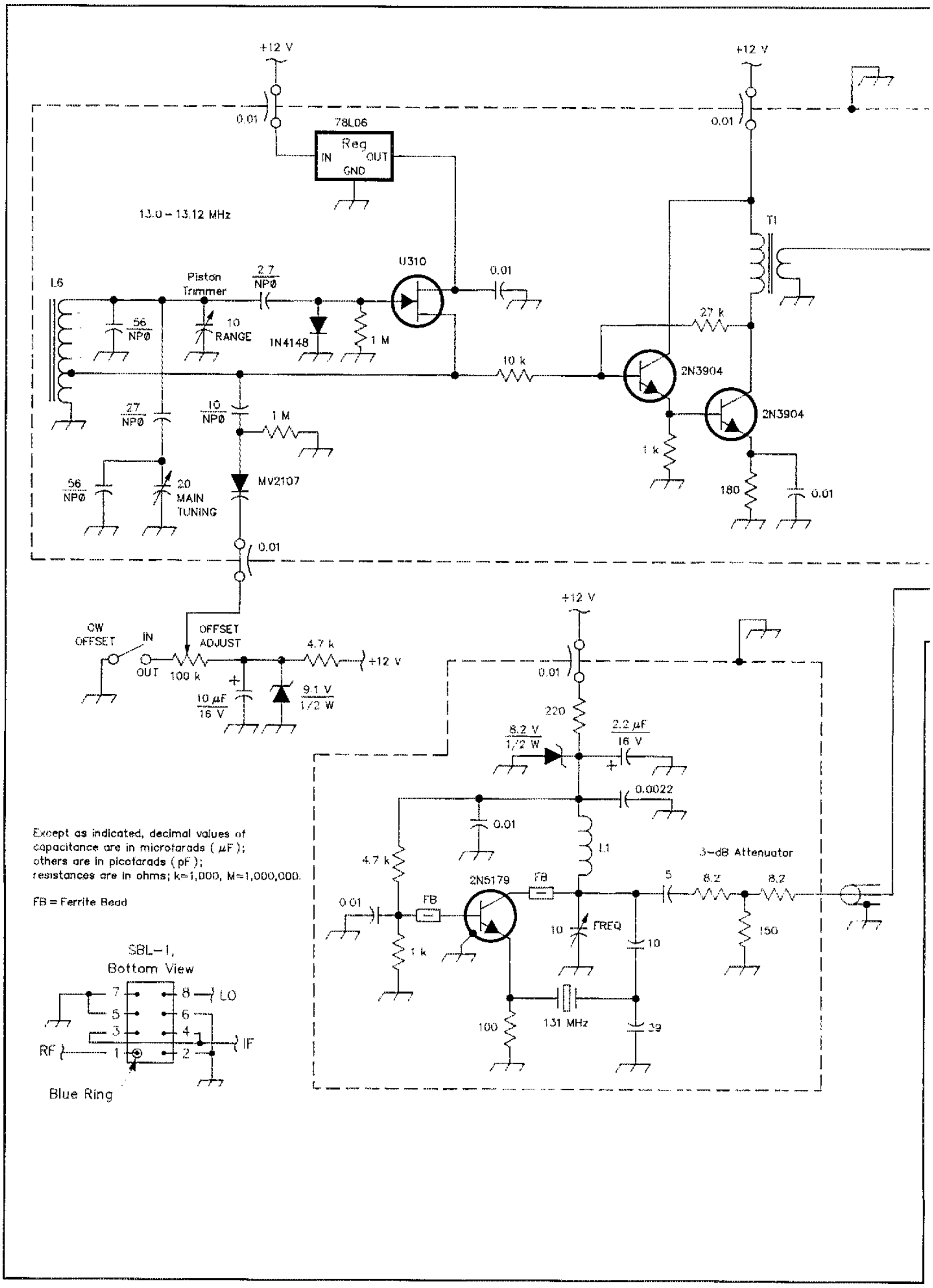
D1—Schottky diode (HP 5082-2835 or equivalent).
L1-L5—108-nH shielded variable inductor (Toko MC122 or equivalent).
RFC—8 turns #24 enameled wire closewound, ⅛ inch ID.
U1—74HC04.
U2—MAV-11 or MSA-1104 MMIC.
Y1—10-pF parallel-resonant crystal.
For 144.2-MHz center frequency, use 20.6-MHz crystal; for 144.1-MHz output, use 20.5857-MHz crystal.

single 90° LO phase shifter with a splitter on each output port to drive the I and Q ports of the R2 and T2 boards. I found that I could adjust the phase-shift network for good opposite sideband suppression from *either* the transmitter or the receiver, but not both! A better approach is to split the LO output first, then use separate phase-shift networks for transmit and receive. This allows independent adjustment of the receiver and transmitter for best opposite-sideband suppression.

Local Oscillators

The superhet system in Fig 2 uses single conversion with a crystal-controlled front end and a tunable IF. To make the IF tunable, we need a tunable LO. It's easy to build a stable VFO for frequencies as low as 5 or 10 MHz, but not quite as easy at VHF. Several approaches to VHF VFOs that are stable enough for SSB and CW have been used in custom and commercial equipment over the years. The simplest approach is a variable crystal oscillator (VXO), and the simplest VHF VXO I've seen appeared on the back of my napkin at lunch a few months ago. The most recent iteration is shown in Fig 5. This oscillator has a 50-kHz range at 144 MHz, reasonable stability, and spurs that are suppressed more than 72 dB. I obtained a CW offset of about 1 kHz by raising the 74HC04 supply from 6.0 to 6.4 volts by switching a Schottky diode in series with the 6-volt, three-terminal regulator's ground lead. This is about the minimum-parts-count CW-offset circuit I've seen, and it emphasizes the need for voltage regulation on crystal oscillators followed by frequency multipliers!

Another tunable VHF local oscillator, a *premixed* VFO, is shown in Fig 6. This circuit achieves oscillator and image spurious levels about 45 dB below the desired output, when properly shielded. A premixed VFO is simply a high-frequency VFO followed by a low-level transverter. *The ARRL Handbook* shows several examples of VFOs and transverters that can be combined to provide a stable, tunable LO for VHF. If the VFO frequency is less than about 10% of the



Except as indicated, decimal values of capacitance are in microfarads (μF); others are in picofarads (pF); resistances are in ohms; k=1,000, M=1,000,000.

FB = Ferrite Bead

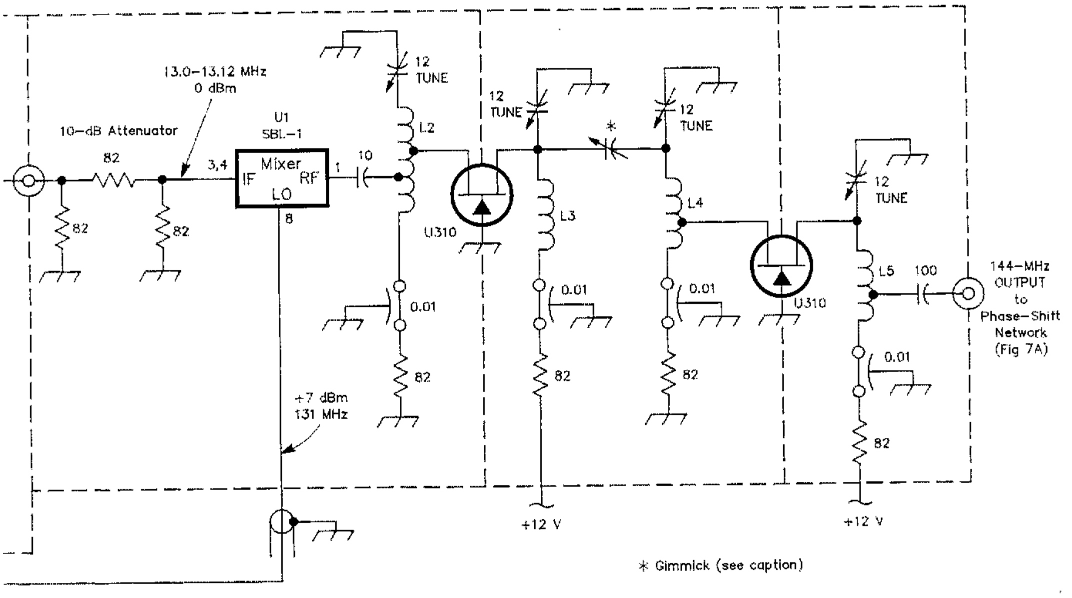


Fig 6—The premixed VFO circuit. The **CW OFFSET** switch is opened to enable the offset. This function can be handled by a relay contact or the front-panel **CW TRANSMIT** switch. The *gimmick* capacitor that provides interstage coupling consists of a ½-inch-long tightly twisted pair of #24 insulated wires.

L1—5 turns #24 enameled wire, ⅛-inch ID, ⅛ inch long.
L2—8 turns #24 bare wire, ¼ inch ID, ⅜ inch long. Tap input 2 turns from ground end; tap U310 source 3 turns from ground end.
L3—8 turns #24 bare wire, ¼ inch ID, ⅜ inch long.
L4—8 turns #24 bare wire, ¼ inch ID, ⅜ inch long. Tap U310 source 3 turns from ground end.
L5—8 turns #24 bare wire, ¼ inch ID, ⅜ inch long. Tap output 2 turns from ground end.
L6—24 turns #24 enameled wire on a T-37-6 core, tapped at 6 turns from ground end. For best stability, boil L6 in water for 2 minutes after winding it.
T1—Core: FT-37-43. Primary: 15 turns #22 enameled wire; secondary, 3 turns #22 enameled wire.

desired output frequency, then the transverter needs very good selectivity to remove the image and oscillator spurs. Helical filters, often used in VHF receiver front ends, can be used to obtain the necessary selectivity.

Another option for a stable variable VHF LO is a frequency synthesizer using a PLL, DDS or some other combination of letters. I have built a few of each type, but they were more complicated and didn't work as well as my VXOs and premixed VFOs, so I stopped playing with them for a while.

Whatever VFO type you choose, use lots of shielding. Commercial radios never have enough shielding, because it's expensive to manufacture and causes problems on the assembly line. Mount your VFO or VXO in a rigid, RF-tight box, and premix circuitry in a separate RF-tight box. Then enclose the whole radio in another RF-tight enclosure. Proper shielding eliminates many of the problems we have learned to tolerate in commercial radios, and once these bugs are gone, it's hard to go back to store-bought gear. Shielding is cheap in a custom radio, so use lots!

The VFO drives a splitter and phase-shift network (Fig 7). This circuit routes the VFO drive to the T2 and R2 boards, as well as handling phase-shifting for the I and Q ports on each board.

Interconnecting T2 and R2 with a No-Tune Transverter Board

The R2 and T2 boards interface directly with the receive and transmit IF ports of the no-tune transverters. The 1296-MHz transceiver in the title photograph follows the block diagram in Fig 2, without the optional IF preamplifier. The R2 noise figure is about 17 dB without the amplifier stage, so it may be useful to add a single MMIC amplifier stage if the transverter gain is low. The R2 receiver can hear signals at reduced sensitivity at odd harmonics of the VFO frequency, so the R2 input needs a simple low-pass filter. Since the 1296-MHz no-tune transverter board includes plenty of RF gain and a low-pass coupler at the mixer output, a piece of RG-174 can be connected directly from the transverter's receive-IF port to the R2's RF input.

The transmitter board contains an MMIC and low-pass filter, so a direct connection works with any of the no-tune transverters.

The no-tune transverters require a continuous 12-volt supply for the LO and multiplier circuitry, and switched 12-volt supplies for transmit and receive. I use the relay circuit in Fig 2, with a spare set of contacts available for hard-switching an external preamplifier and power amplifier.

Fig 8 shows the measured output spectrum of the 1296-MHz transceiver with the premixed VFO. It's a little cleaner than the no-tune 1296-MHz transverter driven by an ICOM IC-202, and *much* cleaner than the transverter slightly overdriven by a badly

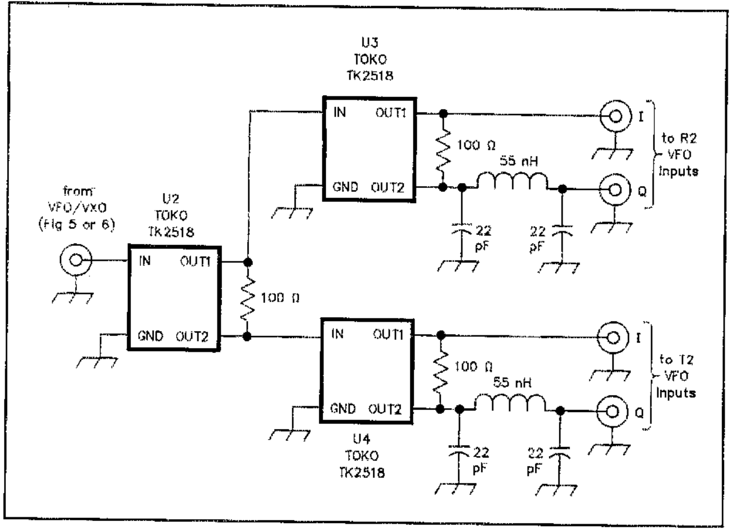
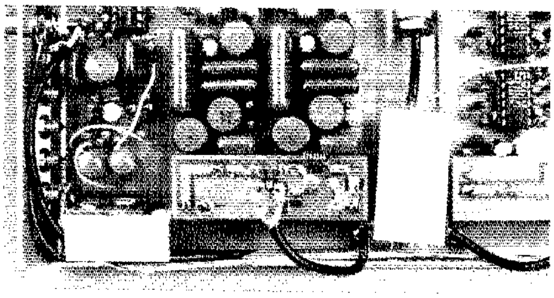


Fig 7—The power splitter and phase-shift network at A drive the R2 and T2 boards with either the VXO (Fig 5) or the premixed VFO (Fig 6). Phase delay in each phase-shift network is optimized by squeezing or spreading the turns of the two 55-nH inductors. Each inductor consists of 7 turns of #22 enameled wire spacewound on a 0.113-inch-ID form (#33 drill). "Piggyback" phase-shift networks are used in IF2 (Fig 4); a photo of the R2 board's network is shown below.



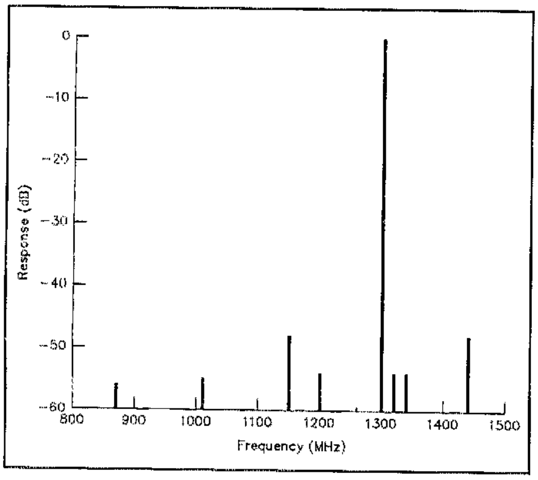


Fig 8—Measured output spectrum of the prototype no-tune 1296-MHz transceiver of Fig 3.

aligned FT-290R. The home-brew setup has higher transmit spurious levels than an ICOM IC-1271, but it also sounds better on both transmit and receive, and it has no birdies over the receiver's entire tuning range.

On transmit, the carrier and opposite-sideband suppression are comparable to VHF multimode radios, and the audio quality is far superior. On receive, the combination of high fidelity, low-distortion audio, decent opposite-sideband suppression, and no AGC is a real eye-opener. Loud, clear signals 40 dB above the noise simply disappear on the other side of zero beat. Weak signals at the noise level may be heard without the distractions of audio distortion and an AGC system that resets the gain on every noise peak.

The radio in the small aluminum box (Fig 4) shown in the photos is the 144-MHz IF for a 903.1-MHz SSB/CW system. It contains only an R2 board, T2 board, VXO and phase shift networks—simple! The no-tune 903 transverter is mounted at the antenna. Separate coax lines connect the transmit and receive IF to the T2 and R2 boards.

I didn't put RIT in either of these radios because I don't use RIT in portable microwave work. Battery-powered microwave radios on cold, windy hilltops tend to be pretty drifty, and it's better to follow other stations around the band than to try operating "random split." For radios with larger front panels, it's easy to add your favorite features.

Conclusions

This article describes one of my favorite applications for the R2 and T2 boards. The January *QST* article showed a 40-meter QRP CW transceiver using an R2 board. With the basic transmitter and receiver circuitry contained on a pair of small printed-circuit boards, the possibilities for custom-built radios are limited only by our imaginations.

I would like to thank the many hundreds of readers who have written to me since the R1 article appeared last August, and to apologize for not having enough time to respond to all of you. I am greatly encouraged by your thoughtful comments; it appears that the technical side of Amateur Radio is very healthy. Keep up all the good work—this stuff is fun!

I would also like to thank my family for tolerating all the evenings I've spent at the bench turning ideas into working radios, and all the times the telephone interrupted a bedtime story.

Notes

[1] R. Campbell, "High-Performance, Single-Signal Direct Conversion Receivers," *QST*, Jan 1993, pp 32-40.

[2] R. Campbell, "A Multimode Phasing Exciter," *QST*, Apr 1993, pp 27-31.

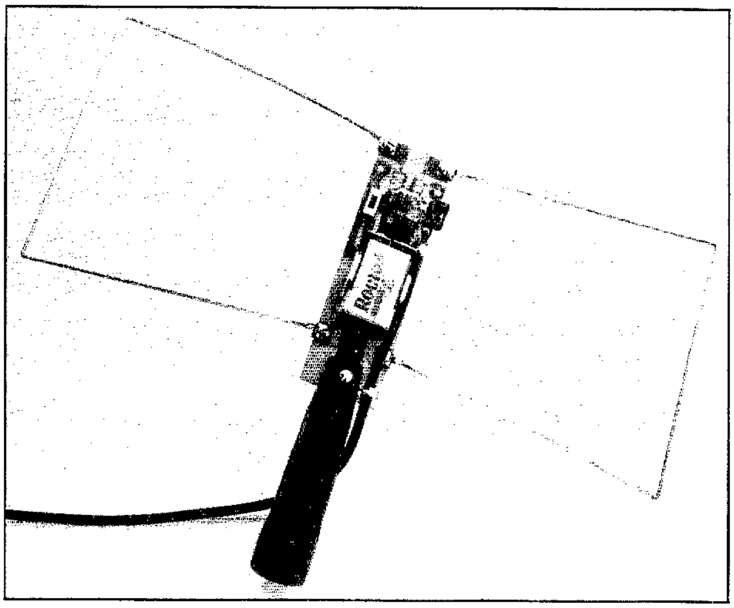
[3] No-tune transverter boards and kits for 432 through 5760 MHz are available from Down East Microwave, RR 1 Box 2310, Troy, ME 04987, tel 207-948-3741, fax 207-948-5157. Catalog available.

[4] R. Campbell, "High-Performance Direct-Conversion Receivers," *QST*, Aug 1992, pp 19-28.

Build the HANDI-Finder!

This hand-held direction finder is great for "fox hunting"! Simply connect it to the antenna input of your H-T or FM scanner and you can locate AM or FM sources over the range of 45 to 470 MHz.

By Bob Leskovec, K8DTS
25884 Highland Rd
Cleveland, OH 44143-2722



Inevitably, every Amateur Radio community experiences its share of repeater jammers. Tempers flare. A new generation of hams gets interested in direction finding. Car roofs start sprouting outlandish antenna arrays, looking more like tuna boats with every antenna addition. Fox hunts are scheduled for practice—and the jammers quickly become more evasive!

When that happened in our area a few years back, Rich James, N8FIL, of the Cuyahoga Amateur Radio Society (CARS), organized members from several area clubs into the "Bozo-Busters," and I resurrected four DOP-SCAN units[1] supplied by the Lake Erie Amateur Radio Association (LEARA). A few dedicated hams soon found themselves getting called out at all hours and driving all over town. When the gasoline bills started mounting up, we figured there had to be a better way!

Thought: Instead of a *few* hams outfitted with *special* equipment, why not have *many* hams equipped with *simple* direction finders? The units would have to be readily available to get a quick bearing whenever some troublemaker appeared with a strong signal on the repeater input near their location.

Inspired by a United States Coast Guard Auxiliary direction finder[2] and *The ARRL Antenna Book*,[3] I designed something simple, inexpensive, compact and practical. I call it the HANDI-Finder:[4] the *HAN*dheld *DI*rection-*Finder*. When equipped with open-loop wire antennas and a short handle, it stores flat (it'll fit in a briefcase) ready for use. An easy-to-build beginner's project, the HANDI-Finder is a good foundation for further experimentation by those with more experience. It's an easy way to introduce FM communications equipment users to the principles of direction finding. The basic circuit board can also be used to implement the similar DFing equipment described in the references, but with much lower power consumption. The HANDI-Finder makes a great club project![5]

How It Works

Refer to Fig 1. An electronic switch, formed by alternately forward-biasing diodes D1 and D2, connects one of two antennas to the feed line attached to the antenna input of an FM receiver tuned to the frequency of interest. Antenna switching is done at an audio rate, well within the receiver's audio passband and usually in the range of 400 to 1500 Hz, nominally 1000 Hz.

If one of the two antennas is closer to an interference source, it receives the wavefront a fraction of a second earlier than the other antenna. This constitutes a *phase difference* between the signals received by each antenna. The switching action imposes *phase modulation* on the incoming signal. This phase modulation is detected by the receiver and you hear it as a tone equal to the switching frequency. As the physical separation of the two antennas increases up to a half wavelength, the audio-signal amplitude increases.[6] The audio-signal pitch remains the same as the switching frequency.

When the antenna array is rotated so that the plane of the two antennas is perpendicular (or broadside) to the signal direction, both antennas receive the signal simultaneously and in phase. Because there is no longer a phase difference, the audio tone disappears. You perceive this as a rather sharp audio null as the array is rotated into a position perpendicular to the signal direction.

One disadvantage of this direction finder is that it exhibits 180-degree ambiguity, but that problem quickly disappears in the process of taking multiple bearings. This single disadvantage is offset by several advantages:

- It works on a *nulling*—rather than a *peaking*—principle. The null is sharp and much easier to detect than a peak in signal strength.
- The *audio tone* is nulled. This is unlike a conventional loop antenna or cardioid array that nulls the *carrier*. It's much easier to determine an audio null than a carrier null, especially if the carrier is unmodulated.
- Because audio is being nulled, you don't have to watch a field-strength meter—you just listen to the tone. That's something you can do while driving or hiking.
- Even with strong signals, *no attenuator is required* because this method uses phase, rather than amplitude, information. By comparison, the signal from directional gain

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[1] Notes appear on page 38.

antennas must be progressively attenuated to keep the received signal within the range of the S meter as you, the hunter, get close to the source.

Technical Overview

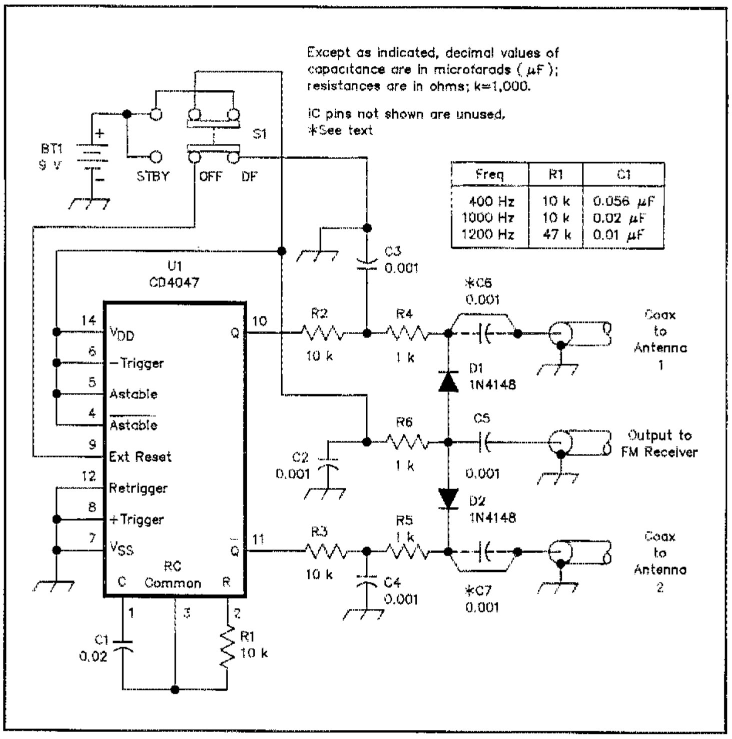
The HANDI-Finder employs a single CD4047B CMOS IC. This IC contains an oscillator and a flip-flop to automatically provide complementary symmetrical square-wave outputs without special adjustments. To set the frequency, you need but a single resistor and capacitor.

Very little current is used to bias the switching diodes, so the total circuit current drain is only 1.7 mA at 9 V. (No power-on indicator is used because even an LED draws 10 mA or more!) You get good service from an alkaline transistor-radio battery, and there's no need for an external power source. This has the additional benefit of eliminating the need for noise filtering and a voltage regulator—requirements when a vehicular battery is used.

All the parts—including the battery and two open-loop antennas—are mounted on a single PC board. There are no ground-plane or vertical dipole elements, no complex packaging items to fabricate and no flexible "rubber ducky" antennas to buy.

Circuit Description

U1's oscillation frequency (determined by R1 and C1) is divided by the flip-flop to produce complementary square-wave outputs at pins 10 and 11. These outputs swing close to the 9-V rail and ground. When pin 10 is high and pin 11 is low, current flows from the battery through R6, D2, R5 and R3, forward-biasing D2. D2 connects Antenna 2 through C5 to the receiver feed line. The 9-V drop across a total of 12 kΩ results in a bias current of only 0.75 mA. That current produces a drop of 0.75 V across R6. Meanwhile, D1 is biased off. No current flows through R2 and R4, so there is no voltage drop across them. As a result, D1's cathode is at a potential of 9 V. The voltage at the anode of D1 is 0.75 V lower than 9 V, so the reverse bias is equal to this difference of 0.75 V.



Freq	R1	C1
400 Hz	10 k	0.056 µF
1000 Hz	10 k	0.02 µF
1200 Hz	47 k	0.01 µF

Fig 1—Schematic of the HANDI-Finder circuit; equivalent parts can be substituted. Unless otherwise specified, resistors are ¼-W, 5%-tolerance carbon-composition or film units. Capacitors are disc-ceramic, 50-V units. C6 and C7 may not be needed; see text. The inset table shows the R1 and C1 values to use for a particular oscillator frequency and corresponding tone. (Note: C1 and S1 are specified primarily because of their physical size and because on-off-on slide switches are not readily available.)

C1—0.02-µF, 50-V metallized-film capacitor (Digi-Key P4517).

S1—DPDT micro-mini slide switch (on-off-on, ALCO MHS-223), available from Newark Electronics. Newark Electronics has many locations throughout the US: check your telephone book for a branch near you. Main office: 4801 N Ravenswood Ave, Chicago, IL 06040-4496, tel 312-784-5100, fax 312-784-5100, ext 3107.

Misc: 14-pin DIP socket, battery holder (Keystone #1291), PC board (see Note 5), 9-V battery, RG-58 coax cable downlead with connector, handle, four #6-32 × ½-inch binding-head screws, flat washers, nuts and lock washers.

Similarly, when pin 10 goes low and pin 11 goes high, D1 is forward biased, D2 is reversed biased, Antenna 1 is connected and Antenna 2 is disconnected. C6 and C7 block current flow from the antenna connection path that would interfere with the biasing action. (In applications where there is no dc path in the antenna circuit, C6 and C7 can be eliminated; see the following "Antennas" section.) C2, C3, and C4 are RF bypass capacitors, producing an ac ground at the bottom of R6, R5 and R4, respectively. Thus, each node in the RF-switching section is 1 kΩ above ac ground. C3 and C4 also act in combination with R2 and R3, respectively, to form a low-pass filter and round off the sharp edges of the switching waveform. This reduces the modulation bandwidth superimposed on strong out-of-band carriers and minimizes radiation back into the receiver passband.[7]

Assembly

Mount the IC socket, then the diodes, resistors, and capacitors, starting with the smallest items first. The **STBY/OFF/DF** switch has two frame tabs that should be soldered to the board, too. (If you use a solvent to remove soldering flux from the board, be careful not to get any on the switch; the solvent may dissolve the plastic parts!)

The battery holder is fastened to the PC board at one end by soldering its positive and negative terminal tabs to the board. The other end of the holder can be fastened to the board using ⅛-inch diameter pop rivets, or small flat-head screws. Whatever you use shouldn't protrude far enough to prevent the battery from seating properly in the holder.

Prepare one end of a length of RG-58 coax and attach it to the board. Cut off the excess braid so that only 3/16 inch lies flat against the PC board near its attachment hole. Fasten a nylon cable tie about ½ inch below the connection point to securely anchor the coax against the board and prevent flexing.

On/Off Switch

The switch has three positions: center is off, up is for DFing, and down is for straight receiving or standby. *Don't transmit through the unit* because that could burn out the diodes, because the switching system doesn't maintain a 50-Ω impedance, and because doing so may generate harmonics.

(I must admit, however, I've often keyed a 2-W H-T even in the DFing position to bring up a repeater for demo purposes and have had no problems—yet!) If you experience diode failure, try using the PIN diodes mentioned by Geiser (see Note 3), or the ECG555 (don't confuse this identification number with that of the popular LM555 timer IC).

Resistors Versus Inductors

Although the HANDI-Finder was originally designed to help locate 2-meter-repeater interference, it operates over a much wider frequency range because the chokes traditionally used in such RF circuits are replaced by resistors (see Note 3).[8] Values in the range of 1 kΩ to 1.3 kΩ will do, but the resistors should all be of the same value. If you want to experiment with increasing the forward bias current through the switching diodes at the expense of power consumption, decrease the values of R2 and R3 toward 1 kΩ, keeping both resistor values equal to maintain symmetry.

Antennas

Because the antennas used in the PC-board version have no path to circuit ground, there's no need for dc-blocking capacitors. On the PC board, the mounting holes for C6 and C7 are shorted by traces. If you want to install C6 and C7, simply cut the traces between the mounting holes of each capacitor.

You can get the HANDI-Finder running without extensive antenna fabrication by using two simple wire antennas (having the appearance of a bow tie—see the title-page photo) attached directly to the PC board. Take two equal lengths of stiff wire about 19 inches long and bend each one into a neat **U** shape. The bottom of the **U** should be 6 inches across. Form each end of the wire ends into a hook so it can be fastened to the screw terminals on the PC board.

The wire you use should be large enough to provide desired rigidity, but thin enough to allow fastening securely beneath the screw terminals. If the wire is too large, solder on spade lugs or similar terminations. Of course, it helps if the wire is a good conductor, but coat-hanger wire (or welding rod) works satisfactorily. Better choices are brass rod or brazing rod, of 1⁄16- to 3⁄32-inch diameter.

There are three terminals along each side of the board. Using #6-32 × 3⁄8- or 1⁄2-inch machine screws, washers and nuts, fasten one end of each antenna to the top terminal on each side of the board. Similarly, attach the other ends of the antennas to the bottom terminals. Don't solder the antennas directly to the board! The heat required may cause the foil to delaminate. Also, if the antennas are bumped, the foil can be pulled off the board.

The bottom antenna terminals are merely mounting points. They are electrically isolated. *Do not ground the bottom of the loop!* If you do, you'll create a closed loop that causes a carrier null in the direction of the signal. Such a loop antenna operates on a principle other than the intended one.

The active antenna element is the vertical portion of the open loop; the horizontal sections are primarily for support. The longer the vertical element, the more signal it receives. This shows up as more carrier, better quieting, or a stronger S-meter reading. A greater separation of the vertical elements produces greater deviation and more audio.

If the vertical sections are supported farther from the PC board to increase the audio amplitude, the longer, unshielded horizontal antenna sections receive more signal as well, degrading the signal separation. You may want to experiment with the proportions of the vertical and horizontal dimensions.

The PC board accommodates connections by coax cable to other kinds of antenna arrays. Just below the top antenna mounting screw on each side of the PC board is a ground pad. This is not used with the open loop antennas, but is used for the coax shield. Thus, if you do want to make a "Double-Ducky" direction finder as described in *The ARRL Antenna Book*, you can connect the two equal-length coax cables to the HANDI-Finder circuit board. When fastening small-diameter or stranded wires to the pads, first install crimp-on spade or ring lugs to the end of the wire. If you must wrap the wire around the screw, place it beneath a flat washer and wrap it in the direction that the screw tightens (clockwise). *Be careful not to over-tighten the screw.* Also, *don't use toothed lock washers on the board surface.* It's better to use slightly longer screws to accommodate flat washers and protect the board surface.

A more extravagant antenna system can be made by positioning two vertically oriented, multielement Yagis at opposite ends of a horizontal boom. Support the boom at its center on a vertical mast so it can be rotated. Use equal lengths of coax from each antenna and connect them to the HANDI-Finder board, which should be mounted in a protective enclosure (more on this later) at the center of the boom. The beams give greater forward gain and reduce the 180-degree ambiguity—but they're a little hard to use while mobile!

A Quick Handle

At the bottom of the PC board is an area for handle attachment. You can mount the board on a short handle or directly to a mast. To get going quickly, attach a piece of wood or metal to the PC board. If you use a round handle, you'll either have to make a flat cutout along one side, or cut a slot down the middle to accommodate the PC board. A round handle is best. The most utilitarian handle is described next.

Paint-Roller Handle

Buy an inexpensive paint roller equipped with a handle that's threaded for use with an extension pole. Most of the handles available are plastic, force-fit directly to the metal roller rod. Clamp the rod securely in a vise and pull off the handle. Hacksaw a slot in the handle, slide the PC board into the slot and fasten it with a machine screw.

Next, get a wood or metal paint-roller extension pole, preferably one made of three screw-together sections. Now, you have a flexible system. You can use the HANDI-Finder with the paint-roller handle alone, or screw on up to three lengths of additional "mast." You may even want to use a second extension for a longer mast. Some extension handles have different threads between the sections than they have at the roller end.

The HANDI-Finder works best when the first 1⁄4 wavelength of feed line to the receiver is kept vertical, parallel to the center line of the PC board. (Measure the 1⁄4 wavelength from the bottom antenna terminal.) If the coax moves, it can throw off the bearing. When using the unit with a short handle, make sure to hold it straight and high so that the cable hangs straight down. When using the unit with a long handle, use tape or cable ties to attach the coax to the mast for the first 1⁄4 wavelength.

Is an Enclosure Really Needed?

The HANDI-Finder can be used without an enclosure. Its long, narrow profile is designed to minimize wind resistance for the benefit of a mobile unit whose DFer may be trying to hold onto the unit mounted on a narrow mast protruding through a window. An enclosure increases drag. With only minimal care, unenclosed units have rattled around in many trunks and back seats without damage.

To fit the PC board into a case, you must cut off the bottom end (an irreversible procedure) and discard the battery holder in favor of a simple pigtail battery connector. I suggest you first become familiar with the HANDI-Finder's operation without a case. After that, you may not want to bother housing it.

If you decide to enclose the PC board, two Radio Shack boxes are good candidates: #270-257, and #270-293. The former is tall and thin and offers the least wind resistance; the latter is shorter and wider, and has a separate cover for battery access.

Mount the PC board with the component side against the inside wall of the case so that the slide switch handle protrudes through a slot. If you're careful to keep the components below the level of the slide-switch body, you can mount the board closely enough (using 5⁄16-inch long × 1⁄4-inch diameter spacers) so that the slide-switch handle is accessible from the outside. Use #6-32 screws through the existing antenna connection holes in the circuit board that are long enough to feed straight through to the outside of the case where the antenna wires can be attached.

Trial Run

It's best to start off with a test situation in which you know the location of the source, and experiment with getting a feel for the null. Hold the unit above your head and position it for the best null. The null is fairly sharp, but doesn't always manifest itself as a *total* null of the audio tone. If a jump in tone one octave upward is observed, or a buzz or a "twiddle" is observed, it indicates the reception of multipath signals. Change your location by a few feet and try again until a clean null is obtained. It's usually not possible to observe a null while moving.

If there's a strong transmitter in the area (such as a paging system a few megahertz from the DFing frequency), the receiver appears to desense when the HANDI-Finder is switched to the DFing mode. The sharp edges of the diode-switching waveform cause it to be rich in harmonics, all of which modulate that strong RF source and broaden its spectrum at the receiver input. (This is the same problem that bothers the more complex DoppleScAnt units that switch 4 or 8 antennas; see Note 3.)

Using the unit with a synthesized scanner or one of the new extended-coverage H-Ts allows you to work the ham bands, and other frequencies as well. Prototypes were evaluated over the range of 49 to 470 MHz with good results. As you get considerably above the 150-MHz design frequency, you may find that your unit nulls signals that aren't perpendicular to the PC-board plane. This is caused by minor differences between the two antenna circuits that can produce a significant phase shift as the frequency increases. Therefore, try to make your antenna structures as symmetrical as possible. However, a nonperpendicular null sometimes occurs in high-reflection areas (especially indoors) because of multipath reception. (Don't try to do any meaningful DFing indoors!) In any event, don't jump to conclusions. Take several measurements at different locations a few feet apart, and consider your findings carefully before deciding whether the problem is caused by location or is characteristic of the way you set up your HANDI-Finder.

Checking Direction Calibration

Once you have a good feel for how the unit operates, you can check the direction calibration. A round handle allows you to make a calibration mark at any angle. Test your unit outdoors in an open field by walking in a circle around a central signal source. A person keying an H-T held at arm's reach is the most expedient, but the rig must be held still. The null should always occur perpendicular to the plane of the antennas, but your radio or other factors may introduce phase shift. If the error is consistent as you walk around the circle, mark the true direction on the handle. If you do this, expect this calibration to change at different frequencies, and possibly if you change the CD4047 oscillator frequency.

If the error is not consistent, and changes as you walk around the source, you are experiencing multipath reflections from the surroundings. Try reducing the power of the source. For example, if a 100-mW H-T is too strong, remove its antenna and use a 51-Ω dummy antenna (such as an inexpensive 2-watt carbon resistor or a Radio Shack #21-506 dummy load). If you still get erratic readings, you'll have to try another location.

Once you've established the calibration mark, fasten the cable along the side of the handle or mast so it runs over the mark. That way, you can feel it in the dark. Now you're ready to do some serious DFing or fox hunting!

What'll You Hear?

When driving through our neighborhoods, it was interesting to scan the bands and see which frequencies are in use. One of our experimenters found several homes with 49-MHz baby monitors. You may also hear cordless-phone conversations. Another application is to locate interference on 145.250 MHz, pinpointing leakage from pole and lawn boxes used for CATV distribution.

If you're a radio-controlled model-airplane enthusiast, consider putting a low-power transmitter and "crash switch" in your plane. Then, if it gets away and crashes in the woods, you can track it down!

Some areas have rather extensive parks, reserves or other areas where naturalists tag wild animals with radio transmitters to track their migration. Naturalists are quite interested in reports on these animals, and may disclose the frequencies to individuals with a genuine interest in helping. Invite your local naturalist to speak to your club and demonstrate how you can help. But keep in mind they don't put radio transmitters on deer and fowl just to make it easy for game hunters to track and kill them!

Acknowledgments

My thanks to Rich, N8FIL, for his nagging; Russ, W8SQY; and Paul, N8HHG, for testing out the prototypes; Jim, WB8WTS; Dave, AD8Y; and Gary, KB3LP; for their encouragement and help with the manuscript.

Notes

[1] "DOP-SCAN" refers to a more complex direction finder called the DoppleScAnt. See T. Rogers, "A DoppleScAnt," *QST*, May 1978, pp 24-28. A multipage template package for this project is available for $1 from the ARRL. Address your request for the DOPPLESCANT CIRCUIT BOARD TEMPLATE to: Technical Department Secretary, ARRL, 225 Main St, Newington, CT 06111. The two PC boards are available from FAR Circuits, 18N640 Field Ct, Dundee, IL 60118-9269; price $30.

[2] Inspiration for this project is based on a design credited to Tom Feierabend, SO/CM 03N18, circa 1979, which appeared in a newsletter published in May, 1980 by Van Field, DCP XVIII, entitled *VHF Radio Direction Finding Manual for Coast Guard Auxiliary Use.* (No longer in print.)

[3] A similar circuit, referred to as the "Double-Ducky" direction finder (DDDF) designed by David Geiser, WA2ANU, is described in July 1981 *QST*, and reprinted in the 14th and 15th Editions of *The ARRL Antenna Book.*

[4] The spelling and form of HANDI-Finder as used throughout this text, and in all references to this design constitutes a trademark pending official registration with the US Patent Office.

[5] A HANDI-Finder kit is available for $27.95. It contains all the parts necessary to build the control circuit and includes an extensive instruction manual. You supply the 9-V battery, coax cable and connector(s) and appropriate antennas. Quantity discounts are available for the kits as well as for bare boards for club projects. Contact Rick Wells, K8SCI, at North Olmsted Amateur Radio Depot, 29462 Lorain Road, North Olmsted, OH 44070, tel 216-777-9460; MasterCard, VISA, Discover charge cards are accepted.

A PC-board template package is available free from the ARRL. Address your request for the LESKOVEC HANDI-FINDER TEMPLATE to: Technical Department Secretary, ARRL, 225 Main St, Newington, CT 06111. Please be sure to enclose a business-size SASE.

[6] Ideally, the amplitude should increase linearly from 0 to 180 degrees of wavelength, then decrease linearly from 180 to 360 degrees, etc.

[7] J. Moell and T. Curlee, *Transmitter Hunting: Radio Direction Finding Simplified* (Blue Ridge Summit: TAB/McGraw-Hill), 1987, page 137, Fig 9-15. Price: $19. This book is available from your local dealer or directly from ARRL. See the ARRL Publications Catalog elsewhere in this issue for ordering information.

[8] If you want to experiment with inductors, you can make them by close-winding one layer of #28 enameled wire on a 1-MΩ, ½-watt carbon-composition resistor. (The resistor value isn't important: It's the *physical size and shape* we're concerned with. The intent is to ensure that the shunt resistance is at least 100 times greater than the inductive reactance. Therefore, any resistor value above 100 kΩ is acceptable. Mouser Electronics carries a line of quality subminiature RF chokes (the 43LQ series) that are reasonably priced (38 cents each). The 1-µH value is part number 43LQ106.

Bob Leskovec has been licensed since 1957 when he got his Novice call sign, KN8DTS, at age 15. In 1958, he upgraded to General; he currently holds the Advanced class license. Also in 1958, Bob obtained a Second-Class Radiotelephone Certificate and the First-Class Radiotelephone Certificate with Radar Endorsement in 1960.

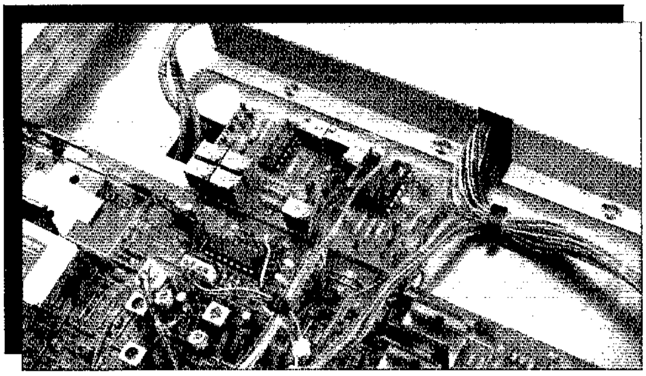
Bob is Director of Electronic Services at Case Western Reserve University, where he designs special electronic instrumentation for research, and consults in electronic product development. He serves as Staff Advisor and Facilities Manager for the Case Amateur Radio Club (CARC), W8EDU.

Bob holds five patents and has earned BS and MS degrees in Physics from John Carroll University. He spent several years as Chief Engineer of WCLV-FM in Cleveland and worked as a consultant in other areas of radio broadcasting. He has also worked with land-mobile, marine, aircraft, RF navigation and communications.

Add Tracking Sidetone to Your ICOM IC-765 Transceiver

With a stock IC-765, adjusting CW receiving pitch leaves the radio's sidetone oscillator pitch unchanged. This circuit generates a sidetone that tracks the '765's receiving pitch to within 1 or 2 Hz!

By Michael A. Mraz, N6MZ
15526 SE 50th St
Bellevue, WA 98006



One of the ICOM IC-765's few weaknesses is that its sidetone oscillator doesn't track its **CW PITCH** control.[1] This means that you can change the frequency of the '765's received CW note, but because the sidetone is fixed at 700 Hz, you have no reference against which you can zero-beat incoming signals if you set **CW PITCH** to anything other than 700 Hz.

Variable CW pitch is an extremely useful feature. But if you can't zero-beat by comparing the incoming CW signal to the sidetone, you'll probably never change the CW pitch. This is really unfortunate: I find a 700-Hz pitch to be very fatiguing after several hours of listening—especially during CW contesting.

Because I find the IC-765 to be an excellent CW transceiver in almost every other respect (especially its receiver's wide dynamic range and superb cascaded 250-Hz CW filters), I decided to design and install a tracking sidetone oscillator in my radio. Here's the circuit I came up with, and the rationale behind it.[2]

How the IC-765 Controls Received-CW Pitch

The IC-765's **CW PITCH** control varies the beat-frequency oscillator (BFO) frequency by "pulling" the radio's BFO crystal with a variable-capacitance diode. ICOM uses a nonlinear potentiometer (R317c) for the **CW PITCH** control. This partially compensates for the variable-capacitance diode's nonlinear voltage/capacitance characteristic. It also means that the BFO frequency is a nonlinear function of the voltage available on the **CW PITCH** potentiometer wiper line (labeled **PIT** in the '765 schematic). Fig 1 shows this nonlinear relationship.

Settling on a Tracking-Sidetone Solution

I had more than one design idea for the tracking sidetone. One idea was to mix the BFO output with a 9-MHz signal from a crystal oscillator—the same method used in the '765 to monitor its transmitted signal in the SSB and RTTY modes. The anticipated advantage to this circuit was that it would track the radio's BFO tuning perfectly. Unfortunately, however, the IC-765's BFO does not oscillate during CW transmit! An audio oscillator generates the '765's **MONITOR** signal on CW.

I decided that the most practical tracking sidetone circuit would be a function-generator IC controlled by the existing, nonlinearly variable PIT voltage. The only challenge was converting the PIT voltage's nonlinear variation into a linear variation suitable for driving the function generator's frequency-sweep input. Because I wanted the sidetone to track the BFO to within a few hertz, the PIT linearizing circuit would have to be quite accurate.

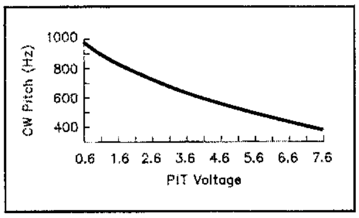


Fig 1—The IC-765's CW receiving pitch is a nonlinear function of the voltage on its **CW PITCH** control wiper line (PIT).

The Tracking Circuit

To linearize the PIT voltage, I used diodes and resistors to imitate a nonlinear dc amplifier. The diode/resistor networks change the gain of the circuit in six steps, as a function of the input (PIT) voltage. After calibration, the circuit is quite accurate and easily tracks the demodulated CW note to within 1 or 2 Hz.

Fig 2 shows the circuit's schematic. U1A-C, R1-R14 and D1-D20 linearize the voltage from the **CW PITCH** potentiometer. U1A and U1B are unity-gain buffer amplifiers. U1C and R8-R14 set U1C's dc offset and gain at the lowest PIT voltage (**CW PITCH** control fully clockwise). At the lowest PIT voltage, none of the diodes conduct, and the R1-R2 voltage divider determines circuit gain. As you turn the **CW PITCH** control counterclockwise past about 2 o'clock, the increasing PIT voltage causes D1 and D2 to conduct, which causes potentiometer R3 to parallel R2, thus decreasing the gain

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[1] Notes appear on page 42.

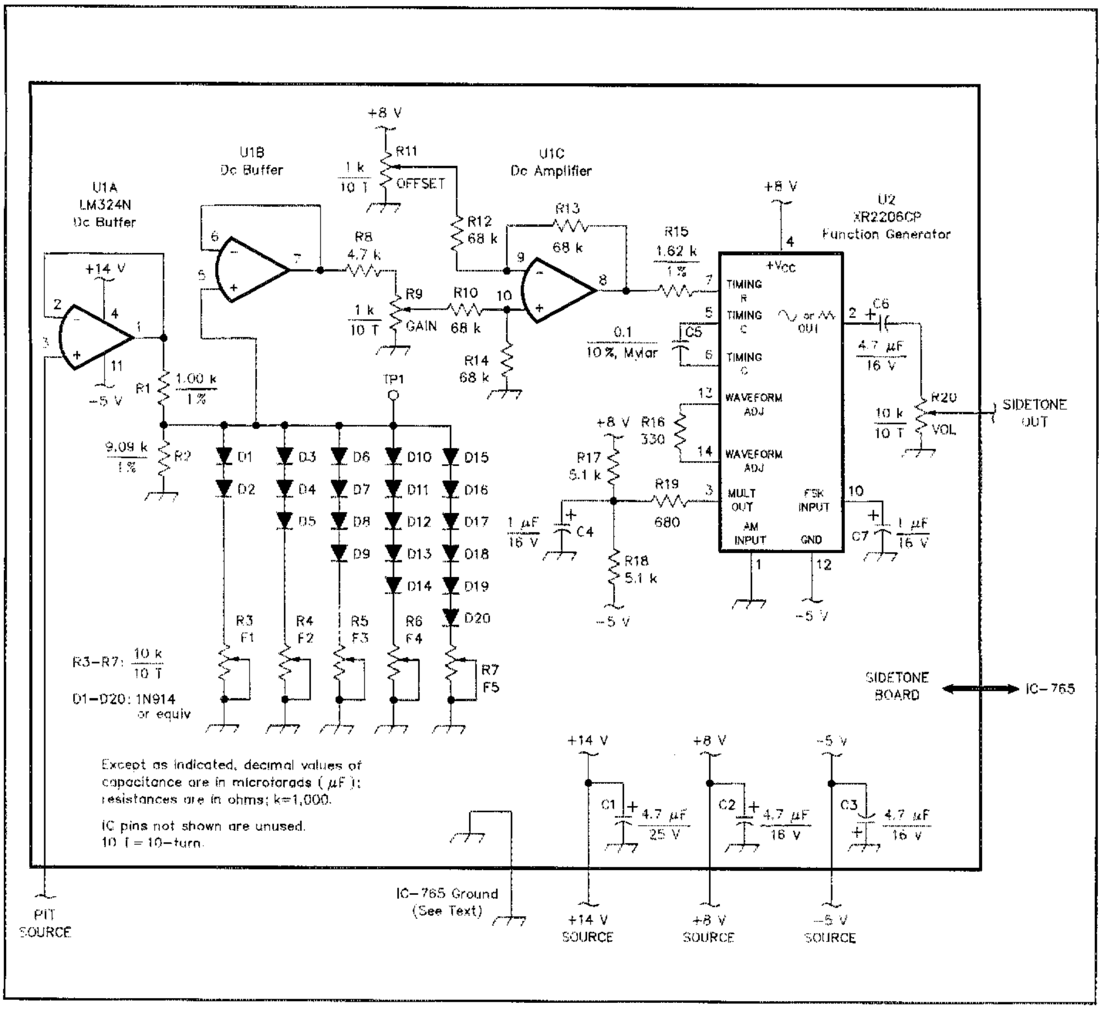


Fig 2—The tracking sidetone circuit uses the IC-765's PIT voltage to drive an Exar XR2206CP function generator (U2) via three sections of an LM324N quad op amp (dc buffers U1A and U1B, and dc amplifier U1C). Diode strings additively switch in compensation potentiometers R3-R7 as the PIT voltage increases, linearizing the voltage applied to U2 in five steps. Aside from the 1%-tolerance units, which are metal-film, all of the circuit's resistors are 5%-tolerance, carbon-film types. Except for C5, which is a 10%-tolerance Mylar type for better temperature stability, the circuit's nonelectrolytic fixed capacitors are general-purpose ceramic. The electrolytic (polarized) capacitors are tantalum. A complete parts kit, including a circuit board, is available for this circuit; see Note 3.

of the circuit. As the voltage increases further, additional diode/potentiometer branches conduct, further decreasing the circuit gain. In this way, the circuit converts the nonlinear PIT voltage into a linear voltage that can be used to drive the function generator's frequency sweep input.

The function generator is an Exar XR2206CP. This 16-pin IC contains a voltage-controlled oscillator, sine-wave shaper and associated bias circuits. The XR2206CP's output frequency is a linear function of the current drawn from pin 7, which is a function of the linearized PIT voltage at U1C pin 8. The '2206's sine-wave output has fairly low distortion, which results in a soft sidetone that is pleasant to listen to.

Construction

The tracking sidetone's circuitry operates at dc and audio frequencies, so you can use whatever construction technique seems appropriate. (I used haywire construction on a piece of perf board.) Complete kits of parts are also available.[3] However you build your version, make it no larger than about 2½ × 5 inches to ensure an easy fit into your IC-765.

Installation

Installing the board involves tapping into the 765's +14-V, +8-V, and -5-V power supplies, connecting the sidetone input to the PIT voltage, removing a resistor on the '765's Main circuit board, and connecting the sidetone output to one of the pads from which the resistor was removed.

You must also decide how to mount the sidetone circuit board. I chose to use two aluminum brackets. These attach the sidetone board to the IC-765 chassis using two of the screws that fasten the '765's Main circuit board to the chassis.

Unplug your IC-765 from ac power and place the transceiver, upside down, on a soft towel. Remove the lower cover from the transceiver. Turn the transceiver on its left

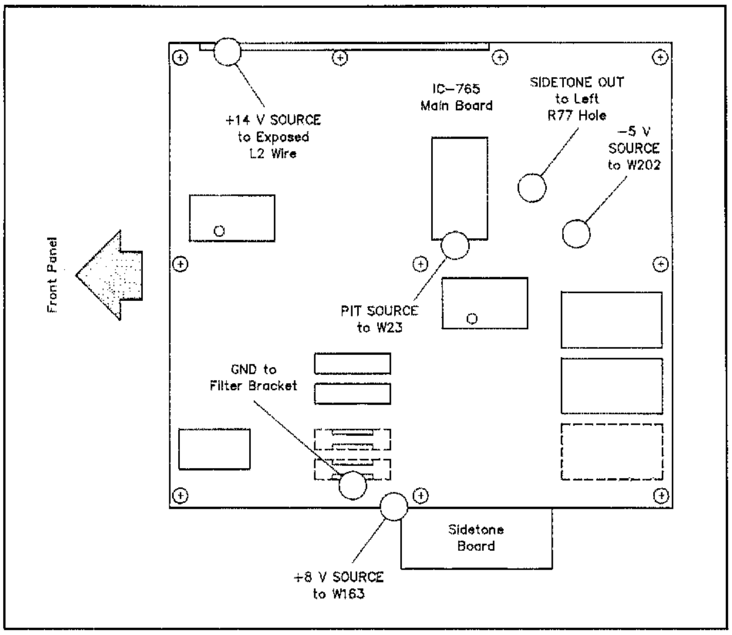


Fig 3—Overview of IC-765 Main board connection points for the tracking-sidetone module's **+14 V SOURCE**, **+8 V SOURCE**, **–5 V SOURCE**, **SIDETONE OUT** and **PIT SOURCE** wires.

side, so that the front panel is to your left as you look at the Main circuit board.

Locate L2, a wire-and-ferrite-bead inductor near the top-left corner of the Main circuit board; see Fig 3. L2 may be hidden by the large wiring harness. Carefully solder one end of a 12-inch length of #24 or #26 hookup wire to the exposed lead of L2. Label this wire **+14 V SOURCE**. Be careful not to melt the insulation on the wires in the wiring harness when you solder the wire to L2. *BE EXTREMELY CAREFUL* not to short the +14-V trace to any other component, or to the regulator heat sink which is immediately above L2.

Remove the 10 screws that attach the Main circuit board to the chassis. Using Fig 3 (and your service manual, if necessary), locate R77, and then carefully unsolder the resistor and remove it from the board. Solder a 12-inch length of hookup wire to the pad of R77 that is closer to the '765's front panel; see Fig 4. Label this wire **SIDETONE OUT**. Replace the Main circuit board; if you'll mount the sidetone board using brackets, do not replace the two screws that will attach the sidetone board to the chassis.

Locate jumper wire W163 on the Main board, using Fig 3 for reference. Solder another 12-inch length of hookup wire to W163. Label this wire **+8 V SOURCE**.

Locate jumper wire W202 on the Main board, using Fig 3 for reference. Solder another 12-inch length of hookup wire to W202. Label this wire **–5 V SOURCE**.

Locate jumper wire W23 on the Main board. Solder another 12-inch length of hookup wire to W23. Label this wire **PIT SOURCE**.

Place the sidetone board in its final location. Carefully cut each of the five new wires (**+14 V SOURCE, +8 V SOURCE, –5 V SOURCE, SIDETONE OUT** and **PIT SOURCE**) to appropriate lengths and strip and tin the ends. Solder the wires to the sidetone board and then double-check to ensure that they are connected to the correct pads.

Mount the sidetone board to the IC-765 chassis.If you home-brew your board and your mounting method completes its ground connection, fine; otherwise, you'll need to connect the sidetone board's ground to the IC-765's ground. (The outer surface of the nearest 9-MHz-filter bracket will suffice.) The LTA kit version requires a separate ground wire. The title photo shows the LTA sidetone board in place. Check all the connections and the wire dressing one more time, before you apply power to the '765.

Calibration

Attach the positive lead of a digital voltmeter (DVM) to TP1; the DVM's ground lead, to any chassis ground point. If you don't have a DVM, you can probably calibrate the sidetone based on the position of the **CW PITCH** control. Apply power to the radio and let it warm up for 10 to 15 minutes. Open the top hatch cover and turn the **MARKER** switch on. Put the '765 in CW mode and tune the radio to receive WWV strongly at one of its exact frequencies (15.00000 MHz, etc). Rotate the top-hatch **CALIBRATOR** control to zero beat the IC-765's marker with WWV.

Rotate the **CW PITCH** control fully clockwise. Tune the '765 to 14.00000 MHz in CW mode. You should hear the radio's marker as a strong (S9+) CW note in the speaker or headphones. Press the **LOCK** button to lock the tuning at this frequency. Push the **VOX** button (to turn VOX *off*) so that you will not transmit a CW signal, and then lock the key. You should hear the 14-MHz CW note and the sidetone beating against each other. If you don't hear the sidetone, adjust the sidetone board's **VOL** trimmer until you hear the beat note. Adjust the sidetone board's **OFFSET** trimmer until the sidetone and received 14-MHz signal are exactly zero-beat with each other.

The voltage at TP1 should be about 0.5 to 0.6. Rotate the IC-765's **CW PITCH** control until the voltage at TP1 is 0.9 (approximately 3 o'clock), and then adjust the **GAIN** trimmer to zero-beat the sidetone and received 14-MHz signal. Repeat this step until the sidetone tracks the 14-MHz signal perfectly between the two positions of the **CW PITCH** control.

Rotate the **CW PITCH** control counterclockwise until the voltage at TP1 is 1.2 (approximately 2 o'clock). Adjust trimmer **F1** for zero beat.

Rotate the **CW PITCH** control counterclockwise until the voltage at TP1 is 1.7 (approximately 12 o'clock, the detent position). Adjust trimmer **F2** for zero beat.

Rotate the **CW PITCH** control counterclockwise until the voltage at TP1 is 2.6 (approximately 9 o'clock). Adjust trimmer **F3** for zero beat.

Rotate the **CW PITCH** control counterclockwise until the voltage at TP1 is 3.1 (approximately 8 o'clock). Adjust trimmer **F4** for zero beat.

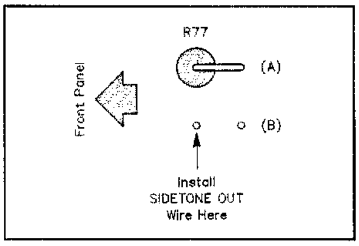


Fig 4—Connecting the sidetone's **SIDETONE OUT** wire requires the removal of the Main board's R77, a 270-kΩ resistor installed vertically. Finding R77 on the component side of the board is easy; finding and desoldering the corresponding pads on the underside of the Main board is more difficult. (The Main board graphic in the IC-765 service manual can help you find them.) As this top view shows, the circuit's **SIDETONE OUT** wire connects to the R77 pad that's closer to the IC-765 front panel.

Rotate the **CW PITCH** control fully counterclockwise. The voltage at TP1 should be approximately 3.4. Adjust trimmer **F5** for zero beat.

Slowly rotate the **CW PITCH** control from stop to stop, listening to the beat note. The sidetone should track the 14-MHz CW note very closely, within 1 or 2 Hz. If it doesn't track, repeat the calibration procedure. You can tweak the sidetone for more accurate tracking using the trimmers that correspond to the TP1 voltage breakpoints in the preceding paragraphs. After several months, I found that I used the 400- to 500-Hz range (approximately the 9 o'clock position) most often, so I tweaked **F3, F4,** and **F5** for perfect tracking in this range.

Turn the **MARKER** switch off, and then replace the top hatch cover. Adjust the **VOL** trimmer for the desired sidetone listening level. (You can use the '765's **MONITOR VOLUME** control to adjust the sidetone level when you're using the radio.) Remove power from the IC-765 and replace the bottom cover. This completes the modification.

Summary

Although this modification is not trivial to do, it's well worth the trouble. I have used the tracking sidetone modification for almost a year, including hundreds of DX contacts and three serious CW contest efforts. The circuit is rock-solid and does not drift with temperature as the '765 warms up. When I rotate the **CW PITCH** control back to the 700-Hz position and make a contact, I can't believe I waited over a year to design this circuit!

I have also installed the PBT modification as described in Hints and Kinks,[4] a mod that I consider absolutely indispensable for the '765. The combination of the IC-765's excellent receiver, the two optional 250-Hz CW filters, and these two modifications, makes it an unbeatable CW transceiver.

Notes

[1] M. Wilson, "ICOM IC-765 160- to 10-Meter Transceiver," Product Review, *QST*, Dec 1990, pp 52-55.

[2] Don't attempt to install this modification unless you're comfortable with invasive surgery on crowded circuit boards. If your IC-765 is still under warranty, a modification of this magnitude will surely void your warranty. As useful as the tracking sidetone is, be sure to consider the implications of voiding your warranty before proceeding.

[3] An etched, drilled circuit board and complete kit of parts for the tracking sidetone circuit is available for $49, plus $2 shipping to points in the US, from LTA, PO Box 77, New Bedford PA 16140, tel 216-565-9950. LTA intends to have tracking sidetone boards and kits available for sale at this year's Dayton HamVention; for details, and for particulars on export orders, contact LTA.

[4] R. Guski, "Variable-Bandwidth Tuning for the ICOM IC-765 Transceiver," Hints and Kinks, *QST*, Mar 1991, p 37.

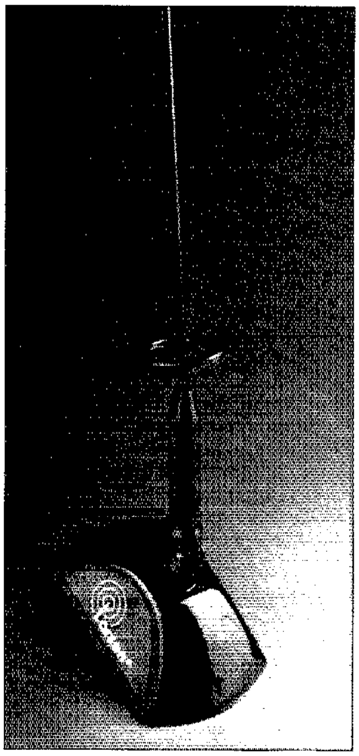
New Products

CORRECTION

◊ The Digi-Field wide-range signal-strength meter was listed on page 46 of March *QST* with a retail price of $79.95. IC Engineering informed us that the correct price is $119.95.

ON-GLASS 23-CM ANTENNA

◊ If you enjoy operating mobile on the "higher" UHF bands, you can do so without a mag-mount or permanent whip antenna. The KG-1290 is the first on-glass antenna for amateur use on 1.2 GHz. The manufacturer claims a low SWR from 1.235-1.432 GHz for this black ½-wave-over-¼ wave Kulrod. The antenna comes with Belden RG-58/U cable, Amphenol TNC or MPL connectors and 3M adhesive. Retail $59.95. Larsen Antennas, 3611 NE 112th Ave, PO Box 1799, Vancouver, WA 98682; tel 206-944-7551, fax 206-944-7556.



HF ANTENNAS

◊ If you're looking for a fresh approach to HF beam antenna designs, Force 12 features multiband no-trap antennas and systems, and monobanders designed for high performance, minimal weight, high strength and a neighborhood-friendly appearance. These contest- and DXpedition-proven Yagis present less tower, tripod and rotator loading, while providing considerable wideband gain and low SWR ratings. Force 12 antennas are constructed with wind-balanced multitapered, riveted elements, requiring no slots, hose clamps or sheet-metal screws. The elements come aligned and riveted to the boom. The unpolished aluminum elements and booms reduce sun glint, which should please the neighbors. Tailgunner technology provides the operator with the ability to immediately reduce the front-to-back ratio for instant listening and/or transmitting in the opposite direction.

Nomad dual-band direct-feed models cover the following frequency bands (boom lengths in feet): 10/12 (12, 16 and 24), 15/17 (16, 20 and 36), 12/17 (16), 10/15 (18), 20/30 (from 20). Nomad dipoles cover 12/17, 12/17/30 and single bands. There are single-band models for 10 meters with 3 or 4 elements on a 12-foot boom, 6 elements on a 24-footer, the Elite 510 with 5 elements on an 18-foot boom for $259, and the 8-element 308 Magnum for $489; 15 meters with 3 elements on a 16-foot boom, the 4-element Elite 415 for $449, and the 357 Magnum with 6 elements on a 36-foot boom for $695; the Elite 320 with 3 elements on an 18-foot boom for $379, the Elite 420 with 4 elements on a 30-foot boom for $549, and the 44 Magnum with 6 elements on a 44-foot boom for $849; 40 meters with a single linear-loaded dipole and the 3-element Elite 340 for $795.

Force 12 packages two Nomads and an Elite 320 antenna into its Strike Force SF-5B five-band package for $995. Short-stacking options, custom orders and consultations are welcome. Tom Schiller, N6BT, Force 12, part of BUY USA Inc, 3350 Scott Blvd, Bldg 6102, Santa Clara, CA 95054; tel 800-248-1985 or 408-988-2766, fax 408-988-2767.

INFORMATIVE CATALOG

◊ There's plenty of fun for browsers in the new Reference Catalog from the Radio Works. It's touted as "The Wire Antenna Discovery Book." Not only does it include specifications for all Radio Works baluns, cable, rope, connectors, adapters, wire antennas and accesories, it also features several pages of practical advice, theory and construction tips. It's an unpretentious 127-page 8½- × 11-inch black-and-white illustrated booklet, written in a friendly, folksy style and helps you to understand the applications for many components and designs, particularly parts and kits available from the Radio Works. There's also an extensive list of Radio Works products, although prices aren't included in the Reference Catalog—those are available in the firm's General Catalog. Retail $5. Jim Thompson Jr, W4THU, The Radio Works, PO Box 6159, Portsmouth, VA 23703; tel 804-484-0140, fax 804-483-1873.

An Introduction to Amateur Television

Part 2: The Basic ATV Station

In Part 1 of this three-part series, I discussed some of the dimensions of ATV activity, the bands used, what sphere of coverage can be expected, and how to eavesdrop on the activities of your local ATV group. This month, I'll discuss a few of the equipment options and technical details involved in getting your ATV station on the air.

By Ralph E. Taggart, WB8DQT
602 South Jefferson
Mason, MI 48854

Fig 1 is a block diagram of an entry-level ATV station. Other than provisions for connecting cameras, VCRs and so on, you can see that the installation is no more complex than that of a VHF/UHF FM transceiver. In part, this is because of the development of compact ATV transceivers available from three vendors. Each of the major elements of this diagram—the ATV transceiver, video sources, antenna, transmission line and optional power amplifier—are discussed briefly in the sections that follow.

ATV Transceivers

The AEA, P.C. Electronics and Wyman Research ATV transceivers (see Tables 1 and 2 and the accompanying photos) take all the fuss and guesswork out of setting up a basic ATV station. Although there are some real differences in engineering details, many functional aspects of these units are quite similar. The transceivers' power output range from 1 to as much as 15 watts PEP in the newer units. All operate from 12- to 14-V dc supplies, making them well-suited for mountaintopping or other portable, mobile or emergency use. Although the equipment dimensions vary, they're all quite compact; the Tridon 2000, for example, measures a mere 2.2 × 7 × 5.75 inches! Each transceiver has provisions for two video inputs and includes a power switch and 10-pin socket to handle cameras, along with a front-panel video-gain control.

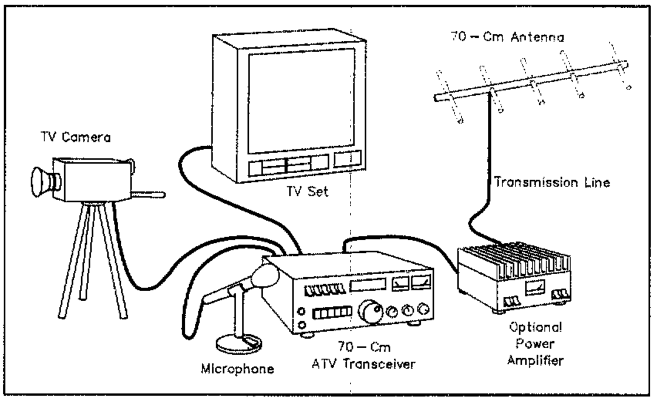


Fig 1—Diagram of a basic 70-cm ATV installation. The various station components are discussed in the text.

> ***ATV transceivers take all the fuss and guesswork out of setting up a basic ATV station.***

All the transceivers have manual TR switches and PTT TR switching. Each unit offers subcarrier-sound transmission, a subject I'll discuss shortly. A single crystal of your choice (439.25 or 434.00 MHz are the usually chosen frequencies) is supplied with each transceiver, and you can switch-select a second operating frequency if you've purchased an optional crystal (for a small additional charge).

Each transceiver is equipped with a tunable converter (AEA's unit has provision for use with a crystal-controlled converter) that covers the entire 70-cm band (420-450 MHz). (Note: ATV transceivers are available for operation on the 900- and 1200-MHz bands, too.) The converters employ low-noise GaAsFET preamplifiers and provide output on TV channel 3. You can change the output frequency to channel 2 or 4, to place the output on a channel unused in your area. The receive and transmit circuits handle full-resolution NTSC color signals with excellent fidelity. For color operation, simply use a color camera or camcorder as your video source and a color TV for display. Because black-and-white operation is far less critical than color, these transceivers do an excellent job with monochrome cameras and displays.

P.C. Electronics and Wyman Research are small, direct-sales operations, but each is operated by an ATV veteran, and each has an exemplary record of product and customer support. Their transceivers feature highly refined (but more or less standard) approaches to generating the transmitted signal: a small, video-modulated UHF transmitter strip, producing a DSB AM signal.

AEA is a much larger company with an extensive dealer network, probably best known for their data controllers, antennas and other products. Their VSB-70 transceiver[7] is a relatively recent introduction and takes a different approach to signal generation.

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[7] Notes appear on page 47.

Although ATV transceivers are definitely the easiest way to put a signal on the air, you have alternatives.

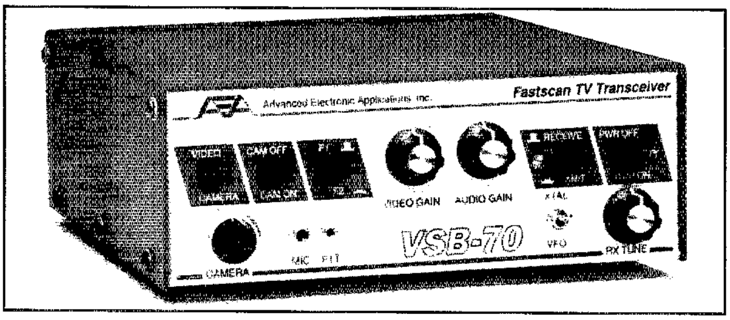
This transceiver is designed for vestigial sideband (VSB) AM transmission, which involves suppressing much of the lower sideband (by more than 60 dBc), just like commercial broadcast TV stations. The VSB-70 transmitted signal is generated at a lower frequency, then converted to 70 cm. The benefits and trade-offs of VSB represent a relatively complex subject that we'll look at in greater detail in Part 3 of this series.

Any of these ATV transceivers will do a good job for you. Chances are, an ATV group has at least one of each type, and you'll most certainly get plenty of advice in making your purchasing decision. Although these ATV transceivers are definitely the easiest way to put a signal on the air, you have alternatives. P.C. Electronics and Wyman Research offer a wide variety of kits; send for their catalogs. Also, low-power UHF transmitter strips are easy to video-modulate (see the Image Communications chapter of any recent edition of *The ARRL Handbook*), and will work well with any of the converters mentioned in Part 1 under "Looking In." If you already have a multimode UHF transceiver or transverter with a tube-type linear amplifier, you can try grid-modulating the final amplifier. Indeed, video modulating a solid-state transceiver's driver stage is *possible*, but few would venture to do the needed surgery! In such a case, operating the transceiver in FM provides on-carrier sound modulation and reception capability as an added bonus (see the discussion in the "Sound" section to follow).

Video Sources

For many ATV newcomers, the most common video source is the family camcorder. Usually, these gadgets just gather dust between family events, so why not get a little more mileage out of the investment and use it for ATV? If you don't have a camcorder, and ATV isn't a sufficient inducement to get one, you have other choices. First-generation portable VCR systems typically had a remote camera that mated with the recorder using a cable equipped with a 10-pin connector. You may be able to get one of these outdated systems at a bargain-basement price by shopping the classified ads, or scrounging through the back room of video-service shops. That 10-pin socket on the AEA, PC-Electronics and Wyman Research transceivers mates with the camera cable and is used to feed power to the camera.

With the consumer rush to buy compact 8-mm and VHS-C models, older, full-sized VHS and Beta camcorders are now passé; you can sometimes find them in the classified ads at attractive prices. Because the item most likely to fail on such systems is the *recorder mechanics*, many owners faced with a big repair bill elected to buy a new unit. Video-service shops frequently have the older camcorders on hand. As long as the *camera electronics* are functioning, they'll work just fine for ATV. Black-and-white surveillance cameras—generally exhibiting excellent resolution and good sensitivity under low-light conditions—can be purchased new for as little as $150; I've seen serviceable units at hamfests for $10-20! One of these will work just fine as your first camera and serve for handling station-ID chores if you later upgrade to color.



Advanced Electronic Applications VSB-70. This unit produces a near-broadcast-standard vestigial sideband (VSB) video signal, in contrast to the double-sideband (DSB) signal output by most ATV transmitters. *(photo courtesy of AEA)*

Table 1
Some ATV Equipment Vendors

Advanced Electronic Applications, Inc
2006-196th St SW
PO Box 2160
Lynwood, WA 98036
Literature request tel: 800-432-8873
Tel: 206-775-7373; fax 206-775-2340

Down East Microwave
RR1 Box 2310
Troy, ME 04987
Tel: 207-948-3741; fax 207-948-5157

P.C. Electronics
2522 Paxson Lane
Arcadia, CA 91007-8537
Tel: 818-447-4565; fax: 818-447-0489
Tom (W6ORG) and Maryann (WB6YSS) O'Hara

Rutland Arrays
1703 Warren St
New Cumberland, PA 17070
Tel: (orders) 800-536-3268,
(info): 717-774-3570

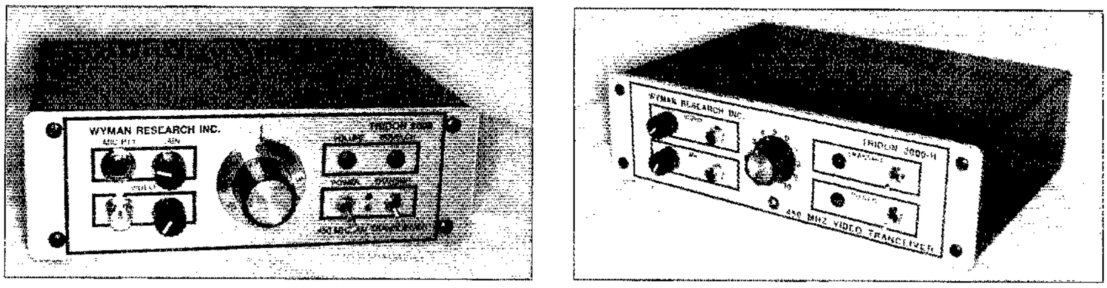
Spectrum International, Inc
PO Box 1084
Concord, MA 01742
Tel: 508-263-2145

Wyman Research, Inc
RR #1, Box 95
Waldron, IN 46182
Tel: 317-525-6452
Don (W9NTP) and Sue (W9YL) Miller

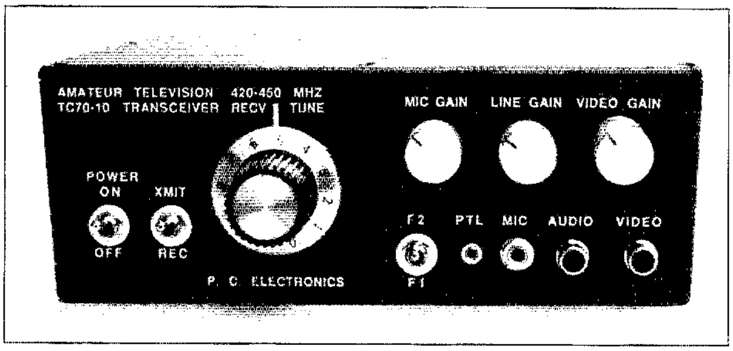
Table 2
A Summary of Features of Three ATV Transceivers

VSB indicates vestigial sideband format; SC = subcarrier FM sound; OC = on-carrier FM sound; NF = preamplifier noise figure (dB); Tune = tunable converter; Xtal = crystal-controlled converter. Prices are list and subject to change; AEA prices are discounted to varying degrees by dealers. Vendors include P.C. Electronics (PCE), Wyman Research (WR) and Advanced Electronic Applications (AEA).

	Transmitter						***Receive Converter***		
Model Price/Vendor	*PEP (Watts)*	*VSB*	*Sound SC*	*Sound OC*	*Video Inputs*	*Supplies Camera Power*	*NF*	*Tune*	*Xtal*
TC70-1D $329/PCE	1	N	Y	N	2	Y	1.0	Y	N
Tridon 2000 $320/WR	3	N	Y	Y	2	Y	0.8	Y	N
VSB-70 $350/AEA	1	Y	Y	N	2	Y	1.5	Y	Y



At left, the Wyman Research Tridon 2000, a 3-W transceiver. The latest addition to the line is the Tridon 2000H (right) capable of 12 W output. These units feature an attractive brushed-aluminum finish. *(photos courtesy of Wyman Research)*



The new P.C. Electronics TC70-10. You can adjust the PEP power output of this 70-cm transceiver from about 2 to 15 watts—quite a respectable signal. *(photo courtesy of P.C. Electronics)*

Table 3

Broadband 70-Cm Antennas Suitable for ATV Use

The models are listed from lowest to highest gain as claimed by the manufacturer over a range from approximately 9 to 16 dBi. Prices vary somewhat between vendors and are subject to change; some include shipping charges, others do not. Vendors include P.C. Electronics (PCE), Wyman Research (WR), Advanced Electronic Applications (AEA), Down East Microwave (DEM), Spectrum International (SI) and Rutland Arrays (RA).

Model Number	*Boom Length*	*Approx. Price*	*Vendor*
KLM 440-6X	28 in.	$ 57	PCE, WR
KLM 440-10X	64 in.	$ 68	PCE, WR
MBM 28-70	48 in.	$ 65	SI
KLM 440-16X	10.5 ft	$119	PCE, WR
MBM 48-70	6 ft	$ 90	SI
AEA 430-16	10 ft	$120	AEA
FO22-ATV	14 ft	$ 99	WR, DEM, RA
MBM 88-70	13 ft	$135	SI

For many ATV newcomers, the most common video source is the family camcorder.

Of course, VCRs and many computers with composite video output (a VIC-20, Radio Shack Color Computer [CoCo], IBM PCjr, CGA-equipped IBM PC and others) are also potential video sources. Computers can be programmed to provide snappy ID displays that are great for use during band openings. Small, dedicated add-on boards with IDs, cartoons and color-bar patterns custom-programmed into EPROM are also available.[8]

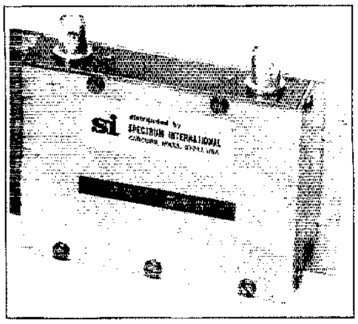
Sound

Commercial TV broadcast stations use a wideband (25-kHz deviation) FM sound (aural) transmitter operating 4.5 MHz above the video carrier frequency. In most amateur systems, a 4.5-MHz oscillator is frequency modulated by the microphone audio. This signal is then mixed with the video going to the modulator to produce an *aural subcarrier* 4.5 MHz above the carrier (and 4.5 MHz below the carrier in the case of a DSB system). When the subcarrier is set to the proper level (−15 dBc), the TV set reproduces the sound signal just as it does for a broadcast station. Subcarrier sound is a standard feature with all of the transceivers mentioned earlier.

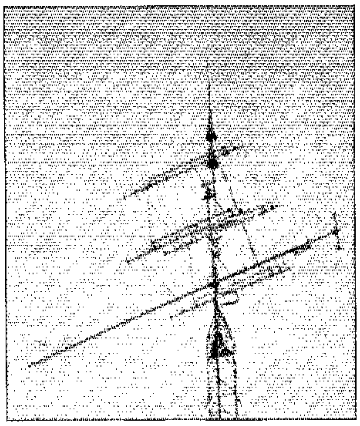
Although subcarrier audio is often used with repeaters and works well for simplex work when signals are reasonably strong, there are many reasons why it's rarely used as the primary audio-intercom link. (You're using the audio here to coordinate things, not necessarily add sound to the pictures.) First, in a typical group of ATV stations, high-gain antennas are moved around to favor some stations, while other stations temporarily receive a weak signal—or perhaps nothing at all. Under such conditions, the subcarrier audio system fails to provide the required audio-intercom function. Because TV sets don't have audio squelch, you're subjected to lots of noise during signal loss. Highly variable signal levels also make subcarrier sound unreliable during openings. Everyone should have subcarrier-sound *capability*, but it's not the best way to handle your primary audio needs.

An alternative to subcarrier sound is on-carrier sound in which the carrier is frequency modulated (narrowband) during video transmission. An auxiliary narrowband FM (NBFM) receiver, tuned to the carrier frequency, is used to copy the audio signal. NBFM has a 25-dB advantage over a wideband video signal, so on-carrier sound provides excellent *audio* copy even when the *video* signal has dropped into the noise! Thus, it provides good results for ATV group communication and during band openings. Unfortunately, not all stations use the on-carrier system, partly because it requires a second receiver. Of the transceivers noted earlier, only the Wyman Research Tridon 2000 and 2000H incorporate on-carrier audio in addition to the subcarrier audio. Wyman also markets a small, crystal-controlled receiver board ($99) that mounts inside the Tridon (along with a speaker and a volume and squelch control) to take full advantage of the on-carrier feature.

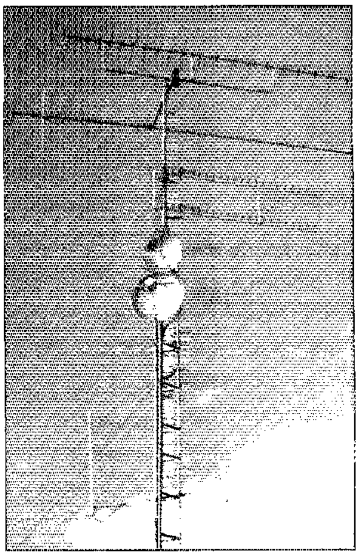
Primarily, audio is transmitted by a 2-meter FM transceiver. Custom varies in different regions of the country but, here in the Midwest, 144.34 MHz is the common ATV intercom frequency. Because of the huge bandwidth differential between ATV and NBFM, a 10-watt transceiver and a modest vertical (or small beam) is entirely adequate for solid copy with any station you can work on TV!



Spectrum International offers a VSB filter that can be added to existing ATV transmitters. *(photo by KB9FO)*



When beginning ATVers grow up: Bob (W5DS) Edlund is prepared for ATV using four stacked J-beams at 100 feet in addition to other aluminum assistants. *(photo by W5DS)*



Badly ATV-bug-bitten W8DMR has a hunk of aluminum to help him snag signals. On top is an 88-element J-beam that can be rotated to provide vertical or horizontal polarization. Next to it is a 2-meter Yagi. Just below that are 23-cm loop Yagis: one for transmit and one for receive. Below them are two dish reflectors for 13-cm operation. At the bottom is a 2-meter coaxial dipole antenna. *(photo by W8DMR)*

Antennas

ATV operation requires high-gain, wide-bandwidth antennas. The models listed in Table 3 have long track records of good service in this mode, and all have a low SWR across the entire 70-cm band. If assembled according to directions, each can be placed into service without further adjustment. The smaller, lower-gain arrays are excellent for local work because of their relatively wide beamwidth; they also do a superb job in portable service. If you have a long path to be covered, one of the longer, higher-gain units is the best choice. The larger antennas also provide an edge during band openings, but their very narrow patterns put a premium on pointing accuracy.

Antenna polarization used to be a matter of local custom, but now most groups use horizontal polarization on 70 cm unless vertical polarization is required for a specific repeater installation. The biggest potential for mutual interference is between ATV and the proliferation of FM repeaters and remote-base installations on 70 cm. Because the FM crowd's signals are vertically polarized, using horizontal polarization on ATV provides at least 20 dB of mutual attenuation between the two modes.

Proper antenna siting is often more important than gain. A high, unobstructed location gives the best results because *any* obstructions, including foliage during the summer months, can severely degrade signal strength. A well-placed, smaller antenna often outperforms a high-gain array that's not optimally located. These antennas are small enough to be turned by a TV antenna rotator, so installation need not be expensive. With the shift to CATV in many areas, you may already have a TV tower or tripod that can be put to good use without disrupting your other antennas. Using commercial power dividers or homemade phasing lines, any of these arrays can be phased to achieve higher gain and greater capture area. Some of the smaller arrays, such as the KLM 440-6X, can be phased to produce a high-gain system that can be set up on a side-arm secured to your tower whereas one of the longer antennas would have clearance problems.

Transmission Lines

In many ways, the single most important element in your station is the transmission line to the antenna. Table 4 lists a variety of common cables in terms of relative line loss. Using any of the readily available RG-58 and RG-8 cables impacts your installation in two ways: excessive power loss on transmit and degradation of receiver performance. The transmission losses are obvious from the table, but the effect on receiver performance is equally serious. Line losses add to the noise figure of the first RF stage. All the benefits of a state-of-the-art GaAsFET preamplifier can be nullified by line loss. The problem with line losses is most serious at weak-signal levels. ATV signals demonstrate a pronounced threshold effect between the point where sync is first detectable and where you can get enough of a signal to confirm a contact. Excessive line losses can raise this critical threshold by decreasing your transmitted and received signal power.

Table 4

Commonly Available RF Transmission Lines

Transmission lines are ranked by their attenuation (dB/100 ft) at 440 MHz. To emphasize the importance of line losses, I've shown the peak power output delivered to the antenna, assuming use of a 100-foot transmission line and a 10-watt-output transmitter. The cables in the upper half of the list are easy to obtain, but generally unsuited to serious ATV work.

Cable Type	*Line Loss (dB/100 ft)*	*Power (Watts)*
RG-58 (solid dielectric, 50 Ω)	12.0	0.6
RG-58 (foam dielectric, 50 Ω)	7.1	2.1
RG-8 (solid dielectric, 50 Ω)	5.0	3.3
RG-8 (foam dielectric, 50 Ω)	3.9	4.2
½-inch Hardline (75 Ω)	2.0	6.7
½-inch Hardline (50 Ω)	1.8	7.0
¾-inch Hardline (75 Ω)	1.6	7.3
½-inch Heliax (50 Ω)	1.5	7.4
¾-inch Hardline (50 Ω)	1.3	7.8

You can optimize all the other components of your station and still suffer from your choice of transmission line!

Although premium transmission lines such as Andrew Heliax or Hardline (aluminum-jacketed coaxial cable) are expensive when purchased new, you have to take a realistic look at the consequences of making do with lossier cable. As an extreme case, compare the results obtained with RG-58 cable and ½-inch Hardline. Using Hardline and a 1-watt exciter, you actually have more power delivered to the antenna (0.78 watts) than you would with a 10-watt amplifier and the RG-58 cable (0.62 watts). You'd also achieve a 10-dB receiver noise-figure improvement by using the Hardline. With the RG-58, you've wasted the investment in the amplifier and completely nullified the effectiveness of a perfectly good preamplifier! It's extremely frustrating to watch snow on the TV screen while neighboring stations are describing the incoming picture from a DX station!

Fortunately, there's a way to obtain premium transmission line without investing a lot of money. CATV companies use large amounts of ½-inch and ⅞-inch 75-Ω Hardline. Cut-off lengths (useless to the cable company, but more than enough for a typical ham antenna installation) can often be had for the asking. To use the cable, you have to solve two problems: Find suitable connectors and determine how to use 75-Ω cable with equipment and antennas designed for 50-Ω systems. Both of these problems are easily overcome using custom-made connectors manufactured by ZD Engineering (605 Balsley Ave, Findlay, OH 45840, tel 419-424-8765). The N connectors, which sell for about $25 a pair (specify the 70-cm band when ordering), feature a built-in balun for efficiently transforming the 75-Ω line to 50 Ω. A pair of these connectors can make a length of 75-Ω cable completely compatible with your ATV system. A short length of flexible cable (such as Belden 9913) is required to connect the Hardline to the antenna, and some short cable runs are required in the shack, but this is the case even if you're using 50-Ω Hardline or Heliax. A good transmission-line installation is a long-term investment. You can change all the other components of your station and still limit your total effectiveness by the decisions you made when you installed your transmission line.

In Part 3 of this series, we'll look at equipment options and the technical details of putting an ATV power amplifier into service and some aspects of the current debate over vestigial sideband. We'll also take a peek at ATV repeaters and experimental subjects such as FM TV transmission.

Notes

[7] R. Taggart, "Advanced Electronic Applications Fast-Scan Television System," *QST*, Product Review, May 1992, pp 56-59.

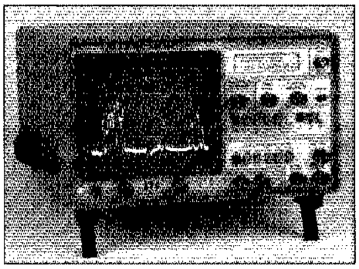
[8] Such ID generators and programmed video sources are available from P.C. Electronics and Elktronics. See Table 1 for addresses and telephone numbers.

New Products

LOW-COST SPECTRUM ANALYZERS

◊ Precision radio servicing and testing are now within reach of amateurs with new equipment from Instruments Technology Corp (ITC). ITC's "ultimate inexpensive analyzer" is designed for two-way radio servicing; tuning cavities, filters and duplexers; EMI and RFI testing; and checking RF systems, cables and amplifiers. ITC SA-series analyzers give the user full control over resolution and frequency bandwidth combinations, vertical position, baseline clipping and sweep speed, and come with an adjustable video filter. Specifications include −110 dBm sensitivity and an 80-dB onscreen dynamic range. The Model SA600A (2-600 MHz) is $1295, the SA1000A (2-1000 MHz) is $1595 and the SA1800A (2-1800 MHz) is $1895.

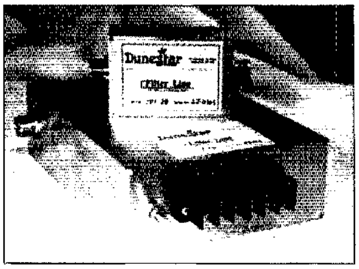
The SA500 adapter turns any oscilloscope into a full-function spectrum analyzer. It covers 2-600 MHz with −110-dBm sensitivity and an 80-dB onscreen dynamic range. It requires only two connections to your scope. Retail $549.95.



Options for the ITC SA series include a center-frequency display and 10-digit frequency counter for $275, ±5 kHz narrowband filter that extends the low-frequency range to less than 200 kHz for $300, a built-in tracking generator with a 60 dB dynamic output range for $250, and a 50-MHz marker generator at no charge (for a limited time). Ron Kunzel, ITC Instruments Inc, 3678 Mt Ariane Dr, San Diego, CA 92111; tel 800-232-3501 or 619-277-4619.

BANDPASS FILTERS

◊ If you need serious RFI suppression or band-to-band isolation, Dunestar's transceiver-bandpass filters might be the solution. On transmit, they help protect against television interference (TVI), intermodulation distortion (IMD), phase noise and harmonics, and on receive they suppress unwanted out-of-band signals. Aside from diminishing interference to entertainment equipment, they can also be used in multitransmitter situations, such as contests, DXpeditions or Field Day.

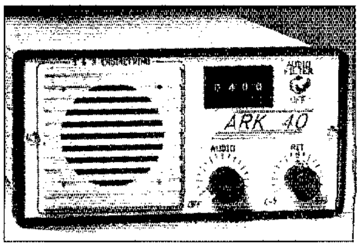


The Model 505 is a five-band remote-switched system with individually selectable bandpass filters for 80, 40, 20, 15 and 10 meters. Built-in direct-bypass and failsafe circuits allow the unit to be left inline even if power to the unit is lost, or when used on 12, 17 or 30 meters or for general-coverage reception. LED indicators show active action. The 505 is designed to track with antenna-switching systems or transceiver bandswitching interfaces. It uses simple closure-to-ground activation to integrate with existing setups. It operates on 12 V dc. Application examples are included with the instructions. Retail price $150.

The Model 101 is a single-band filter of the same design in versions for 80-10 meters, retail price $31.50. Custom frequencies are also available for both series. Dunestar Systems, PO Box 37, St Helens, OR 97051; tel 503-397-2918.

QRP KIT

◊ The ARK-40 synthesized QRP CW transceiver kit covers the 40-meter CW subband in 100-Hz tuning steps. It operates on 11-13.8 V dc, provides adjustable power output from millwatt levels up to 3-4 watts, and has front-panel selectable 200-Hz audio filter and receiver incremental tuning (RIT). Retail $269.95. S&S Engineering, 14102 Brown Rd, Smithsburg, MD 21783; tel 301-416-0661.

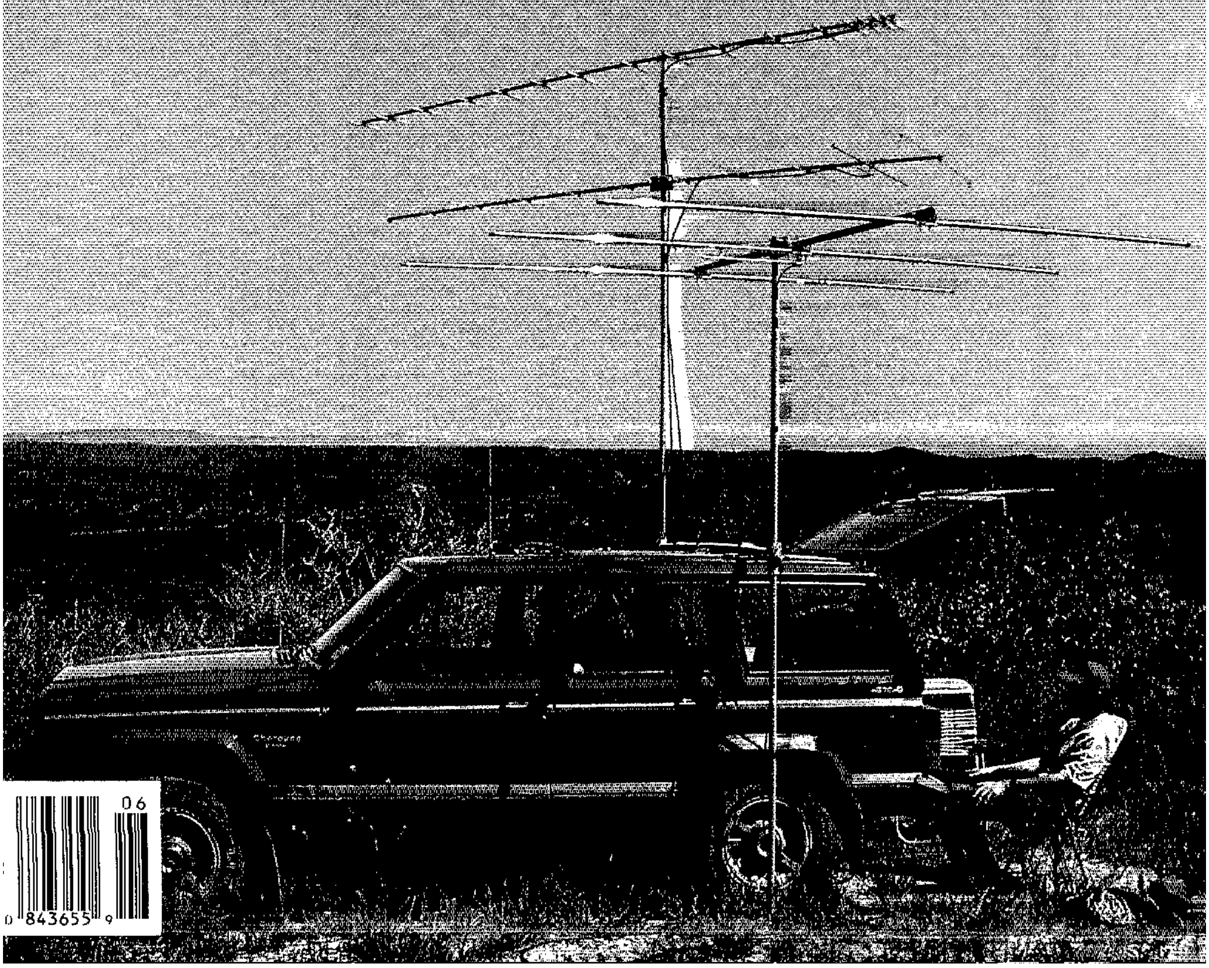


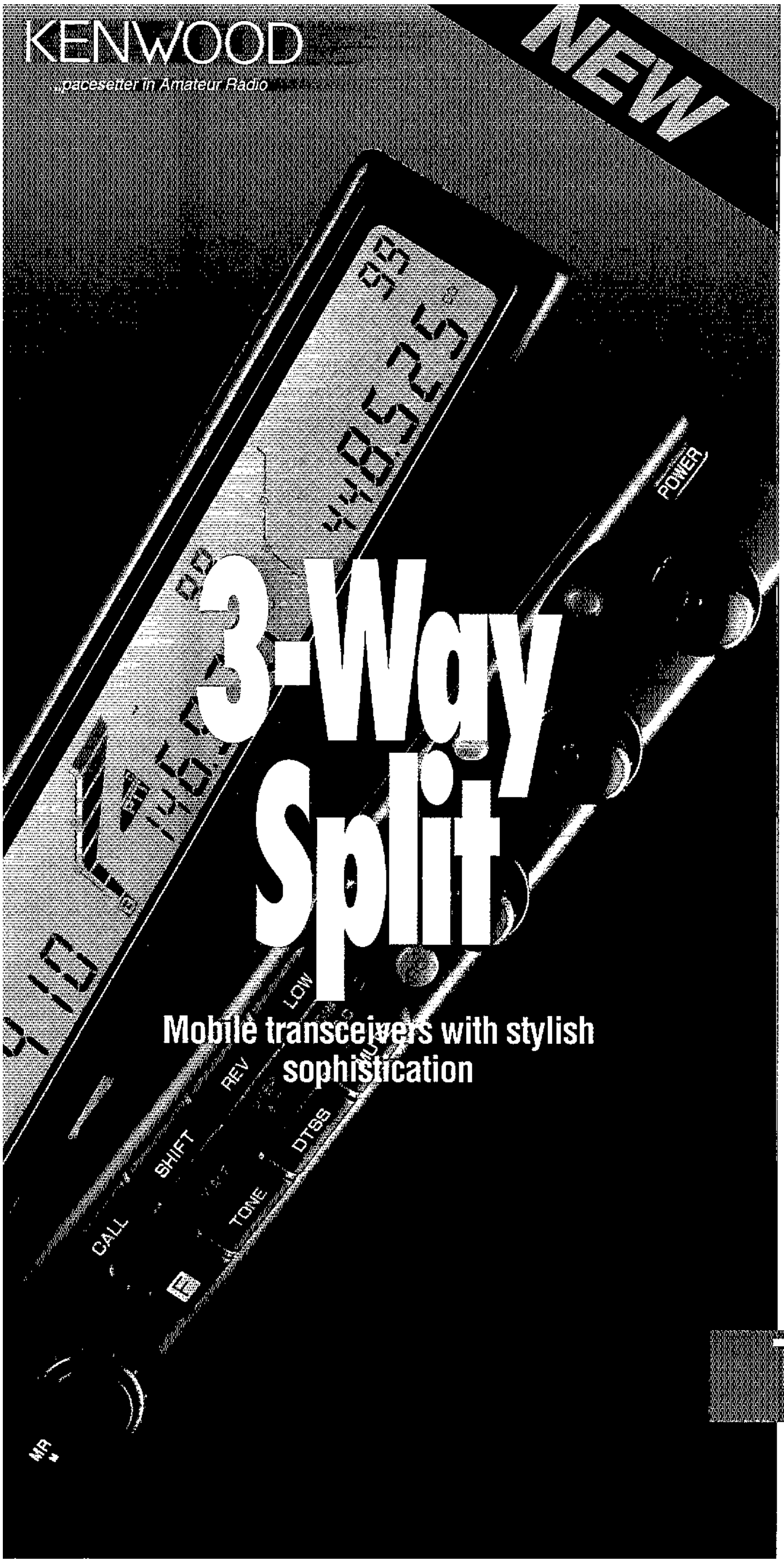
QST

June 1993 $3.00

devoted entirely to Amateur Radio

Roving: VHF/UHF Contesting With A View

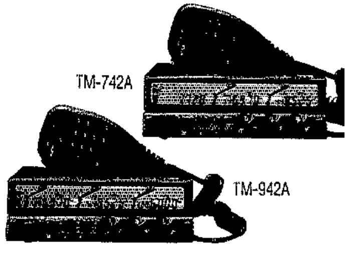




Kenwood's new TM-742A (144MHz/440MHz) and TM-942A (144MHz/440MHz/1200MHz) FM multibanders offer prime performance plus unparalleled freedom of choice for installation (optional kit).

- **High power**
  Maximum RF output is 50 watts (144MHz), 35 watts (440MHz), and 10 watts (1200MHz).
- **Wideband receiver coverage**
  The TM-742A receives from 118 to 174MHz and 410 to 470MHz; transmit ranges are 144–148MHz and 438–450MHz. The TM-942A adds the 1240–1300MHz range.
- **New & improved detachable front panel**
  The display and control sections can be separated for 3-way convenience. (with DFK-3,4,7)
- **100 memory channels**
  The 100 multi-function memory channels (all available for split operation) can be grouped into 5 banks for added convenience.
- **Multiple scan modes**
  Choose from 8 scan modes per band, plus CO (carrier-operated) and TO (time-operated) scan stops.
- **Tri-band receive/display**
  The TM-942A can receive/display all three bands (144MHz/440MHz/1200MHz) simultaneously. For the TM-742A there are four optional band units: 28MHz (50 watts), 50MHz (50 watts), 220MHz (25 watts), and 1200MHz (10 watts).
- **Cross-band repeater, dual-in repeater, fixed-band repeater**
- **Single-bander simplicity**
  Independent SQL and VOL controls for each band enable rapid response.
- **S meter squelch and auto squelch**
  Weak signals can be shut out. Noise squelch is also available.
- **Supplied multi-function microphone**
  Enables direct frequency entry.
- **Clock and timer**
  Includes stopwatch, alarm and on/off timer functions.
- **Wireless remote control function**
  A DTMF transceiver can be used to control various settings on the TM-742A/942A.
- **Built-in DTSS and pager function**
  The TM-742A/942A offers DTSS (Dual-Tone Squelch System) for selective calling and paging using standard DTMF tones. Elapsed time is shown by the tone alert system.
- **Supplied accessories**
  Mounting bracket, DC cable, fuses, mic, mic hook.
- **Choice of accessories**
  A full line of mics, speakers, and other accessories is available. See your authorized Kenwood Amateur Radio dealer for details!

Specifications guaranteed for Amateur band use only.



TM-742A/942A
Mobile Transceivers

KENWOOD COMMUNICATIONS CORPORAT
AMATEUR RADIO PRODUCTS GROUP
P.O. BOX 22745, 2201 E. Dominguez Street
Long Beach, CA 90801-5745

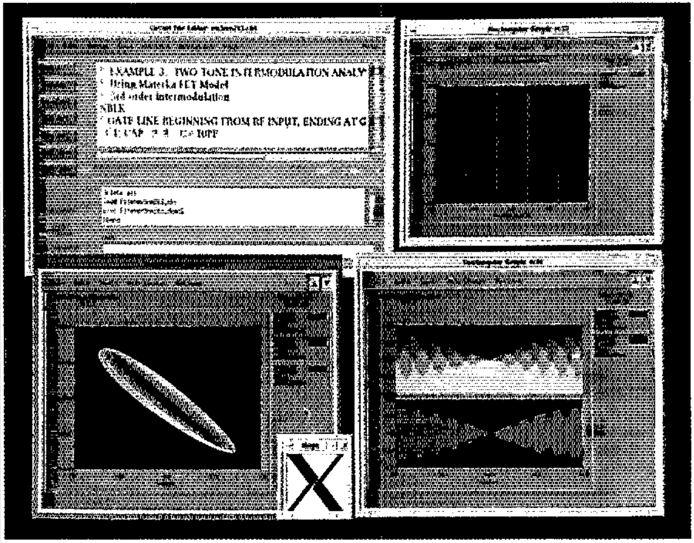
KENWOOD ELECTRONICS CANADA INC.
6070 Kestrel Road, Mississauga,
Ontario, Canada L5T 1S8

93-ARD-0593

A Comparison of Solid-State and Tube-Based Receiver Systems Using CAD

Do today's radios perform better because they use better devices, better design, or both? This study blends computer-aided-design analysis with lab measurements to yield some surprising answers.

By Ulrich L. Rohde,
KA2WEU/DJ2LR/HB9AWE
52 Hillcrest Dr
Upper Saddle River, NJ 07458



The rapid advances in the development of semiconductor devices have allowed us to develop transistors which will work well into the millimeter-wave area. Since 1960, when tubes like the 6CW4 and 417A were the dominant devices for building VHF and UHF receivers, the improvement of noise figure and sensitivity has been dramatic. Yet, controversy still rages in some circles as to whether tubes or transistors can build better radio receivers. Many people have wondered whether or not receivers' actual dynamic range, which is the ratio of the maximum input level (close to the 1-dB compression point) divided by the noise floor, has really been improved.

In the "old days," it was necessary to build all circuits prior to being able to evaluate or simulate them. Today's technologies allow us to do feasibility studies of circuit performance without ever lifting a soldering iron or measuring a noise floor. This article, a nonmathematical adaptation of one recently published in *QEX*,[1] reports the results I obtained in one such study: an evaluation of several top-flight amateur VHF converter designs of the period circa 30 years ago. In that study, I:

1. Obtained high-quality, functional samples of the converters to be tested.
2. Evaluated their theoretical performance limits using computer-aided-design (CAD) tools.
3. Measured the converters' *actual* performance and compared the results with that predicted by the CAD modeling.
4. Reiterating steps 2 and 3, replaced the tubes with, in turn, bipolar transistors, MOS transistors, and GaAsFETs.

The purpose of this study was to determine the advantages, if any, of modern solid-state design over tube design. Would CAD modeling reveal that yesterday's devices could have worked better if they had been more carefully applied? Could today's devices, applied carefully through CAD, improve the performance of yesterday's circuits to a modern level?

Title photo: Screen shot of the output of a modern microwave simulator (Compact Software's Microwave Harmonica) on which linear and nonlinear circuits can be simulated simultaneously. It can show the output spectrum as on the spectrum analyzer (upper right), the beat note for the two tones at the output (lower right), and even the increased level of voltage and current at different points of a three-stage distributed amplifier (lower left). Because of the beat note phenomena, the output load line shows a wider band than one would obtain from a single tone. *(photo courtesy of the author)*

What CAD Must Model to Be Useful

As preamplifiers and mixers form the basis for the converters evaluated and therefore largely determine their overall performance, we must be able to model (that is, predict):

1. The noise figure of low-noise amplifiers.
2. The third-order intercept point of amplifiers.
3. The insertion gain or loss of mixers, including noise figure.
4. The phase noise of oscillators.

Modeling Noise

Every resistor generates noise, and the noise generated by a two-port device—a device with an input and an output—was traditionally thought of largely in terms of its *equivalent noise resistance* (R_n—the value, in ohms, of the resistor that would generate the same noise power). This approach was limited because input-port thermal noise is not the only noise source

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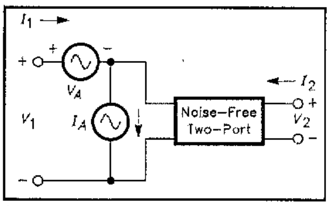
[1] Notes appear on page 28.

Fig 1—Noisy two-port representation in chain-matrix form. The model includes *two* noise sources: a source of noise voltage (V_A) and a source of noise current (I_A). (The device itself is shown graphically as a noise-*free* two-port for the purpose of this representation.) The noise source V_A represents all of the device noise referred to the input when the input is short-circuited. The noise source I_A represents all of the device noise referred to the input when the input is open-circuited. Real-world two-ports operate somewhere between these extremes—at a point affected by input tuning and matching in addition to device characteristics and operating conditions. Variations in these factors cause the two noise sources to contribute noise in varying degrees. Modern CAD techniques can take these factors into account to accurately predict noise performance.



present in a two-port device. Fig 1 shows the equivalent noise circuit of a linear two-port device in chain-matrix form.

The general set of noise-prediction equations for this linear two-port is device independent, but solutions for actual devices include coefficients that reflect the device parameters. Over the last few years, these noise equations have been expressed in *z* parameters, *y* parameters and, today, *S* parameters. Initially, circa 1960, *z* (impedance) parameters were used—modeling based on "open-circuit" models. The accuracy and usefulness of this modeling approach diminishes with frequency because

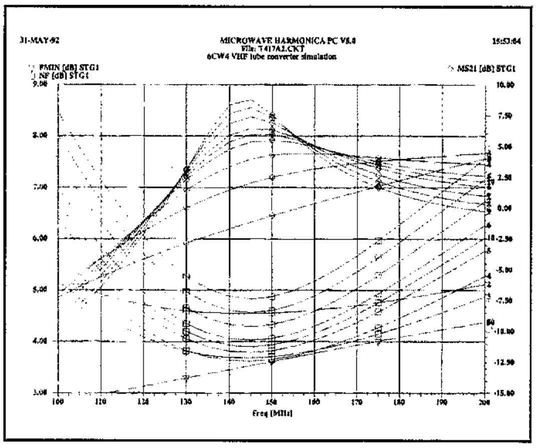


Fig 2—Simulated gain (the scattering parameter S_{21}, denoted on the plot as MS21), noise figure (NF) and minimum noise figure (F_{min}) for the 6CW4 triode VHF tube. Due to the significant feedback without neutralization, there is considerable interaction of noise figure and gain as the input matching changes. Note that minimum noise figure, F_{min} is, according to its definition, matching-independent.

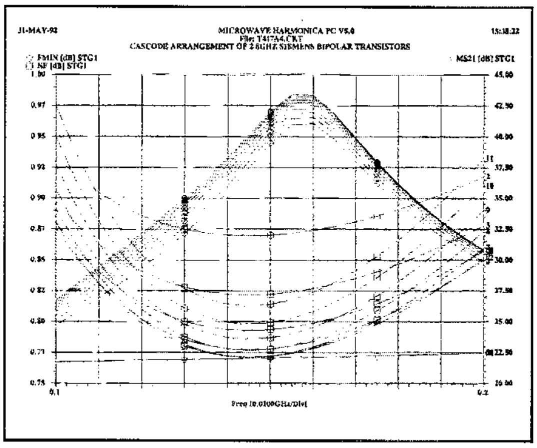


Fig 3—Simulated minimum noise figure, noise figure and gain for the bipolar cascode stage. Note that percent changes of the noise figure are very small because of the high reverse isolation which is common to all cascode arrangements.

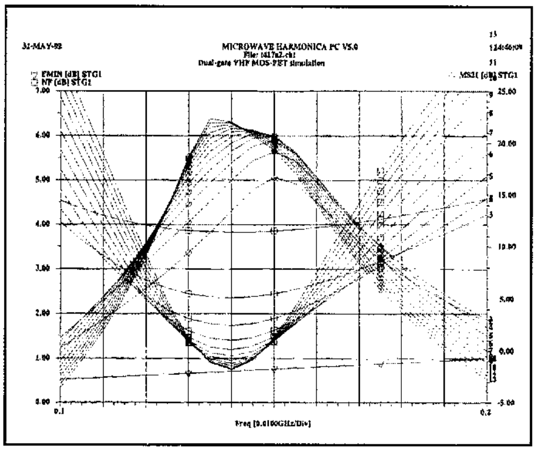


Fig 4—Simulated minimum noise figure, noise figure and gain of a dual-gate MOSFET stage. This arrangement is much more sensitive to changes in input matching, which equally effects noise figure and gain.

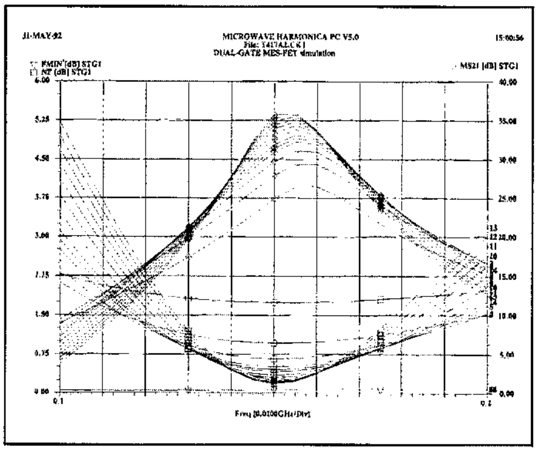


Fig 5—Simulated minimum noise figure, noise figure and gain as a function of input matching for a dual-gate MESFET (a form of GaAsFET). These devices are much less sensitive to matching changes than MOSFETs.

stray capacitance makes truly "open" circuits impossible in the real world.

From 1965 to 1970, the so-called *y* (admittance) parameters, which required a short circuit at the output, were considered. This approach also fails at high frequencies, where real-world parasitic inductances ensure that true short circuits cannot exist.

The *z*- and *y*-parameter approaches assume that the network under analysis is driven by voltages and currents. *S* (scattering) parameter analysis assumes that the network under test is driven by *voltage waves*, a network's behavior being evaluated in terms of how it reflects and scatters the wave energy. Modern transmission lines implemented with microstrip technology and precision 50-Ω terminations allow *S* parameters to work up to 100 GHz. A detailed mathematical derivation of this can be found in *Microwave Circuit Design Using Linear and Nonlinear Techniques*.[2]

The modern CAD programs, like *SuperCompact* and *Microwave Harmonica*, work from the noise parameter values F_{min} (minimum noise figure), Γ_{opt} (optimum source reflection coefficient for best noise figure[F_{min}]), and R_n (equivalent noise resistance) actually measured at one frequency (such as 10 GHz) to predict noise performance over a wider frequency range. This is done by solving a set of linear equations and expanding the frequency range from a few hundred megahertz to an upper cutoff frequency determined by the validity of equivalent circuits of the device.

Modeling reveals that *the optimum condition for minimum noise figure is not a conjugate power match at the input port.* We can explain this by recognizing the different noise sources present as the *two-port* noise, not just the thermal noise of the input port.

Noise in the frequency range above 30 MHz is due to effects either from thermal contribution or Schottky noise (in electron tubes, output-current variation attributable to random cathode emission). The so-called equivalent noise resistance determines the minimum noise figure at low frequencies, but above those at which flicker (1/f) noise or other surface-related noises predominate. The lower the R_n, the lower the minimum noise figure. At higher frequencies, parasitic reactances begin to contribute; in particular, input and feedback capacitance play a role. The feedback and input reactance determines the noise correlation between the Johnson noise (thermal noise) and the Schottky noise.

Analysis suggests that two immediate deductions are possible:

1. Devices with higher gain or higher transductance have lower noise. Also, devices with smaller input capacitance have lower noise.
2. Any parasitic reactances—either capacitance or inductance—which cause unwanted feedback will change the noise.

As a result of this, one can compensate the noise at particular frequencies by applying "noise" feedback. This results in almost simultaneous matching between the operation points for best gain and best noise operation.

Optimizing for low-noise operation assumes an operating point of linear performance, but does not address the issue of large-signal-handling capability. An oscillator or mixer is a hybrid in the sense that an active device is being "pumped" within a large alternating current or voltage swing and, as an effect of this, some of its nonlinear parameters change as a function of drive level. In the case of a mixer, an external oscillator "pumps" and/or switches a nonlinear device or devices between two states (*on* and *off*), and a mixing between several tones occurs. This type of mixing is referred to as *linear mixing* because the basic goal is that the active device is either on or off. Things not being ideal, there is still interaction during both the *on* and *off* conditions and, in particular, during the periods in which the device crosses from one state to the other.

In the *on* mode, modulations due to a change in currents can occur (specifically, modulation of transconductance [G_m]). In the *off* condition, due to large voltage swings, dynamic capacitance varies significantly. In the case of an oscillator, an active device is operated in the region of negative resistance and will start to oscillate into a resonant circuit which determines the resonant frequency. The purity of its signal depends on the flicker noise contribution, specifically due to AM-to-PM conversion.[3]

Converter Evaluations

As promised, we will take a step back into history and look at tube converters. Table 1 lists the small- and large-signal performance of a popular 417A-tube-based converter and 6CW4-tube-based Ameco design used at 144 and 220 MHz. The table evaluates and compares their measured performance with predictions made using Compact Software's *Microwave Harmonica* program for simulation. Table 1 also shows the results obtained by replacing the Ameco's 6CW4s with bipolar transistors, then MOS transistors, and finally GaAsFETs.

Figs 2 through 4 assist in evaluating this simulation by showing the predicted minimum noise figure for the Ameco circuit, first with 6CW4 tubes (Fig 2), and then with cascode bipolar junction transistors (Fig 3) and MOSFET (Fig 4) replacing the 6CW4s, as a function of neutralization setting.

In the case of the 6CW4 (Fig 2), changes in neutralization make dramatic differences in noise matching and also affect stage gain.

By using a cascode arrangement of Siemens bipolar transistors (Fig 3), I obtained the best noise figure thus far (approximately 0.76 dB), along with the highest gain and the least feedback. This is directly attributable to the cascode circuit's high reverse isolation. Subsituting a dual-gate MOSFET (Fig 4) for the bipolar cascode gives about the same noise figure, but the sharper curves indicate that selectivity is vastly improved with the MOSFETs. This means that more interference suppression is possible for the same noise figure. Finally, if we look at the GaAsFET (Fig 5), the minimum noise figure available there is approximately 0.2 dB, but this is not quite achieved in the practical circuit because of input matching losses and lack of input selectivity.

It is also useful to compare the best tube converter (Fig 6, the 417A) with the best

Table 1
Converter Test Data

Ameco (6CW4)		*417A*		*Bipolar-Cascode*		*MOSFET*		*GaAsFET*	
Measured	*Calculated*	*Measured*	*Calculated*	*Measured*	*Calculated*	*Measured*	*Calculated*	*Measured*	*Calculated*
$P_G = 20$ dB	21 dB	$P_G = 20$ dB	21 dB	$P_G = 20$ dB	22 dB	$P_G = 20$ dB	21 dB	$P_G = 20$ dB	21 dB
NF = 4 dB	3.8 dB	NF = 1.6 dB	1.7 dB	NF = 0.8 dB	0.79 dB	NF = 0.7 dB	0.65 dB	NF = 0.4 dB	0.42 dB
$IP_3 = 0$ dBm	1 dBm	$IP_3 = 5$ dBm	5.5 dBm	$IP_3 = 7$ dBm	6.5 dBm	$IP_3 = 10$ dBm	11 dBm	$IP_3 = 10$ dBm	10.5 dBm
Tube Mixer		*Tube Mixer*		*Diode Ring*, +20 dBm		*Diode Ring*, +20 dBm		*Diode Ring*, +20 dBm	
Q = −4	Q = −2.8	Q = 3.4	Q = 3.8	Q = 6.2	Q = 5.71	Q = 9.3	Q = 10.35	Q = 9.6	Q = 10.08

P_G = power gain; 50-Ω input to 50-Ω output; NF = noise figure; IP_3 = third-order input intercept point; Q = figure of merit based on IP_3–NF.

In the case of the bipolar and FET versions, a high-level double balanced mixer was used. Indications are that the limitations in these designs arise from two factors: (1) the IP_3 of the double balanced mixer and (2) the gain distribution of the system, including the matching. For reasons of stability, MOSFETs can be handled better than GaAsFETs at VHF.

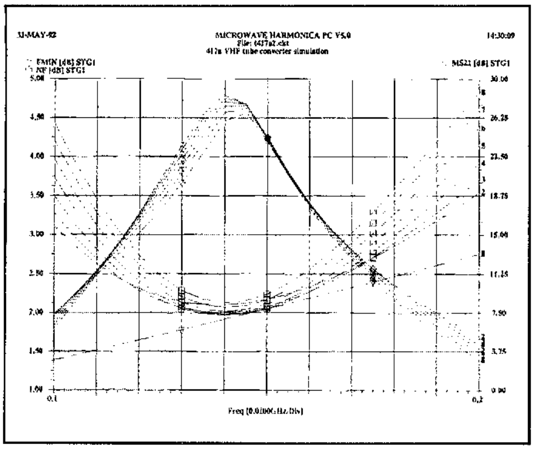


Fig 6—Simulated minimum noise figure, noise figure and gain as a function of input matching for the 417A tube.

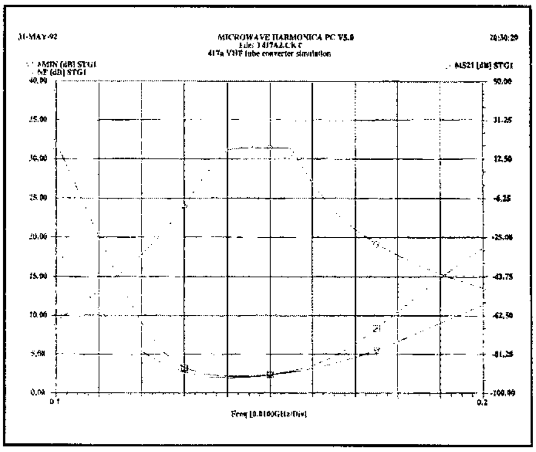


Fig 7—Overall noise figure and gain of the 417A converter. Note that the minimum and predicted noise figures agree quite well.

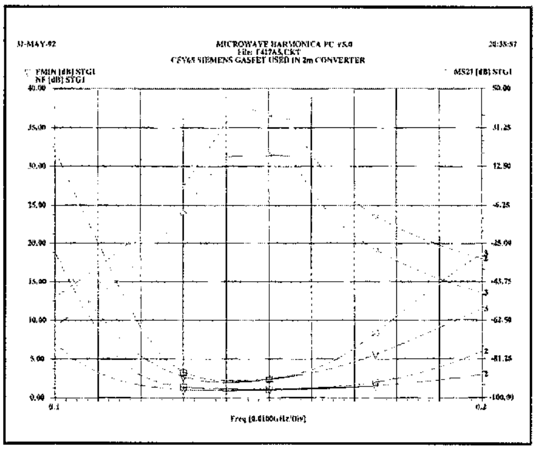


Fig 8—Overall noise figure and gain of the bipolar and GaAsFET versions of the 6CW4 converter. Due to some feedback phenomena, the bipolar version (upper MS21 curve) does not show the same flat performance as the GaAsFET.

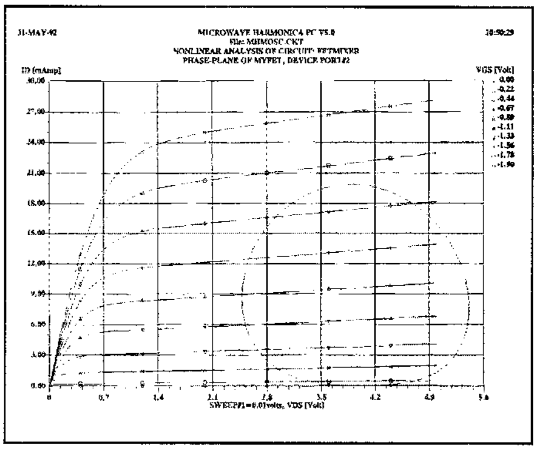


Fig 9—Dynamic load lines of a tube mixer simulated with a modified FET model. The plum-shaped curve is due to the LC tuned circuit at the mixer output.

GaAsFET converter. By substituting and replacing the tubes with GaAsFETs, an overall noise performance is achieved with a system noise figure of approximately 1 dB, versus 1.6 dB with the tubes as shown in Figs 7 and 8.

Simulating Mixer Performance

Since each of these converters typically consisted of a preamplifier and single-ended tube mixer, we will look into the nonlinearities of the mixer. Fig 9 shows the phase plan of the tube mixer, modeled by means of a modified FET model, as seen at its output. If you were to connect a spectrum analyzer to the output of this mixer, you would see a picture like that in Fig 10. Fig 11 shows a simulated output of the high-level diode mixer used in obtaining the bipolar- and FET-preamplifier-based data in Table 1. Its spurious output is considerably lower compared to that of the simulated single-ended tube mixer.

Conclusion

My study suggests that the weakest point in widespread radio design techniques of decades past was the understanding of how to design good *mixers*. It was possible, for instance, to build an MF/HF vacuum-tube receiver with a 30-dBm input intercept 30 years ago, but such examples are rare.

The introduction of the 417A vacuum tube made possible the achievement of good low-noise *amplifier* stages up to 150 MHz; achieving this performance in the microwave region, however, requires modern GaAsFETs.

In the amplifier stages of the converters evaluated, design techniques limited performance more than the devices available. The major differences between tube design and bipolar transistor/GaAsFET design lie in the current matching techniques between the stages and compromises among selectivity, noise figure and losses.

That measurements and computer simulations on these circuits can agree quite well indicates that modern CAD tools are quite capable of making good predictions.

Acknowledgments

I thank all the radio amateurs who supplied me with converters, either gratis or at

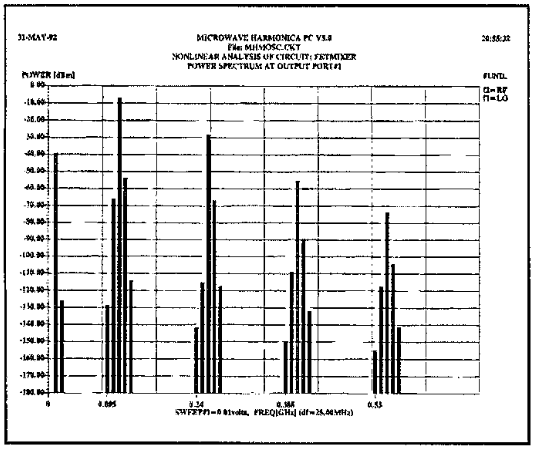


Fig 10—Power spectrum at the output of the Fig 9 mixer with –8 dBm of LO and 10 mV of RF input applied.

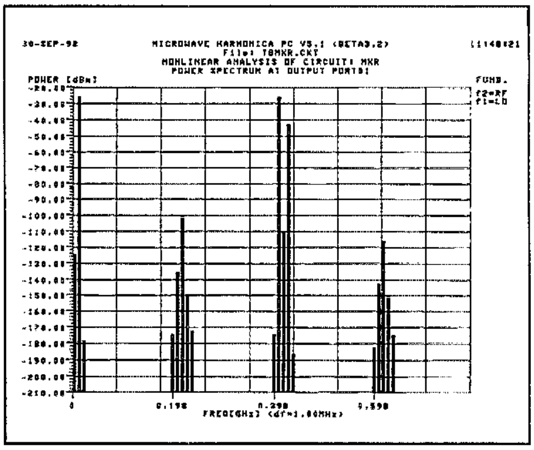


Fig 11—Simulated output of the high-level double-balanced mixer used in the converter characterized by the Bipolar-Cascode, MOSFET and GaAsFET listings in Table 1. This mixer, which uses two diode rings for better harmonic cancellation than the usual single-ring diode DBM, affords significantly better spurious-signal performance than the tube mixer modeled in Figs 9 and 10.

the advertised purchase price, for this project. They are: John Abbruscato, KC5GB; Nelson Dew, K3RGH; Martin Feeney, K1OYB; David Knepper, W3BJZ; Charles Lustick, K3HSS; Jacob Makhinson, N6NWP; John Pivnichny, N2DCH; Earl Shinn, K5KAC.

Additionally, I thank those radio amateurs who responded to my project, but whose converters I had to refuse due to the overwhelming response.

Notes

[1] U. Rohde, "Differences Between Tube-Based and Solid State Based Receiver Systems and Their Evaluation Using CAD," *QEX*, Apr 1993, pp 3-13. This paper includes equations for device noise parameters and schematics of the converters evaluated, as well as those of a 1960s-vintage vacuum-tube MF/HF receiver front end (input intercept, +30 dBm) and a VHF-relay receiver front end with a parallel-MOSFET preamp and diode double-balanced mixer (noise figure, 2 dB; input intercept, +23 dBm). Copies of April 1993 *QEX* are available from ARRL for $1.75 each, postpaid.

[2] G. Vendelin, A. Pavlo, U. Rohde, *Microwave Circuit Design Using Linear and Nonlinear Techniques* (New York: John Wiley & Sons, 1990).

[3] AM-to-PM conversion, a change in output phase angle in accordance with changes in input-signal amplitude, occurs in amplifiers as a direct result of gain saturation. An oscillator is a special form of amplifier.

Recommended Reading

U. Rohde and T. Bucher, *Communications Receivers Principles & Design* (New York: McGraw-Hill, 1987).

New Books

WEATHER RADIO

By Anthony "Tony" Curtis, K3RXK

Published by Tiare Publications, PO Box 493, Lake Geneva, WI 53147, tel 414-248-4845. First edition, 1992. Sofecover, 8½ × 11 inches, 125 pp, $14.95 plus $2 shipping.

VIEW FROM LOW ORBIT

By Thomas S. Lobel, WA1VTA

Published by Imaging Publications, PO Box 66, Hubbardston, MA 01452. First edition, 1992. Softcover, 8½ × 11 inches, 124 pp, black-and-white, $16.95.

Reviewed by Steve Ford, WB8IMY, ARRL Assistant Technical Editor

Few things are more fascinating—and unpredictable—than weather. Our atmosphere is a churning cauldron of clouds, precipitation and wind currents. This chaotic activity can be violent and dangerous, or serene and beautiful. No wonder we're preoccupied with weather—it dominates every facet of our daily lives!

Since the early '60s, weather satellites have been an invaluable meteorological tool. From their orbital vantage point, they allow us to monitor the progress of developing storms and other weather patterns. As satellite technology has advanced, images have become clearer and more detailed. You don't have to be a meteorologist to appreciate the surrealistic beauty provided by the present generation of weather satellites.

View from Low Orbit offers a grand tour of high-resolution imagery from low-earth-orbiting (LEO) satellites. The detailed, black-and-white photographic images were captured during passes over North America, and each includes a caption explaining the image and the weather phenomena illustrated. Special sections are devoted to the violent midwest thunderstorms of 1991 and the spectacular hurricanes of 1990.

If you're like me, you'll want to see the photos that show your particular corner of the country. ("Hey! I can almost see your house from here!") This book is a great conversation-starter. Leave it out on the coffee table for your nonham guests.

Weather Radio spends most of its time on terra firma, delving into the nuts and bolts of weather-related communications. It's a reference book containing everything from the history of weather forecasting (did you know that Antarctica is one of the driest places on earth?) to a long list of weather radio frequencies—including Amateur Radio weather nets. Satellites, WEFAX and NAVTEX are discussed in detail. The book features a complete index, making it easy to locate specific information.

If you enjoy the beauty of weather satellite photos, *View from Low Orbit* is worthwhile. If you're involved in amateur weather forecasting or weather-related emergency operations, *Weather Radio* is a useful reference.

Home-Brewing a 10-GHz SSB/CW Transverter

Part 2—Designed to work with last month's 10.224-GHz local oscillator, this month's mixer, power amplifier and preamplifier round out your narrowband 10-GHz transverter.

By Zack Lau, KH6CP
ARRL Laboratory Engineer

In Part 1,[15] I described a 10.224-GHz local oscillator (LO) designed to drive a dual-mixer board like those used in the no-tune transverters. If you've completed the modules described last month, you should have a working 10.224-GHz LO. The mixer board described this month contains a two-way etched power splitter that delivers equal LO signals to the transmit and receive mixers, which are also etched on the same PC board. On transmit, one of the mixers combines a 144-MHz IF signal with the LO to generate a 10.368-GHz signal; on receive, the other mixer combines the incoming 10.368-GHz signal with the LO to produce a 144-MHz IF output. An external pipe-cap filter (described last month) in each 10-GHz mixer line eliminates the image, passing only the desired signal. Two-stage GaAsFET amplifiers of the same RF design, but using different bias settings, serve as a 10-GHz preamplifier and power amplifier.

Part 1 also shows the transverter block diagram, and covers construction techniques and etching-pattern availability for the transverter's circuit boards, component sources, and performance data for the finished transverter. I suggest that you review that before going ahead with construction of this month's modules.

Mixer Construction and Tweaking

If you're building transverters from surplus hardware, the most difficult module to obtain is not the LO, but the mixer. Builders have gotten widely varying results, even when copying the same design. For most people, 10 GHz is just too high a frequency to accurately build a no-tune mixer that works well. The difficulty is that a full-wavelength microstrip transmission line is only 0.6 inch long at 10 GHz. So, a typical rat-race mixer (which requires signals to be 180° out of phase for proper cancellation) really needs to built with tolerances under 0.005 inch (5 mils).

This problem has several solutions. One is to simply accept the inferior performance. Usually, the conversion loss isn't too bad if you copy a known-good layout, but the LO rejection relative to the PEP output signal can be as little as 10 dB. For receive purposes, LO rejection really doesn't make much difference.

A better solution is to tune the mixer. Once you've etched and assembled the mixer board, terminate all ports in 50-Ω loads or sources. You don't want to look at the mixer though an image-reject filter, unless it is properly tuned. Otherwise, the mixer and filter tuning will interact, making it difficult to adjust the mixer for proper operation. I normally connect the mixer to the LO, attach a 0- to -10-dBm, 50-Ω 144-MHz source at the IF port, and a spectrum analyzer at the RF port.

I usually adjust the LO rejection first. This is done by placing a small piece of copper foil at point A or B indicated in Fig 11—at either side of the junction between the 70-Ω ring and the LO-input line. This shortens the transmission line slightly on one side. Usually, the LO suppression improves with the copper at one point and worsens with it at the other point. True, the copper foil mismatches the amplitude slightly, but this is better than having an improper phase shift. Usually, LO rejection is 17 or 18 dB below the saturated output (this equates to the specification-sheet figure of 27 or 28 dB of LO-to-RF port isolation). Keep in mind that even a lid covered with absorptive rubber or foam affects the tuning slightly. You don't want to tune the mixer to perfection only to have to retune it after installing a cover.

I find that the obtainable LO rejection depends on how well I made the board. Mixer rings that look almost perfect often allow 5 or 10 dB better rejection; ones that look as if they were drawn quickly with a crayon may be almost impossible to tune (though they often work just fine for receive).

Finally, tune the mixer's RF port for maximum output into a 50-Ω load (as described in Part 1 under "Adjusting the 2.556- to 10.224-GHz Multiplier"). I've been unable to etch mixers consistently, so all of my mixers are a little different.

Three short wires, 0.21 inch of #28 enameled wire, serve as 10-GHz RF chokes and 144-MHz shunts at the mixer board's RF and LO inputs (Fig 11).[16] This improves the isolation between the mixer's IF ports. Without them, there is little to stop a 2-meter signal from crossing the power divider. Adding these wires increases the isolation between the IF ports from an almost negligible 4 dB to a decent 40 dB.

It shouldn't be necessary to tune the load termination, though you may want to. As you might guess from the layout, I tacked on the radial stub to ground the 51-Ω chip resistor. Purists may want to use a 68-Ω resistor and tune out the reactance to get a really good 50-Ω load at 10 GHz (as is done in the TNT[17]).

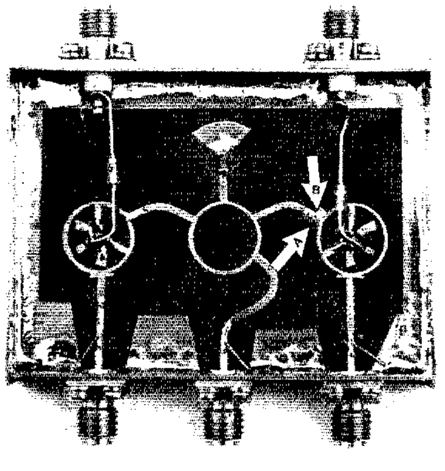


Fig 11—The splitter/mixer board. The IF ports (at top) are connected to the mixers via 0.01-µF encapsulated chip capacitors. Ceramic capacitors can be used instead. The mixers use Hewlett-Packard HSMS 8202 Ku-band diode pairs. The HP HSMS 2822 2-GHz diode pair can be substituted, but this device has more conversion loss at 10 GHz than the HSMS 8202. Mixer tuning for best LO rejection is done by adding a small piece of foil at point A or point B. *(photos by Kirk Kleinschmidt, NT0Z)*

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[15] Notes appear on page 31.

If you have them, you can also use 50-Ω microstrip terminations for this; I've gotten them from surplus isolators.

Power Amplifier and Preamplifier

The same RF design is used for the transmit and receive amplifiers (see Figs 12 and 13). Tripling the bias current from the 10 mA used in the receive-side amplifier to 30 mA in the transmit amplifier increases the circuit's 1-dB compression point from 5.7 to 10.8 dBm. Gain increases from 18 to 19 dB and the noise figure rises by about 1 dB. Computer-modeling results indicate that the Rollett stability factor, K, drops a little, but since it's still above 3 (a K greater than one denotes a stable design), this shouldn't be a problem—even if the amplifier is terminated at the input and output with a high mismatch (such as sharp filters).

It may be possible to get a bit more output by increasing V_{ds} to slightly more than the 3 volts I used, but this would require redesigning the bias circuit (Fig 5). Like many transistor amplifiers, this amplifier's saturated output, typically 14 dBm, is more than twice the recommended output for linear operation.

System Integration

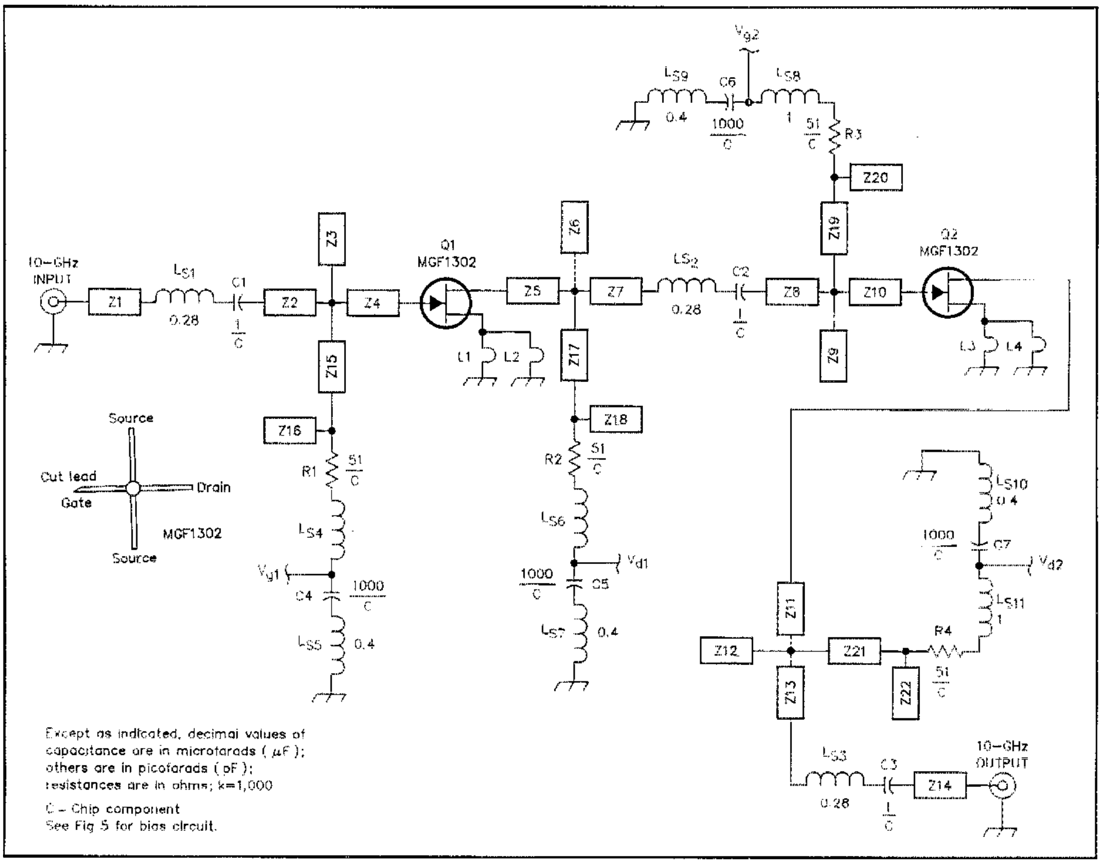
To complete the transverter, build two band-pass filters as shown in Part 1. You can tune them with the aid of the diode detector described in the sidebar, but a few minutes with a spectrum analyzer makes the process easier.[18] Then, following the block diagram of Fig 1, assemble the transverter's blocks. Connect a suitable IF radio, check to make sure the transmit converter and receive converter operate, and you're on the air!

Antenna Thoughts

Most people looking for a high-gain antenna end up with some sort of parabolic reflector. If you put a low-gain horn antenna in the right spot in front of a reflector that is anything close to a parabola, it will probably outperform anything of similar dimensions on this band. People have used everything from metal snow sleds to trash-can covers, in addition to more obvious choices such as light collectors and surplus military/commercial gear. Commercial sources for new dishes exist, but even small dishes are expensive when purchased new. Look for a surplus reflector.

Perhaps the simplest antenna I've seen is a quarter-wave monopole—with a piece of sheet metal as the ground plane! The most complicated is undoubtedly a loop Yagi—it works, but it is more of a curiosity than a practical way of getting 18 to 20 dBi of gain. A horn is much easier to make. *The ARRL Antenna Book*, *The ARRL UHF/Microwave Experimenter's Manual*, the RSGB *Microwave Handbook, Volume 3* and various VHF/UHF/microwave conference proceedings contain duplicable designs. Chapter 18 of the RSGB *Microwave Handbook, Volume 3*, contains all the information you need to get started.

I have yet to adjust one of my 10-GHz antennas with an SWR meter, yet I've made lots of 10-GHz contacts of more than 200 km. Usually, if I do any tweaking at all, such as adjusting the location of a dish feed, it



Except as indicated, decimal values of capacitance are in microfarads (µF); others are in picofarads (pF); resistances are in ohms; k=1,000

C – Chip component
See Fig 5 for bias circuit.

Fig 12—10-GHz low-noise amplifier. Chip components are used unless specified otherwise. L1-L4 are source-lead inductances. L_{S1}-L_{S11} are stray inductances (in nanohenries). Z1-Z22 are etched on the circuit board.

C1-C3—1 pF. Use high-quality, 50-mil ceramic chip capacitors such as ATC 100As.

Q1, Q2—MGF1302. Substitution not recommended. Set bias at 10 mA and V_{ds} = 3.0 V for low-noise receive preamplifier operation. For the transmit amplifier, set bias at 30 mA and V_{ds} = 3 V. For additional biasing information, see the text and Fig 5.

10-GHz Power Measurement

Measuring RF power at 10 GHz presents a challenge; calibrated measurement devices can be very expensive. Fortunately, measuring *relative* power requires only a diode detector and a sensitive dc voltmeter. The 10-GHz power measurements required to optimize this transverter needn't be absolute; relative power measurement is acceptable. A convenient way to measure power in a 50-Ω system is to couple some RF into a low-offset Schottky diode, such as a Hewlett-Packard 5082-2835 (commonly used as a microwave frequency multiplier), filter its dc output, and measure this voltage with a high-impedance voltmeter. This measurement approach gives useful output down to the milliwatt level. Of course, you can also use a commercial diode detector rated to 10 GHz.

To build a detector, etch or cut a 50-Ω microstripline on a small piece of Rogers 5880 RT/duroid with 1-oz copper cladding (the same material used in the transverter's 10-GHz circuits). See Fig A. A 50-Ω trace is 46 mils wide (0.046 inch) on this material. Terminate the microstrip in SMA connectors and enclose the board with brass strip for rigidity, like the transverter. Mount the diode and other components as shown in Fig A. Fig B shows the equivalent circuit. The length of the diode lead that runs along the 50-Ω stripline affects the amount of RF energy coupled into the diode, as does its spacing from the microstrip trace.

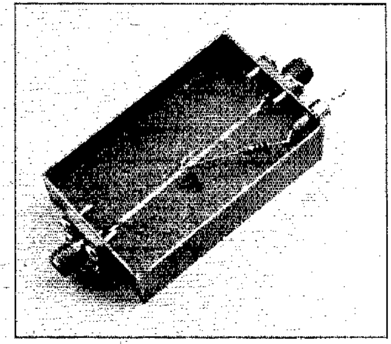


Fig A—This 2- x 1¼-inch diode detector gives useful dc output for 10-GHz power measurement down to about 1 mW. It uses a low-offset Schottky diode (such as the HP 5082-2835), with its anode lead soldered to the ground plane. Its cathode lead follows the 50-Ω microstrip trace for about ¼ inch and is spaced about ⅛ inch from the trace (neither dimension is critical; a longer lead and closer spacing increase coupling). A 1- to 10-kΩ resistor, also soldered to the cathode lead, routes rectified energy to a feedthrough capacitor.

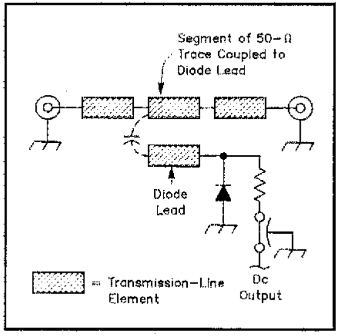


Fig B—Equivalent circuit of the diode detector.

This detector can be used for tuning the transverter's multiplier, filters and amplifiers. To use the detector, terminate one end in a 50-Ω load that's good to 10 GHz. (Alternatively, you can substitute a 50-Ω microstrip load for one of the SMA connectors.) Couple RF into the other port via a 3- to 10-dB attenuator, to ensure that the circuit under test is terminated with a stable 50-Ω load. Measure the voltage on the feed-through capacitor using a sensitive voltmeter or oscilloscope.—*Kent Britain, WA5VJB*

has been for maximum received signal. Similarly, I've adjusted my coax-to-waveguide transitions this way, adjusting tuning screws for minimum loss. Of course, even SWR is no indication of how well an antenna really works. The *real* test is to compare antennas and see which does best.

Summary

Although it takes some effort to build, the transverter described in this two-part article provides useful and exciting 10-GHz SSB and CW capability. Perhaps the best part is that you don't have to hunt through flea markets to find a surplus "brick" LO and filters, or deal with any of the other traditional hassles of getting on this fun band! What hilltop will *you* operate from in this year's ARRL 10-GHz Cumulative Contest in August and September?

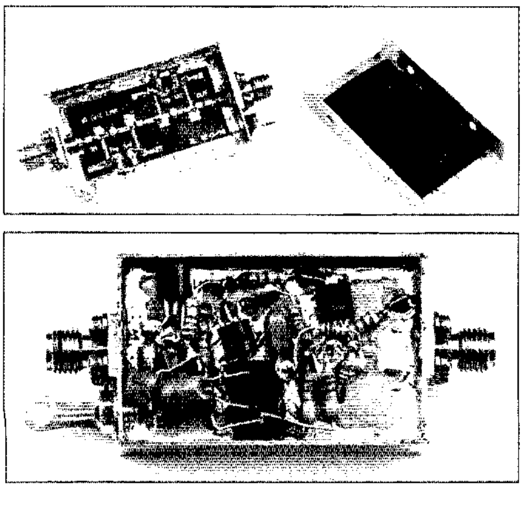


Fig 13—Top and bottom-side views of the completed 10-GHz two-stage amplifier. Microwave-absorptive foam is glued into the lid to suppress waveguide propagation modes inside the amplifier enclosure, which could provide enough feedback from output to input to cause oscillation. The amplifier's bias circuitry, like that of the 10.224-GHz multiplier (Figs 4, 5 and 10) is built on the bottom side of the amplifier board for convenience. External biasing is also acceptable and may be more convenient.

Notes

[15] Z. Lau, "Home-Brewing a 10-GHz SSB/CW Transverter," *QST*, May 1993, pp 21-28.

[16] Part-placement diagrams with component values and more detail for each of the transverter's modules are part of the template package obtainable from the ARRL Technical Department Secretary. See Note 11 for details.

[17] See Note 5.

[18] When tuning these filters, you can use the finished 10.224-GHz LO and a power meter (see the sidebar) to make sure that the filters aren't tuned to the LO or image frequency. To tune a filter, first connect it to the LO and adjust the tuning screw for maximum output at the LO frequency. Then adjust it for peak response at 10.368 GHz by connecting it to the transmit-amplifier output and backing the tuning screw *farther out of the filter cavity* while looking for maximum filter response. —*Ed.*

A Small Loop Antenna for 160 Meters

Effective, low-profile 160-meter antennas present a formidable challenge. Here's one that answers the need.

By Charles J. Mozzochi, W1LYQ
PO Box 1424
Princeton, NJ 08542

For amateurs with limited space, the loop antenna is ideal for 160 meters. Because loops have high Q, they exhibit very narrow bandwidth, suppress harmonics and give a significant receiving-noise reduction compared to dipoles. The loop's apparent simplicity is deceptive, however.

This article describes a practical loop antenna that has worked well for me on 160 meters for more than three years. The design is based primarily on information published by Ted Hart, W5QJR.[1,2]

Theoretical analyses of loop antennas appear in Kraus,[3] Stutzman and Thiele,[4] and Weeks.[5] These authors don't discuss efficiently constructing and coupling the loop to the transmitter, however, and these are not entirely trivial considerations. Practical and design considerations are discussed in David,[6] Hall and Schetgen,[7] Hart,[8] McCoy,[9] Morrow,[10] and Patterson.[11]

For distances of 700 to 1000 miles, my loop has performed interchangeably with my full-size inverted **V** antenna (62 feet high at the midpoint and fed with open-wire transmission line).

During several recent DX contacts with the loop, I've compared it to my inverted **V**. During a QSO with G3PQA, the strength of my signal from the loop was essentially the same as that from the inverted **V**. DL1RK found my signal from the loop slightly stronger than that from the inverted **V**. Both antennas are oriented to favor Europe. Also, the inverted **V** is more than 90% efficient, whereas the loop's efficiency is close to 50% (−3 dB).

Radiation Pattern

I quote from Hart (see "References"):

> The loop has one distinct advantage over conventional antennas due to its radiation pattern. If you envision the pattern to be the shape of a doughnut, and the doughnut is "standing" on the ground, the maximum gain occurs at both low and high angles. In fact, it radiates equally well at all elevation angles in the plane of the loop.

Table 1
Approximate Costs of Materials

Eight 10-foot × 1-inch sections of copper pipe	$50.
Eight 45° elbows	$10.
One PVC insulation section	$10.
One 15-foot, 6-inch × 5/8-inch copper inner loop	$15.
Support rope	$25.
Labor (plumber) to obtain, assemble and solder these components	$100.*
Total	$210.

The total does not include the costs of C1, C2, C3 and the feed line.

*With ingenuity, you may be able to obtain these parts and services for significantly less.

Fig 1, from Hart (see "References"), compares the vertical radiation patterns of the loop with those of a vertical and a dipole. The inner-pattern axis is perpendicular to that of the outer patterns. Note that the loop performs better at low angles, making it a good DX antenna. Fig 2 (also from Hart, "References") compares horizontal patterns at four elevation angles.

Physical Description

Fig 3 shows the antenna schematically and Fig 4 shows the base section of my antenna. The outer loop is made in the form of a hexagon with eight 10-foot sections of 1-inch copper pipe joined with 45° couplers. The inner loop is a circle, 15 feet, 6 inches in circumference, made from ⅝-inch copper pipe. Choosing values and ratings for C1, C2

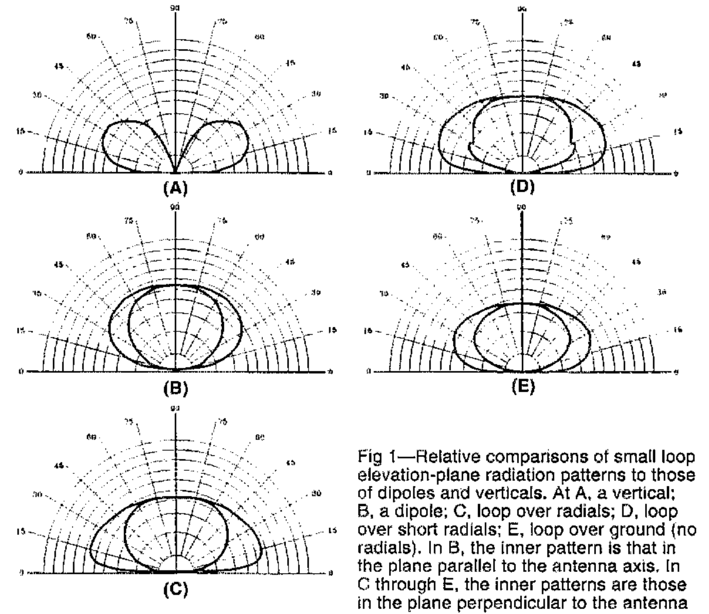


Fig 1—Relative comparisons of small loop elevation-plane radiation patterns to those of dipoles and verticals. At A, a vertical; B, a dipole; C, loop over radials; D, loop over short radials; E, loop over ground (no radials). In B, the inner pattern is that in the plane parallel to the antenna axis. In C through E, the inner patterns are those in the plane perpendicular to the antenna axis.

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[1] Notes appear on page 34.

Table 2

Calculating capacitor ratings

$V_{C1} = 1800\sqrt{P}$
$V_{C2} = 18\sqrt{P}$
$V_{C3} = 17\sqrt{P}$
$I_{C1} = 2.2\sqrt{P}$
$I_{C2} = 0.21\sqrt{P}$
$I_{C3} = 0.15\sqrt{P}$
Example: if $P = 100$ watts, then $V_{C1} = 18{,}000$; $V_{C2} = 180$; $V_{C3} = 170$; $I_{C1} = 22$ A; $I_{C2} = 2.1$ A; and $I_{C3} = 1.5$ A.

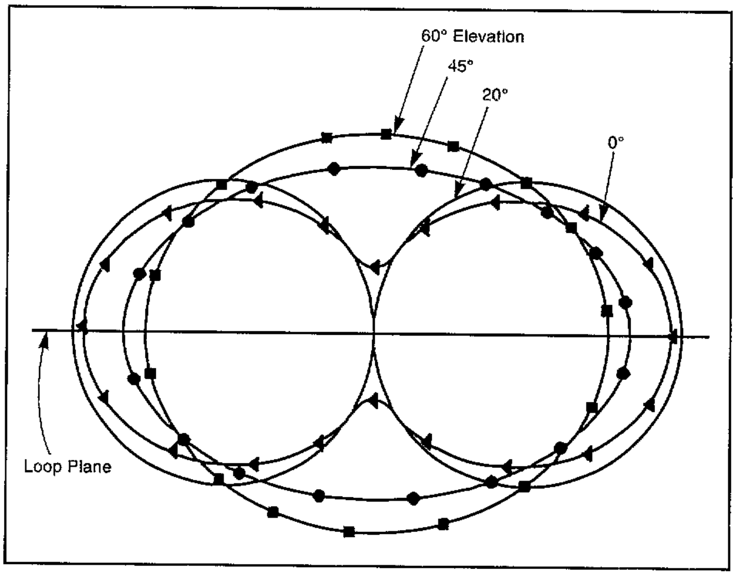

Fig 2—Electrically small loop antenna azimuth-plane radiation patterns at several wave angles.

and C3 is discussed later.

The antenna is mounted vertically, approximately a foot off the ground. I installed 12 quarter-wavelength radials under the loop. These radials are joined under the loop, but are not electrically connected to it. The radials act as a reflective screen to reduce ground loss. The use of radials is not necessary, but they do increase antenna efficiency and thus signal strength. As a compromise, the radials can be as short as twice the loop's height and should be parallel to the loop plane.

Construction

Table 1 lists the bill of materials for my loop. Because the outer loop's radiation resistance is on the order of 0.1 Ω, every effort must be made to minimize losses. Antenna efficiency increases with the diameter of the pipe used in both loops. One-inch copper pipe is used as a compromise in the outer loop; larger pipe would provide better efficiency, but would weigh considerably more.

I support the outer loop with nylon rope strung between two trees. Smaller nylon ropes guy the lower section between trees and a fence. The inner loop is hung from the top of the outer loop with nylon rope, and nylon rope is used to guy and position the inner loop.

Solder all of the loop's joints with rosin-core tin/lead or tin/lead/silver solder. Do not use acid-core solder. Note that McCoy's loop (see "References") was built with aluminum tubing that was bolted together, which accounts for the poor performance noted there.

Capacitors

C1 is a 200-pF vacuum-variable unit. A split-stator capacitor should work in place of a vacuum variable; however, under no circumstances should a conventional variable capacitor be used at C1 because the loss in the wiper contacts is significant compared to the outer loop's radiation resistance. C2 is composed of ten 100-pF "doorknob" capacitors connected in parallel via copper straps, and C3 is a single 740-pF doorknob.

The outer loop radiates and the inner loop functions as a low-loss matching and coupling device. This type of antenna puts very high voltages across C1 and C2 and passes unusually high currents through them. The formulas in Table 2 give approximate currents and voltages for C1, C2 and C3, where P is the applied power in watts.

Table 2's formulas are based on a steady-state analysis, under the assumption that the radiation resistance and the loss resistance in the outer loop are each 0.1 Ω. These resistances can differ significantly from this value depending on how the loop is constructed and located. Furthermore, substantial transient currents and voltages can occur under certain situations. Consequently, I recommend incorporating a safety factor of two to three after calculating the capacitor voltages and currents with these equations.

Tuning and Matching

To adjust the antenna to resonance at 1.85 MHz, follow this procedure:

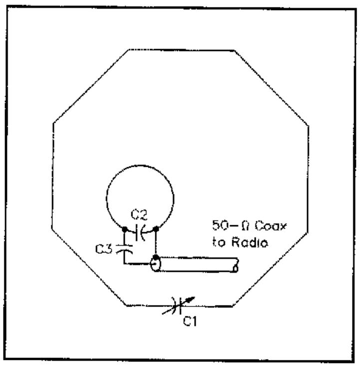

Fig 3—Schematic of the loop antenna.

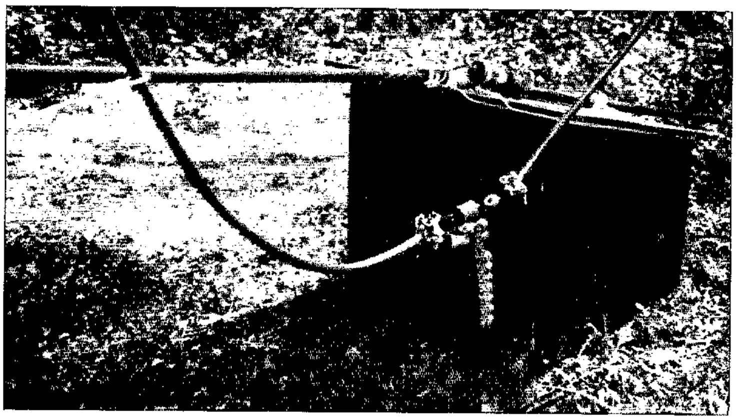

Fig 4—Feed-point detail of the W1LYQ loop. C1, a vacuum variable, is secured to the top of a wooden box. C2 is made of ten "doorknob" capacitors in parallel, and C3 is one large doorknob. The loops are tied off to nearby objects for support.

1) Place the inner loop in either lower corner area of the outer loop, *but not touching it*, in a plane parallel to the outer loop and approximately 20 inches from that plane.

2) Set the transmitter to 1.85 MHz.

3) Adjust C1 for minimum SWR.

By making minor adjustments in the loop position and C1's setting, you should easily get the SWR down to 1:1. Once a 1:1 SWR is achieved at 1.85 MHz, you need only adjust C1 for other frequencies between 1.8 and 2 MHz. The achievable SWR will be 1:1 from 1.8 to 1.85 MHz, and at 1.85 MHz the SWR will climb slowly but steadily to 2:1 at 2 MHz. Changing frequency by more than 2-3 kHz requires readjusting C1 for minimum SWR. Keep in mind that a change of 1 pF in C1 makes a change of 3-5 kHz in the antenna's resonant frequency. I therefore recommend that you use a motor drive with sufficient gear reduction.

For optimum performance, especially at low angles, C1 should be mounted at the top of the loop. Also, it is possible to eliminate C2 and C3 by a suitable choice of the size and position of the inner loop (see Hart, "References," for details). Consider this if you're planning to use high power with the loop.

Warning

Operating a loop antenna indoors or close to dwellings can raise the risk of interference to consumer devices and ham gear. Also, it's prudent to minimize RF exposure to people who may be near the antenna by using the minimum necessary transmitter power to carry on the desired communications. Whenever possible, mount your antennas as far as you can from people and dwellings.

Acknowledgments

I thank Ted Hart for several insightful criticisms of an early draft of this article, and for his permission to quote text and diagrams; and Professor Tom Neuhaus, WB2CLN, and Don Ridley, WT1I, for their comments and suggestions when I was building and testing my antenna. Professor Neuhaus additionally calculated the approximate formulas for the currents and voltages to which C2 and C3 are subjected. I also appreciate the insightful comments and words of encouragement from Vic Misek, W1WCR.

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Notes

[1] See "References."

[2] T. Hart, "Small, High Efficiency Loop Antennas for Transmitting," J. Hall, ed, *The ARRL Antenna Book*, 16th edition (Newington: ARRL, 1991), pp 5-11 to 5-16.

[3] J. Kraus, *Antennas*, second edition (New York: McGraw-Hill, 1988), pp 238-264.

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[7] G. Hall and B. Schetgen, "Six Winners Emerge From the ARRL Antenna Competition," *QST*, Feb 1985, pp 44-47.

[8] T. Hart, "Small High Efficiency Loop Antennas," *QST*, Jun 1986, pp 33-36. Also see Notes 1 and 2.

[9] See "References."

[10] R. Morrow, "The Electromagnetic Antenna," *RadioScan*, Nov 1991, pp 10-13.

[11] K. Patterson, "Down to Earth Army Antenna," *Electronics*, Aug 1967.

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New Books

THE ANTIQUE RADIO BOYS AND THE GARRULOUS GREBE

By Stan Dryer

Published by Rainy Day Books, PO Box 775, Fitzwilliam, NH 03447. Black-and-white illustrations. Retail $6 postpaid First Class mail in the US.

Reviewed By Cynthia Wall, KA7ITT

Holy Hertz! Fans of the original Radio Boys series by Allen Chapman will be pleased to see that Stan Dryer has created their 1990s counterparts in the Antique Radio Boys. The first to be published of the new series is *The Antique Radio Boys and The Garrulous Grebe.*

Back in the 1920s, when radio communications was in its infancy, the Radio Boys solved mystery after mystery using homebuilt crystal sets, spark-gap transmitters and their uncanny sleuthing ability. Now 70 years later, Terry, Jim and Gary, members of Dryer's Antique Radio Club at Cartwright Junior High, retain their predecessors' talents, puns ("Holy Heterodyne—rectify the problem!") and fondness for old radios. Old to them means anything before 1950. In fact, a WWII SCR-5789 Gibson Girl emergency transmitter is featured in the climax of this 50-page story.

If the boys have a few more tools at hand than the original gang, they must still plot their wits against scurrilous Dr Artemisius Krull, thief of Grebe CR-4s. In addition to thievery, Krull is a father who will go to any length to make sure his atonal daughter wins a singing competition against the lovely Lucia Beaucart.

There are other 1920s-type characters who help mix in the action: Krull's accomplice, Verton Gridleak—a thug identifiable by the scar on his hand received in a soldering iron fight; Lucinda's father, Frank Beaucart, proud antique radio collector; and Dick Wood, a Secret Service agent who poses as a mild-mannered radio repairman.

Author Stan Dryer, aka Frank Bequaert, is retired from an electronics, computer and teaching career. He and his wife Lucia operate Rainy Day Books, which offers a wide selection of radio-related technical and historical books.

Reviewer Cynthia Wall, KA7ITT, of Salem, Oregon, is author of several Amateur Radio novels published by the ARRL, including Hostage in the Woods, Night Signals *and* Firewatch.

AMATEUR RADIO RV ANTENNAS

By Robert K. Benson, W2HZF

Published by Tiare Publications, PO Box 493, Lake Geneva, WI 53147. 1992. Softcover, 8½ × 11 inches, 61 pp. $14.95 plus $2 s/h.

Reviewed By Steve Ford, WB8IMY
Assistant Technical Editor

Recreational vehicles (RVs) seem ready-made for Amateur Radio mobile operating. They offer plenty of room for rigs and ample space for antennas. Properly equipped, an RV is an excellent platform for a complete Amateur Radio station.

In *Amateur Radio RV Antennas*, Benson concentrates on HF antennas for RV applications. He discusses vehicle types and antenna designs suited for each application. The discussion includes matching, grounding, single band versus multiband, and so on. (There's also an informative appendix chapter on maritime mobile operating.)

This book, however, isn't all that it could be. *Amateur RV Antennas* avoids detailed evaluations of commercial mobile antennas. Instead, the book speaks in generalities, pointing the reader in various directions, but never quite going the full distance. There are drawings in the book, but no photographs. (Photos of typical RV antenna installations would have been a definite benefit.) And where's VHF/UHF? Excluding antennas for VHF and UHF is a notable omission, considering the popularity of VHF/UHF mobile operating. I'd have even liked to see a few words about antennas for working satellites—an RV could serve as a super mobile OSCAR ground station.

If you're looking for a book that discusses mobile HF antenna *considerations*, you may find *Amateur Radio RV Antennas* worth the price. On the other hand, if you're trying to choose a specific antenna, looking for antenna designs to build yourself or interested in VHF/UHF operating, save your money. You're better off reading *QST* Product Reviews or buying one of the many antenna books on the market. The author knows his subject, but the lack of details leaves the reader hungry for more information.

An Introduction to Amateur Television

Part 3—Part 2 of this series introduced the basic components of an ATV station. Now, we'll look at a variety of technical topics including power amplifiers, vestigial sideband and ATV repeaters.

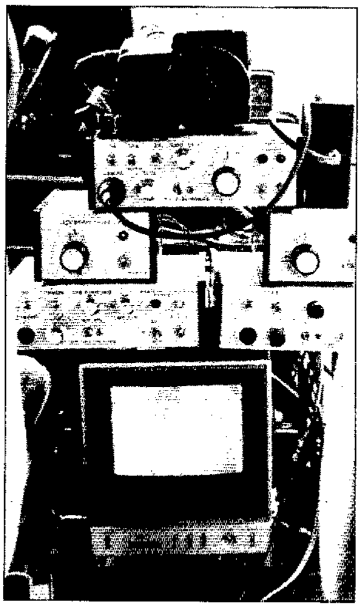
By Ralph E. Taggart, WB8DQT
602 South Jefferson
Mason, MI 48854

Although monitoring an ATV channel with your TV set will help you spot some stations, the directivity of most effective ATV antennas limits the effectiveness of a video CQ. Most day-to-day activity revolves around monitoring the intercom frequency: typically, 2 meters.[9] A short call quickly hooks you up with other stations, providing the necessary information as to where to point the antenna. Our Central Michigan Amateur Television Society (CMATS) group, centered on Lansing, Michigan, now has 13 stations equipped for two-way ATV. Contacts can be had at any time in the evening, and most informal operating sessions involve about half a dozen stations. Another five or six stations at the "looking-in" stage add to the fun.

Getting Organized

Once a local ATV group develops, it's often advantageous to set up a regular time that serves as "local ATV night." It need not be as formal as a net, but it does serve as a focal point for those operators who have to parcel out their operating time. It's also useful for those who are just getting to the looking-in stage. Given the receiving limitations of such stations, being assured of seeing a wider range of stations can help to boost interest on the part of potential new operators.

ATV demonstrations at local radio club meetings can serve to recruit new members. Although video tapes depicting activity are effective, greatest impact is achieved when stations can be worked live at the meeting. This requires a bit of coordination and planning because a complete station—including a moderately effective antenna—must be in place at the club site for the demonstration. The exercise of assembling equipment and coordinating activity can be a useful focus for the local ATV group. Any focussing project can be a great asset in maintaining a high level of local activity and recruiting new members. Providing video communications for public-service activities, building an ATV repeater, mounting a balloon project, or a variety of other activities can serve to keep things lively and interesting.[10]



An ATV station in a car? That's right! At left, Henry (KB9FO) Ruh's well-stacked station. Dave (AH2AR) Pelaez's quick and easy set-up is shown at right. (Better tie down that gear, fellas!) *(photos courtesy of KB9FO and AH2AR)*

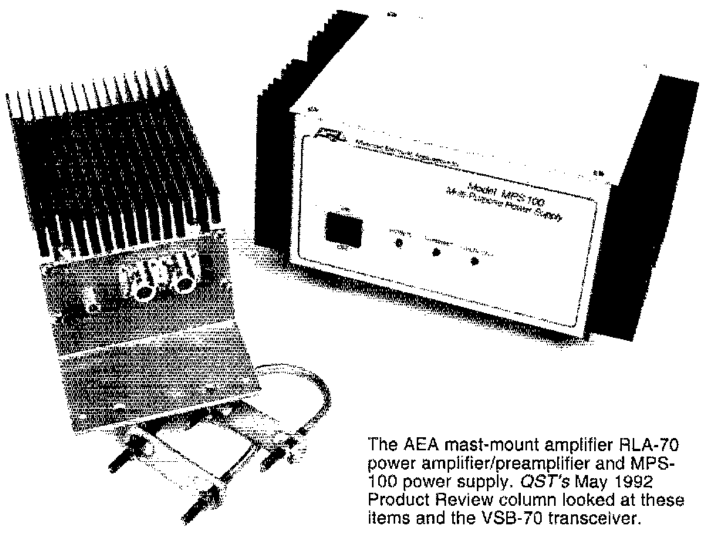
Pleasant Surprises

One spark that can add the element of surprise to day-to-day operations is the tropo DX band opening. The key is in knowing *when* the band is open, or *enhanced*. In most cases, 2-meter band conditions can provide a reliable guide, based on hearing "new" stations on the ATV intercom frequency, or the ability to access repeaters at greater distances. When conditions favor both 2 meters and 70 cm, 2 meters can be used to alert distant stations that you're on, serving as a guide for orienting the more directional arrays used for ATV work. If there are regional variations in ATV intercom frequencies, check out alternative channels on a regular basis.

Although tropo openings typically enhance both 2 meters and 70 cm, this is not *always* the case, nor is the geographic pattern of enhancement equivalent. Repeaters on 70 cm can provide a guide to band conditions, as can DX activity centered on 432-MHz CW and SSB. Stations at the lower end of the UHF broadcast band act as useful beacons for evaluating band conditions. Keeping at hand a log sheet that lists the location of UHF stations[11] within a radius of a few hundred miles will prove helpful in evaluating where to look for ATV activity. When the band is *really* hot, UHF broadcast stations are wall-to-wall well up into the band!

During all but the strongest openings, expect significant signal fading on long paths. If there is a great power/gain differen-

[9]Notes appear on page 41.



The AEA mast-mount amplifier RLA-70 power amplifier/preamplifier and MPS-100 power supply. *QST's* May 1992 Product Review column looked at these items and the VSB-70 transceiver.

tial between two stations attempting to confirm a two-way contact, 2-meter coordination can be the key to success. One strategy that works well is to have the more powerful station transmit, watching for peaks in the received signal. When band conditions peak, a quick shift to transmitting by the low-power station can often result in a readable image. *What* you transmit under such conditions can be critical. Images that work best are large, blocky, black-on-white or white-on-black call signs. Such ID transmissions can be prerecorded on video tape, or generated by computers or dedicated image generators. The universal criterion for a two-way contact is that your call sign can be read—however momentarily—at the other end. Live camera views of the operator work well when conditions are very good, but are not effective when the path is marginal.

One way in which ATV operations differ from the typical CW or voice contact is the use of the "P" system (akin to the better-known RST system) for reporting the strength/quality of the video signal. Reports can range from P0 (some sync, but video not discernible) to P5 (noise-free, "closed-circuit" pictures). The system is certainly subjective (see Fig 16-3 on page 16-4 of the current edition of the *ARRL Operating Manual*), but usage is typically quite consistent, and the reports are a valuable aid in making incremental improvements in your station. Conditions over even relatively short paths can vary quite a bit from day to day, and variations of a P unit are not uncommon. When you have made some change in your station, it is useful to evaluate the effect over several days to ascertain the impact of the change.

RF Power Amplifiers

The most common upgrade of a basic ATV transceiver/antenna combination is the addition of a "brick" amplifier. These are available in a wide range of combinations of power input/output levels, and some models include GaAsFET receiver preamplifiers. Models I know of that have provided good performance in ATV service are listed in Table 5. These amplifiers are quite easy to use because most employ automatic TR switching using RF sensing. This means that you simply stick the amplifier in your transmission line with no switching provisions other than turning the unit on and off (if you don't intend to use it all the time). The RLA-70 looks expensive compared to the other amplifiers listed, but that price includes AEA's MPS-100 power supply. This supply provides power to the VSB-70 transceiver and to the remote RLA-70 amplifier through the coaxial transmission line. Power supplies are *not* included in the prices for the other listed amplifiers.

Where you install the amplifier can greatly impact your system line losses. If an amplifier with an integral receiving preamp is installed close to the antenna, line losses between the amplifier and the transceiver are relatively unimportant. That's because the first RF stage sees only the short length of transmission line connected to the antenna, and the gain of this first stage is usually sufficient to overcome moderate line losses and still control the receiving system noise figure. Line losses impact the drive level to the remote amplifier, resulting in less-than-maximum possible output, but the system is still more effective than if the amplifier were installed at the transceiver end of the line.

The AEA RLA-70 amplifier (see Table 5) carries this strategy to its logical conclusion: The amplifier mounts at the antenna. The 50-watt PEP output of this unit represents an effective power level for local work and will net you *plenty* of contacts during band openings. It's also worth noting that 50 watts generated *at the antenna* is equivalent to a 75- to 100-watt amplifier operated in the shack, given typical line losses! Although many members of our CMATS group can run 100 watts or more into the antenna, they don't do so during local operation. The "afterburners" get fired up during band openings and the occasional test transmission to someone's cable-ready TV set. Otherwise, power levels in the 5- to 15-watt range are the norm.

Lou, K7YZZ ("The Wizard"), built this ATV repeater for the Seattle, Washington, area group. *(photo by K7YZZ)*

Uh-Ohs

One of the major problems—particularly with solid-state amplifiers—is sync compression resulting from amplifier non-linearity. To understand this, we have to delve a bit into the nature of the video waveform. There are four significant signal levels: the sync, blanking, black and white levels. How these are expressed depends on how you analyze or measure the signal. In terms of RF power, a properly modulated signal has the following power-output relationships:

Waveform	*%PEP*
Sync	100
Blanking	75
Black	62.5
White	12

Table 5

Some Solid-State Power Amplifiers Suited for use with Available ATV Transceivers

Model	*Peak Power Input*	*Peak Power Output*	*List Price*	*Vendor†*	*Notes*
D-15N	2	18	$160	PCE,WR	
RFC 4-32	3	20	$183	WR	(1)
RLA-70	1	50 PEP	$699	AEA	(1,3)
D-100-ATVN	1	52	$349	PCE,WR	
D-1010-ATVN-R	4	80	$519	PCE,WR	(2)
RFC 4-110	10	100	$349	WR	(1)
RFC 4-310	30	100	$352	WR	(1)

All of these amplifiers feature RF-sensed switching, so no provisions for manual TR switching are required. Prices shown are manufacturers list prices; discounts are common. Because many of these amplifiers are widely used by SSB and FM UHF operators, they're available off the shelf from local dealers and Amateur Radio equipment suppliers. The D-series amplifiers are produced by Mirage/KLM; the RFC series are manufactured by RF Concepts/Kantronics.

Notes

(1) Amplifier features built-in GaAsFET receiver preamplifier.

(2) Ruggedized model designed for continuous duty/repeater service.

(3) Unit enclosed in weather-resistant housing for mast mounting with the power supply (designed to power the remote amplifier via the transmission line) included in the list price. See the text discussion for the significance of the power-output ratings of these units.

†Contact information for Advanced Electronic Applications (AEA), P.C. Electronics (PCE) and Wyman Research (WR) is given in Table 1 in Part 2 of this article. See May 1993 *QST*, page 44.

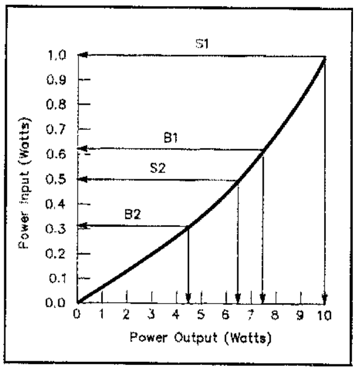


Fig 2—Power output as a function of input for a hypothetical solid-state power amplifier. S1 indicates the input and power output levels on the sync tips for a drive level of 1 watt peak. B1 indicates the relative input and output power for the blanking pedestal (0.5 watt drive). S2 and B2 represent the analogous power relationships when the amplifier is driven at the 0.5-watt input level at a more linear portion of the input/output power curve. The text discusses the effect of amplifier nonlinearity on the transmitted video waveform.

When the transmitter's modulator circuits are properly adjusted for a standard NTSC video signal, the video-waveform sync tips represent 100% transmitter power output (let's say about 1 watt). The blanking pedestal should be 75% of peak output (0.75 watt), black level is at 62.5% (0.625 watts), and white represents 12%, or 0.12 watts. The *difference* of 37.5% between the black level and the maximum power at the sync tips is needed so that the receiver can reliably separate the sync pulses from the rest of the video waveform, locking in (synching) the picture. If we want to amplify this signal, we must do so in a way that preserves the relative power relationships: This is impossible if the amplifier stage(s) *are not strictly linear.*

Fig 2 shows a hypothetical input versus output power curve for a 1-watt input/10-watt output, solid-state amplifier. No matter how carefully you can control the bias in one of these circuits, there's always a certain amount of *gain compression*: The harder you drive the stage, the less power output you get for a given change in drive level at the upper end of the power curve. This nonlinearity matters little for CW or FM service. Some SSB operators see it as a distinct advantage because it affords the equivalent of modest speech compression, if the nonlinearity is not severe. (However, a nonlinear, splattering SSB signal won't endear you to your neighbors on the band.) The graph shows, however, what happens if we drive this amplifier with our hypothetical 1-watt video signal.

With 1 watt of sync (peak) drive from the exciter (S1), the amplifier delivers the expected 10 watts of sync output. With black-level video, the amplifier is driven with 0.625 watts (B1), resulting in a power output of 7.5 watts, 75% of the peak sync level. In this case, the sync pulse has only 25% headroom instead of the proper 37.5% (100−62.5). This *sync compression* makes it difficult to lock the picture (it will roll and tear on the screen) at all but very high signal levels. In effect, the amplifier substantially improves your power output, but at the expense of making the signal unstable at the other end of the path! In practice, the effective sync compression is much worse because the input/output power curve used for Fig 2 is a great deal more linear than that of many "brick" amplifiers, and the problem is often compounded by overzealous operators who overdrive amplifiers!

Remedies

There are only two easy solutions to the problem. One is to reduce the drive level to the amplifier so it's operating in a more linear portion of the input/output curve. If we attenuate the drive level from the exciter by 3 dB, peak sync drive will be 0.5 watt (S2) yielding 6.5 watts of output on the sync tips. The black drive level will have been reduced to 0.31 watt (B2), resulting in a blanking power output of 4.5 watts, about 69% of peak power. Although still higher than the optimum 62.5%, the result is significantly better than the 75% with full drive. Although this greatly improves synchronization, it's a bit frustrating having to operate the amplifier at significantly less than its rated output!

A second solution is to *predistort* the video drive signal so that it has excessive sync, resulting in the proper sync power ratio at the output. The three ATV transceivers discussed previously each have various forms of "sync stretcher" circuits to do just that, allowing you to push the amplifier harder. The solution isn't a perfect one because the nonlinearity affects the video dynamic range and color purity to varying degrees, but it does allow you to get more out of a given amplifier and still maintain acceptable video quality. The major disadvantage is that if you want to use the transceiver on its own, you'll have to readjust the stretcher setting, otherwise the exciter signal will have lower-than-normal video contrast.

AM Linear Service

Perhaps the most critical mode for any amplifier is AM linear service; you can expect to spend some time getting it right if you're using anything other than the AEA amplifier. You can assume that the manufacturer has set up your transceiver properly, but it's almost impossible to adjust an amplifier just by looking at your signal from the shack because you'll be overloading almost everything! Using a *line sampler* and an oscilloscope is ideal, but reports from other local stations (particularly if they can monitor your transmitted waveform) will get you in the ballpark. Line samplers are devices that are placed in the transmission line, sampling the RF envelope and providing enough amplification to drive a video monitor. PC Electronics, Wyman Research and Pauldon Associates market line sampler units; recent editions of *The ARRL Handbook* feature details on the construction of a suitable instrument.

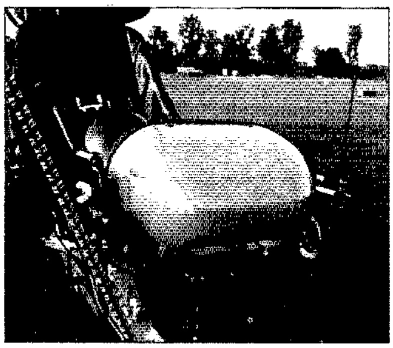
Tube-type amplifiers are less fussy because they are generally more linear than solid-state devices, and tubes are used almost universally if you want power outputs much above 75 to 100 watts peak. Such amplifiers typically require a modest intermediate amplifier, however, if you are to develop maximum possible output from the final when using a 1-watt-output video transceiver. The 15-watt RF Concepts amplifier (and others of equivalent power rating) do the job nicely, allowing you to maximize the power output of your big amplifier.

Some experimentation with settings of the final amplifier's tuning, loading and bias controls is usually required to optimize power output while avoiding sync compression. Some tube amplifier circuits (commonly UHF transmitter strips from surplus commercial equipment) suffer from bandwidth limitations that make it difficult—or impossible—to push a color signal and FM sound subcarrier through the unit.[12] If careful grid tuning can't solve the problem, it's always possible to grid-modulate the final with excellent results, using the TV exciter as a driver with video input disabled. Recent editions of *The ARRL Handbook* feature a grid modulator designed by K2KQZ that can be adapted easily to work with a number of power tubes, with the advantage of significantly more amplifier efficiency than can be obtained in AM linear service. Most of the high-power amplifiers in the Lansing, Michigan, area use a single 4CX250 or 8930 in a coaxial cavity, or two tubes in a parallel stripline configuration (K2RIW design). All of these work well in AM linear service if set up properly, but have a narrow bandwidth.

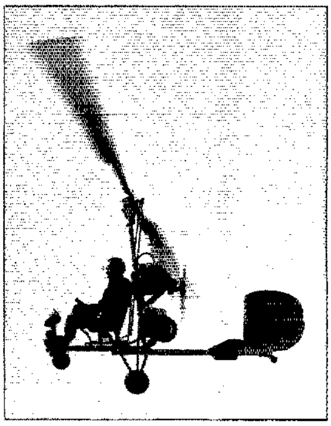
Checking Power Output

Expect to be disappointed if you check the output power level using a wattmeter. The sync pulses, representing your peak power, won't be indicated unless the power meter has been designed to give a true peak power indication. Most "peak power" meters actually indicate on the basis of certain assumptions about audio-power distribution and won't do the job in video applications. Your "indicated" power on a typical wattmeter, often referred to incorrectly as "average" power, will typically be 50% (or less) of the peak power, depending on the modulator setup and drive level. A good wattmeter can be used effectively to help you set up an amplifier (if you don't have access to an oscilloscope and line sampler) and most transceivers come with guidelines that help you out. When noting their power output, most stations specify "average" or "peak" power levels and you can make the rough correction to see how you compare.

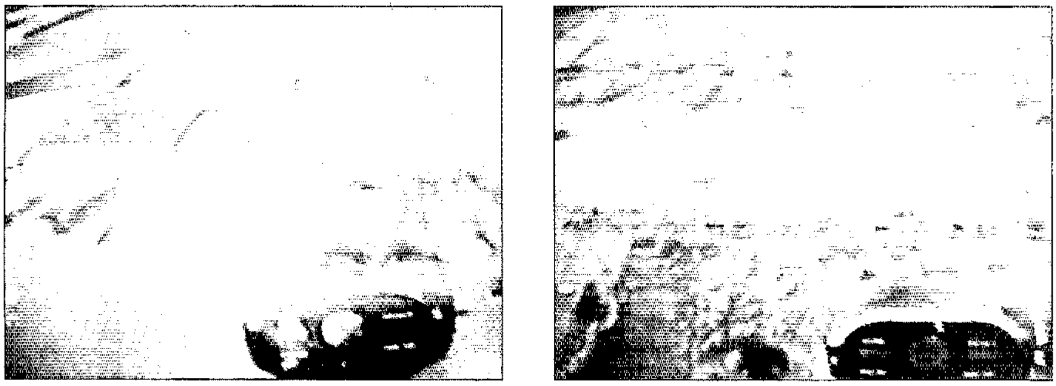
If you're using a vestigial sideband (VSB) filter (more on vestigial sideband in a minute), expect to see at least a 25% decrease in *indicated* power compared to the same amplifier operated with the filter out of line. Most of this power decrease represents energy in the unwanted lower sideband (which the filter is designed to remove) as well as a power reduction resulting from filter insertion loss. The carrier-power loss is equivalent to 0.5-1 dB, about 8 to 16%.



Looking nose-to-nose with my home-built autogyro you see the custom digital/analog instrument pod with a ⅝-wave 70-cm "rubber duck" antenna (off to the right). The antenna is used for ATV and FM communication with a 70-cm H-T. Either the ATV transmitter (a 1-watt P.C. Electronics unit) or the H-T is installed in a small pouch next to the seat. For voice work, the H-T is interfaced to an audio-equipped helmet; the PTT switch is on the gyro's throttle assembly. The ATV transmitter is keyed on and off with a small toggle switch mounted next to the seat. *(photos by WB8RJY)*



Here I am, flying by in my ATV-equipped gyroplane.



A gyro pilot's view of what's below. The photos are of in-flight ATV video received by Jeff, WB8RJY, located about 6 miles away. In this early test of the airborne ATV system, the antenna was attached directly to the transmitter. The pod-mounted antenna, however, has a much cleaner pattern. I was at an altitude of 700 feet when this video was transmitted and the little whip antenna worked well despite cross polarization. (The whip is vertically polarized, but the receiving ground stations use horizontally polarized antennas, as I do at home.)—*WB8DQT*

Vestigial Sideband

One of the most contentious subjects in ATV circles today concerns the merits of using vestigial sideband (VSB) for amateur video transmission. If we're going to approach the subject honestly, it's necessary to recognize that there are three dimensions to the discussion: the marketing climate, the technical side and the operational advantages that attend the use of VSB equipment. Of the various transceivers discussed in Part 1 of this series, only the AEA VSB-70 is designed for VSB service. AEA has invested time and money in the development of the technology, and they are understandably vocal about the advantages. It should not come as a surprise that manufacturers of conventional DSB tend to discount the impact of using vestigial sideband!

Vestigial sideband video transmission has no analogy to other amateur modes. VSB is *not* equivalent to single-sideband because the VSB signal retains the carrier and some lower-sideband components within 1 MHz of the carrier frequency, but with significant suppression of the rest of the lower sideband. That includes the LSB color burst signal and aural subcarrier (3.58 and 4.5 MHz below the carrier, respectively). VSB is used exclusively in broadcast television, allowing for 6-MHz channel separation.

Generating VSB

There are two ways to generate a VSB signal. The approach used by AEA involves the use of a surface acoustic wave (SAW) filter for the extremely low-level channel 3 video signal, followed by amplifier and mixer stages that are all conservatively designed to retain the VSB signal characteristics. The second approach is to amplify the conventional DSB signal to the desired power level, then route it through a sharp, vestigial sideband filter. At 70 cm, these filters take the form of multi-element interdigital filters that pass the upper sideband and carrier, but progressively attenuate the lower sideband components. Suitable filters are available from Spectrum International and International Crystal Manufacturing.[13] The Spectrum International filter is the less expensive of the two, but has a significantly higher insertion loss and lower power-handling capability than the ICM unit. One of the major limitations of add-on VSB filters is that they are single-frequency devices. If you operate more than one frequency, you'll need an individual filter for each of those frequencies. (Ouch!)

Linear amplifiers do provide an additional complication when used in conjunction with a VSB signal. With almost all amplifier circuits, intermodulation products result in the regeneration of the attenuated lower sideband. If a narrow band-pass filter is used to produce the VSB signal, it should be placed in the line *after* the power amplifier if you're expecting to produce a true VSB signal. The RLA-70 amplifier from AEA is specifically designed to maintain a high degree of lower-sideband suppression (equivalent to that obtained with most band-pass filters) when driven by their VSB-70 transceiver.


From this lofty perch, a remote camera and link antenna, part of the Baton Rouge, Louisiana, ATV repeater, allow area ATVers to watch the weather on the Mississippi River and downtown Baton Rouge. The man on the scene is Kenny, WB5JLZ, the repeater trustee and club president. *(photo courtesy of WB5JLZ)*

VSB is mandatory in two situations. One of these concerns the transmitter and receiver of in-band repeaters. Without vestigial sideband filtering, it's almost impossible to reduce desensing and intermodulation effects when the repeater receiver and transmitter are operating in the 70-cm band. Interdigital VSB filters are used in *virtually all* repeater installations, although new repeaters may make increasing use of the VSB capabilities of the AEA equipment line.

The second mandatory use concerns stations north of line A (see note NG135 on page 2-42 of the current edition of the *ARRL Operating Manual*). Basically, stations near the Canadian border can only use the 430- to 450-MHz portion of the 70-cm band. Operation on 439.25 MHz presents no problem, but frequencies below 430 MHz cannot be used. The possibility of operation on 434 MHz is dependent on the use of VSB. A DSB signal has detectable lower-sideband energy (including the 4.5-MHz aural subcarrier) and these components *will be below* 430 MHz—and *illegal*. A VSB signal can operate at a carrier frequency of 434 MHz and not violate the current agreements between the US and Canadian governments.

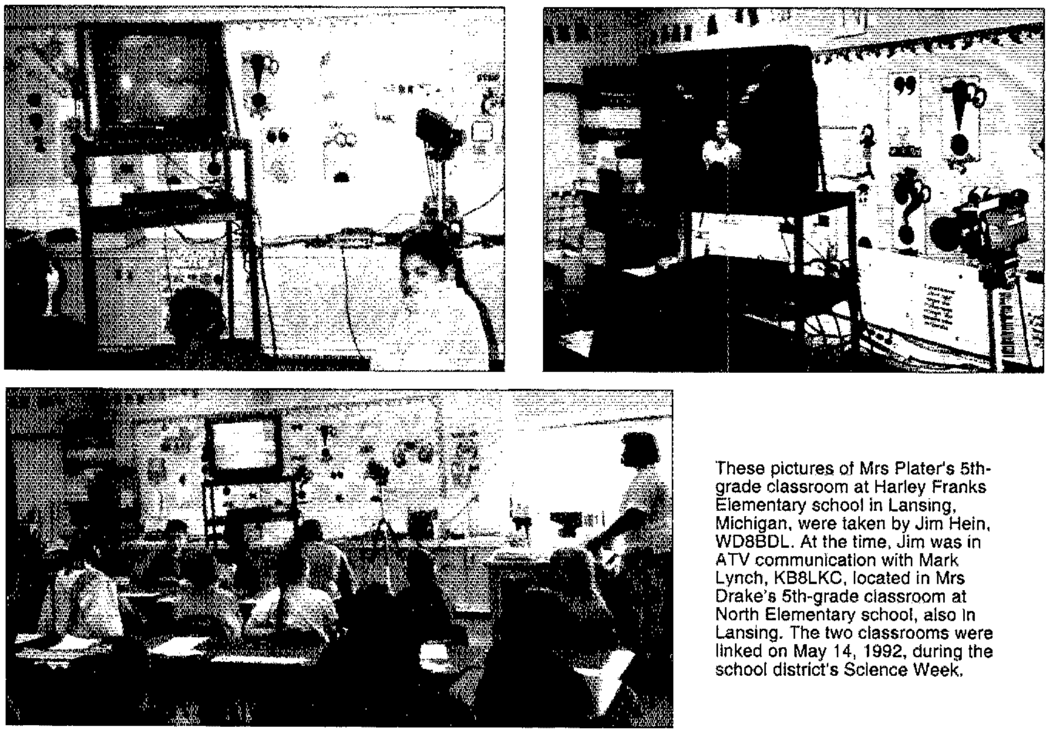
The remaining arguments about VSB are more subjective. The following represent a partial summary where I'll try to be as objective as possible.

- *VSB allows more stations to operate in the 70-cm band.* We can start here with the recognition that although broadcast channel allocations in the US are made at 6-MHz intervals, TV-receiver selectivity limitations result in the FCC avoiding adjacent-channel assignments within a given geographic area. Cable companies achieve adjacent-channel coverage using careful signal filtering and close control of signal levels. Whether or not two channels (say, 439.25 and 434 MHz) can be used simultaneously is very much a factor of: the geographic distribution of the stations in question, how sharp the antenna patterns are, and the relative signal levels. In some cases, where antenna pattern nulls can be used to advantage, two VSB stations can operate simultaneously if power levels are reduced to the minimum necessary to give satisfactory (if not closed-circuit) results.

There are combinations of stations, however, where this isn't possible, given the antenna bearings involved. Also, adjacent-channel operation is usually not possible at all when one or more of the stations is using DSB. Below line A, operation at opposite ends of the band is generally possible with either VSB or DSB as long as power levels are kept to a minimum. VSB can certainly contribute to fitting more channels into the 70-cm band, but this always involves a high degree of coordination and cooperation and due regard to minimize impact on other services within the band.

- *VSB can reduce the potential for interference with other modes.* The potential for interference between operators using different modes varies greatly from one area to another. Obviously, frequency congestion is higher in urban areas. Power density of ATV sidebands tends to be very low beyond 1 MHz (typically −50 dBc) from the carrier frequency, but there are small power peaks associated with the color burst signal (3.58 MHz) and the aural subcarrier (4.5 MHz). The low power density and highly directive pattern of typical ATV arrays minimize the interference potential, but it can be a problem given geographic proximity. VSB, with its reduction of the lower sideband, can be of assistance in terms of minimizing whatever problems might arise with respect to satellite and weak-signal operators. Chronic interference problems tend to escalate into painful confrontations, so the general use of VSB in congested areas may be politically expedient, even if (from the perspective of the ATV operator) there is really not much of a problem. As in all areas of Amateur Radio, there is no substitute for mutual consideration of everyone's interests when frequency congestion is the issue.

Chronic interference problems are emerging in ATV operation on 439.25 MHz because of the proliferation of FM voice repeater outputs in the 440- to 445-MHz segment of the 70-cm band. The results range from mild to severe cross-hatching, loss of color caused by hits near the color burst, loss



These pictures of Mrs Plater's 5th-grade classroom at Harley Franks Elementary school in Lansing, Michigan, were taken by Jim Hein, WD8BDL. At the time, Jim was in ATV communication with Mark Lynch, KB8LKC, located in Mrs Drake's 5th-grade classroom at North Elementary school, also in Lansing. The two classrooms were linked on May 14, 1992, during the school district's Science Week.

of the aural subcarrier, or any combination of these effects. No filtering can mitigate the problem because the offending signals fall within the upper sideband video passband. Those of us located above the line A have no real alternate frequency options without the use of VSB. Some operators have modified the local-oscillator injection of their receive converter to use the lower sideband of the received signal, therefore escaping the bulk of the interference. Unfortunately, such an approach fails if you are attempting to receiving a VSB signal!

- *RF Power Efficiency.* One of the major advantages of SSB over DSB AM is that power amplifiers are used far more effectively in amplifying a single sideband: They don't have to bother with the carrier and unwanted sideband. The result is a definite improvement in signal-to-noise ratio for a given level of output power. There's a slight efficiency gain with VSB, but it's essentially insignificant because the VSB signal has a carrier and a significant LSB power component within the first megahertz below the video carrier.

What, if any, conclusions can be drawn from the comparison of the merits of VSB and the more common DSB format? VSB does represent a noticeable increase in system complexity and cost. The advantages are not nearly as obvious as the differential between SSB voice and DSB AM, but the differences are *real* and can be achieved *without any changes in the receiving system.* Responsible spectrum management is going to become an increasingly important issue, beginning in our congested urban areas. Our most valued ATV band is 70 cm. Although I don't expect the change to be rapid, the trend, in the long run, will be toward increasing use of vestigial sideband, if for no other reason than the fact that it is now feasible and is, ultimately, *the most responsible use* of valuable spectrum.

ATV Repeaters

In general, an ATV repeater is very much like a repeater for any other service. It needs a receiving antenna, receiver transmission-line filters, control circuits and signal patching, an output transmitter, output line filtering and the transmitting antenna. As is the case with other types of repeaters, ATV repeaters are located at the best possible sites to optimize coverage. ATV repeaters are of two general types: *in-band* systems where the transmitter and receiver operate in the 70-cm band, and *cross-band* systems where the repeater input is typically at 70 cm and the output is on 33 or 23 cm (with the latter option being the most common).

Operation of an in-band repeater is convenient for local operators because virtually everyone has tunable receivers, eliminating the need for new transmitting or receiving equipment to access the repeater. Such repeaters represent a major technical challenge, however, because the wide signal bandwidth makes it difficult to achieve the required isolation between the receiver and transmitter. Typically, 439.25 MHz is used for the input or the output, with 421 or 426 MHz for the reciprocal function. Sites above line A effectively cannot use an in-band repeater because it's impractical to achieve the required isolation with the existing frequency limitations.

Cross-band systems are less of a challenge because isolation is easier to achieve, but that approach requires that all stations equip themselves with 23-cm antennas and converters. Cross-band systems are ideal for urban areas because they minimize frequency usage in the 70-cm band, and the reduced coverage of the 23-cm output is not a major limitation. Cross-band systems are the only practical option for sites with line-A limitations. A cross-band repeater advantage is that you can monitor your signal on the repeater output—a great help in setting up amplifiers, modulators and the like.

Actuation of the system is typically achieved using a 15.734-kHz tone decoder connected to the receiver video output. The tone decoder is activated by the received sync signal, thus providing a very high level of immunity from triggering by other modes, unless these options have been included in the system design. Almost all repeaters feed any 4.5-MHz aural subcarrier through to the output, and many will mix the local intercom frequency as well, providing output audio compatible with a very wide range of stations. More-advanced systems incorporate DTMF decoders to enable additional functions, including weather radar displays, external storm-watch cameras, computer games and almost anything else imaginable.

Most repeaters use vertical polarization for ease in obtaining omnidirectional coverage, but this presents some problems. Because most ATV work uses horizontally polarized signals, stations either have to invest in additional antennas for accessing the repeater, or the local group has to go vertical. This makes it very difficult to work out of the local area during band openings because cross polarization results in path losses in excess of 20 dB. That same polarization loss works well to minimize interference from and to local UHF repeaters when horizontal polarization is used for ATV, but the advantage disappears when vertical polarization is used. Some innovative design work is being done to develop more effective omnidirectional horizontal antennas and hopefully these efforts will prove effective.

Summary

This has been a relatively low-level introduction to one of the visual sides of Amateur Radio. The emphasis has been on easy approaches to getting on 70 cm, our most popular ATV band. There are other ways to configure your station and many other equipment sources. Also, there are other bands in the microwave portion of the spectrum and the possibilities inherent in FMTV.

If you want to learn more about ATV and keep up with the latest happenings and equipment in the world of ATV, contact *Amateur Television Quarterly*.[14]

I hope this introductory series has served to whet your appetite to try a new operating mode. ATV's another way of having fun with Amateur Radio and a way of serving your community, too.

Notes.

[9] A listing of ATV intercom frequencies appears on page 9 of *ATV Secrets for Aspiring ATV'ers*, Vol I. See Note 6.

[10] A half-hour VHS video tape for club and public ATV demonstrations is available for $6.50 from *ATVQ*. See Note 14.

[11] Pages 38 and 39 of *ATV Secrets for Aspiring ATV'ers*, Vol 1, have a complete listing of US and Canadian UHF TV stations by channel.

[12] Interelectrode capacitance is the culprit here. Most commercial-service tube amplifiers use tubes designed for class C (FM) service operation. The tubes (4CX250s and others) have high interelectrode capacitance which effectively shorts out the higher modulation frequencies. Use of a tube designed for linear (class A or AB) service reduces this problem, and with the proper pre-emphasis of the video signal, they'll pass good color and sound. Tuning does help on any high-Q tube type amplifier, but you must always reduce the output power to gain bandwidth. Typically, a 1- to 2-dB power reduction is all that's required to gain sufficient bandwidth. This is accomplished by plate loading and plate-coupling adjustments, broadband input/output matching circuits, or use of designs that are inherently more broadband than a quarter wavelength tuned line or cavity.—*KB9FO*

[13] International Crystal Mfg, PO Box 26330, 10 N Lee, Oklahoma City, OK 73126-0330, tel 800-426-9825; 24-hour fax 800-322-9426; 405-235-3741; fax 405-235-1904.

[14] You can contact *ATVQ* at 1545 Lee St, Des Plaines, IL 60018-1950, tel 708-298-2269, fax 708-803-8994. A yearly subscription is $18 in the US, $22 in Canada and $29 for other countries. *ATVQ* offers several books, including a beginner's guide entitled *ATV Secrets for Aspiring ATV'ers*. This 64-page, highly illustrated text contains a wealth of information that new and prospective ATVers need. Price: $9.95 plus $1.50 postage in the US, $2.50 for Canada and $2.90 for foreign orders.

Other ATV publications include: *SPEC-COM*. The SPEC-COM Journal, PO Box 1002, Dubuque, IA 52004-1002, tel 319-557-1002, 24-hour fax 319-583-6462, ($20 per year [six issues] in the US, $25 in Canada and Mexico, $30 in other countries; *Amateur TV Today! Newsletter*, ($18.50 per year for 10 issues), 529 Cedar St, Tipton, IA 52272-1738; *CQ-TV, BATC* (a quarterly publication from the British ATV Club) and *VHF Communications*. In the US and Canada, the latter two publications are available from ATVQ: *CQ-TV* is $21 per year; *VHF Communications* is $27 per year.

New Books

HOW TO GET STARTED IN PACKET RADIO

By Dave Ingram, K4TWJ

Published by the National Amateur Radio Assn, PO Box 598, Redmond, WA 98073; tel 206-869-8052, fax 206-861-580, 1992. Softcover, 6 × 9 inches, 128 pp, black-and-white illus. $9.95, $2 s/h.

Reviewed By Steve Ford, WB8IMY
Assistant Technical Editor

Amateur packet radio is just about 14 years old and still growing at a rapid pace. The flood of Technician-class licensees has added fuel to the fire and there's a constant demand for beginner-level packet literature.

There are many ways to approach the challenge of introducing neophytes to the modern world of packet. Most are valid, depending on what the author feels is most important. Some packet books concentrate on operational techniques, teaching readers how to navigate the packet universe. Others emphasize hardware, illustrating the ins and outs of setting up packet stations. In *How to Get Started in Packet Radio*, Dave Ingram, K4TWJ, takes the latter approach—and does it well.

In the first two chapters, Dave explains the workings of packet and provides easily digestible packet history. From Chapter 3 onward, the focus shifts to hardware. Chapter 4, in particular, offers a guide to several terminal node controllers (TNCs), multimode communications processors (MCPs) and transceivers that can be used for packet. The guide is useful, but not comprehensive. Equipment models change rapidly, and there are a lot of independent manufacturers not covered, so a few recent, lesser-known and specialized packet products aren't included. Missing are the Kantronics KPC-3 and DR SI DPK-2 TNCs; the Gracilis PackeTwin and PacComm IPR-NB96 TNC/transceiver/high-speed modem systems; the AEA PK-900 and Kantronics DataEngine MCPs; and the Kantronics D4-10 and Alinco DR-1200 and DR-119T transceivers. By the time you reach Chapter 6, you're setting up your packet station and tackling a few sticky problems that often occur. Dave does an especially good job of addressing TVI and RFI issues.

The operational side of the book is rather shallow (less than 20 pages devoted to VHF AX.25 operating), but, as I mentioned earlier, the emphasis is on hardware. Dave does a sufficient job of discussing networking schemes such as NetROM and ROSE, but gives only a passing mention of TCP/IP—one of the most popular and powerful forms of packet networking. This is understandable, however, because TCP/IP operation may be beyond the scope of what many hams would consider to be a beginner-level book.

I was pleased to see that Dave included a short chapter on portable packet operating. This is one of the few beginner books that discusses portable operation. It's a scant four pages, but it's four pages more than most books devote to the topic.

The appendix is full of TNC/MCP interfacing diagrams and tables for various rigs. For some packet operators, the information in the appendix alone may be worth the price of the book. (When was the last time you saw packet interfacing information for Drake and Collins equipment?) There's a rudimentary glossary, but one disappointment is the sparse, one-page index.

How to Get Started in Packet Radio is a fine book for the packet newcomer. Dave's writing style is entertaining and easy to understand. (A bit of extra proofreading would have been nice, as there are minor errors and typos, such as mislabeling a photograph of an AEA DSP-2232 as the DSP-1232, and the Kenwood TM-741A as the TR741A.) The only potential flaw of his hardware emphasis is redundancy. With a few exceptions, most TNC/MCP manufacturers already provide adequate setup details in their equipment manuals. The book takes this information several steps further, however, and that may be useful for many readers. If your interest in packet mainly concerns purchasing and connecting your station equipment, *How to Get Started in Packet Radio* is certainly worthwhile. If you're confused about packet *operating* and desire a detailed discussion, you may wish to look elsewhere.

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OUR COVER

You're invited to the 1993 ARRL National Convention August 14-15 in Huntsville, Alabama, site of the US Space and Rocket Center, the largest space museum in the world! Huntsville is famous as the home of NASA's Marshall Space Flight Center and the US Army's Missile Command. See page 185 for more information. *(photos courtesy of NASA, the US Space and Rocket Center, Jared Cassidy, KD4BXO, and Jim Wallace, W4NYC)*

CONTENTS

July 1993
Volume LXXVII Number 7

TECHNICAL

NEWS AND FEATURES

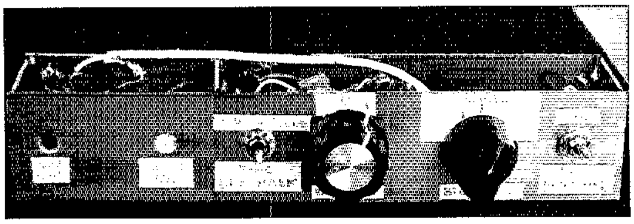
NEW HAM COMPANION

OPERATING

DEPARTMENTS

A Synchronous Detector for AM Transmissions

A "sync" detector far outshines diode detection for good amplitude-modulation reception. Here's one you can build—and all you need to align it is a digital voltmeter.



By Jukka Vermasvuori, OH2GF
Viputie 3
SF-01640 Vantaa
Finland

Do you relax or keep up on the news by listening to shortwave or long-distance mediumwave broadcasts? Are you frustrated by raucous distortion during fading minimums? This article describes how to build one solution to this problem: a simple, effective, 455-kHz *synchronous detector* for your transceiver or receiver. As a bonus, it adds basic CW/SSB reception to receivers unequipped to receive these signals.

Much like switching your receiver or transceiver to SSB and receiving AM as SSB, its carrier at zero beat, synchronous detection overcomes fading-related distortion by supplying an unfading carrier at the receiver. The difference between synchronous AM detection and such simple product detection is that synchronous detection *phase locks* its carrier to that of the incoming signal. There's no tuning error if the received signal happens to fall between your radio's tuning steps, and you don't have to correct for modest tuning drift. The result is a dramatic fidelity improvement over diode-detected AM. Enjoyable music audio is recoverable even from reduced-carrier SSB broadcasts.

The Synchronous Detector Circuit

See Fig 1. The unit uses popular NE602AN (mixer/oscillator) and NE604AN (FM subsystem) ICs to provide both synchronous and quasi-synchronous detection.[1] Operating at a supply voltage of 6, the circuit draws 10 mA. U1, an NE602AN, acts as the BFO and product detector necessary for synchronous detection. Feeding U1's balanced inputs in push-pull helps keep BFO energy from backing out of the input pins and into U3's limiting circuitry. To take advantage of the chip's internal biasing, the input transformer (T1) is isolated with dc blocking capacitors. (U1 also supplies balanced audio output, but usefully reducing this to a single-ended output would have required an operational amplifier. Doing so would reduce even-harmonic distortion in recovered audio, but would not, I decided, justify the increased circuit complexity and power consumption.)

This Synchronous Detector Circuit Affords:

- No signal-to-noise threshold. The incoming signal's IF signal-to-noise ratio is converted to audio as is—something not true of envelope and quasi-synchronous detectors.
- Suppression of overmodulation distortion during carrier fades because the strength of the locally generated BFO remains constant.
- Quality music reproduction with reduced-carrier SSB signals.
- Averaging effect against adjacent-channel splash, partially suppressing it in detection.
- Rejection of phase noise from transmitter and receiver synthesizers.
- Rejection, during reception of reduced-carrier SSB, of phase-modulated second-harmonic distortion and intermodulation-distortion products, therefore giving less distortion in detected audio than simple product detection.
- Rejection of spurious PM sidebands generated in conventional AM transmitters due to improper neutralization or antenna matching.
- Low harmonic distortion (below 1%) in audio recovery, independent of IF level.
- Low AGC-controlled IF input level, therefore needing less pre-IF amplification.—*OH2GF*

U1's oscillator amplitude is optimized to 660 mV P-P (as measured across L1) by the 220-Ω resistor at the oscillator output at pin 7. The BFO frequency is adjusted by two variable-capacitance diodes (D1 and D2) in addition to the tank coil, L1. D2 receives its control voltage via switch S3 (**BFO MODE**), which selects control voltage from either U3's phase detector (**SYNC**) or a constant voltage from a resistive divider (**CW/SSB/TUNE**).

The fixed **CW/SSB/TUNE** voltage (2.16, set by the ratio of R5 and R6) corresponds to the phase detector's optimum output voltage at lock.[2]

R2, **BFO TUNING**, drives D1 to provide manual detector tuning without upsetting D2's control-voltage optimization. S2, **BFO OFFSET**, presets R2's tuning range into the optimum regions for LSB (**−2 kHz**), DSB AM (**±0 kHz**) and USB (**+2 kHz**).[3] The **BFO TUNING** control therefore provides fine adjustment for detector lock around the receiver's tuning steps (coarse—1 kHz—in my receiver).

U3's phase detector requires a 90° phase shift between the incoming and reference phases to give the correct zero-phase output (again, approximately 2.16 V). The all-pass stage, Q1, operates as an isolation stage and adjustable phase shifter to generate the 90° phase shift.

U3 is an NE604N FM IF subsystem IC that contains limiting-amplifier stages (total gain, 101 dB) and a quadrature detector. Band limiting can be inserted between the limiter stages, but experiments with various RC and LC filters brought no improvement, and instead led to increased delay that upset the carrier/sideband phase relationships necessary for good quasi-synchronous detection. The Fig 1 circuit uses U3's quadrature detector as a phase detector that outputs control voltage for D2.

The most difficult aspect of the phase-

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[1] Notes appear on page 33.

locking chain is the selection of a time constant for the locking loop. Signal fading, and the relative absence or presence of phase-modulation components in the transmitted signal, play important roles in detector lock. Were fading not a problem, a short time constant—one allowing fast locking—would suffice for DSB AM. For SSB AM with carrier (which includes a phase-modulation component at all modulation frequencies), however, and DSB AM with fading (during which a fast PLL may unlock on the sudden phase shifts that can accompany fast, deep carrier fades), a long time constant is necessary. Particularly for SSB with carrier, the loop bandwidth must be reduced to below the lowest expected modulation frequency. C1 and R4 set the PLL time constant in the Fig 1 circuit.

The received signal strength indicator (**RSSI**) output at pin 5 of U3 follows the input level logarithmically, giving an output of 1.1 V on noise only (**RF INPUT** shorted in Fig 1) and 3.3 V at an **RF INPUT** level of 3 mV. The **RSSI** output is adaptable as an AGC-detector output, making the NE604AN attractive for simple IF-AGC designs. Because of the NE604AN's high gain, circuit layout can be critical, requiring short leads and physically small bypass capacitors. Coupling must be minimized between pin 9 of U3 and the U1 oscillator components.

U2, an NE602AN, operates as a quasi-synchronous detector. The BFO energy it requires is readily available as a square wave at pin 9 of U3. Except for the fact that its BFO input is derived from a limited input signal instead of a VCO, U2 functions the same way as U1.

Construction

Two evaluation models were constructed using ground-plane construction, mounting the ICs upside down and soldering their ground pins directly to ground with minimal lead length. The later version is constructed onto a long, narrow piece of circuit board intended to be the bottom plate of an add-on box to be fixed under a Sangean ATS-808 receiver. (Fig 2 shows the general layout of this version.) The Sangean ATS-808 (sold under this and other names in the US, including the Radio Shack DX-380), is an "all-band" double-conversion superheterodyne receiver covering 150 kHz to 30 MHz and featuring two selectable IF bandwidths. Adding this outboard detector circuit also allows it to receive CW and SSB.

To avoid crosstalk, I made the receiver-detector IF-AF connections with small-diameter coaxial cable. With these precautions and circuitry arranged as shown in Fig 2, BFO-signal leakage is unmeasurable at U3; that is, the voltage at **RSSI** does not change when the BFO is temporarily disabled under no-signal conditions.

IF signal can be obtained from the ATS-808 via a 56-pF capacitor connected to the hot end of the '808's transformer T9 (at pin 16 of the '808's U1, a TA7758P IC).

Connecting the detector cable detunes T9, which, though difficult to reach, must be retuned by turning its slug outwards a few turns to obtain maximum audio output. The

Quasi-Synchronous Detection

Synchronous detection can be mimicked by amplifying and limiting the AM signal sufficiently (at IF) so that only carrier remains, and substituting this signal for the BFO at the product detector. This *quasi-synchronous detection* acts much like envelope ("diode") detection and works best when the received signal does not fall to zero, as can often occur with SSB and, with AM, during fading. As the signal fades and the carrier-to-noise (C/N) ratio decreases, noise renders the detector's switching action inconsistent, and detection quality deteriorates rapidly. Thus, under conditions of low C/N ratio, quasi-synchronous detection exhibits a distinct detection threshold, as does a diode detector. The chief advantage of quasi-synchronous detection over simple diode rectification is its much lower input level compared to that required by a diode. The detector circuit I present in this article includes a quasi-synchronous detector for flexibility and A/B comparison with the synchronous circuit.—*OH2GF*

Non-Phasing Synchronous Detection: The Better Way?

Amateur Radio transceivers generally select USB or LSB through intermediate-frequency (IF) filtering. Most consumer multiband radios with synchronous AM detectors use *phasing* synchronous detection in which audio-frequency (AF) and IF phasing are used to select the upper or lower signal sideband.*,† Such a system requires *two* synchronous detectors, one that responds to amplitude and another that responds to phase.

The phasing approach has two serious drawbacks. First, even though phasing detection can attenuate opposite-sideband *audio*, it cannot prevent opposite-sideband RF from driving IF-derived automatic gain control (AGC) circuitry and affecting receiver gain. The second drawback is that a phasing synchronous detector detects the phase noise sidebands of transmitted carriers and its receiver's local oscillator (LO) and converts them to audio. (The system's amplitude detector demodulates AM and no phase noise; the system's phase detector demodulates phase noise and quadrature AM. Demodulated phase noise is therefore present in their summed output.)

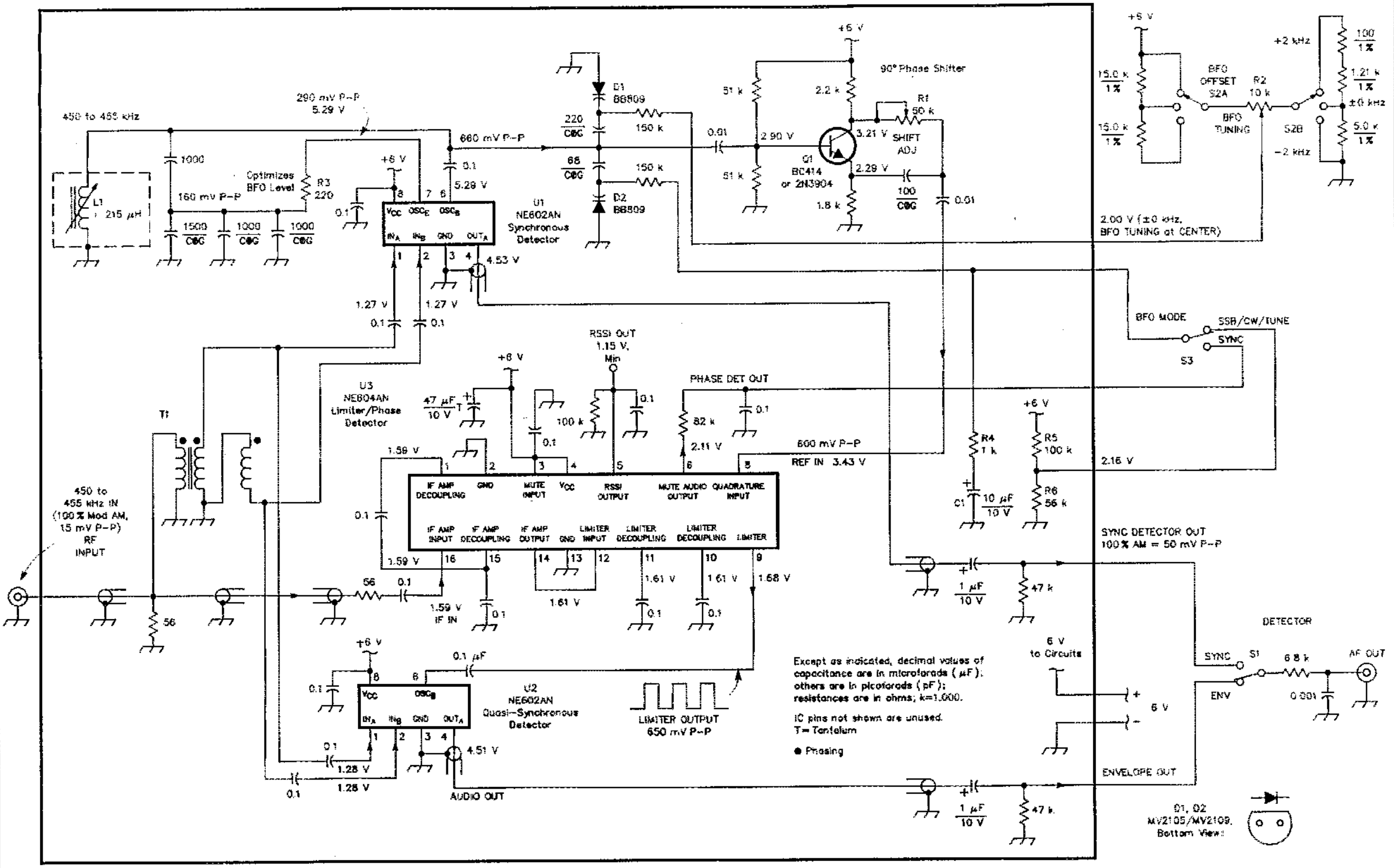
This problem is not trivial. International shortwave broadcasters are working on moving from full-carrier double sideband (DSB) to reduced-carrier SSB transmission by sometime next century. Received with the simple diode detectors long established for AM reception, reduced-carrier SSB may be unacceptably distorted.†† Synchronous detection would seem to solve this, but synchronous detection requires greater receiver stability and tuning accuracy than has ever been necessary with diode detection. PLL synthesis, now used in consumer shortwave receivers even in the US$100 to $200 range, is arguably the best means of achieving these aims economically. But PLL synthesizers economical enough for this service are generally so phase-noisy that they compromise phasing synchronous detection when it is applied. Despite this, most of the radios currently available with synchronous detection use phasing detectors!

Seeking to add a synchronous detector to my Sangean ATS-808 receiver, I therefore decided that a simple basic model—one that uses the receiver's IF filtering to reject the unwanted sideband—would give better overall quality. Whether the marketplace will arrive at the same conclusion remains to be seen!—*OH2GF*

*It follows, of course, that when only one sideband is transmitted, there's no opposite sideband to reject. Even if no other stations are using the frequencies represented by the absent sideband, opposite-sideband rejection is entirely worthwhile because it keeps us from receiving the band noise and static present in that slice of spectrum. When both sidebands are present, the ability to receive either of them at will lets us choose whichever of the two is least troubled by interference.—*Ed.*

†For a nonsynchronous, high-dynamic-range system embodying these principles, see R. Campbell, "High-Performance, Single-Signal Direct-Conversion Receivers," *QST*, Jan 1993, pp 32-40.

††The severity of the distortion depends on how much the carrier is reduced relative to its full-carrier value. Broadcasters intend to use SSB_{-12}—SSB with carrier enough for synchronous detection, but considerably less than that necessary for useful envelope detection. In practice, envelope-detected SSB_{-12} sounds like suppressed-carrier SSB received with the BFO turned off.—*Ed.*



Except as indicated, decimal values of capacitance are in microfarads (µF); others are in picofarads (pF); resistances are in ohms; k=1,000.

IC pins not shown are unused.
T= Tantalum

● Phasing

Fig 1—The OH2GF synchronous detector operates in the 450- to 455-kHz region. Except as otherwise specified, its fixed-value resistors are ¼-W, 5%-tolerance units, and its capacitors' working voltages can be 10 or higher. See the 1993 *ARRL Handbook's* Chapter 35, Component Data, for the full addresses of the part suppliers mentioned below. Be sure to check *QST* ads for additional part sources for this project.

D1, D2—BB809 or BB409 tuning diode. Each of these, a "28-V" diode, exhibits approximately 33 pF at 2 V and an unusually high voltage-versus-capacitance slope of 10. WJ1Z has used two paralleled 30-V Motorola tuning diodes (one MV2109 [≈45 pF at 2 V] and one MV2105 [≈18 pF at 2 V], both with a slope of 3) to replace each BB809 or BB409 in this application. MV2105 and MV2109 diodes are available from Oak Hills Research, tel 616-796-6633 or 800-842-3748; fax 616-796-6633.

L1—Approximately 215 µH. Toko RWRS-T1019Z (nominally 220 µH, Q of 100 at 796 kHz, available as Digi-Key Corporation's #TK1223), suitable.

R1—50-kΩ trimmer.

R2—10-kΩ linear control.

T1—13 trifilar turns of #28 enameled wire, twisted, on an FT-37-77 toroidal ferrite core.

U1, U2—Signetics NE602N, NE602AN, SA602N, SA602AN mixer/oscillator IC. Available from Digi-Key, Oak Hills Research, Ocean State Electronics and others.

U3—Signetics NE604N, NE604AN, SA604N or SA604AN FM receiver subsystem IC. Available from Ocean State Electronics.

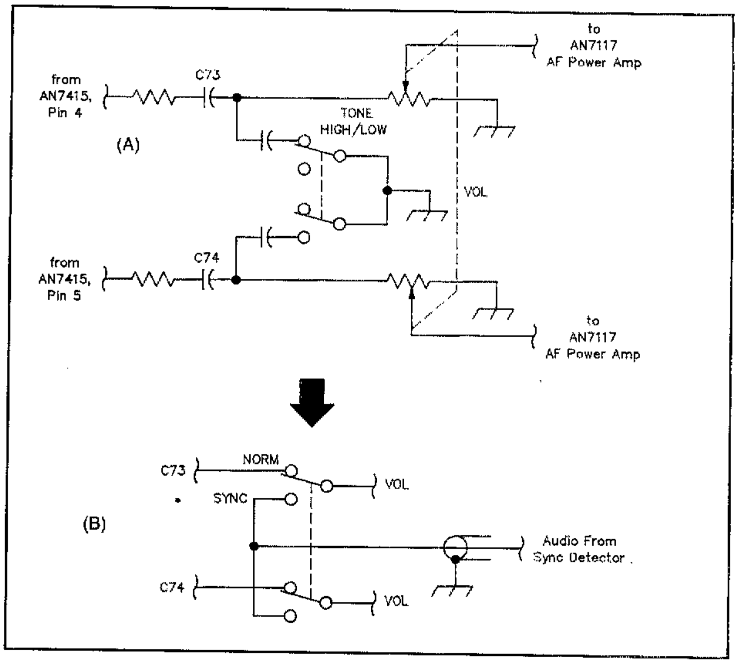


Fig 3—Rewiring the Sangean ATS-808's **TONE** switch (A) to select audio from the synchronous detector (B). The detector receives its power via a 22-Ω resistor connected to the '808's 6-V bus.

detector's audio output (**AF OUT**) returns to the ATS-808 by means of the '808's **TONE** switch, which I rewired to select audio from the outboard detector as shown in Fig 3.

Operation

After checking the circuit, connect it to a 6-V power supply. The total current consumption should be approximately 10 mA. Switch the **DETECTOR** switch to **ENVELOPE**; you should hear band noise. Tune in a strong AM signal, switch S3, **BFO MODE**, to **SSB/CW/TUNE**, and set the **DETECTOR** switch to **SYNC**. The detector may sound very quiet at this point. Adjust L1's core until you hear the signal you were listening to in **ENVELOPE** mode swoop into audibility. Now you know that the BFO is oscillating. If possible, measure the BFO level across L1 with an oscilloscope and 10:1 probe; it should be about 660 mV. (If you can't measure the BFO level, go to the next paragraph.) If it's not, experiment with R3's value to make it so.

Accurately tune the receiver to a strong, pure carrier, such as a beacon. Adjust R2, **BFO TUNING**, for a wiper voltage of 2.00 with S2, **BFO OFFSET**, set to ±0 **kHz**. Mark as **CENTER** this point in its knob's travel. With the **BFO MODE** switch in the **SSB/CW/TUNING** position, use a nonmetallic tool to adjust L1, the VCO coil, for zero beat with the incoming carrier. Returning S3 to the **SYNC** position should allow carrier lock if R1, **SHIFT ADJ**, is reasonably near adjustment. Adjust **SHIFT ADJ** for carrier lock if necessary. This completes coarse adjustment of **SHIFT ADJ**. Return the **BFO MODE** switch to the **SSB/CW/TUNING** position; the BFO

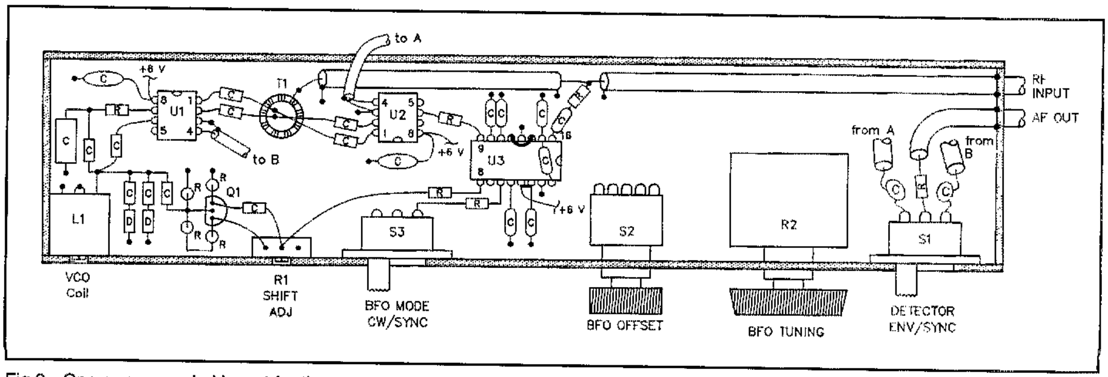


Fig 2—One recommended layout for the synchronous detector—that used by the author to match the footprint of his ATS-808 multiband receiver. U3's high gain requires care in construction—see text.

Notes on Applying OH2GF's Synchronous Detector

OH2GF built this synchronous detector for use with a Sangean ATS-808 portable receiver, but it can also work well with tabletop, communication-quality receivers. I tested my version of the detector (see title photo) with a Drake SW-4A receiver and an ICOM IC-729 transceiver. Fig A shows how I connected the detector to them.

The key to success with this circuit is getting interference-free IF drive. U3's RSSI output can be of critical importance in scoping out possible BFO leakage and/or unwanted signal input. With the detector's RF input shorted, a voltmeter connected to **RSSI** should indicate about 1.1. If it doesn't, U1's BFO signal may be getting into U3. The RSSI indication shouldn't be much higher than this with the detector connected to your receiver and your receiver's RF gain control turned all the way down for minimal noise input to the detector. (Whatever receiver you use must be in "AM" mode—BFO *off.*) If you read an RSSI voltage above the 1-V range at this point, a receiver oscillator or some other signal, however inaudible to you, may be driving U3's limiter.* *The detector will be unable to achieve and hold lock if anything other than the receiver IF signal grabs its limiter.* Just such a condition kept my detector from locking when I drove it with energy from the final 455-kHz IF stage in a Japan Radio Company JST-135HP transceiver. With the '135HP's **RF GAIN** control all the way down and no antenna connected, the detector's RSSI voltage was 3.3—full limiting! Time did not allow the spectrum analysis necessary to characterize the culprit signal, but turning off the receiver returned the RSSI reading to its no-signal value—so the interfering signal is in there *somewhere.*

Ironically, hi-fi may have its price: Listening to the detector with an outboard audio amplifier may reveal distortion and IF hiss inaudible through the driving receiver's audio channel. Many modern radios launder their audio with high-end rolloff between their detectors and AF power amplifiers. A graphic equalizer can do this work for you.

Once I had the detector properly adjusted, I only rarely heard it lose lock on a full-carrier signal, even in one of my toughest subjective tests: Radio Australia's strong but muscularly fady 9580-kHz signal audible in North American mornings. (Judged subjectively, the circuit performed the same with BB409s and MV2105-MV2109 pairs installed at D1 and D2.) It was routinely possible to achieve lock on full-carrier signals right down to the band noise—not that such signals can provide the kind of entertainment-quality listening synchronous detection can provide! I also achieved lock-loss-free listening with HCJB's SSB$_{12}$ transmission at 21455 kHz, and several independent-sideband, reduced-carrier feeders. So OH2GF's synchronous detector means business.—*WJ1Z*

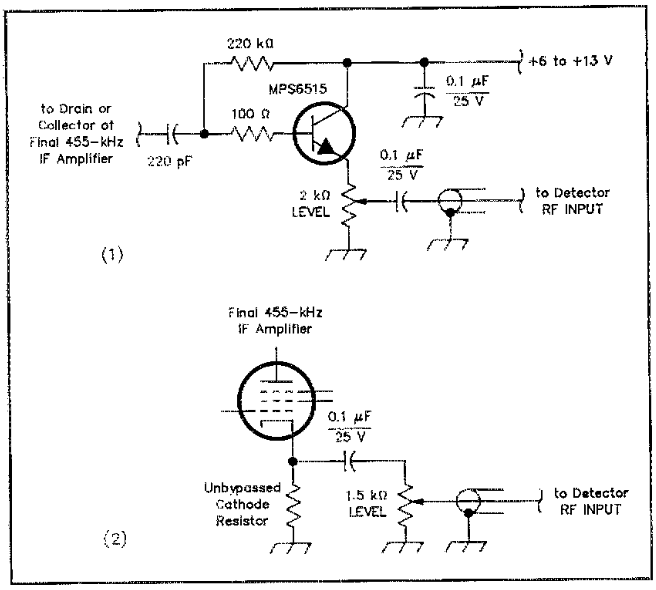


Fig A—A simple bipolar junction transistor emitter follower (1) can connect the synchronous detector to a solid-state transceiver. As shown at 2, the synchronous detector can be driven from the unbypassed cathode resistor of a vacuum-tube receiver's final IF stage. (The WJ1Z Drake SW-4A receiver required other changes, including AGC redesign for low distortion at high signal levels.) If you lack the equipment necessary to adjust these circuits for OH2GF's suggested drive level of 15 mV P-P, just keep the detector drive comfortably below that at which distortion begins. Likewise, if you can't measure the VCO level across L1, just stick with the 220-Ω value Fig 1 shows for R3—it's in the ballpark.

*A local mediumwave broadcaster may even be the culprit, especially if your antenna system is extensive and your detector is relatively unshielded. During tests, I listened to our 1410-kHz local with a turned-off receiver and a powered-up detector switched to **ENVELOPE** and connected to an audio amplifier. Remember: The NE604's limiter stages, specified to work up to at least 21 MHz, are capable of 101 dB of overall gain and specified to be several decibels into limiting with *as little as 3 μV* (–92 dBm) applied across a 50-Ω load at the '604's input!

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should still be at or very close to zero beat with the incoming signal.

Return the **BFO MODE** switch to the **SYNC** position. After the detector locks, fine-tune **SHIFT ADJ** to minimize detected low-frequency hiss. (If a sufficiently strong unmodulated local signal is not available off-air, transmit into a dummy antenna with a PLL-synthesized transceiver and make this adjustment by listening to its signal. You should find a **SHIFT ADJ** setting at which detected hiss distinctly nulls. As a less-desirable alternative, tune in an unfading AM signal modulated with a 1-kHz tone and adjust **SHIFT ADJ** for maximum tone recovery.) Once this is done, the detector's carrier phase is within exactly 0° or 180° of the BFO signal applied to the amplitude detector (U1), and you have minimized the detector's response to phase noise. Significantly, this alignment procedure also sets the detector to lock in a range centered on the control voltage that corresponds to optimum locking sensitivity *and* minimum phase noise demodulation. That's it—you're ready to listen.

To zero-beat and lock a given station: Set the **BFO TUNING** control to its center (2.00 V) position, **BFO MODE** switch to **CW/SSB/TUNE** and **BFO OFFSET** to match the *sideband(s)*—LSB, USB or both—you want to receive. Tune your receiver as close to zero beat as its tuning steps allow. Adjust **BFO TUNING** for zero beat. Switch the **BFO MODE** switch to **SYNC** to lock the detector.

Toggling S1, **DETECTOR**, between **ENVELOPE** and **SYNC** allows you to easily compare the effects of detection mode under adverse propagation conditions. You'll find the synchronous mode to be considerably superior much of the time. The quasi-synchronous (**ENVELOPE**) mode may give crisper audio under average or poor signal conditions; this effect may be due to increasing distortion as the signal approaches the noise floor, however.

The sound picture compared to the original ATS-808 detector, in addition to level difference, is slightly different, probably as a result of the "loudness" band limiting in the original circuit prior to the selection switch. The audio may be processed to individual taste.

Properly adjusted, the synchronous detector operates at less than 1% total harmonic distortion. (The quasi-synchronous detector provides comparable performance—but only on a nonfading test signal.) Measurements confirm the importance of setting R1, **SHIFT ADJ**, properly: Improper adjustment can increase low-order harmonic-distortion products detected from possible phase-modulation sideband components and allow the detection of phase noise.

Conclusion

The synchronous detector I've presented opens new possibilities for simple receiver design by requiring as little as 3 mV of IF output for proper operation. This means that 30 dB less pre-detector gain is required compared to traditional designs. Outfitted with this detector, my Sangean ATS-808 receiver can now receive CW and SSB signals, although the '808's AGC decay time is too short for enjoyable SSB listening.[4] A simple synchronous detector therefore may be the best solution for a compatible, multimode detector.

The modified ATS-808 especially demonstrated its detection power during a trip I made to Springfield, Massachusetts, where it was possible to follow Radio Finland's daily, one-hour SSB $_{,12}$ transmission beamed to West Europe (225°) on 15330 kHz at 1310 UTC with adequate signal quality.

I hope that you'll put this circuit to work with *your* receiver, perhaps modifying it for use at another intermediate frequency. I look forward to hearing of your results.

Notes

[1] This article refers to U1 and U2 as NE602ANs, but NE602Ns, SA602Ns and SA602ANs will work equally well in this application. Likewise, an NE604N, NE604AN, SA604N or SA604AN will work well at U3 in this application.

[2] The resistances given for R5 and R6 set this value only with a supply voltage of 6. If a different supply is used (the NE602s and NE604 can be operated at up to 8 V), change R5's value to return the **SSB/CW/TUNE** voltage to 2.16. This 2.16-V value should itself be considered only as an average for the NE604; the optimum value can be found by using a DVM to measure the voltage across C1 when the receiver is tuned to an empty channel. This optimization is important because manually tuning with **BFO MODE** set to **SSB/CW/TUNE** zero beats the incoming signal at this control voltage. When **BFO MODE** is then switched to **SYNC**, the detector's VCO idling frequency is therefore still almost correct, and easiest locking is guaranteed.

[3] Whether this correspondence between LSB and USB and S2's **–2 kHz** and **+2 kHz** positions holds with receivers other than the author's Sangean ATS-808 depends on whether the radio in question inverts SSB signals in moving them to 455 kHz. At the ATS-808's 455-kHz IF, SSB signals are reversed relative to their on-air sense (USB becomes LSB, and vice versa). For radios in which SSB signals are not inverted at 455 kHz, S2's – position will correspond to USB, + to LSB. During selectable-sideband synchronous reception with communication-quality radios using tighter SSB filtering than that afforded by ATS-808's narrow filter, BFO offsets on the order of ±1.5 kHz will likely be required for optimum carrier lock and tonal balance in recovered audio.—*Ed.*

[4] Proper adjustment of the **SHIFT ADJ** trimmer minimizes the detector's sensitivity to phase noise, but only when the detector is phase-locked. Thus, this phase-noise rejection doesn't apply when the detector is operated in its fix-tuned, unlocked (**CW/SSB/TUNE**) mode. With the ATS-808, receiving strong CW signals therefore includes the addition of keyed noise—the '808's synthesizer phase noise transferred to the CW note. This illustrates how non-synchronous product detection reveals the true quality of a communication system's various oscillators and sets stringent quality requirements for their design and performance.

Jukka Vermasvuori has been an active ham ever since receiving his license in 1956 at age 15. His hobby within the hobby is designing HF receivers and transmitters: He received "First 1970 Award" at RSGB Radio Engineering and Communications Exhibit for a receiver that used dual-gate FETs and a double-balanced mixer.

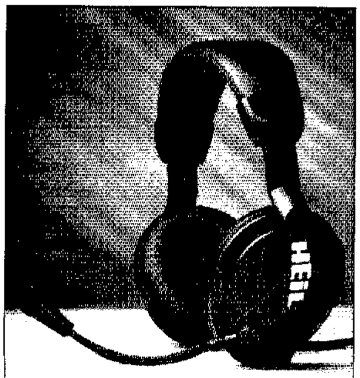
Jukka has spent 30 years with the Finnish Broadcasting Company, purchasing new AM, FM and TV transmitter equipment. Highlights of his career have included the purchase and commissioning of three 1-megawatt-PEP SSB transmitters for shortwave broadcasting, and recently dealing with IOT (inductive output tube) common amplification UHF-TV transmitters.

He has been actively participating in the preparation of "CCIR Report to World Administrative Radio Conference (HFBC-93) Dealing With Matters Connected With The HF Broadcasting Service" (part: SSB System for HF Broadcasting). He is also the author of several technical articles published in Finland and elsewhere.

New Products

HAM HEADPHONES

◊ Designed for the needs of DXers, contesters and DXpeditioners, the Heil PRO-SET boomset headphones are comfortable and efficient. Vinyl earpads snap into the headpiece for easy removal, washing or replacement. The headband is fully padded with detent stops on each side, and features a clever microphone boom that can be placed in almost any position or folded up for storage and traveling. The PRO-SET is available with one of three microphones: the full-range HC-5, the HC-4 "DX Dream Machine" or the HC-6 full-range broadcast-quality element. Suggested retail price is $134.95. Heil Sound Ltd, Heil Dr, Marissa, IL 62257; tel 618-295-3000, fax 618-295-3030.



LiTZ BOX

◊ VHF/UHF FM and repeater operators who want to take advantage of long-tone zero (LiTZ) and other signaling techniques (see the FM/RPT column in October 1992 *QST*) can use wired and tested decoders or "LiTZ Kitz" to set up specialized monitoring capabilities. Based on a project described in December 1992 *73 Amateur Radio Today* ("Build a LiTZ Decoder," page 30), by Marshall Macy, N7IOB, and Paul Holmes, KA5RZI, this compact device connects to your receiver's audio output and can be set to respond to a long "0" or other DTMF tone to trigger an alarm or activate a speaker. An etched and drilled PC board is $9.95; the board with Silicon Systems IC SSI-202 is $16.95; a complete kit (excluding enclosure) is $36.95; and the complete wired and tested LiTZ decoder with case is $56.95. Discounts are available for purchases in quantities of two or more. Marshall Macy, N7IOB, 303 East S Mountain Ave #163, Phoenix, AZ 85040; tel 800-484-9691 (code 7373) or 602-268-3838.

Honey, I Shrunk the Antenna!

Think small! Communicate with multiconductor miniature loop antennas.

By Rod Newkirk, W9BRD
3830 N Richmond St
Chicago, IL 60618

Great energy and ingenuity go into efforts to boost the effectiveness of compact transmitting antennas—*compact* meaning configurations with circumferences of ⅛ wavelength or less. By this definition, a 40-meter design could be a square with 4-foot sides. In the following discussion, I depart slightly from the square. I prefer 3½ feet high by 4½ feet wide, suspended in the vertical plane, for convenience in my antenna's location.

The traditional small-loop approach is to insert capacitance in series with a one-turn loop, tune it to the desired frequency, and then attempt to feed power to it as efficiently as possible—no simple matter. The antenna's Q is astronomical, current and voltage are monstrous, bandwidth is razor sharp, and its radiation resistance is ridiculously low. This lossy situation can be improved by reducing the resistance of the loop material through the use of piping, foil, etc, but the resulting plumbers' nightmares are hardly worth the pains.

A more promising avenue toward practical, simple compact antennas would appear to be in the direction of varied configuration. The goal is to "cool" the system with better power distribution, and at the same time simplify the feed. For example, vast improvement in the key parameters is possible by application of the venerable "folded dipole" technique. This can be done by adding one more turn, tuned identically, connecting the two turns *in series*, and re-resonating the antenna. Now, looking into one of the twin current nodes, the input impedance will have roughly quadrupled. The one-turn loop's skimpy 5 or 6 Ω rises to approximately 25 Ω. Adding yet another such turn doubles this to about 50 Ω—a convenient match to common coax. See Fig 1. As with the folded dipole, more than four turns are unprofitable; diminishing returns raise the input impedance little faster than ohmic resistance. Note that the tuning capacitances, each about 40 pF, must be kept nearly equal for overall system balance.

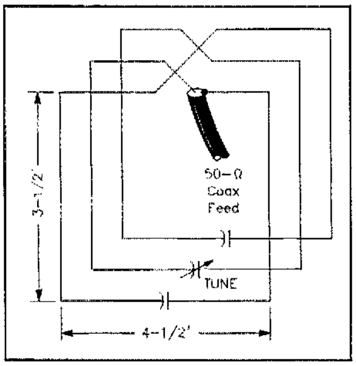


Fig 1—The three-turn loop antenna for 40 meters. Each capacitor is about 40 pF to resonate and match the antenna to a 50-Ω feed line.

Increasing Multiturn Loop Efficiency

Lots of hams use small loops in situations where they can't erect a larger antenna. Making contacts with these antennas largely depends on high radiation efficiency. Here are some things you can do to increase efficiency and effectiveness of such antennas:

- Use large-diameter copper wire. Keep the copper surface clean. Tin-plated wire is noticeably worse than copper in this application because of skin effect—most of the RF travels in the tin, which is far lossier than copper at RF.
- Use low-loss capacitors. The key here is minimizing dielectric loss, which is most favorable in vacuum- and air-dielectric capacitors. You can also parallel several capacitors to decrease losses.

Increasing the wire size used in these loops raises the antenna Q, narrowing the bandwidth and increasing the voltages and currents present. For this reason, you may have to use higher-quality capacitors than those used in the experimental versions if you increase the wire size.

Remember that high voltages can exist on a multiturn loop antenna's wires at power levels of a few tens of watts or more. Thus, it's best to keep the antenna well away from other objects.—*Roy Lewallen, W7EL, ARRL Technical Advisor*

Construction techniques, conductor material, insulation and components now become less critical because voltage peaks and current nodes are distributed throughout the system. I use ordinary zip cord, two lengths entwined, giving four available turns.[1] For proper continuity, the four wires on each side of the loop are clearly tagged at alligator-clip connectors, top and bottom centers. For 7 MHz, one turn is left open (at top *and* bottom), but will become useful on the lower bands as discussed later. A more classy construction might feature braided hook-up wire of contrasting colors. Small cardboard spreaders at the clips keep adjacent capacitors separated.

Midget receiver-type capacitors are adequate for CW power output up to 100 watts. One of these should be conveniently variable, preferably a double-gang or split-stator type, for tweaking the system to the desired center frequency as indicated by a 1:1 SWR. At 7 MHz, each capacitance is about 40 pF in my antenna, so the old war-surplus Hammarlund APC midgets do nicely. Bandwidth can approach that of a normal linear dipole if a slightly higher center-frequency SWR is accepted by stagger-tuning.

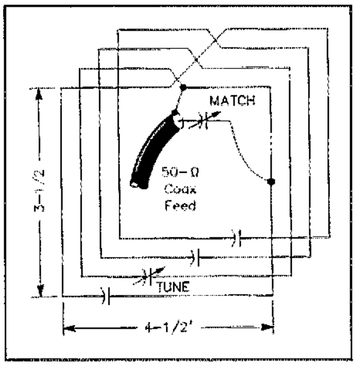


Fig 2—Extending the Fig 1 concept to 80 and 160 meters requires adding another turn to the antenna and increasing the capacitances. See the text for capacitor values.

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[1] Notes appear on page 39.

The Ingredient Called *Belief*

Even though my busy family life keeps hamming time rare, the time I spend hamming has to be spent *right*. The idea isn't just to get on the air and contact anyone anywhere—that's too *easy*. The idea is to do something a hair off-the-wall and see if anyone shares the same orbit. So I'm one of those guys who calls CQ QRP for ten minutes at 10149.5 kHz even as game clogs the band's low end—someone who looks for (and *expects*) ragchews on 3560 at noon. And so I naturally was the guy to call a 5-watt, 3 × 2 × 1 CQ at 7138 kHz at around 10 PM one night last fall when the broadcasters had long since muscled straight-thinking NovTechs off to safer terrain. And so I naturally was *of course* immediately answered by W9BRD.

It figures. We've worked on schedule many times before, but he and I—he's the father; I'm the son—have never had to prearrange each new "first contact" after changes or station setup or locale. *Chance* is too facile a word for this, but then I haven't looked too hard for a better word, either. I don't need to; we just do our individual quirky things and click.

This time, we *both* had new setups: I'd just relocated the armful of gear constituting my "shack" to the bedroom and was talking to a low dipole through 100+ feet of mongrel coax; *he* was trying out an indoor, largely below-ground loop antenna—one of the versions shrunken in this article—with just a few tens of watts, tunneling past those steamroller broadcast signals as soundly as any medium-scale ham signal from Chicago.

Enjoying the magic of yet another first contact, I hardly gave his setup a second thought. But you may wonder if such antennas really work. Gooch's Paradox explains it thus: "RF gotta go *somewhere*." Seeking to soothe the linearless, towerless masses who toil with "*only* 100 watts," Newkirk's Tiresome Chant puts it, "Successful MF/HF radio communication may proceed at astoundingly low received-signal strengths." Reduced to generic essentials, it's this: The most important single factor in success with *whatever* you use is overcoming your own disbelief.—*David Newkirk, WJ1Z*

Excellent 40-meter results indicated that the same loop, dimensions unchanged, could have a shot at 80 meters. At 3.5 MHz, it becomes truly a miniloop, only 1/16 wavelength in circumference. Adding the fourth turn (Fig 2) raises the feed impedance at a slightly higher rate than ohmic resistance, so I included it. Armed with a bargain bag of small 1-kV ceramic capacitors, I reached 3.6-MHz resonance with a value of 100 pF for each of the four capacitances. One capacitor was then replaced with an old broadcast-style, two-gang 300-300-pF variable for tweaking. I measured the feed impedance, at the center of one of the four turns, at about 18 Ω—well within the range of simple gamma matching. So another junk-box broadcast variable, three 350-pF gangs paralleled, was suspended at the loop's top center. Only about 650 pF was needed for a 1:1 match to coax after a 30-inch gamma lead was dangled to a tap at the center of one side of the feed turn. Then the system was tweakable over the entire 80-meter CW range. The remaining ceramic capacitors get slightly warm at 80 watts, which indicates significant power loss. These capacitors should be replaced with low-loss units for best performance, but I left them alone for my experiments. I easily contacted the East Coast and as far west as Arizona in the following few nights on the air.

My next inclination, as you may surmise, was to try the little gem on 160 meters. A four-foot-square transmitting loop for 1.8 MHz? I wasn't overly optimistic. Years of tinkering at W9BRD had failed to produce a decent indoor compact antenna, even of much larger size, for top-band work. But I dug into my bag of ceramics and gave it a go. The four-turn loop, now only 1/32 wavelength in circumference, resonated at about 1.8 MHz with four 350-pF capacitors. One of these was replaced with an old broadcast-type 500-pF variable for tweaking the center frequency between 1.8 and 1.85 MHz. I double- and triple-checked the measured feed impedance—16 Ω. Now I knew it would work. The gamma-match SWR dipped to 1:1 at about 800 pF using the same lead, tap and broadcast capacitor as on 80 meters. Immediate solid 40-watt CW QSOs with WN9W, KC4WWV, WK0B and W4VZB were most gratifying.

Variations and the Higher Bands

If such a 1/16-wavelength loop can function at 80 meters, the same should be true at 20 meters. So I wound a 1-foot-square, three-turn model on a cardboard box, hung it on a wooden bulkhead, and resonated it to 14.050 MHz with three 50-pF ceramic capacitors. This time, instead of the usual fudging capacitor, I spaced the turns 1 1/2 inches apart. Center-frequency tweaking thus could be done by uniformly edging the turns away from or toward each other. This spacing approach works fine for box-wound loops, but for the lower bands, I prefer closewound zip cord. Input impedance turned out to be about 11 Ω—I should have used a fourth turn—and gamma coupling worked okay, as on 80 and 160. The 1:1 feed-line dip occurred with a 120-pF series capacitor, with a gamma lead of 13 inches. In my installation, this antenna is no low-angle DX radiator, to be sure, but I've received S9 reports during many stateside QSOs. Anyone for a 3-inch-square 2-meter version?

Antennas smaller than necessary are interesting stunts that help prove a design. However, for the lowest HF bands I recommend the largest practicable circumference. For operators limited to indoor installations, floor-to-ceiling height is usually the limiting size factor. Thus, a cooler 7-foot by 9-foot job, requiring much smaller resonating capacitances, should be a natural for 80 and 160.

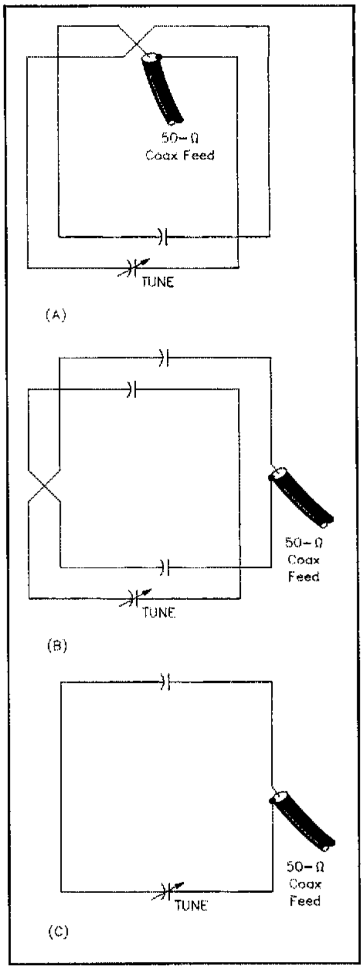
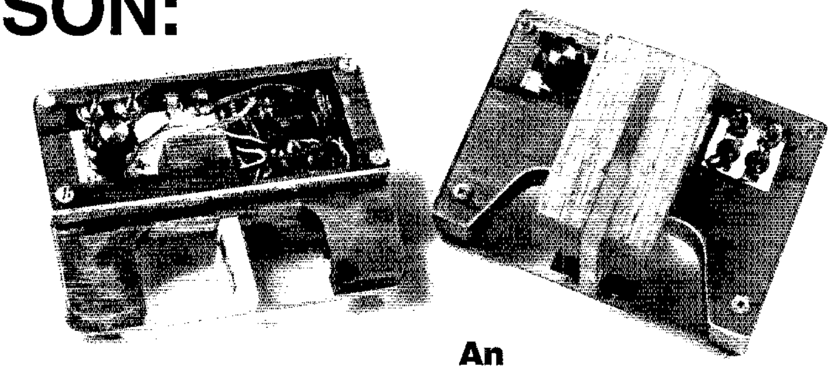


Fig 3—At A, the 30/20-meter loop; at B, the 15/17-meter version, and at C, the 10-meter variant.

Incidentally, the configurations described are quite well balanced. Although in my tests the antennas were placed within tweaking distance of the operator, no shack RFI showed up on any band at the 80-watt level. Careful symmetry helps, notwithstanding the dangling gamma lead on 80 and 160. The coax feeder should run down and away from the top-center feed point, equidistant from the vertical sides. A balun or coaxial RF choke might seem applicable, but I found such isolation unnecessary. The directivity pattern for such a loop is the usual figure-8, with most

(continued on page 39)

IROESON:



An Infrared Optoelectronic Keyer

This offspring of the IROESK straight key has no moving parts except your fingers; it's *doubly* digital.

By L. B. Cebik, W4RNL
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Knoxville, TN 37938-4443

Response to the IROESK straight key has been heartening![1] The infrared optoelectronics that take the effort out of straight keying have proved useful to many hams, especially the handicapped and those troubled by diseases such as arthritis. A friend asked if I could apply the same principle to an electronic keyer—I thought I could. "How will you perform that trick?" he asked. "With mirrors, of course," was my reply. Thus, IROESON was born.

Instead of the usual mechanical-paddle input, IROESON uses a pair of infrared LEDs and phototransistors. This permits a great degree of mechanical design flexibility. The prototype described here can be manipulated by the first two fingers (or the thumb and forefinger) of either hand to key dots and dashes in a vertical, rather than horizontal, motion. Vertically aligned "paddles" require a little more structural work, but when keying no longer requires you to overcome external resistance from a spring-loaded paddle, you no longer need the strength of your opposed thumbs to operate the keyer.

If you haven't read the earlier article, do so. The following circuit and construction notes assume that the earlier article is available to you.[2]

The IROESON Circuit

Two keyer versions are shown schematically in Figs 1 and 2. With these two circuits, all of the Curtis chip variations are provided for. (See the sidebar, "Goodbye, Older Chips!") Although the keyers shown in the title-page photo are built on perfboard using the older 8044 IC, PC boards that accept the 16, 18 and 20-pin chips are available.[3] For experimental purposes and compactness, I omitted the sidetone and weight adjustments. The output circuit is a simple transistor switch that is compatible with my solid-state rigs. If you want to reinstate the sidetone, weight control, or relay output, refer to a recent edition of *The ARRL Handbook*, Chapter 29, for circuit details.

The infrared optoelectronic paddle substitutes consist of two identical circuits akin to the simplest version of IROESK. All of the electronic parts except the keyer chip are available at Radio Shack. One difference between the IROESK and IROESON circuits is use of a lower-value (1-kΩ) resistor in the LED line. Employing mirrors requires more illumination to ensure proper phototransistor operation without finicky alignment. When your finger breaks the invisible light beam, the phototransistor stops conducting, the 2N3904 transistor saturates and grounds the dot or dash line to the 8044. It's that simple.

Goodbye, Older Chips!

Early in 1992, Curtis Electro Devices announced a change in their keyer chip line-up: *Production of all but the 8044ABM ceased!** Until then, Curtis keyer ICs had been available in three models: the 8044, 8044AB and 8044ABM, in 16, 18 and 20-pin packages, respectively, each offering different features. (The older 8043, which preceded the 8044, used a dot/dash symmetry control.) Once existing stocks are depleted (if they aren't already), the 8044 and 8044AB will no longer be available. (By mid 1992, at Curtis, no 8044ABs were available.) Chances are, there are various forms of the 8044 ICs all over the world (probably some 8043s, too!), either in use or lying in a piece of foam waiting for the call to arms.

It's important to note that the first 8 horizontally paired pins of the 16- and 18-pin ICs are *identical*—except the 18-pin IC has two additional pins at the back end of the package used for speed metering. The pin-outs of the 20-pin 8044ABM IC, however, are *incompatible* with those of the earlier 8043, 8044, 8044A, 8044B, 8044M and 8044AB ICs. Curtis produces both DIP and surface-mount versions of the 8044ABM.—*N1FB*

*Curtis 8044 CMOS keyer ICs can be obtained from: Ocean State Electronics, PO Box 1458, 6 Industrial Dr, Westerly, RI 02891, tel 800-866-6626 or 401-596-3080, fax 401-596-3590; Mouser Electronics, 2401 Hwy 287 N, Mansfield, TX 76062, tel 800-346-6873, 817-483-4422, fax 817-483-0931.

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[1] Notes appear on page 39.

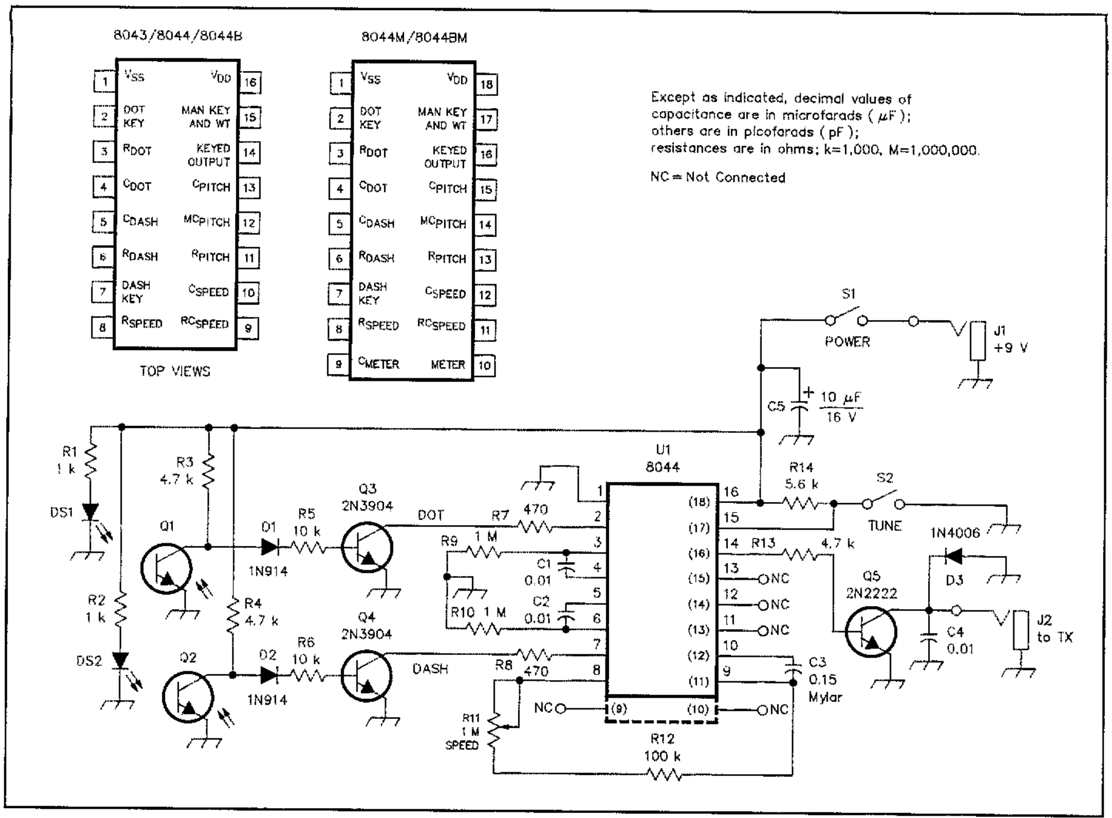


Fig 1—Schematic of the 16- and 18-pin keyer circuit. Part numbers in parentheses are Radio Shack; equivalent parts can be substituted. Unless otherwise specified, resistors are ¼-W, 5%-tolerance carbon-composition or film units.

C1, C2, C4—0.01-µF, 50-V disc-ceramic (272-131).
C3—0.15-µF, 50-V Mylar capacitor (you can use a parallel combination of one 272-1068 and one 272-1069 capacitor).
D1, D2—1N914, 1N4148 silicon switching diode (276-1122).
DS1, DS2—Infrared LED, SY-IR53L or equivalent (276-143).
J1, J2—Two-conductor phone jacks (274-251).
Q1, Q2—Infrared phototransistor, SDP8403-301 (276-245).
Q3, Q4—2N3904 NPN switching transistor (276-2016).
Q5—2N2222 NPN switching transistor (276-2009).
R11—1-MΩ linear-taper potentiometer (271-211).
S1, S2—SPST miniature toggle switches (275-624).
U1—8044 series 16- or 18-pin Curtis Electro Devices CMOS keyer chip (see the sidebar "Goodbye, Older Chips!").
Misc: One 16- or 18-pin DIP IC socket (276-1998 or 276-1992, respectively), wood, Plexiglas, screws, stainless-steel tape, perfboard (276-1396) or PC board, knob.

Physical Construction

IROESON construction follows—with only minor variations—the same pattern I used for his sire: a combination of wood and Plexiglas. The keyer parts fit into a piece of three-quarter-inch-thick oak sandwiched between two pieces of eighth-inch-thick Plexiglas. The top piece covers only the rear part of the board, while the bottom piece forms a base for the whole assembly. For some suggestions on other methods of construction, see the IROESK article.

Fig 3 provides dimensions for an IR-paddle/keyer enclosure suitable for housing the 20-pin 8044ABM FAR Circuits PC board. (This is quite similar to one of the prototypes shown in the title-page photo.)

The chief difference between the IROESK and this keyer is that we need *two sets* of LEDs and phototransistors. As Fig 3 and the photographs show, the LED and phototransistor pairs are mounted at angles to each other so that the LED beam bounces off a reflective surface in the center of the finger area. This arrangement requires a little more planning and more careful routing than the simple IROESK. The LED and phototransistor axes should intersect at the surface of the center barrier. Be sure to consider the sides on which you want the dots and dashes before you connect the IR lines to the main circuit board.

The center barrier can be any strip of material up to about a quarter of an inch thick. I used wood screws to hold my plastic barrier to the Plexiglas base. My mirrors are small pieces of adhesive-backed stainless-steel tape, available from many hardware outlets. You can use thin pieces of real mirrors: Plastic and metal mirrors are most easily cut. Allow for any additional thickness of your mirror when planning the channels for the LEDs and phototransistors.

Mount the LEDs and phototransistors on small strips of perfboard and make a channel in the wood into which they fit (see Fig 4). Allow room for the three wires connecting each assembly to the main board.

Checkout

Test and align the IROESON input circuits *before* inserting the IC into its socket. With power applied, measure the resistance from the paddle transistor collectors to ground. Adjust the LED and phototransistor alignment until the resistance goes from near infinity to near zero as you interrupt the beam with your finger. You can also track the voltage at either of the 1N914 diodes. If the

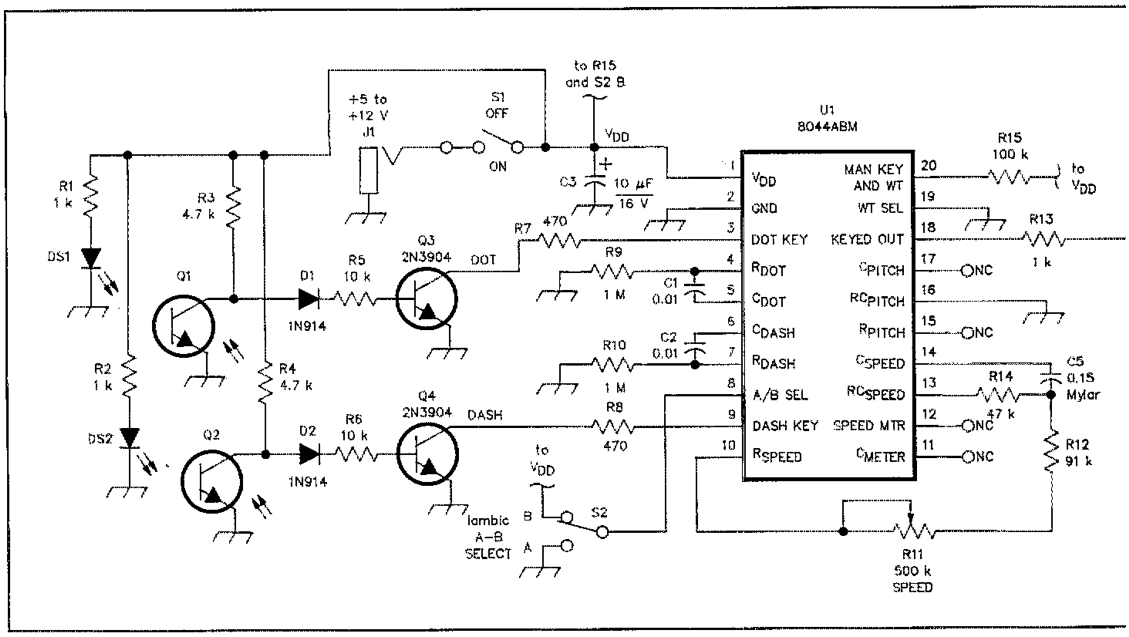


Fig 2—Schematic of the 20-pin 8044ABM keyer circuit. Part numbers in parentheses are Radio Shack; equivalent parts can be substituted. Unless otherwise specified, resistors are ¼-W, 5%-tolerance carbon-composition or film units. See Fig 1 for repeated components.

Q5—2N4401 (276-2058).

R11—500-kΩ linear-taper potentiometer (271-211).

U1—8044ABM 20-pin Curtis Electro Devices CMOS keyer chip (see the sidebar "Goodbye, Older Chips!").

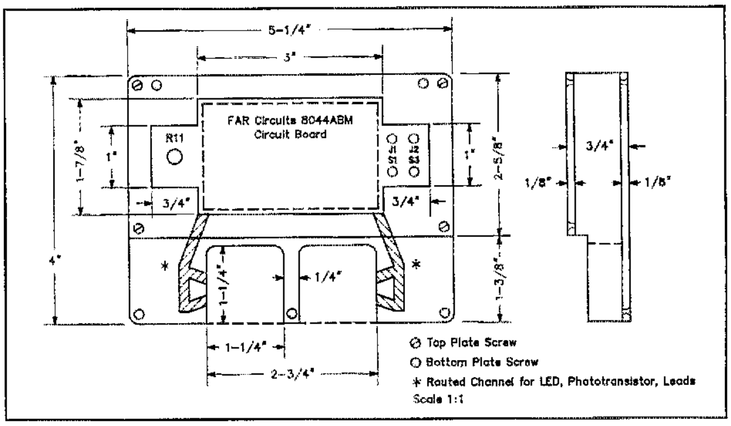


Fig 3—Dimensions for an enclosure that accepts the FAR Circuits 8044ABM PC board. From left to right, the cut-out areas hold the SPEED control, the circuit board and the jacks and switches, respectively.

voltage does not drop well below 1 volt, you may need a shinier mirror, or better alignment.

Next, insert the 8044 chip and check the keyer operation. With the values shown, the keyer has a range of about 8 to 50 WPM. The transistor output will key most modern solid-state rigs, most of which source relatively little key-line current. My Yaesu transceivers list the following key line specs: FT-757, key-up voltage, 5; key-down current, 0.5 mA; FT-767, key-up voltage, 4.5; key-down current, 2 mA. If your rig's key-line current is greater, seriously consider adding a relay, as shown in the *Handbook* circuit.

If you want to install a keyer inside your rig, you can reduce the front-to-back dimension of the keyer by using just the paddle portion of the circuits of Figs 1 and 2. A three-wire line, configured according to your rig's instruction book, lets IROESON key your transmitter as you would with any set of paddles. A 0.01-µF disc-ceramic capacitor from each output line to ground provides a modicum of RF bypassing.

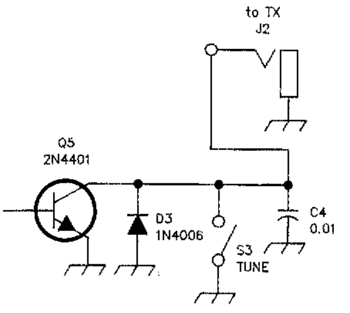
IROESON draws over 20 mA at 9 volts. Therefore, you should power it from either a 9-volt dc wall module, or from an auxiliary dc power jack on your rig, not a 9-V battery.

Using IROESON

Using the thumb and forefinger to operate an electronic keyer paddle derives from the operation of bugs (semi-automatic keys) and sideswipers. With IROESON, you can use any fingers you choose. If you can drum a tabletop with your fingers, you can master the flat version of IROESON. Those who have not yet used a paddle may find learning to use IROESON easier than learning to use standard paddles of any design! A light touch and a relaxed wrist are the prime ingredients to good sending.

With the flat keyer, you can use either your thumb and forefinger (as I do), or your first two fingers (as does my XYL, N4TZP). Whichever feels more comfortable after a 15-minute practice session is likely to be more comfortable in the long run. I suppose it's even possible to use the forefinger of each hand, as do the multitude of hunt-and-peck typists!

The vertically assembled version of the keyer requires more standard keying motions. I tend to use my thumb and second finger, with the forefinger resting on top of the vertical assembly. That technique may



Except as indicated, decimal values of capacitance are in microfarads (µF); others are in picofarads (pF); resistances are in ohms; k=1,000, M=1,000,000.

NC = Not Connected

only be an artifact of the fairly tight quarters I created for keying. You may want to change the assembly design to suit your favorite keying posture.

Summary

Kicking the "brass-and-spring" habit comes hard after many years. But in an age of labor-saving devices, IROESON may be close to the ultimate for manual CW keying. Even electric and electronic typewriter keys load the fingers more than this keyer does! CW will get easier only when someone perfects a voice-to-Morse converter. (With voice-recognition and computer technology already quite advanced, that may well be an idea just awaiting implementation. If you accomplish this, we at *QST* would like to know about it.—*Ed.*)

Notes

[1] L. Cebik, "IROESK—An Infrared Optoelectronic Straight Key," *QST*, Feb 1992, pp 30-34.

[2] Back issues and photocopies are available from the ARRL. Contact the Technical Department Secretary at 203-666-1541; fax 203-665-7531.

[3] Because of the Curtis line-up change, and recognizing that many of the older chips are still available throughout the world, FAR Circuits makes available *three* PC boards. Two boards accommodate the circuits using the IR paddle input, one accepting the 16- and 18-pin ICs, the second is for the 20-pin IC. The third board is for a mechanical-paddle version of the 20-pin 8044ABM keyer *not described here,* but similar to the one in recent editions of *The ARRL Handbook.* PC-board pads are provided to allow access to *all* IC pins and incorporate all the features of a particular chip. PC boards are available from FAR Circuits 18N640 Field Ct, Dundee, IL 60118-9269. PC-board prices: 8044/8044AB, IR input, $4.75; 8044ABM; mechanical-paddle input $4.25. Please add $1.50 for shipping.

PC-board templates, part-placement diagrams and a schematic of the unpublished paddle-input keyer are available from ARRL HQ. Please send a business-size SASE and address your request for the CEBIK IROESON KEYER PC-BOARD TEMPLATE to: the Technical Department Secretary, ARRL, 225 Main St, Newington, CT 06111.

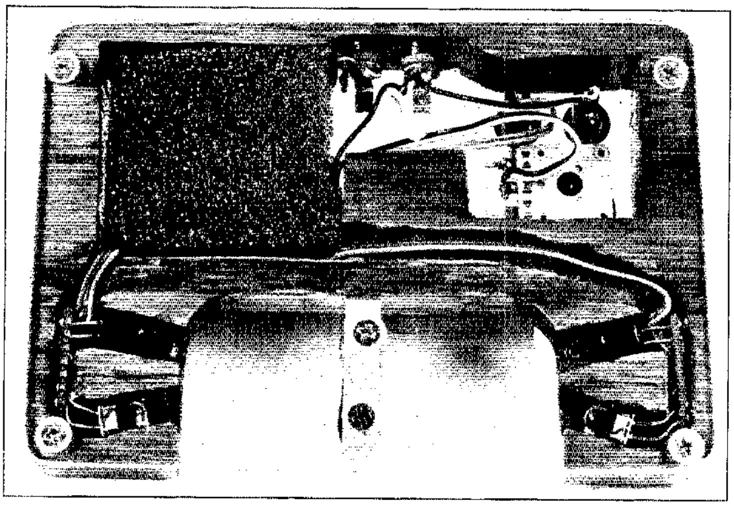


Fig 4—A bottom view of one of the prototype keyers. Note the channels for the wiring to the IR LEDs and phototransistors. Small pieces of perf board hold the LEDs and phototransistors in channels that are routed at an angle to the "paddle."

Honey, I Shrunk the Antenna!

(continued from page 35)

radiation in the plane of the loop. However, there will be sharp nulls perpendicular to the loop's plane, which you should take into account when placing your antenna.

My prime purpose here is to treat loops of ⅛-wavelength circumference or smaller, but you may desire to use the same 3½-foot by 4½-foot dimension on the higher HF bands, as I do. The hook-ups shown in Fig 3 perform well on 30 through 10 meters. Each provides a direct match to 50-Ω coax. True, almost any old piece of wire gets QSOs on 10.1 MHz and above, but a balanced loop radiator is hard to beat for indoor hamming. A few hints and observations:

- At 10.1 and 14 MHz (Fig 3A), just two turns are sufficient to reach a feed impedance of about 50 Ω. The two approximately 20-pF capacitance values are slightly higher for the 30-meter band.
- For 17 and 15 meters (Fig 3B), the loop must be made electrically smaller by splitting it symmetrically. Two split turns and four equal capacitances then bring a close 50-Ω match. Note that the feed point is moved to the center of one of the four vertical sides. Capacitances are in the 15-pF range.
- On 12 and 10 meters (Fig 3C), one split turn will suffice for a near-50-Ω match, with two capacitors of about 12 pF. Here, split-stator midgets can ease adjustment by minimizing hand capacitance, but it's a fancy junk box that includes them. The old APC midgets, whose rotors have little more mass than their stators, will maintain enough system balance.

Cautions and Conclusions

Like any indoor antenna, the loops I describe here generate substantial electromagnetic fields in operation. Thus, they have considerable potential to generate RFI. For this reason and to prudently avoid placing yourself or others in large RF fields, you should keep all antennas as far from consumer electronic devices and people as possible, and use the least RF power necessary to conduct the desired communications. The current editions of *The ARRL Handbook* and *The ARRL Antenna Book* cover this subject in more detail.[2]

A final comment: For overall results with compact antennas, like almost any skyhook, the higher above ground the better. In my case, radiation and reception strongly favor higher skywave propagation angles. All antenna configurations I describe here were tested and operated in a cellar ham shack, where half of the system is below ground level!

Notes

[1] See the sidebar, "Increasing Multiturn Loop Efficiency," for some ideas on optimizing such an antenna's performance.

[2] See Chapter 36 of the *Handbook* and Chapter 1 of the *Antenna Book.*

Reverse-Polarity Protection for Your Gear

Reversing the power-supply polarity to your equipment can be harmful to its health. These simple and inexpensive protection hints can spare you trouble and save you money.

By Michael A. Covington, N4TMI
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Athens, GA 30606

If you've ever destroyed a transistor radio by momentarily connecting its power supply backward, you know that polarity reversal can easily fry semiconductors. Even if you trust yourself to plug batteries in the right way, you're still at the mercy of wall transformers (Fig 1) and accidental connections to other sources of power. In my view, *any* piece of gear worth building or owning is worth protecting against reversed polarity. I'll show you how to accomplish this simply and inexpensively.

Diode Protection

Perhaps the easiest way to keep dc voltage from being applied backwards to your gear's circuitry is to use a diode in series with the equipment's dc-input connections (Fig 2A). A cheap and time-honored means of protection, the lowly diode, adds little cost to the circuit, probably no more than 10 cents. The diode must, of course, handle the maximum current drawn by the load, and its peak inverse voltage (PIV) must be greater than the highest expected applied inverse voltage. A 1N4001 will handle currents up to 1 A at 50 PIV.

Each silicon diode has an intrinsic forward voltage drop of approximately 0.6. Put 12 volts in, and you get 11.4 volts (or less) out. The diode-voltage drop is a problem only if you need those few tenths of a volt, as, for instance, with very low-voltage supplies (3 volts), or if you're supplying voltage to a transistorized transmitter's output stage where a 10% drop in collector voltage causes a 20% drop in output power.

The easiest way to eliminate the diode-voltage drop is to connect the diode *in parallel* with the load instead of in series with it (Fig 2B). That way, the diode normally doesn't conduct, but if the power supply is reverse connected, the diode conducts heavily and blows the fuse or trips a circuit breaker. Naturally, this requires you to use a fuse or circuit breaker. (You do, don't you?) Also, the diode must be able to handle substantially more current than the fuse, otherwise the diode may blow before the fuse does! As a rule of thumb, triple the diode's current rating, ie, for a 1-A fuse, use a 3-A diode.

Fig 1—If you're not (yet) convinced that your circuits need reverse-polarity protection, look closely at the two wall-transformer labels. Do you see the important difference between the two supplies?

Aside from the obvious voltage difference, the output connector pins have opposite polarities.

FET Protection

Another way to eliminate the diode-voltage drop is to use a power MOSFET instead of a diode as the switching element (see Fig 3).[1] When power is applied correctly, the FET conducts and becomes the equivalent of a low-value resistor. If the power supply is applied in reverse, the FET doesn't conduct and no current flows.

Fig 3A shows how to use this approach with cheap, readily available N-channel MOSFETs; Fig 3B shows how to similarly use a P-channel MOSFET. Although P-channel MOSFETs are presently more expensive than N-channel MOSFETs, a P-channel device is often preferred because it interrupts the *positive* supply lead rather than the *negative* one; this lets you use the negative

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[1] Notes appear on page 41.

Table 1
Power MOSFETs for Polarity Protection

Type	*Polarity*	*Max. Voltage Rating (V)*	*Max. Current Rating (A)*	$V_{GS(th)}$ *(V)*	$R_{DS(on)}$ *(Ohms)*	*Max. Current (A) Fig. 3 Circuit*
IRF510	N-channel	100	4	2-4	0.54	1
IRF511	N-channel	80	4	2-4	0.54	1
IRF520	N-channel	100	8	2-4	0.27	2
IRF530	N-channel	100	14	2-4	0.16	3.4
IRF540	N-channel	100	27	2-4	0.09	6
IRF9Z10	P-channel	50	4.7	2-4	0.5	1
IRF9521	P-channel	80	6	2-4	0.6	1

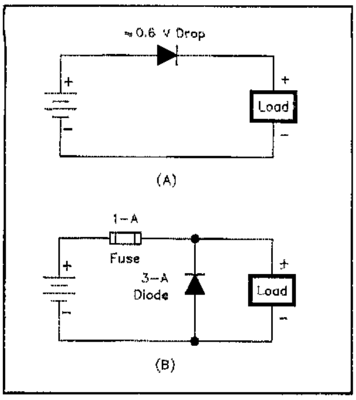

Fig 2—At A, a diode in series with the load protects against polarity reversal, but reduces the input voltage by approximately 0.6. At B, the shunt diode protects the circuit by blowing the fuse if the input-voltage polarity is reversed. The diode must be able to handle more current than the fuse (see text).

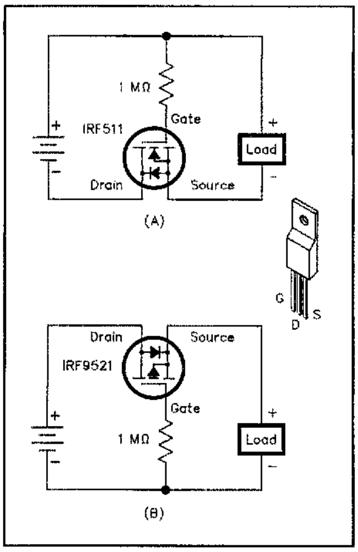

Fig 3—The N-channel MOSFET input-voltage protection circuit (A) turns power on only when the input-voltage polarity is correct. The voltage drop across the MOSFET is low. Using a P-channel MOSFET (B) puts the input-voltage interruption in the positive line. Yes, these FETs are connected "backwards"—the text explains why.

rail as ground in the generally more conventional way.

The FET you use must be an enhancement-mode device, with a turn-on voltage ($V_{GS(th)}$) substantially lower than the power-supply voltage, and with an on-resistance ($R_{DS(on)}$) low enough that the voltage drop is negligible. Because the FET operates as a switch, it dissipates little power and no heat sink is required. MOSFETs suitable for this purpose are listed in Table 1. Right now, power FETs cost about $2 to $5 each, but prices are dropping.

Notice that the current flows backward through the FET (from drain to source). That's okay; FETs are inherently bidirectional devices. Figs 3A and 3B show a diode inside each FET. That's the *parasitic diode* formed by the source and substrate of the FET. Now you see why we have to use the FETs backwards: If we didn't, that diode would be incorrectly oriented and would allow reverse-polarity current to flow.

There are two situations in which the parasitic diode conducts: (1) When the supply voltage is less than about 4, so that the FET doesn't turn on; (2) when the current through the FET exceeds $0.55/R_{DS(on)}$ (the "maximum current for this circuit" in Table 1), because then the voltage across the diode becomes high enough to start conduction. These situations aren't dangerous to the FET; the diode can handle just as much current as the FET itself. It's just that if the diode is conducting, you're not getting the benefit of the FET, whose voltage drop can be much lower than that of the diode.

Protected Voltage Regulators

Some of the newer voltage-regulator ICs have built-in reverse-polarity protection. Table 2 lists some widely available types; Fig 4 shows a practical circuit.[2] These regulators are often known as *PNP* voltage regulators, and the capacitors at the input and output are required for stability.

The main selling point of PNP voltage regulators is that they have lower drop-out voltage than earlier types. For example, the 7812 has a 2-volt drop-out rating. That means that if you want 12 volts out, you have to supply at least 14 volts. But the newer LM2940CT-12 has a drop-out voltage of only *0.5;* you can get 12 volts out with just 12.5 volts in—a real convenience when using battery power!

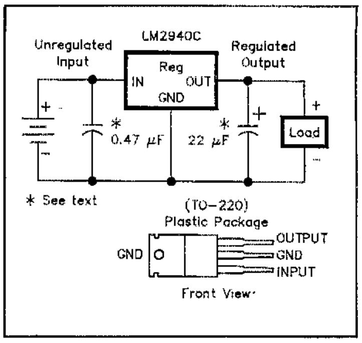

Fig 4—An LM2940, or similar low-voltage-drop PNP voltage regulator, provides built-in reverse-polarity protection. With these regulators, the capacitor shown at the input of the device is required if the distance between the regulator input and the power-supply filter capacitor is greater than 6 inches. For stability, the capacitor connected to the output pin must have a value of at least 22 µF, an equivalent series resistance (ESR) of less than 1 ohm, and be located as close as possible to the regulator output pin. The capacitor must also be rated over the same operating-temperature range as the regulator employed.

Summary

Do *all* projects need reverse-battery protection? Arguably, yes. But in the end, the choice is yours. The leading candidates for such protection are those circuits that employ expensive semiconductors. It would be costly (as well as embarrassing) if you connect the battery backward! Other candidates for protection are those circuits that are used under emergency field conditions (with a high risk of accidental misconnection). A 10-cent diode or a $2 FET is cheap insurance for expensive ICs and PC boards.

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Summary

There you have it: A method that allows you to make inexpensive PC boards with photocopies. I generally use surplus circuit boards which I can buy for a few cents per square inch. This means that I can make a single-sided PC board for less than a dollar. Once I became practiced at using the method, I found that I could reliably produce a PC board on the first try. But it took practice. If authors start publishing PC-board patterns as viewed from the component side, we'll be able to use this method without having to go through an intermediate transparency to reverse the image.

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Notes

[1] C. Nord, "More on PC Boards," *QST*, Aug 1977, Hints and Kinks, p 77; P. Pagel, "G.C. Electronics Lift-It Transfer Sheets," *QST*, Sep 1981, Product Review, p 49; L. Wolfgang, "PC Board Patterns the Easy Way," *QST*, Apr 1982, Hints and Kinks, p 53; T. Desaulniers, Jr, "A Drill Press for Printed-Circuit Boards," *QST*, Jan 1984, Hints and Kinks, p 46; J. Stinson, "A Bubble Etcher for PC Boards," *QST*, Nov 1984, pp 45-47, 56; R. Grabowski, "Etch-Resist Pens for Homemade Circuit Boards," *QST*, Sep 1988, Hints and Kinks, p 44; P. Atkins, "Isolated-Pad Drills Still Available," *QST*, Jul 1989, Hints and Kinks, p 39.

Tek-200 film, which uses a PC-board-making process similar to that described in this article, is available from several sources including Ocean State Electronics, PO Box 1458, 6 Industrial Dr, Westerly, RI 02891, tel 800-866-6626, 401-596-3080; fax 401-596-3590.—*Ed.*

DynaArt Designs has a toner-transfer system that can be used to produce plastic faceplates and decals as well as PC boards. DynaArt Designs, 3535 Stillmeadow Ln, Lancaster CA 93536, tel 805-943-4746. These materials are also available from All Electronics, PO Box 567, Van Nuys, CA 94108, tel 800-826-5432, 818-904-0524, fax 818-781-2653 and DC Electronics, PO Box 3203, Scottsdale, AZ 85257, tel 800-467-7736, 800-423-0070, 602-645-7736, fax 602-994-1707.—*Ed.*

[2] T. Miller and P. Pagel, "PC-Board Production in the ARRL Lab," *QST*, Apr 1988, pp 15-18.

[3] PC-board etching materials are available from most electronic product mail-order distributors. Your local Radio Shack store carries the materials, too: PC-board kit (276-1576); etch-resistant pen (276-1530); etchant, 16-oz container (276-1535).

Bibliography

W. Richardson, "Practical Printed-Circuit Boards," *Modern Electronics*, Aug 1989, pp 14-17, 82.

D. Lancaster, "PC-board Breakthrough," *Radio Electronics*, Hardware Hacker, Dec 1989, pp 68-76.

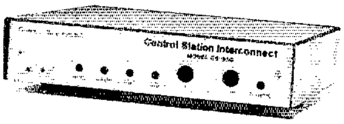
J. Axelson, "Making Printed-Circuit Boards Without Photography," *Modern Electronics*, Dec 1989, pp 16-21.

T. Byers, "Printed-Circuit Boards From Your PC," *Modern Electronics*, Feb 1991, pp 18-73, 74-75.

New Products

SIMPLEX VOX PATCH

◊ The CS-900 is the first low-cost voice-activated (VOX) simplex phone patch to incorporate standard features such as a ½-second electronic voice delay (EVD), built-in programming keyboard with digital display, 90-memory speed dialer, last-number redial, call-progress tone detection, user-programmable CW ID and more. The CS-900 is VOX-operated and doesn't use sampling. There are no sampling noises or interruptions; just smooth, natural audio. The built-in EVD prevents syllable and word clipping caused by slow-switching transceivers. It interfaces with most transceivers, with the only connections required to the mike and speaker jacks. Suggested list price $379. Connect Systems Inc, 2064 Eastman Ave, #113, Ventura, CA 93003; tel 805-642-7184, fax 805-642-7271.



DUAL-BAND "INVISIBLE" ANTENNA

◊ The neighbors will never know you're a ham with the VT-27 dual-band Ventenna for 2 meters and 70 cm. This model follows the original 2-meter single-band Ventenna design, with an antenna sealed between two layers of ABS plastic in a hollow tube that slips over your rooftop plumbing vent pipe. This durable 36-inch-long antenna is virtually indestructible and undetectable by observers. It comes with an attached 2.5-foot length of Belden high-performance 8240 (RG-58) coaxial cable that connects to your equipment with an SO-239 inline connector. Retail price $74.95. Mark Forbes, KC9C, The Forbes Group, PO Box 445, Rocklin, CA 95677; tel 800-551-5156, voice or fax 916-624-7069.

Strays

RADIO-CONTROL NET

◊ The R/C Flyers Net meets on 3.933 MHz every day at 1100-1200Z (following the Southern Coffee Club Net) to discuss all aspects of R/C model aircraft. Fred Lomax, N4KYG, 204 S Claiborne St, Goldsboro, NC 27530; or David Rose, WD4DJA, 817 Hawthorne Rd, Shelby, NC 28150.

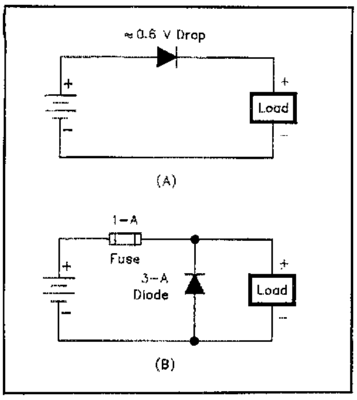


Fig 2—At A, a diode in series with the load protects against polarity reversal, but reduces the input voltage by approximately 0.6. At B, the shunt diode protects the circuit by blowing the fuse if the input-voltage polarity is reversed. The diode must be able to handle more current than the fuse (see text).

rail as ground in the generally more conventional way.

The FET you use must be an enhancement-mode device, with a turn-on voltage ($V_{GS(th)}$) substantially lower than the power-supply voltage, and with an on-resistance ($R_{DS(on)}$) low enough that the voltage drop is negligible. Because the FET operates as a switch, it dissipates little power and no heat sink is required. MOSFETs suitable for this purpose are listed in Table 1. Right now, power FETs cost about $2 to $5 each, but prices are dropping.

Notice that the current flows backward through the FET (from drain to source). That's okay; FETs are inherently bidirectional devices. Figs 3A and 3B show a diode inside each FET. That's the *parasitic diode* formed by the source and substrate of the FET. Now you see why we have to use the FETs backwards: If we didn't, that diode would be incorrectly oriented and would allow reverse-polarity current to flow.

There are two situations in which the parasitic diode conducts: (1) When the supply voltage is less than about 4, so that the FET doesn't turn on; (2) when the current through the FET exceeds $0.55/R_{DS(on)}$ (the "maximum current for this circuit" in Table 1), because then the voltage across the diode becomes high enough to start conduction. These situations aren't dangerous to the FET; the diode can handle just as much current as the FET itself. It's just that if the diode is conducting, you're not getting the benefit of the FET, whose voltage drop can be much lower than that of the diode.

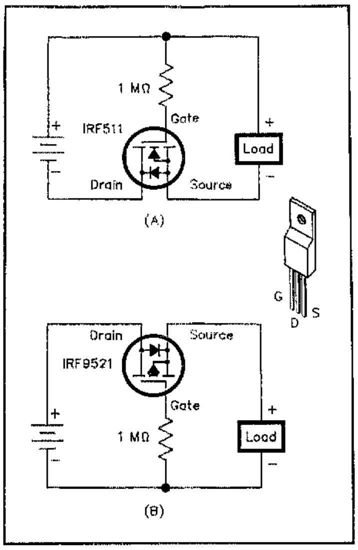
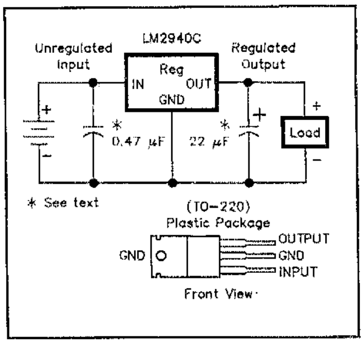


Fig 3—The N-channel MOSFET input-voltage protection circuit (A) turns power on only when the input-voltage polarity is correct. The voltage drop across the MOSFET is low. Using a P-channel MOSFET (B) puts the input-voltage interruption in the positive line. Yes, these FETs are connected "backwards"—the text explains why.

Protected Voltage Regulators

Some of the newer voltage-regulator ICs have built-in reverse-polarity protection. Table 2 lists some widely available types; Fig 4 shows a practical circuit.[2] These regulators are often known as *PNP* voltage regulators, and the capacitors at the input and output are required for stability.

The main selling point of PNP voltage regulators is that they have lower drop-out voltage than earlier types. For example, the 7812 has a 2-volt drop-out rating. That means that if you want 12 volts out, you have to supply at least 14 volts. But the newer LM2940CT-12 has a drop-out voltage of only *0.5;* you can get 12 volts out with just 12.5 volts in—a real convenience when using battery power!



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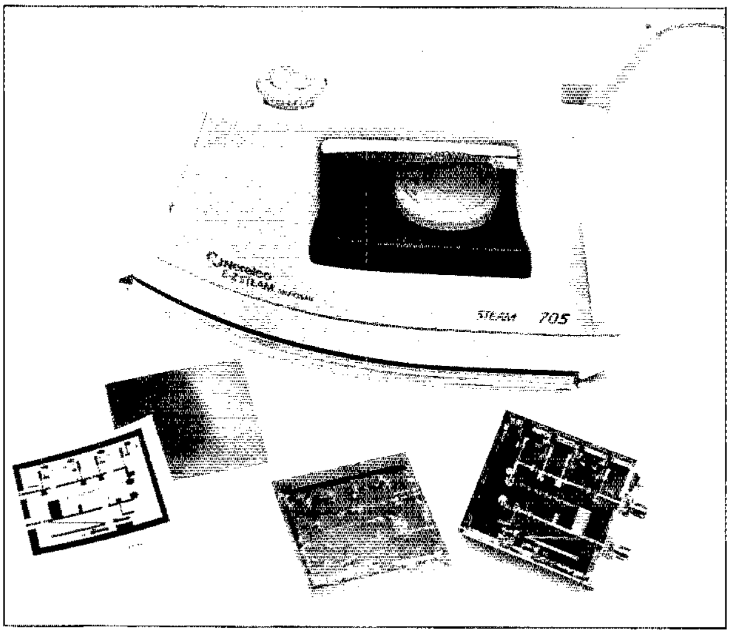
Ironing Out Your Own Printed-Circuit Boards

Here's how you can make professional-looking PC boards with little trouble and minimal cost.

By John Grebenkemper, KI6WX
Tandem Computers
19333 Vallco Parkway
Cupertino, CA 95014



(photos by Kirk Kleinschmidt, NT0Z)

Many construction projects in *QST* and other magazines contain PC-board etching patterns as part of the article, or offer them by mail. Quite often, you can order ready-made PC boards for the project, but if they're not available and you want your project neatly arranged on a PC board, you have to find some means of making the board yourself.

Variety is the Spice of Life

Over the years since PC boards came to be, many methods of making them yourself were developed: the sharp-knife-and-hot-soldering-iron approach, the isolated-pad technique, etch-resist pens of one kind or another, rub-on transfers, adhesive and masking tapes, photographic film and photosensitive materials, to name a few.[1] Somewhere all of these methods are still in use. For good measure, throw in the schematic-capture/PC-board artwork software packages that help design the layout.[2]

Low-Cost Approach

One of the inexpensive ways of making PC boards is to use etch-resistant ink. Many hobbyists hand-draw the circuit on a piece of bare copper board using a pen that dispenses an etch-resistant ink. For simple circuits, this is easy to do and works fine, even though the finished product may not look very professional. Also, the would-be board maker may find it necessary (under some circumstances) to draw a *mirror image* of the pattern, otherwise the finished board comes out backwards. For complicated circuits, the pen-and-ink method is tedious and greatly increases the potential for mistakes.

High-Cost Approach

A more costly alternative is to use a photographic technique: Coat the board with a light-sensitive etch-resistant material, expose the board to light while covered with a transparency of the circuit-board layout, develop the board, then etch it. This method produces high-quality boards even with complicated patterns. For small-quantity production, however, it's time-consuming and a tad expensive: A single board can cost more than 50 cents per square inch!

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But Wait!

If you'd like to combine relative production simplicity, low cost and produce professional-looking boards, I've got a deal for you! I'll show you how to produce boards of a quality approaching that of the photographic technique and at a cost almost as low as the pen-and-ink approach. This method makes it easy to produce any number of single-sided boards—even those with intricate patterns and fine traces.

Photocopier Etching?

The plasticized toner used in plain-paper photocopiers and laser printers is highly resistant to PC-board etching solutions. If you could pass a PC board through a plain-paper copier, you could print the etching pattern on the copper-clad board, then etch it! Unfortunately, photocopiers don't feed PC-board material as well as they do paper....

Where There's a Will, There's a Way

Ah! But, what if you *first* transfer the etching pattern *to paper* and *then* to the copper-clad board! How? By using a process that naturally, unknowingly (and sometimes frustratingly) occurs frequently with copied material. Have you ever placed a photocopy of something into one of the photo-display windows in your wallet? About a month or so later, you probably discovered that some of the toner from the photocopied material had transferred to the display window. To make PC boards, you can employ a similar process to transfer the toner from a paper copy of your etching pattern to the copper-clad circuit board. You do, however, want to control the process more efficiently—and

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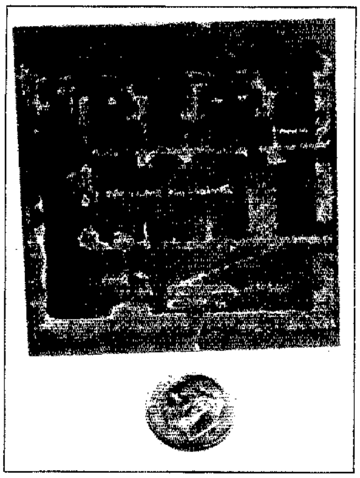
[1] Notes appear on page 44.

rapidly—so don't use a wallet.

The toner used in photocopiers starts out as fine, black particles. Through a relatively complex process using electrical charges, these particles are transferred to a sheet of paper. The particles are melted onto the paper through a combination of heat and pressure. The melting point of the toner is relatively low, only a few hundred degrees Fahrenheit.



This unetched board has had most of the paper (the white, fuzzy-looking stuff) cleaned off.

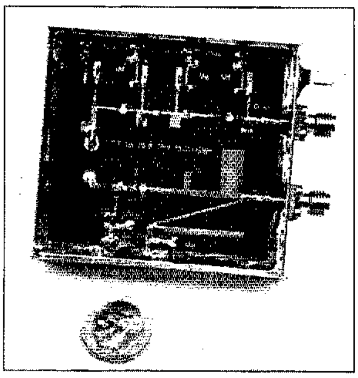
The method is quite simple; the title-page photo gives you an idea of the process. You place a photocopied PC pattern so that the toner is in contact with a copper-clad board. Next, you apply an ordinary household clothes iron to the other side of the paper. You apply heat and pressure until the toner melts and fuses to the copper and the paper. All you then have to do is *remove the paper* by dissolving it in a solution of water and bleach. What remains is a mirror image of the original copy—the *correct* image!

You may not get a perfect transfer of the toner from the copy to the copper-clad board; some pieces of toner may flake off. If that happens, use an etch-resistant-ink pen to touch up these sections. Then, etch the board with a standard ferric chloride[3] or other etchant. The exact procedure followed is very important. Altering the steps can significantly alter the quality of the transfer. The step-by-step procedure for this method is given in the sidebar, "Thirteen Steps to Iron-On PC Boards."

What Sort of Quality Can You Expect?

This process can produce PC boards having trace widths of less than 0.02 inch—more than adequate for most amateur PC-board applications. The method described in the sidebar gives the basic parameters and should work well for most people. If you experience problems, vary the parameters based on the following suggestions.

If the toner doesn't adhere to the copper board, use a higher iron temperature, or longer contact time. Make certain that the copper surface is clean: Toner doesn't like to stick to oily surfaces, so keep your dirty mitts off the board! (Use cotton gloves.—*Ed.*) If the trace width broadens excessively, use a lower iron temperature or a shorter contact time. Excessive trace broadening doesn't seem to be a significant issue when paper is used as the transfer medium. If the toner isn't sticking at the edges or corners of the board, concentrate on applying additional contact time to these areas.



How good is this method? Judge for yourself. The board in this photo is for a 2.556- to 10.224-GHz multiplier. You can see a number of chip capacitors and resistors and three GaAsFETs on the board. Note the board's size compared to the dime.

Thirteen Steps to Black-Belt Iron-On PC Boards

Here's a step-by-step procedure for making a PC board the photocopier way:

1. Most PC-board etching patterns are presented as viewed from the foil side of the board. This method requires that the etching pattern be viewed from the *component side* of the board. Make a mirror image of the etching pattern by copying the etching pattern onto a transparency on a plain-paper copier or a Thermofax machine. Place a blank sheet of paper behind the transparency to obtain the component-side view.
2. Copy the component view of the pattern onto a sheet of paper in a plain-paper photocopier. The copier should be set for as dark an image as possible, but not so dark that the white areas of the image start showing streaks.
3. Thoroughly clean the bare copper board. Remove all surface corrosion using steel wool. I like to make a final cleaning pass by wiping the board with acetone (use caution!). Some builders advocate placing the bare board in ferric-chloride etchant for a minute, but I've gotten acceptable results without using this step.
4. Set up an ironing board and iron. Set the iron's temperature control to the COTTON or LINEN position. This heats the ironing surface to about 300 °F. Turn off the steam. Let the iron fully warm up to a stable temperature.
5. Place a blank sheet of paper on the ironing board (to protect it from toner). Place the clean copper board on the paper, copper side up. Position the plain-paper copy on top of the board, image side touching the board. Put a blank sheet of paper on top of the sheet with the etching pattern. This keeps toner on the copy from transferring to the iron. (When used for its *intended* purpose, an iron with toner on it does a wonderful job of making clean laundry dirty and is certain to raise the ambient noise level.)
6. Carefully place the iron against the paper. S-l-o-w-l-y slide the iron across the paper so that it's flat against the copper board. This causes the toner to start sticking to the copper.
7. Move the iron across the paper in a circular motion with about one cycle every few seconds. Do this for two minutes. Don't neglect the edges and corners; be especially careful about this. If the edges and corners don't get enough heat, the toner won't adhere to the copper when the paper is removed.
8. Let the circuit board cool for several minutes.
9. Prepare a solution of ½ cup of bleach in a gallon of water. Place the circuit board and adhering paper in this solution and let it sit overnight. The solution will start to dissolve the paper. Using a soft brush, *gently* scrub the circuit board to remove the paper. If some paper fibers stubbornly adhere to the toner, place the board back in the solution, let it sit another couple of hours, and brush again. Any paper adhering to the toner may cause the rough trace edges and possible shorts to adjoining traces.
10. Touch up any spots missing toner with an etch-resistant pen. I've found some "etch-resist" pens don't work well (or not at all), whereas permanent-ink marking pens work very well. The Radio Shack etch-resist pen ink works if the etching period isn't too long (say, in excess of 20 minutes).
11. Etch the board in ferric chloride (available at Radio Shack and other electronics stores). Carefully follow the directions on the etchant container.
12. Clean the etched board with acetone or steel wool to remove the toner. The board is now ready to drill and solder.
13. In some cases, the transfer to the board may be poor enough that it's not worth etching the board. If this happens, remove the toner with acetone or steel wool and restart the process.—*KI6WX*

This process can produce PC boards having trace widths of less than 0.02 inch.

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I'd like to thank Russ Mirov and Hal Massey, WB6NNR, for initially showing me the transparency method of making circuit boards.

Notes

[1] C. Nord, "More on PC Boards," *QST*, Aug 1977, Hints and Kinks, p 77; P. Pagel, "G.C. Electronics Lift-It Transfer Sheets," *QST*, Sep 1981, Product Review, p 49; L. Wolfgang, "PC Board Patterns the Easy Way," *QST*, Apr 1982, Hints and Kinks, p 53; T. Desaulniers, Jr, "A Drill Press for Printed-Circuit Boards," *QST*, Jan 1984, Hints and Kinks, p 46; J. Stinson, "A Bubble Etcher for PC Boards," *QST*, Nov 1984, pp 45-47, 56; R. Grabowski, "Etch-Resist Pens for Homemade Circuit Boards," *QST*, Sep 1988, Hints and Kinks, p 44; P. Atkins, "Isolated-Pad Drills Still Available," *QST*, Jul 1989, Hints and Kinks, p 39.

Tek-200 film, which uses a PC-board-making process similar to that described in this article, is available from several sources including Ocean State Electronics, PO Box 1458, 6 Industrial Dr, Westerly, RI 02891, tel 800-866-6626, 401-596-3080; fax 401-596-3590.—*Ed.*

DynaArt Designs has a toner-transfer system that can be used to produce plastic faceplates and decals as well as PC boards. DynaArt Designs, 3535 Stillmeadow Ln, Lancaster CA 93536, tel 805-943-4746. These materials are also available from All Electronics, PO Box 567, Van Nuys, CA 94108, tel 800-826-5432, 818-904-0524, fax 818-781-2653 and DC Electronics, PO Box 3203, Scottsdale, AZ 85257, tel 800-467-7736, 800-423-0070, 602-645-7736, fax 602-994-1707.—*Ed.*

[2] T. Miller and P. Pagel, "PC-Board Production in the ARRL Lab," *QST*, Apr 1988, pp 15-18.

[3] PC-board etching materials are available from most electronic product mail-order distributors. Your local Radio Shack store carries the materials, too: PC-board kit (276-1576); etch-resistant pen (276-1530); etchant, 16-oz container (276-1535).

Bibliography

W. Richardson, "Practical Printed-Circuit Boards," *Modern Electronics*, Aug 1989, pp 14-17, 82.

D. Lancaster, "PC-board Breakthrough," *Radio Electronics*, Hardware Hacker, Dec 1989, pp 68-76.

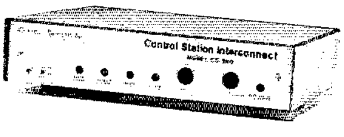
J. Axelson, "Making Printed-Circuit Boards Without Photography," *Modern Electronics*, Dec 1989, pp 16-21.

T. Byers, "Printed-Circuit Boards From Your PC," *Modern Electronics*, Feb 1991, pp 18-73, 74-75.

New Products

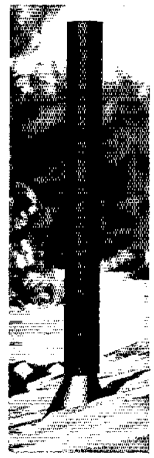
SIMPLEX VOX PATCH

◊ The CS-900 is the first low-cost voice-activated (VOX) simplex phone patch to incorporate standard features such as a ½-second electronic voice delay (EVD), built-in programming keyboard with digital display, 90-memory speed dialer, last-number redial, call-progress tone detection, user-programmable CW ID and more. The CS-900 is VOX-operated and doesn't use sampling. There are no sampling noises or interruptions; just smooth, natural audio. The built-in EVD prevents syllable and word clipping caused by slow-switching transceivers. It interfaces with most transceivers, with the only connections required to the mike and speaker jacks. Suggested list price $379. Connect Systems Inc, 2064 Eastman Ave, #113, Ventura, CA 93003; tel 805-642-7184, fax 805-642-7271.



DUAL-BAND "INVISIBLE" ANTENNA

◊ The neighbors will never know you're a ham with the VT-27 dual-band Ventenna for 2 meters and 70 cm. This model follows the original 2-meter single-band Ventenna design, with an antenna sealed between two layers of ABS plastic in a hollow tube that slips over your rooftop plumbing vent pipe. This durable 36-inch-long antenna is virtually indestructible and undetectable by observers. It comes with an attached 2.5-foot length of Belden high-performance 8240 (RG-58) coaxial cable that connects to your equipment with an SO-239 inline connector. Retail price $74.95. Mark Forbes, KC9C, The Forbes Group, PO Box 445, Rocklin, CA 95677; tel 800-551-5156, voice or fax 916-624-7069.



Strays

RADIO-CONTROL NET

◊ The R/C Flyers Net meets on 3.933 MHz every day at 1100-1200Z (following the Southern Coffee Club Net) to discuss all aspects of R/C model aircraft. Fred Lomax, N4KYG, 204 S Claiborne St, Goldsboro, NC 27530; or David Rose, WD4DJA, 817 Hawthorne Rd, Shelby, NC 28150.

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August 1993 $3.00

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RTTY 60 Receive   UTC 1/2/93 07:17

CQ TEST CQ TEST CQ TEST
DE NF1J NF1J NF1J NF1J KKK
_
 ESC: Menu   Alt-H: Help
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QST

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OUR COVER

As a famous competitor once said, "If you see 'crossed bananas' on you oscilloscope—it must be RTTY Roundup weekend!" At the lower left, John SV1AHV, works RTTY pileups from SV1SV *(photo by SV1BDO)*. At the upper right, Anna (top) and Sandra (seated) get in on the fun from WB2JKJ, the club station of Junior High School 22 in New York City. Check out the results of this year's RTTY Roundup on page 98. *(oscilloscope and screen shots by Kirk Kleinschmidt, NT0Z)*

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August 1993
Volume LXXVII Number 8

TECHNICAL

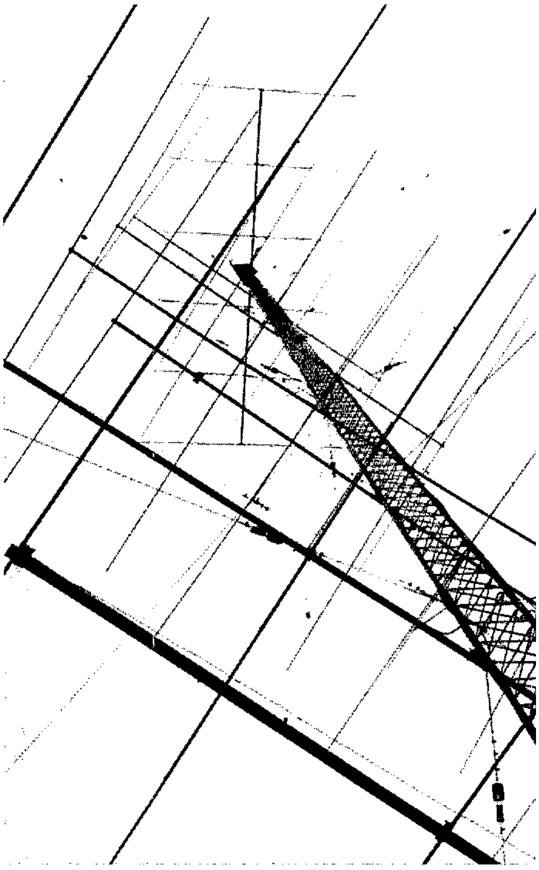
NEWS AND FEATURES

NEW HAM COMPANION

OPERATING

DEPARTMENTS

The Effect of Continuous, Conductive Guy Wires on Antenna Performance



(photo by Rus Healy, NJ2L)

Radio amateurs have long worked to preserve antenna patterns by installing segmented, nonresonant guys. Now, computer modeling reveals that *unbroken* guys can work just about as well.

By Lew Gordon, K4VX
PO Box 105
Hannibal, MO 63401

Ham lore has long preached that tower guy wires not broken up into short, nonresonant lengths by insulators will significantly degrade the performance of antennas on the guyed tower. In contrast, I recall that, a number of years ago, the builder and owner of one of the most successful contest stations in the world, Ed Bissell, W3AU (ex-W3MSK), did *not* use insulator-segmented guys on some of his towers—and his signals were legendary throughout the world on all bands.

Continuous, conductive guys—no segmentation, no insulators—currently support two of my nine towers, and I observe *no discernible performance degradation* attributable to their presence. I hope that the following analysis will dispel some of the myths about conductive guying.

Guying Options

You have three options when constructing a guyed tower. You can:

1. Use continuous, nonconductive guy material, such as Phillystran.
2. Use conductive guys, but break them into nonresonant segments with insulators.
3. Use conductive, *continuous* guys with no insulators or segmentation whatsoever.

For many hams like me, Option 1 is often out of the question because of its cost. I have used Option 2 many times, but with the addition of the 10, 18 and 24-MHz bands at WARC-79, it's difficult to find a guy-segment length that does not resonate in at least one of the ham bands between 3.5 and 29.7 MHz. Also, the cost of Option 2 is now approaching that of Option 1. Option 3 is by far the least-expensive approach—but what about the warnings of ham lore? What does Option 3 cost in terms of antenna *performance*?

I decided to computer-model the antenna-pattern degradation (if any) attributable to continuous, conductive guys. Using *NEC*,[1] I modeled a 109-foot tower constructed of Trylon AB-105, with conductive ⅜-inch-diameter guys connected between the tower (at 53 and 103 feet) and three ground anchors spaced 70 feet from the tower's center. The antennas on the tower comprise two simplistically designed five-element Yagis on 58½-foot booms, one mounted at 54 feet; the other, at 109 feet. (This arrangement is similar to the 20-meter antenna situation at my station.) I modeled the antenna elements as perfect conductors, dividing them into an adequate number of segments for *NEC* analysis.[2] I modeled the ground beneath the tower as exhibiting a dielectric constant of 10 and a conductivity of 13 millisiemens per meter.

For graphing the results, I chose a 0° to 180° linear plot over the more common polar plot to allow detailed side-lobe examination at higher-than-usual pattern attenuations. I plotted the antennas' horizontal and vertical patterns at guy-position intervals of 15° from 0° to 60° off boresight. (The 0° plot places a set of guy wires directly on the Yagis' boresight.) Because the guy geometry repeats after 60°, five modeling passes are sufficient to show the guys' effect on the antennas as they rotate 360°.

Figs 1A and 1B show patterns for the upper Yagi. The horizontal pattern (Fig 1A) shows very little disturbance; the vertical plot (Fig 1B) shows some change in the back-lobe structure. As most of the variation occurs at attenuations greater than 30 dB, this effect can be considered trivial. The forward main lobe at about 9°—the one we communicate with—is not disturbed at all.

Figs 2A and 2B show patterns for the lower Yagi. The back-lobe structure in Fig 2A indicates approximately a 4- to 5-dB variation as the antenna rotates. The higher-

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[1] Notes appear on page 24.

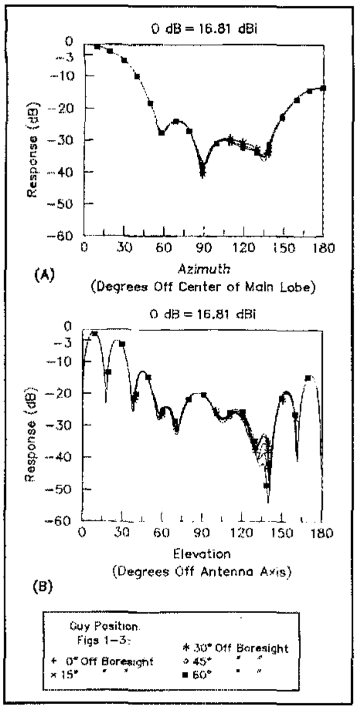


Fig 1—Modeled patterns for the upper Yagi (height, 109 feet). The horizontal pattern (A) shows very little disturbance; the vertical pattern (B) shows trivial change in the back-lobe structure. The main lobe (at about 9°) is not disturbed at all. (*patterns modeled with* NEC *at 14.2 MHz, over ground with a dielectric constant of 10 and a conductivity of 13 millisiemens per meter*)

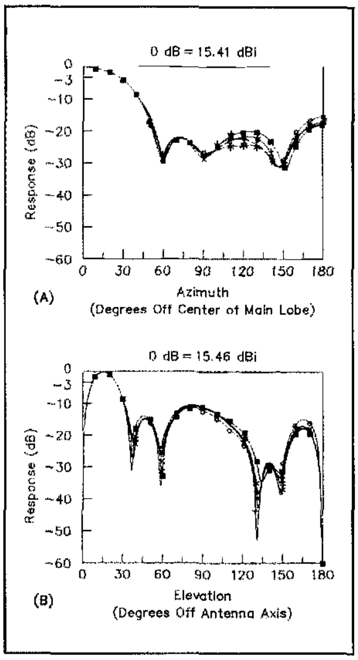


Fig 2—Modeled horizontal (A) and vertical (B) patterns for the lower Yagi (height, 54 feet) show that the practical effect of continuous, conductive guys is slight. Although the higher-angle nulls in B appear to vary over 20 dB, all of these variations occur at attenuations of 25 dB or greater. Most contesters and DXers would find these effects inconsequential.

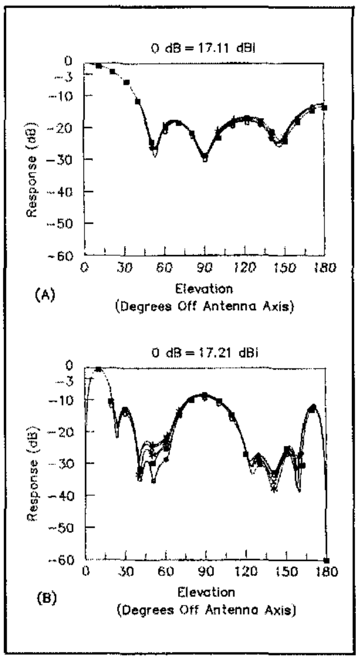


Fig 3—The effect of continuous, conductive guying on the two Yagis driven in phase. As with the previous plots, very little change occurs in the horizontal plane; some variation does occur, however, at higher elevation angles in the vertical plane. The main lobe changes minimally in both planes.

angle nulls in Fig 2B appear to vary over 20 dB, but all of these variations occur at attenuations greater than 25 dB. (In the real world, such deep nulls are unrealizable because of ground variations.) Although this pattern variation may be intolerable to purists, most contesters and DXers would find it inconsequential.

Figs 3A and 3B show patterns for the two Yagis driven in phase. As with the previous plots, very little change occurs in the horizontal plane. Some variation *does* occur at higher elevation angles in the vertical plane, however. As in the previous plots, the guys minimally change the forward main lobe.

Fig 4 plots the stacked Yagis' pattern in the total absence of guys (best case) overlaid with the pattern that results when boresight (0°) guys are present (worst case). Except for the second forward lobe at approximately 30°, all of the variations occur at attenuations of 20 dB or greater.

Although I do not present the evidence graphically, guys directly below the antenna affect the antenna much more than guys directly in front of the antenna. This is so because a Yagi's feed-point impedance changes rapidly as its boom approaches the guy attachment point on the tower. The higher the antenna's feed-point impedance, the wider its SWR bandwidth and the higher its efficiency. *NEC* runs using several different boom-to-guy separations indicate that a separation of at least 1 foot is mandatory at 14 MHz; 3 to 5 feet is preferable. I have observed that guy-wire proximity affects Yagis with low feed-point impedances, such as 10 to 15 Ω, much more than Yagis with impedances of 20 to 35 Ω.

Lower-Frequency Antennas

How might continuous, conductive guys affect the pattern of a half-wave sloper (a very popular low-band DX antenna)? I modeled a *sloping, half-wave*, 75-meter dipole extending, at a 60° angle, from near the top of the 109-foot tower to near the ground. The dipole bisects two of the guy-wire sets. Fig 5A shows the dipole's horizontal-plane patterns (derived at an elevation angle of 20°) with guys (+) and without guys (×). Interestingly, in this specific case, the guy wires actually *increase* the sloper's gain in the desired direction by over 2 dB! In addition, they also enhance its front-to-back ratio. Fig 5B shows the sloper's vertical-plane patterns, which reflect the same gain differences as Fig 5A.

Conclusions

Continuous, conductive guy wires *do*

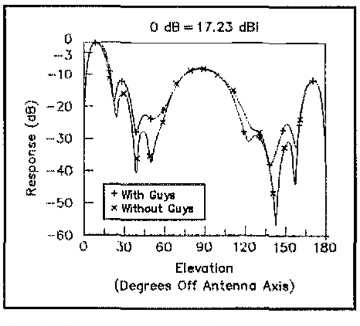


Fig 4—The stacked Yagis' pattern in the total absence of guys (best case) overlaid with the pattern that results with boresight (0°) guys *present (worst case).* Except for the secondary forward lobe at approximately 30°, all of the variations occur at attenuation levels greater than 20 dB.

disturb the patterns of antennas mounted on towers. But unless you're a purist in pursuit of unrealistically clean side lobes, these effects are of no practical importance.

Antennas mounted at the top of a tower (Figs 1A and 1B) show very little effect from guy wires attached several feet below the boom.[3]

Continuous, conductive guy wires may actually enhance the pattern of a low-band wire antenna positioned between them.[4]

As the costs of good-quality insulators, guy-wire preforms (dead ends) and application-specific nonconductive guy cables increase, continuous, conductive guys, installed without insulators, may be a viable guying solution. I hope that this analysis may convince future tower constructors to consider this alternative.

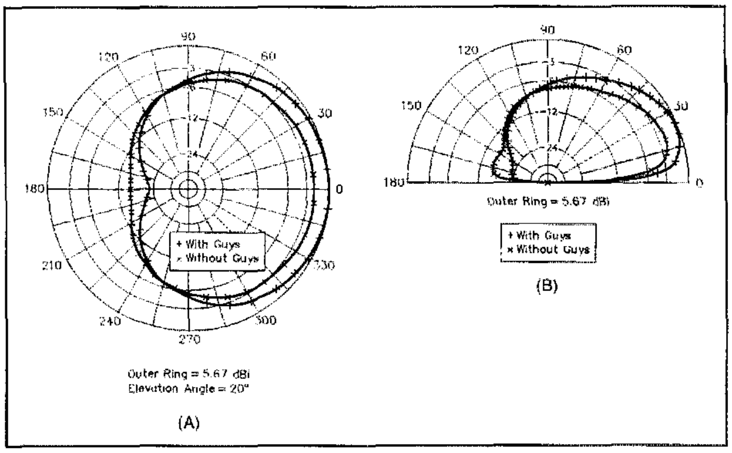


Fig 5—Continuous, conductive guys slightly *enhance* the forward-lobe performance of a 75-meter sloper modeled as extending, at a 60° slope, from near the top of the 109-foot tower to near the ground. The antenna's horizontal patterns appear at A; vertical patterns, at B. (*patterns modeled with* NEC *at 3.85 MHz, over ground with a dielectric constant of 10 and a conductivity of 13 millisiemens per meter*)

Notes

[1] *NEC* is an acronym for *N*umerical *E*lectromagnetics *C*ode, developed by Lawrence Livermore National Laboratory.

[2] All wires use segment lengths of 0.1 wavelength or less. For those readers with access to the *NEC* program, the input and output files are available from me upon receipt of a self-addressed, stamped disk mailer containing a PC/MS-DOS-formatted 1.44-Mbyte, 3.5-inch floppy disk.

[3] Use as much boom-to-guy-wire separation as is practical to allow side-mount rotation of lower antennas. I recommend a minimum separation of 1 ft at 14 MHz; a separation of several feet is more desirable.

[4] This applies to this specific case. It's possible that other tower/guy configurations may produce different results.

New Books

THE HAMSAT HANDBOOK

By Anthony R. "Tony" Curtis, K3RXK

Published by Tiare Publications, PO Box 493, Lake Geneva, WI 53147. First edition, 1993. Softcover, 8½ × 11 inches, 152 pp. $19.95 plus $2 s/h.

Reviewed by Steve Ford, WB8IMY
Assistant Technical Editor

The title of the book is slightly deceiving. *The Hamsat Handbook* is *not* a how-to-do-it guide for operating Amateur Radio satellites. Even so, the book is recommended reading for every satellite-active ham or any ham who is *thinking* about satellite operating.

The Hamsat Handbook is a collection of fascinating information about the satellites themselves. From cover to cover it explores the complete history of Amateur Radio operations in space—including the *Mir* space station and the Shuttle Amateur Radio Experiment (SAREX) missions. (The author even discusses high-altitude balloons, moonbounce and meteor-scatter communications.) You'll learn the frequencies and modes used by satellites, and arcane details, such as power generation, telemetry and so on. In many instances, the author includes brief descriptions of the types of equipment required to work the birds.

Throughout the book there are interesting nuggets of information contained in sidebars. For example, the "Satellite Scoreboard" lists all nations that have objects in space today. (Did you know that Luxembourg has two satellites in orbit?) There's also a list of Amateur Radio operators who have flown aboard space shuttles.

The author doesn't stop with past and present satellites. He also presents the lineup of *future* satellites in as much detail as possible. (Until I read the book, I was unaware of NANOsat—a tiny British satellite that's scheduled to be tossed out of the *Mir* space station in 1994.)

The writing style is informal and conversational. *The Hamsat Handbook* is the type of book that's fun to browse. An extensive glossary is available—a helpful feature for the new satellite user—and the index is one of the best I've seen in books of this type.

The only flaw I could find in *The Hamsat Handbook* was the lack of illustrations. Drawings or photographs of the satellites would have added more value to an already worthwhile book. Keeping up with the ever-changing world of Amateur Radio satellites is a daunting task. The author has fixed his sights on a moving target and it may not be long before another edition of *The Hamsat Handbook* will be necessary to keep pace with current events!

MONITORING NASA COMMUNICATIONS

By Anthony R. "Tony" Curtis, K3RXK

Published by Tiare Publications, PO Box 493, Lake Geneva, WI 53147. First Printing, Jan 1992. Softcover, 8½ × 11 inches. 100 pp. $14.95 plus $2 s/h.

Reviewed By Brian Battles, WS1O
QST *Features Editor*

Here's a handy, practical guide to listening in on space communications that covers just about everything the typical enthusiast will want to know. The author has been writing about science and space for 30 years, and has written almost 70 books, including the 960-page *Space Almanac*. He's been a ham since the 1950s. Curtis knows his material and covers it in great detail.

The book begins with an engrossing history of NASA and the birth of man-made satellites. It explains the work satellites do and why it can be interesting to monitor them. This discussion covers active and passive weather, navigational, deep-space and other types of satellites, and even radio astronomy. The well-researched material cuts through much needless technical detail and provides an unusually lucid explanation of the purposes, results and potential of past, present and future manned and unmanned space missions.

The best and most detailed part of the book is its extensive discussion of manned NASA operations, namely, US space shuttles. Several handy boxes and tables list key facts, shuttle missions and objectives, specific frequencies, and operational details. Curtis unscrambles the meaning of NASA's mysterious acronyms and mission-designation codes. He explains the kind of equipment needed to monitor space communications, frequencies and modes used for particular purposes, and locations of ground stations.

Although the book doesn't focus on Amateur Radio, it does list the frequencies and general schedules of amateur stations that typically rebroadcast NASA communications. These include the outstanding HF transmissions made available by amateurs at WA3NAN at the Goddard Space Flight Center in Maryland, W5RRR at the Johnson Space Center in Houston and W6VIO at the Jet Propulsion Laboratory in Pasadena. An exhaustive table also lists dozens of sources of NASA and amateur VHF/UHF voice and television signals that may be directly received. To round out the book, an excellent glossary and index make it easy to understand and locate details.

If the broadcast news media's coverage isn't sufficient to fulfill your interest in space operations, or if you simply enjoy tuning directly into space operations from the comfort of your home, *Monitoring NASA Communications* belongs on your bookshelf.

A Dual-Range AC Voltage, Current and Frequency Monitor

This easy-to-build project is great for checking ac power sources, including generators. For Field Day, it's ideal!

By Ed Oscarson, WA1TWX
70 Behrens Rd
New Hartford, CT 06057

This microprocessor-controlled monitor can be used in many ways. As a bench instrument, it can monitor the output voltage and load current of a variable-voltage transformer or other power sources. Around the home, you can check household appliance current consumption. The monitor proves its usefulness when setting an electric generator's governor to provide ac output exactly at 60 Hz. It's universal, too, in that it operates over the 30- to 255-Hz range and you can configure it to operate with 120-V or 240-V (US and foreign) ac supplies.

By means of an LCD readout, the monitor displays line voltage, line frequency and the current delivered to the load. Because this circuit is designed primarily for use with a generator, the delivered *current*—not power—is one of the displayed parameters.[1]

General Description

Software in the microcontroller[2] converts the analog inputs to the following ranges: voltage, 0 to 255 ac;[3] current, 0.0 to 25.5 A (with 0.1-A resolution); frequency, 30 to 255 Hz (resolution varies with frequency). Other ranges (such as 400 Hz) can be accommodated with minor software changes.

In addition to voltage, current and frequency displays, the monitor has five alarm outputs to alert you when things go wrong: two outputs each for overvoltage and overcurrent, and one for over or under frequency. These outputs are uncommitted and are capable of driving TTL-level power FETs, or transistors. You can use the outputs to control an alarm or disconnect the output from the load. If needed, multiple outputs can be diode-ORed to control the alarm.

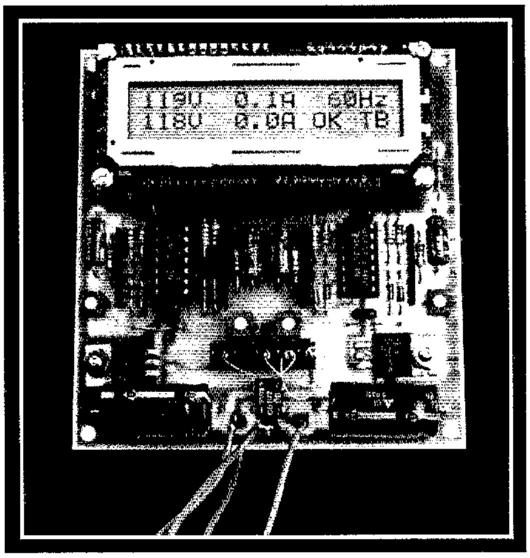
The frequency alarm limits are set to 55 and 65 Hz. A jumper allows selection of an alternative 45- to 55-Hz range. The voltage limit is set between 95 and 130. Voltage channel 1 also works with limits between 200 and 250 V for 220- and 240-V power systems. The current limit is selectable via jumpers. An additional output is active when the current is less than 2 A. This output can also drive a TTL-level power FET. The FET can be connected to a solenoid to control the generator's throttle to cut back the generator speed when the current is below 2 A.

Circuit Description

As shown in Fig 1, the monitor uses an MC68705R3 microcontroller (U1) with built-in analog-to-digital converter (ADC) and timer connected to a liquid-crystal display (LCD). U1 has 4 kbytes of EPROM, 128 bytes of RAM, an internal oscillator, timer and a four-channel ADC. Also available are 24 bidirectional I/O lines split into three 8-bit ports.

Display

Character generation and refresh circuitry are built into the LCD, and it selectively provides for displaying one or two lines of 16 characters each. The display accepts standard ASCII data, including commands for cursor-positioning, display-clearing control, etc. The display's 8-bit data bus also accepts data with two 4-bit writes. This allows one 8-bit I/O port to provide all the control lines and data for the display.



With the two-line liquid-crystal display mounted directly to the PC board, the main portion of the monitor makes a compact package. Several components, including the microprocessor, are beneath the display.

U1's Port C accesses the display. The microcontroller software divides the display into three sections: *xxx*V, *yy.y*A and *zzz*Hz, where the *x*, *y*, and *z* represent the digits for the voltage, current and frequency, respectively. For use with a 240-V generator (two 120-V lines), the second (selectable) display line is used.

Port D of U1 is configured as an 8-bit-resolution ADC. Because the input ranges are 0 to 255 V, or 0 to 25.5 A, the 8-bit ADC data can be converted from binary to binary-coded decimal (BCD) and output directly to the display, with a decimal point inserted at the correct spot for the current value. An active full-wave rectifier converts the ac signals to dc. This rectifier eliminates error introduced by simple diode rectifiers at low voltages. Level shifting is done before the rectifier. This allows for simple calibration to compensate for the various types of transformers you can use with the monitor.

Power Supply

LM7805 (U2) and LM7905 (U3) regulators deliver positive and negative voltages to U1 and the op amps (U4 and U5). These regulators also limit the op amp's output voltage swing to below +5 V (after rectification), which prevents damage to U1's analog inputs. The regulators have a wide input-voltage range, allowing the use of many readily available power transformers.

T1 should have an 18-V center-tapped secondary and a current rating of 300 mA or more. This provides a dc supply of about ±13 V at 120 V ac input, giving the monitor a useful range of about 70 to 140 V

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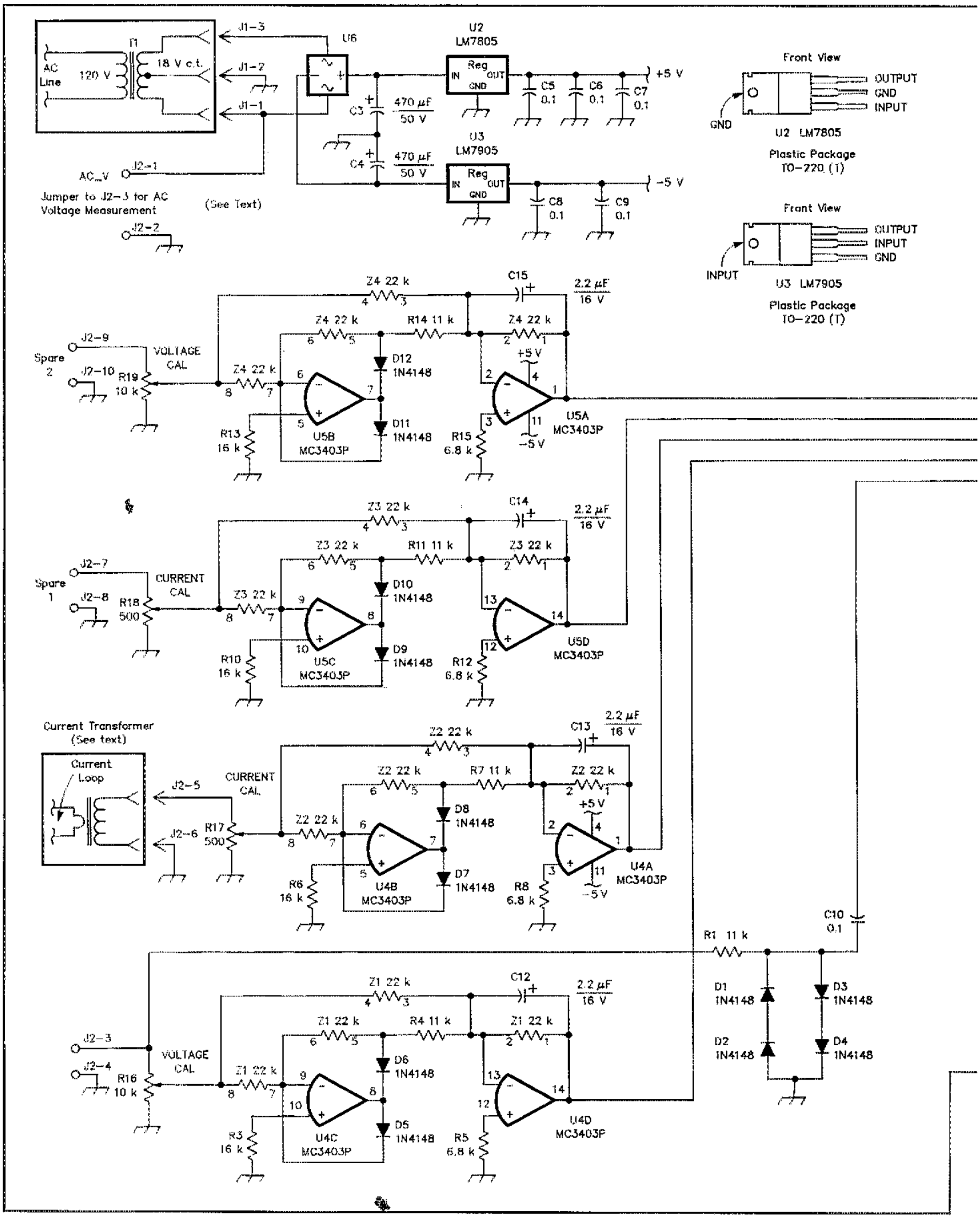
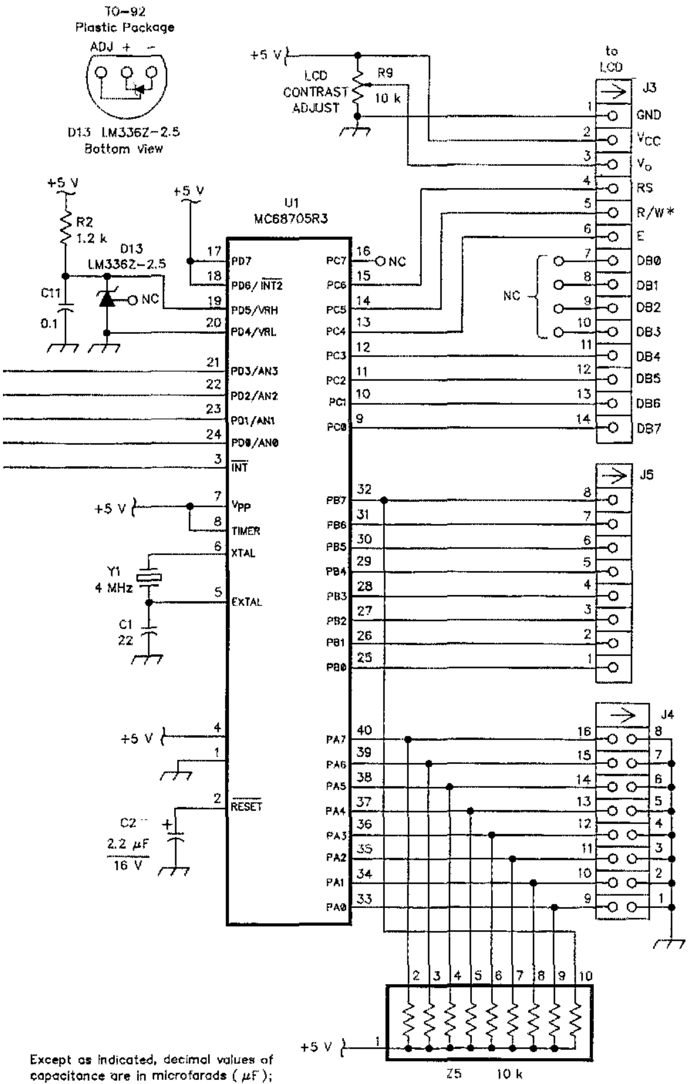
[1] Notes appear on page 30.



ac. You can also use a 25-V, 1-A, center-tapped transformer.[4]

U1 has an internal 8-bit ADC, but requires an external voltage reference; an LM336Z-2.5 (D13). Using a 2.5-V reference results in an ADC input range of 0 to 2.5 V dc. This allows the monitor to use the ADC's full 8-bit range. If +5 V were used as the reference, some op amps wouldn't be able to swing to the positive supply rail, causing loss of the upper part of the ADC's range.

During operation, U1's software samples each ADC input once each input waveform cycle. After conversion, the incoming values are stored in memory and sent to the display



Except as indicated, decimal values of capacitance are in microfarads (µF); others are in picofarads (pF); resistances are in ohms; k=1,000.

NC = No Connection

Fig 1—The monitor schematic. Unless otherwise specified, resistors are ¼-W, 5%-tolerance carbon-composition or film units. A detailed parts list is available with the PC-board template (see Note 1). Resistors identified by Z numbers and numerically identified leads are part of resistor SIPs for which the PC board is designed. Parts are available from Digi-Key Corp, 701 Brooks Ave S, PO Box 677, Thief River Falls, MN 56701-0677, tel 800-344-4539, 218-681-6674; fax 218-681-3880.

DS1—1- or 2-line × 16-character LCD.
R9—10-kΩ PC-mount trimmer potentiometer.
R16, R19—10-kΩ PC-mount trimmer potentiometer.
R17, R18—500-Ω PC-mount trimmer potentiometer.
T1—18-V, 1.5-A center-tapped secondary.
T2—6.3-V, 0.3-A center-tapped secondary. (see text).
T3—6.3-V, 3-A center-tapped secondary (see text).
U1—Programmed MC68705R3 (see Note 1).
Z1-Z4—22-kΩ, 2%-tolerance SIP resistor network.
Z5—10-kΩ, 2%-tolerance SIP resistor network.

once per cycle as a background task.

I/O Access

Monitor inputs are available at a 10-pin header (J2). Each input has a dedicated ground. The secondary of the transformer that supplies power to the circuit is also connected to the header so that it can optionally be connected as one of the voltage inputs. See Table 1; it lists J2's inputs and outputs.

Voltage measurement is made by applying an ac voltage to a resistive divider to produce a level that is easily handled by the op amps. Level-adjustment potentiometer R16 [**VOLTAGE CAL**] allows trimming the voltage during calibration to compensate for variations in the transformers. The level-shifted ac voltage is then applied to an active full-wave rectifier (absolute value) circuit (U4C). This circuit converts the ac to unipolar pulses that are then filtered by the second op-amp section. The filtered dc is applied to the appropriate analog multiplexed input (AN0) of U1.

Current Measurement

For current measurement, a second ac input is connected to a current transformer. The input divider, active rectifier and filter circuits are similar to those used for the voltage-input section. A 500-Ω potentiometer (R17, **CURRENT CAL**) on this input adjusts the value of the current displayed during calibration.

Table 1
J2 Pin Assignments

J2 Pin #	*Signal*
1	Monitor power-transformer ac-output voltage.
3	Voltage-input #1. Also supplies the frequency input.
5	Current input #1.
7	Current input #2.
9	Voltage input #2.
2, 4, 6,	Ground.
8, 10	Ground.

Some "How-To's" About Current Transformers

The voltage output of a current transformer can be determined from the transformer's turns ratio and the load applied. First, we need to determine the turns ratio, because most transformers don't have that information marked on them. One way to get a turns-ratio estimate is to wrap about 10 turns of #22 or #24 wire around the transformer core as an additional secondary winding. Insulate all connections before applying power, then supply line voltage to the transformer primary and measure the secondary voltage. The number of turns in the primary (T_{pri}) is:

$$T_{pri} = (V_{line} \div V_{sec}) \times T_{sec} \qquad \text{(Eq 1)}$$

where T_{sec} is the number of turns you wrapped around the core. The more turns you wrap around the core, the more accurate your reading will be.

When current is flowing in the half-turn of wire used as our sense winding, the primary winding will try to generate a current that is ($0.5 \times I_{SENSE} \div T_{pri}$). For example, if there are 1000 turns in the primary winding, the current will be $0.5 \times I_{SENSE} \div 1000$. If I_{SENSE} is 10 A, the primary current will try to be 0.5×10 A $\div$ 1000, or 5 mA. Because we want to get about 2.5 V at 25 A, 10 A should yield 1 V at the monitor input. This voltage is determined by the load resistance connected to the former primary winding. Because we want 1 V with 5 mA, the resistance is $R = 1 \div 0.005 = 200\ \Omega$. If the primary had 2000 turns, the resistance required would be 400 Ω. This load resistance is supplied by the CURRENT CAL potentiometers. Their 500-Ω value ensures that the board works with most commonly available transformers.—*WA1TWX*

This low-value potentiometer also provides the transformer load. For current measurement, the second analog mux input (AN1) of U1 is used.

The remaining two ADC mux inputs, AN2 and AN3, are available for use with a 240-V generator and are set up as voltage (R19, U5A and U5B) and current (R18, U5C and U5D) inputs. A two-line display is required to use these inputs. (A two-line display is the default; a one-line display is selected by a jumper connecting Port A, bit 7, to ground.)

Frequency Measurement

The frequency-measuring circuit makes use of U1's timer. The ac signal to be measured is derived from the main voltage input on J2, pin 3. This signal is limited by a resistor (R1), diode-clamped to about 1.4 V (by D1-D4), and capacitively coupled by C10 to the external IRQ interrupt input of U1. U1 detects the zero crossings and interrupts the processor. The interrupts are converted to frequency in the software.

The voltage current, and frequency-error-checking routine uses the stored data from the conversions and compares the data against predetermined limits. This happens at a 60-Hz rate using the IRQ interrupt as a time base. If a parameter exceeds its limits for about one second, the appropriate alarm bit is set. The throttle-back control uses the same technique, setting the throttle-back bit when the current is less than 2 A. These bits are TTL-compatible logic levels capable of driving a transistor or logic-level power MOSFET. Throttle control is enabled by a jumper connecting Port A, bit 5, to ground. Frequency check is then disabled because the frequency changes when the generator speed is reduced. Tables 2 and 3 define the port bits and their functions.

Construction

Although point-to-point wiring can be used, I recommend using a PC board. An etched and drilled PC board is available (see Note 1). Use a 30-W grounded-tip soldering iron with a fine tip for best results.

After completing PC-board construction, connect the power transformer to the PC board. J1 pins 1, 2 and 3 are the ac-power inputs. Pin 2 is for the center tap of the secondary winding. Do *not* install U1, U4 or U5 until initial checks are made.

For portability, I built my unit in a black box (one of those things you always hear about in electronic circles...) equipped with two duplex ac outlets and male and female power cords. (You can make the cords simply by cutting in half a heavy-duty extension cord.) This allows easy insertion of the monitor between the generator's power plug and load. If the monitor is used solely with a generator, you may want to permanently mount the monitor on the generator's housing. Remember, though, the monitor will be exposed to *lots* of vibration! For such an installation, I recommend epoxying the large components (such as the filter capacitors) to the PC board and mounting the unit to the housing using shock-absorbent material. *For safety, fuses and properly grounded plugs and receptacles must be used.*

The Current Transformer

When I started this project, I hoped to find a clamp-on, or two-part, core so that the monitor could measure line current without disturbing the generator's wiring. When I couldn't find what I wanted, I tried modifying a Radio Shack power transformer for use as the current transformer. I found that a suitable current transformer can *easily* be made from most 6- or 12-V power transformers: You simply route a length of #10 or #12 wire between the transformer core and its existing windings. An open-frame transformer is easy to work with. You'll probably have to remove the metal case from other transformer types to do this.

The current transformer's primary winding is the *output* and is connected to the monitor's current input (J2-5/J2-6). For the sense wire, one side of the ac line is routed through the transformer core. This provides

Table 2
J5 Pin Assignments

J5 Pin #	*Port B Bit*	*Function*
1	0	Frequency test, 1 = frequency outside limits (Freq = 55 to 65 Hz; optionally, 45 to 55 Hz)
2	1	Voltage test, 1 = voltage #1 outside limits (V1 = 95 to 130 V, or 200 to 250 V)
3	2	Voltage test, 1 = voltage #2 outside limits (V1 = 95 to 130 V).
4	3	Current test, 1 = current #1 outside limits (set by port A bits 0 to 4)
5	4	Current test, 1 = current #2 outside limits (set by port A bits 0 to 4)
6, 7	5, 6	Not used.
8	7	Output to reduce speed, 0 = normal, 1 = reduce speed

Table 3
J4 Pin Assignments

J4 Pin #	*Port A Bit*	*Function*
1-8	—	Ground
9-13	0-4	Sets the current trip value for the current test. The value is equal to the hex value times 0.8 plus 0.7. Example: 01011 = (11 × 0.8) + 0.7 = 9.5 A
14	5	Enable the throttle control, 0 = on, 1 = off (default)
15	6	Frequency select for test, 0 = 50 Hz, 1 = 60 Hz (default)
16	7	Display lines, 0 = one-line display, 1 = two-line display (default)

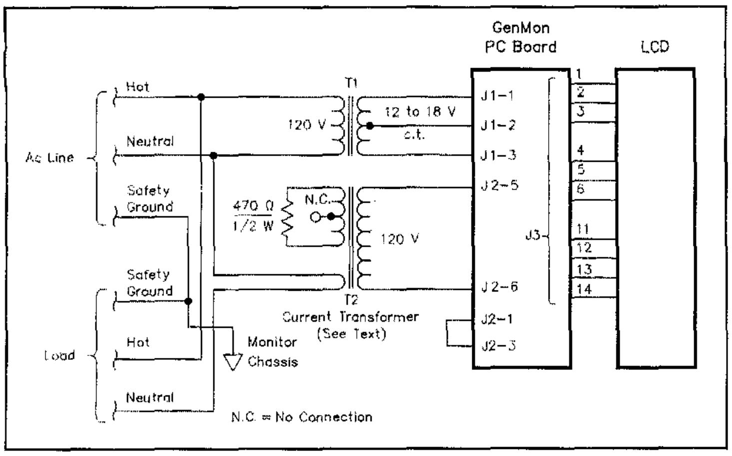


Fig 2—T1 and T2 are as in Fig 1. T3 is used here solely for monitor current-calibration purposes. The text explains the calibration procedure under "*Current Input Calibration Procedure 1.*"

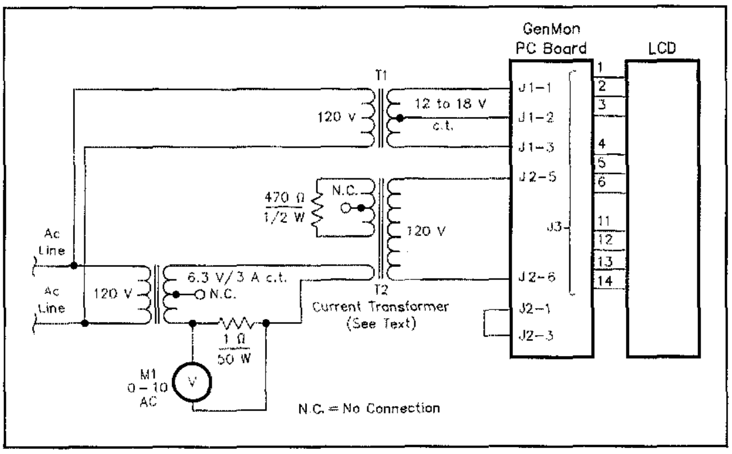


Fig 3—Schematic of the monitor set up for use with a 120-V ac input.

enough voltage (up to 10) on the primary side of most transformers for U1 to measure up to 25.5 A.

Note: Line voltage will be present on the sense winding added to the transformer. Take care that the connections are well insulated and don't nick the wire when you route it through the transformer. Secure the wire to the transformer using epoxy or hot-melt glue. If you're using a 120-V system, route the neutral (usually white) wire through the transformer for added safety.

Testing the Monitor

Ensure the ICs are not installed. If any measurement or operation is incorrect, disconnect the power and troubleshoot the circuit. Turn off power at the source before making any ac connections!

Initial Tests

Attach a plug to the power-supply transformer primary and apply power to the monitor. The potential measured between the positive terminal of C3 and ground depends on the transformer used, but should be between 8 and 18 V dc. The potential between the negative terminal of C4 and ground should be between −8 and −18 V dc. This check verifies correct operation of the main input supply consisting of U6, C3, C4 and associated wiring.

Measure the voltage at pin 4 of U5, with respect to ground. It should be +5 V ±0.3 V. The voltage at pin 11 of U5 should be −5 V ±0.3 V. If these voltages are okay, unplug the power cord.

The Display

You can connect the LCD directly to the PC-board mounting holes provided, or locate the display up to six inches from the board. Most 1-line by 16-character displays have a standard 14-pin I/O connector pin-out, although the connector may not be located in the same place on all displays. Only ten interconnections are required. I recommend using flexible #22 stranded wire if the display is not mounted directly to the PC board. If you have any doubts about the display connections, compare the display's signal names (from its data sheet) to the I/O identifiers in Fig 1. Wire each display pin to the board eyelet with the same name. If you don't want to permanently wire the display, mount a standard 0.1-spaced header on the PC board and a mating socket on the display cable.

Once the display is connected, install U1, U4 and U5. Connect a jumper from pin 1 to pin 3 of J2. This connects the power-transformer secondary to the frequency- and voltage-measurement inputs. Connect the appropriate pins on the port A connector to ground for the operating mode desired (see Table 3).

Apply power to the board and observe the display. **GENMON VER 1.0** is displayed for about one second. Then, the display should clear, and the voltage, current and frequency counts should appear. You may have to adjust R9 (**LCD CONTRAST ADJ**) to make the characters visible.

If no characters are visible at any setting of R9, check U1's reset input (pin 2) using a scope or high-impedance voltmeter. The voltage on pin 2 should be greater than 4. If an oscilloscope is available, check for 4-MHz oscillation at pins 5 and 6. If those inputs seem okay, check the display wiring. With or without displayed characters, adjusting R9 should change the LCD contrast from clear to dark.

If the **GENMON VER 1.0** message appears but the display does not clear after one second, it's an indication that U1 is not getting the IRQ interrupt at pin 3. Troubleshoot the interrupt clamp and limiting-resistor circuit.

Calibration

Voltage-Input Calibration

First, calibrate the voltage input connected to the input power transformer at J2 pin 3 (via the jumper from J2 pin 1). Use R16 (**VOLTAGE CAL**) to adjust the displayed voltage. While measuring the 120-V ac-line voltage, adjust R16 until the LCD display matches the meter reading.

You shouldn't have to calibrate the frequency display; it will read 60 Hz in the US, and 50 Hz in countries using that line frequency. If the readout is incorrect, either the crystal is not oscillating at 4 MHz, or the IRQ circuitry has a problem.

Current-Input Calibration Procedure 1

Refer to Fig 2. This procedure requires external circuitry for the calibration, but is quite accurate. An ac voltmeter is required for the measurements. A low-voltage transformer is used to drive a resistive load. Most 3-A (or greater) transformers can provide much higher currents for a short time, so a 6.3-V, 3-A transformer will suffice. A sense wire is routed through the current transformer, allowing this calibration procedure to be performed with the unit installed in its

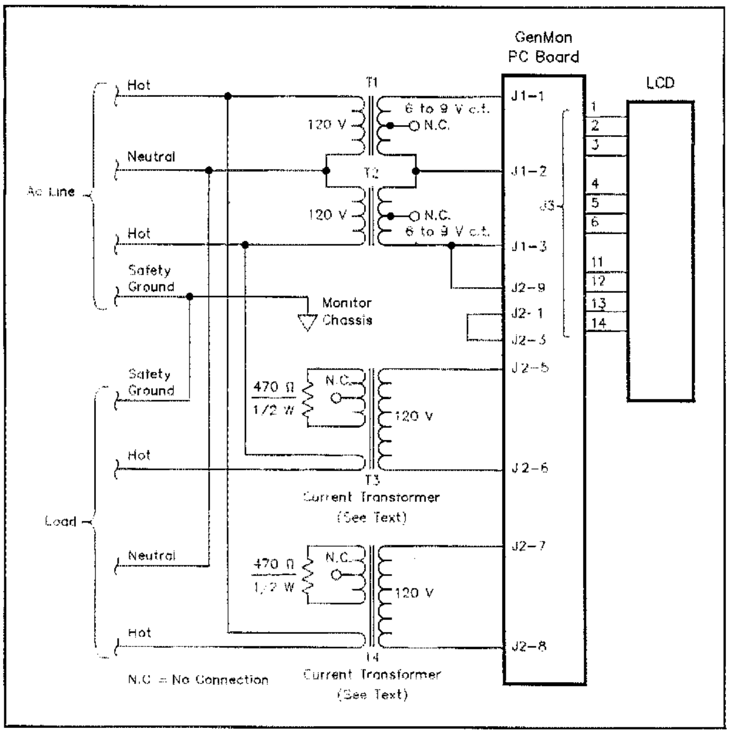


Fig 4—Diagram showing how to use the monitor with a 240-V ac input.

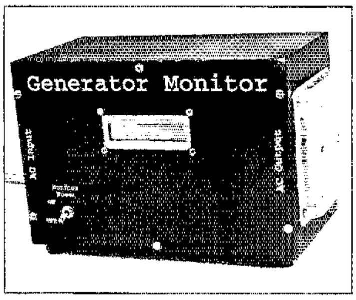


Fig 5—The "black box" monitor. In this unit, the display is located remotely from the PC board. Though labeled **Generator Monitor,** the unit can be used to measure the voltage, current and frequency of other ac power sources. *(photos by Kirk Kleinschmidt, NT0Z)*

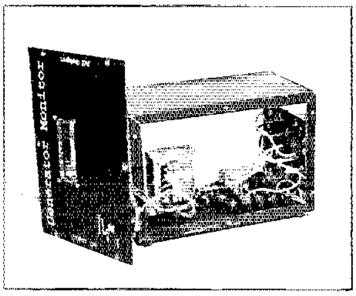


Fig 6—An inside view of the monitor.

final configuration. For best results, don't use a transformer with an output over 12.6 V.

Make the calibration transformer's primary power on/off switchable, or you'll have to plug and unplug the line cord. The calibration transformer secondary is connected to a series circuit consisting of the 1-Ω resistor and the current-sense winding. The resistor should have a 50-W minimum rating and be mounted on a heat sink. The voltage across the resistor is measured and divided by the resistance to calculate the current. When the ac power is connected to the 6.3-V calibration transformer, measure and record the voltage.

With the transformer power *OFF*, calculate the current through the resistor ($I = E \div R$). If a 1-Ω resistor is used, the current is equal to the voltage across the resistor. Reapply power to the calibration transformer and adjust R17 so that the current displayed on the display matches the calculated current through the resistor. When done, remove the extra current-sense wire.

Current-Input Calibration Procedure 2

The easiest calibration method requires a clamp-on ac ammeter. Unlike the first method, no special circuitry is required. Clamp the ac ammeter around one of the current-carrying wires, preferably the sense wire inside the monitor. Connect a load and read the load's current from the clamp-on ammeter's display. Adjust R17 to make the displayed current match the ammeter reading.

At this point, the initial calibration of one channel is complete. Secure the two potentiometer wipers with a dab of paint or nail polish. To calibrate the second channel's inputs (for 240-V systems) simply connect the voltage and current transformers to the appropriate J2 pins and duplicate the calibration procedures described.

Installation

Fig 3 shows the 120-V ac configuration. Two transformers are required: one to supply power for the monitor and voltage sensing, the other for current sensing. The power supply transformer must be center-tapped. The current-sense transformer is located on the neutral wire. This provides an extra measure of safety. Set J4 for a single-line display (see Table 3). If you intend to use the monitor with a two-wire 240-V ac system, use this configuration, but ensure that the power transformer has a 240-V ac primary.

The split 240-V configuration is shown in Fig 4. In this mode, J4 is wired for a two-line display. Note that *two* transformers are used to power the monitor. This allows the transformers to provide power to the monitor circuit and ac to the two voltage-sense inputs. Also, the current-sense lead is now on the *hot* side of each line, so be particularly careful when routing the wires and insulating connections.

Summary

You now have a voltage, current and frequency monitor that will serve you well at home and in the field. I'm sure you'll find this piece of equipment a welcome and helpful addition to your array of test equipment.

Notes

[1] The presence of reactive current in this value means that you can't merely multiply it by the line voltage to determine the real power drawn by the generator load. You must take the load's *power factor* into account to determine real power.

[2] The microcontroller used in this project *must be programmed before use.* Source code is not available. Parts and complete kits are available from: Single Chip Solutions, PO Box 680, New Hartford, CT 06057-0680. Programmed MC68705R3 $45; generator monitor PC Board $25; complete kit (PC board, parts and LCD), $100 (does not include transformers). Assembled and tested PC board with display, $130 (does not include transformers).

A PC-board template package and detailed parts list is available free from the ARRL. Address your request for the OSCARSON GENERATOR MONITOR TEMPLATE to: Technical Department Secretary, ARRL, 225 Main St, Newington, CT 06111. Please enclose a business-size SASE.

[3] The power transformer for this project should be sized for the voltage range in use.

[4] If a 25-V center-tapped transformer is used, mount the +5-V regulator on a heat sink because the regulator's power dissipation can be as high as 3 W at high line-voltage inputs. If the line voltage goes much above 140 V, use a transformer with a 240-V primary (connected to the 120-V line) and a 25-V center-tapped secondary. If only a limited measurement range around 120 V ac is required, a 12.6-V, 1-A center-tapped transformer can be used. This considerably reduces heat dissipation in the +5-V regulator, but limits the working range to between about 90 V and 140 V. QST

A High-Directivity Receiving Antenna for 3.8 MHz

How about a front-to-back ratio of 25 to 45 dB? Three tuned, phased, ground-mounted loops can provide it in less than 100 feet of linear space.

By Floyd Koontz, WA2WVL
6842 Wheeler Rd
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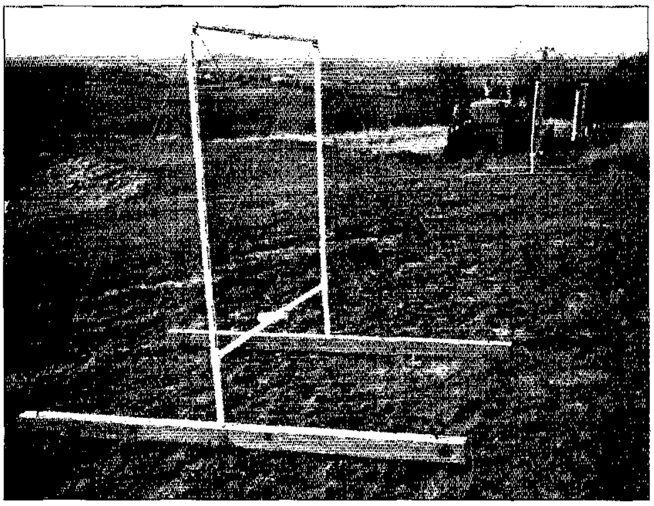

(photos by WA2WVL)

The old saying that you can't work them if you can't hear them certainly applies to 75 meters. Seventy-five-meter DXing from western New York is challenging for this reason. It has been estimated that European DX stations are typically 20 dB weaker in this area than they are near the Atlantic Ocean; farther inland, they are even weaker. Since signal and static levels are generally high at 3.8 MHz, one solution to copying weak European signals is to improve their signal-to-noise ratio with receiving-antenna directivity.

A 400-foot Beverage receiving antenna, directed to Europe, allowed me to work many stations that were not copiable on a ¼-wavelength vertical. Improving my transmitting antenna to a pair of phased verticals reduced the Beverage's advantage to slight. I decided to use Brian Beezley's *MN* antenna analysis program[1] to determine if a better ground-mounted receive antenna could be built.

Nearly all directive antennas have unwanted responses—often less than 20 dB down—on the sides or back of their patterns. Noise and interference contributed by these responses limit the receiving signal-to-noise that can be obtained. In this way, a Beverage antenna's side lobes work against the achievement of high directivity (see Fig 1).

After I reviewed several antenna textbooks,[2-4] the most likely candidate seemed to be the *end-fire binomial array*—an arrangement in which three or more elements are fed with an ever-increasing phase delay (from back to front), with the element currents diminishing from the center element outward. A three-element antenna of this type has appeared in several amateur publications,[5] and usually has ¼-wave spacing with a progressive 90° delay per element and a 1:2:1 current distribution. Computer analysis shows that this antenna can give excellent directivity with an optimized spacing somewhat less than ¼ wavelength and progressive phase delays of 135°.

But what type of element should be used? Mathematicians analyze binomial arrays using what are called *isotropic* elements—idealized radiation sources so small that their dimensions are insignificant relative to wavelength. At 3.8 MHz, a wavelength is about 259 feet, so, seeking practical performance at par with that predicted by theory, and not wanting to put down a large ground system, I chose to use tuned loops instead of vertical elements. The receiving loop has many advantages over monopoles, as discussed by Hart.[6] In addition to being a low-noise antenna, the loop is easily tuned and matched to 50 Ω and requires no ground system. I chose to use square loop elements, each 5 feet per side and mounted 2 feet above ground. Small loops have a null broadside to their plane, so these loops are oriented edge to edge, toward the target.

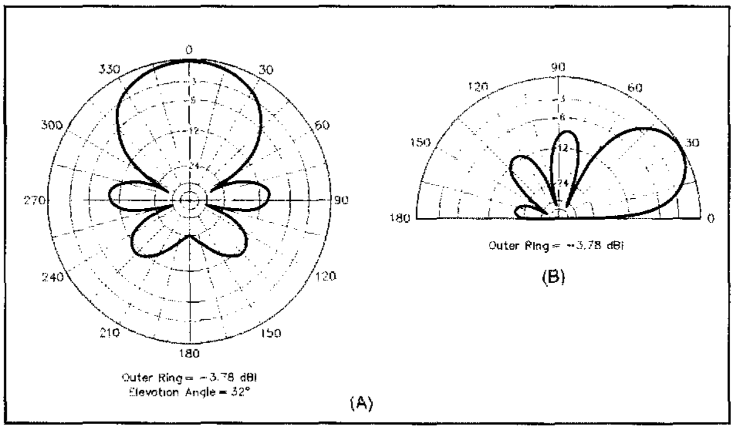

Fig 1—A 400-foot, 8-foot-high terminated Beverage antenna displays significant side lobes that compromise its ability to reject noise and interference in the horizontal plane (A). Evaluating the Fig 1 antenna in the vertical plane (B) tells a similar story. The secondary vertical lobes bring in interference and noise from near-vertical-incidence signals—that is, from sources that are relatively local. (*patterns modeled over average earth at 3.8 MHz with* MN)

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[1] Notes appear on page 34.

Pattern Predictions

I wanted the array's pattern to be easily reversible, so I picked element phase delays of 90° or 135° to simplify the feed system. A 1:2:1 element current ratio gave optimum directivity with 40-foot (center-to-center) element spacing. Fig 2 shows the predicted azimuth and elevation patterns obtainable with an antenna configured in this way. Over average earth, the optimum arrival angle is 29°, but the antenna has an excellent pattern at low and high angles (typical of loops, per Hart). Its horizontal 3-dB beamwidth is 78°.

Electrical Loop Design

Fig 3 shows the schematic of the loops. Small, tuned loops generally have low efficiency, but this is not an important factor for receiving at 3.8 MHz. Using the radiation resistance formula in Hart's book, the loops' radiation resistance computes to about 0.005 Ω each. Measurements have shown that the total feed resistance, which includes the wire loss, transformer and capacitor losses, is about 1 Ω. Based on these numbers, the efficiency works out to 0.5%, or −23 dB. Although a low-noise, high-dynamic-range preamplifier is required to preserve system sensitivity, the normal QRN level at 75 meters, not the array's inefficiency, limits system performance.

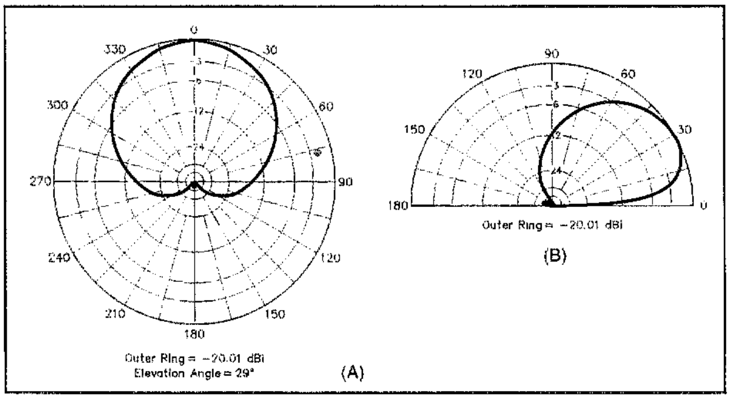


Fig 2—Properly spaced, phased and driven, three ground-mounted loops configured as a *binomial end-fire array* can provide an almost perfect *cardioid* (heart-shaped) horizontal response (A), with a beamwidth almost identical to that of the Beverage modeled in Fig 1. The loop array's lower gain (at least 16 dB lower than that of the Beverage) requires preamplification, but does not compromise weak-signal performance at 3.8 MHz. In the vertical plane (B), the loop array provides *much* better rearward-signal discrimination than the Beverage.

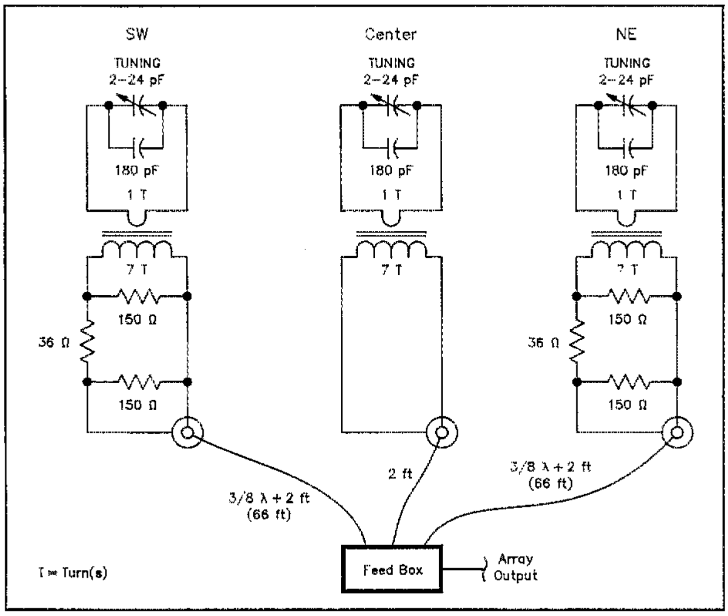


Fig 3—Each loop resonates at 3.8 MHz with a tuning capacitance of about 190 pF; the loops are spaced 40 feet apart (center-to-center). A transformer (turns ratio 7:1) matches each loop to 50-Ω line; 6-dB pads (use noninductive resistors) reduce the current contributed by the outer loops to half that contributed by the center one. The feed-line lengths shown are for RG-58A cable. Fig 5 details the feed-box circuitry.

The array's bandwidth (its region of highest performance) computes to about 20 kHz, although it can be used over most of the 80-meter band. Tuning each loop is accomplished with a low-value trimmer capacitor across a fixed 180-pF capacitor, for a total capacitance of about 190 pF per loop. I recommend using high-quality fixed capacitors, such as those made by ATC, JFD or Vitramon. Ordinary "receiving" variables and dipped micas can be used if *low*-level RF is applied to the array during adjustment. I applied 5 W to my array during adjustment—a level at which more than *700 V* appears across the loop tuning capacitors. (A sudden SWR or return-loss change during loop adjustment indicates probable capacitor failure.)

Each loop's ferrite matching transformer has a one-turn primary consisting of a brass tube passed through stacked toroidal ferrite cores. Seven turns of insulated wire wound through the tubing form the secondary. The primary connects to the loop; the secondary, to 50-Ω coax via coaxial fittings. My transformers are based on commercial assemblies,[7] but you can build your own using hobby brass stock as shown for transistor-final transformers on page 60 of *Solid State Design for the Radio Amateur*.[8] I suggest using four 0.5-inch-OD -43 (μ = 850) toroidal ferrite cores per transformer. PC-board material can serve as tube end plates as shown in *Solid State Design*.

Construction

Each loop consists of #14 copper wire housed in a PVC-pipe frame. About 13 feet of wire is needed per loop half—in all, roughly 80 feet for a three-element array. The loops are tuned at their tops and fed at their bottoms.

Fig 4 and the title photo show the loop partially assembled and completed. One-inch PVC pipe forms each loop frame. PVC pipe comes in 10-foot lengths, so my loop dimension of 5 feet per side allows the use of standard-length sections without waste. Each loop's tuning capacitors mount in a 1-inch cross centered in the loop top. Each loop's feed (a 7:1 transformer and BNC coaxial connector) mounts in an assembly centered in the loop bottom. This assembly consists of 1½- to 1-inch reducers, a 1½-inch cross, a 2- to 1½-inch reducer, a 1½-inch cap and a 2-inch cap. The 2-inch cap holds the loop feed components; the 1½-inch cap closes the uncommitted cross port. The 1½-inch-to-1-inch reducers can be seen on the ends of the two pipes in Fig 4. The necessary PVC cement, crosses, elbows, pipe, reducers and Ts are available where plumbing supplies are sold.

Elbows form the loops' upper corners; Ts, the lower. Sixteen-inch pieces of 1-inch pipe (not visible in Fig 4) cemented into the bottom of the Ts allow the loops to be ground-mounted on timbers.

A Five-Element, End-Fire Binomial Array Design

The three-element end-fire binomial array described in this article easily fits into the space previously occupied by my 400-foot Beverage antenna. A *five*-element array would also fit into that space, so I designed one with the dimensions shown in Table A. Modeling predicts that all of its side and back lobes should down at least 50 dB at all angles of elevation (see Figs A and B). Pattern attenuations of this magnitude may not be achievable in practice because highly precise phasing and element drive are required, but it's comforting to know that there is no *theoretical* reason why directivity exceeding 50 dB cannot be achieved.
—WA2WVL

Table A

Vital Statistics for a Five-Element, End-Fire Binomial Array

Element Position	*Relative Current*	*Relative Phase*
Rear	1	270° (or −90°)
	4	135° (or −225°)
Center	6	0°
	4	−135°
Front	1	−270°

The elements are spaced 50 feet apart (center-to-center) for optimum side-lobe performance.

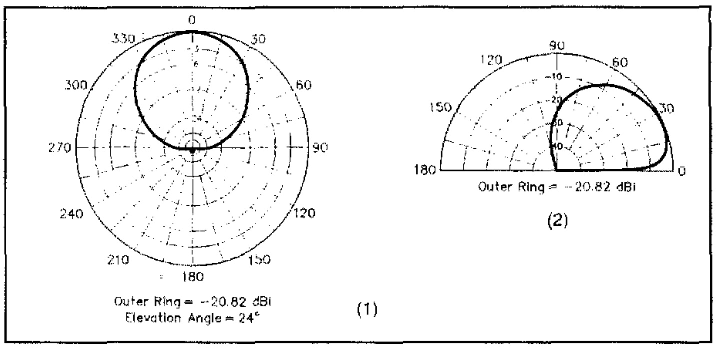


Fig A—Predicted horizontal (1) and vertical (2) performance of a five-element end-fire binomial array. Unlike the other polar plots in this article, which use *QST*'s standard log-periodic scale, the vertical-plane plot uses a linear-decibel scale to highlight the antenna's clean pattern at high attenuations.

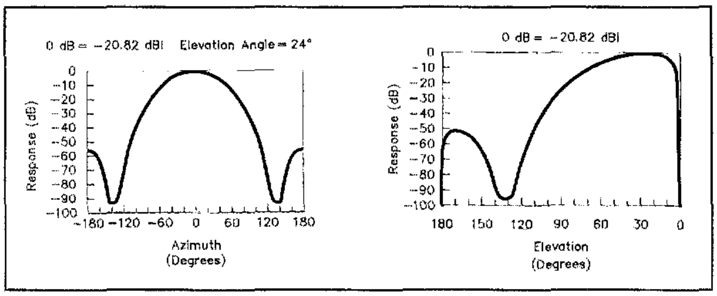
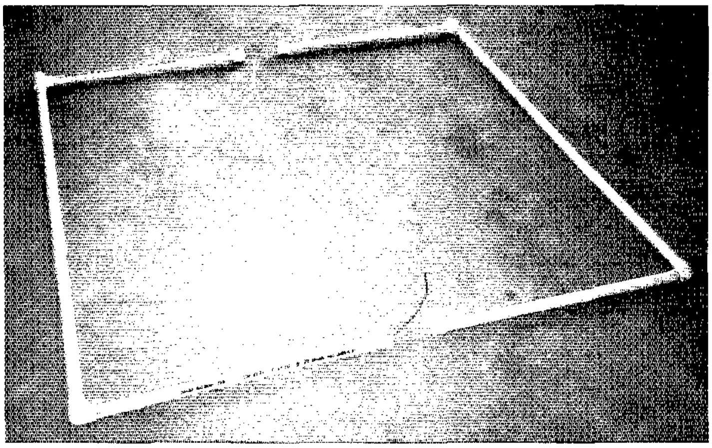


Fig B—Plotting the five-element array's horizontal performance with rectangular coordinates (1) shows its higher pattern attenuations better than the polar plots in Fig A. Critical adjustment of element current and phasing is necessary to approach such attenuations in a practical system. A rectangular-coordinate plot of the five-element array's vertical performance appears at 2.

Build your loops on a flat surface, such as arage floor, to ensure that the loop sides are line. Run the wire through each loop's PVC ces *before* assembling them so you won't ve to push the wire through.

Build each loop in halves, cementing all of its fittings except the center crosses. Once the halves' cement has dried, set the halves upright. *Without using PVC cement*, slip the center crosses on to complete the loop, bringing the loop wires out through the back of the crosses. Tap the sides of the Ts with a block of wood to seat the PVC pipes in the crosses.



4—This loop is ready to receive its center crosses—1-inch for the loop top reground) and 1½-inch for the loop bottom (background, via 1- to 1½-inch adapters).

I drilled small holes in the crosses to secure the loop-wire ends and create tie points. "Flying lead" construction of the transformer/connector and tuning capacitor assemblies allows them to be easily installed or removed. (Be careful not to reverse any of the transformer connections, as this would add an extra 180° of phase lag in that element and spoil your antenna's performance.)

Ground Mounting

Initially, I considered this antenna to be experimental, so I took a mounting approach that allows the loops to be easily repositioned. The local farm-and-garden store had a sale on pressure-treated 4 × 4 timbers, so I used two of these per loop as bases. A 1-inch hole drilled in the center of each timber takes a 16-inch-long piece of 1-inch hardwood doweling. To mount each loop, just slip its legs over the dowels. If you live in a windy area, you can add rope guys down to the base timbers to help support the loops, as shown in the title photo.

Phasing-System Design

The array requires relative element phasing as follows:

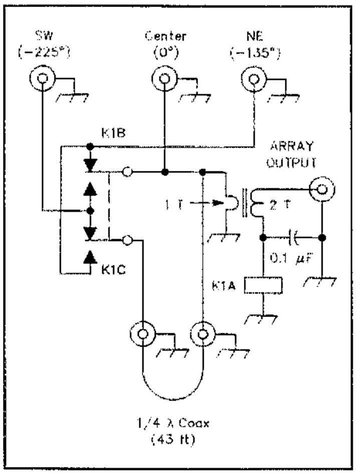


Fig 5—The feed box includes relay switching to reverse the array's pattern. The ¼-wavelength RG-58A section shown is *essential* and works in conjunction with the feed-line lengths shown in Fig 3 to provide proper element phasing. The phase angles shown are for normal operation (northeastern pattern, K1 off); energizing K1 flips the pattern to the southwest by phasing the northeastern loop at −225° and the southeastern loop at −135° relative to the center loop. The transformer (turns ratio, 2:1) is constructed similarly to the 7:1 units used in the loop; see text.

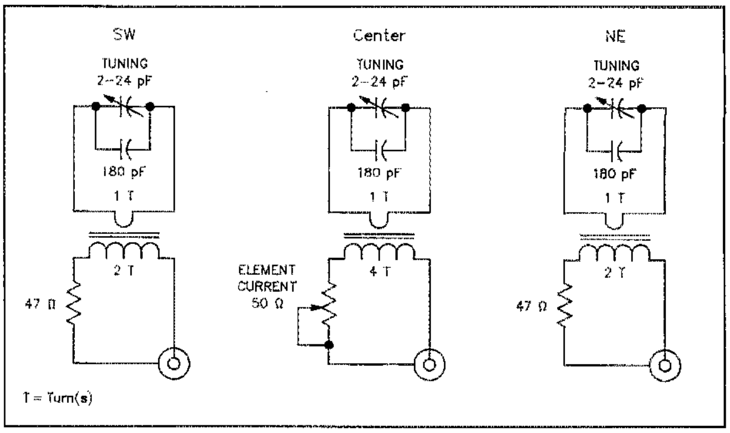


Fig 6—This alternative feed method uses noninductive series resistors and 2:1 and 4:1 matching transformers to allow precise adjustment of element currents for improved front-to-back ratio. Careful adjustment of this system has achieved a peak improvement of 20 dB over the simpler feed method shown in Fig 3. Fig 7 plots its front-to-back ratio versus frequency.

Rear	+135° (or −225°)
Center	0°
Front	−135°

Transposing the front- and rear-element feeds reverses the array's pattern. Fig 5, the schematic of the array's feed box, includes a method of multiplexing the pattern-control relay drive over the box-to-station coax.

Equal lengths of RG-58A coax run from the feed box to the front and rear elements. These cables are ⅜ wavelength long, plus 2 feet. The center element is fed through 2 feet of cable, thus resulting in a differential phase between the center and end elements of 135°. (Assuming that the cables operate at a low SWR, this technique can be used in other phased-array designs where a ¼-wavelength coax is too short to span the physical distance between elements.) Then insert a ¼-wavelength piece of coax (with RG-58A, 43 feet at my array's design frequency of 3795 kHz) in series with the rear element to give a total delay of 225°. A 2:1 ferrite-core transformer matches the three paralleled feed lines to 50 Ω in the feed box.

Adjustment and Measurements

Computer modeling shows virtually no coupling between the loops, so each can be tuned to resonance while ignoring the others. A sensitive SWR meter or return-loss bridge should be used to adjust the variable capacitor for the best impedance match—that is, minimum reflected power or highest return loss. Adjust the loop trimmer via a long, nonmetallic shaft or large plastic knob. Be careful not to get RF burns from the capacitor—high RF voltages appear across it even with only a few watts applied to the loop.

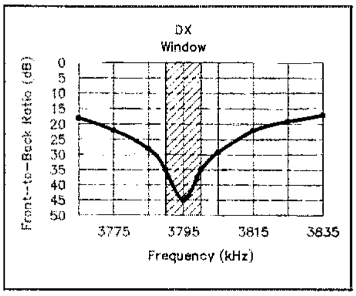


Fig 7—Adjusted for peak performance at 3795 kHz using the feed method shown in Fig 6, the three-element loop array provides a front-to-back ratio of at least 35 dB across the 75-meter DX window.

The loops give less than 1.09:1 SWR (a return loss of 27 dB) when fed via 7:1 transformers. The first three-element array I built used 6-dB attenuators in the feeds to the 1st and 3rd elements to set those elements' current at half that of the center element. After the loops were tuned and hooked to the feed box, a front-to-back ratio (F/B) of 25 dB was measured without further adjustment.

The second three-element array I built used series-resistor feed (Fig 6) to enable precise adjustment of the element current. I varied the outer elements' feed-transformer turns ratio to set their current; then added resistors to bring their input impedance up to 50 Ω. (The resistors also add 7 dB of loss.) I adjusted this array for maximum F/B using a remote test signal. The element trimmers can be used for fine adjustment of the element phase. (If you can't adjust your antenna with a remote test signal, use 7:1 transformers at all three elements and set the outer elements' currents with 6-dB pads. A 25-dB front-to-back ratio is well worth achieving at 75 meters!)

Fig 7 plots the system's measured front-to-back ratio using series-resistor feed. I consider this performance to be well worth the effort necessary to achieve it.

Notes

[1] *MN 4.0* by Brian Beezley, K6STI.

[2] J. Kraus, *Antennas* (New York: McGraw-Hill, 1950), pp 90-97 and 510.

[3] H. Jasik, *Antenna Engineering Handbook*, 1st ed (New York: McGraw-Hill, 19), pp 5-20 to 5-28.

[4] W. L. Weeks, *Antenna Engineering* (New York: McGraw-Hill, 1968), pp 84-97.

[5] J. Devoldere, *Low-Band DXing* (Newington: ARRL, 1986).

[6] T. Hart, *High Efficiency Antennas alias The Loop* (Melbourne, FL 32902: W5QJR Antenna Products).

[7] Transformer assembly #CN20-AT8, manufactured by Ceramic Magnetics, 16 Law Dr, Fairfield, NJ 07004, tel 201-227-4222, fax 201-227-6735. Attn: Jim Florance.

[8] W. Hayward and D. DeMaw, *Solid State Design for the Radio Amateur* (Newington: ARRL, 1986).

Floyd Koontz holds a BSEE in Electrical Engineering and is currently a senior scientist working for Harris Corporation, RF Communications Division, Rochester, New York. He has 35 years experience in the design of high-power transmitters and antenna systems.

For many years, he was active on the bands from 50 through 3456 MHz, including moonbounce communication at 144 and 432 MHz. Prior to moving to New York in 1961, he was licensed in Missouri as KØLTV and, in the late 1950s, as W9JQA in Illinois. In 1965, he was one of the DXpedition operators at XE5L, Revilla Gigedo.

Recently, he has concentrated on 75 meters, operating with various receiving antennas and a six-element phased vertical array for transmitting.

QST

Persistence Gets the Derelict

When a high-altitude balloon failed to respond to its ground control crew's radio commands, the recovery teams embarked on an all-night, 1000-mile chase.

By Warren Williams, N0PBY
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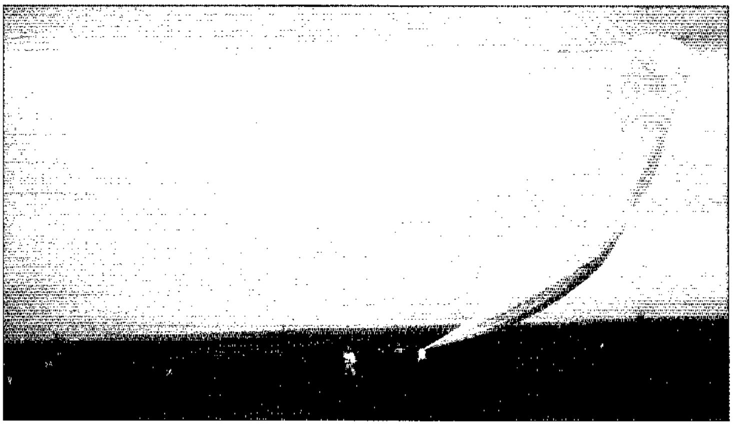

Partially inflated, the balloon is nearly ready for launch. *(photo by Ann Trudeau, KA0ZFI)*

It was with confidence born of five previous successful flights, that we approached the May 30, 1992, flight of our sixth and largest balloon. Edge of Space Sciences (EOSS), a nonprofit group of hams, high school and college students, and other people interested in science, space, radio and astronomy, devoted to educational experimentation, were confident of a spectacular large balloon flight. Little did we know what the next 31 hours would hold. The record for the longest successful balloon chase in the history of Amateur Radio was about to unfold.

Jack Crabtree, AA0P, obtained FAA approval for a flight to more than 100,000 feet with a payload of more than 20 pounds. The launch site was just north of the Terminal Control Area for Denver International Airport. Little difficulty was encountered in getting the approval because the previous five flights were completed without significant trouble. The earlier experiments didn't always work as planned, but we had a perfect record for release and recovery of the payload. The FAA didn't know that it would be dealing with a "derelict balloon" for the next 18 hours.

Balloon and Payload

The balloon was a "special-order Raven" filled with 54,000 cubic feet of helium, 71 feet long and designed to hover at 100,000 feet. The projected size of the balloon at altitude was 37 feet high and 52 feet wide. The 14-pound balloon was the largest launched from Colorado in recent history.

The payload included packet telemetry on 144.34 MHz, sending inside and outside temperatures, altitude and voltage levels. Power output was one watt to a vertical antenna. Amateur television (ATV) on 426.25 MHz was taken with a black-and-white camera that was gimballed and controllable in the vertical direction. Power output was one watt. The ATV audio output was used to send the VHF omnidirectional radio range (VOR) audio to be used for real-time tracking and position plotting.

In addition, there was a 2-meter tracking beacon on 147.555 MHz with a one-second beep and CW ID every 10 minutes. The antenna was vertically polarized and output was 250 mW. Neal Tenhuzen, WB0ADU, designed and built the beacon. It was housed in a styrofoam "Big Wheel," designed by the science students at Heritage High School. The design was a winner of a contest to see which team of high school students could design the best container to withstand a drop from 200 feet and land with the antenna in an upright position. There was also a 10-meter backup tracking beacon on 28.321 MHz with a vertical antenna and 250 mW power output.

The prime experiment was the "Humble II" telescope experiment, designed by Tim Kelliher, N0RHE, a student at the University of Colorado, and students from Green Mountain High School. This was an experiment to record the ultraviolet images of the sun from 100,000 feet through a magnesium filter on 400-speed black-and-white film.

The VOR experiment was designed by Mike Manes, W5VSI. The purpose of this experiment was to see if it was feasible to use three VOR stations in the Stapleton International Airport area for real-time positioning and direction finding (DFing). The directional AM signals from the Gill, Denver and Kiowa VORs were received and retransmitted to ground tracking on the audio signal from the ATV link. The command receiver and flight computer were designed by Bob Schellorn, W6ORE, and a five-cell lithium battery made up the remainder of the payload. The total balloon and payload weight was 35 pounds.

Launch and Chase

The morning of the launch was cloudy, but the surface winds, our biggest concern, were calm. The FAA granted approval for the launch despite the cloud cover, but requested expediting the launch because of incoming air traffic at Stapleton. The tarp material needed to cover the alfalfa field to protect the balloon arrived late. We were feeling time pressure.

EOSS Chief Scientist Dave Clingerman, W6OAL, and Merle McCaslin, K0YUK, began filling the balloon with the four bottles of helium, just enough to lift a 40-pound bucket of sand. The tension and drama began to build. At 0938 MDT, the balloon was sent aloft in calm surface winds. Tom Isenberg, N0KSR, and a group of high school students had already obtained the winds aloft and projected the flight path and touchdown point with Tom's computer program. The foxhunters were already headed downwind with sufficient geographic separation to assure good cross-bearing DFing. There were 20 sophisticated mobile recovery vehicles stationed as far as 50 miles downwind. Some were equipped with packet to receive the telemetry from the balloon, ATV, and 2- and 10-meter DFing gear.

A 2-meter mobile repeater was stationed on high ground southeast of Greeley by the Colorado Repeater Association to provide field communications on 147.835 MHz. The location by grid square from an overlay of the Colorado Recreation Map was called in on this frequency by each station, with their true bearings every 15 minutes, and this informa-

Cooperative RDFing with a Notebook Computer

By Paul Ternlund, WB3JZV
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When Edge of Space Sciences Inc (EOSS) launches a balloon carrying Amateur Radio gear, tracking and recovery becomes the job of a skilled radio direction-finding (RDF) team. A Macintosh PowerBook 140 computer running augmented triangulation software aids the crew.

Background

After participating in previous EOSS balloon chases, I felt there was a better way to apply the bearings from cooperating RDF team members. My goal was to provide the team's field coordinator with accurate transmitter-location estimates quickly, using all available bearings. Before using this technique, the chase team coordinator would gather bearings from his RDF team and then estimate the balloon's position on-the-fly. Although this has worked well—because of the coordinator's skill—it was prone to error. The computer technique handles the number crunching, and consistently does the calculations quickly and accurately. Even though this system was created for chasing balloons, it's equally effective for locating interfering transmitters or hidden transmitters during cooperative foxhunts.

Software Summary

The technique exploits the features of the Microsoft *Excel* spreadsheet program, with two custom macros. To use the system, the RDF team employs a standard map and grid system. All bearings are taken at the same sample times. Using a radio net, each station reports its X and Y coordinates and bearing to a computer coordinator on request. The coordinator enters the data into the computer.

One macro performs the triangulation task that has four automated parts: (1) Triangulation, (2) Position estimation, (3) Plotting, and (4) Station performance monitoring.

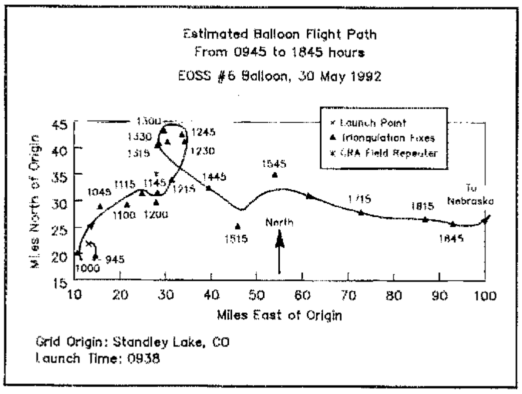

Fig 1—This results chart for the May 30 flight was generated by the laptop computer after the balloon's recovery. It shows the track of the renegade balloon at the DF sample times throughout the flight.

Part 1, *triangulation*, uses trigonometry to determine where lines of bearing cross. The computer performs a triangulation calculation for every unique bearing pair whose bearings converge. Twenty bearings can produce as many as 190 unique points for the beacon's position. In the real world, incorrect bearings result in scattering of these points. There lies the problem: Which point is the best approximation—if any? Many causes, including multipath and human error, prevent triangulated points from superimposing.

Part 2, *position estimation*, determines one location estimate from the multitude produced in Part 1. After experimenting with ways to determine a practical location estimate, I feel double averaging is a winning technique. This technique assumes that *most* bearing data is satisfactory. It's only tolerant of bad bearings if they're in the minority. This is generally the case with our experienced team of 6-20 trackers.

Part 3, *plotting*, produces a chart showing hunter positions, "surviving" triangulated points and the transmitter's estimated position. For this feature, *Excel*'s scatter charting capability provides the solution.

Part 4, *station performance monitoring*, keeps track of points dropped in Steps 1 and 2 because of problematic bearings. From this data, a station-performance report can be produced that shows the bearing-rejection rate for all participating stations.

The second macro saves the results produced by the first macro and computes the balloon's speed and heading.

Software Example

To illustrate the augmented triangulation macro, here's an example: We begin with a piece of graph paper and draw X and Y axes and an origin at the lower left (see Fig 2). The plus and minus X axis denotes miles east and west of the origin, respectively. The plus and minus Y axis denotes miles north and south.

We draw five points to represent five RDFer positions and note their coordinates. Then we draw a circle with a one-mile radius (for realism) that we'll say contains the transmitter to locate.

We draw bearing lines from hunters 1, 2, 4 and 5 to somewhere in or on this circle. Let's say that station 3 has taken a bad bearing and consequently, its bearing passes a

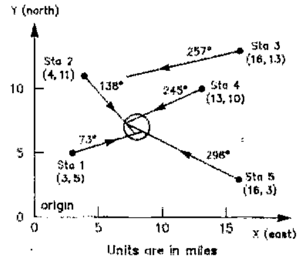

Fig 2—Fabricated example (see text).

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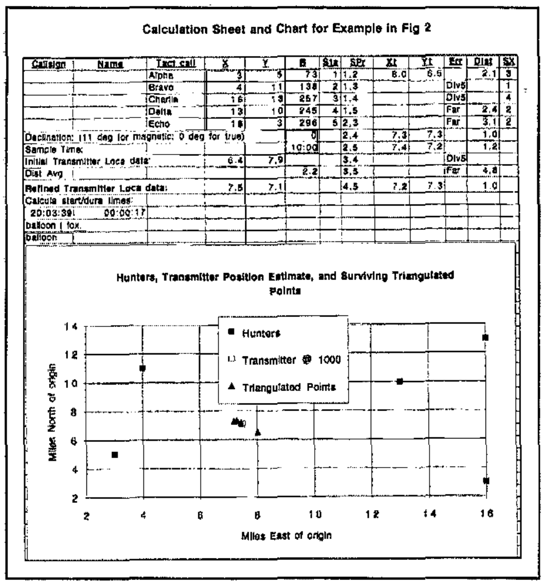
tion was plotted with a computer programmed by Paul Ternlund, WB3JZV. The program threw out inconsistent readings and plotted the estimated position, speed and flight direction of the balloon (see the sidebar). Fig 1 shows the computer-plotted position of the balloon at each reporting period. The flight path takes two interesting loops before taking off to the east. With all this organization and effort, how could we miss?

The shroud lines to the parachute ran through an eyebolt affixed to the bottom of the balloon. Unfortunately, the painter—a rope to hold the balloon while the payload is readied for launch—also ran through the eyebolt. As it turned out, pulling the painter out of the eyebolt at launch had partially cut the shroud lines, but this wasn't known at the time. Within moments after launch, the payload dropped several feet from the bottom of the balloon. It was hanging from the rip panel of the balloon. The sudden jerk separated the power hookup to the camera, telescope and



Calculation Sheet and Chart for Example in Fig 2

Callsign	Name	Tact call	X	Y	B	Sta	SPr	Xt	Yt	Err	Dist	SX
		Alpha	3	5	73	1	1.2	6.0	6.5		2.1	3
		Bravo	4	11	138	2	1.3			Div5		1
		Charlie	16	13	257	3	1.4			Div5		4
		Delta	13	10	245	4	1.5			Far	2.4	2
		Echo	16	3	296	5	2.3			Far	3.1	2
Declination: (11 deg for magnetic; 0 deg for true)					0		2.4	7.3	7.3		1.0	
Sample Time:					10:00		2.5	7.4	7.2		1.2	
Initial Transmitter Loca data:			6.4	7.9			3.4			Div5		
Dist Avg					2.2		3.5			Far	4.8	
Refined Transmitter Loca data:			7.5	7.1			4.5	7.2	7.3		1.0	
Calcula start/dura times:												
20:03:39	00:00:17											
balloon / fox.												
balloon												

Fig 3

Paul Ternlund, WB3JZV (l), teams up with RDF partner Tim Moffitt, N0NXI, who brings his Apple Macintosh laptop computer out for balloon-tracking "foxhunts." *(photo by Stephen Ternlund)*

few miles north of our circle. Using a protractor, we note the bearing angles for all five bearings relative to true north (0° true is straight up along the Y axis).

We enter the hunter's (X, Y) coordinates and their bearings into the computer Calculation Sheet (see Fig 3). After data entry, the *triangulate* button is mouse-clicked. About 17 seconds later, the computer reveals the estimated transmitter-location coordinates at (7.5, 7.1). From the *Err* column in the Calculation sheet, we see station pairs (1,3), (1,4) and (3,4) each have divergent bearings and were ignored. The triangulated points produced by station pairs (1,5), (2,3) and (3,5) were too far from the centroid and were ignored. In this example, the only surviving points were those generated by bearings from station pairs (1,2), (2,4), (2,5) and (4,5). The computer tallies station performances in the *SX* (station "hit") column. A station's *SX* number is incremented by one each time one of its bearings produce a point that the system chooses to ignore. In our example, station 3 was "dinged" four out of four times, making its performance most suspect.

Results of May 30 Flight

Fig 1 shows the calculated balloon flight from launch at 0938 to 1845 Mountain Time. A quick review of the stations with bearing-rejection rates of more than 51% (see "How it Works, Part 4") yields interesting results: Mike Manes' (W5VSI) bearings (tactical call signs *Oscar* and *November*) had rejection rates of 56% and 83%, respectively. Mike was running his own experiment to receive aeronautical VOR signals aboard the balloon. Mike determined later that his telemetry readouts were incorrect. George Reidmuller, N0NJM, had a 57% rejection rate. It turned out that George shut down his station following the 1130 sample. Some unknown station assumed George's tactical call sign without our knowledge; hence we continued to incorrectly use George's fixed coordinates in our calculation! (George's actual performance was solid, with only a 33% rejection rate.) Jack Crabtree, AA0P, revealed that the coax between his rig and DF antenna was too short. This caused him to place his antenna too close to a metal house trailer. He believes that this biased his bearings (he had a 73% rejection rate).

Conclusion

Mobile with a Mac PowerBook running augmented triangulation software compliments other contrivances of an RDF team tracking a balloon. It enhances the output of the team's cooperative efforts. It gives consistent estimates for the balloon's location. It charts the hunter positions with the triangulated points and estimated balloon location. It also monitors each station's bearing performance for follow-on analysis. This gives the team the upper hand, while chasing a balloon that can be as slippery as the foxiest fox.

A Macintosh 3½-inch floppy disk with *Excel* 3.0 and *Excel* 4.0 macros, relevant spreadsheet templates, a user's guide and results of EOSS balloon chases is available from Paul Ternlund for $10.

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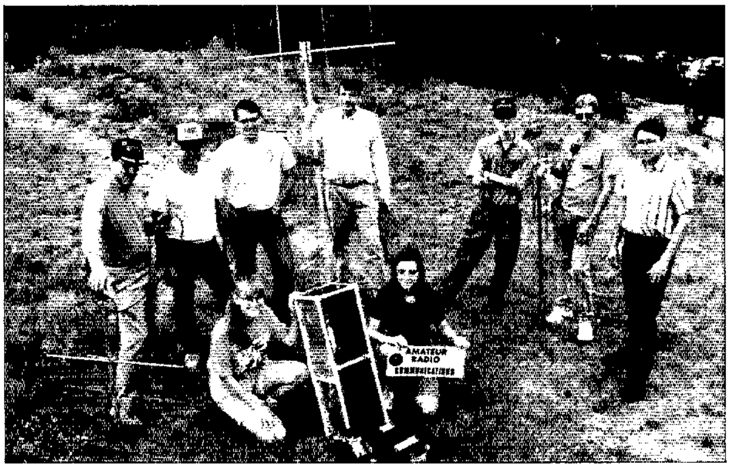
the computer, and the shape of the balloon appeared abnormal.

The balloon rose to 75,000 feet and hovered there. Packet telemetry continued to report temperatures, altitude and voltages. Video of the balloon ascending through the cloud cover was spectacular, including a snowstorm within the clouds. The quality of the video was P5—beautiful!

At 1225, the command for separation was given, but there was no response. At 1300, an automatic timer signalled release to the nichrome wires on the shrouds, but without effect. Several more attempts at commanding release were given, all without effect. We were in for a long haul.

We notified the FAA and the balloon was given a "derelict" designation. We were required to report the position of the balloon every 15 minutes to the FAA at Stapleton. At 1400, the balloon was drifting over Greeley. The chase teams were preparing for the long haul, getting gasoline and food. Shortly thereafter, the balloon began moving east at



The proud chase team displays its direction-finding gear, used to track the balloon's beacons. *(photo by Larry Cemey, N0STZ)*

increased speed, and the chase teams began moving east, but continued to maintain geographic separation. By 1600, the chase teams had moved far enough east that the field repeater was no longer effective. Field communications moved to the 146.94-MHz wide-coverage repeater of the Rocky Mountain Radio League. Eventually, field communications were accomplished by simplex VHF FM in the field and on 40 meters from the field to the base in Denver. As evening approached, communications back to Denver changed to 80 meters for the night.

There were large thunderstorms in the area, and by 1745, the last bearing was taken, showing the balloon moving east of Sterling, 100 miles northeast of the launch point. Shortly thereafter, the 147.555-MHz and 10-meter beacons were lost. Occasional packet bursts were heard, but it was difficult to DF on such a short signal burst. An altitude of 54,000 feet was reported by balloon packet, so it was slowly descending at about 2000 feet per hour.

A repeater trustee, David Richendifer, WD0HNQ, from Denver, called net control to report a "funny sounding" CW beep on his repeater. This was identified as our beacon, which had apparently gotten too cold and lost its phase-lock loop, resulting in an upward-wandering frequency. It's interesting that the frequency rose from 147.555 MHz to as high as 147.91 MHz, so the chase team not only had to DF the balloon, but also had to change frequencies and grope to establish reliable coordination frequencies for the chase teams and long-haul communications to Denver on 80 meters.

The chase continued into Nebraska, and only one chase team, Marty Griffin, WA0GEH, had a Nebraska map. Country roads were difficult to find and navigate at night. All teams were heading east as fast as possible to keep up with the balloon and not allow it to get beyond the horizon. Mike Manes, W5VSI, and Geno McGahey, AL7GQ, remained on 40 and 80 meters in Denver to provide what support they could and to notify the FAA of the location and progress of the balloon and hunters.

At 0100, there were five chase teams active and moving into Nebraska. Calls were made on the 146.94-MHz repeater in North Platte. As the teams neared that area, a local ham, Randy Allen, N0MPF, came to the rescue and helped with road locations. The balloon was plotted to be north of North Platte. The chase teams headed by Greg Burnett, K0ELM, turned north. No main roads heading north could be found, so it was on back roads in knee-deep sand at night that the chase continued.

At 0200 on Sunday, May 31, the 147.555-MHz beacon, back on its proper frequency, was on the horizon, and Bob Ragain, WB4ETT, and his two daughters, Dawn, N0QCW, and Colleen, N0QGH, were picking up the beacon with an S7 signal level with 60-dB attenuation in place. The signal disappeared at 0330, and then was picked up in a large field by the chase leader, Greg. He was unable to tell the other chase teams of his location, however, because of the lack of a map. Bob Ragain did what any good foxhunter would do: He DFed him. The teams finally met at the field and waited for sunup to begin the walk.

At about 0515, the teams walked 1½ miles to the payload. The 2-meter beacon, which dangled below the main payload, was stuck to a yucca plant, and the remainder of the payload and parachute were strung out downwind. In a remarkable feat, one member of the chase team, Dan Griffin, driver and navigator for Marty, hiked several miles downwind and found the balloon—another first.

The five chase teams returned home to a tumultuous greeting on the Denver Radio League's 146.64-MHz repeater. It had been 31 hours since they last saw home, and they had logged more than 1000 miles each in their fine effort.

Epilogue

Final statistics for the flight:

Distance: 219.5 miles (record)
Flight Duration: 18 hours (record)
Launch to recovery time:
  19 hours 30 minutes (record)
Maximum altitude: 75,000 feet
ATV was received with P5 quality.
Touchdown site: 38 miles northwest of North Platte, Nebraska

Postflight payload inspection revealed that the shroud lines to the balloon had been severed, probably by the painter as it was pulled from the eyebolt at launch. The drop of the payload put tension on the power leads to the flight computer and camera, rendering them inoperative. When power was reestablished back at home base, the nichrome cutdown mechanism worked on the first try. If only the painter hadn't cut the shroud lines!

The film in the telescope camera was unexposed, although evidence that the telescope had properly tracked the sun was in evidence by the burned track marks along the plastic body of the camera leading to the lower area of the lens. The mirror had focused the sunlight well enough to produce enough heat to leave its mark.

The FAA later stated that it was pleased with the conscientious effort and persistence of the group. What a magnificent smorgasbord of communications and dedication by a great group of hams.

Strays

SPECIAL-EVENT NET

◊ The National Radio Astronomy Observatory (NRAO) RC will operate a special-event station from the NRAO's Array Operations Center in New Mexico for the dedication of the NRAO's Very Long Baseline Array (VLBA), the world's largest dedicated astronomical instrument. NA5N will be on 10, 15, 20, 40 and 80 meters between 1800-0200Z August 21-22. A net including amateurs from the VLBA antenna sites across North America and other NRAO observatories will be held on 14.250 MHz at 1800Z August 21. For a QSL card, send your QSL and a self-addressed, stamped envelope (SASE) to NRAO ARC, PO Box O, Socorro, NM 87801.

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QST

September 1993 $3.00

devoted entirely to Amateur Radio

Five-Band Worked All States!

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OUR COVER

When he set out to earn an ARRL 5-Band Worked All States award, Leonard Kay, KB2R, of Maynard, Massachusetts, began an absorbing and dedicated endeavor. This is a sampling of the 250 QSL cards he collected to achieve 5BWAS. His account of how he met the challenge begins on page 39.

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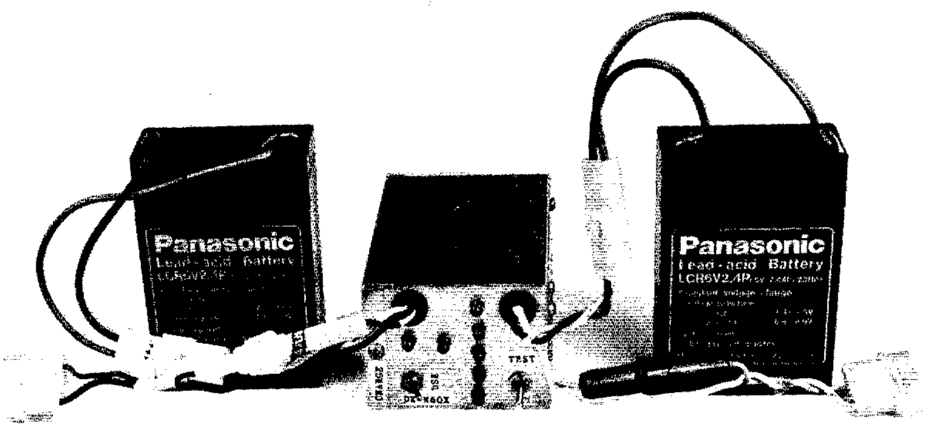

It's inexpensive, portable and you can build it yourself!

By Thurman Smithey, N6QX
56 Center St
Chula Vista, CA 91910

It seems that almost every ham owns a VHF and/or UHF hand-held transceiver (H-T), undoubtedly because they serve so many purposes. Sometimes, however, an H-T's utility is limited by its standard-issue, (usually) short-duration batteries. If you're providing public-service communications for an all-day event, for example, you may find that your battery has died long before your stint is over. There are many other situations, including emergencies of all kinds, where a portable, heavy-duty H-T power source would prove advantageous.

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Reconditioning Small Lead-Acid Batteries

Small lead-acid batteries are available for very little money at surplus outlets, swap meets and hamfests. I have learned a few things about these batteries that I feel are worth passing along.

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Once current flow is started, it can be increased by repeated charging and discharging until the battery begins to act very much like a normal battery. So far, however, I have not been successful in restoring more than about 60% of the original capacity of a battery that has been resurrected in this manner.

When shopping for a lead-acid battery, bring along a small load, such as a small 12- or 6-volt lamp, and use it to test the battery. If the battery lights the lamp, chances are reasonably good that you have a winner. If the battery is completely flat, you have your work cut out for you and may wind up with a mediocre at best. For example, I have one set of used 2.5 Ah batteries that did not require reconditioning. That set puts out as much or more power than the best of three different sets of 4.0 Ah batteries that *did* require reconditioning.

You'll find D-sized cells to be quite popular in the surplus market. They can usually be purchased as individual cells, or as packaged assemblies. I bought one 12-volt assembly (six cells) which I then split to make two 6-volt batteries. The assembly had a decent charge when I purchased it, and made two good 6-volt batteries.

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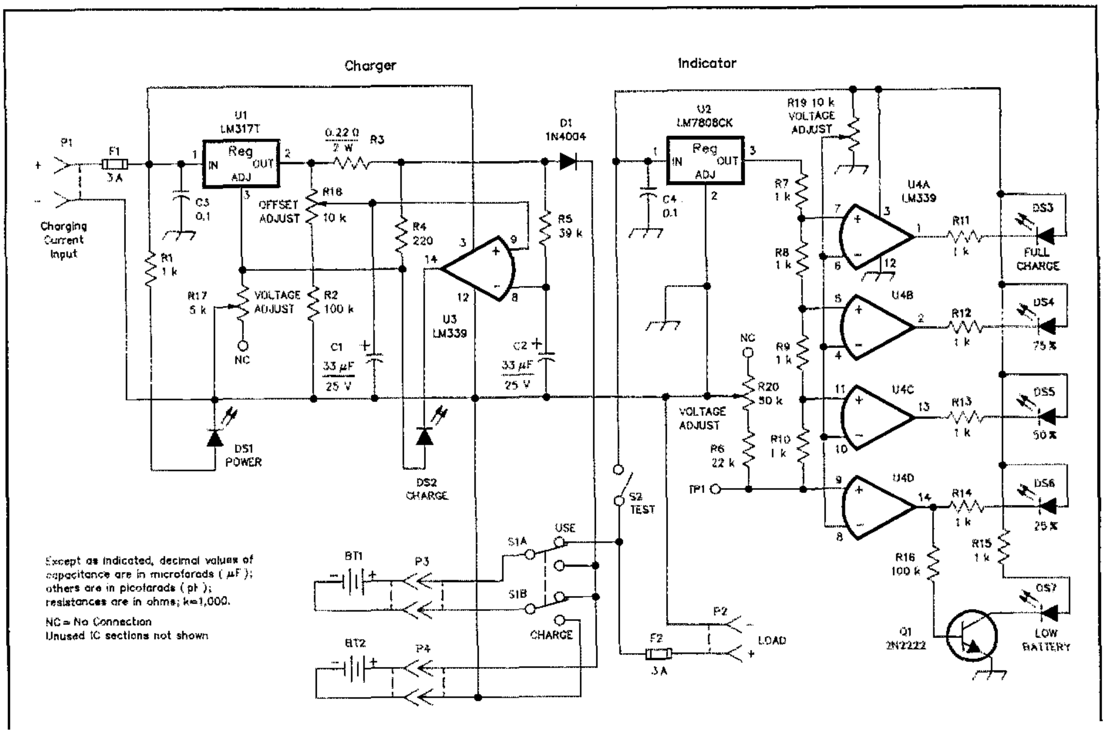

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NC = No Connection
Unused IC sections not shown

with an in-line connector and fuse. Use connectors that mate with whatever dc-power source you intend to use to charge the batteries. I use a cigar-lighter plug which allows me to charge my batteries from an automobile electrical system.

The cable connecting the batteries to your H-T or other device should also include an in-line fuse. If you intend to build the voltage regulator circuit, be advised that the voltage regulator IC should be mounted on a small heat sink. This regulator can deliver up to half an amp or so on transmit.

All parts used in the project, with the exception of the new batteries, are common parts which can be found in any electronics parts store or catalog. A source for the batteries is listed in the parts list (see Fig 1 caption).

Calibrating the Charger

After you've completed construction and checked your work, connect a 12-volt power source to the charger input (P3) *before* you install U3 and connect the batteries. DS1 should glow. Measure the voltage at pin 2 of U1 (referenced to ground) and adjust R17 until U1's output is 8.5 volts. Disconnect the power source and install U3. Reconnect the power source and measure the voltage difference between pins 8 and 9 of U3, adjusting R18 until pin 9 is 100 mV less than pin 8.

Calibrating the Battery-Condition Indicator

I have determined that no single calibration of the indicator unit is truly accurate with several different battery types. Fig 3 shows

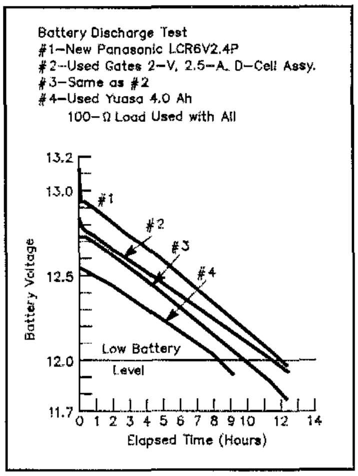


Fig 3—Battery discharge test results.

to 12 or less. This rule seems to apply regardless of the battery brand.

Fully charged battery voltage can vary considerably (see Fig 3). Depending on the brand, battery potential at the beginning of the discharge cycle can vary from 13 to 12.55 volts. Fortunately, the provisions for adjustment of the indicator are flexible enough to accommodate any batteries you're likely to find.

Calibrate the indicator based on the assumptions we've just discussed. In other

brating potential of 12.25 volts. Then, recheck for accuracy on all four LED set points. You should be able to get them all transitioning within 50 mV of the stated voltages.

For maximum accuracy, V_{Ref} should be measured with the same voltmeter you used to set the voltage at TP1. To calibrate the indicator for another battery with different characteristics, merely substitute the appropriate numbers in the equation above. To further improve accuracy, I measured all the 1-kΩ resistors used in the project and selected the four that were most nearly equal for R7 through R10.

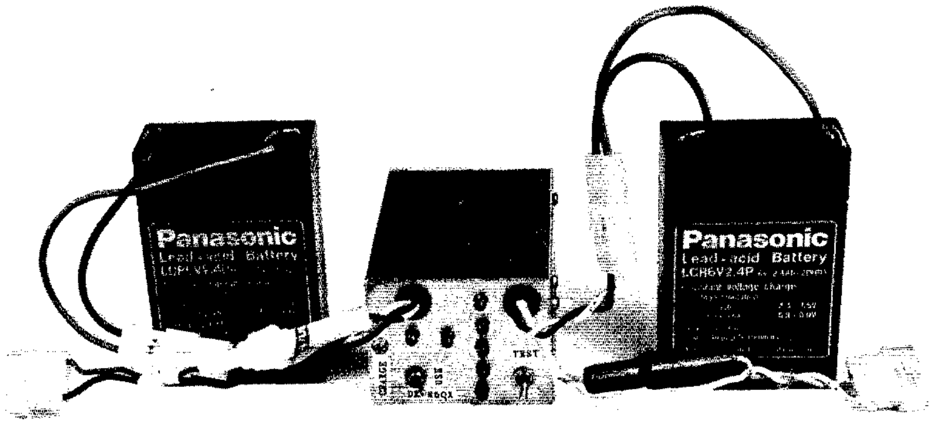
Additional Thoughts

I use a common fanny pack to house the batteries and the charger/indicator when I want to carry the system around. In the one I bought, there's plenty of room for the equipment and accessory cables. Although I elected to discontinue using large batteries, I was able to get a pair of 4.0-Ah batteries in the pack and the weight wasn't too objectionable. The box housing the charger/indicator gets quite warm while batteries are being charged, so it should not be left in the pack when charging is in progress.

Summary

I've had a great deal of enjoyment in developing this low-tech project, and even more enjoyment out of using the long-haul battery system with my own H-T. Try one yourself and I'm sure you'll like it as much as I do.

A Long-Haul H-T Battery System



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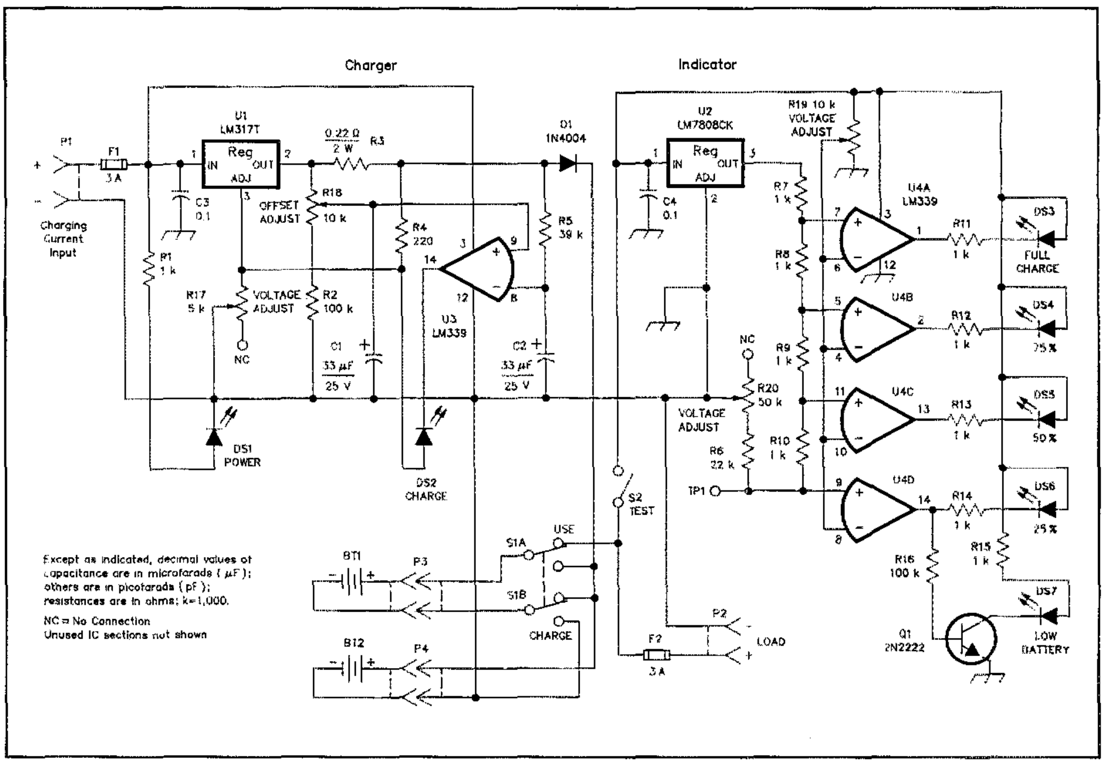

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Fig 1—Schematic of the long-haul H-T battery system. Resistors are ¼-watt, 5%-tolerance carbon-composition or film except as noted below.

BT1, BT2—Panasonic LCR6V2.4P (Digi-Key Corp, 701 Brooks Ave South, PO Box 677, Thief River Falls, MN 56701-0677, tel 800-344-4539. Digi-Key p/n P262), or Radio Shack 23-181.
C1, C2—33-µF, 25-V electrolytic.
C3, C4—0.1-µF ceramic disc.
D1—1N4004.
DS1—T-1¾ yellow.
DS2, DS3, DS4 DS5, DS6—T-1¾ green.
DS7—T-1¾ red.
Q1—2N2222.
R3—0.22 Ω, 2 W (Ocean State RM2-0.22).
R17—5 kΩ, ½ W, linear taper, 15 turn (Digi-Key 3006P-502-ND).
R18, R19—10 kΩ, ½ W, linear taper, 15 turn (Digi-Key 3006P-103-ND).
R20—50 kΩ, ½ W, linear taper, 15 turn (Digi-Key 3006P-503-ND).
S1—Miniature double-pole, double-throw toggle switch.
S2—Miniature single-pole, single-throw momentary normally open switch.
U1—LM317T adjustable voltage regulator (Radio Shack 276-1778).
U2—LM7808CK voltage regulator (Ocean State 7808).
U3, U4—LM339 quad comparator (Radio Shack 276-1712).

admittedly) *continuously* for 2½ days. Charge time for the 2.5 ampere-hour (Ah) battery is only 6 to 8 hours.

System Description

My long-endurance H-T battery system requirements included:

- **Battery Charging**: The battery must be chargeable from any 10- to 15-volt dc source.
- **Automatic shut off**: The charger must shut off automatically when the battery is completely charged. An indicator must be provided to signal when charging is complete.
- **Discharge level indication**: There must be an accurate means to indicate the discharge level of the battery *as it is being used*.
- **Output regulation**: The battery output voltage must be regulated to suit the requirements of any H-T that can't be operated directly from its 12-volt output.

The Batteries

Battery choice is very important. I selected sealed, paste-electrolyte, lead-acid types. They hold their charge better than NiCds and they're readily available at reasonable prices. I chose two 6-volt batteries which are paralleled for charging, then connected in series to provide a 12-volt source for powering H-Ts. An added benefit of this switchable series/parallel approach is that it allows the use of either battery to supply 6-volt loads (video cameras, video lights, portable electric lanterns and so on).

In addition to the battery, the other three parts of the system are the charger, the battery-condition indicator and the output regulator.

Charger Circuit Description

A sealed, 12-volt lead-acid battery (2 to 4 ampere hours [Ah] capacity) is fully charged when its terminal voltage reaches about 15 volts and the charge current has dropped from its initial value to about 0.25 amperes. This assumes that the charging source maintains a constant voltage at the end of the charge cycle. The charger shut-off circuitry uses this current drop to define the full-charge condition.

The batteries are connected through connectors P3 and P4. S1 is placed in the **CHARGE** position to connect the batteries to the charger circuit. Charge current is supplied through connector P1 and applied to the input (pin 1) of the LM317T regulator, U1. A yellow LED (DS1) lights to indicate the application of charging power. R17 sets the output of U1 to 8.5 volts. R18 is adjusted so that 100 mV appears between its wiper and the junction of R3 and pin 2 of U1. With power applied to the circuit and no current flowing in R3, this offset voltage appears between pins 8 and 9 of U3, an LM339 quad comparator. In this state, the voltage at pin 9, the noninverting terminal, is 100 mV less than the voltage on pin 8, the inverting terminal. Therefore, the comparator output at pin 14 is *low*.

With batteries connected to the charger, however, the initial charging current flow-

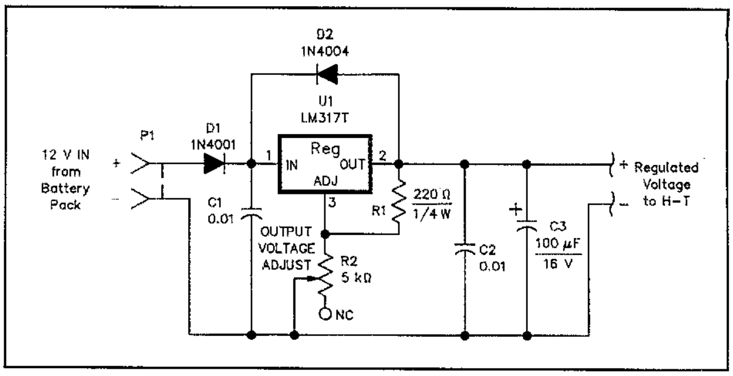

Fig 2—Voltage regulator schematic. Resistors are ¼-watt, 5%-tolerance carbon-composition or film except as noted below.
C1, C2—0.01-µF ceramic disc.
C3—100-µF, 15V electrolytic.
D1—1N4001.
D2—1N4004.
R2—5 kΩ, ½ W, linear taper, 15 turn (Digi-Key 3006P-502-ND).
U1—LM317T adjustable voltage regulator (Radio Shack 276-1778).

ing through R3 and D1 is approximately 0.8 amp, resulting in a voltage drop across R3 of about 176 mV. The inverting terminal of comparator U3 is now negative with respect to the noninverting terminal by 76 mV. As a consequence, the comparator output switches to *high*. C1 and C2 prevent a racing condition that might otherwise cause the comparator to change state before the charging current is established. At the outset, the output of U1 is less than 8.5 volts because R3 is used in the voltage-determining circuit in a negative feedback (current limiting) mode. U1's output voltage rises as the battery voltage increases, until it reaches about 8.4 volts. The charging current, and the voltage across R3, remain nearly constant at less than their initial values for most of the charge cycle. When the batteries approach their full-charge condition, their voltage rises. This decreases the charging current through R3, which results in decreased voltage across R3. The voltages appearing at pins 8 and 9 of U3 become equal when the voltage drop across R3 is reduced to the amount of the offset (100 mV). This equals a charging current of 455 mA, shared between two batteries, or about 227 mA for each battery.

When the voltage across R3 becomes *less* than 100 mV, comparator in U3 changes state and pin 14 goes low. This draws current through R4 and lights DS2. R4 is in the voltage-determining circuit of U1. The additional current drawn through R4 by U3 reduces the voltage at pin 3 of U1, dropping the output voltage lower than the battery voltage. D1 prevents current from the batteries from flowing backward in the circuit, so there is essentially no current through R3.

With no current flowing through R3, pin 9 of U3 is lower than pin 8 by 100 mV. The comparator output remains low and no additional charging takes place. The lighting of DS2 signals that the charge cycle is complete. At that point, the charging power source is disconnected, S1 is switched to the **USE** position and the batteries are available to power whatever device is connected to P2.

Battery-Condition Indicator Circuit Description

In Fig 1, U2, an LM7808CK 8-volt regulator, provides a stable reference voltage when S2 is closed. The string of equal-value resistors (R7 through R10) functions as a four-way voltage divider. Since the resistor values are equal, the resulting voltage drops across each resistor are equal. Even so, the voltage drops can be increased or decreased (as a group) by adjusting R20. A simple computation and a voltage measurement at TP1 determines the R20 adjustment—as we'll see later.

The reference voltages are applied to the noninverting terminals of the four comparators of U4, another LM339. The inverting terminals are all connected to a common voltage which is referenced to the battery voltage. (The ratio is adjusted by R19.) Four green LEDs (DS3 through DS6) are connected to the comparators.

When S2 is closed (placing power on the indicator), the battery voltage is applied to pin 1 of U2 as well as the LEDs. If the referenced battery voltage is greater than a given comparator's reference voltage, that comparator's output is low and its LED glows. If the referenced battery voltage is less than a given comparator's reference voltage, that comparator's output goes high and its LED does not illuminate.

If the reference voltage is correct and R19 and R20 are properly adjusted, all four LEDs glow when the battery is fully charged. The first LED (DS3) switches off when 75% of the charge remains. The second LED is extinguished when a 50% charge remains. The third LED winks out at 25% and the fourth switches off when the low-voltage condition is reached (when the battery should be taken out of service and recharged). As a safeguard against the problems of negative indications, a fifth LED (DS7) was added (with Q1 and U4D acting as a switch) to provide a constant **LOW BATTERY** indication. This red LED lights when the fourth green LED is extinguished.

Output Regulator Circuit Description

Many H-Ts are designed to be operated on 12 volts and thus don't require this regulator circuit. For this reason, the regulator is not included on the same board with the charger/indicator. (The regulator schematic is shown in Fig 2.) If your H-T requires less than 12 volts, the regulator can be set to provide the required voltage. U1, an LM317T adjustable regulator, is at the heart of this simple circuit. R2, a 5-kΩ potentiometer, adjusts the output of U1 to suit your H-T.

You'll need to provide a connector to fit your H-T. I made an adapter for my Kenwood TR-2500 using the case of a defunct battery pack. This is a good way of making the power connection, since it is reliable and attractive. In addition, the empty case provides space for the voltage regulator.

Construction

The charger is built on a 2¼- × 3½-inch circuit board. The prototype was constructed on perf board, but I highly recommend that you use a printed-circuit board. You can make your own or order the PC board from FAR Circuits.[1]

The circuitry is housed in a 4 × 2 × 2¼-inch aluminum box. The circuit board is supported inside the box on three angle brackets made of stiff steel wire bent to shape. The circuit board is mounted flush with the end of the box that is farthest from the indicator. This leaves space between the box and the board edge at the indicator for wires to pass to and from the switches, LEDs and grommets. The board edge on which U1 is mounted must also be flush with the side of the box so that U1 can use the box as a heat sink. The LEDs are cemented into their 1/16-inch holes using epoxy adhesive. U1 is mounted near the board edge abutting the side of the box and is bolted to the box using an insulating kit.

Four connectors are used—one for each battery, one for the charging source, and one for the device to be powered by the batteries. Select connectors that fit your requirements.

I recommend you make a power cable

---

[1] A PC board and part overlay are available from FAR Circuits, 18N640 Field Court, Dundee, IL 60118; price $4.50 plus $1.50 shipping and handling per order. Check or money order only; credit cards not accepted. The PC-board template and part overlay are available free of charge from the ARRL Technical Department Secretary. With your request for the SMITHEY LONG-HAUL H-T BATTERY SYSTEM PC BOARD TEMPLATE PACKAGE, send a #10 SASE.

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Calibrating the Charger

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Calibrating the Battery-Condition Indicator

I have determined that no single calibration of the indicator unit is truly accurate with several different battery types. Fig 3 shows the discharge characteristics of four different batteries, all with the same 100-ohm load, all having just been charged using the charger. The variations are great enough to significantly affect the accuracy of the indicator.

If you buy new batteries of the type shown in the parts list, you can be confident using the discharge characteristics of battery #1 in Fig 3 to calibrate your indicator. If you have elected to use batteries that have seen previous service, I recommend that you run a simple discharge test on the batteries before performing final calibration of the indicator. (Charge the battery using the calibrated charger, then attach a 100-ohm load and plot the discharge characteristic as was done for Fig 3.)

If you have more than one set of batteries with different discharge characteristics, I recommend that you calibrate for the best set, and take into account the difference when you read the charge remaining in the weaker set(s). That way, the charge remaining at any time will be nearly equal for the different batteries.

Given a small, constant load—such as an H-T in the receive mode—the discharge voltage curve over time is nearly linear until the battery voltage drops to about 12 volts. A battery should be taken out of service and recharged when its voltage under load drops to 12 or less. This rule seems to apply regardless of the battery brand.

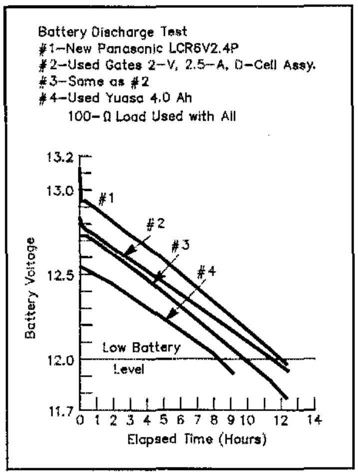

Fig 3—Battery discharge test results.

Fully charged battery voltage can vary considerably (see Fig 3). Depending on the brand, battery potential at the beginning of the discharge cycle can vary from 13 to 12.55 volts. Fortunately, the provisions for adjustment of the indicator are flexible enough to accommodate any batteries you're likely to find.

Calibrate the indicator based on the assumptions we've just discussed. In other words, assume that its voltage-versus-time curve will be linear with a constant, small load, and that it will need recharging when the voltage decreases to 12 volts. Compute the voltage you want to see at TP1 as follows:

$$V_{TP1} = V_{Ref} \times \frac{V_{Low}}{V_{Full}}$$

V_{Ref} is the voltage output of U2 at pin 3, in this case 8.0 volts. V_{Low} is the voltage selected for **LOW BATTERY** indication (12 volts). V_{Full} is the battery voltage at full charge.

To calibrate the indicator, you'll need a variable-voltage power supply with a range of 11.5 to 13 volts. After determining the voltage that you want at TP1, apply 12 volts to pin 1 of U2 and adjust R20 until the voltage at TP1 reaches the desired level.

Now reduce the voltage to the V_{Low} value and adjust R19 until DS6 goes out and DS7 comes on. That's all there is to it! DS3 should now go out at a battery voltage of 12.75, DS4 at 12.50, DS5 at 12.25 and DS6 at 12.00. When DS6 turns off, DS7 lights to tell you your battery needs to be charged.

While you have your variable-voltage supply connected, check that all four green LEDs do in fact go on and off at the correct voltages. If not, you may want to try calibrating using the transition of DS5 and a calibrating potential of 12.25 volts. Then, recheck for accuracy on all four LED set points. You should be able to get them all transitioning within 50 mV of the stated voltages.

For maximum accuracy, V_{Ref} should be measured with the same voltmeter you used to set the voltage at TP1. To calibrate the indicator for another battery with different characteristics, merely substitute the appropriate numbers in the equation above. To further improve accuracy, I measured all the 1-kΩ resistors used in the project and selected the four that were most nearly equal for R7 through R10.

Additional Thoughts

I use a common fanny pack to house the batteries and the charger/indicator when I want to carry the system around. In the one I bought, there's plenty of room for the equipment and accessory cables. Although I elected to discontinue using large batteries, I was able to get a pair of 4.0-Ah batteries in the pack and the weight wasn't too objectionable. The box housing the charger/indicator gets quite warm while batteries are being charged, so it should not be left in the pack when charging is in progress.

Summary

I've had a great deal of enjoyment in developing this low-tech project, and even more enjoyment out of using the long-haul battery system with my own H-T. Try one yourself and I'm sure you'll like it as much as I do.

Thurman Smithey, N6QX, was introduced to Amateur Radio in the late 1930s as a high school student. It wasn't until his retirement from the Navy, 30 years later, that he finally obtained his license. He was first licensed to the General class as WA6FUY in 1968. Thurman acquired a sailboat the same year and has enjoyed operating maritime mobile while doing some blue water sailing. Thurman holds a Master of Science degree in Engineering Electronics.

Strays

HAM AWARENESS GALA IN WESTERN THEME PARK

◊ It's fun for the whole family at Scottsdale, Arizona's, Family Amateur Radio Event (FARE), Sunday, September 19, at North Scottsdale's Rawhide Western Village.

Held in conjunction with the ARRL's Ham Radio Awareness Day, 13 Arizona radio clubs are participating in the get-together, which will feature an HF special-event station signing K7UGA (held by former US Senator Barry Goldwater) from 1700 to 2400Z.

For more information, contact Len Winkler, KB7LPW, at 602-861-0303.

A Simple Broadband Dipole for 80 Meters

Turn your existing 80-meter dipole into a broadband antenna by simply modifying the feed line. Multiband operation is an option.

By Frank Witt, AI1H
20 Chatham Rd
Andover, MA 01810

A conventional coax-fed, half-wave dipole doesn't provide a low SWR over the entire 80-meter band—an inconvenience for those of us who like to operate phone and CW on that band. Several approaches to overcoming this limitation, short of an antenna tuner in the station, have been described.[1,2] The antenna system described here is simpler than any of its predecessors and has the following features:

- A 2:1 SWR or better is achieved over all or most of the 80-meter band.
- Antenna length and appearance are the same as those of a conventional half-wave dipole. Consequently, it's lightweight and has small wind and ice loading.
- The antenna configuration permits multiband operation with a single feed line.
- The losses due to broadband matching are acceptable.
- The cost is about the same as a conventional half-wave dipole.

All the SWR data given in this article were measured at the transmitter end of the feed line. The reference impedance is 50 Ω, since most equipment is designed for this impedance. The term *antenna system* as used throughout this article includes not only the radiating wire, but also the feed line, balun (if used), any lightning-protection measures, antenna tuner and so forth.

The dipole antenna itself is not broadband; the system uses a broadband *match*. The key broadbanding element of this antenna system is the *transmission-line resonator*: Part of the transmission line compensates for the reactance presented by the dipole away from its resonant frequency. This part of the line is a multiple of an electrical half wavelength. Another part of the line presents an appropriate source impedance to the transmission-line resonator (TLR).

First I'll describe a version of the broadband antenna system, along with some practical results. Then I'll cover the important matter of antenna-system loss. Following that are some variations to suit specific requirements, and a method for using the antenna for several bands. I'll also compare transmission-line-resonator broadbanding to other broadbanding methods.

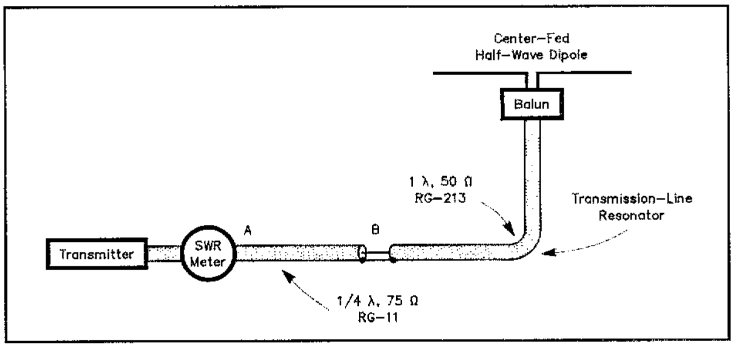

Fig 1—One form of the simple broadband antenna system. It resembles a conventional dipole except for the ¼-wavelength, 75-Ω segment. Points A and B are discussed in the text.

Table 1
Calculated and Actual Lengths of the Broadband Dipole Antenna at AI1H

	Calculated	*Actual*
¼-λ Coax	43.3 feet	43.3 feet
1-λ Coax	173.1 feet	170.5 feet*
Dipole	124.5 feet	122.7 feet

*Includes 11 inches for balun.

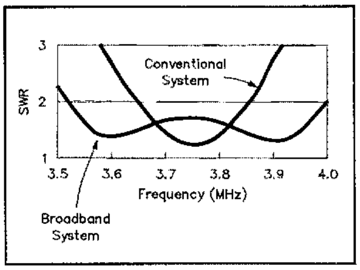

Fig 2—Measured SWR versus frequency for the broadband and conventional antenna systems.

The 80-Meter Broadband Antenna System

Fig 1 shows the simple broadband antenna system as used at my station. The antenna proper is a center-fed half-wavelength dipole. The transmission line is segmented into one electrical wavelength of 50-Ω coax and an electrical quarter wavelength of 75-Ω coax. The calculated and actual lengths are shown in Table 1. Lengths were calculated using the formulas given later in this article, using a center frequency (F_0) of 3.75 MHz and VF (velocity factor) of 0.66. The actual lengths resulted after I performed the tuning procedure described later. Manufacturing variations from the published cable velocity factors, and some stretching of the coax, contributed to the differences between actual and measured values. (The actual lengths were measured on untensioned cable.) The antenna is installed as an inverted V with a 140° included angle and an apex height of 60 feet. The wire size is #14, but is not critical.

This system's SWR (at the transmitter) as a function of frequency is shown in Fig 2. For comparison, the SWR for the same dipole fed with about ⁵⁄₄ wavelengths (214

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[1] Notes appear on page 30.

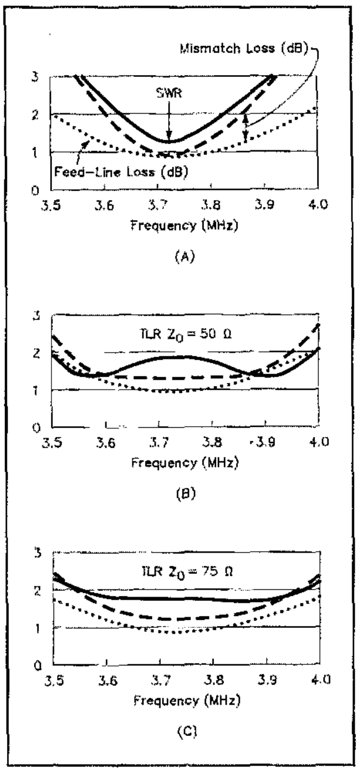

Fig 3—Antenna-system configurations for long feed-line runs. The solid lines are SWR; the dotted lines are feed-line loss; and the dashed lines are feed-line loss plus mismatch loss. At A, a conventional system using an RG-213 feed line; at B, the feed line is a ¼-λ section of RG-11 (75 Ω) followed by 1 λ of RG-213; and at C, a ¼-λ segment of RG-11 is followed by 1 λ of RG-11 (one 5/4-λ piece of RG-11). The total feed-line length in each case is 216.4 feet.

feet) of RG-213 coax is also shown. (This is the same total length as the RG-213 and RG-11 segments used in the broadband system.) The broadband system's 2:1 SWR bandwidth is 2.2 times that of the conventional system—and the *only* difference is the feed-line configuration!

The radiating properties of the broadband antenna over the 80-meter band are essentially identical to those of a dipole cut for any specific frequency in the band. Also, since the antenna system is designed for a 50-Ω transmitter, the feed-line length may be extended by adding the required length of 50-Ω coax between the transmitter and the quarter-wave segment (point A in Fig 1).

A 1:1 current balun should be installed at the antenna's feed point. I use the balun on general principles. Often, it provides no visible difference in operation, but the balun does minimize feed-line radiation. You can determine whether your antenna needs a balun by measuring the SWR versus frequency with and without a balun installed. If the balun is not needed, the two sets of data will be identical.

Antenna-System Losses

It's important to know the losses in any antenna system. This is especially true for broadband antennas, because loss alone can broadband an antenna system. As the next section shows, the configurations presented in this article do not yield a significant loss penalty. Although other loss contributors exist in antenna systems, we will focus on the primary ones: *feed-line loss* and *mismatch loss*. Other losses, such as ohmic loss in the antenna wire, are the same for both the conventional and broadband systems described here.

Feed-line loss is the easiest to understand. It is unavoidable, and is lowest when the feed line is *flat* (when the line SWR is close to 1:1). At HF, feed-line loss results primarily from ohmic losses in the copper conductors.

Mismatch loss occurs when the impedance seen by the transmitter is not the complex conjugate of the transmitter's impedance (when the line SWR at the transmitter is not 1:1). For a 50-Ω transmitter, the mismatch loss is 0 dB when the load impedance is 50 Ω. When the load impedance is not 50 Ω, the mismatch loss can be made to be 0 dB if a transmitter with a tunable output stage (such as a conventional tube-type linear amplifier) is tuned for a conjugate match. An antenna tuner can also provide this match. In this case, however, the antenna-tuner loss (perhaps as much as 1 dB) replaces the mismatch loss in the total-loss equation. That subject isn't discussed here.

If you don't use an antenna tuner and the transmitter has a fixed-tuned 50-Ω output, loads that present the transmitter with an SWR under 2:1 are highly desirable. The impact of high SWR on mismatch loss will become clear in the next section.

Loss must be kept in perspective. All of the broadband antenna systems described here have a worst-case total loss of less than 3 dB—not enough to notice in many 80-meter QSOs. (If the loss is 3 dB, half of the transmitter's output power is radiated and half is lost elsewhere.) The main effect of loss is stress on system components: that on the transmitter due to the mismatched load, and that on the transmission line due to heating.

Variations

The broadband antenna system described above is well-suited for the installation at my station, where the distance between the shack and the antenna is relatively long (more than 200 feet) and because I use a 1-kW amplifier. Other feed-line combinations are better suited to other installations. Some of these are shown in Figs 3 through 5, along with calculated SWR and loss data. From this information, you can select an appropriate feed-line combination for your needs.

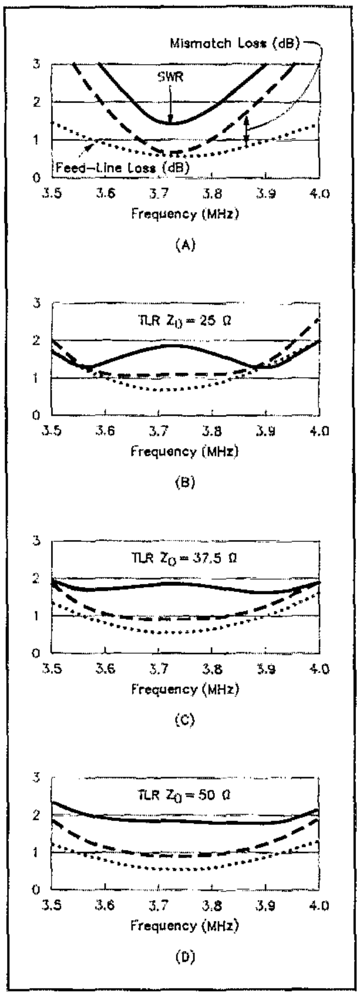

Fig 4—Antenna systems for high power and shorter feed-line runs. The solid lines show SWR; the dotted lines represent feed-line loss; and the dashed lines show feed-line loss plus mismatch loss. At A, the feed line is RG-213; at B, it's a ¼-λ section of RG-11 followed by two paralleled ½-λ lengths of RG-213; at C, a ¼-λ segment of RG-11 is followed by two paralleled ½-λ lengths of RG-11; and at D, ¼ λ of RG-11 is followed by ½ λ of RG-213. The total feed-line length in each case is 129.8 feet.

The figures also show the characteristics of conventional dipole antenna systems. If you compare them, you'll see that the transmission-line resonator provides broadbanding without a significant loss penalty. I haven't tried all these combinations, but based on my experience, they should perform as predicted in most situations if the radiator doesn't deviate significantly from the model I used in my calculations: a dipole 125 feet long, 40 feet high, and made of #14 wire. This model is based on data provided by Walt Maxwell, W2DU, in his book, *Reflections*.[3] I chose his data since it is typical of many 80-meter installations.

All of the broadband antenna systems use a ¼-wave section and either a ½- or 1-wavelength section. Fig 3 illustrates a system for long feed-line runs. It uses RG-11 and RG-213 cable and should be considered for all power levels. Fig 3B covers the case shown in Fig 1 and used at my station. The feed line of Fig 3C is a continuous length of RG-11 cable 5/4 wavelengths long. The transmission-line resonator is the 1-wavelength section of the cable nearest the antenna.

This approach would also work with surplus 75-Ω CATV Hardline. A ¾, 5/4 or 7/4-λ section of ½-inch Hardline yields less than 2 dB feed-line loss plus mismatch loss over the entire band, and less than 1 dB total loss over any 300 kHz of the band. This configuration is particularly attractive to contesters and DXers, because even a fairly long line—7/4 λ is 372 feet of ½-inch CATV Hardline—gives low loss and a very good match over, say, the 3.5- to 3.8-MHz range.

Three broadband antenna systems are shown in Fig 4. All of these are candidates for applications requiring shorter feed-line lengths. Figs 4B and 4C show the performance realized when coax cables are paralleled to achieve a low equivalent characteristic impedance. Fig 3B, which results from a 1-wavelength RG-213 transmission-line resonator, and Fig 4B, are very similar. The latter system uses the same amount of cable, but it's cut in half and parallel-connected. This will become clear in the sidebar, "How It Works." The configuration in Fig 4D is attractive because of its simplicity.

Lower-power applications without long feed-line runs can use RG-58 and RG-59 coax. Fig 5B shows how excellent broadbanding is achieved with a remarkably simple feed line. Again, no loss penalty results from the broadbanding.

Adjusting the Broadband Antenna System

The antenna system is easy to build and adjust. First calculate the lengths (in feet) of the transmission-line segments:

$$L_{quarter} = \frac{245.9\ VF}{F_0} \qquad \text{(Eq 1)}$$

$$L_{half} = \frac{491.8\ VF}{F_0} \qquad \text{(Eq 2)}$$

$$L_{full} = \frac{983.6\ VF}{F_0} \qquad \text{(Eq 3)}$$

where

$L_{quarter}$ = length of quarter-wave segment
L_{half} = length of half-wave segment
L_{full} = length of full-wave segment
VF = velocity factor
F_0 = center frequency in MHz

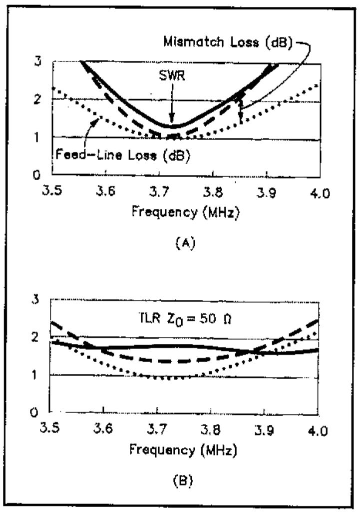

Fig 5—Antenna systems for low power and shorter feed-line runs. The solid lines show SWR; the dotted lines represent feed-line loss; and the dashed lines show feed-line loss plus mismatch loss. At A, the feed line is RG-58; at B, it's a ¼-λ section of RG-59 followed by ½ λ of RG-58. The total feed-line length in each case is 129.8 feet.

A good starting point for the dipole wire length (in feet) is:

$$L_{dipole} = \frac{467}{F_0} \qquad \text{(Eq 4)}$$

For the 80-meter application, I suggest using an F_0 of 3.75 MHz. It's a good idea to cut the wires so that the overall length is 4 feet longer than necessary, in case you need to lengthen the wire during tuning. Pass 2 feet of the extra wire through each end insulator and wrap it back around the antenna wire.

To tune the antenna system, you'll change only the dipole and transmission-line-resonator lengths. The best approach is to build the antenna system as I have outlined here and to measure the SWR at the transmitter end of the system. Any tilt or frequency offset in the SWR characteristic can be removed by increasing or decreasing the dipole or transmission-line-resonator length. Start by changing the length of the dipole. To improve the SWR at the high end of the band, the dipole must be shortened; to improve the SWR at the low end of the band, the dipole must be lengthened. Progressively add or subtract 6 inches from both legs of the dipole until the SWR curve is symmetrical about the center frequency.

Frequency offset may be required to center the SWR characteristic in the 80-meter band. You can move the entire curve along the frequency axis without causing asymmetry by changing both the dipole and transmission-line resonator lengths using the following equation:

$$L_{New} = L_{Old}\left(\frac{3750 - \Delta F}{3750}\right) \qquad \text{(Eq 5)}$$

ΔF is the required frequency offset in kilohertz. Shortening the dipole and resonator moves the curve center up in frequency, and lengthening them moves the center down. The length of the quarter-wave segment need not be changed, since the SWR characteristic is not very sensitive to its length.

Lightning Protection

Every antenna system should be designed to minimize the likelihood of a lightning strike. One part of this is keeping all parts of the antenna proper at ground potential. The grounding should be done *outside the shack*, by means of a good ground rod.

I recommend that you install a coaxial lightning protector, which bleeds any static charge from the center conductor, at point B of Fig 1. The protector (and therefore the feed-line shield) should be connected to a high-quality ground rod (the kind electricians use) driven 8 feet into the ground.

Conversion of Existing 80-Meter Dipoles

A study of the cases shown in Figs 3B, 4D and 5B suggests that it's possible to easily convert many existing 80-meter half-wave dipole antennas. Because the most popular way to feed an 80-meter dipole is with a 50-Ω coaxial feed line, the conversion to a broadband antenna system is straightforward. First trim the dipole for resonance at about 3.75 MHz. Then cut the 50-Ω feed line at a multiple of an electrical half-wavelength (at 3.75 MHz) from the antenna. Calculate this length using Eq 2 or Eq 3. Add the 75-Ω quarter-wave section, then complete the run to the shack (if necessary) with 50-Ω coax. Then use the tuning procedure described earlier to optimize the system.

Multiband Operation

Most broadband 80-meter antenna systems are usable only on the 80-meter band, because the broadbanding elements do not allow efficient power transfer on other bands. This is not true with the approach described here, since the structure consists only of a center-fed dipole and a transmission line. Moreover, the transmission-line segments are close to multiples of an electrical half-wavelength near 40 meters and other bands. This opens the possibility for paralleling other half-wave dipoles with the 80-meter dipole and sharing the feed line.

To minimize their interaction, the various dipoles should be spaced from each other away from the feed point. Of course, some interaction will occur and you must tune the multiband system to meet your requirements. I recommend first tuning the 80-meter

How It Works

A fundamental way of achieving a broadband match to a resonant dipole antenna involves a parallel-tuned LC network and an appropriate source resistance. In an *RF Design* article,* I described the method for designing such networks, even with lossy resonators. The top of Fig A shows the equivalent circuits of the antenna and matching network. The bottom of Fig A illustrates the corresponding elements in the antenna system.

The role of the resonator is played by the transmission-line segment nearest the antenna. It must be a multiple of an electrical half-wavelength. The quarter-wavelength "Q"-section, made from 75-Ω coax, transforms the 50-Ω transmitter resistance to 112.5 Ω $(75^2/50 = 112.5)$. I won't go into the design details here; they're the subject of another article, "Broadband Matching Using the Transmission-Line Resonator," in preparation for *The ARRL Antenna Compendium, Volume 4.*

For the structure of Fig A to yield a broadband match, the characteristic impedance of the transmission-line resonator and the transmitter resistance must be within a range of values. Fortunately, commonly used transmission lines, which are available in 50- and 75-Ω characteristic impedances, work well in this application. The broadband systems of Figs 3 through 5 show the usefulness of this approach.

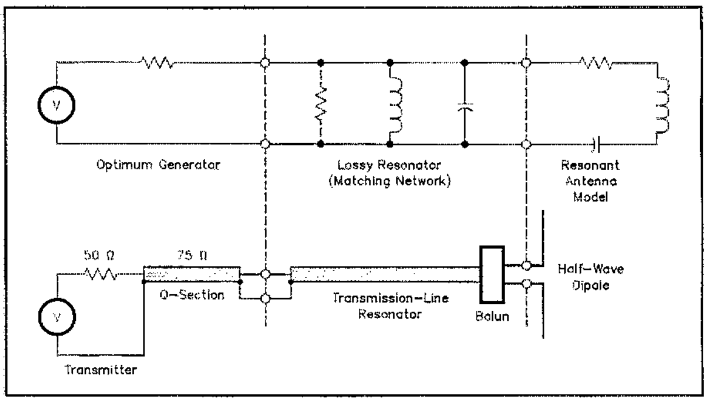

Fig A—Lossy broadband-matching-network equivalent circuit (top), and corresponding simple broadband antenna system elements (bottom).

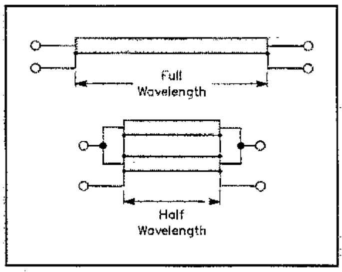

Fig B—These two transmission-line resonators behave essentially the same in this application. The characteristic impedance of each cable segment is the same, making the characteristic impedance of the lower resonator half that of the upper one.

Fig B makes another significant point. For this application, the network parameters of a one-wavelength transmission-line resonator (top) are similar to those of a half-wavelength resonator (bottom) with half the characteristic impedance of the upper resonator. Parallel-connecting two identical cables is a convenient way of achieving lower characteristic impedances. This explains the similarity of Figs 3B and 4B and the similarity of Figs 3C and 4C.—*AI1H*

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*F. Witt, "Optimum Lossy Broadband Matching Networks for Resonant Antennas," *RF Design*, Apr 1990, pp 44-51, and Jul 1990, p 10.

broadband system and then the next-highest-frequency dipole, and so forth. Only the 80-meter antenna will be broadband, but such broadbanding is not required on the other bands. Fig 6 shows the result of adding a 40-meter dipole to the Fig 1 antenna. Each dipole leg is 34.4 feet long. Note that the SWR on 80 meters changes very little compared to Fig 2. No change was made to the 80-meter dipole or the transmission line.

The multiple-dipole approach described above achieves resonance on several bands and eliminates the need for an antenna tuner on those bands. Of course, if you use an antenna tuner, operation on all HF bands should be possible, but this arrangement is usually not as effective as the multiple-resonance antenna system described here because the feed-line loss is much higher.

Comparison with the Coaxial-Resonator Match

How does the simple broadband dipole described here stack up against other approaches for achieving a good match over the entire 80-meter band? The coaxial-resonator match broadband dipole[4,5] represents one of the more efficient designs published to date. It achieves broadband matching at the antenna by the integration of ¼ wavelength of coaxial cable as a part of the antenna.

Since the coaxial-resonator match achieves a good match at the antenna, the SWR on the feed line is low and the feed-line loss is about the same as its matched loss. However, the coaxial cable in the match itself increases the system loss. The net result is that the total loss is about the same with the coaxial-resonator match, but the SWR at the transmitter is lower, never exceeding about 1.6:1 between 3.5 and 4 MHz. Once the SWR is less than 2:1, however, a lower SWR has little value unless you're using a transmitter that significantly reduces power at such SWRs.

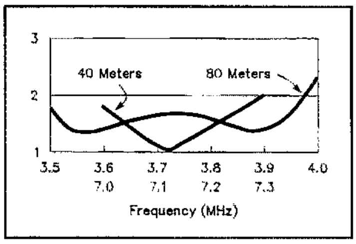


Fig 6—Measured SWR for the 80- and 40-meter multiband antenna system.

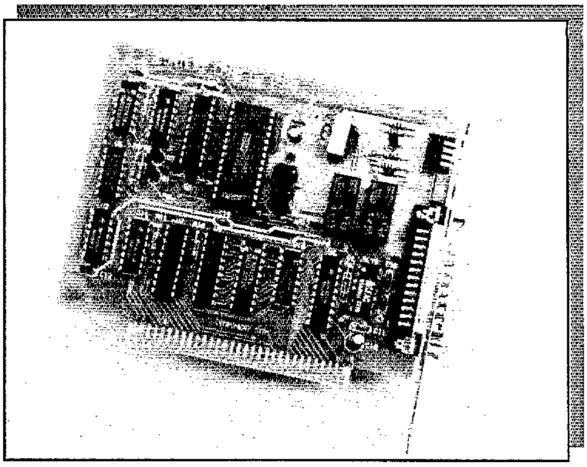
Note that the approach described in this article uses a thin wire for the antenna. Most other broadbanding approaches use additional wires or radiators made partly from coaxial cable and are vulnerable to damage from wind and ice loading. Their additional weight and complexity are also limitations.

From the above comparison, the simple broadband antenna system has, by its very simplicity, an edge over the coaxial-resonator match, at least in applications where the simpler approach is feasible. Because of the limitations of available coaxial cables, the opportunity for a satisfactory design is constrained. On the other hand, the coaxial-resonator match has more adjustment para-

(continued on page 76)

The Contest Card

This 5- × 4½-inch plug-in turns your PC into a high-powered contest machine that does voice recording and playback, *and* CW keying!



By Gary C. Sutcliffe, W9XT
3310 Bonnie Ln
Slinger, WI 53086

A major trend in recent years is for ham-shack computers to control more and more station equipment. For contest operators, logging programs started by sending CW, then controlling digital voice keyers, and later controlling transceivers, all via the PC's serial and parallel ports.

So far, this trend has also meant increased clutter on the operating table—not to mention higher cost, since each accessory requires its own interface, cabinet and power supply. Now there's a better way: Put *all* of the accessories right *inside* the computer.

This article describes a PC-compatible, plug-in card that contains a voice recorder and CW interface. Dubbed the Contest Card, it's software-compatible with the most popular contest logging programs, and can be built for less than most standalone voice recorders cost.

During the past couple of years, several articles about voice keyers based on the Information Storage Devices ISD-1000-series chips have appeared in Amateur Radio and hobby-electronics magazines.[1,2,3] These ICs incorporate many of the functions required to implement high-quality voice storage and playback at low cost.

I felt that these designs lacked several important features needed by serious contesters. This started me thinking about designing my own voice keyer based on the ISD-1016A. The Contest Card is the result.

Design Considerations

After hearing some voice keyers on the air, I decided that it was very important for such devices to accurately reproduce the operator's voice. It's confusing to answer a CQ and hear a different voice coming back to you! The designs I've seen so far use inexpensive electret microphones for recording. To minimize the difference between live and recorded audio, the Contest Card uses the mike normally used with the transmitter.

The design must also be compatible with K8CC's *NA* and K1EA's *CT* contest programs. These programs use the same technique to key external voice keyers via one of the computer's LPT (parallel) ports.

It soon became obvious that the whole unit could mount inside the computer. The card's hardware would "look like" an LPT port to the software. Since *NA* and *CT* can also key a transmitter on CW with a suitable interface, I also included this feature in the design. Fig 1 shows the Contest Card in block-diagram form.

Putting the entire unit on a PC plug-in card offers several advantages. The first is price. A plug-in computer card is more expensive than a box-mounted board because of the card's complex shape and gold-plated contact fingers. This higher cost, however, is more than offset by the savings gained from the lack of packaging and power-supply requirements. Since the card is controlled by software, fewer buttons and switches are needed, further reducing cost. Putting the card inside the computer also frees up desk space and eliminates the need to share an LPT port between a printer and voice keyer—important for PCs with only one parallel port.

Putting the circuit inside the computer challenged me to develop an efficient way to record messages. You don't want to have to exit your contest program to record a new message! I included two ways around this, as I'll describe later.

The ISD-1016A Integrated Circuit

The heart of the card is the ISD-1016A. Other articles[4] have published the details of this chip, so I'll hit only the high points here.

The ISD-1016A can store 16 seconds of audio. It samples the audio signal at its microphone input at an 8-kHz rate. The chip's EEPROM stores the sampled values in analog form. The main advantage of this technique is that the message is not lost when power is removed, as it is in traditional RAM-based voice recorders.

The Contest Card supports up to four messages. These correspond to the messages started by pressing the **F1**, **F2**, **F3** and **F4** keys on the keyboard while running *NA* or *CT*. Each key is assigned a specific type of message by the software, such as CQ, contest report, QRZ mes-

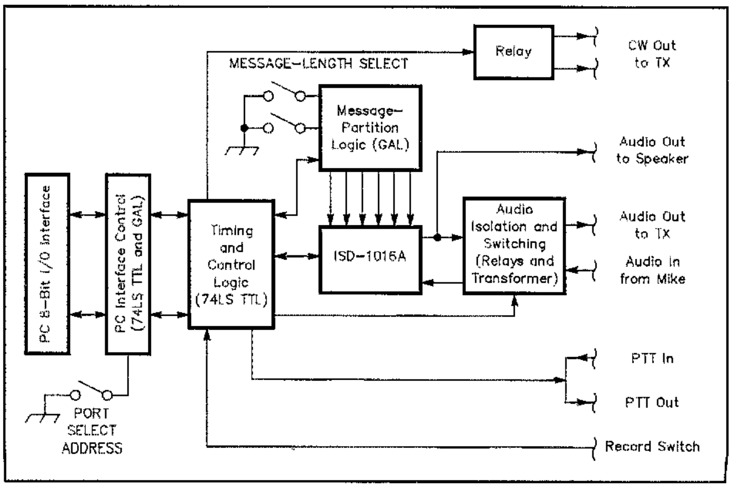


Fig 1—Block diagram of the Contest Card.

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[1] Notes appear on page 32.

sage, and your call sign. The chip's 16 seconds of recording time must be divided between the messages. You select how this time is divided by setting two switches.

During playback, a message continues until it reaches the end of the message. You can interrupt playback in the middle of a message by pressing **ESC** in *CT* or *NA*.

Recording resembles playback, except that it requires a record-control signal that can be generated by two sources: an external switch, or software control. The transmitter's PTT line is not activated during recording.

Audio to and from the ISD-1016A is switched by a pair of DPDT relays. When the Contest Card is not recording or playing back a message, the microphone audio input and microphone ground are switched directly to the transmitter. This provides complete isolation from the computer, which is required to prevent hum on live audio. You can leave your microphone connected through the Contest Card even when the computer is powered off.

The card's audio output is isolated by a transformer. Without this isolation, voltage differences between the computer and the transmitter cause current to flow on the ground connection. This current generates hum on the transmit audio.

On CW, the card uses a relay to key the transmitter. I chose relay keying to allow operation with positive- and negative-keyed rigs with a single circuit.

Construction

Detailed construction information, including schematics for the Contest Card and interface cable, are available from the ARRL.[5] The card uses entirely off-the-shelf parts, with the exception of two generic-array logic (GAL) ICs that manage some of the ISD-1016A's functions. As a result, you can build a Contest Card yourself, from scratch, if you have GAL-programming capabilities. If you don't, you can obtain programmed GALs from me. A PC board, kits of parts and the completed Contest Card are available through Unified Microsystems.[6]

In order for the Contest Card circuit board to be offered for sale, I had to obtain FCC Class B certification. The testing procedure revealed that unacceptably high levels of RF emissions were being picked up and radiated through the cable. To eliminate this problem, a special DB25 connector is needed. Each pin of this connector contains a 470-pF bypass capacitor to ground. Such connectors are very expensive in small quantities, so they're included in the kits and assembled version.

If you decide to wire-wrap or design your own Contest Card PC board, you should either use the specified connector or install 470-pF bypass capacitors on every signal line leaving the connector. Place the capacitors as close to the connector as possible, and ground the connector shell. Under FCC regulations, homebuilt electronics for personal use do not require certification, but must use "good engineering practice." Aside from the FCC regulations, you don't want computer-generated noise covering up weak DX stations answering your CQs!

To use the Contest Card, you'll need to construct an interface cable. All connections to the card are made through a single DB25 connector. The circuit uses only 14 of this connector's pins, but a 25-pin connector provides a bit of extra space to ease construction. Use only shielded cable for all connections. The PTT ground and microphone ground *must* be separate shields or severe hum is likely to appear on the transmitted audio. Small coaxial cable, such as RG-174, works well for this application.

Table 1
Contest Card Message Partitioning

Switch Selection	*Message Length (Seconds)* #1	#2	#3	#4
1	6.4	2.4	4.0	3.2
2	5.6	7.2	3.2	0.0
3	8.0	0.0	4.8	3.2
4	4.0	4.0	4.0	4.0

Some transmitters (notably ICOMs) place +8 V on their microphone-audio lines. If you have such a rig, put a dc-blocking capacitor in series with the audio line to the transmitter. A 0.22-µF or larger cap should work fine, and can be mounted in the cable's DB25 shell. Interface cables for most popular transceivers are available from LTA Industries.[7]

Using the Contest Card

You'll need to decide how to partition the card's memory space. This is handled by two switches on the card, accessible through the card's mounting bracket. Table 1 shows how the messages can be divided. The simplest way to divide the time between four messages is to give each one 4 seconds. This is fine for some uses, but poor for others. Your contest exchange for the ARRL Sweepstakes, for example, may be too long to fit comfortably in 4 seconds. You can set the switches so that one of the messages is zero seconds long, giving the extra time to the remaining messages.

You can record messages in two ways. Perhaps the easiest method is to use your logging program and an external switch (normally open, momentary) connected to pins 8 and 9 of the DB25 connector. To record with the external switch, bring up *CT* or *NA*. In the set-up screen, be sure to set the Contest Card to use the correct parallel port. In *CT* and *NA*, function keys **F1-F4** correspond to messages 1 through 4, respectively. Pressing a function key starts playing back the appropriate message. If the record switch is held closed before the function key is pressed, the corresponding message can be recorded. Recording continues until the record button is released. I chose this method of recording messages so that it would not be necessary to quit the logging program to record new messages. (*NA* Version 8 also supports recording without the switch, and *CT* may soon do so as well.)

The other way to store messages is to use a program that supports software-controlled recording. Such a program can be easily written in most common computer languages. The card's documentation (and the information package available from ARRL[8]) explains the requirements. A standalone program for Contest Card control, DVRC.EXE, is also available from me and the ARRL BBS.[9] It controls record and playback of voice messages on the Contest Card. It also controls playback on other voice keyers that follow the same interface standards.

It's important that you not record messages longer than the length of time specified for each message (see Table 1). A message that's too long will overrun part of the next message's space, recording over anything already stored there. Also, if you change the switch settings, you will usually need to rerecord each message.

Last, you'll need to adjust the card's audio-output level. First, record a long test message. Then adjust your transmitter's microphone gain to the appropriate level while speaking into the microphone. Next, play back the test message and adjust the potentiometer accessible through the card-edge bracket for the proper level for your transmitter. The ISD-1016A has an excellent AGC circuit, so it should not be necessary to adjust this pot again unless you switch to a significantly different microphone.

Summary

The Contest Card offers a low-cost solution for integrating your computer, transceiver, digital voice recorder and CW-keying interface. It does this while maintaining software compatibility with the standards set by *NA* and *CT*. Other applications for the Contest Card abound. For example, a PC-based repeater controller could use the Contest Card for IDs and special messages. VHF operators can also use it for their transmit sequences during meteor-scatter schedules.

Thanks to Bruce Herrick, WW1M, for laying out the Contest Card circuit board. Additional thanks to Paul Hellenberg, KS9K, and Scott Ellington, K9MA, for their help and suggestions.

Notes

[1] B. Brown, "Voice ID on a Chip," *73 Amateur Radio Today*, Nov 1991, pp 11, 12, 61.

[2] J. Jarrett, "Chip Talker," *QST*, Dec 1991, pp 17-22.

[3] R. Tenney, "Single-Chip Messaging System," *Radio-Electronics*, Mar 1992, pp 59-64, 92.

[4] See Note 2.

[5] A package of information including GAL program listings, detailed construction information, circuit theory and schematics for the Contest Card is available from the ARRL Technical Department Secretary. Request the *September 1993* QST *Sutcliffe Information Package*. Include a large SASE with your request. Source-code listings for the GAL equations and the DVRC.EXE program are available from the ARRL telephone BBS (203-666-0578) as W9XT.ZIP.

[6] A partial kit consisting of a blank, double-sided PC board with plated-through holes, solder mask and silk-screened part-placement labels, as well as the preprogrammed GAL chips, RF-bypassed DB25 connector, and the DVRC.EXE control program on a 5¼-inch MS-DOS disk is $58.95. A kit of the harder-to-find parts (PC board, GALs, ISD-1016A, isolation transformer, prepunched mounting bracket, relays, RF-bypassed DB25 connector) and the DVRC.EXE program is $94.95. A complete kit of all parts (except interface cabling) is $119.95. The assembled and tested Contest Card, without interface cable, is $179.95. Check or money order only to Unified Microsystems, PO Box 133, Slinger, WI 53086, tel 414-644-9036. Please add $5 shipping per order to US and Canada, and $20 for overseas orders. Wisconsin residents must add 5% sales tax.

[7] These cables cost $40 each, plus shipping and handling. Contact LTA Industries, PO Box 77, New Bedford, PA 16140, tel 216-565-9950, for more information.

[8] See Note 5.

[9] See Note 5.

The ⅓-Wavelength Multiband Dipole

There's plenty of good antenna installation weather left this season. This compact, high-performance multiband antenna beckons. Try it!

By Andrew S. Griffith, W4ULD
203 Lord Granville Dr, Rte 2
Morehead City, NC 28557

Are you looking for a relatively simple, inexpensive antenna for the HF bands? Are you considering an antenna fed with open-wire (ladder) line, but do you dread the idea of bringing such line through the wall into the shack? Well, relax while I describe a simple antenna system that works well on several ham bands, and is fed with coaxial cable. When properly installed, the antenna system shows an SWR of less than 2.3:1 at its feed point near the middle of the 7, 14, 18, 21, 24 and 28-MHz bands.

The SWR at 3.5 MHz is approximately 5:1. Thus, you may need an antenna tuner (built into your transceiver, or external) on 80 meters. You may also need a tuner on other bands when using a solid-state transceiver because such rigs often reduce their output power when operating into even a moderate SWR, such as 2:1. If you use a tuner, the rig will see a proper load and operate at its full power level. Except perhaps on 80 meters, tube-type transceivers and amplifiers should work into the antenna without a tuner. Since the SWR in the coaxial portion of the feed line is low on most bands, losses are low and efficiency is high.

Fig 1 shows the antenna. Its only unusual aspect is the 375-Ω balanced line between the center of the dipole and the balun. This line is constructed by reforming commercial 450-Ω ladder line for reduced conductor spacing. Reforming the line is a simple procedure that takes about 30 minutes, as I'll describe later.

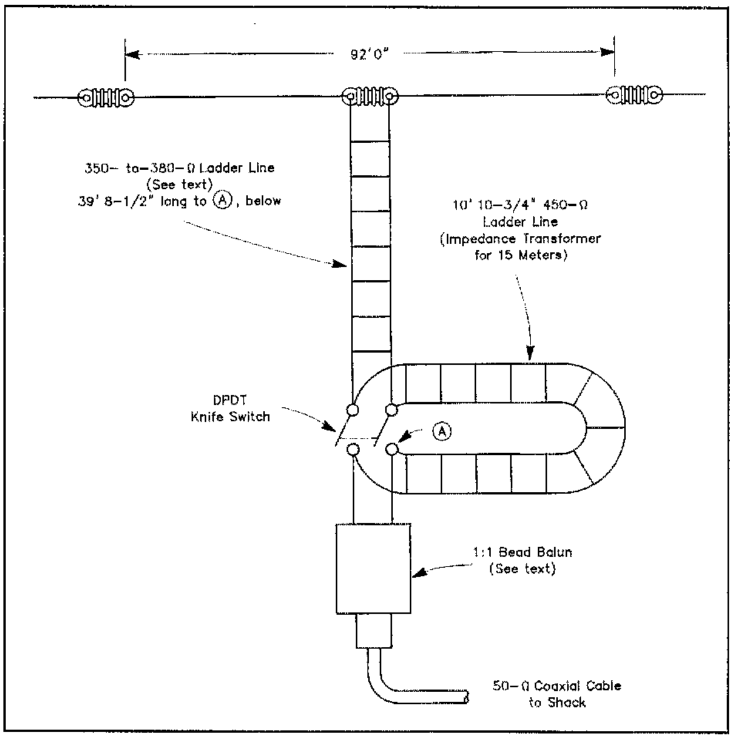

Fig 1—The ⅓-wavelength multiband dipole, with 15-meter impedance transformer and switch. This simple, efficient antenna system has SWR characteristics compatible with most transceivers.

The 1/3-Wavelength Principle

The ⅓-λ dipole was described by Taft Nicholson, W5ANB, in November 1981 *QST*.[1] Nicholson showed mathematically that a ⅓-λ-long dipole, fed with ⅙ λ of balanced line of the proper impedance, exhibits a low impedance at the input (the transmitter end) of the balanced line at the fundamental frequency, and at the second, fourth, fifth, seventh, eighth and some higher harmonics of the fundamental frequency.

Note that the third and sixth harmonics are missing. This means that such an antenna system, cut for 3.5 MHz, will work at approximately 3.5, 7, 14, 17.5, 24.5 and 28 MHz. At 10.1 and 21 MHz, the input impedance will be very high. However, the input impedance on these bands can be transformed to a low value by switching in an extra ¼ λ of feed line at the coax-to-ladder-line transition. Fig 1 shows such a scheme for 15 meters.

Computer Study

To achieve a reasonable SWR on several bands, two factors necessitate a compromise in the lengths of the dipole and balanced line. First, a fixed-length antenna system doesn't resonate exactly at the harmonics of the fundamental frequency.[2] Second, the HF ham-band centers are not exact multiples of any single frequency. So, I performed a computer study to determine the best compromise among dipole length, balanced-line length, and balanced-line impedance. A ⅓-λ dipole at 3.55 MHz is 87.8 feet long, assuming the usual 0.95 K factor for thin wire antennas.[3] One-sixth wavelength of ladder line is 46.2 feet, assuming, for the time being, a velocity factor of 1. I used the *ELNEC*[4] antenna-analysis computer program to determine the impedance at the center of dipoles from 85 to 95 feet in length. For the models, I used an antenna height of 35 feet and #14 wire.

The study showed that the best compromise is a 92-foot dipole, fed with 45 feet of balanced line having a characteristic impedance of 375 Ω and a velocity factor of 1. (The actual line used has a velocity factor of 0.886,

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[1] Notes appear on page 35.

Table 1
***ELNEC*-Calculated Data**

Frequency (MHz)	*Dipole Feed-Point Impedance R ±jX ohms*	*Ladder-Line Input (Balun Output) Impedance R ±jX ohms*	*50-Ω SWR on Coaxial Feed Line Measured at Balun*	*Ladder-Line Loss (dB)*	*Total Loss, Ladder Line and Coax (dB)*
3.55	24 − *j*580*	9 +*j*8	5.7	1.1	1.8
7.18	329 + *j*819	52 + *j*40	2.1	0.3	0.7
10.13	4071 − *j*216	56 − *j*0**	1.1	0.7	1.1
14.25	96 − *j*517	33 − *j*0	1.3	0.6	1.2
18.11	340 + *j*765	59 + *j*34	1.9	0.4	1.7
21.30	2396 − *j*2320	53 − *j*0**	1.1	0.9	1.6
24.93	133 − *j*447	59 − *j*7	1.2	0.5	1.3
28.5	345 + *j*647	75 − *j*39	2.1	0.4	2.2

**ELNEC* says the resistive portion of the impedance is 11 Ω. I raised this to 24 Ω because of the low antenna height. See text and Note 5.

**These are transformed impedances with proper lengths of 450-Ω ladder line added at the input to the 375-Ω line. See text.

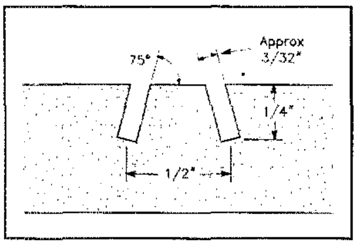


Fig 2—Slot dimensions for the reforming jig, cross-sectional view through board. The 3/32-inch width of the saw blade is perfect for making the ladder-line slots.

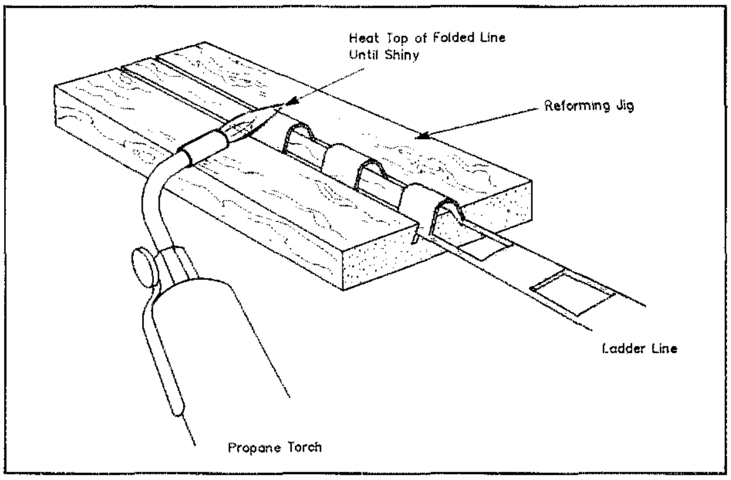


Fig 3—The reforming jig in use. Push and pull the ladder line through the jig as the reforming proceeds. An assistant is helpful for this operation.

so the physical line length must be 45 feet × 0.886, or 39.87 feet.) The lengths are somewhat critical, but balanced-line impedances between 350 and 380 ohms are acceptable. Standard 300- and 450-Ω lines are not acceptable. A summary of the *ELNEC*-calculated data is given in Table 1. Since *ELNEC* predicts erroneously low feed-point resistance values when a horizontal antenna is modeled at heights below approximately 0.2 λ,[5] the resistive part of the antenna impedance at 3.55 MHz was estimated from actual SWR measurements.

Construction

The antenna is easy to build. The only unusual aspect of its construction is reforming the 450-Ω ladder line, which you can buy from antenna-material suppliers. The center-to-center spacing of the line as supplied is approximately 25/32 inch. To make 350- to 380-Ω line, the spacing must be reduced to about ½ inch.[6]

To make a reforming jig for commercial ladder line, use a 2-foot length of wood. Saw two slots lengthwise in the board according to the dimensions in Fig 2. For this task, you'll need access to a table saw or radial saw.

Clamp the reforming jig to a table in a well-ventilated area. Start the reforming by folding toward each other the two wires on one end of the line. Insert the end of the folded line about four inches into the slots at the right end of the jig as shown in Fig 3. Using a propane torch, heat the top of the folded polyethylene until it develops a shiny, almost liquid, appearance. Hold the flame about two inches from the polyethylene and nearly horizontal to the board to minimize charring the line and the board. Push another four inches of line into the slots. Heat the new four-inch section as just described. Continue this process until the reformed line reaches the left end of the jig. By that time, the first four-inch section of the line should be cool. Additional four-inch sections of line can be pulled into the slots from the left while pushing from the right. It's helpful to have an extra pair of hands on the left side to pull the line and coil it into a three-foot-diameter coil. This process may sound tedious, but it's really quite simple. It took me about five minutes to make the reforming jig, and 20 minutes to reform the line without assistance. Fig 4 shows what the reformed line should look like when you're done.

The impedance of my line measures approximately 350 Ω. The velocity factor measures 0.886, which is the same as the stock 450-Ω line. The reformed line is not as flexible as the original, but can be bent easily to a 6-inch radius. Otherwise, the line should be handled like any other ladder line. Make sure that it drops straight down from the connection to the dipole for the greatest possible distance. Keep the line at least 2 inches away from other structures, especially metal ones. These precautions preserve feed-line balance and minimize line radiation. Solder the ladder line at the dipole feed point.

The impedance transformer for the 15-meter band is a loop of 450-Ω ladder line (not reformed). It is switched in and out (shorted) with a small DPDT knife switch such as the Radio Shack 275-1537. I put my switch in a small birdhouse for protection from the weather, but this is not necessary. You can use a remote-controlled relay instead of the manual knife switch; I leave such details to your ingenuity. The form of the loop is not critical, but it cannot be tightly coiled. The dimension given is for an open loop; when I formed the loop into a three-turn helix, the resonant frequency shifted upward from 21.2 to 21.3 MHz.

Instead of the 15-meter transformer, you can use one cut for 10.1 MHz. The transformer should be approximately 25 feet, 3 inches in length. Of course, the antenna can be used without the transformer if an antenna

Table 2

Effect of Coax Feed-Line Length on SWR at Transmitter

Length of RG-8X (feet)	*Frequency (MHz)* 3.55	7.18	14.20	18.11	24.93	28.50
42	4.4	1.3	1.6	2.3	1.3	1.3
52	4.6	1.5	1.2	1.6	1.1	1.5
54	4.6	1.6	1.1	1.7	1.2	1.3
57	4.8	1.6	1.1	1.8	1.3	1.2
64	4.5	1.6	1.5	2.1	1.0	1.5
69	4.3	1.4	1.6	1.7	1.3	1.2
79	3.7	1.2	1.0	1.9	1.0	1.2

Note: These measurements were made without the 15-meter impedance transformer installed.

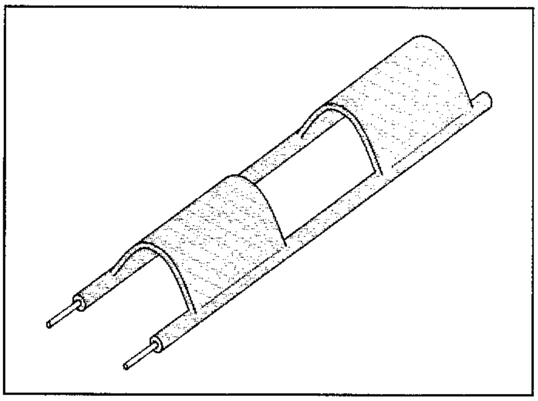


Fig 4—The appearance of the ladder line after reforming.

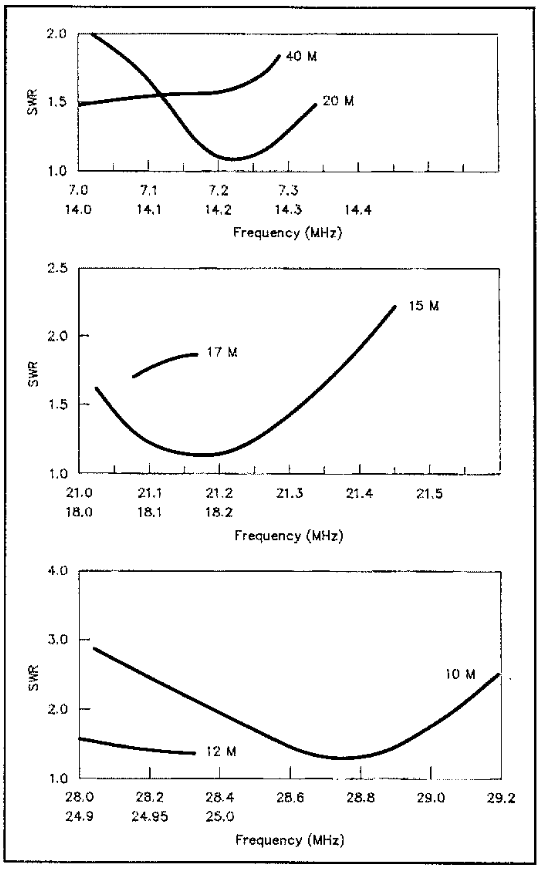


Fig 5—SWR curves for the antenna as built. You can achieve a low SWR on 30 meters by adding an appropriate impedance transformer. See text.

tuner is used. However, the SWR on the coaxial feed line will be quite high, and antenna system efficiency will be low.

The balun consists of 10 ferrite beads (Amidon Associates FB-77-5621) slid over a 13-inch length of RG-8X coax. Solder the coax to the end of the ladder line, and seal it with silicone or windshield sealant. Install a PL-259 connector on the transmitter side of the balun. After installation, seal the balun and the coax connector. I used vinyl electrical tape with a thin coat of sealant. A commercial 1:1 current balun should be acceptable in place of the home-brew unit just described, but I have not tested one.

Evaluation and Operation

After installation, I pruned the ladder line to a length of 39 feet, 8½ inches for the best SWR compromise across the bands. I also pruned the 15-meter impedance transformer to 10 feet, 10¾ inches for best 15-meter SWR. Some pruning of these two lengths may be necessary, so be prepared to experiment a little.

Fig 5 shows the SWR curves for the various bands. I obtained these data with a Bird 43 wattmeter, and with the 15-meter impedance transformer and knife switch installed. I used 57 feet of RG-8X from the balun to the transmitter. Apparently, there was some coupling between the antenna and the coax shield, since varying the length of coax between the balun and the transmitter affected the measured SWR. Table 2 shows this phenomenon. To reduce the variation in SWR with feed-line length, I tried a 1:1 voltage balun, and 10 additional beads on the current balun. These efforts were not particularly successful. In any case, the SWR at the coax input remained reasonable regardless of coax length.

The antenna performs as I expected it would. At 3.5 MHz, signals are not quite as good as with a ½-λ dipole because of the relatively high line losses. At 7 MHz and above, the antenna exhibits some gain over a ½-λ dipole because the antenna is more than ½ λ long. *ELNEC* plots of the radiation pattern show maximum radiation broadside to the antenna at 3.5, 7 and 14 MHz. At 18 MHz and above, the patterns break into four major lobes: 40° from broadside at 18.11 MHz, 35° at 21.3 MHz, 27° at 24.93 MHz, and 53° at 28.5 MHz. The software predicts approximately 3 dB of gain over a ½-λ dipole at 18 MHz and above.

In summary, the antenna works quite well. At 7 MHz and above, it's more efficient than my G5RV. Try it; you'll like it!

Notes

[1] T. Nicholson, "A Compact Multiband Antenna Without Traps," *QST*, Nov 1981, p 26.

[2] J. Hall, ed, *The ARRL Antenna Book*, 16th ed, (Newington: ARRL, 1991), p 2-8.

[3] *The ARRL Antenna Book*, p 2-3.

[4] *ELNEC* antenna analysis software, R. Lewallen, PO Box 6658, Beaverton, OR 97007.

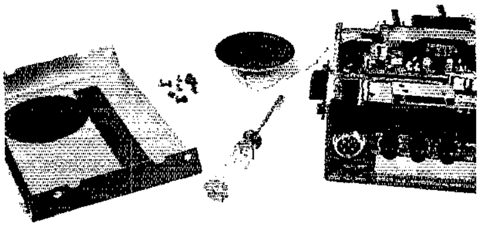
[5] R. Lewallen, "*MININEC:* The Other Edge of the Sword," *QST*, Feb 1991, pp 18-22. See also Feedback, *QST*, May 1991, p 46.

[6] Although I feel that reforming commercially available ladder line is easier, you can build 360-Ω feed line from scratch. Use #12 wire with plastic insulators every 6 inches or so along the line. The conductor spacing should be 0.81 inch. If you choose this alternative, make the line length 42 feet, 9 inches.

Radio and electronics have been Andrew Griffith's hobby since he was 15. He was first licensed in 1951, received his Advanced class license in 1952, and graduated to Extra Class in 1983. In 1943, he earned a BS degree in Chemical Engineering from Virginia Polytechnic Institute. After a tour of duty with the Army in Europe, he received his MS degree in 1947. He spent his industrial career with the Du Pont Company.

Now retired, Andrew devotes his free time to ham radio, golf, and staying out of his wife's hair. His primary radio interests are its technical and building aspects. He still uses a homemade linear amplifier, but most of his home-brewed equipment has been relegated to the attic. He occasionally chases DX and is active in the Volunteer Examiner program.

Under the Hood



Why *not* pop the top off that multihundred-dollar radio and see the work of industrial art you paid for? Follow along as one of *QST*'s resident radio geeks reveals the first steps in moving from "Huh?" to "Gee...."

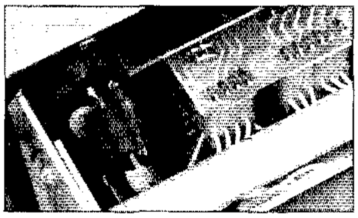
By David Newkirk, WJ1Z
Senior Assistant Technical Editor

I admit it: I'm one of those people who insists on keeping his plane-window shade open a few inches while everybody else in the cabin watches *Rocky III*. The way I see it, I don't pay an airfare price just to get from A to B while yawning through a movie I could just as well see on the ground. Included in the ticket cost, I figure, is several hours' worth of viewing the world from five miles up.

Likewise, I figure that buying one of today's radios includes ogling rights. You've bought a piece of modern industrial art—an *artifact* in the most positive sense—that looks as good as it works, and has a story to tell. All you need to start appreciating it is a few tools and your imagination.

Open Sesame

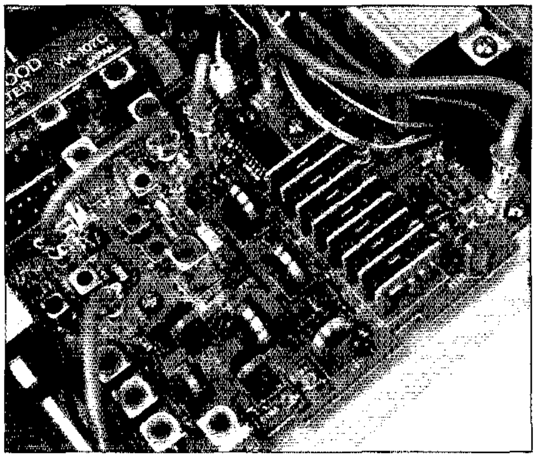
Look in your radio's instruction manual for details on how to open its cabinet. You'll probably find these instructions somewhere in the vicinity of text and drawings about setting internal switches, or installing optional filters or an FM or tone-encoder board. The tool of choice will probably be a Phillips screwdriver. Don't be bashful—study the manual and dig right in, putting the screws into a small box or cup as you go. Keep paper and pencil handy to take notes in case you run into something that might make reassembly tricky. A penlight and magnifying glass can help in scoping out really neat and/or hard-to-see stuff. One caution: In some radios, speakers mount on the top cabinet piece. The speaker wires may be *just* long enough to let you lay the cabinet top flat. Keep this in mind so you won't yank the wires off the speaker or out of their connector or circuit-board holes. If the speaker plugs into the radio circuitry, so much the better: Unplug it.



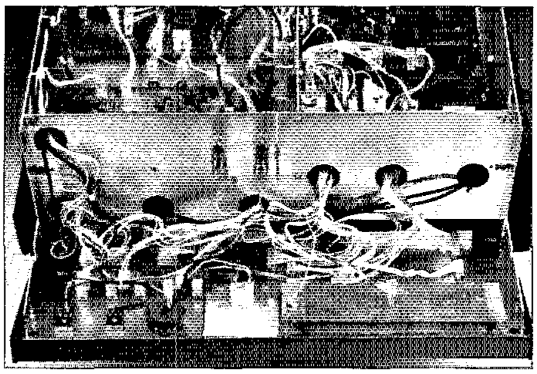
We're always telling you to knock down RFI by shoving audio, power and control lines through ferrite beads and toroidal cores. The pros do it, too—it's standard practice in making sure RF goes only where it's supposed to go in today's compact equipment designs. This is the mike-jack corner of a Ten-Tec OMNI VI. *(photos by Kirk Kleinschmidt, NT0Z)*

Be a Tourist

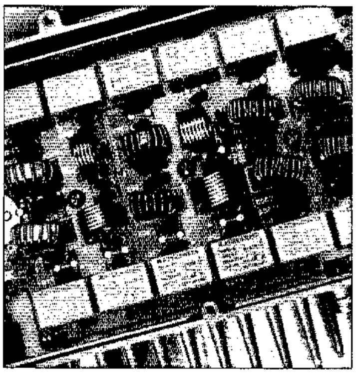
You need no goal more honorable than tourism to snoop into your radio, but practical goals lend focus. Maybe you're a QRPer and want to locate power-output control VR57 to set your radio's minimum output to 5 watts. Maybe you've bought an aftermarket CW filter and intend to install it. Or maybe you're like me: You can't stand those darned keypad beeps, and the manual tells you that flipping switch SW4 will make them go away. Goals like these force you to inspect your radio and manual *together* to find landmarks that lead to your target. In working this process, you take an important step in transforming your radio



Each card in that little sheaf of daughterboards contains one of the Kenwood TS-50's eight receiver input filters. Surface-mount components keep each filter small—about 1 × 1.5 centimeters!



Remove two screws and loosen two others, and the Ten-Tec OMNI VI's front panel swings down for inspection. What, no tuning capacitor? Nope! As in all of today's fully synthesized radios, a *shaft encoder*, often no larger than an audio gain control, translates tuning-knob rotation into pulses that command the radio's computerized tuning system where you want it to go.



Miniature relays (the light-grey *MATSUSHITA* blocks) select the TS-50's transmit filters (rows of capacitors and coils) band by band. This circuitry helps the TS-50 meet FCC's rules for transmitted-signal purity.

from an inscrutable black box into something you can know and understand: *You identify and locate a particular part of the radio that performs a known function.* If you can really understand the function of just *one* component in a radio, you can work your way up to understanding what *any* of its components do.

See How Form Follows Function

Much of your radio's status as an object of industrial art comes from the fact that no aspect of its design is accidental. Every detail is the result of care—arguably, in many instances, love—and decision. That those decisions had to be made with cost and profit in mind make the result all the more impressive and deserving of respect. The key to enjoying your radio as a work of industrial art is to *how* its designers shaped it to do its intended job. So let's quickly overview main transceiver functions and how they commonly translate into real-world "modularity."

Radio energy enters and exits your transceiver through at least one *antenna jack*—an *SO-239* "UHF" connector on MF/HF and some VHF radios, a *BNC* connector on others. Somewhere quite close to this, you may be able to find a *TR* (transmit-receive) *relay* that switches the antenna between the transceiver's transmitter and receiver circuitry. (You'll know you're in the transmitter zone when you find yourself regarding the radio's massive and finned *heat sink*, which may also be accompanied by a compact cooling fan.)

Receiver signal flow will likely proceed through a module named something like *RF Board*. This is the home of the receiver's *front end*—the first-line circuitry that performs the stressful, all-important job of turning radio signals from the wide world into something your radio can digest, control and ultimately provide to your ears or modem in ready-to-use form.

Further along in the receive chain, you'll probably find an *IF* (intermediate frequency) or *Main board* containing two or more *crystal filters*—oblong metallic modules, often accompanied by spaces for you to add more, that your radio uses to separate closely spaced signals from one another. Somewhere in this vicinity, your radio likely generates an *AGC* (automatic gain control) signal. This is important: AGC keeps signal-strength variation from blowing you out of your chair by turning up your radio's gain for weak signals and reducing its gain for strong ones. A front-panel **AGC** control may allow you to tailor your radio's AGC *speed*—how fast it readjusts its gain in response to signal variations.

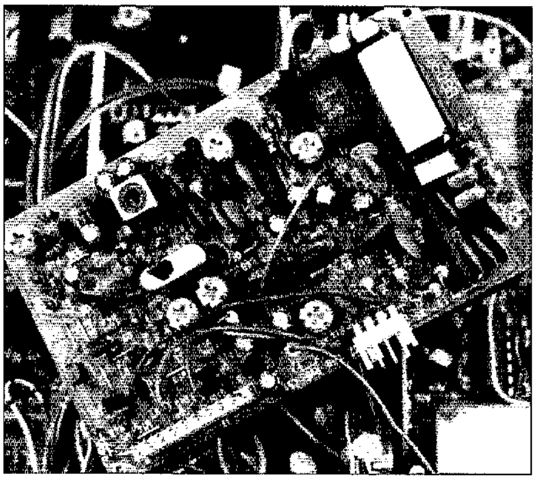
Pretty soon after IF filtering, a radio receiver ultimately transforms its IF signals into audio-frequency (AF) energy (you may see an *AF* or *Detector Board*) for application to your speaker, headphones or modem. This audio usually appears at a **PHONES** jack (front panel), **EXTERNAL SPEAKER** jack (rear panel) and one or more **ACCESSORY** jacks (rear panel). Panel-mounted subassemblies often group these jacks with others—jacks that let you control external amplifiers ("linears"), modems and so on.

Hands Off the Handhelds

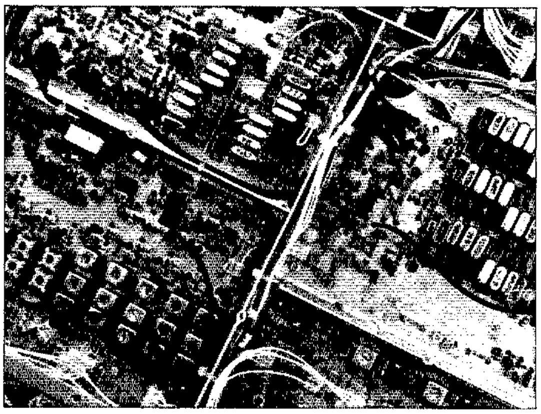
This article talks about looking inside radios *intended to be opened by you.* That generally includes MF/HF transceivers and VHF/UHF radios designed for mounting under your car's dashboard. It *doesn't* include handheld VHF/UHF transceivers and similar microgear. *Don't* open these marvels of miniaturization unless their instructions tell you exactly how to. Speaking from experience, I can tell you that they usually come apart a lot easier than they go back together!
—WJ1Z

Radio front panels can be particularly marvelous. In some radios, one large circuit board backs the front panel and carries its display (liquid crystal, light-emitting diode [LED] or fluorescent) and supporting logic, *and* all of its knobs, jacks, switches and buttons.

As you investigate your radio's transmit chain, look for a board or boards with names like *TX*, *Generator*, or *Driver*. (Most radios' *Final Amplifier* boards are hidden within metallic shields on or inside the radio's heat sink.) These modules generate your code, voice or digital-data signal at a very low level, amplify it, and ultimately cast it into the world at a globe-spanning power level commonly at tens or hundreds of watts. A *reflected power* sensor monitors your antenna system performance



Pull out the ICOM IC-737's optional FM board and look closely. How many of its components can you identify? Don't forget to take in the surrounding scenery when you reinstall it. Our radios don't just make QSOs: they're works of industrial and technological art.



A technician's-eye underchassis view of the Ten-Tec OMNI VI (outer and inner covers removed). This circuitry—much of it receiver IF amplifier and filtering—plays a large part in how a radio feels and sounds as you listen to signals on the air. Some Very Serious Hams might get huffy if I said that this looks like a city—so I won't.

and tells the transmitter to throttle back if too much transmitted energy reflects back into the final amplifier. Many radios start turning their output power down at reflected powers that equate to *standing-wave ratios* (SWRs) of 2:1 or so.

A transceiver is called a transceiver because much of its circuitry does double duty—in *trans*mit and re*ceive*. Modern radios use computers—*microprocessors*—to oversee and regulate the switching and command sequencing necessary to make the radio's modules work together. Your radio's computer likely lives on a board named *Logic* or *Control*, perhaps associated with (or including) a function called *Display*. Your radio's LCD or LED or fluorescent display—and all of its various frequency, band, mode and status indicators—operates under the direct control of the microprocessor. (So does your radio's *frequency synthesizer*, which may reside on a board or boards labeled *Synthesizer*, *DDS* [direct digital synthesis] or *VCO* [voltage-controlled oscillator].) The wires connecting a radio's computer to its front-panel display can radiate radio interference, so designers usually keep their radios' control and display modules close together to minimize wire lengths. Some designs put their control microprocessor on the front-panel board!

Returning to Base

It's time to end your tour when you find yourself feeling bored ("Okay, nice, but I'd rather be scanning 40") or nervous ("Maybe I should put this back together now and make sure it still works"). Don't ignore either feeling. Just retrace your disassembly steps, taking care not to pinch wires or leave out any screws. And don't forget to plug that speaker back in!

You Survived

So—your radio's back together, it still works, you survived the experience, and nobody (except you) is the wiser. Now you've taken your first look under the hood and added value to your equipment dollar. Whether you built it or bought it, you own a darned amazing radio. It's art you can *use*.

Boards, Boards, Boards

Most of our radios' action happens on various printed circuit (*PC*) boards. These phenolic or glass-epoxy sheets carry wiring in the form of etched copper *vias* or *traces*. Single-layer boards are everywhere; multilayer boards—boards with sandwiched wiring layers and interlayer connections made with *plated-through* holes—are increasingly common in high-component-density designs. The board surface is often *screened* or *masked* with transparent green paint and labeled with *component designators* (R13, T1, Q56—see the table) that tag parts with their schematic designations.

Many components are *through-hole* mounted—their wire leads pass through PC-board holes and are soldered to PC-board traces. But more and more of our radios' parts are *surface-mount devices* (SMDs) that have very short leads or no leads at all. Their solderable ends or tabs connect them directly to PC-board traces. As if SMDs don't exemplify miniaturization enough, PC boards sometimes carry yet other PC boards—daughterboards—as subassemblies. Even if your radio's motherboard doesn't contain SMDs, one of its daughterboards probably does.—*WJ1Z*

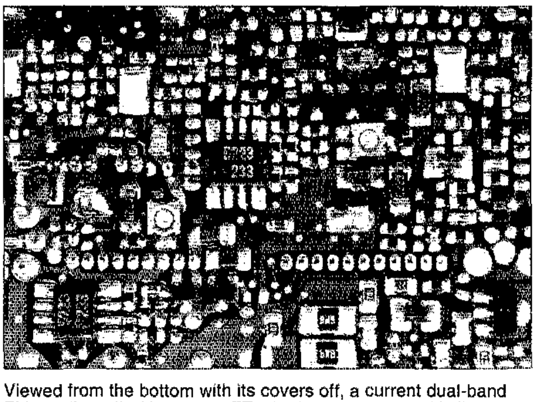
Common Component Designators

C—Capacitor
D—Diode
IC or U—Integrated circuit
J—Jack or wire jumper
K, RY or RLY—Relay
L—inductor or coil
P—Plug
Q—Transistor
R—Resistor
S or SW—Switch
T—Transformer
TC—Trimmer capacitor
VR—Variable resistor
W—Wire jumper

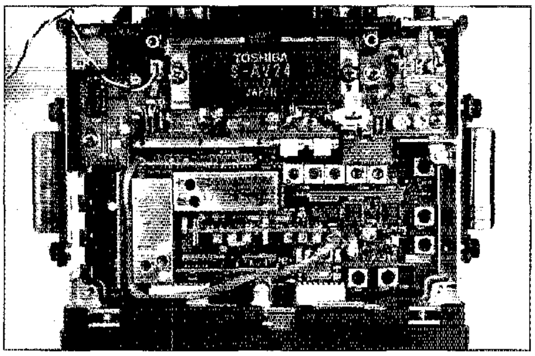
Homework

Aside from just snooping into your radio for the heck of it, you can use your radio as a springboard for learning more about the world. Try these teasers:

- Locate and identify your radio's *heat sink*. Its job is to conduct heat away from the radio's power-amplifier transistors (or else they'd die) and radiate it safely into the air. It's probably made of cast or extruded aluminum. How hot was that aluminum when molten? Extra credit: Even flawless castings often have to be machined for installation. Research the role played by *CNC* in machining.
- If your radio is at all modern, its front panel and/or display carries an abbreviation for the last name of a famous scientist. What is that abbreviation, what does it stand for, who is the scientist, and what did he or she achieve in science to merit mention on your radio? Extra credit: Your radio's *DIN* jacks take their name from an acronym of three words in that scientist's native language. What are those words?
- Spread out your radio's schematic and identify a *diode* (—▶|—). Pick a 4 × 4-inch section of the schematic that seems to contain quite a few diodes. Count them and multiply this number by 0.945. What you get is how much heater ("filament") power (in watts) you'd need to light their vacuum-tube equivalents. Extra credit: Assuming a diameter of 3/4 inch for a tube containing *two* diodes, how wide would your radio have to be just to contain, side by side, all the equivalent tubes for the diodes in that 4 × 4 schematic chunk?—*WJ1Z*



Viewed from the bottom with its covers off, a current dual-band FM mobile radio (the Yaesu FT-5100) presents a starkly linear mosaic of tiny surface-mount components. Those little white squares are *coils* wound on plastic bobbins!



Discrete transistors are *out* in many RF-power applications at VHF and UHF. Like many other radios on these bands, the FT-5100's transmitter ends up with a one-piece *hybrid power module*—the rectangular black block. Nonamateur two-way VHF/UHF radios use them, too.

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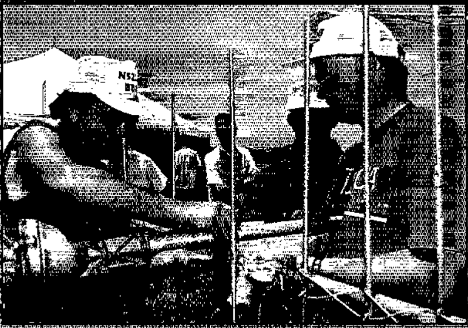

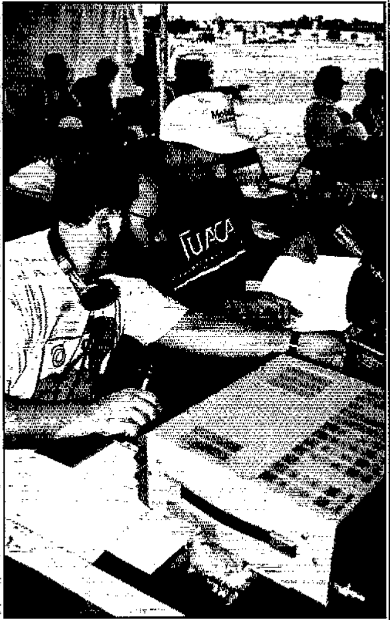

FIELD DAY '93

QST

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OUR COVER

Field Day '93 in Plano Texas: It was a fair-weather Field Day for the 75-or-so ops and visitors who participated at AA5NT, the 4A station fielded by the Plano Amateur Radio Klub. The photos were taken by Patrick Nugent, of nearby Allen, Texas. He's an electronic-imaging specialist for Kodak—and a brand new ham: KC5CMK. Congrats!

CONTENTS

November 1993
Volume LXXVII Number 11

TECHNICAL

NEWS AND FEATURES

NEW HAM COMPANION

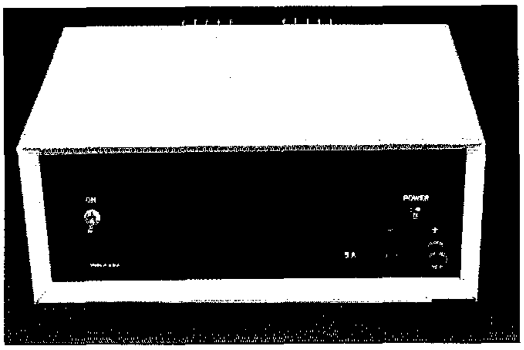
OPERATING

DEPARTMENTS

A 13.8-V, 5-A Regulated Power Supply

This power supply's features, keep itself—and the equipment attached to it—safe!

By Ben Spencer, G4YNM
100 Linsdale St
Swindon, SN2 2BN
England



(photos by Kirk Kleinschmidt, NT0Z)

Many commercially built 13.8-V power supplies are available. Some are robust and equipped with protection circuits; others have clearly been "built down" to meet a specific selling price. Most hams and electronics hobbyists want a well-designed power supply—but at the lowest (read "cheapest") price. Cheap power supplies, however, usually have no protection circuitry and exhibit poor regulation. Worse, regulator failure can damage the equipment to which the supply is connected. Think about it: With the gear you have attached to the supply, is it worth taking a risk like that?

This power supply provides 13.8 V dc at 5 A. It features time-dependent current limiting and short-circuit protection, thermal-overload protection within the safe operating area of the regulator IC, and overvoltage protection for the equipment it powers.

Construction, testing and calibration are straightforward, requiring no special skills or equipment. Many of the components can be found in junk boxes, or purchased at hamfests or from mail-order suppliers.[1]

Circuit Description

Fig 1 is the power-supply schematic. Incoming ac-line current is filtered by a chassis-mounted line filter (FL1) and, after passing through the fuse (F1), is routed via **ON/OFF** switch S1 to T1.

U1 rectifies, and C1 filters, the ac output of T1. U2 is an LM338K voltage regulator. This IC features a *continuous* output of 5 A, with a guaranteed peak output of 7 A, on-chip thermal and safe-operating-area protection for itself, and current limiting. U2's output voltage is set by two resistors (R2 and R3) and a trimmer potentiometer (R8), which allows for adjustment over a small range. U2's input and output are bypassed by C2, C3 and C4. D1 and D2 protect U2 against these capacitors discharging into it.

Overvoltage protection is provided by an SCR, Q1, across the regulator input. Normally, Q1 presents an open circuit, but under fault conditions, it's triggered and short-circuits the unregulated dc input to ground. This discharges C1 and blows F2, reducing the possibility of damage to any connected equipment.

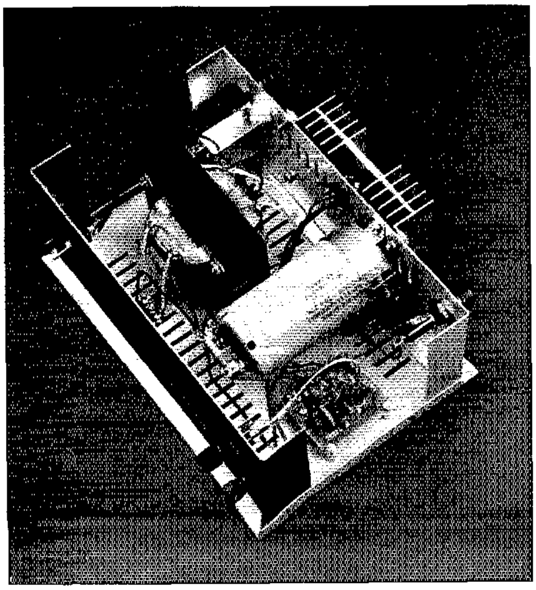
Overvoltage-protection IC U3 continuously monitors the output voltage. When the output voltage rises above a predetermined level, U3 starts charging C5. If the overvoltage duration is sufficiently long, U3 triggers Q1. This built-in delay (about 1 ms) allows short transient noise spikes on the output voltage to be safely ignored while still triggering the SCR if a true fault occurs. The monitored voltage is set by R5 and R6 and trimmer potentiometer R9, which allows for adjustment over a limited range.

D3 protects the supply from reverse-polarity discharge from connected equipment. The presence of output voltage is indicated by an LED, DS1. R7 is a current limiting-resistor for DS1.

Construction

How you construct your supply depends on the size of the components and enclosure you use. General physical layout is not important, although there are a couple of areas that require some attention. In the unit shown in the photographs, FL1, the fuse holders, S1, the heat sink, DS1 and the binding posts are mounted on the front and rear enclosure panels. T1 and the PC board[2] are secured to the enclosure's bottom plate. C1's mounting clamp is attached to the rear panel. Bleeder resistor R1 is connected directly across C1's terminals. D3 is soldered directly across the output binding posts.

U2, D1, D2, and R3 are all mounted on the heat sink as shown in Fig 2. (It's



Physical layout of the power supply. On the rear panel, left, are the ac-line filter and F1. The regulator's heat sink is at the middle of the panel and F2 is to the right. At the bottom of the enclosure, in front of T1, is the diode bridge rectifier. Because C1 is too tall to mount vertically within the Hammond #14260 cabinet, its mounting clamp is secured to the inside rear panel. Immediately to the right of C1 is the FAR Circuits PC board. On the front panel are the on/off switch, LED power-on indicator and output-voltage binding posts.

---

[1] Notes appear on page 29.

important to keep R3 attached as closely as possible to U2's terminals to prevent instability.) Use a TO-3 mounting kit and heat-conductive grease to electrically isolate U2 from the heat sink.

Cover all ac-input wiring (use insulated wire and heat-shrink tubing) to prevent electrical shock, and route the ac wiring away from the dc wiring.

Mount U2, C1 and the PC board close to each other; keep the wire runs between these components as short as possible. Excessively long wire runs may lead to unpredictable behavior.

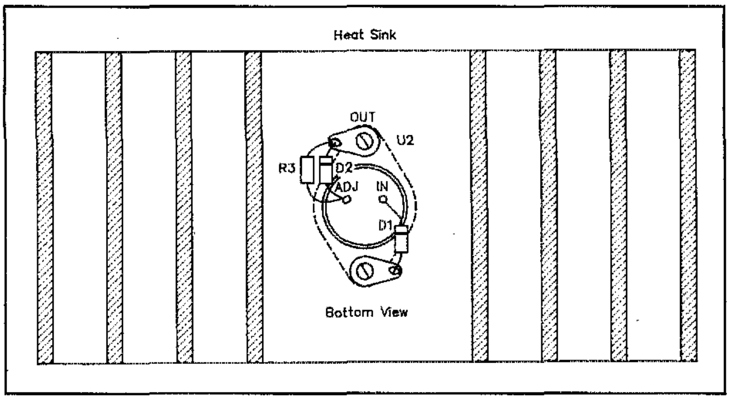


Fig 2—Here's the regulator's heat-sinking arrangement, showing U2, D1, D2 and R3. U2 is electrically isolated from the heat sink by a TO-3 mounting kit.

Test and Calibration

An accurate analog or digital multimeter, covering ranges of 30 V dc and 10 A dc, is required. A variable resistive load with a power rating of 100 W is also needed; this can be made using a heat-sink-mounted 2N3055 power transistor and a couple of components as shown in Fig 3.

First, set R8 (**OUTPUT VOLTAGE**) fully clockwise and R9 (**OVERVOLTAGE**) fully counterclockwise. Insert a fuse in the dc line at F2. Connect the ac line, turn on S1 and check that DS1 lights. Measure the output voltage; it should be about 12; adjust R8 counterclockwise until you obtain 14.2 V output; this sets the trip voltage.

While monitoring the output voltage, gradually adjust R9 clockwise until the voltage suddenly falls to zero.[3] This indicates that the SCR has triggered and blown F2. Disconnect the ac line cord from the wall socket. *Don't make any adjustment to R9!* Instead, adjust R8 fully clockwise.

With the ac line cord removed, check that F2 is open. Replace F2 with a new fuse (now you know why two of the three fuses are called for). Reconnect the ac line cord, and while continually monitoring the output voltage, gradually adjust R8 until the Q1 again triggers at 14.2 V, blowing F2.

Again, disconnect the line cord from the wall socket, set R8 fully clockwise, replace F2 (there's the third fuse!) and reset R8 for 13.8 V. This completes the voltage calibration and overvoltage protection tests. The power supply is now set to 13.8 V output, with the overvoltage protection set for 14.2 V.

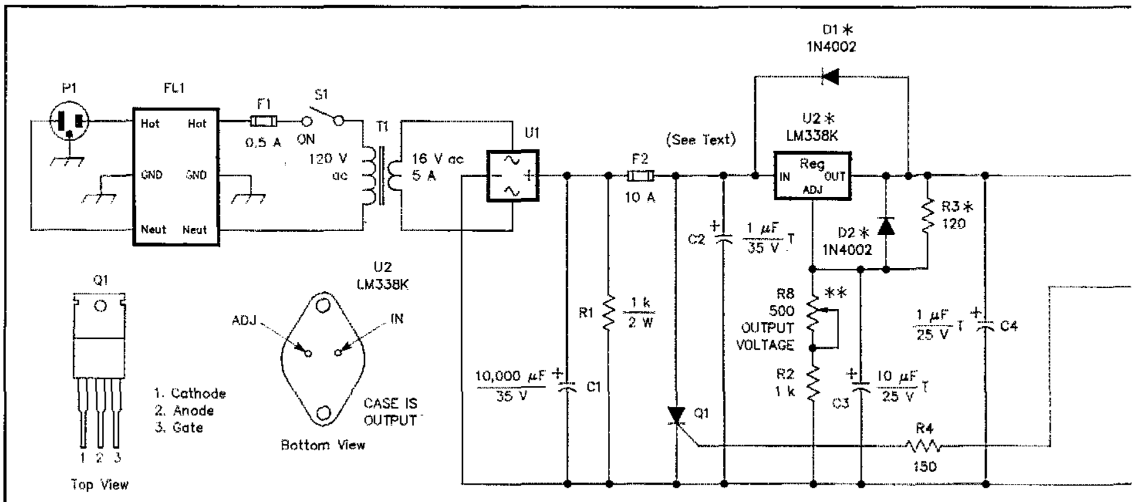
Adjust the variable resistive load to maximum resistance and connect it to the power-supply output in series with the ammeter. Turn on the power supply and gradually adjust R10 until a current of 5 A flows. Decrease the resistance further and check that the current limits between 5.5 A and 6.5 A.

Finally, be thoroughly unpleasant and apply a short circuit via the ammeter. Check that the current-limiting feature operates correctly. My prototype limited at approximately 3.5 A. Disconnect the ac line cord and test equipment, close up the case and your power supply is ready for service.

Summary

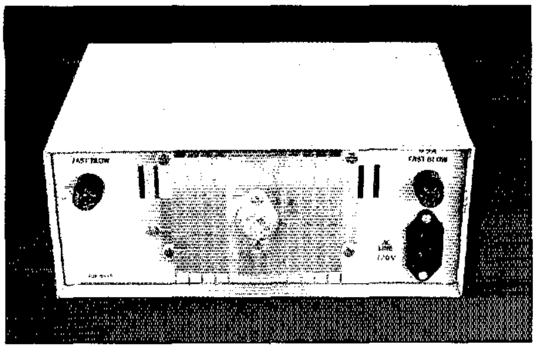
My prototype supply powers a 25-W transmitter, which continually draws 4.5 A. Under these conditions, the entire power supply gets hot, so ventilation holes were eventually drilled into the case (in my prototype, the heat sinks are mounted *inside* the cabinet). The heat sink gets hot enough to scorch the skin of the unwary, so be careful!

This power supply is suitable for a range of tasks. I've used my prototype supply for some time. It's proved itself to be reliable and is a money-saving alternative to purchasing a power supply of similar quality. Not only that, I had the fun of building it myself—so can *you!*



Except as indicated, decimal values of capacitance are in microfarads (µF); others are in picofarads (pF); resistances are in ohms; k=1,000.

T=Tantalum
N.C.=Not Connected
* Mount on Heat Sink
** See Text and Caption



A rear-panel view of the power supply. Louvers in the enclosure bottom and on the rear panel provide convective cooling.

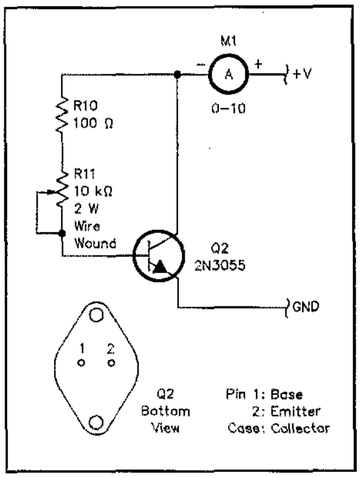


Fig 3—An active resistive load to use in testing the supply. Adjustment of R11 is sensitive.

- M1—Multimeter or ammeter capable of measuring 10 A.
- Q2—2N3055 (Radio Shack 276-2041); mount on heat sink.
- R10—100-Ω, ½-W carbon composition or film resistor (271-012).
- R11—10-kΩ, 2-W wire-wound potentiometer.

Notes

[1] Recent editions of *The ARRL Handbook* list parts-suppliers in Chapter 35; in the 1993 *Handbook*, it's on pages 35-38 to 35-40. Here are some places to shop for transformers and other components used in this project: American Design Components, 400 County Ave, Secaucus, NJ 07094, tel 800-776-3700, 201-601-8999, fax 201-601-8990: T1 can consist of two #15724 transformers ($4.95 each). These transformers have a 120-V primary and dual 18-V, 3.1-A and 24-V c.t., 3.5-A secondary windings; connect the two 18-V windings in parallel to produce an 18-V, 6.2-A secondary.

Mendelson Electronics Co, Inc, 340 E First St, Dayton, OH 45402, tel 800-344-4465, 513-461-3525, fax 513-461-3391: For T1, a #600-0190F transformer can be used (120-V primary, 35-V c.t. 10-A secondary, $14.95).

Fair Radio Sales Co, Inc, PO Box 1105, 1016 E Eureka St, Lima, OH 45802, tel 419-227-6573, fax 419-227-1313; For T1, a #5608798 transformer with a 120-/240-V primary and 19.5-V, 6-A secondary, $7. Connect the primary windings in parallel (observe proper phasing) for 120-V operation, or in series for use with 240-V lines. This is the transformer used in the unit shown in the accompanying photographs.

[2] PC boards and overvoltage protection ICs are available from FAR Circuits, 18N640 Field Ct, Dundee, IL 60118-9269. Price: PC board, $3.50, plus $1.50 shipping; MC2423P1 overvoltage IC, $3. The PC board has mounting holes and pads to allow for handling different trimmer-potentiometer footprints.

A PC-board template package is available free from the ARRL. Address your request for the SPENCER 13.5-V, 5-A POWER SUPPLY TEMPLATE to: Technical Department Secretary, ARRL, 225 Main St, Newington, CT 06111. Please enclose a business-size SASE.

[3] If you find the adjustment of R9 to be too sensitive, use a 500-Ω potentiometer in its place and reduce the value of R5 (if necessary) to provide the required adjustment range.

Ben Spencer received his Class B (VHF only) license, G6VAN, in 1984. His first rig was an FM CB radio with a modified 10- to 2-m transverter providing eight VHF channels and 0.4 W output. After upgrading to a Class A license in 1985, he borrowed an 80/40-m 2-W CW rig "that drifted all over the band during a QSO (even the Russians complained)."

After six years in a nontechnical trade in the Royal Air Force, Ben retrained for electronics and got a job with Racal working on military communications equipment. After an 18-month spell in the Middle East, he moved to Plessey Semiconductors working on wafers and device technology. Presently, Ben runs his own electronics design consulting firm. Now, Ben spends less time operating and more time designing and building "odds and ends." These vary considerably in nature and usually because he needs "a thing that does this or that."

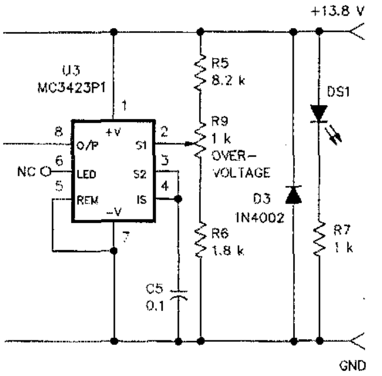


Fig 1—Schematic of the power supply. Part numbers in parentheses are Radio Shack; equivalent parts can be substituted. Unless otherwise specified, resistors are ¼-W, 5%-tolerance carbon-composition or film units.

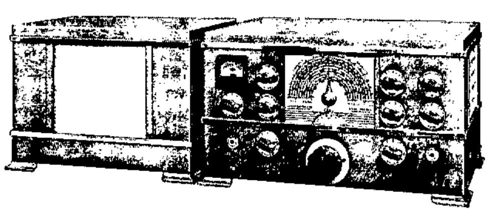
- C1—10,000-µF, 35-V electrolytic.
- C2, C4—1-µF, 35-V tantalum (272-1434).
- C3—10-µF, 35-V tantalum.
- C5—0.1-µF, 25-V ceramic disc (272-135).
- D1-D3, incl—1N4002 (276-1102 suitable).
- DS1—Red LED (276-041).
- F2—Fast-acting 10-A fuses; three required—see text; (270-1325).
- F1—Fast-acting 0.5-A fuse (270-1271).
- FL1—Ac-line filter.
- Q1—BT152 400-V, 25-A SCR in TO-220A package (ECG 5554).
- R1—1 kΩ, 2 W.
- R8—500-Ω, single-turn trimmer potentiometer.
- R9—500-Ω or 1-kΩ, single-turn trimmer potentiometer (see Note 3).
- S1—SPST panel-mount switch (275-634).
- T1—120-V primary, 16- to 20-V, 5-A secondary (see Note 1).
- U1—100-PIV, 6-A bridge rectifier (276-1181 suitable).
- U2—LM338K 5-A adjustable power regulator in a TO-3 package; available from Digi-Key and other sources.
- U3—MC3423P1 overvoltage protection IC (see Note 2).
- Misc: two panel-mount fuse holders (270-364, 270-365, 270-367, or 270-368); IEC line cord; heat sinks for TO-3 case transistors; TO-3 mounting kit and heat-sink grease (276-1371 and 276-1372); two binding posts, one red, one black (274-662); chassis or cabinet, PC board (see Note 2), hardware, rubber hoods, heat-shrink tubing or electrical tape for F1 and FL1, hook-up wire.

Hello Again, to an Old Friend

Old friends and old radios—they're often one and the same!

By Bart Brady-Ciampa, N7NJL
4610 Washington St
Vancouver, WA 98663

The National NC-240-D, circa 1948.



My folks tell me that when I was old enough to pull a chair over to the old RCA Victor, I'd climb up and twist the knobs to tune in foreign broadcasts during WWII. They must have thought it was funny, because they let me continue without fearing that I'd damage the radio. What they didn't realize was that they were sowing the seeds for a lifetime of radio "knob-twisting."

We always had an all-band radio in the house, as did many families in those days. The big dials, printed with the names of exotic countries, were fascinating. My fingers became adept at fine-tuning the "magic-eye" tuning bulb.

When the 1950s rolled around, my Dad brought home a small shortwave receiver that someone had given to him. I was in heaven! I saved my allowances, did odd jobs—anything short of begging—until I had enough money to "upgrade" to a better superhet.

During my "radio craze," my parents voiced the typical concerns: complaints about all the wires coming out of my bedroom window; Mother chiding me to be careful not to put my hands inside that radio "box" while it was plugged in; and Dad rushing into my room when he thought the house was on fire because he smelled the fumes from the soldering iron. (I was forbidden to solder in my bedroom after that!)

Finally, in 1958, when I was a 16-year-old high school student in the Boston area, I had saved enough to buy a "real" receiver: a 1948-vintage National NC-240-D. National radios were quite common in Boston because the factory was in neighboring Malden, Massachusetts.

For the next 10 years, the old National and I were never far apart. Through high school, college, graduate school, marriage, the birth of two children and the beginning years of a career, I could always count on it to provide the solace, comfort and relaxation that we all need from time to time.

But with all the other priorities of young adulthood, the radio, now 20 years old, somehow got tucked away and stored at my folks' house when it came time to pick up stakes and seek my fortune in the world.

As 23 more years passed, and we moved to New Hampshire, Maine, England, Nebraska, and finally to Washington State, every time I'd visit my parents I would make sure that the old National was still in its plastic bag in the garage.

Finally, when we visited last fall, I decided to ship my 43-year-old "boat anchor" back to Washington. With my Dad's help we carefully packed up my treasure, which weighed in at a hefty 67 pounds—just three pounds shy of the shipping company's 70-pound weight limit. We had to ship the speaker in a second box; it weighed another 20 pounds. (It cost only $35 to ship both boxes from coast to coast. What a deal!)

Home at Last

Once home, I could hardly contain myself. Would the carton arrive in one piece? Would any of the tubes be broken? And—miracle of all miracles—would the radio still work?

When the day finally came, I carefully chose a spot on the floor of my shack that would be close enough to an ac outlet so I could plug it in without too much delay. After stripping the layers of cardboard away, collecting every one of those plastic peanuts and carefully removing the radio from its plastic cover, I connected the speaker and flipped the power switch. The tubes started glowing and I could hear the crackling and hissing sound of an antenna-less receiver. It worked!

A few days of fun with the '240-D passed and the excitement wore off a bit as my old friend assumed a position of prominence among the solid-state rigs that are the mainstay of my shack.

As I was casually thumbing through the owner's manual, however, something on the front cover caught my eye. In the upper right-hand corner of the manual's front cover, crossed out and barely discernible, was a name and call sign: Ralph Morris, W1QUE.

I was amazed that I had never taken the trouble to decipher this before! This ham must have been the original owner of the National and had probably put his name there more than 40 years ago. He had sold me the radio when I was 16 years old! I wondered...could he still be listed in the *Callbook?*

I was struck by the irony of old technology meeting new as I slid the *Hamcall* CD into my computer's CD-ROM drive and typed his call sign. Incredible! His name and address were still listed! I wrote down his address, picked up the phone, dialed information for New Bedford, Massachusetts, and immediately got a phone number. I dialed the number and a man answered the phone!

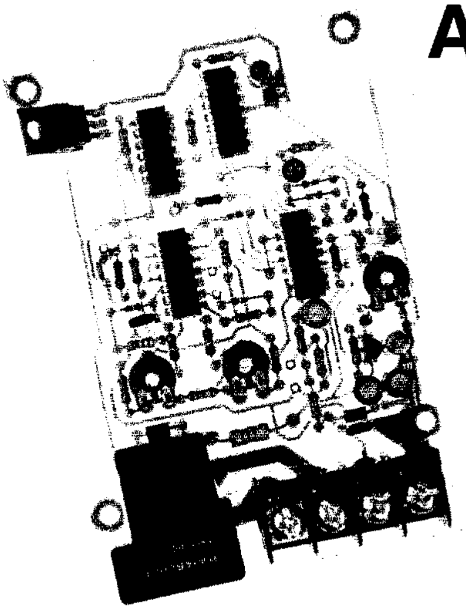
"I'm trying to track down W1QUE," I said. The spry voice on the other end of the phone line answered, "Well, you've found him!" I asked, "Ralph, did you ever own a National NC-240-D?" He said," Yes, that was a great radio. I sold that to a kid about 35-years ago."

I couldn't believe it! After all these years, not only did I have my old radio back, but I had found its original owner and was about to make a new friend. Ralph and I reminisced about ham radio and "back-home" stuff for about 15 minutes. I told him I would send him a letter with all the details about where the old National had been during all those years. I also enclosed a QSL card and asked him for one in return, just to have as a keepsake.

Several months passed and I received a letter in an envelope with a return address of Margo the Magician. Yes, it was from Ralph. It was a long newsy letter and tucked in the envelope was a QSL card and some show tickets and memorabilia. It seems that he and his wife used to travel quite a bit and had a professional magic act that took them far and wide. He told me about his personal life and that he's still active on the bands. It was a super letter and caused me to reflect deeply.

Now, when I go to a hamfest or a flea market and see all the old "boat anchors" for sale, I say to myself: *If those old radios could talk, what great stories would they tell?*

The next time you're at a hamfest, try to imagine the people those old radios have brought together over the years and think about the quality they have added to their lives.



A Low-Voltage Disconnect

Whether you're operating a repeater, operating on emergency power or just watching TV in your RV, this little gadget protects your batteries from damage.

By Michael Bryce, WB8VGE
2225 Mayflower NW
Massillon, Ohio 44647

Many Amateur Radio stations use batteries to power their radio equipment during commercial electric power outages. Some of us use battery power all the time in the shack. Keeping an eye on the battery's charge sometimes can't be done (or is forgotten altogether) until you are unexpectedly—and unwillingly—off the air!

What you need is a battery watchdog—something to keep track of the battery and disconnect loads when the battery just about goes kaput. No matter what your use of battery power—whether you own a camper, RV, or just fish on the lake beside your cottage—this contraption does the battery monitoring for you and protects your battery from severe discharge as well. Repeater owners and operators may find the device an ideal way to extend operating time while on emergency power.

What's It Do?

The low-voltage disconnect (LVD) automatically disconnects a load from the battery before damage is done to the battery or the load. The potential across a discharging battery's terminals depends on the battery's state of charge and the load's discharge rate. This circuit monitors the battery terminal voltage, and when it reaches a preset level, a relay is de-energized disconnecting the load from the battery. There's an approximate 5-second delay before the device senses the low-voltage set point and the relay drops.

Take a look at some of the features of this watchdog:

- User-adjustable turn-off voltage.
- User-adjustable reset voltage.
- Built-in delay to ignore temporary low-voltage conditions.
- 30-A-capacity relay contacts.
- Low current consumption.
- Easy construction using a readily available PC board and components.

The load can be connected directly to the LVD's relay, or its relay can drive an off-board, heavy-duty power relay, if need be. You can also use the PC-board-mounted relay to control a 120-V ac load (I'll talk about that later). The LVD can also supply logic to a repeater controller, too.

Circuit Description

To see how the circuit works, refer to Fig 1. U5, an LM317LZ 100-mA adjustable-voltage regulator, creates a reference voltage for the comparators. This voltage (4.00) is set by R30 (**REFERENCE ADJ**), a 1-kΩ potentiometer. R27 places a 4-mA load on the regulator's output and improves regulator stability. U2C buffers the reference voltage. From here, the reference voltage goes to the voltage comparators. D1, a 1N4001 diode, protects U5 from reversed power-supply voltage polarity.

R1 and R2 halve the battery terminal voltage. R3 and C9 help filter out battery-line noise. U2B acts as a buffer between the voltage divider and the battery sense line.

Two set points are needed to control the LVD. One turns on the LVD. That causes K1 to drop out, disconnecting the load. The other set point turns off the LVD, closing the relay contacts and reconnecting the load. If it weren't for the two different set points, the LVD would constantly switch on and off. The difference in the set-point voltages allows the battery to recover before the load is reconnected. (This assumes you have some means of recharging the battery after the load has been disconnected.)

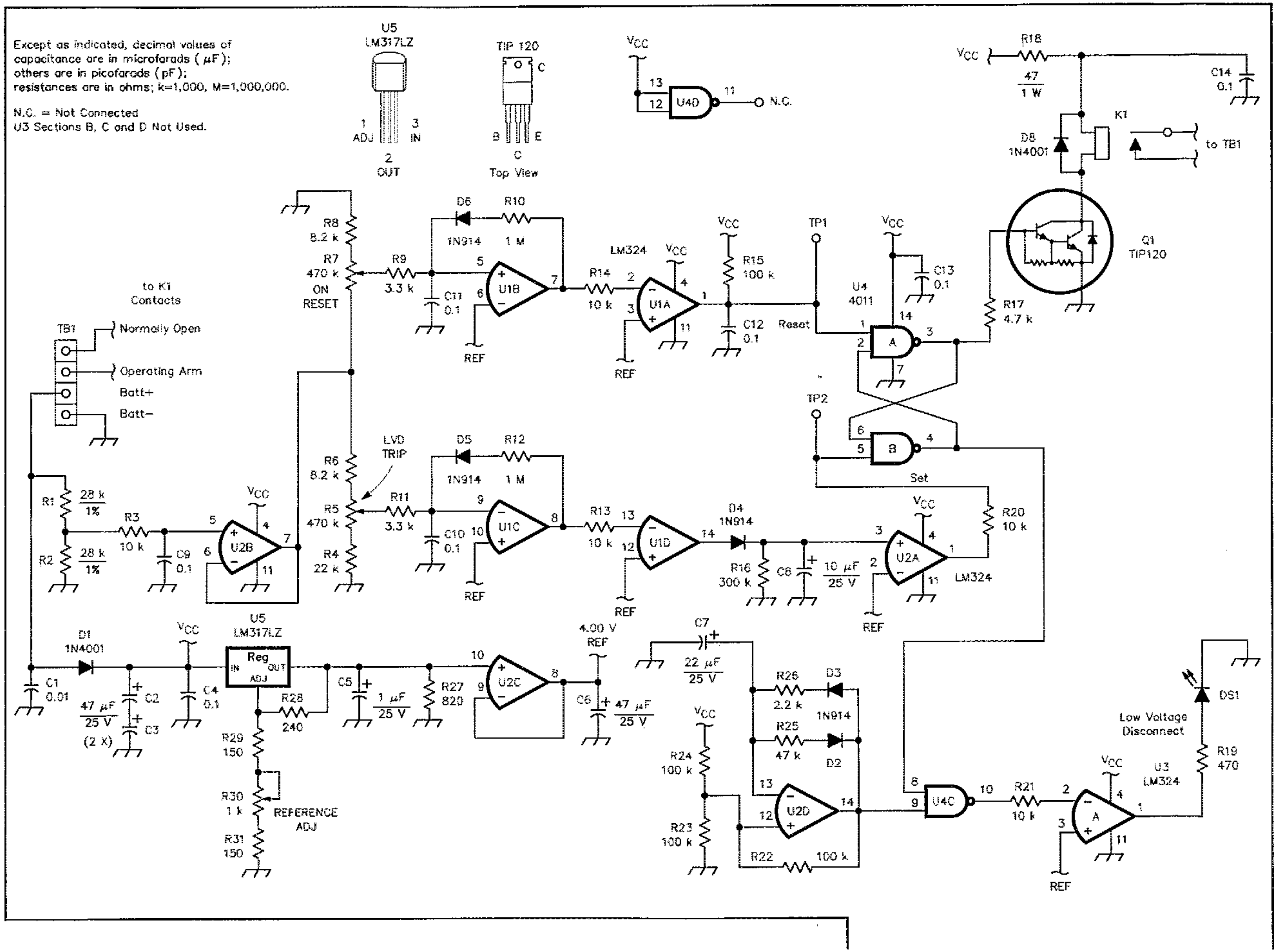
The voltage comparators are nearly identical. The battery voltage, now divided by two, is applied to two voltage dividers. Let's first look at the trip comparator.

The battery's output voltage is divided in half and applied to comparators U1C and U1B through U2B. If the battery was discharged to 10 volts,[1] the trip comparator (U1C) input would see less than 5 volts. The reference voltage at pin 10 is 4 volts. By adjusting R5 (**LVD TRIP**), we can set the comparator to switch states when the input voltage at U1C pin 9 equals the reference voltage. D5 and R12 provide a bit of hysteresis to keep U1C from oscillating. Because U1C may not provide the needed high-to-low positive switching action, a second op-amp section (U1D) is used. U1D provides the switch-like on/off state needed by the delay circuit composed of D4, R16 and C8, which provides a delay of about 5 seconds. Again, to provide the required logic levels, U2A is used. Its output goes to the SET point of the R/S latch, U4A and B.

The **ON RESET** circuit (U1B) works similarly, but has no delay circuit. U1B's output is routed through U1A and goes to the RESET of the R/S latch, U4A.

When the battery voltage is above the LVD trip point, the output of the R/S latch (U4A pin 3) is high. This turns on Q1, a TIP-120 NPN Darlington power transistor, which energizes K1. A 1-W, 47-Ω resistor (R18) limits relay current. This minimizes the overall current demand of the LVD. D8 protects Q1 from the back EMF produced when the relay coil's magnetic field collapses.

[1] Notes appear on page 34.



Except as indicated, decimal values of capacitance are in microfarads (μF); others are in picofarads (pF); resistances are in ohms; k=1,000, M=1,000,000.

N.C. = Not Connected
U3 Sections B, C and D Not Used.

Fig 1—Schematic of the low-voltage disconnect circuit. Equivalent parts can be substituted. Unless otherwise specified, resistors are ¼-W, 5%-tolerance carbon-composition or film units.

U1-3—LM324 quad op amp (Mouser 511-LM324; Digi-Key LM324N).
U4—4011 quad gate (Mouser 511-4011; Digi-Key 4011CD).
U5—LM317LZ 5-V, 1-A adjustable regulator (Mouser LM317LZ; Digi-Key LM317LZ).
K1—Potter and Brumfield T-90 series; 12-V dc, 155-Ω coil, SPST, 30-A normally open contacts (Digi-Key PB110-ND).
TB1—Terminal block (Mouser 506-8PCV-04).
R5, R7—470-kΩ trimmer potentiometer (Mouser 531-PT15D-470K).
R30—1-kΩ trimmer potentiometer (Mouser 531-PT15D-1K).
Q1—TIP-120 Darlington power transistor (Mouser 511-TIP-120; Digi-Key TIP120PH-ND).

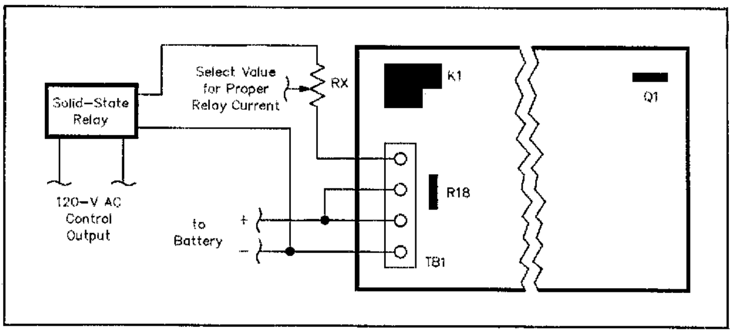


Fig 2—An off-board solid-state relay provides an excellent means of controlling 120-V-ac operated loads. Here, the relay derives its operating current and voltage from the battery being monitored. The resistance value and power rating of RX is chosen for proper relay operating current.

As the battery discharges, its terminal voltage falls. When the LVD turn-on voltage is reached, there's a 5-second delay, then the output of U2A goes low, setting the R/S latch. This removes Q1's base drive, causing K1 to de-energize. The output of U4 pin 4 connects to U4 pin 8 to allow the oscillator (U2D) to output through U4C to U3A, which turns on the **LOW-VOLTAGE** LED, DS1. DS1 flashes at a rate determined by R25, R26 and C7. With the values shown, the on-time is about 1/20th of a second. Otherwise, DS1 remains dark. This arrangement reduces the circuit current drain during a low-voltage battery condition.

Construction

There's nothing critical here: perfboard, wire-wrap, dead-bug or PC-board construction are all suitable. A PC board is available[2] as well as a complete kit of parts.[3] Using a PC board speeds construction and makes troubleshooting easier.

You can buy most of the parts from a well-stocked Radio Shack store. Mouser Electronics[4] can supply the parts "the Shack" does not carry. In both cases, the single exception is the relay. Obviously, the one specified fits the PC board. Secondly, it has a hefty contact rating (30 A) and is inexpensive (less than $4). This relay is available from Digi-Key.[5] Certainly, you can use a different relay, but you'll probably have to mount it off-board. Also, you may have to change R18's value, as it's dependent on the relay's coil current. The value shown is calculated for the relay identified in the caption. (By the way, mount R18 with a ¼- to ½-inch clearance between the resistor body and the PC board. This allows air to circulate around the resistor and prevents the dissipated heat from discoloring the PC board.)

Component values aren't critical. Use equivalent parts you have on hand. If you don't have a 300-kΩ resistor, a 270-kΩ will work just fine. (The ±10% rule of thumb can be safely applied.) Use sockets for the ICs and be careful when handling U4: It's subject to damage by static discharges, so use a wrist strap. Install all the parts—*except for C8*. After you've assembled the PC board, check your work to ensure the diodes and capacitors are properly polarized. Ensure you have the ICs oriented properly before you apply power to the board.

Set Up and Test

You'll need a digital VOM[6] and a variable-voltage power supply to adjust the LVD. Connect your power supply to the terminal block at the battery terminals.

Set the power supply output to 14 volts. Apply power to the board. Probe pin 10 of U2. Verify the presence of the reference voltage; in all probability, it's not 4.00 volts—yet. Move the probe to U2 pin 8 and verify the reference voltage is there, too. Now, adjust R30 (**REFERENCE ADJ**) until the voltage at U2 pin 8 is 4.00 volts. Next, set your power supply to 10.5 volts. Measure the voltage at pin 7 of U2. It should be 5.25 volts (half of 10.5 volts).

Adjustment Method

First, *remove C8* (in case you forgot the earlier warning and soldered it in) from the PC board. This defeats the delay circuit. Turn both R7 and R5 fully counterclockwise. Set your power supply output to 13.5 volts. Connect the power supply to the LVD, then turn it on. The **LOW VOLTAGE DISCONNECT** LED (DS1) should be blinking. Slowly adjust R7 (**ON RESET**) until the relay closes and DS1 goes dark. Reduce the power supply output to 10.5 volts. Slowly turn R5 (**LVD TRIP**) until K1 drops out and DS1 begins to flash. Verify the two set points by raising the power supply output voltage to 13.5 volts. K1 should energize. Reduce the voltage to 10.5 volts and K1 drops out. *Now* you can install C8! You're done!

Troubleshooting

If you can't get the circuit to work, first check for the presence of the reference voltage. Without it, you'll be dead in the water from the start. Check for 4.00 volts on pins 3, 6, 10 and 12 of U1, and at U2 pin 2 and U3 pin 3.

If the comparators won't switch (and you have the proper reference voltage) check the battery sense line by checking the output voltage at U2 pin 7. (This voltage should be one-half of the power-supply voltage applied to the battery sense line.) As you can see, replacing U1 replaces *all* the battery sense comparators.

When the battery voltage is at 10.5, you'll be able to see the delay action by probing U2 pin 3. At this pin, you'll see the voltage slowly drop during the 5-second delay period.

If you used a relay other than the one specified, the value of the current-limiting resistor (R18) may be too high to allow the relay to energize. Try reducing R18's value (or short it out).

Hooking Up the LVD

With only four wires, hook-up is a breeze! Simply connect the battery you intend to monitor to TB1's battery terminals. K1's contacts are completely isolated from the battery. By connecting a jumper from the +12-volt battery terminal to one of the relay contacts, you can deliver battery power through the relay contacts to your load.

Remember, although K1's contacts can carry 30 amperes, voltage drop caused by long wire runs can have an effect on the load. If you need to control heavy current loads, use K1 to control a power relay located right at the load. Install a protective diode across the power-relay's coil terminals to prevent inductive kick-back.

You can use K1's contacts to control logic levels to a repeater. Connect K1's contacts to ground or +12 V via a current-limiting resistor. If you have a repeater controller and it requires a logic input, this is one way to go.

Controlling a 120-V AC Load

Although K1's contacts easily handle a 120-volt load, having an exposed 120-volt line connected to TB1 would keep me up at night! A safer way to control such a load is to use an off-board solid-state relay (see Fig 2).[7] (A solid-state relay is an optically coupled device that provides excellent isolation between the load and the driving source. In this case, between the 120-V ac mains and your battery.) Solid-state relays with various control voltages are readily available. RX, an external resistor, serves to limit the current flowing to the solid-state relay. By properly altering the value of RX, you can use a 5- or 12-volt control line. As mentioned earlier, the battery can supply power to operate the solid-state relay via K1's relay contacts.

Life with an LVD

While the LVD certainly protects your battery from deep discharge, it's not perfect. (What is?) Every LVD consumes *some* power from the battery it's trying to protect. In this case, when the LVD has the relay pulled in, it draws about 90 mA. If you run the LVD 24 hours a day, you have a 2.16-Ah load just for the LVD. Even with the relay off, and the battery at 10.5 volts, the LVD draws about 12 mA. So, if your battery is being charged by a solar array, be sure to include the LVD load requirements when performing your sizing calculations. To save power when you're not using the load, turning off the LVD automatically turns off the load connected to the relay.

If your battery is charged from a 120-V ac charger, you'll not have to worry about the extra LVD load. Repeaters operators normally have a battery back-up system constantly being charged. When the grid power fails, the battery takes over. When the battery discharges to the point that the LVD trips, the LVD can then take the power amplifier off line to extend battery operation until the grid power comes back on.

The LVD load should not be your main load. *Shedding* loads is the main job of the LVD. You don't want to have it shut down *everything*, but disconnect what you can live without. For instance, in an RV, you may want to connect the running lights to the LVD and leave your TV bypassed. When the battery becomes so low as to trip the LVD, the running lights will be disconnected. (Given the nature of what's on TV these days, it's probably a better idea to take the TV off line and keep the running lights on!)

Here's another example of choice: You have a sailboat docked in the lake. A bilge pump is connected to the LVD. If too much water leaks into the boat and the pump is running all the time, the LVD will disconnect the pump from the battery protecting the battery from damage. With the pump disconnected from the battery, the pump won't work any more and before you know it, your sailboat has become a submarine!

A much better way to prevent your sailboat from sinking is to have the LVD warn you of the discharged battery. The warning could be as simple as a flashing light or a buzzer. If you really want to go to the extreme, you could combine the LVD and the METCON II[8] for telemetry. The LVD keeps a constant eye on your batteries, while you work the world on your radio, or just fish in the lake by your cottage.

Notes

[1] A lead-acid battery is generally considered dead when the terminal voltage is 10.5 under load.

[2] A PC board for this project is available from FAR Circuits, 18N640 Field Ct, Dundee, IL 60118-9269. Price: $12, plus $1.50 shipping. A PC-board template package is available free from the ARRL. Address your request for the LOW VOLTAGE DISCONNECT TEMPLATE to: Technical Department Secretary, ARRL, 225 Main St, Newington, CT 06111. Please enclose a business-size SASE.

[3] A complete kit of parts, including the PC board and relay, is available from SunLight Energy Systems, 2225 Mayflower NW, Massillon, OH 44647. Price: $55 plus $3 shipping.

[4] Mouser Electronics, 2401 Hwy 287 N, Mansfield, TX 76062; tel 800-346-6873, 817-483-4422, fax: 817-483-0931.

[5] Digi-Key Corp, 701 Brooks Ave S, PO Box 677, Thief River Falls, MN 56701-0677, tel 800-344-4539, 218-681-6674, fax 218-681-3880.

[6] You can use an analog meter to calibrate the LVD, but a digital voltmeter provides better resolution.

[7] Available from All Electronics Corp, PO Box 567, Van Nuys, CA 91408-0567, tel 800-826-5432, 818-997-1806, fax 818-781-2653.

[8] P. Newland, "Introducing METCON, a New Remote control and Telemetry System," *QST*, Jan 1993, pp 41-47.

New Products

DC POWER PACK

◊ When you plan a portable operation that may last all day and into the night, your hand-held transceiver's standard battery may not be able to keep you on the air. Innova's popular DC PowerPack can run your rig for days—it can even start your car. Now Innova has a scaled-down version available for lower-current applications. The Innova DC PowerPack /S (model 06-1130) uses a sealed lead-acid battery to provide 3.0 Ah at 12 V dc (36 W/h). It's rated at 250-1000 recharges (from a car, ac outlet or with optional solar charger) and uses a standard 10-A automotive fuse for output protection. It weighs three pounds and has a universal automotive adapter to plug in a ham radio, cellular telephone, emergency lighting, notebook computer, etc. It can run a hand-held transceiver at high power for hours. Suggested retail price $61.52. Innova Electronics Corp, 17287 Mt Herrmann St, Fountain Valley, CA 92708; tel 714-432-1184, fax 714-432-7910.

CW FROM YOUR PC

◊ If you have an IBM or compatible computer running DOS, Windows, DesqView or OS/2, you can feed your radio Morse code with a small, but feature-filled terminate-and-stay-resident (TSR) program. *Pop-Up Morse* uses less than 8 kbytes of RAM, and enables you to bring up its screen whenever you need it. *Pop-Up Morse* uses the standard computer-to-transceiver interface, compatible with K1EA's *CT* contest-logging program, and keys a rig at 10-60 WPM through any serial or parallel port. It has eight 50-character memories and sends the full range of English and foreign-language characters, including Russian and amateur prosigns. Price $29.95 postpaid (specify 3½- or 5¼-inch high-density diskette). Pete Smith, N4ZR Software, 2003 Sarazen Pl, Reston, VA 22091.

New Books

HF ANTENNAS FOR ALL LOCATIONS

By Les Moxon BSc CEng MIEE, G6XN

Radio Society of Great Britain (RSGB), Cranborne Rd, Potters Bar, Herts EN6 3JE, England. Second Edition, revised and expanded 1993. Paperback, 7½ × 9⅝ inches, 20 chapters, 322 pp, including an eight-page index, many diagrams and photos. £13.50. (Distributed in the US by the ARRL, $20 plus s/h.)

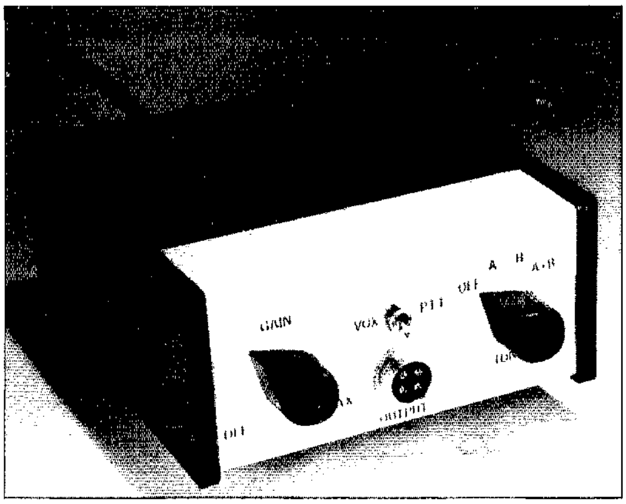
Reviewed By Jerry Hall, K1TD
ARRL Associate Technical Editor (Retired)

Les Moxon published the first edition of *HF Antennas* in 1982, and the book quickly became a valuable reference for antenna builders and experimenters around the world. Practical information was presented on the common antennas used in Amateur Radio and on many antennas that aren't so common. Les included a wealth of ideas and tips on antenna design, construction and installation, and thought-provoking suggestions for experimenters wishing to further develop antenna ideas presented in print. There was also extensive information on feeding and matching.

Amateur Radio has changed a bit in the years since the first edition appeared. The most significant with respect to amateur antennas is that hams acquired three new bands (10, 18 and 24 MHz), based on resolutions made at the 1979 World Administrative Radio Conference. The "all-band" HF antennas of yesteryear no longer cover all available bands. In addition, antenna designs have become more complex, and the need for small antenna systems has increased as society becomes more urbanized worldwide. New designs meeting these changes have evolved.

The 1993 second edition of *HF Antennas for All Locations* still has all the good information and ideas of the first edition, but goes on to present an abundance of new information to meet the needs of today's antenna enthusiast. The text has been revised and greatly expanded. Although many basic discussions in the book draw from the previous edition, the text is updated and in many cases clarified, "to leave fewer questions unanswered so that the experimenter is fully equipped for evaluating new ideas and dealing with whatever problems may come his way," as Les states in his preface.

There are more novel antenna designs in this edition, including beams that cover more bands with fewer problems and no trap losses. A new chapter focuses on small antennas, with particular emphasis on small transmitting loops. There's plenty of practical information on the choice and construction of antennas to suit most locations and requirements. This book explains the "why" and the "how" of HF antennas, and takes a critical look at existing designs in light of the latest developments. If you're an antenna builder or experimenter, you'll find this new edition to be a valuable reference.



(photo by the author)

A Simple, Two-Tone Audio Generator

Three transistors, a handful of parts and a few hours of building time from now, you can stop whistling into your microphone forever!

By James G. Lee, W6VAT
1060 Big Oak Ct
San Jose, CA 95129-3105

"*Heeeeeello*, testing one, two, three." Tuning up with an SSB transmitter gets old fast if you have to make audio tones yourself. Here's a simple circuit that generates one or both of two audio tones for such activities as SSB transmitter testing, power-amplifier linearity checks, and other general audio troubleshooting and adjustments.

The Generator Circuit

See Fig 1. The tone generator uses two JFETs (Q1 and Q2) in a slightly modified Hartley configuration.[1,2] Small audio output transformers (T1 and T2) serve as the oscillators' tuned-circuit inductances. The oscillators operate at about 600 and 2100 Hz, which keeps them nonharmonically related, yet spaced a suitable distance apart in the voice frequency range to be useful for testing and alignment. The precise frequencies are not critical, and you can change them as described later.

Normally, a Hartley oscillator's feedback is set by varying the placement of its inductor tap. Because the transformers' taps are fixed, I used adjustable resistors in the JFETs' source leads (**FEEDBACK** controls R1 and R3) to set the oscillators' feedback at the proper value. Too little resistance may allow such strong oscillation that the transistors destroy themselves by gate overvoltage. Too much resistance prevents oscillation. The optimum resistance range lies between 910 and 1800 Ω, so 910-Ω resistors (R13 and R14) in series with R1 and R3 keep the feedback-adjustment range under the safe limit.

The transformers' secondary windings provide output coupling for the oscillators. Switching (via S1, **MODE**) allows the selection of either tone (**A** or **B**), both tones (**A + B**), or none (**OFF**). In the **A + B** mode, the oscillators must be sufficiently isolated from each other to prevent interaction. The summing network (R2, R4, and **GAIN** control R5) conveys the oscillator output to the output amplifier (Q3). This approach allows control over output voltage without interaction. The output amplifier provides further buffering, as well as impedance matching for driving a variety of loads. It uses a Darlington transistor connected as an emitter follower.

Q3's output impedance is about 2 Ω—too low for some applications. The test generator's overall output impedance (Z_{out}) is roughly determined by the value of R9 and the reactance of C9 in series with Q3's output impedance. Using a 560-Ω resistor at R9 sets the tone generator's output impedance at about 600 Ω.[3]

As a convenience, S3 (**PTT**) can be connected to the radio under test so you can control the rig from the tone generator's front panel.

Power for the unit can be supplied by batteries or, as shown in Fig 1, a small ac-line-operated supply. The supply provides two voltages: 5 volts, regulated, for the oscillators, and 9 volts, unregulated, for the output amplifier. For portability, you can use batteries in place of the ac supply. If you use batteries, limit the oscillators' supply voltage to +5 or less to keep the oscillator transistors' feedback within the safe zone. The output stage runs satisfactorily at a collector voltage between +6 and +12.

The tone generator operates only at audio frequencies, so constructing it successfully doesn't require much care in part placement. If you keep the wiring to its controls short, you don't even have to use shielded wire. Just keep the circuit's ac-supply wiring as far away from the oscillators as possible. It's a good idea to house the generator in a metal box to shield it against ac hum and RFI. I built my version in a Ten-Tec enclosure.

Adjusting and Using the Tone Generator

Only two simple adjustments are necessary. R1 and R3 must be adjusted to supply approximately 0.4 volt peak-to-peak (about 140 millivolts RMS) from each oscillator to the **GAIN** control. To do this, use a sensitive ac voltmeter or, ideally, an oscilloscope. (A digital voltmeter or multimeter capable of reading 200 millivolts full-scale will work.) Set both pots to midrange. Put the meter leads (or your scope probe) from the top of the **GAIN** pot to ground, and set S1 to connect the proper signal to the **GAIN** control. Slowly adjust each pot for equal meter readings in the range of 140 to 160 millivolts. There may be no indication on the meter until the pots have been adjusted well past their midpoints. Then there should be a sudden rise in voltage as each oscillator begins working. When each oscillator is adjusted to the proper value, the unit is ready for use.

You can easily change the oscillator frequencies by changing the value of the capacitors (C1 and C3) across T1 and T2. R13 and R14 should protect Q1 and Q2 as you adjust R1 and R3 for optimum, but just to be on the safe side, adjust them from their highest-resistance settings *downward*.

Using the Tone Generator

Covering the details of SSB transceiver or amplifier alignment is beyond the scope of this article, but the basic steps in using the

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[1] Notes appear on page 36.

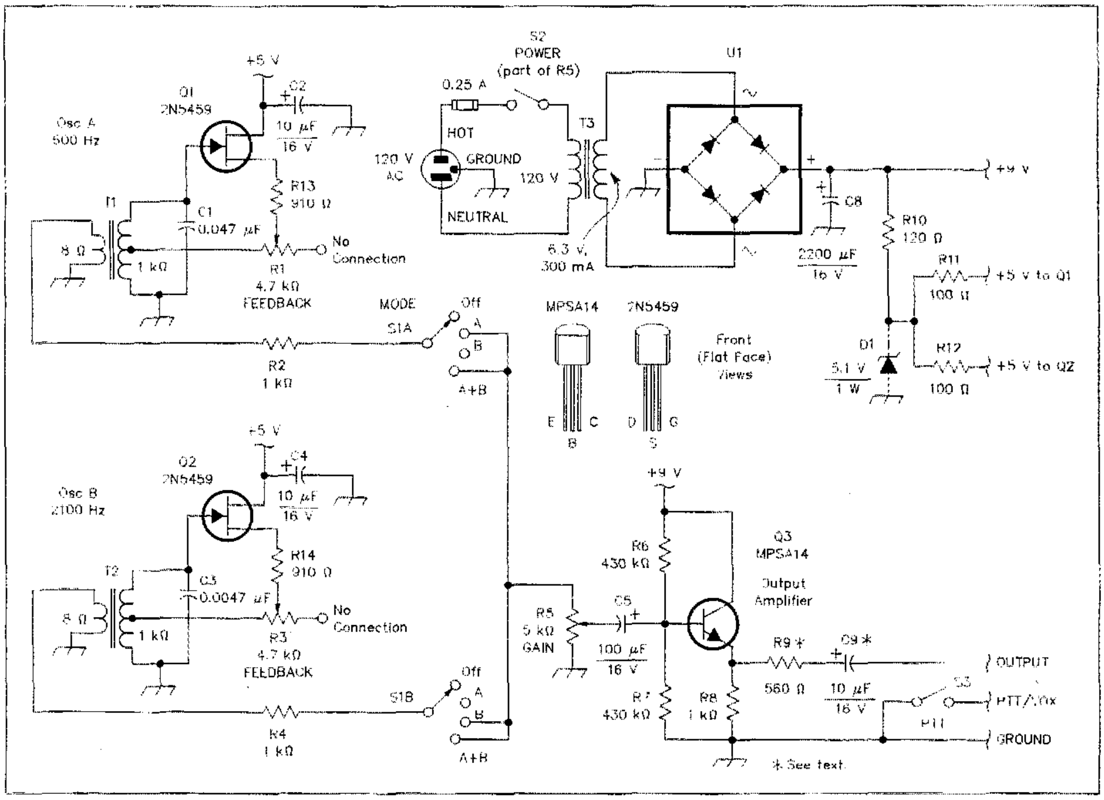

Fig 1—The two-tone audio generator uses just three transistors and makes a good project for beginning builders. The electrolytic capacitors in this circuit are rated at 16 WVDC; the non-electrolytic capacitors are 50-volt plastic-film or ceramic units; the fixed-value resistors are ¼-watt carbon composition or film. Use the connector(s) appropriate for your application for the generator's **OUTPUT**, **PTT/VOX** and **GROUND** lines. For more on how to get parts for and build circuits like this, see Bruce Hale's "Build It Yourself From *QST*" series in the April through July 1992 issues of *QST*.

D1—5.1-V, 1-watt Zener diode (1N4733—Radio Shack 276-565 suitable).
Q1, Q2—2N5459 JFET—any general-purpose N-channel JFET (MPF102 [276-2062] usable.
Q3—MPSA14 NPN Darlington transistor (Digi-Key Electronics, Ocean State Electronics).
R1, R3—4.7-kΩ PC-board-mount control (271-281).
R5—5-kΩ, audio-taper control (271-1720).
S1—2-pole, 6-position rotary switch (275-1386), 2 positions unused.
S2—SPST switch (mounted on R5) (271-1740).
S3—SPST toggle switch (275-612).
T1, T2—1-kΩ-center-tapped to 8-audio transformer (273-1380).
T3—6.3-V, 300-mA power transformer.
U1—50-PIV, 1-A bridge rectifier (276-1161). Four discrete diodes (1N4001 [276-1101]) suitable as a substitute.

tone generator stay the same no matter what you use it for. Make sure the tone generator is turned off by setting the **GAIN** control to that position. Hook up all necessary cables, turn the unit on but leave the **GAIN** control fully counterclockwise (at minimum output). Then proceed with whatever settings are necessary on the equipment you are testing. Often, manufacturer's literature tells you which controls to set and how to set them.

Either of the generator's tones should suffice for single-tone testing. Once you have properly set the transceiver controls, simply advance the **GAIN** control until you see a reading on the meter. Then follow the appropriate instructions in your transceiver manual. Once the transceiver has been set up correctly for CW, switching to either sideband and setting the **MODE** switch to **A + B** will allow you to examine its two-tone-test patterns on an oscilloscope. Note that the tone generator's output level does not change when you switch between single and two-tone operation.

The *ARRL Handbook* covers SSB-transmitter alignment and linearity testing. Back issues of *QST*, and some older ARRL publications, such as *Single Sideband for the Radio Amateur*, also contain detailed information on using two-tone generators for amplifier testing. Be careful when testing equipment if you have to remove any covers, since there really is no such thing as low voltage. Remember your ABC's—Always Be Careful—when testing equipment that is powered up.

Notes

[1] H. Harris, "Simplified Q-Multiplier," *Electronics*, May 1951, p 130.

[2] D. Tong, "A Stable Q-Multiplier," *Wireless World*, Dec 1971, p 600.

[3] C9 is necessary: It keeps Q3's emitter current from flowing through R9 and the device under test. To minimize C9's effect on the generator's output impedance, choose its value so that its reactance is less than 10% of the desired output impedance at the lower of the two tone frequencies. I selected an output impedance of about 600 Ω, 10% of which is 60 Ω. At 600 Hz, the lower tone frequency, the reactance of 10 µF (a commonly available value) is 27 Ω—acceptably low. I made R9 560 Ω. Using the impedance equation for resistance and reactance in series ($Z = \sqrt{R^2 + X^2}$), the impedance at both tone frequencies of R9 and C9 in series therefore equates to just over 560 Ω. This value, added to Q3's output impedance (2 Ω), results in an output impedance of about 562 Ω—close enough to 600 Ω for my purposes. A 620-Ω resistor at R9 would make it closer still.

Jim Lee holds a Bachelor of Arts degree in Mathematics and Physics (Electronics) from UCLA. He has been licensed since 1944, and enjoys DXing and public-service operating when he's not designing and building gear for the shack and workshop. Jim recently retired from GTE-Sylvania where he was a test engineer and a satellite ground systems engineer.

QST

An Update on Compact Transmitting Loops

What did you say you were using on 40 meters? A one-meter loop?

By John S. "Jack" Belrose, VE2CV
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Aylmer, PQ J9J 1G1
Canada

Electrically small transmitting loops have been around since 1957,[1] but this type of antenna has not been widely used by radio amateurs, except perhaps in Europe. This is unfortunate, given the fact that a compact loop can be both an effective and an inconspicuous, neighbor-friendly radiator.

Fig 1A sketches the method of tuning and matching used by Patterson. Subsequent experiments by Lew McCoy[2] patterned after the Patterson loop did not shed a favorable light on the performance of such an electrically small loop for use by radio amateurs for the 75- and 40-meter bands. McCoy cited the critical elements of loop conductor and connection losses and the difficulties in keeping these low with typical construction techniques available to most radio amateurs. In Germany, Christian Käferlein, DK5CZ, developed the AMA series of electrically small loops, and has been manufacturing them since 1983. AMA stands for *Abstimmbare Magnetische Antennen* – tunable magnetic antennas. The electrical design for these loops was worked out by Hans Wurtz, DL2FA. AMA loops are currently available for the 160- to 10-meter amateur bands. Four sizes are sold, ranging in diameter from 0.8 to 3.4 meters.[3] In 1986, Robert T. (Ted) Hart, W5QJR,[4] independently analyzed the compact loop and described practical designs for the radio amateur. But still it has not been widely used in Canada and the United States. I have not made an on-the-air contact with any North American amateur who was using a compact loop, but I suspect that will change. Recently, two US companies catering to the Amateur Radio market have developed small loops for the 30- through 10-meter bands: the Isoloop (AEA[5]) and the Super Hi-Q Loop (MFJ[6]). Brian Battles, WS1O, reviewed the AEA Isoloop for *QST*, and ran a number of comparison tests between it and a reference dipole.[7]

Table 1 summarizes the various types of compact loops currently available. High radiation efficiencies, 38% to 95% depending on the size of the loop and the frequency, are claimed, with conductor losses better than −4.2 dB. However, there are conflicting opinions concerning the performance of such loops, and overly enthusiastic claims sometimes heard about the compact loop can be misleading. This article is meant to give an overview of the actual characteristics and expected performance of electrically small transmitting loops.

Characteristics of Compact Loops

Electrically small loops (perimeters 0.04 to 0.1 λ) are characterized by a very small radiation resistance. Therefore, such loops must be fabricated from large-diameter tubing to keep conductor losses small. In many designs, 2.5- to 10-cm-diameter aluminum tubing has been used. Since the loops are inductive, they can be readily tuned by means of a series capacitor. Power is coupled into them by means of a small coupling loop, a more convenient method of tuning and matching than the original method of Patterson. See Fig 1B.

The radiation efficiency and gain of elec-

Table 1
Some Examples of Commercially Available Compact Transmitting Loops for the Radio Amateur

Type	*Diameter (m)*	*Conductor Diameter (mm)*	*Frequency Range (MHz)*
AMA-7	3.4	32	1.75-8
AMA-2D	1.7	32	6.9-16
AMA-10D	1.3	32	7-22
AMA-3D	0.8	32	13.5-30
AEA Isoloop	0.89	Flat Strip 1.5 × 38.1	10-30
MFJ Super High-Q	0.91	26.7	10-30

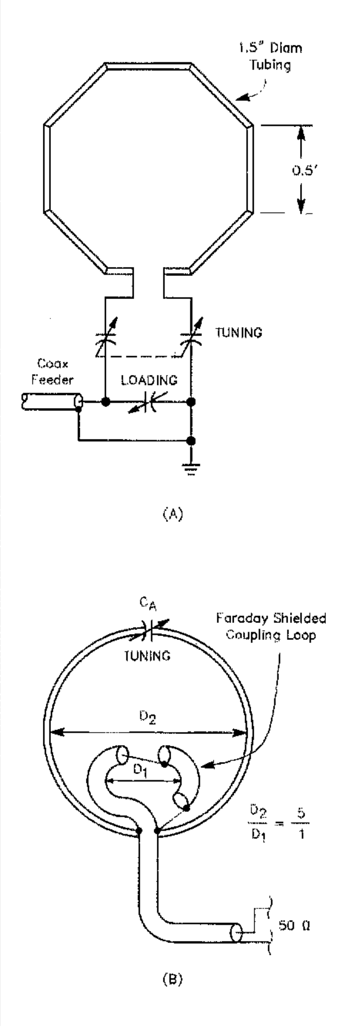

Fig 1—Equivalent circuit of compact loops, showing two methods of tuning and matching. At A, original Patterson loop. At B, AMA loop with coupling link.

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[1] Notes appear on page 40.

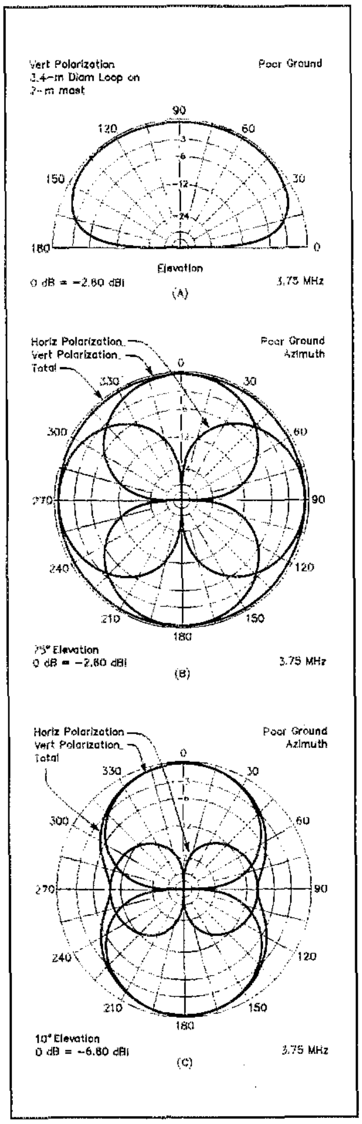

Fig 2—Elevation and azimuth plots for a 3.4-m-diameter vertically polarized hexagonal loop mounted 2.5 meters (height to center of loop) over poor ground, with a conductivity of 3 mS/m and a dielectric constant of 13. The loop is in the 0° to 180° plane. The elevation pattern at 0° azimuth is shown at A, where the radiation is mainly going straight up. B shows the total azimuthal pattern for the loop at a takeoff angle of 75°, plus the horizontally polarized and the vertically polarized field components, which vectorially add to create the total pattern. At this high angle of elevation, the two contributory fields are almost identical, but rotated in azimuth by 90° from each other. C shows the total azimuthal pattern, plus the overlaid horizontal and vertical patterns, for an elevation angle of 10°. Now the vertical field dominates the plot.

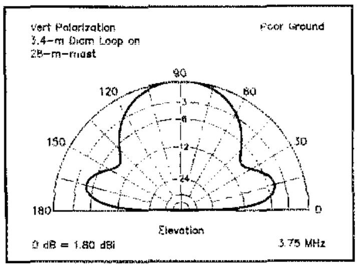

Fig 3—Elevation plot for a 3.4-m-diameter loop at 3.75 MHz, oriented vertically at 28 m in height above poor ground (conductivity of 3 mS/m and dielectric constant of 13). Much of the signal is directed upwards at high elevation angles, but some is radiated at lower angles as well.

trically small loops has been analyzed using simple equations based on standard classical analysis. Equations can be found in various texts for the radiation resistance, conductor loss resistance, and loop inductance.[8] But modern computational electromagnetic codes developed for wire antennas, such as *MININEC*[9] or *NEC*, can be used to calculate efficiency, gain and pattern. The effects of the tuning, feeding and mounting methods on gain, pattern and the bandwidth of the antenna can also be calculated with these programs.

As an example, consider a 1-meter-diameter hexagonal loop using a 2.54-cm aluminum tube as a conductor. Wire models using *MININEC* code must be made up from straight wire sections; therefore, round loops must be modeled as regular polygons. For this loop, the free-space gain is −2.88 dBi at 10 MHz; −0.22 dBi at 14 MHz; 1.14 dBi at 21 MHz; and 1.42 dBi at 30 MHz. The free-space gain of a dipole antenna is 2.15 dBi, and therefore the loop's gain at 10 MHz is 5 dB below that of a dipole. The gain increases and the operational bandwidth decreases as the conductor diameter is increased. I will discuss this in more detail later on.

The disadvantages of using small loops are their high currents and narrow bandwidths, a natural consequence of the very low radiation resistance characteristic of physically small antennas. High-voltage capacitors must be used to tune compact loops. This means that the power used must be limited to about 150 watts, unless expensive vacuum or ceramic variable capacitors (rated at 10 to 20 kV) are used. The problem becomes worse at lower operating frequencies. For the 1-m loop above, which has an impedance of 0.088 + *j* 161 Ω at 10 MHz, the loop current for 150 watts of transmitter power is 41 amperes, and the voltage across the capacitor is 6.6 kV rms. Don't touch your loop when transmitting!

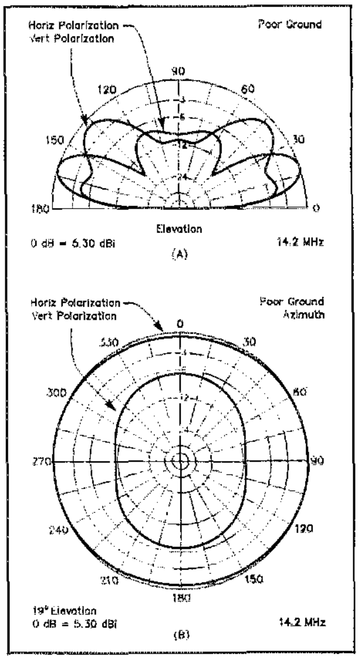

Fig 4—Comparison between horizontally and vertically polarized 1-m-diameter loops placed 15 m over poor ground. The operating frequency is 14 MHz. The horizontally polarized loop has more gain at lower elevation angles than the vertically polarized loop.

The Q of such a small transmitting loop is extremely high (1824 at 10 MHz), and this results in a small antenna bandwidth — only 5.5 kHz. When the loop is fed by a transmitter, the operating bandwidth is twice the antenna bandwidth. For maximum power transfer, the effective source impedance of the RF amplifier is conjugately matched to the load impedance of the antenna. A more practical operational bandwidth is 1.5 times the antenna's 3-dB bandwidth, which for our 1-m loop at 10 MHz is 8.25 kHz. Either way, the bandwidth is narrow, so the loop must be tuned with care. The operational bandwidth of a 1-m loop increases rapidly with frequency, reaching 590 kHz at 28 MHz.

Radiation Patterns of Compact Loops

The desired radiation pattern depends on frequency and path length. For the 160, 80 and 40-meter bands, high-angle skywave is often a requirement, except for those who wish to work mainly DX on these lower frequencies. For bands higher than 10 MHz, low-angle skywaves are a requirement.

Fig 2 shows the elevation and azimuth patterns for a 3.4-m-diameter loop which has been mounted vertically on a 2-m mast over average ground (conductivity 3 mS/m, dielectric constant 13). The patterns were calculated using the *NEC2* computer program, which incorporates a sophisticated Sommerfeld/Norton ground model that takes into account ground interactions at low heights. For an electrically small loop near

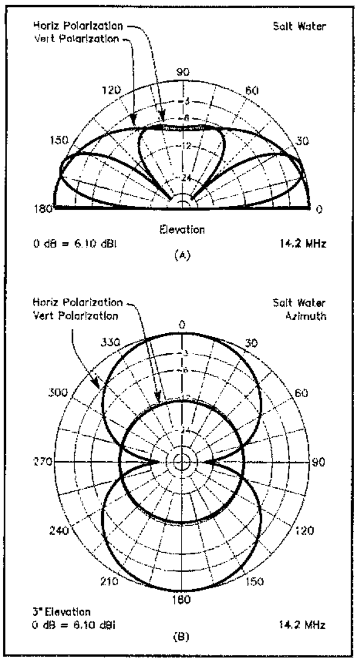

Fig 5—Comparison between horizontally polarized 1-m-diameter loop 15 m over salt water and vertically polarized 1-m-diameter loop mounted on a sailboat transom over salt water. Now the vertically polarized loop has superior gain even at low elevation angles compared to the higher horizontal loop.

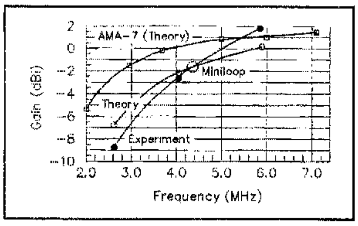

Fig 6—Measured and theoretical NVIS gain for the Antenna Research Associates Miniloop over poor ground, compared with the theoretical gain of the AMA-1 loop (3.4-m diameter).

the ground, with its plane oriented vertically, the total azimuthal radiation pattern (resulting from the combination of the vertically and horizontally polarized fields emanating from the loop) is omnidirectional at high elevation angles. See Fig 2B. The skywave polarization is characterized as being vertical, because the dominant component, the magnetic component, is parallel to the earth's surface.

In Fig 2C, the azimuthal pattern at an elevation of 10° is shown. The overall pattern is no longer omnidirectional, since the vertical pattern dominates over the horizontal pattern. At low elevation angles, the maximum gain is reached when the loop is approximately 0.35 wavelength above the ground. This would require a 28-m tower at a frequency of 3.8 MHz, a height which is not too practical for most hams. Fig 3 shows the elevation plot for a 3.4-m-diameter loop mounted vertically 28 m above the ground. When the conductivity of the ground is finite, the pattern "cut back" at low elevation angles, characteristic of vertical polarization, makes it difficult to launch skywaves at elevation angles less than about 10°, so there is little advantage in using a vertical loop mounted any higher off the ground.

When the loop is mounted horizontally (ie, parallel to the ground), the polarization for the skywave is horizontal, and the azimuthal response is omnidirectional. As with any horizontally polarized antenna, the takeoff angle will depend on the height of the antenna above ground. As the height is increased from 3 to 30 meters (0.14 to 1.4 λ for a 14-MHz loop) the gain increases from 0.5 to 5.5 dBi, and the takeoff angle decreases from 42° to 10°. Fig 4 shows the elevation and azimuth patterns for a 1-m-diameter horizontal loop mounted at a height of 15 m above ground, compared to the patterns for a vertical loop at the same height.

Clearly, the preferred polarization for working DX is horizontal, and the antenna should be mounted as high as possible, as with any dipole or Yagi antenna. The exception is the case of salt water in the direction of propagation, where vertical polarization can be used effectively. See Fig 5A and 5B, where a vertically polarized loop mounted on the transom of a sailboat in sea water is compared to a horizontally polarized loop which is 15 meters high. The vertical loop's response at a 3° elevation angle (close to the horizon) is far better than the much-higher horizontal antenna.

Measured/Predicted Performance of a Compact Loop

The performance of a commercial high power (1 kW) HF Miniloop[10] has been measured[11] on a near-vertical-incidence skywave (NVIS) link, where the distance between transmitter and receiver was 100 km. This loop has a diameter of about 1.6 m, and extreme measures were taken to reduce the conductor loss so that the antenna could be used even at low frequencies: the conductor was made of 10.2-cm aluminum tubing. ARA specifies the operating band to be 2 to 16 MHz.

The measured and theoretical gain for the Miniloop mounted 3 meters above poor ground (conductivity of 1 mS/m, dielectric constant of 13) is shown in Fig 6, compared with the theoretical gain of a larger AMA-7 (loop diameter 3.4 m, conductor diameter 4.8 cm). For reference, the gain of a full-size 4-MHz half-wave dipole at 9.1 m over poor ground is about 4 dBi. Clearly, if a compact loop is used on 75 or 40 meters, the largest diameter loop possible should be used.

The bandwidth of a compact loop depends on its operating frequency and its overall size, as well as on the conductor diameter and the conductor resistance. The bandwidth also depends on the degree of inductive coupling between the coupling loop and the main loop—and of course on the minimum SWR at resonance. Obviously the bandwidth for SWR less than 2:1 cannot be defined if the minimum SWR is greater than 2:1. Fig 7 shows the measured bandwidth (SWR < 2:1) versus frequency for well-matched AMA 3.4-m, 1.7-m, and 0.8-m-diameter loops. Fig 8 shows the theoretical free-space gain for these loops.

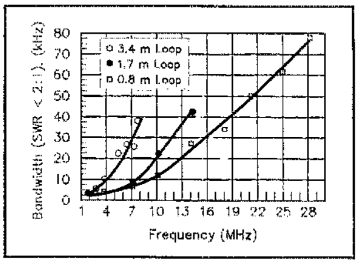

Fig 7—Measured bandwidth versus frequency for three different loop sizes.

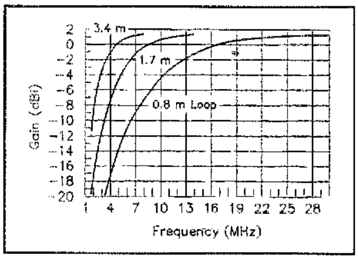

Fig 8—Free-space gain versus frequency for three different loop sizes.

The data shown in Figs 7 and 8 allow you to choose the loop diameter which best meets your operating requirements, including the bands you use most often, and your favorite application (for example, Field Day or base station use). Obviously, the gain of a 1-m-diameter loop at 3.75 MHz (a band of interest to me) is not impressive, but your outlook on loop gain will be highly dependent on the antenna used as a reference. If the reference antenna is a half-wave dipole, a small loop's gain seems poor. But if the reference antenna is a 2.8-m center-loaded mobile whip, the loop's gain is comparable, and a vertical loop is clearly superior for NVIS paths.

Operational Experience

I have carried out a series of experiments to evaluate operationally the performance of several loops. The AMA-7 and the AMA-2D loops were mounted in a vertical plane on pipe masts at the corners of a field laboratory. The height was 3.6 meters, and the comparison antenna was a stagger-tuned droop-

Loops Really Work!

For me, one of the more interesting experiences during September's IARU Region 1 Conference in De Haan, Belgium, was a demonstration by Gaston Bertels, ON4WF, of his home-brew, 2-meter-diameter magnetic loop antenna. Gaston is president of the UBA, the IARU member-society for Belgium.

Gaston's version of the loop antenna uses a conductor made of RG-213 coaxial cable, held in shape by ordinary plumbing parts: eight PVC spokes radiating from a center hub. His tuning capacitor is homemade, and is tuned remotely by a 1.5-volt dc motor designed for a barbeque rotisserie. Unlike the commercial versions of the loop antenna described here, Gaston's use of simple yet highly efficient homemade capacitors does limit frequency coverage; his basic capacitor tunes the loop throughout the 20-meter band, and he plugs in a parallel fixed capacitor, also homemade, to tune the antenna to 40 meters. It is designed for portable operation and can be set up in about five minutes.

Gaston demonstrated the antenna during a lunch break. He set it up in a parking lot, surrounded by cars and buildings, with the bottom of the antenna perhaps a foot above the ground. Using a 100-watt mobile rig on SSB, his first call on 20 meters netted a station in Thessaloniki, Greece. Switching to 40 meters, his first call on that band brought back an Italian station more than 500 miles away. Both contacts were ragchew quality, with signal strengths in each direction well above the noise.

Reading an article about the effectiveness of these small antennas is one thing; seeing a live demonstration is a lot more persuasive!—*Dave Sumner, K1ZZ*

ing dipole for 80, 40 and 20 meters, with an apex height of 15 m. Numerous test QSOs were conducted during daytime on 80 and 40 meters. As shown in Fig 8, the free-space gain of the AMA-7 for these bands is about −3.6 dB and −1 dB respectively, referenced to the comparison dipole, and the 40-meter gain for the AMA-2D is −3.4 dB. By and large, test QSOs confirmed these gain differences. An S unit is about 6 dB, and when comparing antennas with different polarizations and patterns, the signals typically fade differentially on the two antennas.

I took an AMA-6, a 0.8 m diameter loop which tunes 6.7 to 25 MHz, on a camping trip during July and August 1993. I mounted it vertically on a 3-m mast attached to the side of the travel trailer. Initial tests confirmed that it did indeed work on all bands. From a site in the San Fernando Valley in Southern California, schedules were kept with radio amateurs in Toronto, Woodstock, Smith Falls and Ottawa on 20 meters, and in spite of the poor gain of this antenna on 40 meters, QSOs were made with several amateurs in San Francisco, Phoenix and Tucson under daytime propagation conditions. Clearly, the AMA-6 is far easier to install at a campsite than is a dipole. If I had purchased an AMA-13, which is a 0.8-m-diameter loop that tunes from 3.5 to 21.5 MHz, I could have had a bit of fun working the 80-meter band as well.

Concluding Remarks

The Patterson loop was developed for a military tactical communications requirement, where the interest was in NVIS links. Hence the interest was to use the low end of the HF spectrum, between 2 to 8 MHz, and to use high takeoff angles. Notwithstanding limitations on transmitting efficiency and bandwidth, the compact loop can be used with advantage for particular applications. The motor drive for a remotely tuned loop must have no mechanical backlash, and should have a slow fine-tuning speed. A stepper-motor drive is unlikely to prove satisfactory with small loops, since no matter how small the step, you never seem to be able to get the SWR down to exactly 1:1.

Undoubtedly, small transmitting loop antennas will see application for Field Day, camp, maritime communications, apartments, and residential use where antenna height and the "look" of the antenna are factors to consider.

Acknowledgment

I work for the Communications Research Centre, Shirleys Bay, ON, and I wish to acknowledge the use of their laboratory equipment to measure antenna gain and bandwidth. I also want to thank Peter Bouliane, VE3KLO, for making the bandwidth measurements.

Notes

[1] K. Patterson, "Down-to-Earth Army Antennas," *Electronics*, Aug 21, 1967, pp 111-114.

[2] L. McCoy, "The Army Loop in Ham Communications," *QST*, Mar 1968, pp 17-18, 150. (See also *QST* Technical Correspondence, May 1968, pp 49-51 and Nov 1968, pp 46-47.)

[3] Magnetische Kurzwellenantennen "AMA," Christian Käferlein, Weinbergstrasse 5, D-6100 Darmstadt, Germany (tel 49-61-51-61272, fax 49-61-51-663009). An English-language brochure is available.

[4] T. Hart, "Small, High-Efficiency Loop Antennas," *QST*, Jun 1986, pp 33-36.

[5] The IsoLoop 10-30 HF antenna, Advanced Electronic Applications, Inc., PO Box C2160, 2006 196th St SW, Lynnwood, WA 98036. tel 800-432-8873.

[6] The Super Hi-Q Loop Antenna, MFJ Enterprises, Inc., Box 494, Miss. State, MS 39762, tel 601-323-5869, fax 601-323-6551.

[7] B. Battles, "AEA Isoloop Antenna," *QST* Product Review, Nov 1992, pp 71-72.

[8] F. Terman, *Radio Engineering Handbook*, (New York: McGraw-Hill, 1943), pp 35, 52, 53, 814.

[9] For example, I use the *ELNEC* version, by Roy Lewallen, W7EL, PO Box 6658, Beaverton, OR 97007.

[10] Antenna Research Associates Miniloop, see P. Wahi, "HF Miniloop Antennas," *RF Design*, Jan 1992, pp 73-77.

[11] J. Belrose, G. Royer, L. Petrie, "HF Wire Antennas Over Real Ground: Computer Simulation and Measurement," AGARD LS 165 on Modern Antenna Design Using Computers and Measurement: Application to Antenna Problems of Military Interest, Sep 1989, pp 6-1 to 6-30. Available from NTIS, Springfield, VA (Ref NASA Access No N90-17932).

New Products

MAP SOFTWARE

◊ Personal computer users can run software that puts the time, day/night indications, distance between two points, phone numbers in a selected city, and notes on a full-color world map. *Sun Clock* version 4.0, developed by MLT Software of Tigard, Oregon, is available for Macintosh and IBM-compatible PCs. Both support a mouse and the IBM-compatible version runs under *Windows 3.1*. As many as 20 user-embedded readouts show the time in specific places, a notepad, pop-up windows linked to map locations, and a "different day" features lets users see the night shadow on any date. Retail price $39.95. The Mac version requires 1 Mbyte of RAM and comes with *Pyro* and *After Dark* screen savers. The *Windows* version requires 640 kbytes of RAM. Palo Alto Software, 2641 Columbia St, Eugene, OR 97403; tel 503-683-6161.

LOW-LOSS COAX

◊ Amateurs looking for an alternative to Hardline for VHF/UHF/SHF applications might want to try AIRCOM Plus cable, manufactured by Germany's SSB-Electronic GmbH. Its loss is claimed to be lower than other cables of similar size, including RG-213. It has a copper foil shield lined with plastic foil to minimize tearing, and 75% copper braid over the foil to increase mechanical stability. The center conductor is supported by a continuous plastic spreader and embedded in plastic to prevent corrosion and to keep it from moving when it's bent, to maintain nominal impedance. The specially developed N connector provides strain relief and is compensated for better return loss at frequencies above 3 GHz. Distributors: In the US, SSB Electronic, 124 Cherrywood Dr, Mountaintop PA, 18707, tel 717-868-5643, fax 717-868-6917. In Canada, Manfred Zielinsky, VE3ZIE, MAS Enterprises, Import-Export UHF Technik, 104 King St S, St Jacobs, ON N0B 2N0, Canada; tel 519-664-1273, fax 519-664-3082.

Under the Hood II: Resistors

Resistors are one of the most basic building blocks in our communication gear—but do you know one when you see one? Here's what they look like, and how they're built to do what they do.

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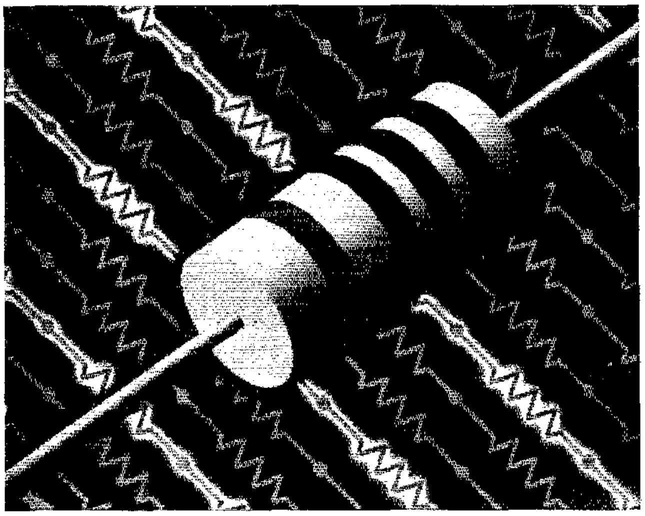

(Art by NU1N)

Resistors are so common in electronic gear that their basic job may surprise you: They waste power by converting electrical energy into heat! But resistors don't just uselessly throw energy away. They allow us to insert *resistance*—opposition to current flow—into circuits to control how current or voltage is used. Thus, a peek into almost any electronic gadget reveals resistors doing many jobs, including: reducing, dividing and equalizing voltages and currents; controlling gain; setting bias and feedback levels; setting impedance levels; loading amplifiers, oscillators, switches and transmission lines; and discharging capacitors and setting time constants. Resistors may even serve as coil forms or fuses!

In a given piece of ham gear, most of the resistors are *fixed*: their resistance is not adjustable. Our radios and allied gear also include *variable* resistors intended to be user- or factory-adjusted. Here's a look at resistor basics.

Resistor Construction

The fixed resistors common in our gear come in three basic types: carbon composition, film (carbon or metal) and wire wound.

Carbon Composition

A "carbon-comp" resistor consists of a rod of resistive material with leads attached to each end (Fig 1). The resistivity of the core, which is made from carbon granules mixed with a resin and silica binder, is determined by carbon particle size and density. An insulating jacket, made of the same resin and binder used in the core (minus conducting carbon particles) provides mechanical stability.

Film Resistors

The resistive element in these resistors consists of a thin film of resistive material on an insulating rod. Silver cement holds lead-equipped end caps to this structure (Fig 2). The type of film and how it's deposited on the rod define the resistor's characteristics. Ceramic-metal (*cermet*) designs use relatively thick film for durability and survival in environmental extremes.

Change the thickness and/or composition of the film, and you change the resistor's resistance. Because these factors can be varied only so precisely, this approach is generally acceptable for resistance values up to 1 kΩ or so. Higher precision and/or higher resistance values can be obtained by cutting a spiral groove, which lengthens the resistance path, through the film until the resistance reaches the desired value . A protective coating of resin, epoxy or ceramic, marked to indicate the resistor's characteristics, completes the part.

Wire-Wound Resistors

Wire-wound resistors come in where considerable power dissipation (generally, 3 watts and more) is required. The basic wire-wound resistor (Fig 3) consists of a resistive wire or ribbon wound around an insulating

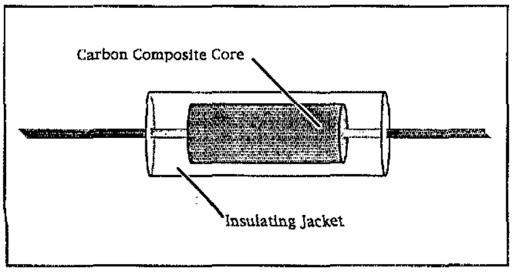


Fig 1—Beneath its colored bands and insulating jacket (usually colored a medium brown or dark green) a carbon-composition resistor contains a leaded resistive core.

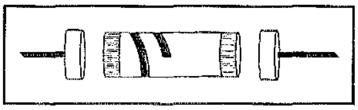


Fig 2—Details of film resistor construction, shown with end caps and leads removed from the resistor body. A spiral groove cut through the film and into the core trims the resistance to its final value. After spiraling, the resistor body and end caps are coated with resin, epoxy, or ceramic and labeled with the resistor's value.

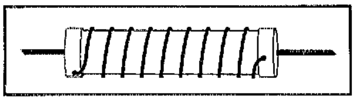


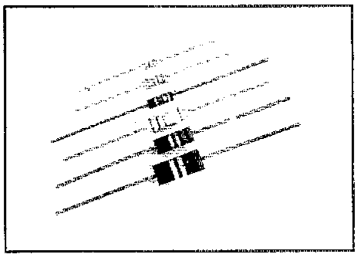
Fig 3—A single-layer spiral of high-resistance wire forms the basis of many wire-wound resistors. After winding, the resistance wire is usually encapsulated in ceramic, enamel or a similar hard, heat-tolerant substance.

Fig 4—This "noninductive" wire-wound-resistor design uses two resistance wires wound in opposite senses and connected in parallel. A disadvantage of this design is that the highest resistance value that can be obtained with a given finished size is halved because of the parallel connection.

Table 2
Value Color Code for Carbon Composition and Film Resistors

Color	*Significant Digit*	*Multiplier*	*Tolerance*
Black	0	10^0	—
Brown	1	10^1	±1%
Red	2	10^2	±2%
Orange	3	10^3	±3%
Yellow	4	10^4	±4%
Green	5	10^5	—
Blue	6	10^6	—
Violet	7	10^7	—
Gray	8	—	—
White	9	—	—
Gold	—	0.1	±5%
Silver	—	0.01	±10%
No color	—	—	±20%

Example: A carbon-composition resistor marked with brown (1), black (0), red (10^2) and gold is a 1-kΩ, 5%-tolerance part.



Wire leads connect these fixed-value resistors to the outside world; color bands indicate their resistance values. As a general rule, the larger the resistor, the more power it can dissipate. This selection includes resistors covering a span of types and ratings, from a ¼-watt, carbon-film unit (top) to a 2-watt, carbon-composition unit (bottom). *(photo by Kirk Kleinschmidt, NT0Z)*

rod. Wire-wound resistors commonly contain a single-layer spiral of high-resistance wire. "Wire-wounds" intended for use above a few kilohertz may use inductance-canceling windings (Fig 4) or multilayer bobbins.

A high-power wire-wound resistor may be brought into tolerance during manufacturing by monitoring its resistance during winding; the wire is then welded to the end caps. Lower-power, precision wire-wounds may be trimmed (upward in value) by sandblasting their wires after end caps have been attached. A protective coating electrically insulates the windings and helps distribute heat away from the resistance wire. Precision wire-wound resistors may be baked for several days after encapsulation to stabilize their resistance.

Wireless Resistors?

The resistor packages I've described so far all have wire leads, yet a look inside many new radios reveals resistors that don't. Such *surface-mount* devices are now standard in many electronics manufacturing applications, as I described in "A Surface-Mount Primer" (December 1990 and January 1991 *QST*). Surface-mount fixed resistors now account for over 50% of the market for all fixed resistors!

Resistor Characteristics

Standard Resistance Values

Resistors are available with resistances ranging from about 0.01 Ω through 10^{12} Ω—ten million million!—ohms. Most resistor types are available in standard values. The resistance values selected for low-power carbon composition and film resistors are such that adjacent values have relative ratios of about 10% for 5%-tolerance types and 20% for 10%- and 20%-tolerance types. Adjacent standard values for precision metal-film resistors differ by approximately 2%.

Tolerance

At purchase time, a resistor's marked and actual resistance must reasonably agree. How far a part's actual value can acceptably deviate from marked value is known as *tolerance*. A resistor's *purchase tolerance* is the permissible resistance deviation from the marked value at the time of purchase.

Standard tolerances for carbon composition resistors are 5%, 10%, and 20%. For example, a resistor marked as 47 kΩ, 10% tolerance, should have an actual resistance of 47 kΩ ± 4.7 kΩ—in other words, anywhere in the range from 42.3 kΩ to 51.7 kΩ—at the time of purchase. Carbon-film resistors are typically available at 5% tolerance. Wire-wound and metal-film resistors are available in precision (1% or less) tolerances.

Labeling

Many wire-lead resistors add considerable color to circuit boards because they carry color bands that indicate their values. Figs 5 and 6 show the standard Electronic Industries Association/Military (EIA/MIL) resistor color code scheme used on many carbon composition, film, and wire-wound resistors.

To read a resistor's color code, begin with the band nearest one end. The first two bands indicate one digit each; the third, a *multiplier*; the fourth, tolerance (20% if this band is absent); and (sometimes) a fifth band indicates other information. Variations in the width and additional bands identify other resistor attributes; for example, a wide white

Table 1
Typical Characteristics of Common Resistor Types

	Tolerance (%)	Temperature Coefficient (ppm/°C)	Power Rating (W)	Values (Ω)	Operating Temperature (°C)	Low Noise	Stability	RF Use	High Voltage Use	Relative Cost
Carbon Composition	10	1200	1/4	10-100 M	–40 to +100	—	—	✓	✓	$
Carbon Film	5	500	1/4	1-100 M	–40 to +125	✓	✓	✓	✓✓	$
Metal Film	0.5	50	<1	0.1-5 M	–55 to +125	✓✓	✓✓	✓✓	✓	$$
Precision Wire-Wound	<0.5	<10	<1	10-301 k	–55 to +150	✓✓	✓✓✓	—	✓	$$$
Power Wire-Wound	5	50-250	5-200	0.1-10 k	–55 to +200	✓✓	✓✓	—	✓	$$
Power Metal Film	1	50	15	0.1-5 M	–55 to +200	✓✓	✓✓	✓	✓	$$
Surface Mount	2	200	1/4	1-10 M	–55 to +125	✓✓	✓✓	✓✓	—	$

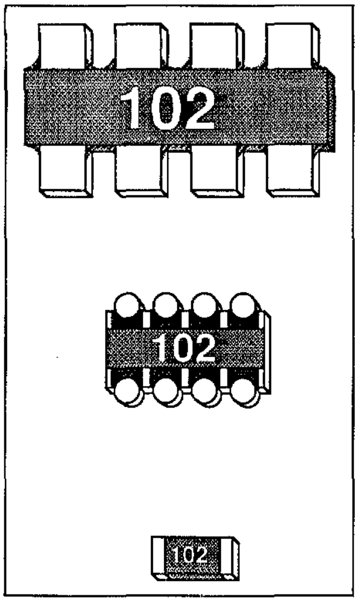
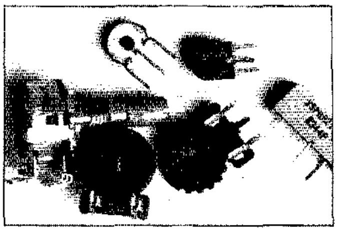


Fig 5—Resistance value markings for leaded (top) and unleaded (middle) surface-mount resistor networks; and individual resistors (bottom). Each of the upper two units consist of four unconnected 1-kΩ resistors. The bottom part is a single 1-kΩ resistor.

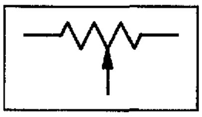
Variable Resistors



(NU1N photo)

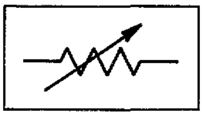
Many resistor applications require *variable* resistance. Like fixed resistors, variable resistors are available in many styles that span a wide range of electrical characteristics. Depending on their construction and intended application, variable resistors are classified as potentiometers (*pots*) or rheostats.

A *potentiometer* (A) acts as a variable *voltage divider*, varying its end-to-wiper resistance according to a mathematical function (linearly or logarithmically, for example) commonly referred to as *taper*. Potentiometers are commonly used for volume, tone and power-output controls. *Trimmer* potentiometers, by comparison, are not as precise, handle less power, and are generally not intended to be adjusted often. You'll often find them mounted on transceiver PC boards, where they're used for set-and-forget adjustment of, for example, CW sidetone level, power output and mixer or modulator balance.



(A)

A *rheostat* (B) is a two-terminal variable resistance. Any potentiometer can be used as a rheostat (one end terminal unconnected), but a rheostat can't work as a potentiometer.



(B)

Rotary or linear motion may set the resistance of a variable resistor. Rotary controls may be *single-turn* (adjustable over about 270° of a circle) or *multiturn* (for precise resistance control). Linear-motion controls include adjustable wire-wound power resistors and the slide pots so common in portable audio/radio gadgets.—*NU1N*

band after the tolerance band denotes solderable (as opposed to weldable) leads.

Precision metal-film resistors are usually marked with four-digit codes. The first three digits denote a value, and the last digit is the multiplier: for example, *1372* indicates 13.7 kΩ. For values less than 100 Ω, *R* signifies a decimal point; for example, *51R1* indicates a 51.1-Ω resistor. Compared to the EIA/MIL color-band scheme, the four-digit code supports an additional significant digit of information.

Surface-mount resistors are labeled with a three-digit coding scheme, in which 100 denotes 10 Ω; 102, 1 kΩ; 105, 1 MΩ, and so on. That is, the first two digits correspond to the basic resistor value, and the third digit is the multiplier. The multipliers are based on powers of 10; for instance, a multiplier of 0 corresponds to 10^0, which equals 1, so a resistor labeled *100* is 10 Ω (10 × 1). A *2* therefore signifies a multiplier of 100 (10^2); a 5, 100,000 (10^5).

Power Dissipation

Carbon-composition resistors are available at ratings up to 2 watts; wire-wound and metal-film resistors, at considerably higher ratings. Wire-wound resistors are the usual choice for power dissipations greater than 2 watts.

How much power a resistor can *actually* dissipate depends on its construction and the temperature and state of the surrounding air. A resistor capable of handling ¼ W in still 70 °C air may have to be *derated* to ⅛ W at 95 °C. (Relatedly, moving the air with a fan can raise the power rating. This is why our 100-W radios commonly include fans.) Nearness to other heat-producing components may require further derating.

Operating a resistor "too hot" can permanently change its resistance or cause it to fail open. Different resistor types tend to respond to extreme overload in predictable ways. Carbon-composition resistors may bake and smoke before failing; flameproof wire-wound resistors tend to open cleanly.

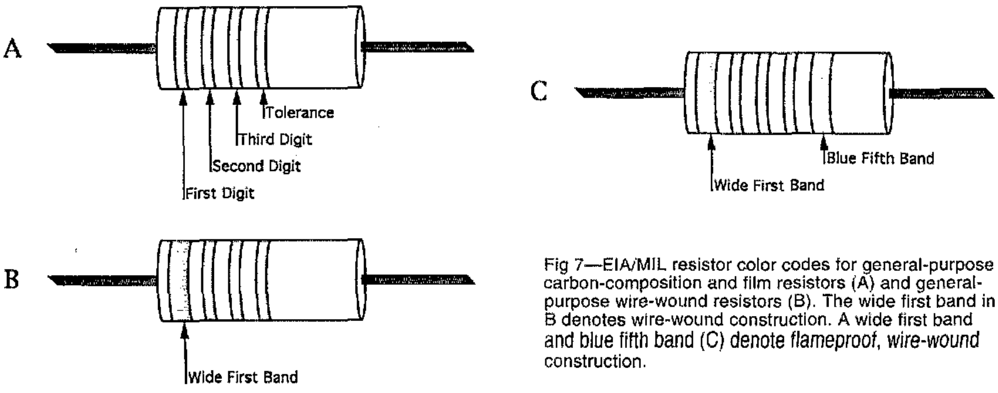
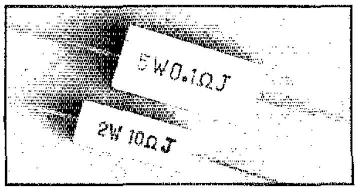


Fig 7—EIA/MIL resistor color codes for general-purpose carbon-composition and film resistors (A) and general-purpose wire-wound resistors (B). The wide first band in B denotes wire-wound construction. A wide first band and blue fifth band (C) denote flameproof, wire-wound construction.



Ceramic blocks and heat-resistance cement package these fixed-value power resistors. Borrowing from a value-marking system we'll study when Under the Hood investigates capacitors, the *J* on these parts indicates a tolerance of 5%. *(NU1N photo)*

Carbon-film resistors, which are very intolerant of overloads, open so cleanly and quickly that they can be used as fuses.

Some electronic parts catalogs include *derating curves* with their resistor listings to help you determine which power rating to specify. One rule of thumb long applied by ham constructors is to operate resistors at half (or less) of their rated dissipation.

Like most electronic components, resistors don't last forever. Resistor *load life*, the number of hours at which a resistor may dissipate rated power under specified operating conditions, is primarily a function of operating temperature. Lower operating temperatures extend load life; higher operating temperatures decrease it.

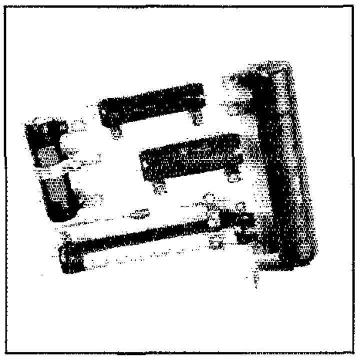
Maximum Applied Voltage

You may be surprised to learn that resistors also have *voltage* ratings. For example, carbon-composition resistors rated at 2, ¼ and ⅛ watts commonly have maximum applied voltage ratings of 750, 250, and 150 volts, respectively. (Application-specific resistors are available for use at voltages exceeding 100,000!) The penalty for overvoltage can be total failure or permanent resistance change.

A resistor's voltage rating can sometimes be more important than its power rating. If, for instance, you need to install 3 megohms in series with a voltmeter to monitor a 1000-V line, Ohm's Law calculations suggest (applying the "half or less power" rule) that three seriesed 1-megohm, ¼-watt resistors would do the trick. In this case, each resistor would dissipate only a tenth of a watt. Assuming that each resistor's voltage rating is 250, though, the voltage rating of our three-resistor string is only 750. For long-term accuracy in a metering circuit, 2-watt resistors (total rating, 2250 volts) wouldn't be overkill in our 1-kV circuit.

Stability

Resistor values shift somewhat with age and stress, including extreme temperatures, moisture, voltage and/or current overloads, and soldering heat. Wire-wound resistors are the most stable, followed by metal-film, carbon-film, and carbon-composition resistors. Whatever the type used, its stability must be sufficient for the expected conditions.



These fixed-value and adjustable power resistors can dissipate powers from 2 to 50 watts. The adjustables' sliding contacts them to be used as potentiometers or rheostats. *(photo by Kirk Kleinschmidt, NT0Z)*

Noise

All conductors, including resistors, generate electrical noise—small random fluctuations of voltage or current. Some types of resistors generate more noise than others. Carbon-composition resistors are noisiest; wire-wound resistors are usually quietest. Resistor noise may be of little importance in many hobby applications, but designers of sensitive amplifiers and instrumentation must take it seriously.

Frequency Effects

A resistor may exhibit different ac and dc values. This is so because resistors are not always purely resistive. Depending on its design, a resistor is also *reactive*—that is, it acts like a coil and/or capacitor to some degree. Because the reactance of coils and capacitors changes with frequency, a resistor's effective value also changes with frequency.

This characteristic is often important in Amateur Radio applications, especially at higher frequencies. Up to at least 30 MHz, carbon-composition, carbon-film and metal-film resistors generally perform comparably. Standard wire-wound resistors, which tend to be inductive at audio frequencies and capacitive at radio frequencies, exhibit the poorest high-frequency performance in terms of reactance. This is why connecting five 10-Ω wire-wound power resistors in series doesn't make a good 50-Ω dummy antenna for Amateur Radio frequencies. Even if their dc resistance totals exactly 50 Ω, their reactance makes an SWR meter read high.

Resistors: Simply Important

So common are resistors in our electronic gear that the simplicity of their basic job—wasting electrical power as heat—may seem paradoxical. But the importance of resistors in electric and electronic technology goes far beyond their presence in our radios. The next time you crank up your stereo, toast a piece of bread or turn on an incandescent lamp, think of what life might be like without resistors!

New Books

THE COMPLETE AMATEUR TELEVISION WORKBOOK

Assembled By Mike Stone, WB0QCD

Published by QCD Publications, 1492 Highway 30, Clarence, IA 52216-9335. Second edition, 1993. Softcover, 8½ × 11 inches, 216 pp. $18.95 plus $1.50 s/h.

Reviewed by Steve Ford, WB8IMY
Assistant Technical Editor

If you're looking for the definitive ATV reference, this may be it. The 2nd edition of *The Complete Amateur Television Workbook* is 216 pages of ATV information distilled into a format that any ham can comprehend. Forty-six pages larger than the first edition, the *Workbook* is a must-have resource for ATVers or anyone considering ATV. (The ample reference section at the back of the book nearly justifies the price by itself.)

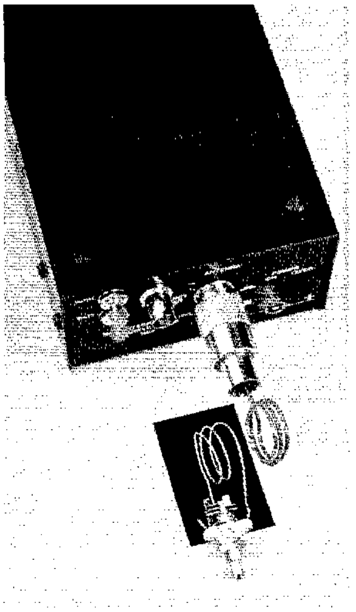
Here's a sample of the diverse information you'll find in this book: Chapter 9 offers complete project diagrams and instructions for an ATV transmitter, a 10-watt amplifier and a receiver downconverter. Is there an ATV repeater or club in your area? Check Chapter 12 and you'll find the most comprehensive list of clubs and repeaters available. If ATV ballooning tickles your fancy, it's discussed in Chapter 13.

Other sections of the *Workbook* are devoted to topics such as choosing the best coaxial cable, TV DX and propagation, worldwide TV standards, antenna polarization, a fascinating discussion of the Alford slot antenna (a horizontally polarized, omnidirectional gain antenna marketed by Mike Stone's company) and more. There's even a section for ATV beginners.

Photographic quality is good and the schematic diagrams are easy to read. The description of the ATV transmitter includes a well-illustrated, step-by-step tutorial on soldering surface-mount components. With the increasing popularity of surface-mount capacitors, resistors and so on, this tutorial is worthwhile for any ham. The *Workbook* is sprinkled with advertisements throughout. Some readers may find this objectionable, but I found the ads valuable for determining the availability of off-the-shelf gear.

The only drawback worth mentioning is the lack of an index. With the expanded size of this second edition, an index would certainly make it easier to find your way around.

If you own the first edition of the *Workbook*, QCD Publications is making an unusual buy-back offer. If your first edition is still in mint condition, you can return it to QCD and buy the second edition at a $5 discount. Not a bad deal at all!



An Accurate Dip Meter Using the MFJ-249 SWR Analyzer

Here's how to add dip-meter performance to several popular antenna-system testers.

By David M. Barton, AF6S
14842 Nelson Way
San Jose, CA 95124

The *dip meter* (often called *grid-dip meter* because vacuum tubes formerly implemented its function) has long been a de rigueur tool in amateur RF work. Not only is a dip meter useful for finding a "tank" circuit's resonant frequency, or an unwanted parasitic resonance, it can also be used to find the electrical length of a transmission line. Modern dip meters are battery operated, making them easy to take up a tower.

But dip meters have always had two major drawbacks. First, they generally aren't very accurate. You're lucky to guess a 14-MHz frequency within 200 kHz on an analog dipper dial, and when you dip a resonant circuit, the dipper's oscillator shifts frequency ("pulls"). You can minimize pulling by minimizing coupling, but this results in a barely discernible dip.

Hams have traditionally augmented dipper operation by using their receivers as calibrated frequency meters. But finding a warbly grid dipper's signal can be tedious, especially when it pulls during the dipping process. The slow tuning rates of modern receivers make this worse—and who wants to take a receiver up the tower, anyway? You could modify your dip meter, giving it a buffered oscillator output for a small frequency counter. That would be fine on the bench, but cumbersome up a tower.

MFJ has inadvertently provided part of the answer. The MFJ-249 HF/VHF SWR Analyzer contains nearly everything you need for accurate dipping: a bandswitched variable-frequency oscillator with built-in frequency counter covering 1.8 to 170 MHz, and an SWR meter. By adding the simple plug-in probe I'll describe, you can use an MFJ-249 as a dip meter![1]

The MFJ-249 as Dipper

Let's look at how the MFJ-249 operates. If you connect a 50-ohm resistor to the MFJ-249, it indicates a 1:1 SWR at all frequencies (ideally, anyway). If you put an inductor in series or parallel with that resistor, the indicated SWR can be anywhere between 1:1 and infinity, depending on the inductor's value, parasitic capacitance, and the measurement frequency.

Suppose you connect an inductor that reads between 2:1 and 5:1 in the frequency range you wish to use. If you couple that inductor to the inductor in a resonant tank circuit, energy extracted by the circuit under test reduces the SWR reading.[2] (Anything that removes energy from a badly matched RF circuit improves the match, lowering the SWR.) So as you tune through the tank's resonant frequency, the meter reading dips—like a dip meter.

With a coil in parallel with a 50-ohm resistor, the SWR decreases as you tune upward in frequency, eventually approaching 1:1. As you tune downward, the SWR approaches infinity. For dipping, the resting meter indication must not be at either extreme.

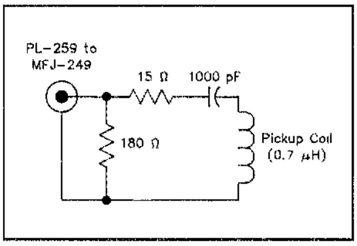


Fig 1—Schematic of the wide-range dipper probe for the MFJ-249 HF/VHF SWR Analyzer.

You can try this yourself with the coils from a conventional dip meter. The coils on my Heathkit Model HD-1250 dip meter have phono plugs. I tested the method with them and a Tektronix through-line 50-ohm terminator for the resistor, via UHF-to-BNC and BNC-to-phono adapters.

The green Heath coil (3 µH) covers 1.8 to 9 MHz. The blue coil (0.7 µH) covers 6.5 to 32 MHz, and the violet coil (0.16 µH) extends from 26 to 90 MHz. But the little 100- to 250-MHz Heath coil did not couple sufficiently. (It doesn't work well on Heath's dip meter, either.)

A Wide-Range Dipper Coil

I was unhappy with the limited frequency range of each coil (although I've had to put up with this dip meter "feature" for years). Unlike its role in a dip meter, the coil doesn't serve as part of the test oscillator in the MFJ-249 system, so one might expect its value to be uncritical. I set up an *Excel* spreadsheet to show the SWR over the whole frequency range, tried various circuits and parts values, and soon discovered a way to cover the entire range with *one* coil!

The circuit, Fig 1, uses a 0.7-µH coil (this value is not critical), a 1000-pF capacitor and two ¼-watt carbon-composition or metal-film resistors (15 and 180 ohms). As *Excel* predicted, the MFJ-249 SWR indication stays almost constant (about 3.6) over the entire frequency range when the probe isn't coupled to a resonant tank. (The probe's track on a 50-ohm Smith Chart is a circle centered on the middle of the chart.)

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[1] Notes appear on page 46.

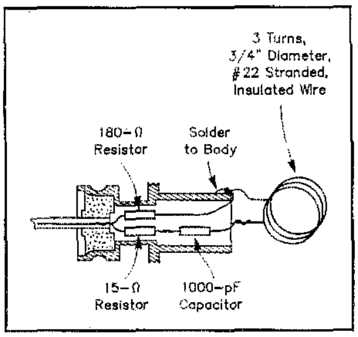

Fig 2—How the probe goes together.

I feared that coupling might be insufficient for a solid dip at the ends of this super-wide frequency range, but that fear proved unfounded. I built the circuit in and on a PL-259 connector, and its dips were weaker than those of my Heath dip meter only at the very low end of my range of interest—below 80 meters. An open-ended length of RG-58 coax with a small loop at the near end gave a strong dip at its 130-MHz quarter-wave resonance. You could increase the coil inductance to favor the low end of the range a bit more—presumably at the price of reduced performance at the high end.

Another worry was that the MFJ-249's oscillator might pull. It does, slightly, but you can always read the frequency at the bottom of a dip. As a plus, the MFJ-249 never exhibits the "stretch, then break" phenomenon of a dip meter.

Constructing the Probe

Fig 2 shows the physical construction of my add-on gimmick. I twisted and soldered the two resistors' wires to each other, poked them into a PL-259's pin from the inside, and soldered them to the pin. I folded the 180-ohm resistor's other lead back against the outside of the connector's body and soldered it. Then I cut the 15-ohm resistor's remaining lead and both leads of an axial-lead 1000-pF ceramic capacitor to ¼ inch, soldered the capacitor to the resistor, and soldered an 8-inch length of #22 insulated, stranded hook-up wire to the capacitor.

I slipped heat-shrink tubing over the resistor and capacitor and wound the wire three times around my finger before soldering its other end to the PL-259's body. Three small strips of tape and a tie-wrap completed the assembly.

If you have the parts in stock, you can build the gimmick in less time than you've spent reading this. That effort will get you a better dip meter than any you can buy. It's portable, battery-operated, accurate and completely self-contained, and you never have to change coils. With it, you even avoid the necessity of carrying another instrument up the tower, because you can pull off the probe and use the MFJ-249 in its intended role as an SWR meter.

So if you've been coveting the MFJ-249 but couldn't quite justify its expense, now you can. And sell that old dip meter while it's still worth something!

How to Dip a Dipole

You can dip a coax-fed dipole by plugging it right into the MFJ-249. Sometimes, though—when setting up parasitic elements in a beam antenna, for instance—you may want to dip a dipole that isn't coax-fed. According to dip-meter lore, you can dip a dipole by inserting a small pickup loop at its center. The problem is that the loop's inductance detunes the dipole.

To avoid detuning the dipole, the loop itself must be resonant at roughly the dipole's resonance frequency. You can accomplish this with either a series or parallel trimmer capacitor (Fig A). The loaded Q of the loop must be low, so small errors in its tuning won't affect the dipole being measured. For low Q in the parallel connection, the inductor's reactance must be higher than the dipole's radiation resistance—say, over 100 ohms. If you choose the series circuit, make the inductor's reactance less than 50 ohms.

Either way, you must tune the loop when it is *not* connected to the dipole. The series circuit requires a removable shorting strap for tuning. For optimum coupling, the pickup-loop diameter should be similar to your dip meter's coil diameter. At 2 meters, a hairpin loop ½ inch long and ¾ inch wide works fine in the parallel circuit. At lower frequencies, you can use multiturn loops to bring the inductance into range.

Form the loop and its trimmer capacitor to fit the dipole's terminals. Remove it and adjust the trimmer for resonance at the test frequency. Then reinstall it and dip the antenna.

The low Q and initial tuning guarantee low error, but you can do even better. Follow the above procedure and record the dipole's resonance frequency. Then remove the pickup loop and readjust it to exactly the same frequency. Now reinstall the loop on the dipole and dip once more.—*AF6S*

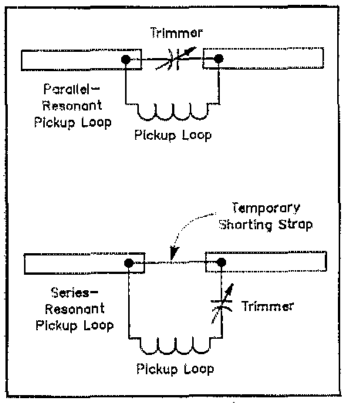

Fig A—If you can't plug your antenna into the MFJ-249 to check its resonance by measuring its SWR, the MFJ-249/dipper can help—if you equip the antenna under test with a resonant coupling loop as shown here.

Notes

[1] The technique and coupling probe described also work with the '249's superseded MFJ-247 sibling, and with the AEA SWR-121 HF Antenna Analyst. The MFJ-247 and -249 simulate the *feel* of a classical dip-meter somewhat more closely than the AEA device because they include tuning knobs and analog SWR meters, but all provide a useful dip meter *function* when used with Dave Barton's probe.—*Ed.*

[2] This requires magnetic coupling between the dipper and the circuit under test. This is easy to achieve merely by bringing the dipper coil close to the coil in the circuit under test if that coil is a solenoidal or hairpin inductor. Dipping a *toroidal* tuned circuit is problematic, however, because a toroidal coil has little external field and therefore affords insufficient coupling for a good dip. As shown in Fig 69 on page 2-35 of the 1993 *ARRL Handbook*, link coupling can allow you to dip a toroidal tuned circuit—but see Charles J. (W7XC) Michaels's "Beware the One-Turn Loop!" (Technical Correspondence, *QST*, Oct 1986, p 50) for a caution on how this technique can introduce error.—*Ed.* QST

Strays

◊ As part of its Field Day preparation, San Angelo (Texas) ARC Activities Chairman Dave Dobbins, KB5UBW (l) requested Mayor Dick Funk (center) to proclaim Amateur Radio Week, and at the urging of the city's emergency management division, presented club president Glenn Miller, AA5PK, with a plaque in appreciation of the club's "many hours of volunteer service to the citizens of San Angelo."—*thanks, ARRL West Gulf Division Assistant Director Noel Johnson, KE5NO*

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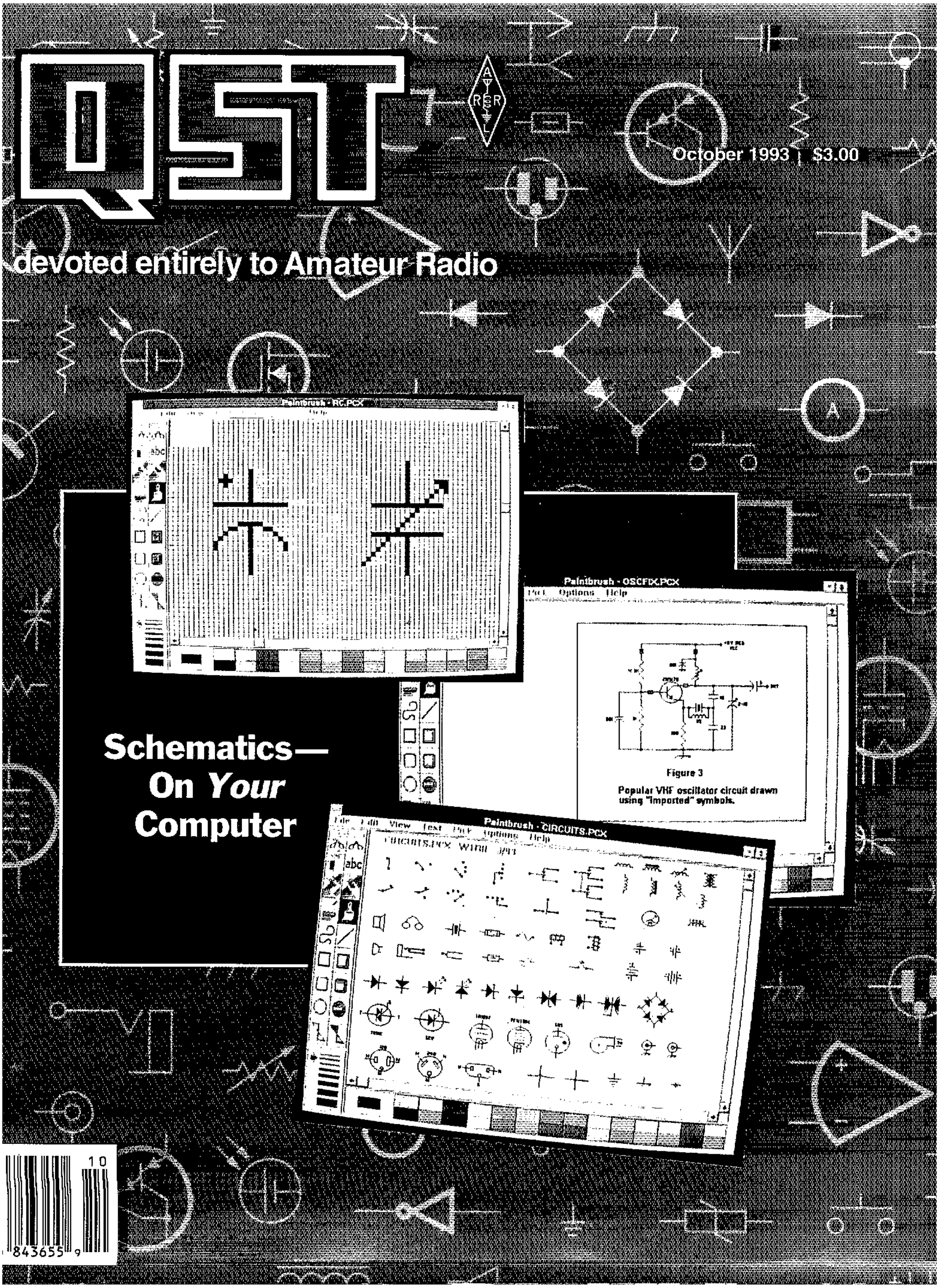

Figure 3

Popular VHF oscillator circuit drawn using "imported" symbols.

QST

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OUR COVER

Now there's another practical use for your personal computer: creating precise schematic diagrams painlessly with Microsoft Windows Paintbrush! See the article by Ken Schofield, W1RIL, on page 39. *(photos by Kirk Kleinschmidt NT0Z)*

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October 1993
Volume LXXVII Number 10

TECHNICAL

NEWS AND FEATURES

NEW HAM COMPANION

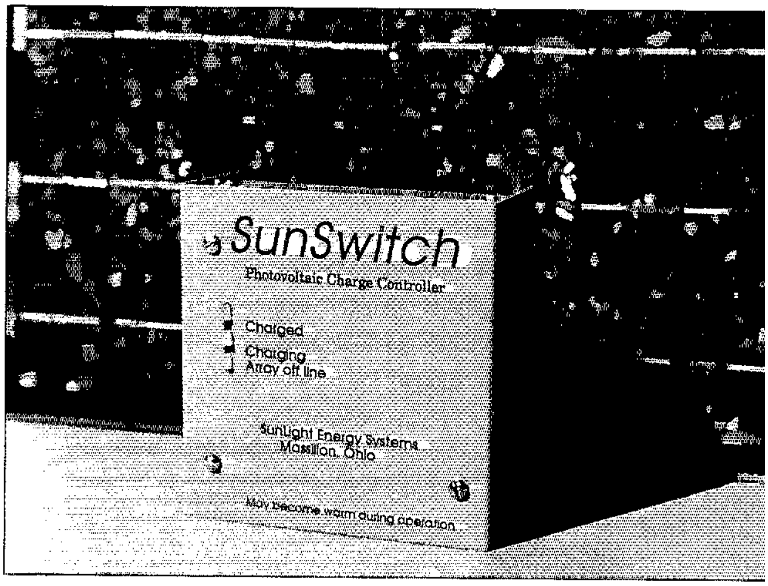
OPERATING

DEPARTMENTS

The SunSwitch

Here's a simple, low-cost and easy-to-use charge controller you can build during part of a rainy day. Then, when the sun comes out, you'll be ready to charge and protect your batteries!

By Michael Bryce, WB8VGE
2225 Mayflower NW
Massillon, OH 44646



(photo by Mike Bryce, WB8VGE)

Many people enjoy powering their radio, boat, RV or cottage by the lake using solar power. The appearance of the popular FET Charge Controller[1] triggered a wealth of correspondence from people asking for a smaller, easier-to-build unit. My fellow QRPers asked for a charge controller small enough to carry in a backpack and one that didn't discharge their batteries at night. In answer to these requests, I developed the SunSwitch. I designed the SunSwitch so you can also charge batteries using a commercial ac-line-operated power supply.

A General Description

Small enough to fit in the palm of your hand, the SunSwitch, like its bigger brother, uses two power MOSFETS to control the flow of array current from a solar panel to a battery. A series controller, the SunSwitch uses the same pulse charging as its sibling. With the two FETs specified, this controller handles array currents of up to 6 amperes—that's enough to tackle two 48-W photovoltaic (PV) panels.[2] If you don't plan on controlling that much current, you can build a 3-A version and save a few dollars. Or, you can easily increase the SunSwitch's current-handling capacity, as you'll see later.

Three LEDs monitor circuit operation and keep you informed of what's going on. A four-position terminal block makes it a snap to connect the PV array and battery. Because it's easy to build and adjust, this may be the perfect *first* project for newcomers to building. A wide part-selection tolerance makes this project junk-box friendly. At $40—even less with the help of your (or a friend's) junk box—the SunSwitch is affordable. To make things ultra convenient, PC boards and complete kits are available. And, if you don't like to bend and drill metal, a punched and silk-screened enclosure is available, too.[3]

How It Works

Refer to Fig 1. The battery under charge supplies the current to run the SunSwitch. In standby mode, the SunSwitch draws 15 mA. D6 provides reverse-polarity protection for the ICs. U4 has two functions: (1) It serves as the SunSwitch's voltage regulator; (2) in case of operator error (such as mistakenly hooking the array to the **BATTERY** terminals), the regulator protects the ICs from array overvoltage. You can use the SunSwitch in a 24-V system: U4 keeps the circuit supply voltage at 12 and the ICs happy.

Because the normal battery voltage of 12.6 is insufficient overhead for the 78L12 input, U4's output runs about 10.5 volts, but that's more than ample for the CMOS chips. When the battery is charging and its terminal voltage is near 14.3, U4 supplies a steady 12 volts to the ICs.

A second regulator, U1, serves as a stable 8-volt reference for the voltage comparator, U2. Capacitors connected to U1's input and output lines ensure stability.

Battery voltage is sampled by a 20-kΩ trimmer potentiometer R6 (**STATE OF CHARGE**), which sets the battery's state of charge. U2A, half of an LM393 dual-voltage comparator, monitors the battery's terminal voltage. U2 pin 2 connects to the 8-V reference. R10 pulls the output of U2A high. When the battery voltage is low, and while the battery is charging, U2A's output is pulled low. This low is applied to U3B. C10, at the output of U2A, keeps the chip stable.

As the battery charges, its terminal voltage increases. When the battery voltage increases to the charger's off point (14.3 V), U2A switches states and pin 1 goes high. A small amount of this output is fed back to the comparator's input. A 470-kΩ trimmer potentiometer, R12 (**HYSTERESIS**) adjusts the circuit hysteresis which provides the basics for pulse charging. (See the sidebar, "What's Hysteresis?") When the battery voltage reaches the off point, FETs Q1 and Q2 are turned off. The battery voltage then begins to fall. But—because of the hysteresis—the comparator won't switch the charger on until the battery voltage drops to a *preset* value, lower than the turn-on point set by R6. When this occurs, the comparator turns on the FETs and the cycle repeats. Then, the battery charging is pulsed on and off. Pulsing is controlled by the **HYSTERESIS** control setting, the rate of charge and the battery capacity.

U2B monitors the PV array output. When the PV array is active, a small current flows in the array's negative lead. U2B detects the current and its output switches states from low to high. At night, when there's no array output, U2B's output is low. This low causes DS1 (**PV READY**) to glow when the array is inactive. During battery charging, U3C pin

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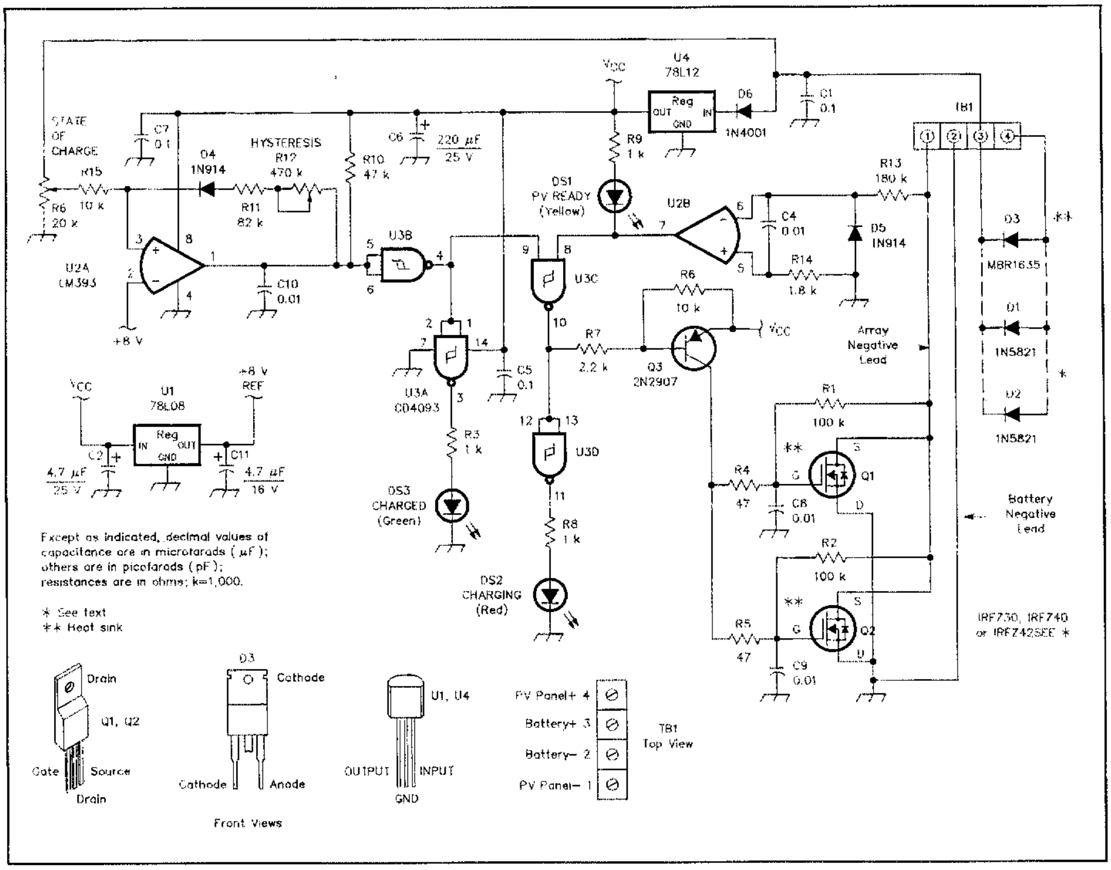
[1] Notes appear on page 27.



Except as indicated, decimal values of capacitance are in microfarads (µF); others are in picofarads (pF); resistances are in ohms; k=1,000.

* See text
** Heat sink

Fig 1—Schematic of the SunSwitch circuit. Part numbers in parentheses are Mouser or Active Electronics; equivalent parts can be substituted. Unless otherwise specified, resistors are ¼-W, 5%-tolerance, carbon-composition or film units. Mouser Electronics, 2401 Hwy 287 N, Mansfield, TX 76062, tel 800-346-6873, 817-483-4422; fax 817-483-0931. Active Electronics, 11 Cummings Park, Woburn, MA 01801, tel 800-677-8899, 617-932-4616; fax 617-933-8884. Mouser and Active Electronics have several locations throughout the US.

D1, D2—1N5821 30-V, 3-A rectifier (Mouser #333-1N5821).
D3—MBR1635 35-V, 16-A Schottky diode (Mouser #584-18TQ045 or Active Electronics #MBR1635).
D4, D5—1N914 silicon switching diode.
D6—1N4001 50-PIV, 1-A diode.
DS1—Yellow rectangular LED; see text (Mouser #ME3516231).
DS2—Red rectangular LED (Mouser #35BL510).
DS3—Green rectangular LED (Mouser #ME3516221).
Q1, Q2—IRFZ30 50-V HEXFET (Mouser #584-IRFZ30; see text) or 584-IRFZ40; Active Electronics #26058.
Q3—2N2907 PNP switching transistor.
R6—20-kΩ multiturn pot (Mouser #594-64W203).
R12—470-kΩ trimmer pot (Mouser #324-CP10H-470K).
U1—MC78L08, 8-V, 100-mA three-terminal positive voltage regulator (Mouser #333-78L008AP).
U2—LM393 dual voltage comparator (Mouser #511-LM393N).
U3—CD4093 quad two-input **NAND** Schmitt trigger (Mouser #570-CD4093BE).
U4—LM78L12 12-V, 100-mA three-terminal positive voltage regulator (Mouser #333-78LO12AP).
Misc: Terminal block (Mouser #506-4PCV-04), heat sinks (Mouser #567-7-199BA), enclosure (see Note 3).

10 is low. This low turns on Q3, which drives FETs Q1 and Q2. FET gate drive is 10 V.

A Complete Charging Sequence

When the sun comes up, the solar-panel array starts producing current. This is sensed by U2B, which switches states and outputs a high at pin 7. DS1 goes dark. Because the battery voltage is low, the output of U2A is also low and its output is inverted by U3B.

U3B's output at pin 4 is routed in two directions: (1) It's applied to pins 1 and 2 of U3A and inverted again. Because a high is applied to its inputs, the output is low and DS3, the **CHARGED** LED, remains dark; (2) U3B's output also goes to U3C pin 9. Since we now have highs on both inputs of U3C, the gate switches states and the result is a low at pin 10. This low is inverted by U3D. The result is a high, which lights DS2, the **CHARGING** LED.

U3 pin 10 supplies a low to the base of Q3. The transistor saturates and supplies gate voltage to Q1 and Q2. The FETs conduct and connect the negative lead of the PV array to the negative side of the battery.

When the battery is almost fully charged, U2A pin 1 goes high. This high is inverted by U3B. U3A sees this as a low and outputs a high to illuminate DS3, the **CHARGED** LED.

Because U3C pin 9 is now low, U3C pin 10 goes high, turning off Q3 and removing gate drive from the FETs. DS2, the **CHARGING** LED, goes dark because U3D pin 11 is no longer high.

With the FETs off, the battery voltage starts to drop. Because of the hysteresis provided by R12, the voltage can be adjusted to drop almost 1.5 V before the comparator switches states. When the battery voltage does drop below the hysteresis point, the comparator switches and the cycle repeats.

As the battery charge nears 100 percent, DS2 and DS3 blink on and off as the SunSwitch sends current pulses to the battery.

Towards dusk, the current pulses begin to lengthen. Finally, DS2 remains on. When U2B no longer detects output from the PV array, its output goes low. This fires DS1, the **PV READY** LED, and turns off the FETs by way of U3C pin 8. At dusk, DS2 extinguishes and DS1 lights.

Even though the FETs have been turned off, they still allow current to flow from the batteries back to the array. D3, an MBR1635 16-A Schottky diode, prevents this nighttime current leakage. D3 also protects the FETs in the event you reverse-connect the array to the SunSwitch.

A Word About Power MOSFETS

For your SunSwitch, you have a wide selection of suitable FETs from which to choose. IRFZ30s can handle a drain current of over 50 A (more than three times the maximum this circuit can handle). The $R_{DS(on)}$ is a scant 0.045 ohm. IRFZ30s are inexpensive and easy to obtain. Other good choices are the IRFZ42 and the IRFZ44. The IRFZ44 or IRFZ42 have an $R_{DS(on)}$ of 0.028 ohm. If you plan on running just 1 or 2 amperes of charging current, then an IRF531 or IRF511 can be substituted. Although more expensive and a bit difficult to obtain, the Siliconix SMP60N06 is an excellent power MOSFET for use in the SunSwitch.

Different Switching Methods

Q1 and Q2 are in the battery's negative lead. There is, however, one drawback to this scheme: You can't connect the negative side of the array and the negative side of the battery and then ground them. If you do, you'll defeat the SunSwitch and overcharge your batteries. Fig 2 details where the FETs are located in the circuit, along with the array and the battery. The blocking diode is in the positive side of the array and battery.

Three Amps or Six?

Do you want to build the 3- or 6-A version of the SunSwitch? In the 3-A version, D3, the 15-A blocking diode, is replaced by two 1N5821 Schottky diodes, D1 and D2. If the current you'll be controlling is less than 1 A, you can use two 1N4001 diodes. With the 3-A version, there's no need to use two FETs, so you can eliminate one. The basic operation of either version is the same.

Putting It Together

There's nothing critical about the SunSwitch assembly—you can build it on perfboard, if you choose. Using a PC board (see Note 2), however, is the easiest and neatest way to build your SunSwitch. Take precautions to avoid static build-up and discharge when handling U3 (the CMOS 4093) and, of course, the FETs. I recommend use of a wrist ground strap. Use IC sockets for the chips. Once the devices are mounted on the PC board, they'll withstand a lot of abuse. With all the parts on hand, you should be able to assemble and test the SunSwitch in a few hours.

What's Hysteresis?

To get an idea of what hysteresis is, let's look at the operation of your home heating thermostat. When the thermostat demands heat from the furnace, a set of contacts close. As the house warms, the contacts stay closed until the temperature is a few degrees above the thermostat setting. If the thermostat is set at 68°, the furnace may heat the house to 72°. When the thermostat turns off the furnace at 72°, the house cools (slowly, we hope). The thermostat allows the house to reach, say, 66° before calling on the furnace to start heating the house once again. If it weren't for this hysteresis, the furnace would oscillate on and off at the desired 68°. The thermostat's action is exactly what the hysteresis control imitates for the SunSwitch. By adding hysteresis, we keep the SunSwitch from oscillating at the battery's turn-off point.—*WB8VGE*

Start by stuffing the PC board with the IC sockets and then the resistors. Install U1 before you mount C2 and C11; it's a tight fit there! Install U4 and Q3. Don't forget the wire jumper (W1) near Q3. Finish the PC board by mounting the terminal block.

Note: If you're going to use the 15-A Schottky diode at D3, you must follow one of these mounting methods to avoid shorting PC-board traces: *Don't drill through the PC board to mount the TO-220 case and heat-sink combination.* Attach the diode to a heat sink with #4-40 hardware; be certain to use thermal compound between the diode and heat sink. Bend the diode leads toward the PC board. While holding the heat sink level on the PC board, solder D3's leads to the board. Next, place a small square of double-stick foam tape beneath the heat sink. Push the heat sink onto the foam tape. That holds the heat sink to the PC board without the need to drill any holes. Otherwise, you *must* use insulating hardware (such as nylon) to secure the diode/heat-sink assembly.

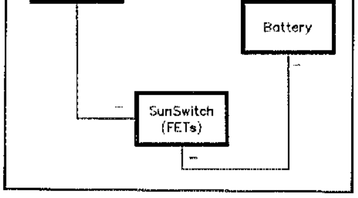


Fig 2—A block diagram of a typical SunSwitch system. The blocking diode is physically mounted on the SunSwitch PC board.

Bend the leads of one FET 90°, apply some thermal compound to its heat sink and bolt the combination to the PC board in the Q1 position. Follow the same procedure for Q2. Solder both FETs in place.

Testing and Adjustment

To adjust the SunSwitch, you'll need a variable-voltage power supply, a battery and an accurate digital VOM (you can use an analog VOM, but you'll lose some resolution). Set the power supply to 14.3 V. Adjust R6 and R12 to midrange. Connect the power supply to the battery terminals on TB1. DS1 (**PV READY**) should light. DS3 may be on; DS2 should be off. If it's on, you've either got a solder bridge at the 4093 or LM393 pins, or one (or both) of the chips is defective.

Check for +12 volts at U3 pin 14. Check for +8 V at U2 pin 2. Monitor the voltage at U2 pin 3 while adjusting R6, the **STATE OF CHANGE** pot. Verify that the voltage increases and decreases as you rotate R6's control shaft.

With the power supply still set to 14.3 V, adjust R6 until DS3 glows. Drop the supply voltage to 12, then *very slowly* increase the voltage to 14.3. At that point, DS3 should light again. (You may want to repeat this step, as you'll encounter the effect of the hysteresis circuit.)

For the rest of the testing, use a small battery. I use a 6.5-Ah lead-acid battery. Connect the battery to your SunSwitch. Adjust the power supply for an output of 16 to 19 V. Connect the power supply to the TB1 **ARRAY INPUT** terminals and turn on the power supply. DS1 (**PV READY**) will extinguish and DS2 (**CHARGING**) will light. If possible, monitor the charging current as well as the battery voltage. When the battery's terminal voltage hits 14.3, DS3 should light and DS2 extinguish. With DS2 off, the current going into the battery *must* be zero! If not, you may have an oscillating FET, or a solder bridge in the FET area.

If the current drops to zero with DS2 off, the SunSwitch sends current pulses to the battery. The pulse length is determined by the charge current, battery condition, and, of course, the setting of the **HYSTERESIS** control, R12. (Don't worry, the setting of R12 is not critical.)

If all's fine so far, slowly reduce the power-supply voltage. At about 13 V, DS1 lights and DS2 goes dark. This completes the testing and adjustments. If everything works as outlined, you're ready to hook the SunSwitch to your solar panels!

Hooking It Up

First, connect the battery to the SunSwitch, then connect the array. Charging is fully automatic and requires no action on your part. At dawn and dusk, DS1 and DS2 may light simultaneously. This is normal, as the comparator monitors the output of the array. With a full 6 A of array current, the heat sinks become "politely" warm to the touch.

The full state-of-charge voltage of 14.3 is about right for just about everyone. However, if you have a high charge current (6 A) and a small-capacity battery (1.2 Ah), you may want to increase the set point a bit. Likewise, if you're charging a 105-Ah battery with only 600 mA, you'll need to lower the set point to 14 volts or so.

You can also use the SunSwitch to charge small lead-acid batteries. Use your variable-voltage power supply as the current source. Set the power supply output for 16 to 19 V and connect it to the **ARRAY** positions on the terminal block. It's best if you use a current-limiting power supply—one that can safely limit its output to a preset value. (A really dead battery draws *all* the current it can handle! You could toast the power supply if the current drain is excessive.) The SunSwitch is happy with currents of up to 7 A.

Uh-Oh! It Doesn't Work!

The most common failure of the SunSwitch is sloppy soldering around U1. It's close quarters there and a solder bridge is easy to make. If DS3 lights, but current is still flowing into the battery, check for a solder bridge near the FETs. When DS3 is on, the base of Q3 must be high. Ensure that you installed C10, it keeps U2 stable.

If there's no charging current flowing when DS2 is on, check for +10 V at the FET gates. Work your way back toward Q3. Did you remember to install the jumper wire (W1)? Did you install the blocking diode (D3) backwards?

If DS1 doesn't extinguish, check for the proper placement of D5. Also, it's possible to have both DS1 and DS2 on at dusk when the SunSwitch is connected to a solar panel. Long runs of too-small-gauge wire between the SunSwitch and the array may keep DS1 on. The SunSwitch will still function if this happens. Mistakenly swapping the 1.8-kΩ and the 180-kΩ resistors is quite easy to do.

Part Substitutions

I understand that not everyone has junk boxes or part bins overflowing with the exact parts for every project. The SunSwitch is quite tolerant of part substitutions. Your local Radio Shack outlet can supply almost every part you need. You can use IRF510 or IRF511 FETs at Q1 and Q2 for currents of 1 to 3 A, if you don't mind them running a bit hotter than the FETs specified.

If you can't obtain the 8-V regulator, an LM78L05 5-V regulator can be substituted. In fact, you can use just about *any* regulator in place of the LM78L08 as long as its output voltage is below 9 V and you can squeeze it into the available board space. Also, just about any PNP switching transistor will work at Q3.

If you can't find—or just want to eliminate—the **HYSTERESIS** pot, use a 470-kΩ pot and adjust the hysteresis to your liking. Then, remove the pot, measure its value and install a fixed-value resistor in its place.

You don't need to use the commercial heat sinks listed, either. They fit the board nicely, but a bent-up piece of aluminum works just fine. If you mount the two FETs and the diode on the heat sink, *insulate the diode from the heat sink*. The cases of the diode and the tabs of the FET are electrically hot. Don't allow any other wires to touch them.

A ready-to-go case is available for the SunSwitch. Because the case requires a different method of mounting the board, you can't use round LEDs; they won't fit the case—it's designed for rectangular LEDs. Solder the LEDs to the board from the *foil* side, spacing them about ¼ inch from the board. Because the case has silk-screened panel labels, the LEDs can all be of one color, although the color variations are attractive.

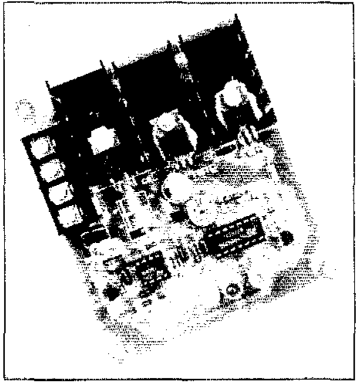
Increasing SunSwitch Current Capacity

Three factors limit the current capacity of the SunSwitch: (1) The capacity of the heat sinks to keep the active devices cool; (2) the PC-board traces carrying the current to and from the battery; and (3) the FETs. The PC-board traces can carry a little over 7 A before they become rather warm. To increase the current-carrying capacity of the SunSwitch, you'll need to reinforce the current-carrying PC-board traces with some solid copper wire; #14 solid wire works well. Four places need to be beefed up—they're the main circuit traces. They're easy to spot, as they go to the FETs and back to the terminal block. By doing this—and increasing the heat-sink capacity—the SunSwitch can handle up to 14 A of array current, the current-handling limit of the blocking diode, D3.

If you choose to use the IRF510 or IRF511 FET instead of the IRFZ30, you'll not be able to draw 14 A, as these FETs won't handle the current without meltdown.

Odds and Ends

Although the SunSwitch is designed for use with 12-V lead-acid batteries, you can use it to charge wet pocket NiCds, too. A wet pocket NiCd requires up to 16 V or so before it's fully charged. Simply adjust the SunSwitch to reflect the higher turn-off voltage. You can also use the SunSwitch in a 24-V lead-acid system. Increase the value of R12 to, say 50 kΩ, and adjust the state of charge to your needs.



The SunSwitch PC board. *(photo by Kirk Kleinschmidt, NT0Z)*

Summary

The next time you're looking for a charge controller that fits in the palm of your hand, controls 6 A (or more) of array current and can be assembled in half a day, look no further than the SunSwitch!

Notes

[1] M. Bryce, "FET Charge Controller," *QST*, Jan 1992, pp 45-50. See also "Charge Controller Comments and Catches," *QST*, May 1992, Technical Correspondence, pp 95-96 and Feedback, same issue, p 96.

[2] For periods of less than one hour, an unmodified PC-board-based SunSwitch can handle over 8 A. PC-board trace size is the limiting factor.

[3] Kits for the SunSwitch are available from SunLight Energy Systems, 2225 Mayflower NW, Massillon, OH 44646. A complete parts kit including the 3½ × 4-inch PC board (but not the enclosure): $40 plus $2.50 shipping. A punched, drilled and silk-screened enclosure: $12. PC board only: $8.50 plus $1 shipping. PC boards are also available from FAR Circuits, 18N640 Field Ct, Dundee, IL 60118.

A PC-board and part-overlay template is available free from the ARRL. Address your request for the BRYCE SUNSWITCH PC-BOARD TEMPLATE to: Technical Department Secretary, ARRL, 225 Main St, Newington, CT 06111. Please enclose a business-size SASE.

New Products

PREMIUM COAX

◊ To achieve better results on HF, VHF and UHF without the expense of Hardline, consider The Wireman's X-4XL series coaxial cable, available with cutoff frequencies up to 16 GHz, 20+-year weather-resistant polyethylene jackets with integral vapor seal, 95% tinned copper braid and 100% bonded aluminum shields. The cable is designed to eliminate moisture problems associated with semi-air dielectric cable, and its low-loss characteristics make it an ideal transmission line for UHF work.

X-4XL series cable is available in a variety of sizes. Models CQ1008 (solid center conductor) and CQ1009 (stranded) are 0.590-inch OD and have loss characteristics similar to ½-inch Hardline.

Special N connectors are available. Prices range from $0.70 to $1.50 per foot. Press Jones, N8UG, The Wireman Inc, 261 Pittman Rd, Landrum, SC 29356; tel 800-433-WIRE (9473) or 803-895-4195, fax 803-895-5811.

Inside the Grounded-Grid Linear Amplifier

Operating a linear amplifier is a pretty simple matter. But the technical details of what happens *inside* amplifier tubes and impedance-matching networks can be mysterious. Let's unravel those mysteries.

By Warren Bruene, W5OLY
7805 Chattington Dr
Dallas, TX 75248

To many Amateur Radio operators, linear amplifier operating characteristics are mysterious. To help you better understand amplifier operation, I'll discuss the factors that determine the optimum anode (plate) load resistance and the functions of the RF output network, using a grounded-grid linear amplifier as an example. I'll also discuss some modern concepts for tuning and loading adjustment.

Ham stations have gravitated toward the use of a 100-watt-class transceiver followed by a grounded-grid linear amplifier. Grounded-grid amplifier operation is a bit more complex than grid-driven operation because of the effects that "fed-through" drive power cause in the input circuit. Tube operation is easiest analyzed with the aid of an "idealized" set of tube curves. These curves, shown in Fig 1, roughly approximate those of the Eimac 3-500Z. The anode-current lines are spaced so that there is no odd-order intermodulation distortion.[1] In this tube, the minimum idling current for distortion-free operation is 0.1 A. The curves are drawn so that zero bias is required for a dc anode potential of 3 kV. The left ends of the lines stop approximately where flat-topping and excessive grid current occur. The use of these idealized curves simplifies calculations and makes understanding the fundamental principles of operation a bit easier.

The horizontal scale represents the instantaneous anode-to-grid voltage. The vertical scale represents the filament-to-grid voltage. Note that the values are relative to the voltage on the grid, which is zero in this case. The filament is driven negative (upward on the diagram) to increase anode current. The sine wave to the left of the grid represents the filament-to-grid RF drive voltage. Below the graph, centered at 3 kV, is a sine wave representing the RF anode voltage. These two sine waves are in phase when the anode circuit is resonant.

The Anode Load Line

The heavy line extending from point A through point Q and into the current cutoff region is called the *load line*. The locations of points A and Q determine tube operation. Point Q represents the zero-signal or idling point; point A determines operation at maximum linear PEP output. This point should be at the edge of the linear range as indicated by the left ends of the current lines. It should also end at a current value that results in the desired dc anode current, RF power output and anode dissipation.

Input and output sine-wave RF voltages are assumed. (RF circuits with sufficient Q maintain nearly sine-wave voltages in real amplifiers.) Also, the voltages are assumed to be in phase, since the load line is elliptical when the anode circuit is out of resonance. Out-of-resonance operation causes reduced power output and efficiency.

With these curves, we can calculate the amplifier's specific operating characteristics based on these givens:

$E_B = 3000$ V (dc anode potential)

$i_P = 1.2$ A (peak anode current read at point A)

$e_k = -105$ V (peak input voltage read at point A)

$e_P = 2500$ V (peak RF anode voltage [3000-500])

$I_Q = 0.1$ A (idling anode current)

These basic tube-operating characteristic calculations represent a single-tone signal at maximum PEP output in class-B operation:

$$I_B = i_P/\pi = 0.382 \text{ A} \qquad \text{(Eq 1)}$$
(dc anode current)

$$P_O = (i_P\, e_P)/4 = 750 \text{ W} \qquad \text{(Eq 2)}$$
(RF power output[2])

$$P_{FT} = (i_P\, e_k)/4 = 31.5 \text{ W} \qquad \text{(Eq 3)}$$
(fed-through drive power)

$$P_T = P_O - P_{FT} = 718.5 \text{ W} \qquad \text{(Eq 4)}$$
(generated by the tube)

$$P_{dc} = I_B\, E_B = 1146 \text{ W} \qquad \text{(Eq 5)}$$
(dc anode input power)

$$P_{diss} = P_{dc} - P_T = 427.5 \text{ W} \qquad \text{(Eq 6)}$$
(anode dissipation)

$$\text{Eff} = 100\, P_T/P_{dc} = 62.7\% \qquad \text{(Eq 7)}$$
(anode efficiency)

$$R_L = 2\, e_P/i_P = 4167\ \Omega \qquad \text{(Eq 8)}$$
(anode load resistance)

$$R_K = 2\, e_k/i_P = 175\ \Omega \qquad \text{(Eq 9)}$$
(filament-to-grid input-load resistance)[3]

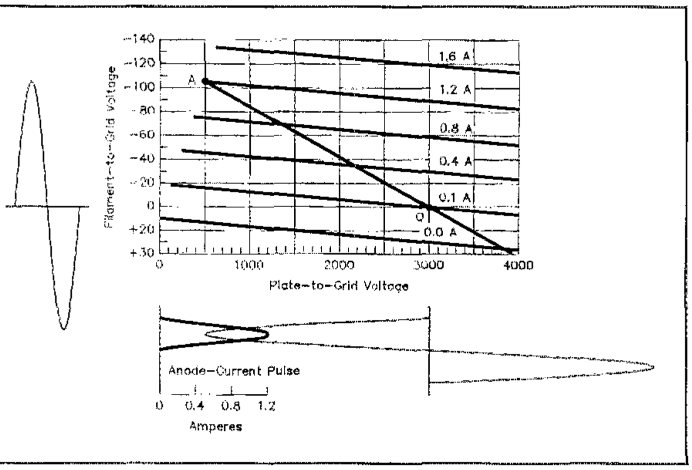


Fig 1—Idealized anode-current characteristic curves for a tube roughly the size of an Eimac 3-500Z. The input-voltage waveform appears at the left side of the graph; anode current and voltage waveforms are shown below it. The heavy line originating at point A and traveling through point Q is called the *load line*.

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[1]Notes appear on page 30.

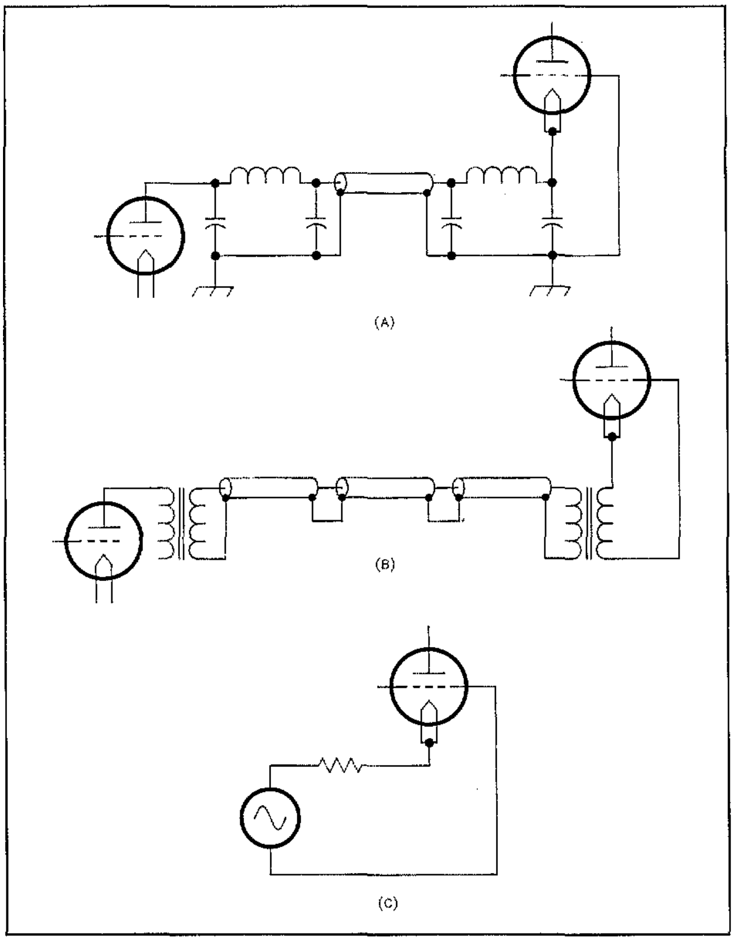


Fig 2—Equivalent driver circuit (at the fundamental frequency). At A is the output network of the driver and the amplifier input network, connected by a length of transmission line. This network can be treated as three lengths of transmission line with transformers at each end (B). From this, the equivalent generator voltage and line impedance can be derived (C).

The anode-current calculation is actually for class-B operation, which would exist if the 0.1-A anode-current line represented zero current instead. The curves represent the real-world case of class AB, however, so the actual dc anode current is slightly higher.

The anode-current pulse shape is shown in the lower left part of Fig 1. Note that the edges of the pulse are flared slightly. The difference between this curve and a true half-sine-wave causes a slight increase in dc anode current. Using a computer to integrate this pulse shows that the dc anode current is actually 0.389 A—7 mA more than the 0.382 A calculated above. Anode dissipation is also higher (by 0.007 A × 3000 V, or 21 W), and anode efficiency is reduced to 61.6%. The simple formulas get us close to the correct values, however.

The amount of anode dissipation with zero signal has only a small effect on the anode dissipation at maximum PEP output, although it contributes most of the anode dissipation over fairly long time periods with normal speech signals, because the average signal level is quite low.

Effect of Moving Point A—Tuning, Loading and Drive

The effect of moving point A can be determined from Fig 1. Moving it upward exceeds the tube's anode current and/or anode dissipation ratings. A downward shift reduces power output. A leftward move enters the flat-topping distortion region, and a shift to the right reduces power output and efficiency. Point A is moved up and down in practice by adjusting the drive level, and is moved right or left by adjusting the loading control (which, as noted earlier, must be followed by retuning for resonance).

Tube Anode Resistance

The effective internal resistance, R_S, of the amplifier is determined by the tube's anode resistance and the effective source resistance of the driver (as seen across the tube's grid-to-filament connection).

The tube's anode resistance, R_P, is determined by the *slope* of the anode-current lines. In Fig 1, find the intersection of 1.5 kV and −40 V, which falls on the 0.4-A anode-current line. Now extend the 0.8 A line to the right until it intersects an extension of the −40 V grid-voltage line (off the graph). This intersection occurs where the anode potential is 6 kV. Thus, with −40 V input, the anode voltage must be increased by 4.5 kV (6 kV − 1.5 kV) to get an anode-current increase of 0.4 A (0.8 A − 0.4 A). The tube's anode resistance is, therefore, 4500/0.4 = 11.25 kΩ in the linear part of the tube's characteristic curves. The tube conducts only the equivalent of half of the time, so the effective R_P, per Eq 8, is 2 × (4500/0.4), or 22.5 kΩ.

Inherent RF-Current Feedback

In a grounded-grid amplifier, all of the anode current in the output circuit also flows through the input circuit. Any change in the anode circuit's load resistance (loading) causes a change in the input resistance. Any internal resistance in the driver allows some variation in the drive voltage. The change is small if the driver resistance is low, but if the driver resistance is high, the voltage change can be nearly as large (in percent) as the resistance change. Fig 2 helps clarify this point. On any given frequency, a lossless RF network can be represented by an ideal transformer and a length of transmission line. The first network is the output network of the driver transmitter, and the second is the input circuit of the grounded-grid amplifier. They are connected with a length of transmission line (Fig 2A).

In the equivalent circuit (Fig 2B), let's assume that the sum of all line lengths is some multiple of 180°, such as 360°. In this case, the input impedance of the three lines in series is always exactly the same as the load impedance because of the half-wavelength transformer effect. The input transformer steps the coax impedance up to the correct load resistance for the driver's output device(s), and the output transformer steps the coax impedance up to the grounded-grid tube's input resistance (175 Ω for the tube in the earlier example). Next, let's eliminate both transformers by normalizing the generator voltage and line impedance to 175 Ω, as illustrated in Fig 2C.

The internal R_S of a driver that uses RF-voltage feedback (such as a Collins KWM-2) is relatively low. Measurements on one transceiver[4] gave approximately 25 Ω, relative to a 50-Ω coax impedance, which translates to 87.5 Ω for our example. Broadband, solid-state transceiver amplifiers probably have R_S values closer to 50 Ω. On the other hand, tetrodes with no voltage feedback have quite high R_S values—typically more than 200 Ω. In this case, it may be better to use a total phase delay of an *odd* multiple of 90° to take advantage of the transmission line's impedance-inverting properties.

The amount of effective RF-current feedback is determined by the effective source resistance, R_S, at the driver tube's anode terminals. If this source resistance is zero (as in a constant-voltage source), there can be no feedback because variations in anode current cannot affect the drive voltage. On the other hand, if the source resistance is very high (like a constant-current source), any anode-current increase (caused by a loading change or tube nonlinearity), causes the input resistance to decrease, reducing the drive voltage in proportion. The effective feedback would be very high and would *completely cancel* the tube's anode-current nonlinearity. Intermodulation distortion at the amplifier output would be only that of the driver!

The reason that we don't take advantage of this feedback technique is that nonlinear grid-current circuit loading is usually a much greater cause of distortion. Grid current (plus screen current, in a tetrode) places a nonlinear resistive load across the grid-to-filament circuit. Grid current rises quite steeply as the anode voltage swings down near the grid (or screen) voltage. This nonlinear grid-current loading causes flat-topping and becomes the major cause of intermodulation distortion at large signal levels.

The effect of this nonlinear grid loading can be reduced by providing a low driver source resistance. This is the reason for the special length of coax specified to connect a Collins KWM-2 to a 30S-1 amplifier, for example. This special length, plus the phase delay in the KWM-2 output network and that of the 30S-1's cathode circuit, approximately equals some multiple of 180° on each band. This provides a low source resistance, which reduces the effect of nonlinear screen-current loading in the 30S-1's 4CX1000A or 4CX1500B tetrode, which is cathode-driven and operates in class AB_1. The length of connecting coax is therefore important for minimizing inter-modulation distortion. This is why a KWM-2 driving a 30S-1 can produce intermodulation products 41 dB down from PEP output. As stated earlier, solid-state transceivers likely have an R_S near 50 Ω. Intermodulation distortion is acceptable, but not as good as it could be. A properly adjusted ALC circuit should always be used to avoid overdriving the amplifier, thereby keeping distortion acceptable.

Adjusting Anode Load Resistance

As stated earlier, the anode circuit should *always* be kept in resonance. The load impedance is correct when the desired single-tone dc anode current and dc grid current, or screen current for a tetrode (as given on the tube's data sheet for typical operating conditions), occur at the same time. The drive level must be adjusted along with loading to reach that point.

Tuning and Loading Detectors

The RF anode voltage is in phase with the RF filament-to-grid voltage when the anode circuit is in resonance. Therefore, a *phase discriminator* can be used to indicate resonance. The discriminator output can feed a zero-center meter to indicate the direction and magnitude of tuning error. The direction of error is independent of signal amplitude, but, of course, its magnitude depends on the signal amplitude. A linear amplifier can be tuned to resonance during voice transmission using such a device.

A similar approach to loading adjustment uses a detector to monitor the correct ratio of RF anode voltage to RF grid-to-filament voltage. In the example, this ratio is 2500/105 = 23.8. This ratio can be monitored by dividing both voltages down to the same level (such as 5 volts) and using a diode detector to produce a positive voltage from one sample and a negative voltage from the other. When adding the two produces a 0-V output, the loading is properly adjusted. Incorrect loading will show either a positive or negative output voltage, indicating which way the loading should be adjusted. This adjustment can also be made while transmitting. Compared to the traditional method of tuning with a key-down signal, this is a much more friendly way to tune an amplifier!

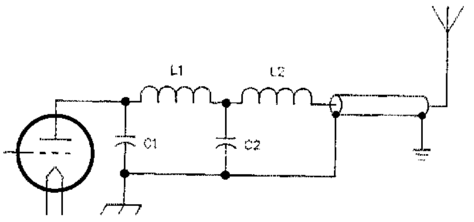


Fig 3—The amplifier's RF-output network matches the antenna's feed-line input impedance to the amplifier's anode load resistance.

RF Output Circuitry

The RF circuitry for transforming the antenna impedance to the correct load resistance (R_L) for the tube typically includes the amplifier's output network and a transmission line. Additional circuitry is often employed to match the antenna to the transmission line, and/or a network to match the coax input impedance to the desired 50-Ω load for a broadband, solid-state transceiver. For simplicity, I'll describe a coupling system consisting of a tuned output network and a length of 50-Ω transmission line connected directly to a vertical antenna.

Fig 3 shows the circuit. The vertical antenna is assumed to have an input impedance of $30 - j40$ Ω, which causes a 3:1 SWR on the transmission line. The transmission line is assumed to have an electrical length of $2\frac{1}{4}$ wavelengths, which results in a coax-input impedance of $30 + j40$ Ω.

RF Output Network

The output network is required to perform several functions:

1) Transform the input impedance, $r_a + jx_a$, to 4167 Ω, which is the anode load resistance in our example. (A practical network must be tunable over the desired frequency range and also over the desired load-matching range, such as 3:1 SWR.)

2) Provide sufficient input-circuit Q (or circulating energy) to maintain a nearly sine-wave voltage across the tube. (The tube generates a pulse of anode current each RF cycle.) A Q of 10 or more is adequate.

3) Provide sufficient harmonic attenuation. The second-harmonic content of the anode-current pulses is approximately 7 dB down at maximum output. The transmitter's second-harmonic output should be at least 50 dB down, which a pi-L network can provide with a large margin.

The tuning and loading functions are normally achieved by varying C1 and C2, respectively. Inductors L1 and L2 are preset for each band by switching taps on the coils. C1 is tuned for resonance, assuring that the tube sees a resistive load. C2 is adjusted in small increments (followed by retuning C1) until the correct load resistance is obtained. When the tuning and loading adjustments are completed, the output network transforms the coax input impedance ($30 + j40$ Ω in our example) to the tube's R_L, 4167 Ω.

If we pulled the plug on our linear amplifier after correctly tuning it, we could connect an RF impedance meter from the tube anode to ground and measure an impedance of 4167 Ω. If we looked back into the output network from the coax output terminal (with the amplifier turned on but with no signal applied) we would observe a very high SWR, as previously reported.[5] Of course, this has no effect on the transmitted signal.

Summary

The operating conditions of a grounded-grid linear amplifier are determined by the anode load resistance and drive-voltage amplitude. The anode load resistance determines the slope of the load line. There is practically no relationship between the correct tube load resistance and the effective internal resistance, R_S, of the amplifier.

Thanks to Dr T. R. Cuthbert and S. E. Bonney, W5PAQ, for their helpful comments during the preparation of this manuscript.

Notes

[1] E. W. Pappenfus, W. B. Bruene and E. O. Schoenike, *Single Sideband Principles and Circuits* (New York: McGraw-Hill, 1964), pp 188-190.

[2] This equation does not take into account output-network loss.

[3] Grid current adds input loading and typically reduces R_K to approximately 70% of this value.

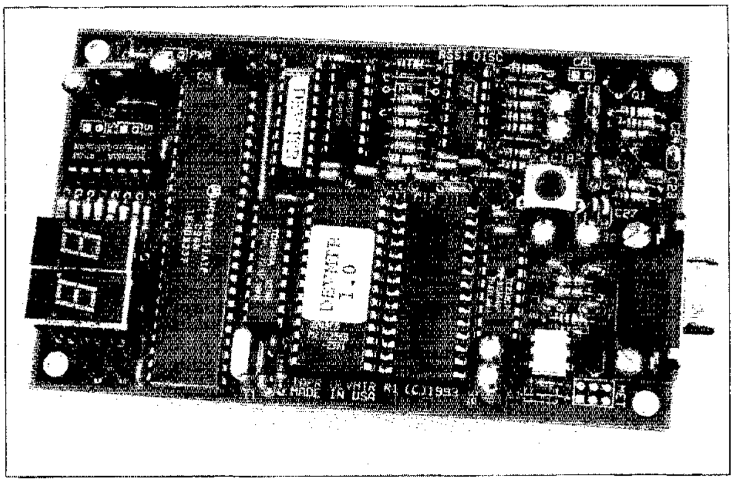
[4] W. Bruene, "RF Power Amplifiers and the Conjugate Match," *QST*, Nov 1991, pp 31-32, 35.

[5] See Note 4.

TAPR's Digital Deviation Meter

Checking your rig's frequency deviation need no longer be a task requiring lots of expensive test equipment. It's now in the do-it-yourself domain!

By Lyle Johnson, WA7GXD
9991 E Morrill Way
Tucson, AZ 85749-9568



(photo by Kirk Kleinschmidt, NT0Z)

Most amateurs lack the equipment to determine FSK or FM-transmitter frequency deviation. As a result, they usually don't think about it, or try to convince themselves that it doesn't matter. "After all, the deviation was set at the factory. And besides, other stations seem to be able to copy me most of the time..."

If frequency deviation (see the sidebar, "What is Frequency Deviation?") were easy to measure, most amateurs would take a moment now and then to verify that their signal deviation is correct. But deviation-measuring equipment typically costs thousands of dollars; most of us either can't afford to purchase it, or don't know someone who can measure the transmitter's deviation for us.

In this article, I'll present an inexpensive, *self-calibrating* instrument that enables you to *easily and automatically* perform deviation measurements![1] It's flexible enough to work with most receivers and is configurable for operation as a community resource. It's an excellent club project brought to you by those folks who were primarily responsible for making packet radio readily available to everyone: TAPR—the Tucson Amateur Packet Radio club (see the sidebar, "What—or Who—is TAPR?").

Why Measuring Transmitter Deviation is Important

An FM signal consists of the carrier and multiple sidebands. For example, transmitting a 2200-Hz tone (the high tone in a 1200-baud AFSK packet system) creates sidebands displaced from the carrier frequency by ±2200 Hz, ±4400 Hz, ±6600 Hz, and so on.

As you increase the deviation, the amount of signal in the farther-out sidebands increases, and the amount in the closer-in sidebands and the carrier decreases. When the FM signal passes through a receiver's IF filter, the farther-out sidebands are eliminated; they fall outside of the filter's passband. If the deviation is too high, there will be too little signal left in the close-in sidebands to properly represent the signal. Worse, the sidebands won't be in the correct proportions to represent the transmitted signal. The result is distortion: The received-signal waveform doesn't look much like that of the transmitted signal, and your packet TNC can't properly demodulate the signal—you get bit errors. So, too much deviation is just as bad as too little deviation; worse, as it may cause interference to stations on adjacent channels.

A shift of two-thirds the data rate for simple FSK results in transmitted spectra that are flat across the top and drop down fairly rapidly through the first sidebands. Narrower shifts result in intersymbol interference (ISI), making it harder for the receiving demodulator, while increased deviation makes the signal wider than necessary. At 9600 bits/s, a shift of 6.2 kHz—a deviation of 3.1 kHz—is about optimum.

For packet satellites, it may be even more critical to have the transmit deviation set correctly. Further, the uplink signal may have to be offset-tuned to accommodate the Doppler shift seen by the satellite.

For these reasons—and more—it's necessary to be able to set your transmitter's deviation to the correct level for optimum performance of your packet or FM voice station.

How Can I Measure Deviation?

Frequency deviation is difficult to measure because it's a variable RF phenomenon and requires an instrument capable of responding to RF to measure it. Because an FM signal is of constant amplitude, simple diode detectors used for field-strength measurements won't help.

The ARRL Handbook contains a simple deviation-meter circuit diagram.[2] It's intended to connect to the output of a receiver discriminator and measures deviation directly. Its accuracy requires that (1) the FM detector is linear and (2) you have a means of calibrating it. Not many hams have the gear necessary to calibrate such a meter. It's less clear whether those who *do* have the equipment *also have the inclination* to calibrate the meter!

What commonly available item can you think of that reflects the deviation of an incoming FM signal? Answer: an FM receiver's detector output. The idea of using an existing FM detector is a seductive one. Cheap scanners—even the ubiquitous weather-cube receivers—have FM detectors that can be pressed into service. An FM demodulator can be used to measure deviation if it (1) has an output that can be related to deviation, and (2) is calibrated, or can be *easily* calibrated and provide meaningful results.

Calibration 101

If you want to calibrate a voltmeter, you must give it a known voltage. If you want to

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[1] Notes appear on page 38.

What is Frequency Deviation?

When an FSK or FM transmitter is keyed without modulation applied, a carrier is transmitted at a single frequency.* At this point, you can: amplitude modulate the signal (including keying it on and off for CW), or vary its frequency to convey information.

If you choose to vary its frequency, the result is *frequency modulation* (FM), or *frequency-shift keying* (FSK). Which it is depends on the nature of the modulating signal and how you apply it. Let's use analog modulation, such as your voice, or a packet, to create an FM signal.

As you vary the transmitted carrier frequency, you create sidebands. The greater the frequency swing of a given audio signal, the greater the number of sidebands you produce and the wider the signal becomes. Thus, a 1-kHz tone may produce a pair of sidebands above and below the carrier, or it may produce several sidebands, spaced at 1-kHz intervals above and below the carrier frequency. The sideband *spacing* is determined by the modulating-signal *frequency*; the number of sidebands is determined by the *modulating-signal amplitude*. Thus, a loud 1-kHz tone may produce many sidebands 1 kHz apart, while a soft 1-kHz tone may only produce two sidebands of low amplitude.

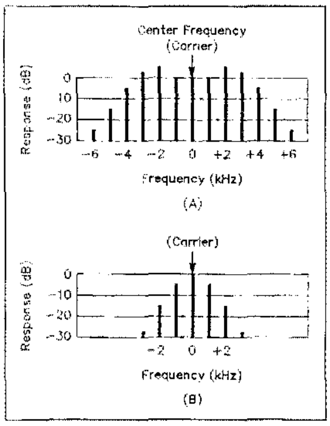


Fig A—A shows the sidebands produced by 3 kHz of deviation produced by a 1-kHz tone. Note the number of sidebands and their spacing from the carrier, and that the sidebands rapidly diminish in amplitude the farther they are in frequency from the carrier. At B we see that reducing the deviation reduces the number of sidebands. Compare this to A and you'll see that this signal is significantly narrower, occupying less spectrum.

Now, let's modulate the FM transmitter by applying a sufficiently strong 1-kHz tone such that the signal swings 6 kHz. We then speak of our signal as having a positive peak deviation of 3 kHz and a negative peak deviation of 3 kHz. (This assumes that our modulating process is linear—which is often not an accurate assumption in the real world). These positive (above the carrier frequency) and negative (below the carrier frequency) peak deviation values are referred to as positive and negative deviation. Likewise, one half the total 6-kHz deviation is equal to the deviation value of 3 kHz.

A mathematical discussion of this interaction is beyond the scope of this article. The bottom line, however, is that sidebands are produced, and that the width of the resulting signal is dependent on both the frequency of the modulating tones and the amplitude of those tones. Finally, deviation measurements are related to *peak* values, not averages.
—*WA7GXD*

*We'll ignore the effects of drift, PLLs, VCO noise and other hopefully minor influences on your transmitter's purity for the purposes of this article.

calibrate a deviation meter, you must give it a signal of known deviation. Here, we'll exploit a feature of common FM detectors.

An FM detector outputs a voltage proportional to the frequency input. The voltage may drift over time, but over the short term, the voltage readings are repeatable. So, if you inject an unmodulated signal offset from the detector's center frequency, you'll obtain a dc voltage that corresponds to this frequency—that is, deviation!

To eliminate the need to know the ultimate transmitter frequency (be it 10 meters or 70 cm), you can use a simple signal generator that operates at the receiver's IF. This signal generator can be stepped through the IF passband in increments of, say, 1 kHz. The dc voltage corresponding to the given signal frequency is then tabulated. Finally, you compare on-the-air signals to your table and interpolate between the 1-kHz readings to get accurate deviation measurements down to 100 Hz or so.

An Overview of the DevMtr

The TAPR Deviation Meter (DevMtr) is a microprocessor-based instrument designed to measure the deviation of an applied signal in an IF range of about 10 to a little over 11 MHz. (The standard FM IF is 10.7 MHz.) The meter is self-calibrating and simple to use. It requires detector-voltage output, a signal-strength voltage (or squelch voltage), a connection to the IF amplifier for the calibration signal, and an 8- to 20-V power source. The DevMtr visually displays deviation readings using a two-digit seven-segment LED readout (kilohertz and hundreds of hertz deviation). It also provides continuously updated deviation readings in the forms of RS-232-C serial data, and a pulse train suitable for directly driving a METCON counter input for remote-site readings.[3] (See the sidebar "Using the Deviation Meter with TAPR's METCON.")

Circuit Description

A detailed circuit description is provided with the TAPR construction/instruction manual, so I'll provide a shortened explanation here.

Fig 1 block-diagrams connections between the five major DevMtr subsections: CPU and serial I/O, signal conditioning, calibration synthesizer, local and METCON I/O and the power supply.

Analog-Signal-Processing Section

The analog-signal-processing section (Fig 2) uses a quad op amp (U7). U7C and U7D act as peak detectors (remember, deviation is a peak-related phenomenon) to momentarily trap the highest and lowest detector-output voltages. These voltages correspond to the positive and negative frequency deviation. U7A buffers the received-signal-strength-indication (RSSI) voltage.

Digital-Signal-Processing Section

Refer to Fig 3. The detector-output voltages are passed to the microcontroller, U3. Five times per second, an internal analog-to-digital converter (A/D) measures the peak-detector and signal-strength buffer outputs. If the received-signal strength is too weak, the detector readings are ignored. If the signal is sufficiently strong, the peak detector voltages are processed and the deviation is calculated.

A simple three-terminal linear regulator, U1, is used in the power supply. A charge pump contained within the RS-232-C interface chip (U11) provides ±9 volts for the op amp and serial port.

DS1 and DS2 display the calculated deviation (Fig 4). For example, a 3.8-kHz deviation of the received signal equates to a reading of **3.8.** This value can be passed to a computer through the serial-port connection of J1 (Fig 3) at 9600 bits/s. On the computer's display, this data appears as **kHz> 3.8.** The **kHz>** prompt indicates the DevMtr is operating in the deviation measurement mode; the **3.8** corresponds to the measured deviation.

Additionally, 500-µsec pulse bursts are gated out of optoisolator U12 at 380 pulses per second. A METCON unit attached to P5 can provide a frequency counter reading of **00380** to show a 3.8-kHz deviation.[4]

Calibration oscillator U6 of Fig 5 is a phase-locked loop (PLL) synthesizer operating at an IF of 10 to 11 MHz. Here's the neat touch: The *only* deviation meter adjustment is L1, the tuning coil for the synthesizer's voltage controlled oscillator (VCO). During alignment, you simply adjust L1's tuning slug for the proper displayed calibration value and the DevMtr is ready for use! U7B buffers and filters the VCO tuning voltage and the DevMtr LED displays it. On the computer, the tuning voltage is displayed along with a **VCO>** prompt.

Every time power is applied to the deviation meter, the calibration oscillator runs. The DevMtr sets the VCO 10 kHz below the IF center frequency. After a short pause to allow the VCO to stabilize, the DevMtr reads its peak detectors, averages them (to remove the effects of any residual FM on the VCO), then stores the reading in a table. The VCO is stepped up 1 kHz and the process repeated.

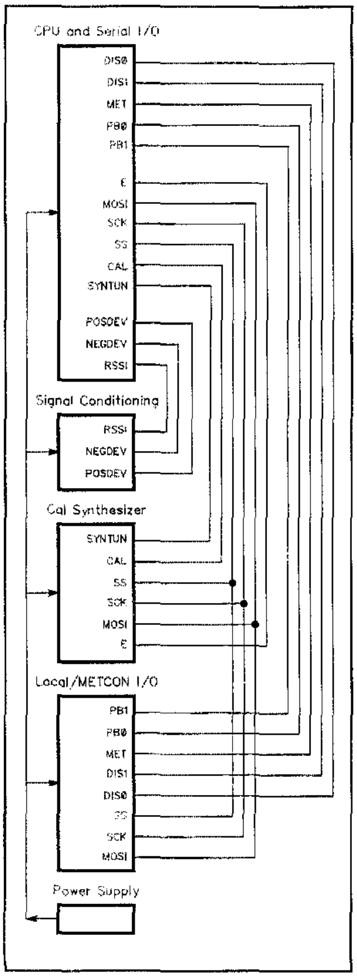


Fig 1—A block diagram of the deviation meter.

What—or Who—is TAPR?

Tucson Amateur Packet Radio (TAPR) is a nonprofit research and development organization with worldwide membership.

Since 1981, TAPR has been at the forefront of developing amateur packet-radio technology. TAPR is probably best known for its development of the TNC 1 and TNC 2, licensed versions of which form the basis of virtually every packet controller available to radio amateurs today.

TAPR's volunteers have been involved in hardware, software and protocol efforts. TAPR helped fund the AO-16 PACSAT, launched in January 1990.

TAPR's quarterly newsletter, the *Packet Status Register,* has carried news of packet-related developments since 1982. All of TAPR's work is done by dedicated volunteers.

How Do You Get a Project Funded by TAPR?

Tucson Amateur Packet Radio (TAPR) has an open policy to support Amateur Radio hardware and software experimentation and development. If you have an interesting project and the qualifications to carry it off, but lack some resources (PC board layout, funding, certain tools, etc), you can apply to TAPR for assistance.

In general, TAPR honors those requests that appear credible within the limits of TAPR's ability to assist. Your providing a detailed project description, how the project applies to packet radio or other digital modes, milestones and timetables, and a verifiable track record in doing similar volunteer efforts will get the attention of the TAPR Board of Directors, who then consider the request. (The DevMtr project, for example, was funded entirely by TAPR.) For more information about TAPR, please write to TAPR at PO Box 12925, Tucson, AZ 85732-2925, or call 602-749-9479.—*WA7GXD*

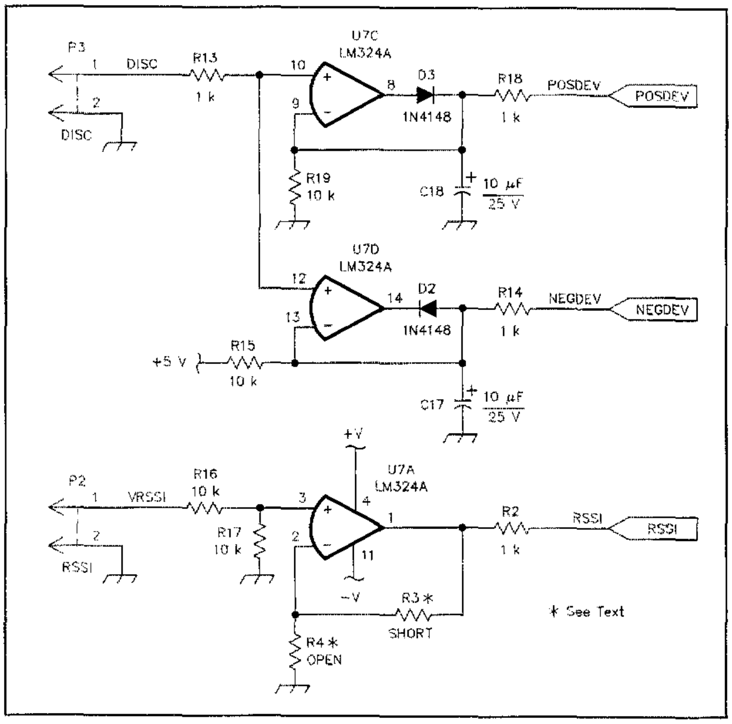
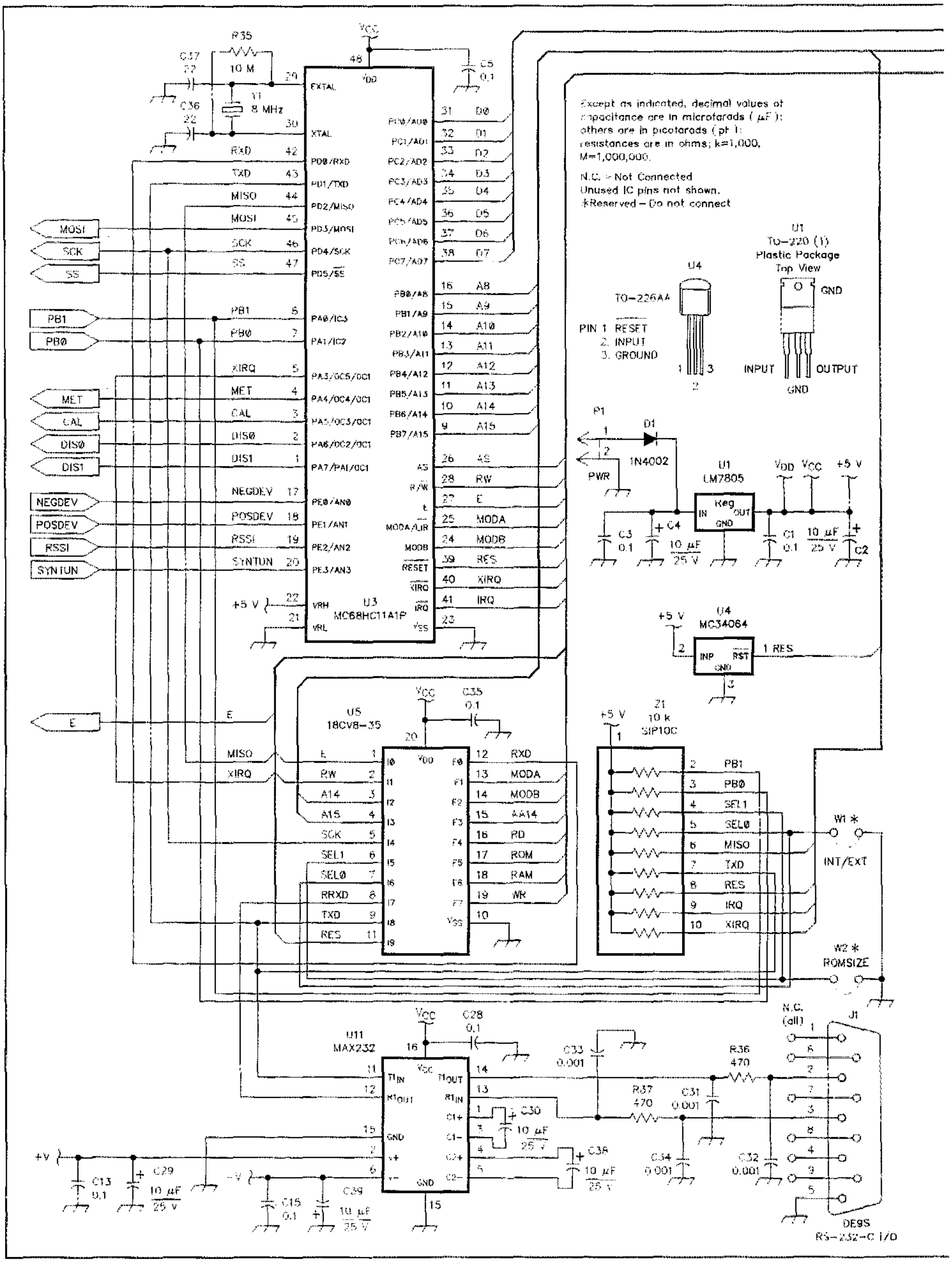


Fig 2—Schematic of the analog-signal-processing section of the deviation meter. Equivalent parts can be substituted. Unless otherwise specified, resistors are ¼-W, 5%-tolerance carbon-composition or film units.

D2, D3—1N4148 silicon switching diode.

P2, P3—2-pin SIP male header.

U7—LM324 quad op amp.


Except as indicated, decimal values of capacitance are in microfarads (µF); others are in picofarads (pF); resistances are in ohms; k=1,000, M=1,000,000.
N.C. = Not Connected
Unused IC pins not shown.
*Reserved – Do not connect
U3
MC68HC11A1P
U5
18CV8-35
U11
MAX232
U4
MC34064
TO-226AA
PIN 1 RESET
2. INPUT
3. GROUND
U1
TO-220 (1)
Plastic Package
Top View
GND
INPUT
OUTPUT
U1
LM7805
Reg
D1
1N4002
P1
PWR
Z1
10 k
SIP10C
Y1
8 MHz
R35
10 M
C37
22
C36
22
C5
0.1
C35
0.1
C28
0.1
C3
0.1
C4
10 µF
25 V
C1
0.1
C2
10 µF
25 V
C33
0.001
R36
470
R37
470
C31
0.001
C34
0.001
C32
0.001
C30
10 µF
25 V
C38
10 µF
25 V
C13
0.1
C29
10 µF
25 V
C15
0.1
C39
10 µF
25 V
MOSI
SCK
SS
PB1
PB0
MET
CAL
DIS0
DIS1
NEGDEV
POSDEV
RSSI
SYNTUN
E
W1 *
INT/EXT
W2 *
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VDD
+5 V
+V
-V

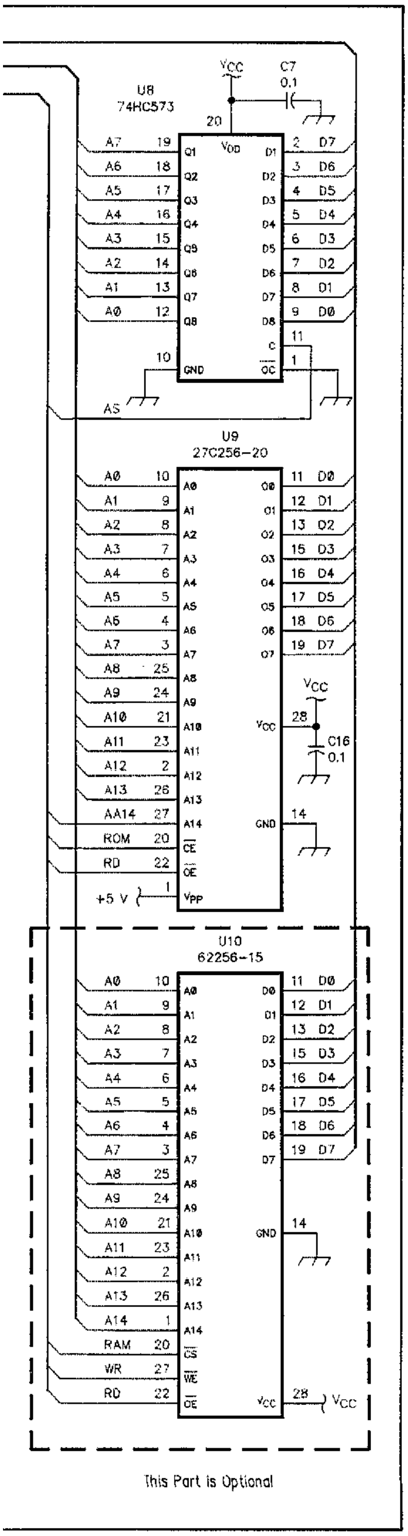


Fig 3—Diagram of digital signal-processing section and power supply of the TAPR deviation meter circuit. Equivalent parts can be substituted. Unless otherwise specified, resistors are ¼-W, 5%-tolerance units.

C1, C2, C4, C29, C30, C38, C39—10-μF, 25-V radial electrolytic.
D1—1N4002 1-A, 100-PIV silicon diode.
J1—DE9S.
P1—2-pin SIP male header.
U1—LM7805 positive 5-V, 1-A, three-terminal voltage regulator.
U3—MC68HC11A1P microcontroller (see Note 1).
U4—MC34064 reset generator.
U5—18CV8 programmable logic array.
U8—74HC573 octal latch.
U9—27C256-20 *programmed* EPROM (see Note 1).
U10—62256-150 (optional; see text).
U11—MAX232, Maxim RS-232-C interface IC.
W1, W2—On-board jumpers: *Do not connect!* (See text).
Z1—10-kΩ SIP resistor package.

After 21 readings are taken (the carrier frequency and ±10 kHz), the VCO is turned off and the DevMtr is ready to automatically measure and display incoming-signal deviation.

The EEPROM

Four basic variables that must be handled by the DevMtr to allow it to work with an existing receiver: the receiver's IF center frequency, signal-strength threshold, signal-strength polarity and detector output-voltage polarity. Fortunately, there is an easy way to store this information: electrically erasable programmable read-only memory (EEPROM). EEPROM remembers things when power is removed.

U3 contains 512 bytes of EEPROM. The firmware contained in the external EPROM (U9) is linked through the EEPROM. This means that you can tell the microprocessor to use certain portions of the program in the EPROM and ignore other portions. The firmware also provides a means of telling the DevMtr about the receiver IF and the polarity of the signals being measured.

Whatever the receiver IF, you can tell it to U3, which remembers the value until you change it. Likewise, if your signal-strength voltage increases or decreases, you can tell it to the EEPROM. You can set the voltage threshold you want to use to discern if your receiver is listening to a signal or only to noise. Finally, you can tell U3 if the detector output voltage increases or decreases with increasing input frequency.

U9 contains the DevMtr code (see Note 1) and a simple monitor program: *BUFFALO.*[5] If S2 is held closed when the DevMtr is powered up, it runs the monitor program. You then key your receiver information into the monitor, which, in turn, stores them in U3's EEPROM. When the DevMtr is next powered up, it self-calibrates and is ready to measure deviation! Pretty neat, isn't it?

Using the Deviation Meter with TAPR's METCON

The DevMtr has special provisions for use with METCON, TAPR's remote telemetry and control device.*

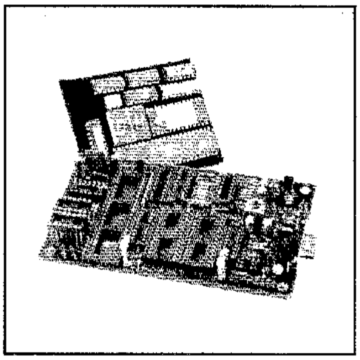
P5, **MET**—Brings the S1 and S2 switch inputs as well as a special pulse output to one convenient location on the DevMtr PC board.

Pin 1—One terminal of an optically isolated pulsed output. It should be tied to a METCON GND terminal at one of the digital inputs of a METCON board.

Pin 2—Second terminal of the optically isolated pulsed output. This ties to any digital input of a METCON device. The input should be configured as a frequency counter.

When pins 1 and 2 are connected as described, the corresponding frequency counter displays a reading of 1/10 the measured deviation. For example, if the pulsed output is connected to METCON input channel 1, and a signal of 2.8-kHz deviation is measured, the AMAP output from METCON indicates a value of 00280.—*WA7GXD*

*P. Newland, "Introducing METCON, a New Remote Control and Telemetry System," *QST*, Jan 1993, pp 41-47.



Just about at the finish line! The next construction step is to plug in the ICs and readouts.

Hooking It Up

Naturally, the DevMtr requires a few connections to your receiver. It must provide the DevMtr with detector voltage output, RSSI voltage output, and dc power. The DevMtr must be able to inject its calibration signal into the receiver's IF amplifier. Here's a guide to these connections:

P1, **PWR**—For attachment of an unregulated source of +8 to +20 V. Current requirement is under 100 mA; typically about 55 mA.

P2, **RSSI**—The DevMtr defaults to a range of 0 to +10 V dc at the RSSI input. A voltage above about +2.5 V dc indicates a

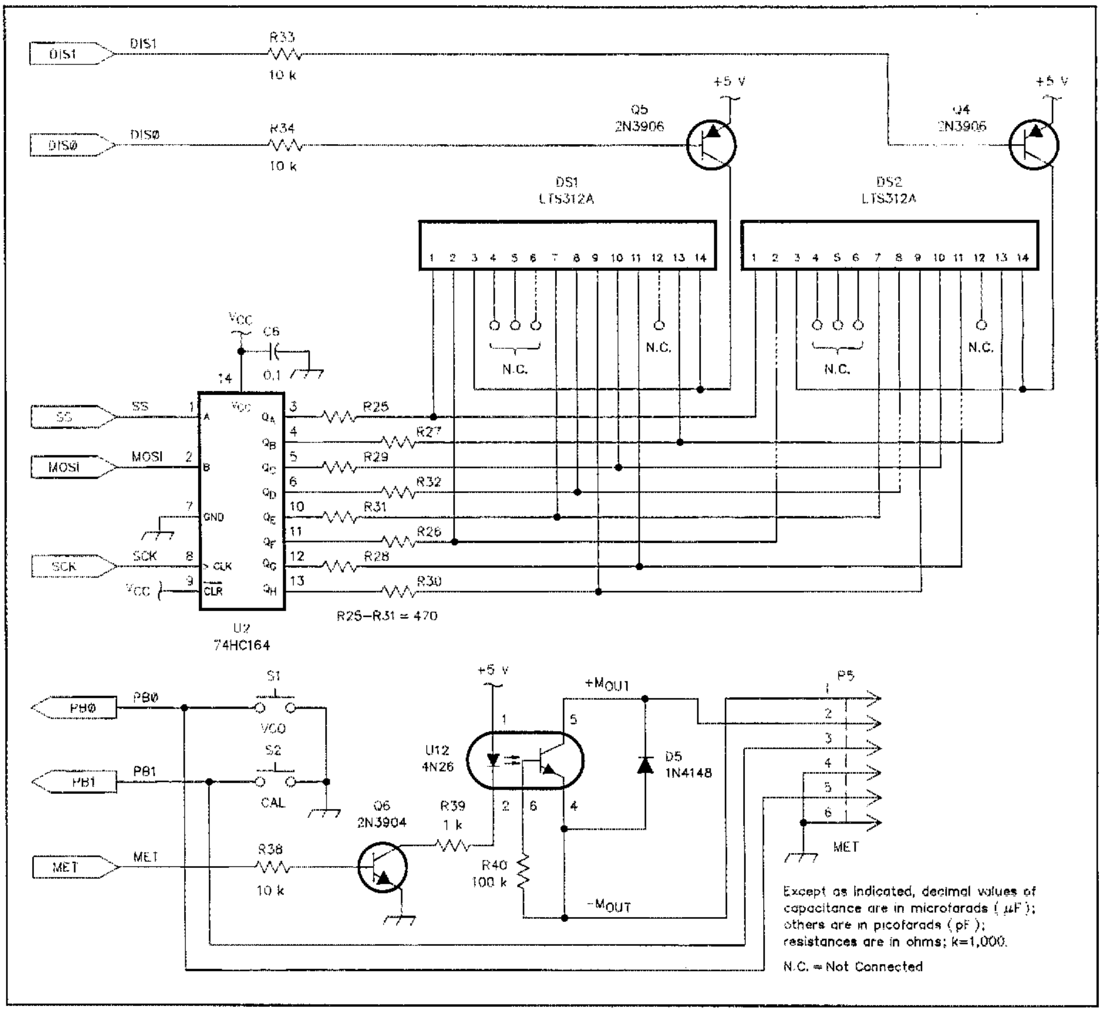


Fig 4—Schematic of the DevMtr's digital display section and METCON interface. Equivalent parts can be substituted. Unless otherwise specified, resistors are ¼-W, 5%-tolerance carbon-composition or film units.

D5—1N4148 silicon switching diode.
DS1, DS2—LTS3132A 7-segment LED display.
P5—6-pin SIP male header.
Q4, Q5—2N3906 PNP transistor.
Q6—2N3904 NPN transistor.
S1, S2—Single-pole normally open push-button switch.
U2—74HC164 shift register.
U12—4N26 optoisolator.

signal is present; a voltage below this level indicates the input is noise and is ignored. (This input has a nominal impedance of 20 kΩ to ground.) If the receiver provides a different output than this, it can be accommodated.

P3, **DISC**—The output of your receiver's FM signal detector (or discriminator) attaches here. DevMtr expects a 0 to +5-V range from the receiver's FM signal detector under all conditions. The meter cannot accept negative voltages, nor can it tolerate voltages in excess of +5 dc. If your detector's output falls outside this limitation, you must construct additional circuitry to scale it within these bounds. This is a high-impedance input, on the order of several hundred kilohms.

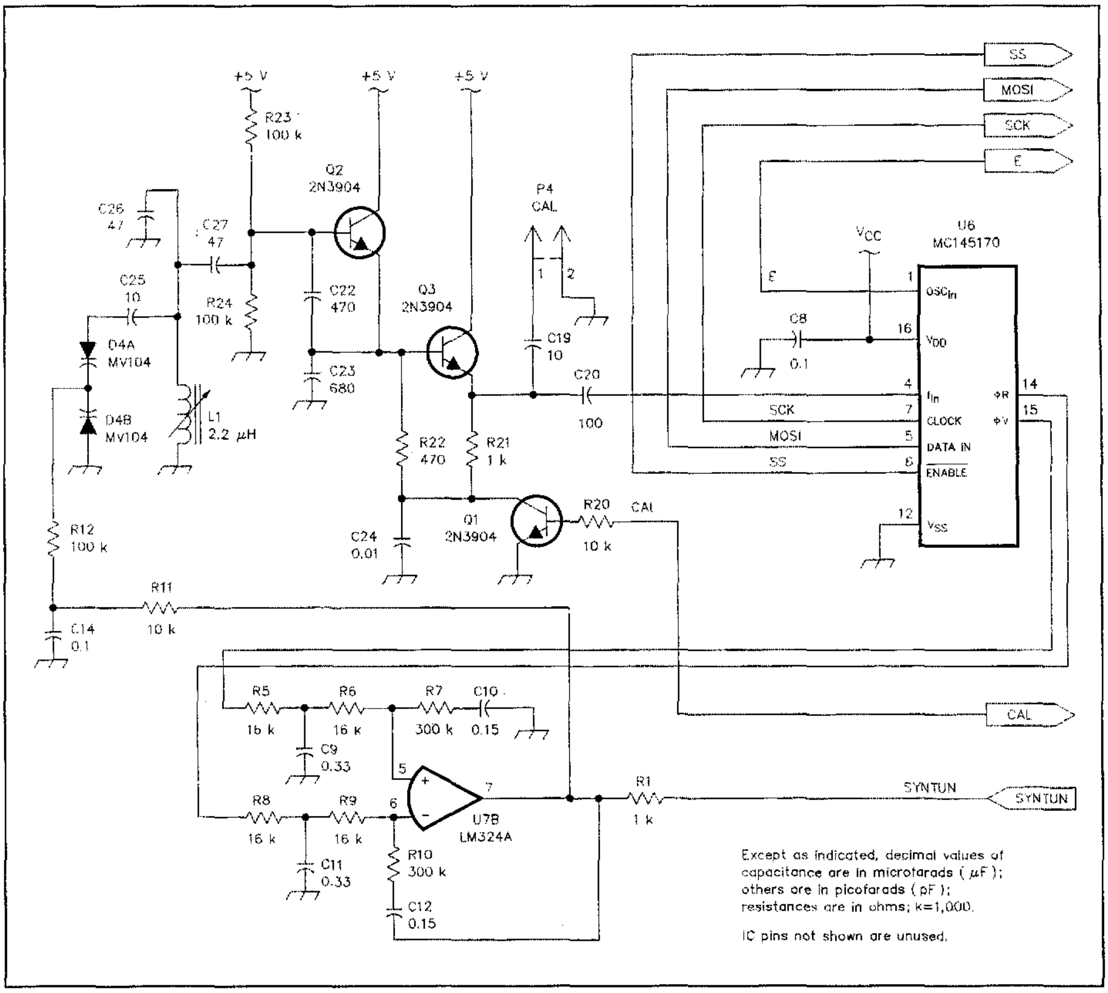
If your receiver has a decreasing output voltage for an increasing input signal (the calibration routine results in a smoothly decreasing voltage at P3 rather than a smoothly increasing one), that can be altered following instructions in the manual.

P4, **CAL**—The DevMtr outputs a several-hundred-millivolt signal here in the IF range of your receiver. The signal is coupled through a 10-pF capacitor to minimize receiver-loading effects. Install a 10-kΩ series resistor at the attachment point within the receiver. This helps to minimize the effects of the cable attaching the receiver to the DevMtr.

This signal is output during VCO calibration and self-calibration of the DevMtr, which occurs at every power-up. The default IF is 10.850 MHz.

Using a Different RSSI-Voltage Range

The DevMtr cannot accept negative voltages. If the incoming positive voltage is greater than 10, alter the values of R16 and R17 to divide the voltage at the output of U7A so it does not exceed +5 V dc. If the maximum output is less than about +3 V dc,



Except as indicated, decimal values of capacitance are in microfarads (µF); others are in picofarads (pF); resistances are in ohms; k=1,000.

IC pins not shown are unused.

Fig 5—Diagram of the deviation meter's PLL VCO. Equivalent parts can be substituted. Unless otherwise specified, resistors are ¼-W, 5%-tolerance carbon-composition or film units. Adjustment of L1 is the only alignment step required.

L1—2.2-µH adjustable iron-core coil (Toko TK1413).
D4A, D4B—MV104 varicap.
P4—2-pin SIP male header.
Q1-Q3—2N3904 NPN transistor.
U6—MC145170 PLL

adjust the values of R3 and R4 to provide more gain from U7A.

Using Different Logic

If your receiver's RSSI is such that a voltage below the threshold means a good signal and a higher voltage means noise, a patch must be made to the DevMtr's EEPROM. The procedure is explained in the instruction manual. A similar procedure is used if your discriminator voltage decreases with increasing signal frequency.

Using a Different Threshold Value

The DevMtr uses an 8-bit analog-to-digital converter (A/D) to measure the RSSI voltage. If your receiver provides a simple squelch output (open or closed), the default threshold of +2.5 V dc will probably work just fine. If it doesn't—or if you want to change the threshold for any reason—that can be done, too.

Using a Different IF Center Frequency

The calibration oscillator can be retuned to other IFs in the range of 10 to 11 MHz by merely adjusting L1. Other IF ranges can be accommodated by further changes in the calibration oscillator component values. (Such changes are beyond the scope of this article.)

If your receiver IF is in the range of 10 to 11 MHz, you can alter the DevMtr's calibration constant located in the 68HC11's internal EEPROM. The DevMtr will use the IF center you've entered for its calibration routines. You must perform the VCO calibration and tune L1 *before* you try to use the DevMtr at this new frequency!

The DevMtr in Action

The DevMtr can be used to set your transmitter's frequency deviation and to check the frequency deviation of other stations. The signal to be checked *must be "full quieting."* Noise detected on the received signal gives a false deviation reading. Further, use of the DevMtr in a strong RF field can result in faulty operation and erroneous readings.

For a voice FM transmitter, follow the manufacturer's directions in setting deviation and deviation limiting. In general, deviation should not exceed about 4.5 kHz.

As an example, for 1200-bits/s AFSK/FM packet (the most common type), use the following general procedure:

- Use a dummy antenna!
- Tune the DevMtr to an unused channel and set the packet transmitter to the same frequency.
- Key the transmitter with the TNC in CAL mode. Select the higher modulating tone (usually 2100 or 2200 Hz).
- Adjust the TNC's output control for a deviation reading of 2.8 kHz. Greater deviation won't help and may make your signal harder for other stations to copy (even though it may sound louder to you).
- Unkey the transmitter and exit the TNC's CAL mode. For 9600 bits/s packet, using direct FSK (the most common type), set things up as for 1200 bits/s packet, then adjust the 9600 bits/s modem for a reading of 3.0 kHz.

Be sure to listen to the receiver's audio as you measure deviation! Your ear is a *valuable* piece of test equipment! For instance, transmitted ac hum may fool the DevMtr into reporting adequate deviation on a signal that's undermodulated with packet audio. *You* can hear the difference and take appropriate action.

And, you'll likely hear stations with excessive deviation needing more retries to get their signals through the network than those which are properly adjusted.

Summary

The DevMtr is simple to construct and use. It provides a reliable, self-calibrating instrument of sufficient accuracy for Amateur Radio needs.

Please let me know if you build this project. I'm particularly interested in compiling interfacing instructions to a variety of radios so the information can be shared with others.

Notes

[1] A *four-layer* PC board and complete kit of parts including a *preprogrammed IC* is available from Tucson Amateur Packet Radio Corporation (TAPR), PO Box 12925, Tucson, AZ 85732-2925, tel 602-749-9479. The kit is supplied with a detailed construction and technical information manual. Because the meter is designed to be added to an existing receiver, the kit does *not* include a case, power connector, fuses, etc. A PC-board template package is *not* available from the ARRL. Complete kit, $90.

All software for this project is also available from TAPR. Source code plus assembler on a 5.25-inch PC-/MS-DOS formatted diskette, $2; 3.5-inch PC-/MS-DOS diskette, $3. Order from the TAPR address given earlier in this note.

[2] *The 1993 ARRL Handbook* describes this deviation meter in Chapter 18 on pages 18-20 and 18-21.

[3] P. Newland, "Introducing METCON, a New Remote Control and Telemetry System," *QST*, Jan 1993, pp 41-47.

[4] A METCON unit and a TNC can be used as a community resource. A packet station can ask the METCON unit for a memory display (=AMAP is the default METCON command). By simply asking for this information, METCON automatically reports the deviation of the inquiring station's transmission!

[5] *BUFFALO* is an acronym for Bit User Fast Friendly Aid to Logical Operation.

Lyle Johnson was first licensed as WN6JLR in 1964 at 13. He received his present call sign in 1967 and holds an Extra Class license. He is a Life Member of ARRL and AMSAT and a member of TAPR and QCWA. He has donated time and expertise to AMSAT and TAPR on a continuing basis. The UoSAT-OSCAR 11 Digital Communications Experiment is largely his design, as are the flight computers aboard the MicroSats (AO-16, DO-17, WO-18 and LO-19). For TAPR, he was heavily involved in the design of the Alpha and Beta TNCs, TNC 1, TNC 2, PSK modem, 9600 bits/s modem, NNC and other projects.

Lyle is a past member of the ARRL Digital Committee (1984-1991) and currently serves on the Future Systems Committee. Lyle was the first recipient of the Dayton HamVention Technical Excellence Award in 1984 for his leadership role in the TAPR TNC 1 project.

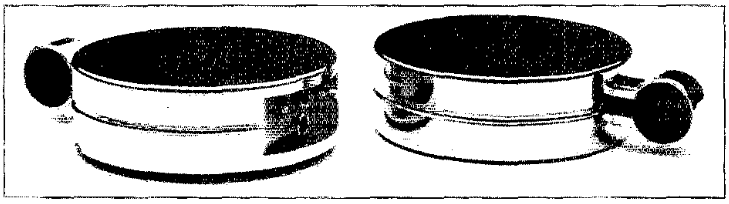
Professionally, Lyle is Vice President of Hardware Engineering at Modular Mining Systems, a company that provides mine management systems worldwide. Oddly enough, these systems are based on packet radio techniques...

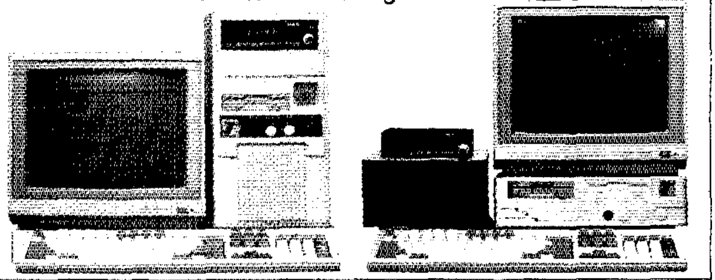
Lyle is very happily married to Heather, N7DZU. They have six children, three of whom are licensed amateurs. They are working on the other three!

New Products

"SILENT KEY"

◊ A code key with no moving parts? The Spirit series dual-paddle Morse code keys (below) use solid-state force sensors to trigger keying signals. The independent left and right paddles have infinite adjustments can be preset to any operator's touch, and when no sidetone is used, operation is completely silent. Each limited-production key is individually serialized and can be customized to user specifications: The standard polished solid-brass instrument can be engraved with name and call sign, and is also available with chrome or gold plating. Detachable cabling accommodates a variety of keyers. Prices start at $430. Ken Hirschberg, K6HPX, CAL-AV Labs Inc, 515-B Westchester Dr, Campbell, CA 95008; tel 408-369-1000, fax 408-371-0672.





Tucker Packet Systems

INTEGRATED PACKET PC

◊ There's a simple way to get on packet radio with no fuss, no need to select components or prepare custom cables, and no frustration with conflicting with hardware arrangements. The Tucker System 1 (above) is an IBM-compatible XT-class computer with a mini-tower case that houses a 25-watt Alinco DR-1200T 2-meter data transceiver and a DRSI PC*Packet Adapter; it's completely self-powered, and all the operator adds is an antenna. The System 2 has a subminiature 3 × 12 × 14-inch desktop case with an external transceiver and Astron RS-10A power supply. Each standard system comes with 640 kbytes RAM, monochrome monitor, 3½-inch floppy disk drive, serial and parallel ports, and keyboard. Custom configurations are available, including upgrades to a 12-MHz 80286 CPU, a 40-Mbyte hard drive, 1 Mbyte RAM and other options. Retail prices are $699 for System 1 and $795 for System 2. Alan Benoit, WQ5W, Tucker Electronics and Computers, 1717 Reserve St, Garland, TX 75042-7621; tel 800-527-4642 ext 230 or 214-348-8800, fax 214-348-0367.

Schematics at Your Fingertips

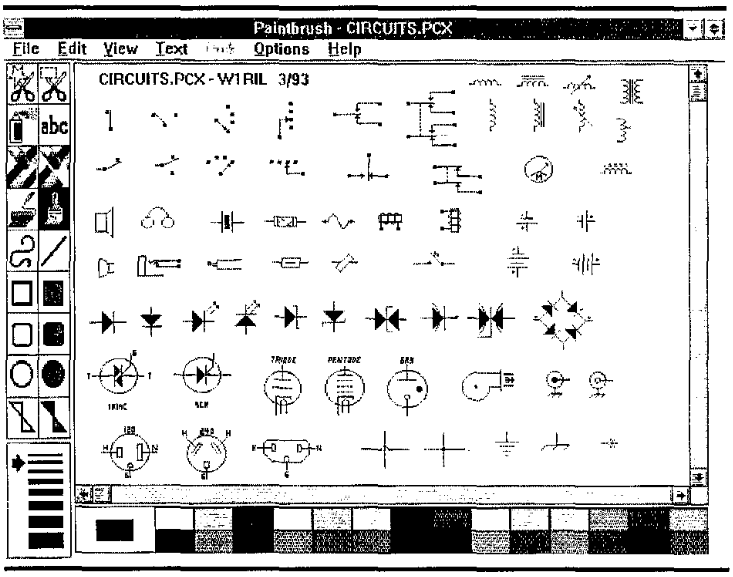
If you can run *Microsoft Windows Paintbrush*, you can move those dog-eared schematics off the backs of envelopes and into your computer where they belong.

By Ken Schofield, W1RIL
21 Forestdale Rd
Paxton, MA 01612

Did you ever wish you could sit down at the computer and draw the schematic of that weekend gem you just completed at the workbench? You can—with *one of those expensive drawing programs* you've drooled over in the software ads. But those programs cost from around a hundred to thousands of dollars—quite an investment for a weekend engineer! In this article, I'll tell you how you can get started with basic schematic CAD (computer-aided design) on a budget. The secret is an application that comes bundled with *Microsoft Windows*.

Windows?

Many IBM-compatible PC systems come with a program called *Windows*, the latest version of which is 3.1. *Windows* is an operating *environment*: It manages the affairs of, and gives a consistent look and feel to, *Windows*-compatible programs—*applications*—that run within it. The *Windows* package comes with a set of basic applications that include ASCII text editing (*Microsoft Windows Notepad*), word processing (*Microsoft Windows Write*), calendar, calculator, telecommunication, drawing and game software. Drawing? Bingo! The *Microsoft Windows* bundle includes *Microsoft Windows Paintbrush*. *Paintbrush* allows you to draw on-screen in monochrome or color (if you're fortunate enough to have a color system). You can draw straight lines, boxes, circles, rectangles, triangles—you name it! You can even magnify (zoom in) on an area in the drawing and get right down to the picture-element (*pel*, a shortening of the older term *pixel*) level, adding, erasing or changing pieces of the drawing to meet your requirements.



Pel by Pel

A pel is small—about the size of a speck of dust on your screen. But you don't need a magnifying glass to paint images pel by pel with *Paintbrush*. Fig 1, a resistor I drew with *Paintbrush*, shows an example of this. Darkened pels form the resistor symbol. Now that you've seen what a *Paintbrush*-drawn schematic symbol looks like, you're ready to draw one yourself. Here's how.

Start *Windows* and run *Paintbrush*. (If you haven't moved the *Paintbrush* program item from its default location, you'll find it in *Windows'* Accessories program group.)

Most of *Paintbrush's* startup defaults—Brush tool, white background and black foreground colors—are just what you need to start drawing schematics. (If you're already in *Paintbrush* and have changed these settings to something else, change them back now.) One, the View default (Zoom Out) *isn't* what we need, so change it to Zoom In via the View menu. This changes the cursor to a rectangle. Position the Zoom In rectangle where you want it—right in the middle of the drawing area is fine for now—and click the left button on your mouse. (A right click at this point drops you back to Zoom Out.) Choose the minimum-width line in the Linesize box.

Move the cursor to the pel you want to change and left-click the mouse. This colors the pel black—the foreground color currently in force. (Right-clicking sets the pel to the background color—currently, white. So you can use the right mouse button as a form of Undo if you darken the wrong pel.) Do this pel by pel until the image—let's say you're doing your own resistor—is finished. If you have a steady hand, you can hold down the left mouse button and drag the pointer through the pel blocks you want to blacken. (Likewise, you can right-click and drag through pels you want to whiten.) Selecting Zoom Out lets you see the image full-scale.

Snip, Snip, Snip

What if your finished resistor isn't quite where you want it? No problem! With Zoom Out selected, click on the Scissors tool in the Toolbox. Position the crosshairs cursor at one corner of the drawing section you want to move. Left-click and drag the cursor around the image you want to copy. (The Pick tool does essentially the same thing except that where you left-click anchors an expandable box that you drag larger to enclose what you want to copy. You may find this more comfortable to use than the Scissors.) Unclick. The enclosing line is now a dashed box. Position the cursor inside the dashed box, and drag the resistor image to where you want it. Magic! You've moved the resistor to a new location. Now select the Paintbrush tool.

Paintbrush also lets you *copy* images. So if you need more than one resistor, you only have to draw one. Select the resistor you drew with the Scissors. Click on Copy in the Edit Menu. Then click on Paste. Magic! Another copy of the resistor appears in the upper left corner of the image area (anything already there won't be affected) ready to be dragged to its final destination. Once you've moved it where you want it, click on the Paintbrush tool and you're ready to keep drawing. Or

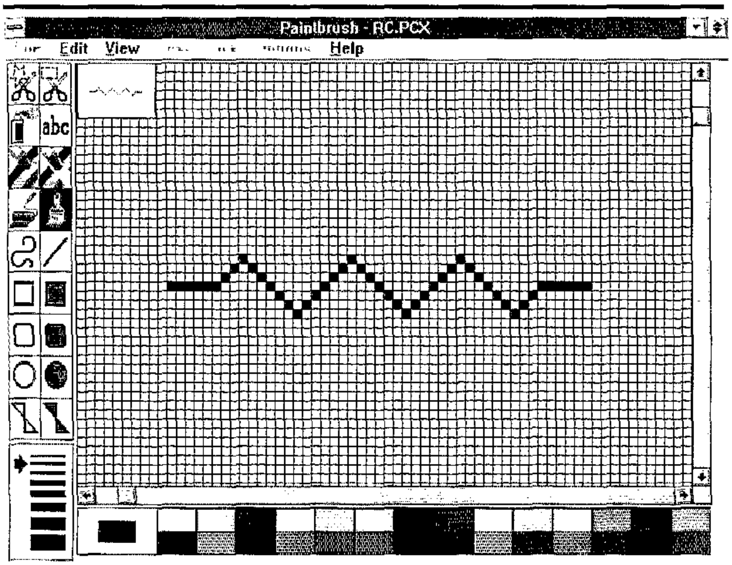


Fig 1—Darkened pels look like paint clogging up a window screen! *Microsoft Windows Paintbrush's* larger-than-life Zoom In view lets you draw components pel by pel with simple mouse commands. (It's possible to navigate *Paintbrush* with just a keyboard, but a mouse is recommended.)

choose Paste again to import another copy of the image. The Cut and Copy commands move or copy the selected pels to computer memory *Windows* calls the Clipboard. The image stays there until you overwrite it with something else or exit *Windows*. (You can view the Clipboard contents whenever you want by opening *Windows Clipboard Viewer* through *Windows's Program Manager*.)

Once you've got two or more symbols where you want them, you can interconnect them with solid lines. You also can draw boxes, curves, circles and slanted lines—check out the Toolbox. You can even enter and edit text for drawing labels. I won't go into the details; the idea is to run *Paintbrush*, refer to its documentation (on-line Help and the *Windows* manual) as necessary, and *play!*

Working Within *Paintbrush's* Limitations

With full-blown CAD packages, you don't need to draw most of the symbols you need. *Symbol libraries*—files containing predrawn components—account for part of their purchase price. You just import symbols from the symbol libraries as you need them. You can usually *scale*—resize—and rotate images until they appear as you want them. And big-time schematic CAD packages often contain PC-board drafting capability that can be linked to their schematic-drawing functions.

Paintbrush approximates *some* of this utility. There's currently no way to use a *Paintbrush* schematic file in generating PC-board artwork from *Paintbrush*, for instance. On the other hand, you can scale *Paintbrush* images via the Pick menu's Shrink & Grow command. You can flip them upside down (Flip Vertical) or turn them into mirror images (Flip Horizontal). One limitation that's important in schematic drafting is that you can't rotate *Paintbrush* images 90°! *Paintbrush's* ability to import images can help you get around these limitations. The key is to develop symbol-library files that contain all the components you need in horizontal *and* vertical formats. Then you use Copy and Paste to move the images you want to and from the *Windows* Clipboard.

Windows itself lets you end-run another *Paintbrush* limitation that might complicate this process. *Paintbrush* can open only one image file at a time, so it might seem that you'd have to endlessly open, change, save and close files to make maximum use of the Clipboard. But since *Windows* lets you run more than one *Paintbrush* session at the same time, all you have to do is start and run as many *Paintbrush* sessions as you have symbol-library files, opening each additional file per each additional *Paintbrush* window. Then you can hop from session to session—in effect, from library file to library file—via *Windows*' Task List (callable by hitting **Ctrl-Esc**), or by **Alt-Tabbing** through them in turn.

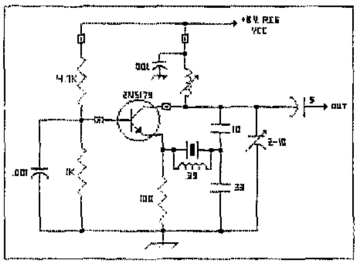


Fig 2—Drawing this VHF crystal oscillator is easy with symbol-library files in place. *Paintbrush* can't equal a full-blown CAD package, but it gives you enough functionality to get started in basic computer graphics.

Ready-Made Paintbrush Symbol Library Files

I've saved the best news for last. Developing symbol libraries takes time—time you'd probably rather spend on drawing schematics than building basic components pel by pel. You can get the symbol library files I've developed from the ARRL telephone BBS or directly from me.[1] The files contain all the symbols I thought would be useful in making electronic schematics, including transistors and logic symbols. Most of the symbols are identical to those found in the *ARRL Handbook*. The files are named as follows:

ANTSYS.PCX	Antennas and dummy antennas, RF bodies and switches
RC.PCX	resistors and capacitors
CIRCUITS.PCX	tubes, diodes, chokes, switches, relays
TRANSLOG.PCX	transistors and logic symbols

The title graphic shows some of the symbols as they appear in *Paintbrush*. If the files don't contain a symbol that you need, you can draw it and add it to the library file or files of your choice. And of course you can customize my symbols as you wish. I encourage you to share your symbol-library developments with others via the ARRL BBS.

Putting It All Together

Fig 2 shows a schematic of a popular VHF crystal oscillator circuit assembled from components in my symbol-library files. Going back and forth between drawing and symbol-library windows can be time-consuming. You can speed the process by predetermining which symbols, and how many of each, you'll need for a given drawing. Then you can repeatedly Copy (dragging each imported symbol out of the upper left drawing corner each time) as many times as necessary before moving on to the next symbol and overwriting the previous symbol on the Clipboard.

Once you've brought in all the schematic pieces and wired them together, you can save your weekend project's schematic for posterity as a new .PCX file. If you own a *Windows*-compatible printer, you can also print a paper copy for your records. *Paintbrush* allows you to resize the drawing when printing.

Now there's no excuse for not keeping your schematic records up to date!

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[1] The ARRL telephone BBS awaits you at 203-666-0578. Copies of the symbol-library files are also available by mail from the author. Include with your request an IBM-compatible-formatted diskette (1.2-Mb [5¼-inch]; or 720-kB or 1.44-Mb [3½-inch]) in a suitable disk mailer (return postage required). QST

The Earth Detunes My Antenna

You know that the electrical quality of the soil beneath a horizontal antenna affects its impedance. But do you know how the resonant frequency is affected?

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Recently, through computer analysis, I studied the effects of varying soil conditions on the feed-point impedances of dipole antennas over "real earth." Discovering how much the soil characteristics *really* affect antenna impedance was fascinating. I also found a few surprises in the information I gathered about the dipole.

Let's say we plan to install a half-wave, 80-meter dipole in our backyard. We have sufficient space to install it as a straight wire in a horizontal plane (how fortunate!). It will be at a modest height, 50 feet, and constructed with #12 copper wire. We want it to resonate at 3.75 MHz, the center of the 80/75-meter band. To find the antenna length in feet, we plug this frequency into the familiar equation for a ½-λ dipole, 468 divided by the frequency in megahertz, and come up with:

$$\text{Length} = \frac{468}{3.75} = 124.8 \text{ feet} \qquad \text{(Eq 1)}$$

So we cut our dipole to this length, placing 62.4 feet of wire each side of the feed point. Now, by analysis, let's see if this antenna actually resonates at 3.75 MHz.

In this study I used *NEC*-2, an exceptional moments-method computer program. With its Sommerfeld/Norton method of determining ground interaction, *NEC* provides reliable impedance results for varying soil characteristics.[2] (*MININEC* and its derivative programs, such as *ELNEC* and *MN*, do not accommodate such impedance calculations, permitting evaluation only for free space or perfectly conducting earth.) For this antenna I used 11 segments (segment length is approximately 0.045 λ), and neglected copper wire losses.

To obtain meaningful information from any analysis of this nature, we must first establish a *reference*, or starting point. A logical reference here is a dipole in free space, where there are no earth effects. In practice, one could probably come close to reaching free-space conditions by locating the antenna in a desert with nothing but clean, dry sand to a depth of several dozen feet.

Results from *NEC* (rounded to the nearest tenth of an ohm) indicate the free-space feed-point impedance of our dipole is $67.5 - j41.8\ \Omega$. This is *not* resonant; by definition, there is zero reactance at resonance. The 41.8 Ω of capacitive reactance indicates the applied frequency is *below* that of the antenna's resonance. To reach resonance, we could either raise the applied frequency (to obtain a shorter wavelength and thus make the antenna electrically longer), or else make the antenna physically longer. *NEC* indicates that for resonance at 3.75 MHz, we must lengthen the antenna by about 2.5%, for a total length of 127.9 feet. That works out to be $479.6 + f_{MHz}$, rather than the familiar 468. The calculated feed-point resistance for an antenna of this length with no. 12 wire is 72.2 Ω. (The 73-Ω value you see quoted so often for free space is for an infinitely thin conductor.) This free-space-resonant dipole, 127.9 feet long, will serve as our reference antenna.

Earth Effects

Now let's see how the earth affects this antenna. From an analysis standpoint, the ground provides an *image antenna* beneath the real antenna, as far below the earth's surface as the real antenna is above it. For an antenna height of 50 feet, it is exactly the same as if we had two dipoles in free space, 100 feet apart. Mutual coupling, along with the earth's "efficiency" in creating the image antenna, affects the feed-point impedance of the true antenna. Because of mutual coupling, current flowing in the dipole induces current into the image antenna, which in turn induces current into the real dipole. Because the dipole's feed-point impedance equals the applied voltage divided by the *total* (incident and induced) current, you can see that a change in the induced-current component causes a change in the antenna's impedance.

If the earth beneath the antenna were a perfect conductor, without losses, then the image antenna would be identical to the true antenna. The currents flowing in the two antennas would be of equal amplitude, but of opposite phase. The phase difference exists because there is a 180° phase shift in the lossless-earth reflection of a horizontal antenna. If the earth is imperfect (that is, real earth with dielectric losses), then the image antenna will have a lower current amplitude, and the phase shift will depart from 180°. This subject is discussed in more detail later in this article.

First let's consider our reference dipole over perfectly conducting earth, approximately equivalent to a solid sheet of copper extending an infinite distance in all directions. With regard to soil conditions, this is the opposite extreme of free space; instead of having no earth effects, now *all* rays that strike the earth are reflected. In practice, saltwater is quite close to a perfect conductor. We could probably come close to the same impedance conditions on dry land with a 1-λ-square copper screen on the ground beneath the antenna.[3]

If we install our reference antenna at 50 feet, *NEC* reports the resulting impedance over perfect earth to be $63.6 + j38.1\ \Omega$. This means the earth is reflecting an inductive reactance into the antenna, and we must make it shorter to maintain resonance at 3.75 MHz. Analysis reveals the resonant length to be 125.2 feet, which works out to be $469.5 + f_{MHz}$. The feed-point resistance for this resonant length is 59.6 Ω. (Shortening the antenna to make it resonant lowers its radiation resistance slightly. We saw the same effect when we shortened our finite-diameter, free-space dipole slightly to make it resonant; its radiation resistance went from 73 to 72.2 Ω.)

Likely few amateurs have a backyard with conditions close to either extreme (a great depth of dry sand or an expanse of saltwater). Most of us have soil with electrical characteristics somewhere between very good and poor. Table 1, taken from *The ARRL Antenna Book*, defines the electrical characteristics for various types of soil. The analysis results so far indicate that for a dipole at 50 feet over real earth, we must cut

[1] Notes appear on page 44.

Table 1

Conductivities and Dielectric Constants for Common Types of Earth

Surface Type	*Dielectric constant*	*Conductivity (S/m)*	*Relative quality*
Fresh water	80	0.001	
Salt water	81	5.0	
Pastoral, low hills, rich soil typ Dallas, TX to Lincoln, NE areas	20	0.0303	Very good
Pastoral, low hills, rich soil typ OH and IL	14	0.01	
Flat country, marshy, densely wooded, typ LA near Mississippi River	12	0.0075	
Pastoral, medium hills and forestation, typ MD, PA, NY (exclusive of mountains and coastline)	13	0.006	
Pastoral, medium hills and forestation, heavy clay soil, typ central VA	13	0.005	Average
Rocky soil, steep hills, typ mountainous	12-14	0.002	Poor
Sandy, dry, flat, coastal	10	0.002	
Cities, industrial areas	5	0.001	Very Poor
Cities, heavy industrial areas, high buildings	3	0.001	Extremely poor

its length for resonance somewhere between $469.5 \div f_{MHz}$ and $479.6 \div f_{MHz}$, depending on the soil conditions beneath the antenna. Additional *NEC* analysis reveals that for very good earth we should use $470.8 \div f$, $472.7 \div f$ for average earth, and $473.9 \div f$ for poor earth. This amounts to an overall length change of about 10 inches from very good to poor earth, everything else being equal!

Although that's not a whole lot of wire, according to *NEC*, it represents a change of 25.4 kHz in the antenna's resonant frequency, *just because the soil ranges from very good to poor*. Surprise! I used to think hams were joking when they told me their antennas' SWRs changed during periods of heavy rain, but now I'm not so skeptical. Fresh water, and presumably water-soaked ground, are high on the list of "good soils." An antenna that was resonated over poor earth could be detuned by the improved soil characteristics resulting from rain, and this would cause the SWR to rise. (If the SWR was 1:1 in 70-Ω line when the dipole was resonated, a resonance shift of 25 kHz would raise the SWR at the original frequency to about 1.2:1.)

Keep in mind that the above values apply only for a dipole height of 50 feet, which is about 0.19 λ at 3.75 MHz. Even then, those values are not *exact* because factors such as wire loops through the supporting end insulators, coupling to other objects in the vicinity, and so on, are not taken into account. It is not my intent here to provide cookbook values for precise dipole resonance for all possible situations, but rather to illustrate how soil conditions affect resonance of an antenna at a fixed height.

Other Heights

The foregoing example shows that the earth, perfect or imperfect, reflects an inductive reactance into the antenna's feed-point impedance when the antenna is 0.19 λ high. Although interesting and illustrative, this analysis of a horizontal dipole at a fixed height provides information of rather limited value. A thorough analysis should examine both the resistance and reactance results for a range of antenna heights over various kinds of soil. So let us look further into earth effects.

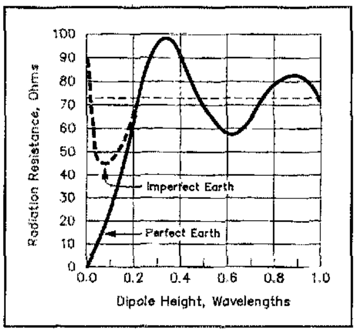

Fig 1—Variation in feed-point resistance of a horizontal half-wave antenna at various heights above ground, after *The ARRL Antenna Book* (15th and 16th eds, p 3-11). The same information, less the broken line for imperfect earth, appears in Kraus (reference of Note 1, p 463).

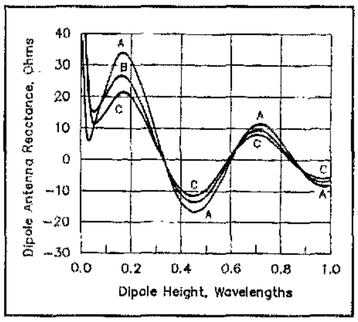

Fig 2—Variation in feed-point reactance of a free-space-resonant horizontal half-wave antenna at various heights above imperfect ground. Curve A, very good earth; curve B, average earth and curve C, poor earth. See Table 1 for definitions of these earth types.

We have seen that the antenna at 50 feet must be shortened from the free-space resonant length to maintain resonance. Is this true for all heights? Certainly not! The mutual impedance from the image antenna changes with height, and therefore the resulting resistance and reactance at the feed point also change. At heights between approximately 0.33 and 0.6 λ, we'd have to *lengthen* the free-space-resonant antenna to maintain resonance. The same is true for heights between about 0.85 and 1.1 λ, and again for even greater heights. The exact heights where the shifts occur depend on the soil conditions, but for ground ranging from extremely poor to very good, the height difference is only a few feet at 3.75 MHz.

Information on feed-point resistance as a function of horizontal antenna height has been available in the literature for decades. Fig 1 shows the damped-sine-wave curve that has become familiar to many amateurs. But lacking in the literature is much information on feed-point *reactance* changes with height. An excellent article that touches briefly on this subject has recently been published,[4] but that treatise primarily discusses feed-point resistance. To help fill the void in reactance information, I offer Fig 2, showing the feed-point reactance as a function of height and soil characteristics. Curves for three types of soil are included: very good, average and poor. For this portion of the analysis, the free-space resonant length (actually 127.892 feet) was held constant while the height and soil characteristics were varied.

At first glance, the reactance curves of Fig 2 appear to resemble the resistance curve of Fig 1. But take a closer look: The peak excursions of the resistance curve do not occur at the same antenna heights as the peak excursions of the reactance curves. Instead, the zero-crossing points of the reactance curves (where the reactance goes through zero) coincide with the heights producing peak excursions in the resistance curve. And a converse situation is also true; the peak excursions of the reactance curves occur at heights where the feed-point resistance crosses 72 ohms, the resistance value for free-space.

We would expect the curve for average earth (curve B in Fig 2) to fall between curves A and C for very good and poor earth. This is indeed what happens at heights above 0.07 λ, but (surprise) does *not* hold true for heights below 0.07 λ. Here the reactance change with poor earth is close to that of very good earth, while that for average earth is not so close.

Earth Efficiency

Earlier I mentioned that the efficiency of the earth in creating the image antenna affects the actual feed-point impedance of a dipole antenna. Consider for the moment just the image antenna that exists below the earth's surface. It's like looking into a mirror and seeing an image of yourself behind its surface. If the earth is a perfect conductor, then as stated before, the image antenna is identical to the true antenna and the ratio of the current magnitudes in the two antennas is 1:1. And as discussed earlier, there is a 180° reversal of the current phase. Again consider the mirror image. As you stand before the mirror, notice any object that exists to your left. Now consider yourself as being behind the mirror. There is a similar reversal here: that same object is now to your right, a 180° reversal.

To obtain information on earth efficiency, I treated the dipole and its image as two antennas in free space. By working from the impedances that *NEC* reported for the actual dipole over lossy earth, I calculated the magnitude and phase information for the current in the image antenna. (Thanks to Roy Lewallen, W7EL, for leading me to the development of a straightforward procedure to accomplish this.[5]) The method is described in the Appendix.

The 1:1 current ratio and the 180° phase difference do change when the earth beneath the dipole has dielectric losses. The situation is complex, but this broad generalization states it fairly well: With the antenna at a fixed height above 0.05 λ, the current magnitude in the image antenna decreases with poorer earth. There is no surprise here; we would expect that increasing earth losses reduce the current.

The biggest surprise of this whole study came when I examined the current magnitudes in the image antenna. Without giving it really serious thought, I had always assumed that the current decreased somewhat linearly with greater antenna heights. But not so, according to calculations based on *NEC* results! Above 0.05 λ, the current amplitude *rises* logarithmically with antenna height.

I also expected the phase difference to change with height, but the results show that the phase angle remains almost constant at heights above 0.1 λ. (But magnitude and phase vary with different soil types.) Fig 3 shows the image-antenna current changes versus height for various types of soil. I did not carry the study to heights above 0.65 λ.

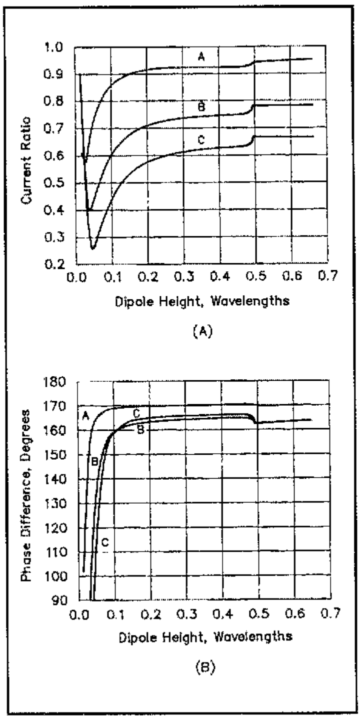

Fig 3—Current flowing in the earth image of a ½-λ dipole over real earth versus dipole height. At A, the current magnitude is shown relative to the current flowing in the dipole itself, and at B the phase angle is shown relative to that at the dipole's feed point. For magnitude and phase, curve A is for very good earth, curve B for average earth, and curve C for poor earth. The text explains the steps in the curves.

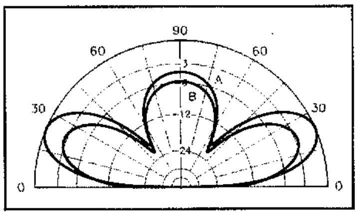

Fig 4—Elevation-plane radiation patterns calculated with *ELNEC*. Curve A, the pattern of a 0.6-λ-high dipole over average earth (see Table 1) at 3.75 MHz. Curve B, the pattern of an array of two dipole elements in free space, 1.2 λ apart, and using a phase difference determined from Eq 3. This array represents the dipole and its earth image. In theory, Curve A should exhibit 3 dB more gain at all angles because of earth reflectivity, but earth losses reduce that value. Here the difference at 23° is 2.6 dB, and at 90° is 1.5 dB. The favorable *NEC* and *ELNEC* comparison of the patterns and feed-point impedances at this and other heights indicates the current analysis method of the Appendix is sound.

Refer to Fig 3A. For a given height, the current magnitudes for poor earth (curve C) are approximately 65% of that for very good earth (curve A), while for average earth (curve B) they are approximately 80% of that for very good earth. Depending on antenna height, very good earth falls in the 90% to 95% range of that for perfect earth (a constant ratio of 1.0 for all heights).

The phase angle of the image-antenna current, Fig 3B, does not change appreciably with changes in antenna height except below 0.1 λ. Rather, it depends primarily on the dielectric characteristics of the earth. For very good earth, the values hover around 170, and 165 for average and poor earth.

All the curves of Fig 3 show steps at a height just below 0.5 λ. *NEC* actually switches from the Sommerfeld integration method to the Norton approximation when the antenna and its image are 0.95 λ apart. The steps in the curves result from this switch at a corresponding dipole height of 0.475 λ; small discontinuities in the dipole impedances are magnified in the current analysis.

Frequency Effects

The curves of Figs 2 and 3 result from using a frequency of 3.75 MHz. For a given type of soil, earth losses become greater as the frequency increases. For frequencies higher than 3.75 MHz, the curves of Fig 2 therefore exaggerate the amount of antenna reactance. Stated another way, the peak excursions for a given type of soil diminish with increasing frequency.

Additional *NEC* calculations disclose that, for the same electrical heights, average earth at 14 MHz produces a curve very close to curve C in Fig 2 (poor earth at 3.75 MHz). The greatest departure is for a height of 0.08 λ, where the 14-MHz curve dips to about +9 Ω reactance before rising to join curves B and C.

One thing to keep in mind concerning frequency effects is that, for a fixed physical antenna height, increasing the frequency also increases the electrical antenna height. And as an antenna is raised to greater electrical heights, the earth effects diminish. So there is a counteracting effect: higher dielectric losses from increased frequency, but lower dielectric losses because of increased electrical height. The height increase wins out; from an impedance standpoint, over real earth, the net effect is less change at the higher frequencies. Stated another way, for a fixed physical height over imperfect earth, a dipole antenna that's pruned to resonance will be closer percentage-wise to its free-space-resonant length at the higher frequencies.

Why isn't reactance information of this nature widely available in the literature? Perhaps because it is not of great consequence. Suppose you do put up an antenna and find that it's not resonant. It's generally a small task to prune the antenna to resonance, and you don't need reactance curves to aid the process. But on the other hand, you have very little control over the feed-point resistance; you must take what you get (discounting small changes with length adjustment), so resistance information is worth having for reference, to facilitate matching and the like.

So of what practical value is all this information? For one thing, this study emphasizes why the familiar dipole length equation, 468 ÷ f_{MHz}, is presented in the texts as only *approximate*. Use it only as a starting point. This study has shown that with only a few height exceptions, the antenna *will* need to be pruned—made either longer or shorter—for exact resonance. This information may give you an idea of which way you'll need to go to adjust your antenna length, longer or shorter, for its height. It might even be possible for the enterprising amateur to gain an idea of the type of soil under his antenna by judicious use of these data. My hope, though, is that from this information you'll gain a better idea of *how* the earth really does detune your horizontal antenna.

APPENDIX

The analysis of the current ratios as graphed in Fig 3 is based on the relationship from Kraus:[6]

$$V_1 = I_1 Z_{11} + I_2 Z_m \qquad \text{(Eq 1)}$$

where

All terms are complex, and where

V_1 = the applied voltage at the dipole antenna terminals

I_1 = the dipole antenna current

I_2 = the image antenna current

Z_{11} = the self-impedance of the dipole in free space

Z_m = the mutual impedance between the dipole and its image at a spacing of twice the dipole height

From this relationship,

$$Z_1 = \frac{V_1}{I_1} = Z_{11} + \frac{I_2}{I_1} Z_m \qquad \text{(Eq 2)}$$

where Z_1 is the dipole impedance above real earth (as calculated by *NEC*). Rearranging terms, the relationship becomes

$$\frac{I_2}{I_1} = \frac{Z_1 - Z_{11}}{Z_m} = \text{the ratio of currents} \qquad \text{(Eq 3)}$$

Remember, all the terms in each of these equations are complex, with a real and an imaginary component (such as R + *j*X for an impedance.)

Mutual impedance and mutual coupling are discussed in Chapter 8 of *The ARRL Antenna Book*, 15th and 16th editions (1988 and 1991), p 8-4. Also see the *Antenna Book*'s Fig 19 for a graph of mutual resistance and reactance values versus spacing.

To verify the results of this technique, I used *ELNEC* to calculate and plot the broadside elevation patterns of a half-wave dipole at several heights above real earth. Then I calculated and plotted the impedances and patterns for the dipole and its image as an array in free space, with element separations of twice the various heights. For each height I set the element current ratios at the magnitude and phase angle determined from Eq 3. The impedances of the upper element in the free-space array (the "dipole") reported by *ELNEC* agree well with the impedances *NEC* reports for a dipole over real earth.

The patterns also agree well. A comparison of one pair of patterns from the two calculations is given in Fig 4. The patterns over real earth contain a theoretical 3 dB gain from earth reflectivity (the dipole pattern is contained in only the upper hemisphere, whereas the array in free space is not limited in this way). Otherwise the patterns compare very favorably. The excellent agreement of the dipole impedances and patterns indicates that this method of current analysis is valid.

Notes

[1] See J. D. Kraus, *Antennas*, second edition (New York: McGraw-Hill, 1988), pp 384-397, for a discussion of moments-method analyses.

[2] K. A. Norton, "The Propagation of Radio Waves Over the Surface of the Earth and in the Upper Atmosphere," *Proceedings of the IRE*, Vol 26, #9, Sep 1937, and A. Sommerfeld, *Partial Differential Equations in Physics* (New York: Academic Press, 1964). The Sommerfeld/Norton *NEC* option uses a numerical evaluation of the Sommerfeld integrals for ground fields when the interaction distance is below 1 λ, and uses Norton's asymptotic approximations for larger distances.

[3] Although a reflector screen as described in the text may provide *impedance* conditions similar to perfect earth, it could *not* provide the same far-field radiation pattern or the same gain from earth reflections. To obtain equivalent far-field performance, the screen would require a radius of approximately 15 to 20 λ, depending on antenna height.

[4] C. Michaels, W7XC, "Dipoles above Real Earth," Technical Correspondence, *QST*, Nov 1992, pp 67-68. Also see Feedback in the January and February 1993 issues.

[5] Private communication. The method I used assumes uniform sinusoidal current distribution in both antennas, a situation that may not be precisely true in reality.

[6] See the referent of Note 1, p 463.

New Products

Ham Windows UPDATE

◊ This full-featured, integrated Amateur Radio software package, *HamWindows Plus 3.0*, includes the *Radio Construction Set* (*RCS*), which makes it easy to "design a transceiver" graphically on a computer screen, controlling all of a rig's major functions. *HamWindows* features dynamic data exchange (DDE) that lets its *TNCWindow* automatically or manually select stations spotted on a DX *PacketCluster*, tune an HF transceiver to the spot and place the proper information in the *LogWindow* module. It includes awards tracking for DXCC, WAS and WAS; the extensive *Almanac Window*, a data base of facts on the countries of the world from information published in the *CIA World Fact Book*; *SWLWindow* with shortwave listening frequency charts and schedules from the *World Radio and TV Handbook*; ham radio keyword lexicons for popular foreign languages; *MapWindow* with grayline maps; and more. The new *Sunspot Breaker* module will allow hams to monitor all amateur bands and graphically depict which bands appear to be open, in real time. The software is designed to function as a personal communications control center, and the code can be freely modified with *Toolbook* software by Asymetrix (not included).

HamWindows 3.0 is the fourth upgrade for this full-featured program. It runs on IBM-compatible PCs with 80386 or 80486 CPUs, at least 8 Mbytes of RAM, a bus mouse, SVGA and two or more serial ports, operating under MS-DOS with *Windows 3.1*. Retail price is $189.95, and other options are available. California Software Inc, 2121 E Pacific Coast Hwy, Corona del Mar, CA 92625-3235; tel 714-729-4222, fax 714-644-4222.

ANTENNA-DESIGN SOFTWARE

◊ The creator of *MININEC*, *YO (Yagi Optimizer)* and *AO (Antenna Optimizer)* has released *NEC/Yagis 2.0*, a new program especially useful for VHF and UHF antenna design. Because it uses the same data files, it's simple to verify Yagi designs optimized with *YO*. It calculates 20-39% faster than *NEC/Yagis 1.0* (even faster for stacked Yagis), providing as many as 2000 segments, and can model real-world ground characteristics. *NEC/Yagis* features high-performance math and graphics libraries optimized for 80386+387 and 80486DX processors, and requires a hard drive, 640- × 480-pixel resolution, 256-kb color VGA, and provides .PCX file output and improved prints. Price is $100 for amateur use only.

PLOT 6.0 is a general-purpose antenna-plotting computer program that displays and prints far-field antenna radiation patterns in polar or rectangular coordinates. It can quickly synthesize the radiation pattern and calculate the gain for a rectangular array, making it useful in designing large arrays for applications like moonbounce (EME). *PLOT*'s array synthesis feature works with fee-space, single antenna plot files generated by *NEC/Yagis*, *YO*, *AO* or *MN* antenna-analysis programs. *PLOT* generates smooth pattern animation at 640 × 480-pixel resolution with any VGA card. Brian Beezley, K6STI, 507½ Taylor, Vista, CA 92084; tel 619-945-9824.

Yagi: The Man and His Antenna

While Armstrong, De Forest and Sarnoff were warring over Stateside radio inventions, the Yagi-Uda beam antenna was quietly emerging in Tokyo. With nary a shot fired, it took its place in radio history.

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Every ham seems to know about the Yagi antenna, but few know how it came about and who really did the experimental work.

The Yagi, widely used in Amateur Radio and commercial communications, is known as a beam or parasitic array, and occasionally as a Yagi-Uda, reflecting the names of its creators.

It's important to note that many scientists have contributed to the development of radio communications, and no single scientist has been credited with "inventing radio." We associate Morse with the telegraph, Bell with the telephone and Marconi with wireless telegraphy, but the development of radio came from the combined efforts of many researchers working simultaneously around the world.

Early History

Two Japanese electrical engineers[1] on the faculty of Tohoku University at Sendai, Japan, investigated methods of improving transmitted signals from shortwave oscillators. These two engineers were Hidetsugu Yagi (1886-1976) and Shintaro Uda (1896-1976). The name "Yagi" is a household name to radio amateurs and communications engineers the world over; but Yagi's student and assistant, Shintaro Uda, is relatively unknown among the world's amateurs.

Hidetsugu Yagi was born in Osaka, Japan, on January 28, 1886, two years before Heinrich Hertz began his experiments on the generation, transmission and detection of electromagnetic waves. When Yagi died 90 years later, he had more honors than there are directors on the biggest parasitic beam ever erected.

Yagi preferred the arts to the sciences in his early years, but slowly became enthusiastic about science as he learned more about physical phenomena. He earned an engineering degree from Tokyo Imperial University (now Tokyo University) at age 24. He then began a career as a teacher at the Sendai Engineering High School. After four years of teaching, the Ministry of Education sent him on a European tour to further his education. This common practice put Yagi in touch with some of the great European experimenters of that time.

Yagi first went to Germany to study resonant transformers of the type used in transmitters. Yagi's work with Heinrich Barkhausen involved generating CW oscillations by electric arcs; that is, spark-gap transmitters. Barkhausen's studies of oscillations led to the construction of an early vacuum-tube VHF oscillator (which was later used by Uda in his VHF propagation studies).

Yagi's work was interrupted by the outbreak of WW1 in 1914. He left Germany and headed for England. The war caused such a scarcity of young talent that Yagi had no difficulty in obtaining employment with J. A. Fleming in London. Fleming had invented the vacuum diode (Fleming Valve), which was beginning to be used as a detector in radios.

Following two years of experimental studies with Fleming, Yagi came to the United States to work with George W. Pierce at Harvard University. Pierce was known for his research on crystal rectifiers and was the inventor of the Pierce oscillator, an early crystal-controlled oscillator capable of generating a continuous wave. During his stay in the United States, Yagi joined the Institute of Radio Engineers (IRE), now the IEEE, which meant that he would have access to its *Proceedings* when he returned to Japan.

While in the US, he submitted two papers in English on his recent research: the first one discussed his work with resonant transformer circuits and the second dealt with comparing tones in wireless telegraphy using ac and dc sources applied to different spark gaps.

By the time Yagi returned to Japan to resume his teaching duties, he was well versed in the most up-to-date RF technologies. Also at this time, the Sendai Engineering High School merged with the Tohoku Imperial University; Yagi received his doctorate two years later in 1921. With his research background, it was natural for Yagi to continue his earlier line of research on the generation of radio waves.

He gathered an excellent team of student researchers who furthered his own studies. It was at this time that he learned of the invention of the magnetron tube as a means of generating extremely short wavelengths. The magnetron was immediately incorporated into Yagi's research.

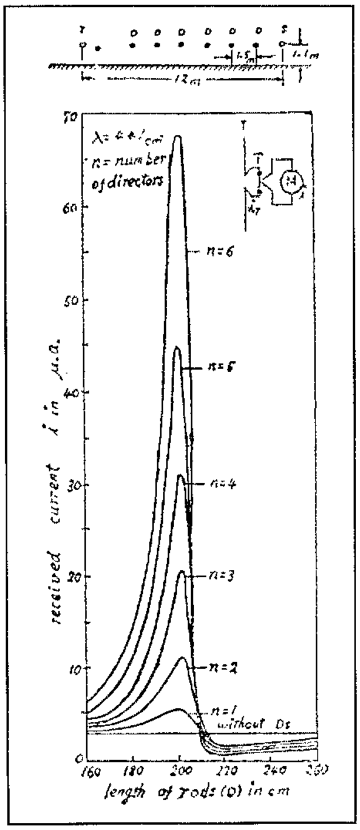

Fig 1—The effect of varying the number and length of directors in Wave Canals on received current. (What we call a "Yagi," Yagi called a Wave Canal.)

The diagram in the upper right shows a resonant antenna element bridged by an RF ammeter. This simple, yet effective, device was used by Yagi and Uda to investigate and construct directive antennas at VHF/UHF.

Note: All figures are photographed from Yagi's landmark article, "Beam Transmission of Ultra Short Waves," in Volume 16 of the *Proceeding of the Institute of Radio Engineers*, June 1926.

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[1]Notes appear on page 47.

Yagi's Students

One of Yagi's students, Kinjiro Okabe, was able to generate even shorter wavelengths (a few tens of centimeters) and produce greater output power by splitting the magnetron anode into segments—the invention of the split-anode magnetron.[2]

Another student, Shintaro Uda, experimented with vacuum-tube triode oscillators at wavelengths of about 4.4 meters and formed the waves into a directional beam with an antenna now known as the Yagi or Yagi-Uda.

Uda began his antenna studies by measuring the radiation pattern of a single-wire resonant loop. True to the experimenter's spirit, he tried to improve the directive pattern by placing a parasitic loop, tuned below resonance, near the driven loop. Next he replaced the parasitic loops with parasitic rods, finding that the field strength increased with increasing numbers of parasitic rods. Does this sound familiar? This was the birth of the Yagi-Uda antenna!

Uda's antenna went from a loop to a dipole configuration, with the dipole and parasitic rods vertically polarized. What followed was an in-depth analysis of the variables that controlled directivity such as parasitic element lengths, spacing and geometric arrangement of parasitic elements, the effect of receiving antenna height, the effect of transmitting antenna height and the effect of varying the number of directors.

Uda's experimental method was simplicity itself. He used a dipole receiving antenna to which he connected a crystal detector and an ammeter to record the current as a function of the variable of interest. He plotted the results in graphical format that looks identical to the data plotted in any antenna engineering textbook today, and in *QST*.

Uda found the greatest gain with a reflector of about one-half wavelength in length and spaced about a quarter wavelength from the driven element, with director lengths of 10% less than a half-wavelength spaced at about a third of a wavelength.

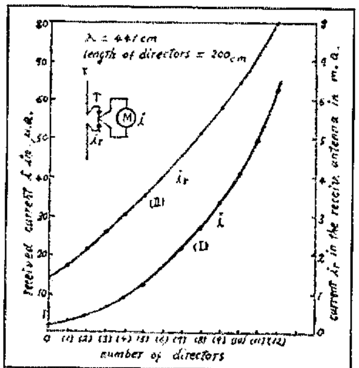

Fig 2—The effect of varying the number of directors on received current.

These spacings and lengths compare favorably with modern commercial and home-brew antennas. The experiments were performed at 4.4 (68 MHz) and 2.6 meters (115 MHz). In those days, wavelengths shorter than 10 meters were considered to be "ultra short waves."

The first publication to come out of Uda's research appeared in the *Proceedings of the Imperial Academy of Japan* in February of 1926. Entitled "Projector of the Sharpest Beam of Electric Waves," this paper contained a theoretical analysis of a vertical dipole with a parasitic rod. In the next publication in the *Journal of the IEE of Japan*, March 1926, Uda described an experiment in which he used three parasitic rods in close proximity to a driven half-wave dipole. He suggested that this arrangement be called a "trigonal reflector" as a transmitting antenna and a "trigonal collector" as a receiving antenna.

Additional publications followed, including one in the *Proceedings of the IRE* in March of 1927. In this paper, entitled "High Angle Radiation of Short Electric Waves," Uda fully described the new antenna, including the terminology used to describe its operation.

For example, following the description of the trigonal reflector, Uda stated that the addition of a number of parasitic directors in line with the reflectors and driven element "remarkably improved" the antenna's directivity. He went on to suggest that a grouping of directors be called a "wave canal" and a combination of the wave canal and trigonal reflector be called a "wave projector."

Yagi had applied for a patent on this antenna on December 28, 1925. The patent, "Denpa-shikou-houshiki," was issued as number 69115 the following year. Yagi then applied for a US patent in September of 1926, and on May 24, 1932, he was issued patent number 1,860,213: "Variable Direction Electric Wave Generating Device." Yagi assigned the US patent to the Radio Corporation of America (RCA), where it became the basis for all VHF television receiving antennas after the war.

Publication of a Classic

In 1928, Yagi again visited the US, giving speeches to IRE members in New York City, Washington, DC, and Hartford, Connecticut. He traveled to Schenectady to speak at the GE Research Laboratory, where he described Okabe's split-anode magnetron. His disclosures on the antenna work and the split-anode magnetron led GE's radio engineers to construct a 400-MHz magnetron and a Yagi-Uda antenna system in the summer of 1928—which was probably the first use of this system in the US.

At about the same time, Yagi contributed a paper to the IRE for publication in its *Proceedings*. This paper, entitled "Beam Transmission of Ultra Short Waves" was published in June of 1928.[3] It is a classic scientific paper.

In Part I of the paper, Yagi describes the mechanics of the antenna through text, drawings and graphs. He explains the testing procedures and the results of each test. Included in this paper is a brief discussion of the effect of the earth on the transmitted energy—a topic recently discussed in the pages of *QST*. Part II describes the magnetron oscillators that were developed to generate the CW signals. This part also contained text, drawings and graphs showing how the magnetron operated.

Remembering that this was published in 1928, the shortest wavelength claimed to be generated was 12 cm (2.5 GHz); a second magnetron generated 19-cm waves (1.6 GHz). In today's terminology, these signals are microwaves.

In the "Introduction," Yagi clearly ascribes the success of the split-anode magnetron to his student Kinjiro Okabe, and in the first sentence of "Part I," he ascribes the 4.4-meter antenna research to "Mr S. Uda, assistant professor at the Tohoku Imperial University." Professor Yagi ends this paper with the following statement:

"The experiments described in Part I were made by S. Uda, and those of Part II by Mr K. Okabe, to the ingenuity of both of whom the successful development of the beam system is mainly due."

Yagi's paper is then followed by a brief discussion written by J. H. Dellinger, Chief of the Radio Division, US Bureau of Standards. Dellinger comments on the impact that "directive radio" will have on communications, closing with the sentence "...I have never listened to a paper that I felt so sure was destined to be a classic."

The research group at Tohoku University continued its radio experiments in the form

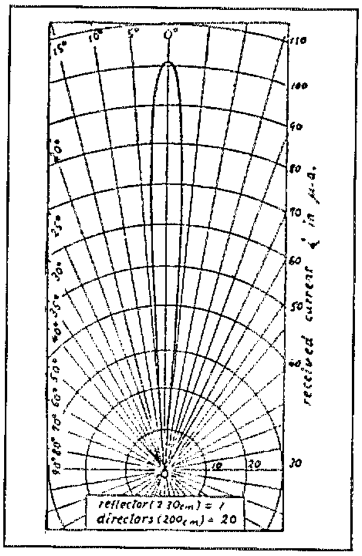

Fig 3—Beam radiation from a radiator using a wave canal. Note the similarity to modern radiation-pattern display methods!

of propagation studies. Uda developed a regenerative detector for wavelengths of 40 to 80 cm (375 to 750 MHz). This new receiver, when used with the Yagi-Uda antenna, extended the operational range of early VHF/UHF radio systems from meters to kilometers.

In 1933, the new antenna came into its first commercial use in Japan. A government radio link was installed between Sakata and Tobishima Island over a distance of 40 kilometers (about 24 miles). The system used a radiotelephone operating at wavelengths of between 2 and 10 meters.

At about the same time, the Fisheries Research Institute at Watanoha considered the use of this system between fixed land stations and shipboard stations. For this purpose, Uda's research group investigated the effects of antenna polarization, transmitter power and the screening formed by small islands on the success of the communications link.

The commercial success of the Yagi-Uda antenna was ensured by the wide introduction of television after WW2. In 1951, Professor Yagi acted as a consultant to the Japanese government on the formulation of television standards. The Yagi Antenna Company, Ltd, was organized with Professor Yagi as its president. This antenna has become the de-facto standard television antenna around the world.

Oddly enough, the Yagi-Uda antenna received very little attention by amateurs during the 1930s in *QST*. The only steerable, directive array that was popular during that time was the flat-top beam invented by Dr John Kraus, W8JK, of Ohio State University.

The Antenna Systems chapter of the 1943 *Radio Amateur's Handbook*[4] devotes five and a half pages to directive-driven arrays, of which the flat-top beam is one type; and two and one-half pages to directive arrays with parasitic elements. The name Yagi was not used by the ARRL to describe the parasitic antenna.

According to Toshiki Komatsu, JF7WED/1, of Japan's University of Tsukuba, Professor Yagi held the call sign J7AA prior to WW2 and was a member of the JARL. It's amazing that it took the introduction of television to bring this antenna into Amateur Radio.

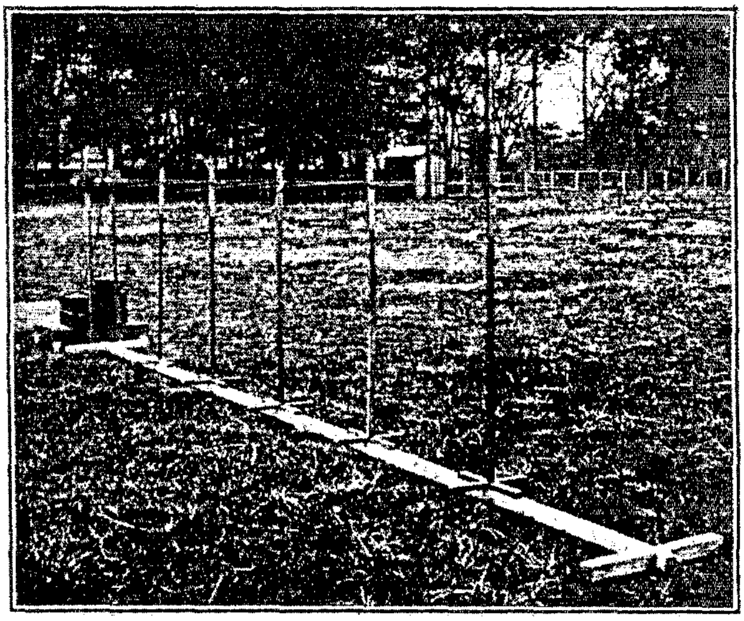

Fig 4—A 7-element "Yagi" for a wavelength of 260 cm (115 MHz). Note the wooden supports and the positioning track on the ground.

Epilog

Professor Yagi's home and personal library, which included 30 years of research data, were destroyed in the fire-bombing of Tokyo in April of 1945. After retiring as Emeritus Professor of Tohoku University and Osaka University, he remained active as a special advisor to the Yagi Antenna Company, the Japan television broadcasting network, the Tokyo Express Electric Railway and various scientific councils and government agencies. He was awarded Japan's Order of Cultural Merit by Emperor Hirohito.

Professor Uda continued his research into the Yagi-Uda antenna after WW2, coauthoring a textbook with Y. Mushiake in 1954 entitled *Yagi-Uda Antenna*. During the 1950s, he performed research on the traveling-wave tube until his retirement from Tohoku University in 1960. This work included several years (1955-1958) in India doing research on microwave communications for UNESCO.

Those of you who are interested in microwave propagation would do well to read Uda's experiments.[5] He investigated path profiles, diffraction losses, day versus night signals, fading and threshold signals. He operated in the 2-GHz range at distances up to 224 kilometers (134 miles).

Ever the tireless researcher, Uda joined the faculty of Kanagawa University in 1960 to study the ruby laser at a time just after its invention in the US.

Having built numerous Yagi-Uda antennas in my 33 years as a radio amateur, I gave little thought to Yagi and didn't even know of Uda. As an electronics and physics teacher of 20 years, however, I deeply appreciate their use of the scientific method. As amateurs, we owe both of them a great debt of gratitude for what they have given our hobby.

Let me suggest that the next time you describe your station to someone, don't tell them that your antenna is a "beam"; tell them you're using a Yagi-Uda.

Thanks

I would like to thank those on packet and on the Internet who responded to my questions. Special thanks to Frank Donovan, W3LPL; Professor James E. Brittain, PhD, Georgia Institute of Technology; and Professor Gentei Sato PhD, Sofia University in Tokyo, for their gracious assistance.

Notes

[1] J. E. Brittain, PhD, "The Yagi-Uda Antenna," *Proceedings of The Radio Club of America*, Vol 58, No. 2, Nov 1985.

[2] W. A. Atherton, "Pioneers," *Electronics and Wireless World*, Jan 1989.

[3] H. Yagi, "Beam Transmission of Ultra Short Waves," *Proceedings of the IRE*, Vol 16, No. 6, Jun 1928.

(Reprinted in *Proceedings of the IEEE*, Vol 72, May 1984.

[4] *The Radio Amateur's Handbook*, American Radio Relay League, West Hartford, CT, 1943, pp 184-190.

[5] S. Uda, *Short Wave Projector: Historical Records of My Studies in Early Days* (Copyright 1974 by Shintaro Uda). (From a copy provided by the Uda family.)

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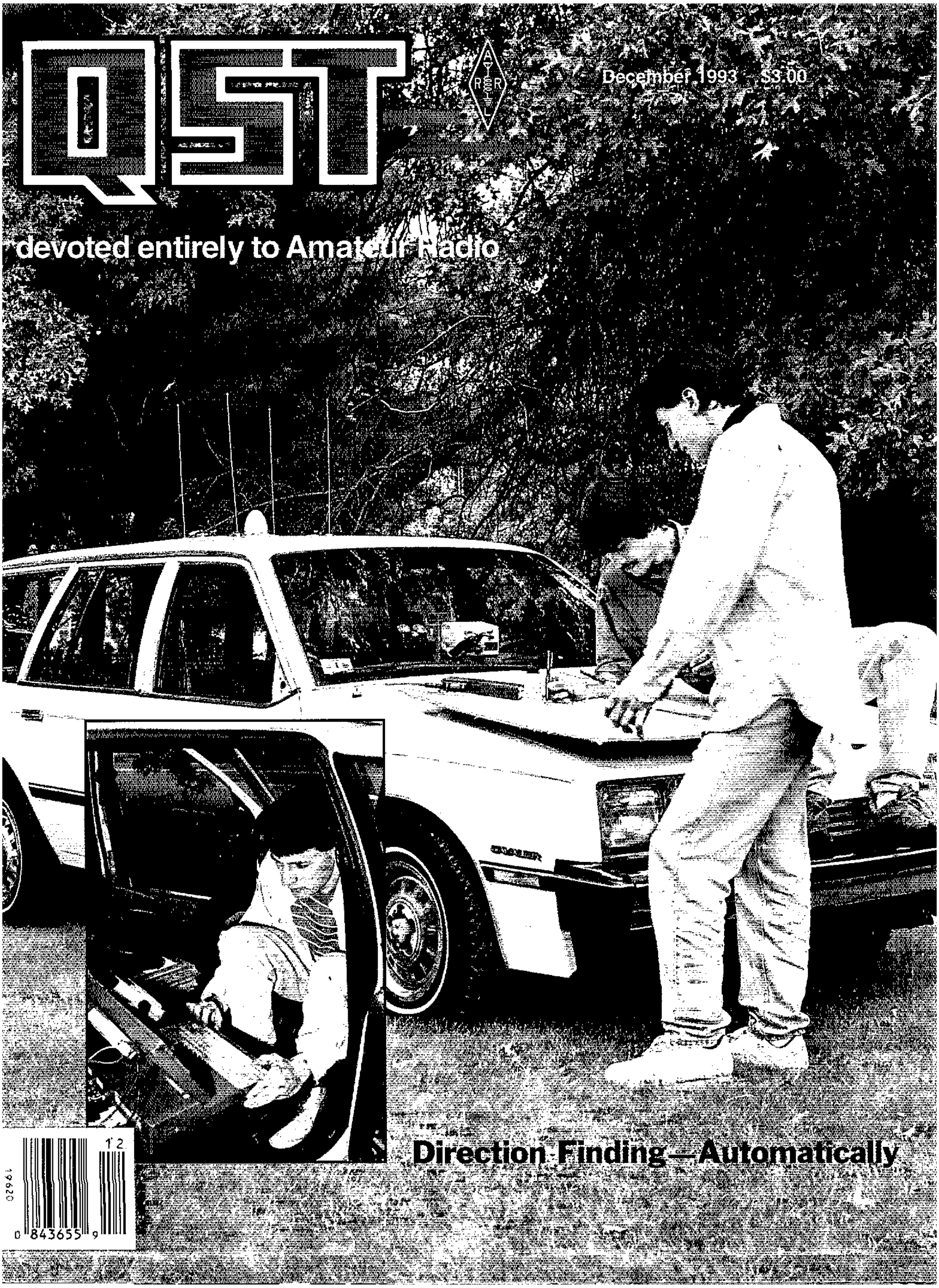
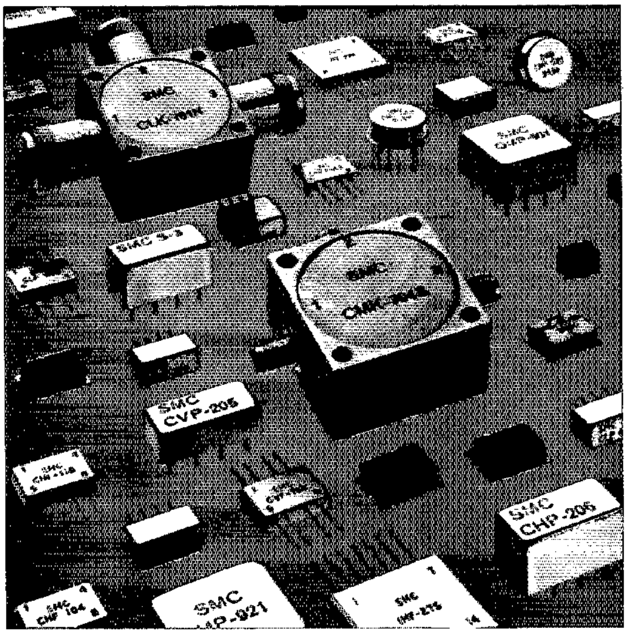


Direction Finding—Automatically



(photos courtesy of Synergy Microwave)

Taking the Mystery Out of Diode Double-Balanced Mixers

Here's a look into the workings of a common signal-processing block many modern telecommunication systems couldn't do without.

By Shankar Joshi, Chief Engineer
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Double-balanced mixers (DBMs) are common in modern communication and signal-processing systems. They can serve as mixers (including image-reject types), modulators (including single- and double-sideband, phase, biphase, and quadrature-phase types) and demodulators, limiters, attenuator switches, phase detectors, and frequency doublers. In some of these applications, they work in conjunction with power dividers/combiners and hybrids.

A DBM type commonly used in Amateur Radio and industry consists of little more than a diode ring and coupling transformers. Radio amateurs have been using such DBMs in homemade and commercial forms for at least 25 years, yet these DBMs' characteristics remain mysterious to some. This article explores these mixers' construction, characteristics and application.

How Diode Double-Balanced Mixers Look and Work

The Basic DBM Circuit

A diode double-balanced mixer (Fig 1) contains two or more unbalanced-to-balanced transformers and a diode ring consisting of 4 × *n* diodes, where *n* is the number of diodes in each leg of the ring. Each leg commonly consists of up to four diodes. Input and output *ports*—commonly named local oscillator (*LO*), radio frequency (*RF*) or *signal*, and intermediate frequency (*IF*)—connect the DBM to its associated circuitry.

Basic DBM Function

See Fig 1. Applied to the diode ring's D4-D1 and D2-D3 terminals via T1, a 1:4 transformer, LO energy forces D1 and D2 into full conduction on each positive half cycle of the LO waveform. D3 and D4 conduct on each negative half cycle. T2, also a 1:4 transformer, applies the RF-port signal to the ring's D1-D2 and D3-D4 terminals. The LO (f_{LO}) switches the RF signal (f_{RF}) to the IF port at the LO frequency, resulting in RF/LO frequency *multiplication*. The resultant IF-port signal includes two dominant components, $f_{LO}+f_{RF}$ and $f_{LO}-f_{RF}$, in addition to harmonic products of f_{LO} and f_{RF}.

The degree to which a mixer is *balanced* depends on whether either, neither or both of its input signals (RF and LO) emerge from the IF port along with mixing products. An unbalanced mixer suppresses neither its RF nor its LO; both are present at its IF port. A *single-balanced* mixer suppresses its RF or LO, but not both. A *double-balanced* mixer suppresses its RF *and* LO inputs. Diode and transformer uniformity in the Fig 1 circuit results in equal LO potentials at the center taps of T1 and T2. The LO potential at T1's secondary is zero (ground); therefore, the LO potential at the IF port is zero.

Balance in T2's secondary likewise results in an RF null at the IF port. The RF potential between the IF port and ground is therefore zero.

The Fig 1 circuit normally also affords high RF-LO isolation because its diode

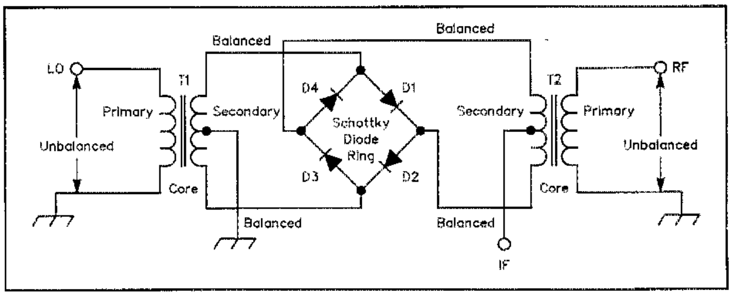


Fig 1—A diode double-balanced mixer consists of two unbalanced-to-balanced transformers and a Schottky-diode ring.

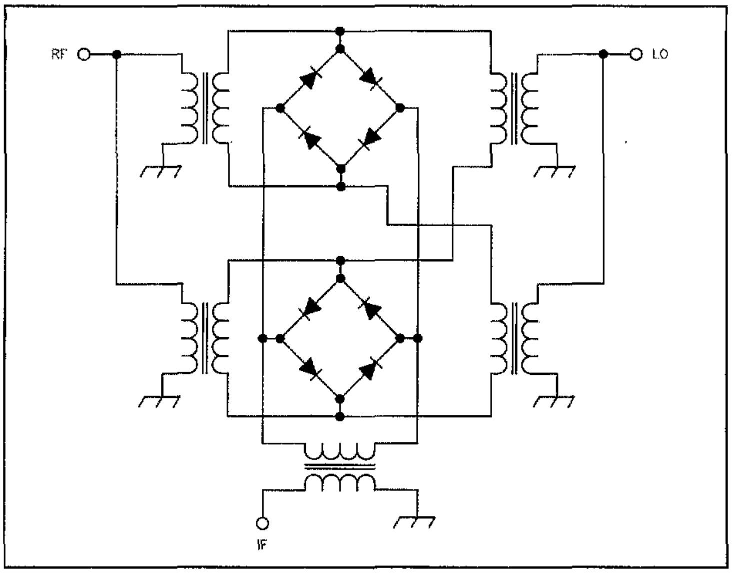


Fig 2—Five transmission-line transformers and two Schottky-quad rings form this double double-balanced mixer (DDBM). Such designs can provide lower distortion, better signal-handling capability and higher interport isolation than single-ring designs.

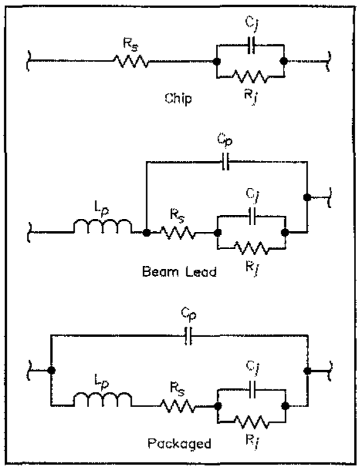


Fig 3—Its semiconductive properties aside, a Schottky diode can be represented as a network consisting of resistance, capacitance and/or inductance. Of these, the junction capacitance (C_j) plays an especially critical role in a double-balanced mixer's high-end response. (R_j = junction resistance, R_s = contact resistance, L_p = parasitic inductance and C_p = parasitic capacitance.)

Table 1
Typical Parameters of Schottky Diodes

Part	*Typ V_f @ 1 mA (mV)*	*Max C_j @ 0 Volts (pF)*	*Max VΔ @ 5 mA (mV)*	*Typ R_S @ 5 mA (Ohms)*	*Approx Max Operating Frequency (GHz)*
MP2400	290	0.10	15	20	35
MP2401	290	0.10	15	20	40
MP2402	290	0.15	15	20	26
MP2403	290	0.20	15	20	18
MP2407	270	0.20	15	15	18
MP2410	240	0.25	15	10	18
MP2412	240	0.30	15	10	12
MP2414	240	0.45	15	10	8
MP2417	220	0.55	15	8	6
MP2418	220	0.60	15	8	4

courtesy of M-Pulse Microwave

switching precludes direct connections between T1 and T2. A diode DBM can be used as a current-controlled switch or attenuator by applying dc to its IF port. This causes opposing diodes (D2 and D4, for instance) to conduct to a degree that depends on the current magnitude, connecting T1 to T2.

One extension of the single-diode-ring DBM is a *double* double-balanced mixer (DDBM) with high dynamic range and larger signal-handling capability than a single-ring design. Fig 2 diagrams such a DDBM, which uses transmission-line transformers and two diode rings. This type of mixer has a higher 1-dB compression point (usually 3 to 4 dB lower than the LO drive) than a DBM. Low distortion is a typical characteristic of DDBMs. Depending on the ferrite core material used, frequencies as low as a few hundred hertz and as high as a few gigahertz can be covered.

Diode Double-Balanced-Mixer Components

Commercially manufactured diode DBMs generally consist of: (A) a support base on which the unit is constructed; (B) a diode ring; (C) two or more ferrite-core transformers wound with transmission lines, two or three twisted-pair wires, or a combination of both; (D) encapsulating material; and (E) an enclosure.

Diodes

Hot-carrier (Schottky) diodes are the devices of choice for diode-DBM rings. The forward voltage drop, V_f, across each diode in the ring determines the mixer's optimum local-oscillator drive level. Depending on the forward voltage drop of each of its diodes and the number of diodes in each ring leg, a diode DBM may be categorized as a 0, +3, +7, +10, +13, +17, +23 or +27 dBm-level mixer.

Schottky diodes are characterized by loss and contact resistance (R_s), junction capacitance (C_j), and forward voltage drop (V_f) at a known current, typically 1 mA or 10 mA. The lower the diode-to-diode V_f difference in millivolts, the better the diode match at dc. (Some earlier diode DBM designs used diodes in series with a parallel resistor/capacitor combination for automatic biasing.) Better diode matching (in V_f and C_j) results in higher isolation among the ports. Diodes capable of operating at higher frequencies have lower junction capacitance and lower parasitic inductance. Fig 3 shows the equivalent circuit for Schottky diodes of three package types.

Manufacturers of diodes suitable for DBMs characterize their diodes as low-barrier, medium-barrier, high-barrier and very-high-barrier (usually two or more diodes in each leg), with typical V_f values of 220 mV, 350 mV, 600 mV and 1 V or more, respectively. Fig 4 shows a typical current-voltage (I-V) characteristic for a low-barrier Schottky quad capable of operating up to 4 GHz. Note that as current through the diodes increases, the V_f difference among the ring's diodes also increases, affecting the balance.

At higher frequencies, diode packaging becomes critical and expensive. As the frequency of operation increases, the effect of junction capacitance and package capacitance cannot be ignored. Part or all of the capacitance can be compensated at the mixer's highest operating frequency by properly designing the unbalanced-to-balanced transformers. The transformer inductance and diode junction capacitance form a low-pass network with its cutoff frequency higher than the frequency of operation. Compensated in this way, diodes with a junction capacitance of 0.2 pF can be used up to 8 to 10 GHz.

Table 1 shows typical parameters for Schottky ring diodes.

Transformers

From the DBM schematic shown in Fig 1, it is clear that the LO and RF transformers are unbalanced on the input side and balanced on the diode side. The diode ends of the balanced ports are 180° out of phase throughout the frequency range of interest. This property causes signal cancellations that result in higher port-to-port isolation. Fig 5 plots LO-RF and LO-IF isolation versus frequency for Synergy Microwave's CLP-4A3 DBM. Isolations on the order of 70 dB occur at the lower end of the band as a direct result of the balance among the four diode-ring legs and the RF phasing of the balanced ports.

Transformer efficiency plays an important role in determining a mixer's conversion loss and drive-level requirement. Core loss, copper loss and impedance mismatch all contribute to transformer losses.

Ferrite in toroidal, bead, balun (multihole), or rod form can serve as DBM transformer cores. Balun and toroidal cores are most commonly used, and are available from TDK, Siemens, Ferroxcube, Krystinel and other vendors.

Before selecting a material type (or magnetic permeability, µ), its temperature coefficient and Curie temperature (the temperature at which the core loses its magnetic properties) must be considered. In some cases, µ may change drastically across the desired temperature range; this would cause a transformer's frequency response to shift with temperature. Once a suitable core material and form have been selected, the transformer's size and frequency requirements determine the necessary core size.

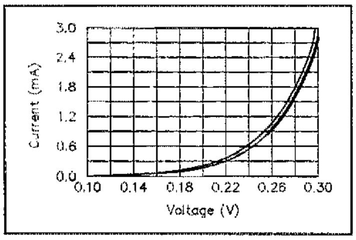


Fig 4—Current-voltage (I-V) characteristic for Schottky diode quad, showing worst-case voltage imbalance (the spread between the two curves) among the four diodes.

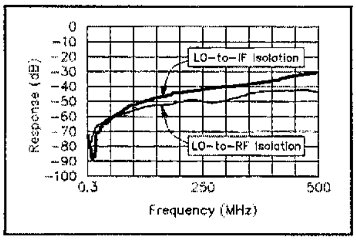


Fig 5—A diode DBM's port-to-port isolation depends on how well its diodes match and how well its transformers are balanced. This graph shows LO-IF and LO-RF isolation versus frequency for a Synergy Microwave CLP-4A3 mixer.

For a given core shape and size, the number of turns, wire size, and the number of twists determine transformer performance. Wire placement also plays an important role. The wires' insulation thickness determines the isolation between the wire and the core. Care must be taken to ensure that the insulation does not break during tight winding or due to mishandling. Sharp-edged cores should be avoided; these can break insulation and nick wires. Subjected to repeated hot-cold flexing, nicked wires may fail open.

RF transformers combine lumped and distributed capacitance and inductance. The interwinding capacitance and characteristic impedance of a transformer's twisted wires sets the transformer's high-frequency response. The optimum number of turns necessary to achieve best performance at the upper end of a transformer's operating range differs with the core type used. The core's µ and size, and the number of winding turns, determine the transformer's lower frequency limit. Covering a specific frequency range requires a compromise in the number of turns used with a given core. Increasing a transformer's core size and number of turns improves its low-frequency response. Many times, cores may be stacked to meet low-frequency performance specs.

Inexpensive mixers operating up to 2 GHz most commonly use twisted trifilar (three-wire) windings made of a wire size between #36 and #32. The number of twists per unit length of wire determines a winding's characteristic impedance. Twisted wires are analogous to transmission lines and can be analyzed in terms of distributed interwinding capacitance. Decreasing the number of twists lowers the interwinding capacitance and increases the frequency of operation. If the wists per inch are fewer than four, handling becomes difficult.

The transmission-line effect predominates at the higher end of a transformer's frequency range. If two impedances, Z_1 and Z_2, need to be matched through a transmission line of characteristic impedance, Z_0, then

$$Z_0 = \sqrt{Z_1 \times Z_2} \qquad \text{Eq 1}$$

Fig 6 shows two types of transformers using twisted wires: (a) a three-wire type in which the primary winding is isolated from the secondary winding with a center tap, and (b) a two-wire (transmission-line) type in which two sets of transmission lines are interconnected to form a center tap at the secondary with a direct connection between primary and secondary. The primary-secondary turns ratio determines the impedance match, as shown in Eq 1. The properties of these two transformer types can be summarized as follows:

1. By virtue of its construction, the three-wire transformer is more unsymmetrical at higher frequencies than the transmission-line type.
2. The transformers' lower cutoff frequency (f_L) is determined by the equation

$$\omega L > 4R \qquad \text{Eq 2}$$

where

L = inductance of the winding
R = system impedance (for example, 50 Ω, 75 Ω, etc) and
$\omega = 2\pi f L$

3. The transmission-line transformer's upper cutoff frequency (f_H) is determined by the highest frequency at which its wires' twists (that is, the coupling between them) allow it to function as a transmission line of the proper characteristic impedance.
4. Transformers convert one impedance, Z_1 (primary) to another, Z_2 (secondary) according to the relationship

$$Z_2 = Z_1 (N)^2 \qquad \text{Eq 3}$$

where

N = secondary-to-primary turns ratio

Within certain limits, if Z_1 is varied, Z_2 also varies to a new value multiplied by N^2. Thus, a mixer designed for a 50-Ω system may work in a 75-Ω system with minor modifications.

Support Base

The base serves as a support for all of the mixer's components and connecting leads

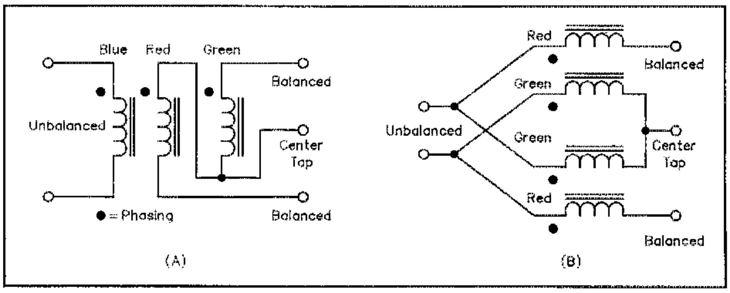


Fig 6—Transformers for DBMs: three-wire (A), and transmission-line (B).

while providing whatever mechanical strength the application requires. The base can be made of plastic (or any other insulator) or metal.

Plastic packages are commonly used in cost-sensitive systems. Plastic packages are inexpensive, lightweight, can be easily molded, and can be formed into complicated shapes. Offsetting these advantages, they are unsuitable for very high frequency operation, susceptible to cracking, easily destroyed by chemicals, non-hermetic, poorly resistant to heat and mechanically weak. Further, the tendency of plastic packages to warp and crack limits their maximum size.

Metal packages can be made from a variety of metals and alloys, such as brass, copper, cold rolled steel, Kovar and aluminum. (Different applications often require that specific platings be applied to these base metals.) Glass-to-metal or ceramic-to-metal seals transfer the mixer-port leads through the package. Metal mixer packages can be hermetically sealed. Hermetic packages are suitable for harsh environments and are extensively used in military applications.

Another advantage of metal packaging is that it serves as a conductive (and therefore shielding) plane that resists electromagnetic interference (EMI) and radio-frequency interference (RFI). Unlike plastics, metal packages do not warp, and need not be restricted in size for this reason.

Encapsulating Material

Commercial mixers are often encapsulated—potted—to be sure that their components cannot be dislodged during vibration, impact, thermal stress and so on. Many different encapsulants are available on the market; however, some of them crack when heated. Each of these products is unique in the sense of hardness, weight, dielectric constant, dielectric loss, thermal conductivity, elasticity, adhesion to different surfaces, chemical resistance, fungus resistance, brittleness and so on, so they must be evaluated carefully before specification. Some encapsulants can act as a heat sink to remove component heat.

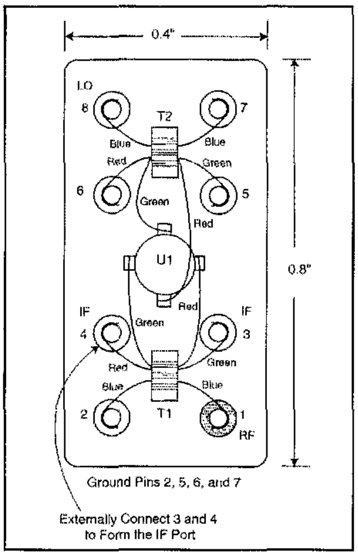


Fig 7—How a typical commercial DBM is wired. The use of different wire colors for the transformers' various windings speeds assembly and minimizes error. U1, a Schottky-diode quad, contains D1-D4 of Fig 1.

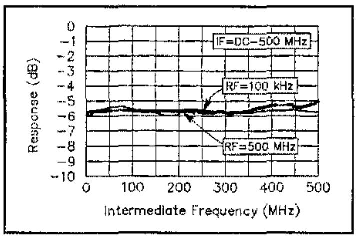


Fig 8—Conversion loss versus frequency for a typical diode DBM. The heavier curve shows the loss for RF = 100 kHz; the lighter, for RF = 500 MHz. The LO drive level is +7 dBm.

Diode DBMs in Practice

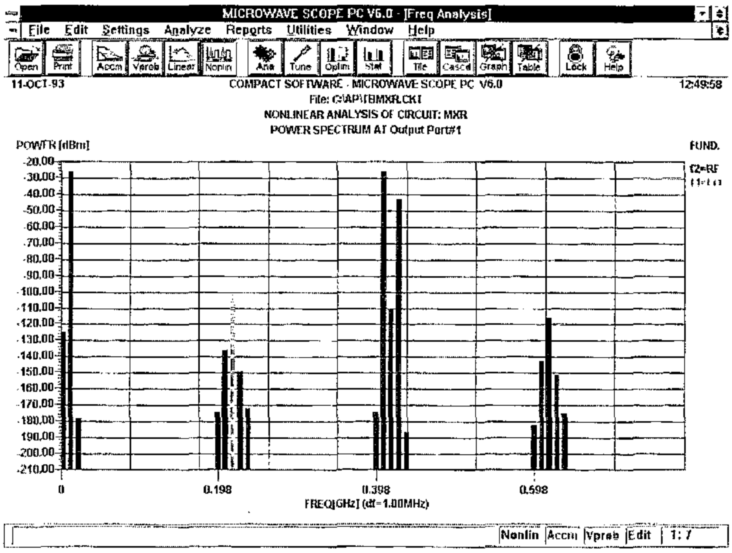
Fig 7 shows the wiring of a typical DBM that uses toroidal cores. The wires are wrapped around the package pins and diode leads, and then soldered. In this unit, the primary winding of the LO transformer connects across pins 7 and 8; the RF-transformer primary, across pins 1 and 2. The pin pairs 3-4 and 5-6 are connected externally to form the transformers' secondary center taps, one of which (5-6, that of the LO transformer) connects to a common ground point while the other (3-4, that of the RF transformer) serves as the IF port.

The DBM shown in Fig 7 has a dc-coupled IF port. If necessary, this DBM can be operated at a particular polarity (positive or negative) by appropriately connecting the LO, RF, IF and common ground points.

DBMs and DDBMs are available in plug-in and surface-mount packages. Plug-in packages commonly come in metal cans measuring (for eight-pin packages) 0.8 inch (length) × 0.4 inch (width) or (for eight- or four-pin packages) 0.5 inch × 0.25 inch, with heights ranging from 0.25 to 0.4 inch. Flat-packs measuring 0.51 × 0.385 × 0.15 inch are metal packages, classified as surface-mount types. Synergy's surface-mount DBMs and DDBMs are available in leaded and non-leaded versions, and come in two configurations: ceramic base, with a ceramic or plastic cover, and glass-epoxy base, with a plastic cover. The Synergy surface-mount line includes a patented hermetic package with excellent EMI/RFI suppression.

DBM Specification

The typical parameters involved in selecting a diode DBM are: (A) conversion loss and its flatness (ripple) across the required IF bandwidth; (B) variation of conversion loss with RF frequency; (C) variation of conversion loss with LO drive, (D) 1-dB compression point; (E) LO-RF, LO-IF and RF-IF isolation; (F) intermodulation products;



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NONLINEAR ANALYSIS OF CIRCUIT: MXR
POWER SPECTRUM AT Output Port#1

Fig 9—Simulated diode-DBM intermodulation distortion. Note that the desired output products (the highest two products, RF–LO and RF+LO) emerge at a level 5 dB below the mixer's RF input (–20 dBm). This indicates a mixer conversion loss of 5 dB. (*Microwave SCOPE* simulation)

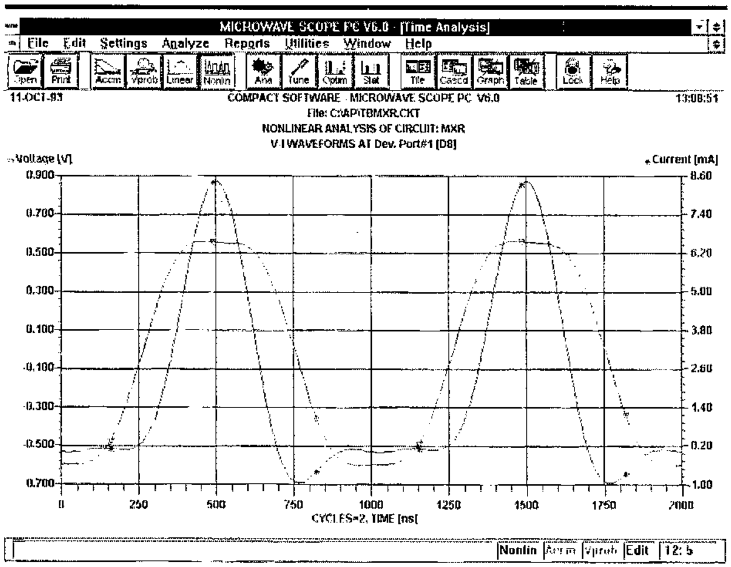


Fig 10—Sampled current and voltage waveforms across one of the diodes in a DDBM ring reflect the mixer's nonlinear action. (*Microwave SCOPE* simulation)

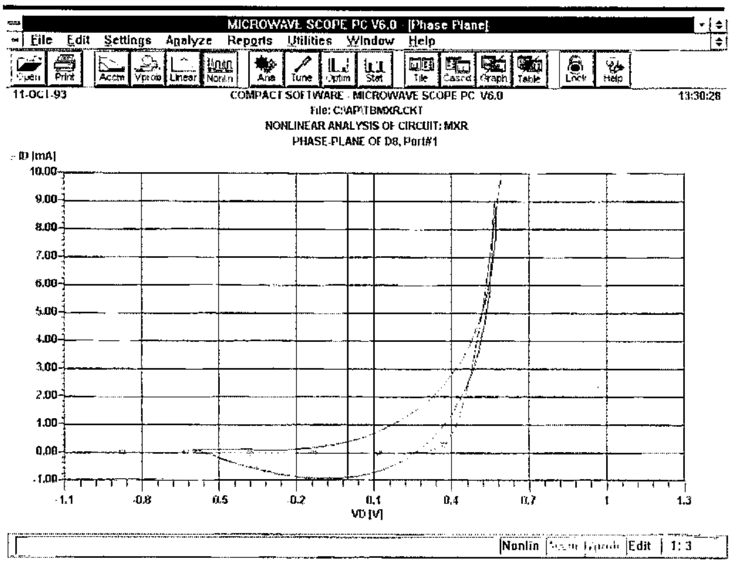
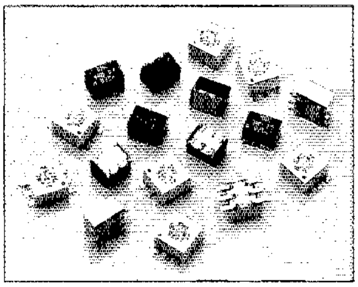


Fig 11—Phase plane of the one of the diodes in a DDBM ring. The curves show the diode's dc and dynamic load lines across the operating range, and reflect the effect of energy storage in the diode's junction capacitance. (*Microwave SCOPE* simulation)



Many DBM applications call for surface-mount devices. These are Synergy Microwave SMD mixers.

(G) noise figure (usually within 1 dB of conversion loss); (H) port SWR; and (I) dc offset, which is directly related to isolation among the RF, LO and IF ports.

Fig 7 shows a plot of conversion loss versus intermediate frequency in a typical DBM. The curves show conversion loss for two fixed RF-port signals, one at 100 kHz and the another at 500 MHz, while varying the LO frequency from 100 kHz to 500 MHz.

To give some additional insight, Fig 9 plots a diode DDBM's simulated intermodulation distortion. Note that the RF input is −20 dBm and the IF output (the frequency difference between the RF and LO signals) is −25 dBm, implying a conversion loss of 5 dB. This figure also applies to the *sum* of both signals (RF + LO).

Fig 10 shows the waveforms across one of the diodes in a DDBM ring. Note the signal distortion, which indicates high harmonic content. Despite this distortion, the mixer's output is relatively clean because of the cancellation due to balance in the mixer's diodes and transformers.

Fig 11 shows the diode's phase plane. It indicates the RF swing across the diode based on the energy stored in the diode's capacitance.

Conclusion

I hope that this article sheds some light on the workings of diode double-balanced mixers, which commonly serve as signal-processing building blocks in Amateur Radio and industry. Whether you build your own DBMs or buy them ready-made, their design, specifications and construction need not be mysterious.

Acknowledgments

I thank the following individuals for their contributions to this project, without whose assistance this effort would have been unsuccessful: Dr. Ulrich L. Rohde—invaluable advice and counsel; Mike Vogas—technical expertise and creation of the drawings; and Merle L. Cioban-Chandler—word-processing support.

Shankar Joshi, Chief Engineer for Synergy Microwave, is responsible for overall product design. Previously, Mr. Joshi was employed with Polarad, on Long Island, a subsidiary of Rohde & Schwarz, where he was responsible for the design of microwave spectrum analyzers. Mr. Joshi has an MSEE from Indian Institute of Technology, Bombay, India, and over 20 years of design experience in RF and microwave transceivers, and components. QST

Measuring and Compensating Oscillator Frequency Drift

How stable is the VFO in your rig? A frequency counter, a simple environmental chamber, and a home-built electronic thermometer can allow you to measure and even compensate the drift in your oscillators.

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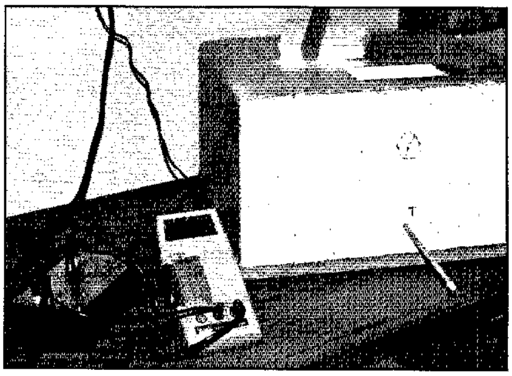


(photos by W7ZOI)

An oscillator's *stability* characterizes its relative freedom from undesired frequency change. Many factors can cause unwanted frequency shift. Variations in loading (termination) can cause *pulling*. Altering the oscillator power supply voltage can cause *pushing*. Circuit noise can modulate an oscillator, leading to *phase-noise* sidebands.

The most common form of oscillator instability, however, is that related to temperature. The inductors, capacitors, crystals, and/or transmission lines that determine oscillator frequency have values that depend upon temperature. Temperature changes arise either from internal heating, which causes *warm-up drift*, or from changes in the environment surrounding the circuitry. Both of these are called *thermal drift*.

Variable frequency oscillator (VFO) drift has always been a topic of vital concern to the radio amateur. Articles on the subject abound, with each author presenting his or her own recipe for achieving oscillator stability. Some of this lore is well-founded—that is, based on careful observation and sound thinking, and correspondence between the two. Other contributions depart from science. The collective result is confusing, if not chaotic. Believing the lore may cause us to build ill-conceived circuits while we ignore others that might offer wonderful stability.

VFO design is an ever-evolving pursuit. Internal heat generation was the major problem we fought with vacuum-tube oscillators. Although less severe, these thermal effects persist with solid-state oscillators. A more recent difficulty adds further complication: Some critical VFO components are no longer commercially available. Stable fixed-value capacitors and high-quality variable capacitors are relatively difficult to find. (Some fixed-ceramic capacitors advertised as having an NP0 temperature characteristic do not really qualify. More later on this.) The smooth, double-bearing "variables" used in our vintage rigs have been largely replaced by smaller, but more rugged and robust tuning diodes. The diodes are easy to use and offer wonderful mechanical stability, but they also tend to have terrible temperature characteristics and be lower-Q than the mechanical variables they replace. They can also generate noise.

Thermal stability problems can be solved through careful measurements. The requisite tools for oscillator measurement include a frequency counter and a simple environmental chamber with a controlled, measured temperature. An initial measurement (at power up) determines the oscillator's "cold" output frequency. Further measurements at the same chamber temperature quantify what warm-up drift occurs. Heating the chamber then adds the factor of rising ambient temperature. Further temperature and frequency measurements then characterize the oscillator's thermal stability.

Many experimenters already own frequency counters, so this article will describe a simple thermal chamber I built for oscillator evaluation. I'll also present several methods for temperature measurement—a vital part of the process—and show how to use the counter, temperature chamber and thermometer for oscillator evaluation.

Once the measurement tools are in place, the details of temperature compensation can be explored. I'll discuss some of the things that cause drift in oscillators, and how components are specified for temperature stability. I'll then present some examples that illustrate the compensation process.

Temperature Measurement

The ambient temperature in a ham shack is usually around 20 degrees Celsius (°C). A reasonably useful ham environmental chamber would produce temperatures from –20 to +70 °C. The chamber I'll describe is less ideal; it operates only at room temperatures and above. It is still nonetheless useful to be able to measure temperatures on either side of 20 °C.

A mercury thermometer, available from a chemical supply house, serves as an easy-to-use and traditional solution. For some experiments, I've used a Curtin Matheson Scientific Inc type CM-8 (catalog #248-422), which covers –20 to +150 °C. A hole in the chamber passes the thermometer. Most of the thermometer remains outside the chamber; its mercury bulb enters the chamber at the same height above the bottom of the chamber as the circuit under test.

A photographic shop may also carry a thermometer you can use: Photo thermometers, typically priced around $20, operate over a temperature range that includes most of the one we need. Some home thermometers may even work, although many lack useful resolution. The restricted range of medical thermometers makes them unsuitable for our purposes.

If you already have a digital multimeter (DMM), you can cheaply and easily build the ultimate solution, an electronic digital thermometer. (If you're an avid experimenter who doesn't have a DMM, put one on your "wish list.") Two possible circuits (Fig 1) use National Semiconductor's LM3911 IC temperature sensor. The simpler circuit, Fig 1A, provides a readout in kelvins. The LM3911 has an output of 10 mV/°C, so this circuit provides 0.1-°C resolution when used with a DVM capable of 1-mV resolu-

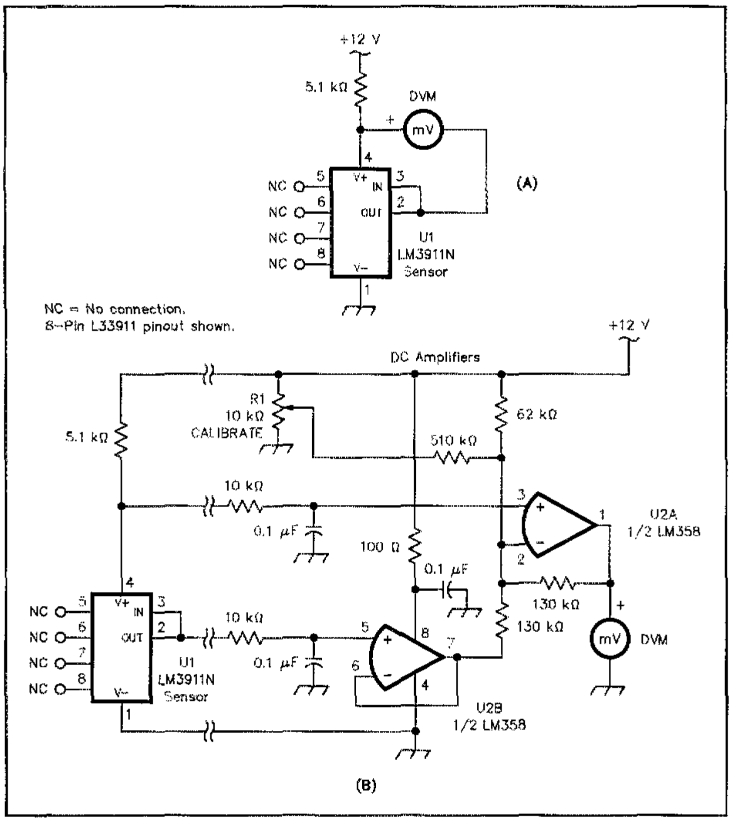
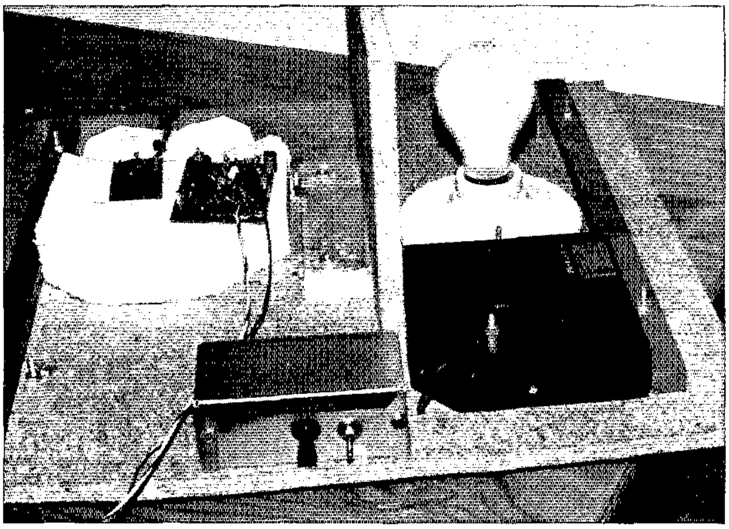


Fig 1—Two solid-state thermometers based on the LM3911 temperature sensor IC, one with readout in kelvins (A) and another (B) with readout in degrees Celsius. The circuit's fixed resistors are ¼-watt carbon-film; the capacitors, general-purpose ceramic. R1, **CALIBRATE**, is a trimmer potentiometer.



Inside the chamber, a vertical partition separates the incandescent-lamp heat source from the circuit under test. The switch box controls the lamp and exhaust fan.

tion at the 3-V level. Fig 1B shows a version with direct readout in degrees Celsius.

The TO-46 metal-can LM3911 I used in my thermometer needed a small heat sink to guarantee accuracy. Without it, self-heating in the sensor IC can generate errors of several degrees, even with the IC dissipating only 5 mW. An 8-pin plastic LM3911 (LM3911N) is available from Digi-Key for about $2. This plastic IC can be epoxied to a small, junk-box heat sink. Alternatively, the LM3911N may provide accurate results without a heat sink if its unused pins (5, 6, 7, and 8) are soldered to the surface where temperature is being measured. Some experiments may be required here.

I built the Fig 1B circuit in two sections. A small scrap of PC-board material carrying the LM3911 resides in the temperature chamber. A box away from the chamber contains the offset amplifier. There is nothing critical about the offset circuit and any construction form is suitable. R1, the **CALIBRATE** pot, should afford high resolution.

The electronic thermometer can be calibrated at room temperature using a chemical thermometer as a standard. Absolute temperature accuracy is not especially important for this application. What's more important is the ability to measure temperature *changes* with accuracy and repeatability.

The Chamber

The thermal chamber (see title photo) can be as simple or as elaborate as you would like it to be. The oven described here emphasizes simplicity, but it still provides a great deal of information.

A thermal chamber consists of a source of heat and an enclosure to spatially confine the heat. A 60-watt light bulb serves as the heat source. The only control is a toggle switch. The bulb resides in a ceramic socket mounted on a piece of fiberboard about 16 inches square (see photo this page). Another piece of fiberboard, placed vertically between the light bulb and the part of the chamber containing the circuit under test, prevents direct bulb radiation from reaching the circuit under test.

The chamber itself is a 9-inch deep Styrofoam box originally used as a shipping container for tropical fish. (At a supermarket, I located a suitable substitute box in the form of a Styrofoam picnic cooler. A usable substitute could be fabricated from plywood, or even cardboard.) The box is inverted and placed over the fiberboard base. Wood strips on the base locate the Styrofoam, simplifying the box's quick removal and replacement when inserting test circuits.

The box confines the light-bulb heat through the excellent insulation properties of Styrofoam. Lining the inside surface with aluminum foil would further improve the chamber's thermal performance by providing reflection as a second isolation mechanism. The metal would, however, act as a thermal load and slow the rate at which the

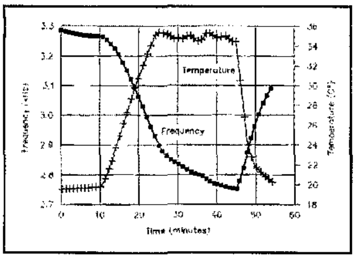


Fig 2—Results of a drift run made on the Ugly Weekender (JFET Hartley) VFO operating with its enclosure closed.

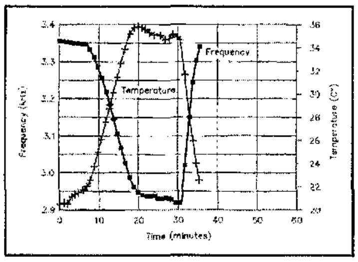


Fig 3—An Ugly Weekender drift run made with the transceiver's enclosure open. The quicker response of the VFO components to test-chamber temperature changes is clearly evident.

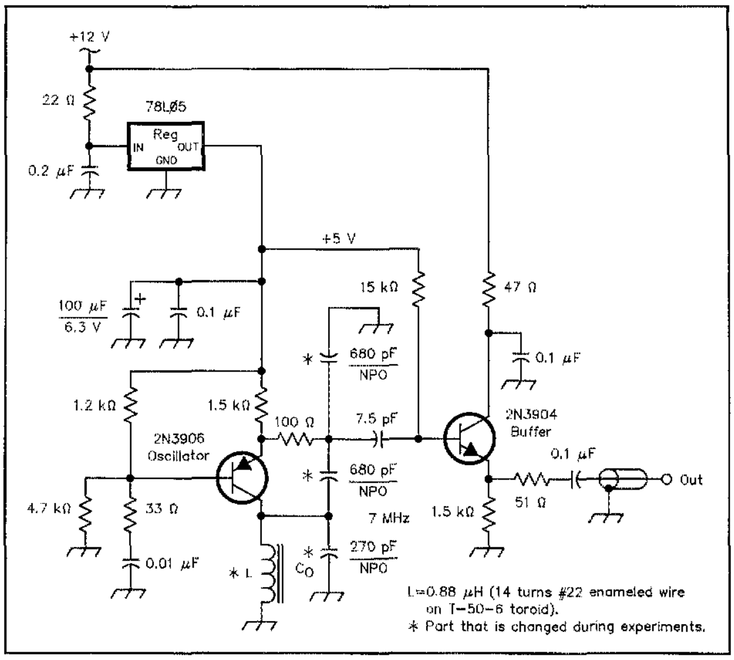


Fig 4—The 7-MHz bipolar-junction-transistor Colpitts VFO used for temperature-compensation experiments. Its resistors are ¼-watt carbon-composition; all of its nonpolarized capacitors not marked NPO are general-purpose ceramics.

chamber temperature can be changed.

My chamber also contains a small fan configured to exchange room and chamber air. This allows quick cooling of the chamber air to ambient temperature. Air exits through a small outlet hole in the Styrofoam. A Styrofoam plug normally closes a hole in the top of the box; removing the plug allows room air to enter when cooling is desired. (The cooling fan is a convenience and is not essential. The chamber would be effective without it.)

Another useful chamber refinement would be a small internal fan to circulate the air within the chamber. Such a fan must be carefully mounted to avoid inducing mechanical vibrations that might compromise oscillator performance.

Thermometer measurements indicated that the chamber temperature is higher at the top when the heat source is on. It's therefore useful to put the circuit under study on a small pedestal made from a piece of Styrofoam packing material. The thermometer should be close to, and at the same height as, the circuit under study.

Testing and Using the Chamber

My initial experiments with the chamber dealt with thermometer calibration. I began oscillator testing only after confirming proper operation of the chamber and thermometer.

The first oscillator tested was a 7-MHz VFO from a portable QRP transceiver. This circuit, a design based on earlier lore that promoted Micrometals SF (-6 designation) toroid inductors and NPO capacitors,[1,2] was built in a small aluminum box. Although drift has certainly been observed with this circuit, it has been tolerated. The transceiver has seen considerable use over a decade-long period from locations as thermally diverse as sun-drenched ocean beaches and mountaintops to snow caves in winter.

I placed the entire VFO box in the chamber with the electronic thermometer in close proximity. Cabling carried signal and power connections for the thermometer and circuit under test. After placing the Styrofoam box over the chamber base, I applied power to the VFO and measured its frequency with the counter. The frequency dropped by about 150 Hz over the first 10 minutes of operation; this is the *warm-up drift*. A slight increase in chamber temperature accompanied oscillator warm-up. Fig 2 shows the experiment's frequency and temperature versus time for this period.

Ten minutes after powering up the VFO, I turned on the chamber heat. As Fig 2 indicates, the temperature immediately began to increase. The accompanying frequency change, however, was not immediate. Only after a couple of minutes of increasing temperature did the VFO frequency begin to drop. The time delay results from the thermal isolation provided by the VFO box.

I kept the heat on until the chamber temperature reached 35 °C—for 24 minutes. Then I cycled the heat on and off in short bursts to maintain a chamber temperature around 35 °C. The VFO's frequency decreased more slowly during this interval, but never reached equilibrium. At 45 minutes into the test, I turned on the fan. The thermal insulation properties of the VFO box again complicated the measurements by slowing the onset of cooling inside the box. Such thermal isolation is very useful; it is a major reason for building a VFO in a separate box!

I removed the VFO from the environmental chamber immediately after obtaining the data for Fig 2. Then I removed the transceiver cover, returned the VFO to the chamber, and repeated the measurements. Fig 3 shows the results on the same time scale used in Fig 2. The VFO was still warm from the previous experiment, so the starting temperature was higher. The warm-up was similar to the earlier one, but when I applied oven heat at 7 minutes into the experiment, the VFO started to change frequency almost immediately. The frequency changed in direct response to changing temperature. Moreover, the frequency was relatively stable as soon as I turned the heat off at 20 minutes. Also, the chamber took less time to reach 35 °C than in the earlier run. Clearly, VFO enclosures should be removed or opened for temperature compensation measurements.

A small circuit board carrying an experimental oscillator should be much more responsive to thermal changes than one in an enclosure. I confirmed this with the design

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[1] Notes appear on page 41.

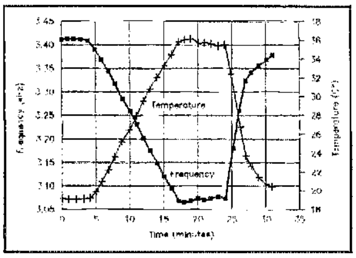


Fig 5—One temperature-drift run made with the oscillator of Fig 4. The circuit was not enclosed in a shield box.

presented in Fig 4. A PNP transistor serves as a Colpitts oscillator. The PNP circuit's collector operates at dc ground when powered from a negative-ground supply—convenient when biasing a tuning diode. I used NP0 ceramic capacitors for the circuit's three frequency determining elements. The inductor value chosen produced oscillation very close to 7 MHz, the same value as was used earlier. The inductor was wound on an SF toroid, a T-50-6.

Fig 5 shows frequency and temperature versus time for this circuit. I turned the oven on at 4 minutes and off at 17 minutes into the test. The oscillator seemed to follow the chamber temperature with virtually no time lag. Additional experiments to higher temperatures showed that the frequency continued to drop as temperature increased to values as high as 50 °C. The run shown in Fig 5 went from room temperature to 35 °C, producing a frequency change of about 3.5 kHz.

Fig 6 shows the results of a first attempt at temperature compensation. I replaced a 270-pF NP0 capacitor with a 100-pF NP0 capacitor paralleled by a 150-pF polystyrene capacitor. This resulted in a drift of 1.3 kHz for the same temperature swing that initially produced 3.5 kHz.

These measurements suggest a simplified way of using the oven: Rather than recording data at one-minute intervals, record the initial frequency and temperature in your notebook. Then apply heat to the oven for a long enough time to produce a temperature rise of 10 to 15 °C. Record the final conditions. This is enough information to calculate a temperature coefficient of frequency (TCF) for the circuit. I used this abbreviated procedure for almost all of the measurements reported in this article.

Temperature Compensation by Numbers

Frequency drift with temperature is usually specified in terms of a fractional change per degree of temperature change. The run in Fig 6 produced a change of about 80 Hz for each degree-Celsius of temperature swing with the oscillator operating at 7 MHz. This corresponds to a change of 0.0011 percent per degree. A more convenient specification replaces percentages with parts per million (ppm). The TCF for the VFO of Fig 6 is −11.4 ppm/°C. The negative sign indicates that frequency decreases with increasing temperature.

Components are specified in a similar way. The *temperature coefficient of inductance* (TCL) for a coil wound on a carbonyl SF toroid (Micrometals −6 designator) is +35 ppm/°C. That is, the inductance will increase by 35 millionths of the total inductance for each degree-Celsius increase in temperature. Note, however, that this value predominantly describes how the *core* permeability changes. The inductor TC is often higher, usually because its wire is not wound tightly against the core. (Heavier wire results in higher Qs but is harder to wind tightly than thinner wire; a more stable coil may therefore result from the use of smaller wire than might be optimum for Q.)

The TCF is related to the coefficients of the individual components by the equation given in the sidebar, "Calculations in Temperature Compensation." There is a factor of one half in the sidebar's Eq A; The TCF is half of the temperature coefficient of the components. The sidebar example analyzes a parallel tuned circuit with one inductor and two capacitors with combined value of C_{total}. The overall temperature-compensating effect of a capacitor diminishes by the ratio of its value to the circuit's total capacitance.

Capacitor manufacturers specify their parts' thermal characteristics, but this data may be omitted in distributors' catalogs to save space. The most stable capacitors have a temperature coefficient of *NP0*. NP0 is a shorthand designation for *negative-positive-zero*. This characteristic depends mainly on the material from which the capacitor is made. Because manufacturing processes operate within tolerances that allow some degree of error and variation, temperature coefficients are sometimes specified in greater detail. These specifications include terms like C0G, C0H and C0J, all of which are variations on the NP0 specification. A C0G capacitor has a TC of 0 ppm/°C with an uncertainty of ±30 ppm/°C.

Temperature compensation involves counteracting an oscillator's inherent thermal instability by adding instability equal in value but opposite in sign. Drift attributable to an oscillator's tuned-circuit inductor usually plays the largest role in thermal instability, although this may not be true if some especially "bad" or poorly specified capacitors are present. Inductor TC is usually positive and can be compensated by a capacitor with a negative TC. A common temperature-compensating capacitor type (a ceramic "mix") has a TC of −750 ppm/°C; this is designated as a *N750* part. The compensating capacitor's coefficient is usually much larger than that of the inductor to be compensated because the compensating component replaces only a tiny fraction of the total resonator capacitance.

As an example, consider the oscillator evaluated in Figs 4 and 5. Based on data recorded at 5 and 17 minutes, the oscillator TCF (as determined using the sidebar's Eq A) was −28.0 ppm/°C. The circuit used all NP0 capacitors. Assuming that these parts are drift-free, the inductor has a temperature coefficient of +56 ppm/°C.

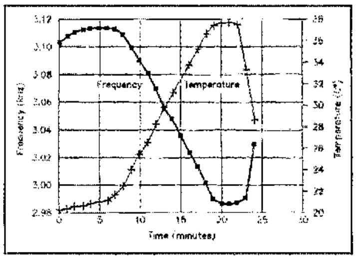


Fig 6—Thermal drift characteristic of the Fig 4 oscillator with some of its NP0 tuned-circuit capacitance replaced by polystyrene-film units. The polystyrenes' negative temperature coefficient offsets drift in the oscillator inductor. (In this case, the result was modest *overcompensation*; see text.) As in the Fig 5 test, the oscillator was not enclosed in a shield box.

The modified oscillator that produced the data of Fig 6 used parallel capacitors, 440 pF of NP0 plus 150 pF of polystyrene (TC, −150 ppm/°C). According to the sidebar's Eq A, the approximately compensated oscillator should have a TCF of −9 ppm/°C. The measurements on which Fig 6 is based reflect a TCF of −11.5 ppm/°C. Compensation would be exact with 220 pF of N150 and 370 pF of NP0 capacitance.

Observations on the Method and Its Limits

The methods presented are incomplete because the test chamber is capable only of positive temperature excursions. Extending the process to cooler extremes should not be difficult. For example, a small thermal chamber could be put inside a household freezer. After a suitable cooling period, the heat source could be turned on to bring the oscillator under test up to room temperature.[3]

The compensation process should not be asked to deliver miracles. The folly of such an effort is evident from the sidebar's Eq A, which shows how two or three thermal effects are balanced against each other. As one component becomes less stable, the others must be made similarly unstable, but in the opposite direction, such that the overall result is zero net drift.

As a better design procedure, use the most stable inductor available. Resonate it with the most stable capacitors available to you. Then use compensation only to make this good design even better. This approach illustrates a concept that finds its way through much of science: Two large effects are not expected to balance each other unless there is good physical justification for the balance. This idea, applied to VFO design, was emphasized by Roy Lewallen, W7EL, in his earlier work.

Ham junk boxes typically do not contain capacitors with negative temperature coeffi-

Calculations in Temperature Compensation

If a tuned circuit consists of an inductor with a temperature coefficient of TC_L that is paralleled by two capacitors, C_1 and C_2, with temperature coefficients TC_{c1} and TC_{c2}, respectively, the temperature coefficient of frequency is then TCF, where

$$TCF = -\frac{1}{2}\left(TC_L + TC_{c1}\frac{C_1}{C_{total}} + TC_{c2}\frac{C_2}{C_{total}}\right) \qquad \text{Eq A}$$

If a compensating capacitor, C_c, is paralleled with a capacitor C_p and the combination is placed in series with a capacitor C_s, the combination has a net capacitance C with a net temperature coefficient TC_n, where

$$C_{net} = \frac{1}{\left[\frac{1}{C_s} + \frac{1}{(C_p + C_c)}\right]} \qquad \text{Eq B}$$

$$TC_n = \frac{C_{net}\,C_c}{(C_c + C_p)^2}\,TC_c \qquad \text{Eq C}$$

cients. One major exception to this is common: polystyrene film capacitors. Most of the polystyrene capacitors I've used have come from Mouser Electronics. They were specified to have a temperature coefficient of −150 ppm/°C. Such relatively stable, low-TC performance should be used in preference to smaller-valued, but higher-coefficient N750 parts.

Examining compensation more closely further complicates the process. Especially at higher frequencies, the parasitic reactances of a circuit's active devices become significant. The analysis presented in this article assumes that drift is linear with temperature. Linearity implies that a component temperature coefficient is constant, but this only approximates reality. Compensating an oscillator over a wide temperature range can be a very difficult chore.

Conclusions and Quasi-Lore

I found this project very enlightening, and it suggests the following conclusions and observations. Some of these are rather specific, often pertaining to technical lore long believed by amateur constructors, while others are more general.

The first, and certainly the strongest, conclusion that emerges is that temperature compensation is neither difficult nor expensive. The equipment needed is usually available in an experimenter's collection of goodies. While lore provides a starting point for design, it is no substitute for measurement.

I'll now present as lore some of the generalities that have come from the experiments, with the thought that they can and will be confirmed through additional reader measurements:

- Excellent components are available to radio amateurs of today. This allows good compensation results to be realized with *single-point* methods. That is, we can do one TC measurement between two relatively close temperatures and be assured that the results will apply over a wider range.
- There is no fundamental thermal stability difference between a Hartley or a Colpitts oscillator, or any other topology that I have tried. It's important that the oscillator operate with current limiting.[4] Voltage limiting can severely degrade resonator Q, leading to compromised stability.
- Micrometals -6 toroid material, when used with high-quality NP0 capacitors, form the most stable starting configuration I have found. The run of Fig 5 is typical (TCF = −28 ppm/°C). After compensation, several oscillators have yielded TCFs between 1 and 2 ppm/°C. I have not yet investigated the newer -7 material that some builders have found useful.
- Warm-up drift is interesting as a measure of self-heating. However, it is a poor indicator of stability when compared with a temperature run in a simple thermal chamber.
- A search for readily available, inexpensive NP0 capacitors that perform as advertised produced two viable sources. One part tested was a 470-pF monolithic ceramic (Mouser Electronics 21RD647). The other part tested was a 220-pF ceramic (Digi-Key P4460, manufactured by Panasonic). Other parts from the same catalog groups are probably as stable. This experience is much more encouraging than a similar experiment done a few years ago.
- A tuning diode I examined (a Motorola MV209) exhibited, as expected, a high TC (+442 ppm/°C). Work continues on active compensation circuits, however. Voltage-regulator circuitry used with tuning diodes must be designed with care because Zener-diode temperature drift can confuse compensation results.
- Some builders (amateur and otherwise) have been building oscillators from available, but apparently mismatched components. Experiments confirmed the variable performance of these methods. One oscillator, built using a -6 toroid and polystyrene capacitors alone (no NP0 units), was overcompensated (TCF = +24 ppm/°C—better than expected, but not recommended). Another test oscillator, built with a -2 ("carbonyl E") core, exhibited a TCF of −63 ppm/°C when tuned with NP0 capacitors. Replacing the NP0 capacitors with polystyrenes yielded a TCF of +8.2 ppm/°C. This topology is also not recommended, because its temperature depends on two relatively large coefficients working against each other. As mentioned earlier, better designs aim for the best stability with each and every component.
- Coils should be conditioned for best stability. Bending and stretching the coil wire during winding creates mechanical stresses in the windings. Heating the coil, and then allowing it to cool to ambient, relieves those stresses. I readily observed this effect with the temperature chamber: A new coil, just wound and subjected to a thermal cycle, starts at a frequency that is different than the final one. Subsequent cycles, however, show repeatable, reversible behavior. Temperature-chamber tests confirmed a method of annealing an inductor in boiling water as suggested by W7EL.[5]
- This article considers only toroidal inductors. They are presently available, offer higher Q for a given volume than other forms, and are well-characterized. Some workers continue to advocate slug-tuned or air-core coils, but have not offered specifications or measurements.

Science can help you build better oscillators if you give it a chance. You can have "rock stable" oscillators if you're willing to build and apply some simple measurement gear.

Acknowledgments

I'd like to extend my thanks to numerous ham colleagues who have shown an interest in this project, especially W7EL; Terry White, KL7IAK; and K8BHZ. Thanks also to Steve Bingham, who contributed an LM3911 from his junk box.

Notes

[1] R. Lewallen, "An Optimized QRP Transceiver," *QST*, Aug 1980, pp 14-19; also see Feedback, *QST*, Nov 1980, p 53. The Optimized QRP Transceiver article also appears in re-edited form in the 1992-1994 *ARRL Handbooks*.

[2] R. Hayward and W. Hayward, "The 'Ugly Weekender,'" *QST*, Aug 1981, pp 18-21. (The Ugly Weekender also appears in re-edited form in the 1992-1994 *ARRL Handbooks*.) Also see oscillator comments in W. Hayward and D. DeMaw, *Solid-State Design for the Radio Amateur* (Newington: ARRL, 1977 and 1986).

[3] Brian Mattson, K8BHZ, has suggested using the Peltier-effect coolers that are now available as 12-volt-powered mini-refrigerators. Relatively inexpensive and available in a size intended to cool six-pack-sized objects, they might be an ideal basis for a small environmental chamber.

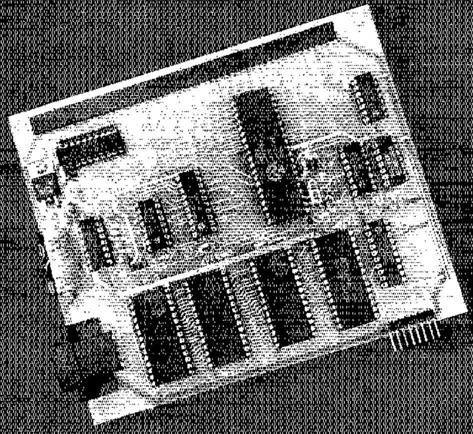
[4] W. Hayward, *Introduction to Radio Frequency Design* (Englewood Cliffs, NJ: Prentice-Hall, 1982), Ch 7.

[5] See Note 1.

Computer-Controlled Electronic Test Equipment

—*Part 1*

Here's a way to *really* put your PC to work! You can turn it into a control center for a wide variety of test equipment. You'll do so at relatively low cost—and completely *externally!*



By Ron Portugal
52 Susan Lane
North Haven, CT 06473

Radio amateurs long ago found ways to use their PCs for more than wordprocessing, checkbook balancing, recipes and games. Ham applications span the range from CW to TV. In this article, I'll show you how to make your PC even *more* useful: how to use it as the control center for an array of test equipment.

If you've investigated adding commercially available plug-in test-equipment cards to your PC, you already know that route is far from inexpensive! Another detraction is each card's use of a PC's expansion slot (there never seems to be enough of them!). On top of that, without using an external card cage (cost: $600 and up), you wind up with *more* bundles of wires and cables dangling from the back of your PC—as if you didn't have enough already! Then there's the software—another $100 (probably more) at a whack. There's got to be a better alternative. You've found it. This series of articles is about computer-controlled test equipment that is:

- External to your PC. You won't have to plug boards into your PC's expansion slots. Input/output (I/O) communication is done through the PC's serial and parallel ports.
- Inexpensive.
- Easy to build. You'll enjoy the one or two evening's work per PC-board-based instrument.
- Programmable in BASICA, GWBASIC, QBASIC, TBASIC or just about *any lan-*

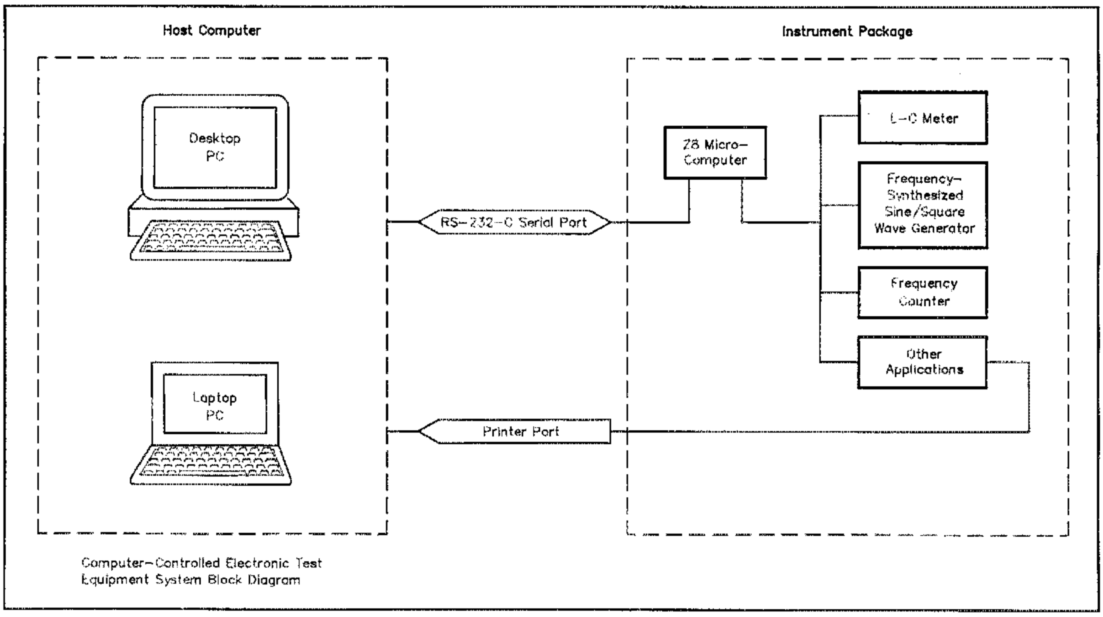


Fig 1—Block diagram of the computer-controlled electronic test equipment setup.

guage that can send and receive data through a PC's serial and parallel ports.

- Flexible. The base unit—the microcomputer board presented in this installment—can be used in virtually any control application: a burglar/fire alarm monitoring system for your home or business; monitoring your weather-sensing gadgets on the roof of your home; managing track switches and keeping tabs of the trains in a model railroad layout, monitoring the water level in your pool (or basement!) or computerizing your sauna or your outdoor grill.

Even if you feel you don't know anything about microcomputers, armed with the information in this and future articles, you can rest assured you'll be able to handle the computerese. To whet your appetite, here's a list of the equipment I'll describe in future installments:

- Completely monitored, triple-section power supply.
- Inductance/capacitance (L-C) meter.
- 1-Hz to 10-MHz frequency-synthesized sine-/square-wave generator.
- 500-MHz frequency counter.
- And...

General Description

Fig 1 shows the system consists of two basic parts: a host computer (your PC) and an instrument package consisting of a Z8671 microcomputer board and a group of add-on instrument boards. The ZILOG Z8671 we'll be using has been around a long, long time. It's a favorite among hobbyists and experimenters because it's cheap ($8 to $15), you don't need special equipment to program it and...*it's easy to use.*

Any IBM PC-compatible computer, (desktop, laptop or hand-held) can act as the host as long as it's equipped with an RS-232-C serial port and a Centronics-compatible parallel port. The host and the instrument packages communicate with each other over these two bidirectional channels. Why bother, you ask, with the serial port when the parallel port is so fast? One reason is that the Z8671 *has a built-in serial port;* second, it is *factory programmed* to communicate with the outside world *through the serial port.*

We'll use the serial port for data and instruction transfers and employ the parallel port for high-speed data transfers. Programming is done in BASIC—QBASIC to be exact. Actually, the host programs are written in QBASIC; the Z8671 programs are written in BASIC/Debug, a watered-down version of BASIC. (More about BASIC/Debug later.) As I mentioned before, if you're familiar with BASIC, you'll be able to do all the programming involved in this project series. To make things simpler, programs for each project are readily available on floppy disk, and everything from the PC board to a complete kit is easy to get, too.[1]

What About Cost?

That depends on what you may or may not have on hand. Each instrument project cost ranges from $50 to $150. If you have to buy everything, this project (with the PC board, all components, cable, ac/dc adapter and software) costs about $150.

Z8671 Functional Description

Fig 2 is the functional block diagram of the Z8671 MCU (microcomputer unit) that's at the heart of this project. The same block diagram applies to the entire Z8 series of ZILOG microcomputers. The only difference between the Z8671 and the other chips in the family is that our chip's 2 kbytes of internal ROM have been *factory programmed* by ZILOG with BASIC/Debug.

I suggest you get a copy of the *ZILOG Technical Manual.*[2] It contains everything you need to know about the Z8 family of microcomputers. The Z8671 material it contains is directed at the software aspect of the

[1] Notes appear on page 50.

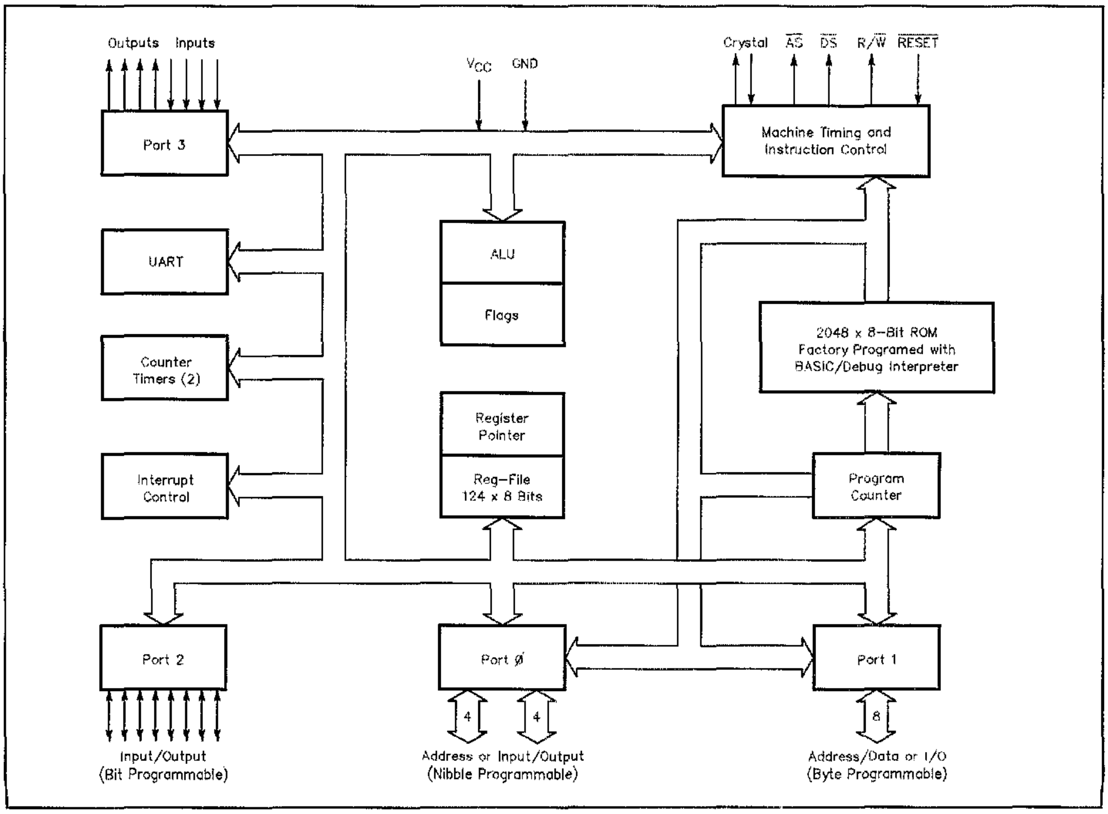


Fig 2—Functional block diagram of the ZILOG Z8671 microcomputer with BASIC/Debug interpreter.

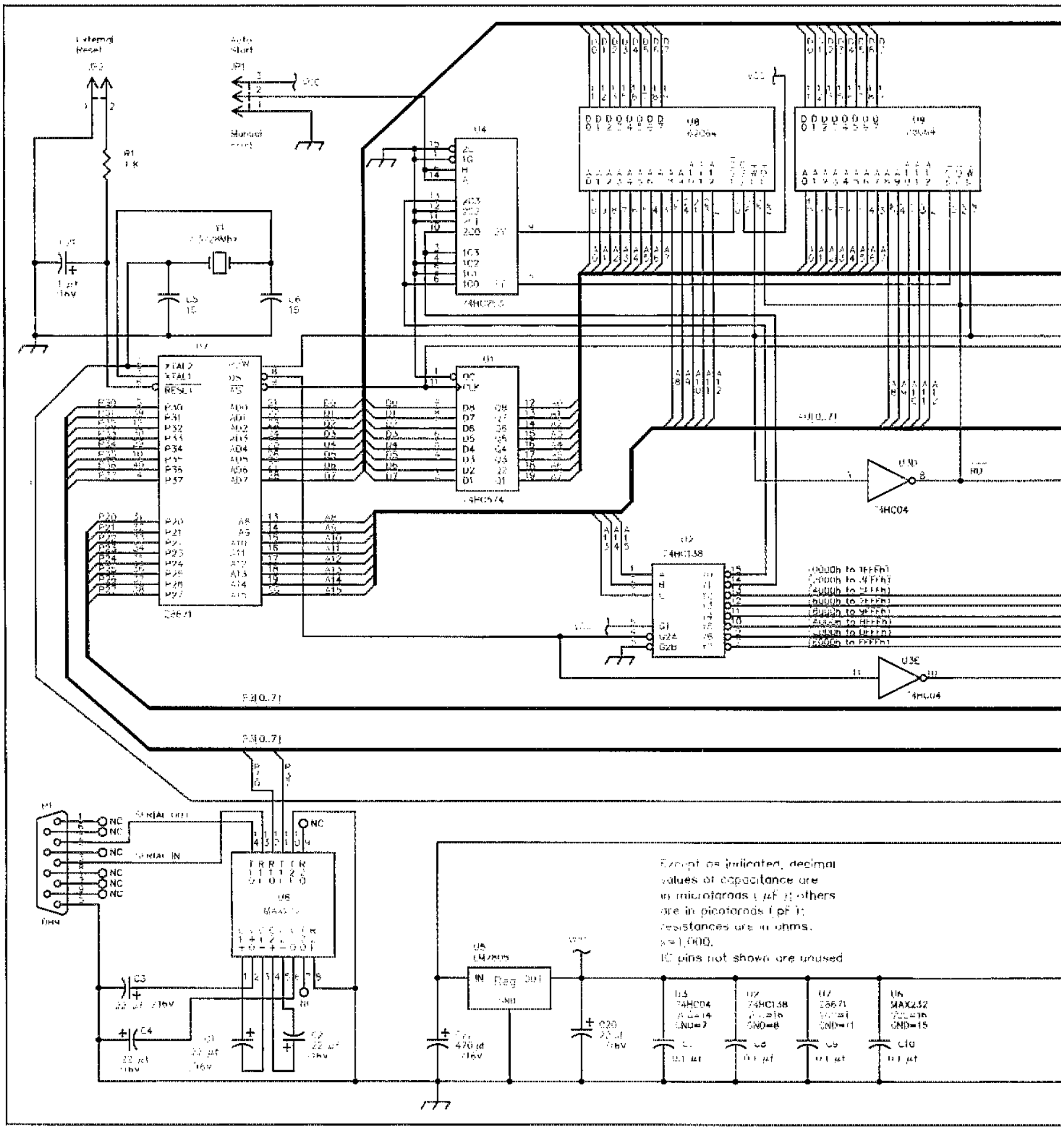


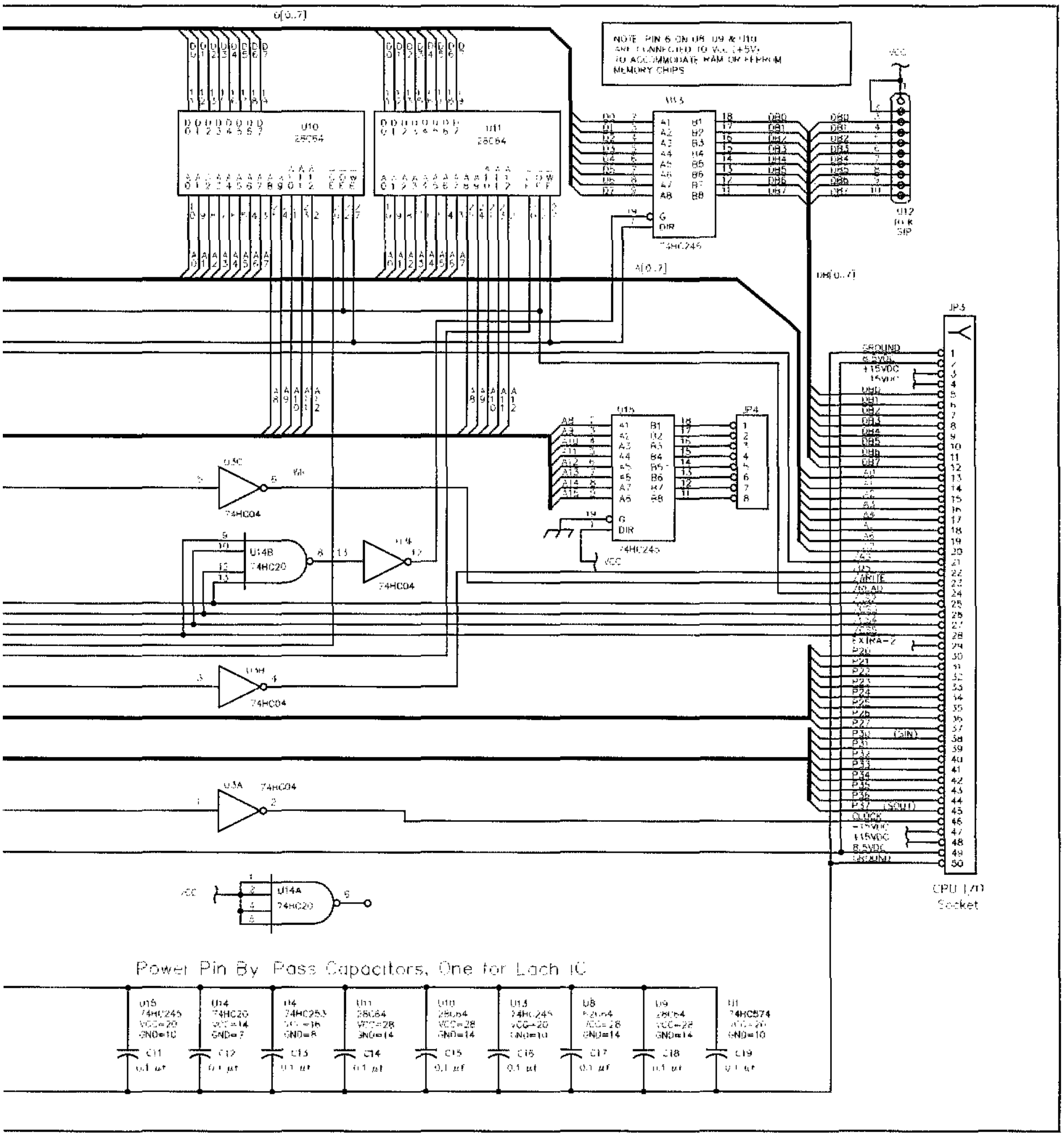
Fig 3—Schematic of the microcomputer circuit. Equivalent parts can be substituted. Unless otherwise specified, resistors are ¼-W, 5%-tolerance carbon-composition or film units.

chip and is very light on hardware-interfacing details.

The Z8671 clock frequency is limited to 8 MHz; in this project we'll be using a 7.3728-MHz controlled clock. Why such a "strange" frequency? Because this crystal frequency makes it easy to obtain exact values of common data rates, ie, 19,200, 9600, 4800, 2400 bits/s.

During power up, BASIC/Debug programs the universal asynchronous receiver/transmitter (UART) for one start bit, one stop bit, eight data bits and no parity. It looks for a data-rate code at address FFFDh[3] and then sets up the count/timer to generate the proper clock frequency for the selected data rate.

What are the data-rate codes? They're the least-significant three bits of the byte at memory location FFFDh that correspond to eight data rates as shown in Table 1. Later, I'll discuss how you get the appropriate bit pattern into address FFFDh. For now, just remember that this UART sends and receives 10-bit words consisting of one start bit, one



stop bit and eight data bits at any one of the data rates shown in Table 1.

Circuit Description

Refer to Fig 3. The $\overline{\text{RESET}}$ line (pin 6) of U7 is connected to a 1-µF capacitor and a 1-kilohm resistor to JP2, the **EXTERNAL RESET** connector. Internally, U7 has a high-value resistor (100 kΩ) connected from the $\overline{\text{RESET}}$ pin to the +5-V supply. A capacitor is connected to the $\overline{\text{RESET}}$ pin so that at power up, the $\overline{\text{RESET}}$ pin is held low for about 100 ms resulting in a power-on $\overline{\text{RESET}}$ signal.

For hot resets (power already applied to the board), just ground the 1-µF capacitor (C21) through R1 by shorting pins 1 and 2 of JP2. JP1, the **AUTOSTART, MANUAL-PROGRAM** connector lets you *electrically switch the positions* of the first two memory chips. When pin 2 of JP1 is connected to pin 3 of JP1, U8 occupies the first 8 kbytes of U7's memory space and U9, an EEPROM, resides in the second 8-kbyte block. By connecting pins 1 and 2 of JP1, the EEPROM occupies the first 8 kbytes and the RAM occupies the second 8. This reversing is done

Table 1

Contents of Memory Location FFFDh

LSB	*Data Rate (bits/s)*
xxxx,x 1 1 1	300
xxxx,x 1 1 0	110
xxxx,x 1 0 1	1200
xxxx,x 1 0 0	2400
xxxx,x 0 1 1	4800
xxxx,x 0 1 0	9600
xxxx,x 0 0 1	19,200
xxxx,x 0 0 0	150

Table 2

Memory Map of the Z8671 Microcomputer of Fig 3

HEX	*BINARY*	*DEC*	
FFFFh	(1111)(1111)(1111)(1111)	65,535	(Memory ends here)
	{111} Block Number		8192 bytes EEPROM
E000h	(1110)(0000)(0000)(0000)	57.344	Fig 3, U11, 28C64.
DFFFh	(1101)(1111)(1111)(1111)	57,343	8192 bytes EEPROM
	{110} Block Number		Fig 3, U10, 28C64.
C000h	(1100)(0000)(0000)(0000)	49,152	
BFFFh	(1011)(1111)(1111)(1111)	49,151	8192 unassigned
	{101} Block Number		input/output ports
A000h	(1010)(0000)(0000)(0000)	40,960	
9FFFh	(1001)(1111)(1111)(1111)	40,959	8192 unassigned
	{100} Block Number		input/output ports
8000h	(1000)(0000)(0000)(0000)	32,768	
7FFFh	(0111)(1111)(1111)(1111)	32,767	8,192 unassigned
	{011} Block Number		input/output ports
6000h	(0110)(0000)(0000)(0000)	24,576	
5FFFh	(0101)(1111)(1111)(1111)	24,575	8192 unassigned
	{010} Block Number		input/output ports
4000h	(0100)(0000)(0000)(0000)	16,384	
3FFFh	(0011)(1111)(1111)(1111)	16,383	8192 bytes EEPROM
	{001} Block Number		Fig 3, U9, 28C64.
2000h	(0010)(0000)(0000)(0000)	8192	
1FFFh	(0001)(1111)(1111)(1111)	8191	6144 bytes RAM
	{000} Block Number		Fig 3, U8, 62C64.
0800h	(0000)(1000)(0000)(0000)	2048	
07FFh	(0000)(0111)(1111)(1111)	2047	2048 bytes ROM used
	{000} Block Number		for BASIC/Debug. ROM
			inside the Z8671 chip.
0000h	(0000)(0000)(0000)(0000)		(Memory Starts Here)

by U4, a 74HC253 multiplexer chip.

P1 is the RS-232-C serial-port connector. Only three wires are needed: serial-in (read), serial-out (write) and ground. The *IN* and *OUT* pins go directly to U6, a MAX232 level shifter. This IC changes the ±12-V RS-232-C levels to 0 and +5-V logic levels, and the 0 and +5-V logic levels to ±12-V RS-232-C levels. The negative voltage required by the MAX232 is generated by the IC itself from the +5-V line. The two lines from the MAX232 find their way back to port pins P30 and P37 of the Z8671.

At the bottom of the schematic is a string of 0.1-µF capacitors. These are the bypass capacitors connected to the power pin of each IC. U5, the LM7805 voltage regulator, gets its power from pin 2 and 49 of JP3, a 50-pin I/O connector. The recommended input voltage for the LM7805 is 8.5 V; however, any value from 8 to 13 V will do.

U8, U9, U10 and U11 are memory ICs. U8 is static random access memory (SRAM); the rest of the memory chips are electrically erasable, programmable read-only memory (EEPROM). EEPROM is essentially the same as SRAM. You can read data from, and write to, the device. However, when you turn off the power, the EEPROM retains the data last written to it, but the SRAM *loses all of its data*. The only catch is, writing data to an EEPROM is relatively time consuming: It takes 10 ms per byte to write to EEPROM as compared to 150 ns per byte for SRAM.

With a little bit of programming, we can make the EEPROM-writing operation almost transparent to us. I say *almost* because when you're writing 8 kbytes of data to EEPROM, it's going to take about 1 minute and 22 seconds! This presents no real inconvenience as most of the programs that we'll be dealing with are only a few hundred bytes long.

How are the memory chips connected to U7? First, you can see that the data lines of each memory chip (D0-D7) feed into bus D and that they are driven by the AD0-AD7 outputs of U7. Port 1 of U7 is a multiplexed port, so address data A7-A0 is delivered to the port during the address strobe ($\overline{AS}$), and data is delivered to the port when the $\overline{DS}$ (data strobe) occurs. Note also that the AD0-AD7 lines feed into U1, 74HC574, an 8-bit latch that's clocked by the $\overline{AS}$ signal coming from U7. And, the outputs of U1 feed into the A0-A7 address lines of each of the memory chips. A9-A12 just go directly to U7's corresponding address lines as shown.

At U8, $\overline{WE}$ (write enable) is the write signal and $\overline{OE}$ (output enable) is the read signal. Since we need only one chip select ($\overline{CS}$) for this chip, and the CE driver provides a high-going-low signal (asserted low), the CS2 signal, the one that's asserted high, is connected to +5 V, so it's always on. (Chip enable [CE] and chip select [CS] perform similar functions.) The real control line here is $\overline{CS1}$, and it comes from our fancy reversing switch, U4. The inputs to U4 come from U2, a 74HC138, a 3- to 8-line decoder.

U15 is a 74HC245 used as an 8-bit buffer. Its buffering address lines A15-A8, and then feeds the buffered address lines to JP4. Why aren't these lines on main connector JP3? Quite simply, there's not enough room. We really won't have too much use for them, so they've been relegated to a separate connector. U13 is another 74HC245. The 74HC245 is really a three-state buffered receiver/transmitter. It can feed data in two directions, from the A pins to the B pins, or from the B pins to the A pins under the control of the DIR (direction) input, pin 1. A second control input to the chip, pin 19, selects the chip or puts it into the three-state or high-impedance mode, effectively disconnecting it from the circuit.

Notice that the data lines that feed the memory chips also feed U13. Furthermore, the direction-control pin is driven by the R/W pin of U7. And, the gate pins come from U3F, an inverter, that is driven by a four-input **NAND** gate, U14B, that in turn is connected to Y2 through Y5 of U2, our 3- to 8-line decoder. Direction of the data into and out of U13 is determined by the R/W control line of U7. The bidirectional data-bus lines of U13 (DB0-DB7) feed JP3.

U12 is a pack of 10-kΩ resistors used as data-bus pull-ups. The **NAND** gate inverter combination determines *when* the data lines from/to U7 are connected to the outside world. When U7 is addressing memory, we don't want the external data bus to be active. That's what the **NAND** inverter combination takes care of. The only time the external data bus is active is when U7 is not addressing any of the four memory chips that are, by the way, selected by the decoded outputs of U2.

Let's look a little closer at this decoder chip. Notice that Y0 and Y1 of U2 feed either U8 or U9 through U4, that in turn is controlled by the setting of JP1. Also, U10 and U11 are controlled by U2 outputs Y6 and Y7. U2 outputs Y2, Y3, Y4 and Y5 are not connected to any memory chips and are brought directly to JP3 pins 25, 26, 27 and 28 as well as to U14B.

Y0 through Y7 are asserted low and only one of these lines is low (logic 0) at any point in time. When neither Y2, Y3, Y4 or Y5 are active, the output of gate U14B is low and through inverter U3F, the three-state control pin of U13 (pin 19) is high, disabling U13. Once again: If Y2 through Y5 are high, the

Table 3

Terminal Program Number 1 for Communications with the MCU

```
COLOR 7, 1: CLS                         'clears screen, blue background
                                        'white letters
Quit$ = CHR$(0) + CHR$(68)              'F10 key quit
LOCATE 24, 1, 1                         'sets up
PRINT STRING$(80, "-")                  'prompt
LOCATE 25, 1, 1                         'message for quit key
PRINT TAB(30)                           "Press <F10> key to quit";
VIEW PRINT 1 TO 23                      'Print between lines 1 & 2
'************************************************************
'open communications port "COM2", 9600 baud, no parity, 8 data
'bits, 1 start & stop bits, 8,192 byte rec./tran. buffers
Com$ = "COM2:"                          'user selectable port COM1:, COM2:, COM3:
baud$ = "9600"                          'Data rates:
  '110, 150, 300, 1200, 2400, 4800, 9600, 19200
'************************************************************
comlink$ = ",N,8,1,DS,RS,CS,CD,RB8192,TB8192"
OPEN Com$ + baud$ + comlink$ FOR RANDOM AS #1
DO                                      'main communications loop
keyin$ = INKEY$                         'check the keyboard
IF keyin$ = Quit$ THEN                  'exit the loop if the user
  EXIT DO                               'pressed the F10 key
  ELSEIF keyin$ <> " " THEN             'otherwise, if user has pressed
  PRINT #1, keyin$;                     'a key, send character to the MCU
END IF
'Test "data in buffer". If characters are waiting (EOF(1) is
'true), then filter backspaces & line feeds & prints them to the
'screen
IF NOT EOF(1) THEN
Z8$ = INPUT$(LOC(1), #1)
DO
Backspace = INSTR(Z8$, CHR$(8))     'test for backspace
IF Backspace THEN                   'change backspace
MID$(Z8$, Backspace) = CHR$(29)     'code so it works
END IF                              'on display
LOOP WHILE Backspace
DO
Lnfd = INSTR(Z8$, CHR$(13))         'test for line feed
IF Lnfd THEN                        'remove line feed
Z8$ = LEFT$(Z8$, Lnfd - 1) + MID$(Z8$, Lnfd + 1)
END IF
LOOP WHILE Lnfd
PRINT Z8$;                              'print filtered string to display
END IF
LOOP                                    'go back to beginning of program
CLOSE : CLS
END
```

gate inverter combination U14B and U3F disables the external data bus. On the other hand, if either Y2, Y3, Y4 or Y5 are asserted, the gate inverter combination enables U13.

This decoding business is more difficult to describe than to understand. Basically, all U2, U14B, U3F and U13 are doing is activating the external data bus when U7 is not accessing the memory chips.

U3 is a hex inverter and only one inverter has been accounted for. Starting with U3A, its input comes from the XTAL2 pin of U7 and its output goes to pin 46 of JP3. That's easy, U3A is acting as a clock buffer. U3E and U3B are connected in series and U3B's input comes from U7's $\overline{\text{DS}}$ pin. The output of U3B runs to pin 22 of JP3 providing a buffered version of the $\overline{\text{DS}}$ (data strobe) to the outside world. The last two inverters, U3D and U3C, are also in series. U3D is driven by U7's R/W line and its output drives, the memory chips $\overline{\text{OE}}$ line, the input to U3C and pin 24 of JP3 where it is labeled $\overline{\text{READ}}$. The output of U3C runs over to pin 23 of JP3 where it is labeled $\overline{\text{WRITE}}$.

Earlier I mentioned that the Z8671 is capable of addressing up to 64 kbytes of external memory. Actually, with some machine-language witchcraft, it can access about *double* that. For our purposes, 64 kbytes is more than enough. If you're interested in learning about the data memory portion of U7, refer to the *Z8 Technical Manual* (see Note 2).

The memory chips on the MCU board are all 8 kbytes wide, and from the wiring of the chips shown in Fig 3, it's easy to see that only 13 address lines are needed to specify any one of 8192 bytes of data in each chip. Since we are dealing with 8-kbyte memory chips, it seems sensible to divide U7's 64 kbytes of memory into eight 8-kbyte blocks, and that's what U2 is really all about. U7's upper three address lines, A15, A14 and A13, are fed to U2 where they are decoded into 8 discreet outputs. Remember, three binary bits can represent 8 decimal values, 0d (000b) through 7d (111b).

At the bottom left of U2 are three gating inputs, G1, G2A and G2B, that determine whether the chip's outputs, Y0 through Y7, are active. G1 is asserted high, while G2A and G2B are asserted low, so the outputs of the U2 can be controlled by the states of the gating inputs. Memory and I/O ports can only be accessed when the $\overline{\text{DS}}$ (data strobe) signal is asserted. This is assured by connecting the $\overline{\text{DS}}$ signal directly to one of the inverting inputs of U2 (G2A) while the remaining gating inputs, G1 and G2B, are hard wired to their *on* state. U2 performs a dual function for us; first, it decodes the upper three address lines, dividing the memory space into eight blocks of 8 kbytes each and, second, it gates the decoded outputs, Y0-Y7 with the $\overline{\text{DS}}$ signal. Let's try to picture U7's memory layout by using the memory map shown in Table 2. Starting at the second column from the right at the bottom, under the DEC heading, is a series of decimal numbers ranging from 0 through 65,535 representing the entire program memory space of U7. The column to the left, headed BINARY, is the binary equivalent of the decimal number shown in the column headed DEC, and the left-hand column headed HEX is the hexadecimal equivalent of the original decimal addresses.

Each of eight blocks contains a *block number* that corresponds to the three most-significant (the left-hand three) bits of the binary address. They also correspond to the decoded outputs of U2, the 3- to 8-line decoder. For example, the top 8 kbytes of memory have a block number of 111 binary or 7d, while the third block from the top has a block number of 101 binary or 5d.

The first two blocks, 0000h to 07FFh and 0800h to 1FFFh, are a little different than the rest. The first 2048d memory locations are eaten up by the BASIC/Debug software inside the Z8671 chip. The remainder of the first 8 kbytes of memory, 2048d (800h) to 8192d (1FFFh) resides in U8, the 62C64. H-m-m…it appears we've lost 2048 bytes of memory somewhere! You're right. The first 2048 bytes of memory in the RAM chip are not addressable and are effectively lost. So, we only get 8192d – 2048d = 6144d bytes of RAM from the 62C64.

Of the remaining 8-kbyte blocks, we lose nothing. From the memory map, you can see that this particular hardware setup for the Z8671 can access four blocks of I/O ports, each block containing 8192 possible ports, a total of 32,768 I/O ports—more than enough for our purposes.

Automatic Startup

One Z8671 feature is *automatic startup*. A program stored in ROM (in our case, EEPROM), starting at memory location 1020h (4128d) and beginning with a line number between 1 and 245, *automatically executes on power-up/reset* without operator intervention. Is that why we have that reversing arrangement for U8 and U9? Yep! You develop a program in RAM. When it's operating, you transfer it to EEPROM starting at memory location 2000h + 1020h (that's where U9 lives), then change the jumper position on JP1, push the **RESET** button and your program runs. This feature is important in stand-alone applications such as alarm systems, machine control, etc. In our application—electronic test equipment—it probably won't be of use. However, the feature is accessible with just a jumper change.

By the way, memory block one (001b) can be populated by another 62C64 of RAM; likewise, block 6 (110b) can also be RAM if you so desire. On the other hand, block 7 (111b) *must be* EEPROM. Why? Remember the data-rate code located at FFFDh? If the right code is not located in the three least-significant bits of memory location FFFDh, the board can't communicate with the host computer.

Building the Microprocessor Board

Using perf board or wire-wrapping techniques for this project would be frustrating, so use a PC board—and IC sockets! (I like the screw machine pin sockets because they are low profile and keep the ICs close to the board.) The 5- × 6-inch MCU PC board (see the title-page photo) is double-sided and has plated-through holes. In addition, you'll need mounting hardware, an RS-232-C three-wire cable (send, receive and ground wires), a reset push-button/cable assembly and a power source: an ac to 9-V dc, 500-mA wall transformer will suffice (the board draws about 150 mA when idling).

Don't, don't, don't pop in the ICs as soon as you finish soldering in the other components! Check the part values and polarities (tantalum and electrolytic capacitors, diodes) at least twice. Then, install the 7805 regulator and apply power to connector JP3 pins 1 and 2 or 49 and 50. Use your trusty DVM to make sure that +5 V appears where it should be on each of the ICs, and that the polarity of the voltages at the tantalum and electrolytic caps are correct. With these preliminaries taken care of, you can now install the rest of the ICs.

The Serial-Port Cable

Now that the board is ready, we'll build a serial-port interface cable and develop a terminal emulator program (in QBASIC of course) so your computer can talk to the Z8671 board.

Most, not all, 9-pin output connectors for serial computer ports are males. I chose to use a 9-pin female connector for the MCU board. Pins 2 and 3 of the serial port at the computer are usually the Rx (serial-in, received-data) and Tx (serial-out, transmit-data) pins respectively. At the MCU end, serial out is pin 2 and serial in is pin 3.

Computer	*Z8671 MCU*
Transmit, serial out (pin 3)	Receive, serial in (pin 3)
Ground (pin 5)	Ground (pin 5)

All you need is a male-to-female interface cable with three wires and two DB9 connectors, one male and one female. To keep things neat, use a hood at both ends of the cable. Use a length of two-conductor, flexible shielded cable; it keeps the radiation from the data transfers to and from the host and MCU board down to a minimum.

Table 4
Simple Test Program for the MCU Board

```
10  "Hello world"
20  "I'm your brand new Z8 microcomputer"
30  "and I'm ready to do your bidding."
40  GO @%61, %7: STOP
```

You need a **RESET** pushbutton connected to JP2. Solder an 8- to 12-inch length of two-wire cable (24- to 28-gauge) to the pins of a dual 0.025-inch box connector using shrink tubing to insulate the pins. Connect the other cable end to a normally open push-button switch, again insulating the solder joints with shrink tubing. Next, add a shorting block across pins 1 and 2 of JP1 to select the RAM first, EEPROM second, setup for the memory.

Communicating with the PC

Now, we must have U7 communicate with your PC via the serial port. But how? We need a terminal emulation program. You can use a standard modem communications program such as *PROCOMM* or others, if you like, or you can write your own; it's really easy. As a matter of fact, a simple first cut at a terminal emulator for your Z8671 MCU board written in QBASIC appears in Table 3. The text to the right of the single-quote (') marks are comments and are ignored by the program.

Things to Note

The program segment between the asterisks sets up the communications link between the Z8671 and your PC. Com$ must be equated to the COM channel you're using. I use COM2; you may want to use, or may only have, one COM port on your machine. Therefore, you'd set Com$ up as follows:

`Com$="COM1:"` And don't forget the quotes or the colon!

Next comes baud$. Earlier in the article, I listed the data rates and the corresponding data-rate codes that U7 responds to. In order to set up memory location FFFDh, the data-rate code address, you first have to get U7 on line. This starts to sound like a Catch-22 situation. If you don't know what's stored at FFFDh you don't know the data rate, and so on around the loop. What to do? There are two solutions: Manually program the EEPROM with the desired code: tedious, but doable. Second, cut and try, or rather type and try, since you can do the trial-and-error routine with software.

This is how the trial-and-error routine works. Assume that your MCU board is correct and ready to go. Hook up the power lines, connect the serial port cable, connect your **RESET** pushbutton and turn on the power. Start QBASIC, type in the terminal program and set Com$ to the appropriate port. Then set "baud$ = " to any one of the eight listed data rates. Now run the terminal program.

The display screen should clear, and when you press U7's **RESET** pushbutton, a colon ":" *may* appear in the upper left-hand screen corner. (Note the "may." If you've picked the right data rate, it will; if you haven't, it won't!) If the colon is nowhere to be seen, exit the terminal program by pressing the F10 key. Then, change baud$ to another data rate, run the terminal program, press U7's **RESET** pushbutton and see if you get the colon. Repeat this procedure until the colon appears.

Once you get the colon, you have to program memory location FFFDh. I used 9600 baud for the U7-to-PC link. Assuming you want to do the same, here's how. Enter the following instruction via your keyboard: @%FFFD = 2. Make sure it's correct, then press the **Enter** key. To be sure that the "2" got to where it is supposed to be, enter the following instruction, then press the **Enter** key:

```
PRINT @%FFFD
```

A 2 should appear on the screen indicating that the content of memory location FFFDh is "2", the data-rate code for 9600 bits/s. That's it!

Here's what happened. The @ sign signals U7 to use the characters to its right as an address location. The % sign tells U7 to interpret the characters to its right as a hexadecimal number. The equals sign tells U7 to place the next character, 2, into the previously designated memory location FFFDh. The same reasoning goes into the interpretation of the print instruction. Pretty simple, isn't it? Now, exit the terminal program (press F10) and change baud$ (in the terminal program) to: baud$="9600."

Run the terminal program, reset U7, and key in the contents of Table 4. Now, type **RUN** and press the **Enter** key. (BASIC/Debug does not differentiate between upper- and lowercase characters.) The message program that you just typed should appear on the screen and your computer will beep at you (the beep is from the "GO @%61, %7" instruction, %7 is the code for bell or beep). Also note that BASIC/Debug has a binary output driver starting at address %61 that lets you write a single character to the UART. There is another binary driver, this time an input driver, located at %54. To use the input driver, use an instruction such as:

```
X = USER (%54)
```

What happens is this. The routine at %54 looks for a character in the UART buffer and assigns its value to X. (The real usefulness of the driver becomes apparent when we get into some interface programming in upcom-

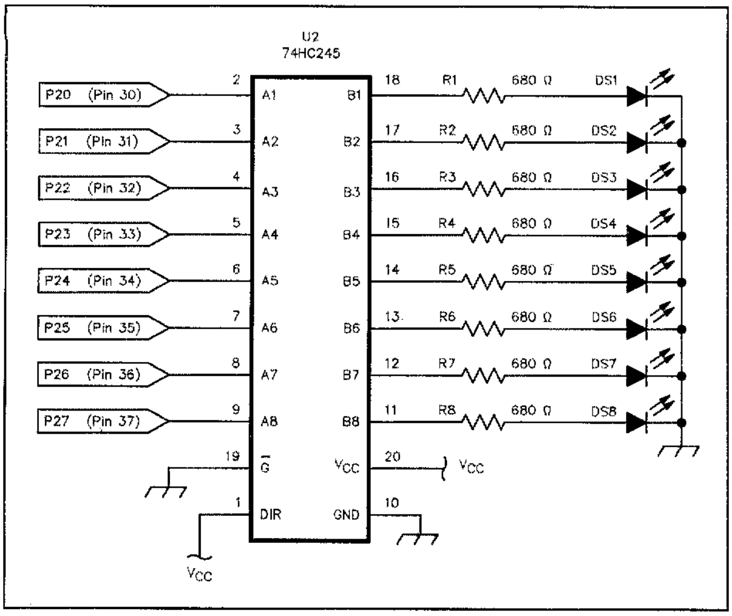


Fig 4—A simple LED-illumination string you can use to familiarize yourself with the operation of the Z8671.

ing articles.) Type **LIST,** press the **Enter** key and a reproduction of what you typed into U7 appears on the screen.

OK, you're up and running. Now relax and get a drink, or whatever you desire...you've earned a well-deserved break!

A Brief Introduction to the Z8671 Instructions, Keywords

BASIC/Debug is a subset of the original Dartmouth BASIC, has a similar syntax and it's easy to use. Many of the common features: FOR .. NEXT, DO WHILE or UNTIL LOOP, trig and transcendental functions, string- and array-handling commands, fractional and floating-point arithmetic have been omitted from BASIC/Debug. However, each of these (and other unsupported BASIC commands) can be duplicated by BASIC/Debug with combinations of the available instructions.

As Table 5 shows, BASIC/Debug supports fifteen commands, a few functions and ...well, you'll see shortly. First, let's see how BASIC/Debug works. It executes commands in two modes, RUN or IMMEDIATE. What's the difference? In the RUN mode, the program has to be stored in memory; in the IMMEDIATE mode, the IMMEDIATE instruction(s) are executed following a pressing of the **Enter** key. The data-rate code "save" and "read back" commands you just used are examples of the IMMEDIATE mode, while the message program used the RUN mode. Some commands are almost always used in the IMMEDIATE mode: RUN, LIST and NEW. GOTO, LET, LIST and others are used in both modes.

Table 5

Keywords for the Z8671 MCU

GO @	Unconditional branch to machine-language program
GOSUB	Subroutine call
GOTO	Unconditionally changes program sequence
IF ... THEN	Conditional operations and branches
IN	Input data to Z8671
INPUT	Input data to Z8671
LET	Assigns expression value to a variable or memory
LIST	Lists the current program to the computer display
NEW	Resets Z8671 to beginning of user memory
PRINT	Prints arguments to computer display
PRINT HEX	Prints the hexadecimal value of an argument
REM	Inserts comments, remarks into code.
RETURN (RET)	Last instruction of a subroutine
RUN	Initiates sequential execution of a Z8671 program
STOP	Ends program execution and clears the GOSUB stack
USR	Machine-language subroutine call

Unfortunately, space does not permit a detailed explanation of each instruction. However, complete explanations of each of the BASIC/Debug KEYWORDS are included in the Z8671 software guide found in the *ZILOG Technical Manual* mentioned earlier.

Implementing the RUN mode is easy, just type **RUN,** then press the **Enter** key. After the program in user memory is complete, or is interrupted by an error, U7 returns to the IMMEDIATE mode. Programs are a series of instructions that perform a specific task. Most computer instructions are entered into memory one line at a time and then executed sequentially by the machine. Programs for the Z8671 consist of a line number followed by a key word and perhaps an address, a data value or nothing. The STOP instruction is not followed by anything; if it were, the computer would ignore it. The message program gives you an idea of how BASIC/Debug instructions appear in a simple program. More than one instruction can be placed on one line, as long as they are separated by a colon (:)—just like in BASIC.

Variables are the fodder of most computers, and in the Z8671, there are 26 two-byte variables available (A through Z). If you need more you can always use part of the RAM that's outside the range of your program.

Operators: +, -, *, / are all self-explanatory. Arithmetic operations are performed from left to right; multiplication and division are performed first, followed by addition and subtraction. There's even a special backslash "\" division operator that performs unsigned division.

The *relational operators* are:

=	equal to
<=	less than or equal to
<	less than
<>	not equal
>	greater than
>=	greater than or equal to

Functions: BASIC/Debug includes two functions: AND and USR. The AND function, like its BASIC counterpart performs a logical AND, like so:

A = AND (expression 1, expression 2)

The USR function calls machine-language subroutines:

USR (%2000, 2980d,C)

The %2000 is the address of the machine-language subroutine, while the 2980d and the "C" are arguments that are needed by the subroutine. Note: Arguments are word size, consisting of two bytes.

Your First Interface

Now that the MCU board is up and running, what can you do with it? As a simple demonstration, let's light some LEDs. Refer to Fig 4. You can build this circuit on perfboard or a solderless breadboard. To keep the wiring to a minimum we'll program Port 2 as an output port. Then we take the eight Port 2 lines, pins 30 through 37 on JP3, to an octal buffer. A 74HC245 does the trick nicely and it can directly drive low-power LEDs.

Dragging the eight Port 2 lines over to the 74HC245 can be a chore. An easy way of getting the job done is to use an 8- or 12-inch length of flat multiconductor cable stripped down to 8 wires. Solder 8-pin, 0.1 on-center, 0.025" square-pin headers to both ends of the cable. Be sure that the pins on the header you use are long enough to provide a good mechanical and electrical contact to the MCU-board connector.

Attach the finished circuit to the MCU board and a power supply. Remember to connect a common ground between the MCU board and the circuit! You're now ready to write your first program of this series.

First, start up the terminal program and reset U7. Now, set up Port 2 as an output port with active pull-ups by entering the following instruction:

Instructions	Comments (Do not enter)
@246 = 0	sets PORT 2 for outputs
@247 = %41	sets PORT 2 for active pull ups

Next let's light up LEDs 0 and 7.

LED #	7	6	5	4	3	2	1	0
on LEDs	1	0	0	0	0	0	0	1 = 81h

And the instruction we need is:

```
@2 = %81
```

The 2 is the address of the destination of the data (81h) which, in this case, is Port 2. With this information as a starting point, you should be able to write and run a few simple programs that make the LEDs turn on and off in binary sequence, first counting up, then down, and then perhaps a marquee-type display where one LED at a time turns on in sequence.

When you start experimenting with LED port programs, you'll quickly realize that you don't have a way of saving a program to a file. That can be annoying. What you need is a terminal program that has a built-in save and load instruction so you can preserve your masterpieces for posterity.

Other discoveries: There's no way to clear the display screen, and using the Z8671 line editor is a wee bit awkward. These and other minor annoyances are eliminated by a more comprehensive terminal program, that I call *Z8_OPS*. The program listing is long and has not been included in this article; however, it is available for those who are interested (see Note 1).

Another answer to the need for a terminal program is to use one of the many commercially available communications programs such as *PROCOMM*. These programs, usually used with modems, take care of some of the keyboard annoyances. Uploading and downloading programs take a little keyboard tap dancing and the saved files usually have to be manipulated a little before they can be reloaded into U7. However, it's still better than retyping the same program a dozen and a half times.

A Preview of Things to Come

Next month, I'll present an autoranging L-C meter with these specifications:

- Capacitance range 1.0 pF to 10,000 µF
- Inductance range 1.0 µH to 1 H

Along with the L-C meter, you'll build a motherboard for the Z8671 MCU and other instrument boards. This motherboard contains a triple-voltage power supply: 8.5 V at 2 A and ±15 V at 450-mA. The 8.5-V supply is unregulated; the other uses a switching regulator to simplify the power transformer requirements. There's lots more to follow, so stay with us!

Notes

[1] PC boards, complete kits, manuals and software are available: (1) *ZILOG Technical Manual* containing Z8671 software description and general information on Z8 MCUs, $5; (2) 1:1 positive PC-board artwork including the component-side silkscreen, component- and solder-side artwork, part-placement and drilling diagrams, $10; (3) double-sided, silkscreened, solder-masked PC board with plated-through holes, $30; (4) *Z8_OPS* software disk (please specify your choice of 1.2 Mb, 5.25- or 1.44 Mb, 3.5-inch diskette), $25.

A complete kit including the *ZILOG Technical Manual*, PC board and part-placement diagram, *Z8_OPS* software (please specify disk size), all on-board components, ICs, IC sockets, connectors and hardware, reset push-button cable, a 4-ft RS-232-C cable, connectors and wire. (The ac/dc adapter is *not* included); price $150. Shipping and handling: On orders under $50, please add $4; on orders over $75, please add $6. Foreign orders add an additional $3. Please make your check or money order payable in US funds to: Ronald J. Portugal. Mail all orders to: Ronald J. Portugal, 52 Susan Lane, North Haven, CT 06473. Allow 4 to 6 weeks after receipt of order for delivery; tel 203-239-0942.

[2] *ZILOG Technical Manual*, part number DC 8291-02, available from ZILOG, Inc, 210 E. Hacienda Ave, Campbell, CA 95008-6600, tel 408-370-8000, fax 408-370-8056/8027. Other publications include *Zilog Microcomputers*, part number DC 8275-04 and *Z8671 Single Chip BASIC Interpreter BASIC/DEBUG Software Reference Manual*, May 1989, part number 03-3149-03 and *Z8 Universal Object File Utilities User's Guide*, Nov 1989, part number 03-8236-04.

[3] Addresses shown with the suffix *d*, are decimal; *h* denotes hexadecimal; *b*, represents binary.

Ron Portugal's had a life-long interest in electronics: He holds a BA in electrical engineering from the Polytechnic Institute of New York, has done graduate work in electrical engineering at the University of Pennsylvania; in physics at Drexel University; and in mathematics at Columbia University. Ron has been a member of the IEEE since 1958. In addition to working as an electrical and electronic engineering consultant, Ron teaches—or has taught—electronics and computer science at Southern Connecticut State University and Greater New Haven State Technical College. Yes, Ron was a ham, but let his license lapse many years ago. (We're twisting his arm about that!)

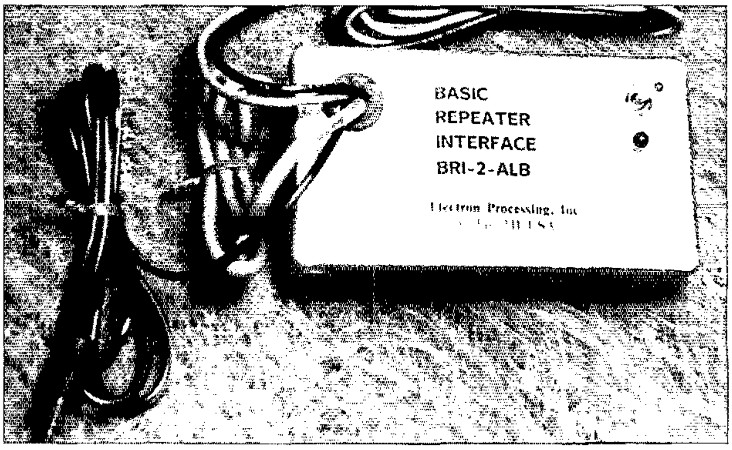
In addition to working for several industrial firms, Ron founded three companies of his own: E & L Instruments, Global Specialities Corporation (yes, that *one!*) *and RSP Electronics Corporation.*

If Ron's name is familiar to you, it may be because you've read work he's published in ComputerCraft. *We welcome Ron to the pages of* QST *and hope our readers will find his project series interesting and instructive.*

New Products

REPEATER DELAY BOARD

◊ To eliminate clunking, snapping and crashing noises on a repeater, the UAD-100 Universal Audio Delay Board is inserted in the machine's receiver audio path before the audio-switching circuitry. It then delays incoming audio before the transmitter with DTMF-tone muting and squelch-tail elimination to provide a pleasant-sounding transmitter drop. Assembled and tested, the price is $99, with manual and one-year warranty. Larry Brooks, WB0ECV, Creative Control Products, 3185 Bunting Ave, Grand Junction, CO 81504; tel 303-434-9405.



An Automated Mobile Radio-Direction-Finding System

Off-the-shelf high-tech and ham know-how are now working together to provide a real-time RDF system that would have been science fiction only a few years ago.

By Robert H. Flanagan, KA1RBH
48 Manchester Terr
Springfield, MA 01108
and
Louis A. Calabrese Jr, N1LZD
62 Glynn Farms Dr
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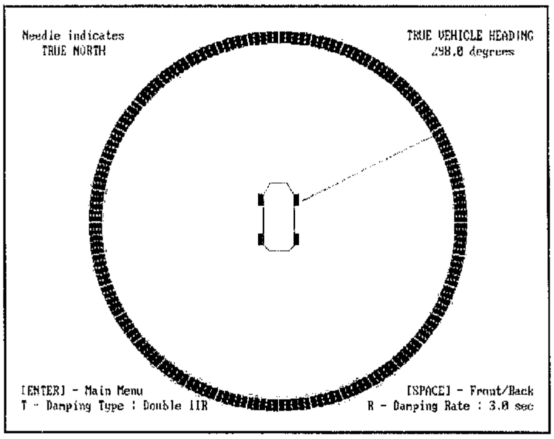


```
Needle indicates                         TRUE VEHICLE HEADING
   TRUE NORTH                                 298.0 degrees

[ENTER] - Main Menu                          [SPACE] - Front/Back
T - Damping Type : Double IIR         R - Damping Rate : 3.0 sec
```

The hardware in this RDF system currently costs well into four figures, so you may not want to build your own version next week or even next year. But now's the time to learn more about some of the RDF technology and techniques we'll be taking for granted tomorrow.

This article describes an automated, computerized, mobile radio-direction-finding (RDF) system we developed while participating in 2-meter mobile foxhunts sponsored by the Mt Tom Amateur Repeater Association (MTARA) in western Massachusetts. The system automates the task of obtaining and plotting RDF bearings—all while the vehicle is in motion!

As the "Plotting Mobile Bearings" sidebar relates, manually plotting RDF bearings isn't difficult, it's just tedious and time-consuming, and time is particularly critical in foxhunting. Every time you have to stop to "shoot" a stable bearing, you lose time. Further, the pressure of competition can lead to human error in bearing calculations. A computer is therefore the perfect tool for streamlining the RDF process. But a new problem arose as we brought computing into the RDF process: that of giving the computer accurate heading and position information fast enough to allow continuous bearing computation and display.

Letting the Computer do the Plotting

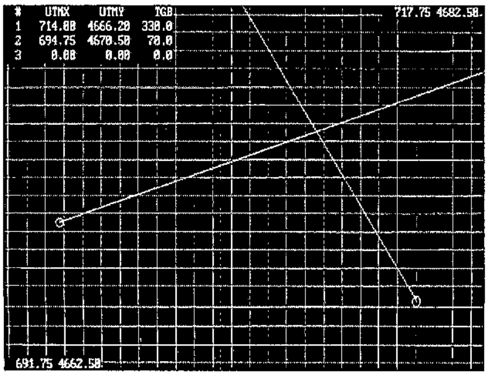
At first we developed a simple BASIC program that would run on a laptop PC and plot bearings on the computer screen. Our intent was to eliminate the hassle of drawing them by hand using a pencil, protractor, and map while driving. The result was a short program that prompts the user for the XY map coordinates of the position from which the bearing is taken, the direction finder or antenna bearing, and the vehicle compass heading. The program uses the USGS maps' Universal Transverse Mercator (UTM) grid, which consists of 1 × 1-km grid squares. We augmented the program by making a special graduated clear plastic template for determining position within the grid squares.

The computer screen (Fig 1) displayed the bearings, their intersection point and the map grid squares. The program worked well, but it required a lot of coordination to juggle all of the input data that it required. Determining our map coordinates, "eyeball" averaging the fluctuating headings provided by our Doppler RDF unit, reading the compass *and* driving were enough to make our heads spin! We felt that the system would be a lot easier to use if we could automate more of its input and calculation functions.

Making the Computer Work Harder

Our current system consists of an IBM-PC-compatible laptop computer with two serial ports and a VGA display, a Doppler RDF unit, a flux-gate compass, and a Global Positioning System (GPS) receiver. All three devices communicate with the computer via RS-232-C serial data; because the computer has only two serial ports, a data switch lets one port do double duty. Fig 2 shows the system block diagram.

A BASIC control program[1] receives data from these three devices and accurately determines the transmitter position. The screen displays the bearings and predicted transmitter location in color. The computer updates and displays the vehicle position *in real time*, showing the path taken from the starting point. The result is a user-friendly system that allows the quick location of



```
#   UTMX     UTMY    TGB                717.75 4682.58.
1   714.00  4666.20  330.0
2   694.75  4670.50   70.0
3     0.00     0.00    0.0

691.75 4662.58
```

Fig 1—How the radio-direction-finding (RDF) program displays two intersecting bearings. This is the basic method of locating a hidden transmitter.

[1] Notes appear on page 55.

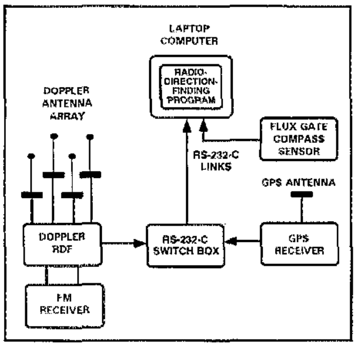


Fig 2—System block diagram. A laptop computer running direction-finding software communicates with three devices—a Doppler direction finder, a flux-gate compass and a Global Positioning System receiver—to gather the information necessary to plot mobile RDF bearings.

foxes, hidden transmitters, and interference.

The Direction Finder

We use a Doppler Systems DDF-4003.[2] Capable of operation from 30 to 500 MHz, it has a three-digit bearing display in addition to a 16-element circular LED readout that shows the relative bearing to the transmitter (see Fig 3). This system must be used with a receiver that operates in the band of interest. It uses four ¼-wave whip antennas magnetically mounted in a square pattern on the car roof. Via its RS-232 interface, the DDF-4003 provides the computer with ASCII information containing the transmitter bearing (in degrees) approximately twice per second at 300 bits/s.

The DDF-4003's bearing accuracy is rated at ±5° in the 144-148 MHz band. It's important to note that this accuracy figure applies to signals received directly, without reflections. Urban and residential areas are full of objects (buildings, cars, hills) that cause reflections in the 2-meter band. This can cause apparent bearings to differ significantly from the true bearing to the transmitter—an effect that's especially noticeable while taking bearings from a mobile. As a vehicle moves, it passes in and out of zones of variable reflection, causing RDF bearings to fluctuate constantly.

With experience, these varying bearings can be averaged by eye, but we decided to filter out reflections via software, averaging the DDF-4003's bearings over time. This consistently yields greater accuracy than is possible by eye. During a one-minute transmission period (common in foxhunting), for instance, the DDF-4003 outputs over 100 individual transmitter bearings at regular intervals. The computer plots these onscreen as vectors (Fig 4A). At the end of the transmission, the computer scales the vectors' lengths according to their frequency of occurrence and redraws them (Fig 4B). The longer the line, the greater the frequency of that bearing value.

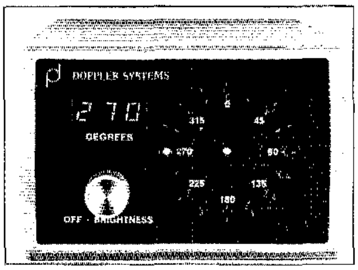


Fig 3—The DDF-4003 radio direction finder. This unit is capable of DFing signals in the 30- to 500-MHz range and features RS-232-C output, 16-element LED and digital bearing displays. *(photo courtesy of Doppler Systems)*

Despite the presence of the computer, human judgment remains an important part of this process. Viewing all of these data on the screen simultaneously, the operator can usually determine which bearings are reflections and which bearings are direct. (Direct-path bearings tend to bunch in a narrow sector; reflections tend to be somewhat random.) The operator chooses a sector, all bearings outside of which are ignored and all bearings within which are averaged. The resulting single bearing value is used for plotting (Fig 4C).

The Flux-Gate Compass

We use a KVH C100 Compass Engine (Fig 5).[3] Consisting of a detachable toroidal flux-gate sensing element and a small electronics board, the C100 is capable of ±0.5° accuracy. Its sensor element is a saturable ring core that free-floats in an inert fluid within a cylindrical Lexan housing. The floating ring keeps the sensing element horizontal with respect to the earth; the housing is surrounded by windings that drive the coil into saturation. Two secondary windings measure the amplitude of induced pulses, which are proportional to the Earth's magnetic field.

The C100's flux-gate sensor takes 10 measurements of the Earth's magnetic field every second. These signals are converted to an equivalent dc level and sent to a microprocessor that uses filtering and averaging algorithms to translate the signals into extremely accurate heading data. Automatic compensation corrects for deviation caused by soft and hard iron objects. Because this compass can compensate for static magnetic fields, such as those created by a car body, it can be mounted inside a vehicle. We mounted our compass on a nonmetallic plastic fixture in the rear of our car.

The RDF program reads the true vehicle heading from the flux-gate compass and adds it to the direction-finder bearing to establish a true bearing to the hidden transmitter.

The Global Positioning System (GPS) Receiver

Based on spread-spectrum data signals from a network of Earth satellites transmitting near 1227 and 1575 MHz, the Global Positioning System allows the determination of terrestrial position to within 100 meters. GPS receivers are available in many different configurations. Some, intended for marine use, are meant for use in sophisticated navigation systems. Others are small hand-held units that can be used for hiking or boating.

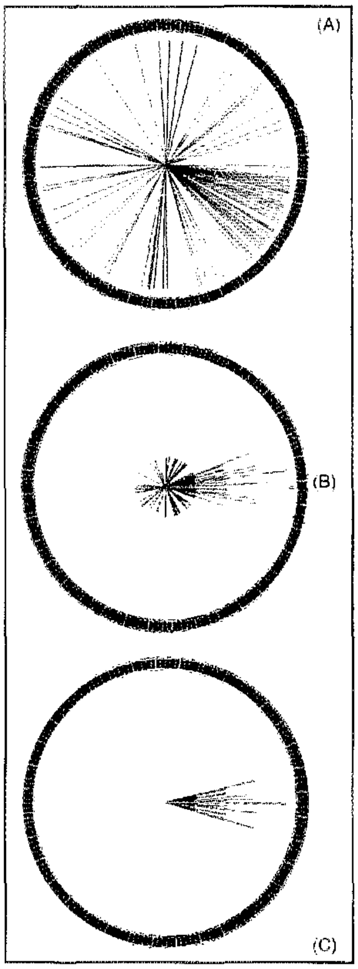


Fig 4—How the *RDF* program time-averages Doppler RDF data. (A) A full minute's worth of bearings produces a major grouping mixed with random reflections from nearby objects. (B) The bearings are replotted with their lengths corresponding to their frequency of occurrence. (C) Bearings more than 15° from the cursor are removed and the remaining data are averaged to yield one overall bearing.

The most significant factor in choosing a GPS receiver for land-vehicle tracking is how many GPS channels it can receive, because more channels allow better satellite reception in rough terrain. Our system also requires a GPS receiver with data output. Most GPS receivers capable of this use the National Marine Electronics Association (NMEA) 0183 protocol, an RS-232-C-based system that sends GPS position information in ASCII sentences.

We use a Magellan NAV 5000D (Fig 6).[4] This is a five-channel, hand-held receiver with NMEA data output. Our software reads the 5000D's raw GPS data and uses it to update our vehicle's position in real time on the computer screen. The GPS vehicle position is used each time a bearing is taken. The result is a complete plot of our path, from our starting point to the hidden transmitter. The resolution is fine enough such that road features such as curves are easily discernible and can be referenced to user maps (Fig 7). This helps us determine which direction to take when closing in on the signal source.

GPS data includes a wealth of information besides latitude and longitude. Among other things, it can also indicate UTC, your

Plotting Mobile Bearings

A *heading* is just that: a reading, in degrees relative to some external reference, of which way you're heading. A *bearing* is your target's direction relative to your position. Plotting a bearing on a hidden transmitter from a moving vehicle requires three important pieces of information: vehicle location, transmitter bearing with respect to the vehicle, and vehicle heading with respect to true north.

The simplest way to find your precision location is find it on a map. We use US Geological Survey (USGS) maps. These are extremely accurate, have excellent detail, and line up perfectly when placed edge to edge. Widely available at camping/hiking stores or directly from the USGS,* these maps cover the entire US.

Next, using your RDF equipment, you must determine the bearing to the hidden transmitter (0 to 359.9°) with respect to the vehicle *centerline*—an imaginary line running through the vehicle's center from front to back. Your RDF equipment should be calibrated to indicate a 0° bearing on signals coming from directly in front of the vehicle. Signals from directly right of the vehicle should indicate 90°, from behind, 180°, and from the left, 270°.

Finally, using a compass, you must determine your vehicle's *true heading*—its heading relative to *true north*. To do this, you must take magnetic declination into account.

Compass needles point to *magnetic north* and yield *magnetic* headings. Because magnetic north shifts over time, maps refer to *geographic north*. Translating a magnetic heading into a true heading therefore requires that you add a correction factor, *magnetic declination*. Magnetic declination can be positive or negative, and its actual value depends on your location on the earth. Once you find out the value of magnetic declination for your area—it's shown on the USGS 7.5′ × 7.5′ maps commonly used by foxhunters and rover contest stations—you can add it to your magnetic compass headings to convert them to true headings. (Some compasses have a declination adjustment that allows them to directly display true headings.)

Let's illustrate how to plot a bearing with the following example. Assume that we've determined that the transmitted signal arrives at 30° with respect to the vehicle centerline, that our compass indicates that the vehicle's magnetic heading is 15°, and that we know the magnetic declination in our area is +15°. First we add the declination (15°) to the vehicle's magnetic heading (15°) to yield the true vehicle heading (30°). To this we add the transmitter bearing (30°) to yield the overall true transmitter bearing of 60° with respect to true north. Fig A illustrates this example.

To draw this bearing, first find the vehicle position on the map. Center the protractor on this position and draw a dot 60° clockwise from true north. Remove the protractor and draw a line, starting from the vehicle position, through the dot drawn at 60°. Ideally, the hidden transmitter will be located somewhere along this line. Subsequent bearings can allow you to determine *where* along this line the transmitter lies: Several bearings taken from different locations will intersect in the general area of the hidden transmitter. *The ARRL Handbook for Radio Amateurs* contains further discussion of direction-finding techniques.
—KA1RBH and N1LZD

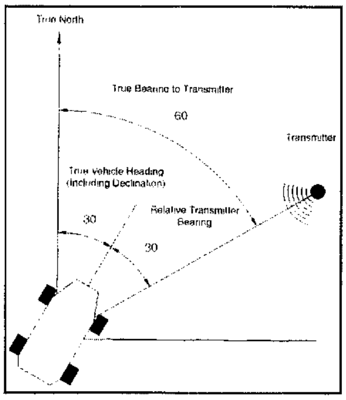


Fig A—How to plot a mobile bearing. The three pieces of information necessary to plot a mobile bearing are the vehicle map position, the angle between the arriving signal and the vehicle centerline, and the angle between the vehicle centerline and true north.

*USGS topographical maps are available at most camping/hiking stores or may be ordered directly from US Geological Survey, Denver, CO 80225 or Reston, VA 22092.

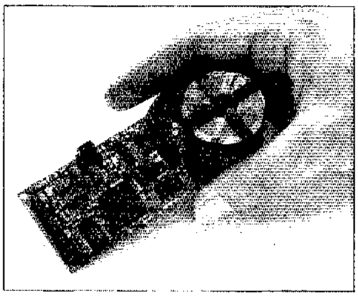


Fig 5—The CV-100 flux-gate compass engine. This electronic compass has a heading accuracy of ±0.5° and features digital, analog and RS-232-C outputs, and automatic calibration. *(Photo courtesy of KVH Industries)*

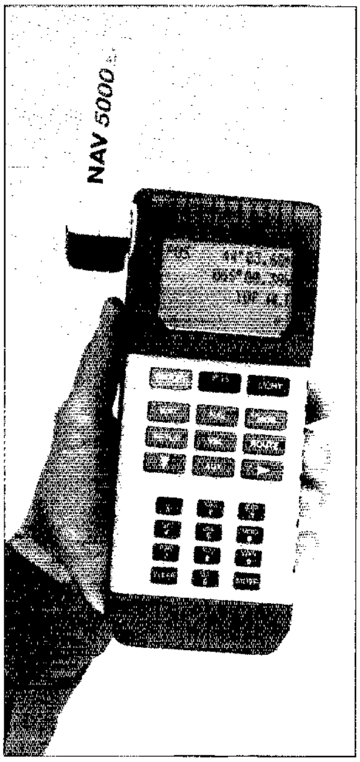
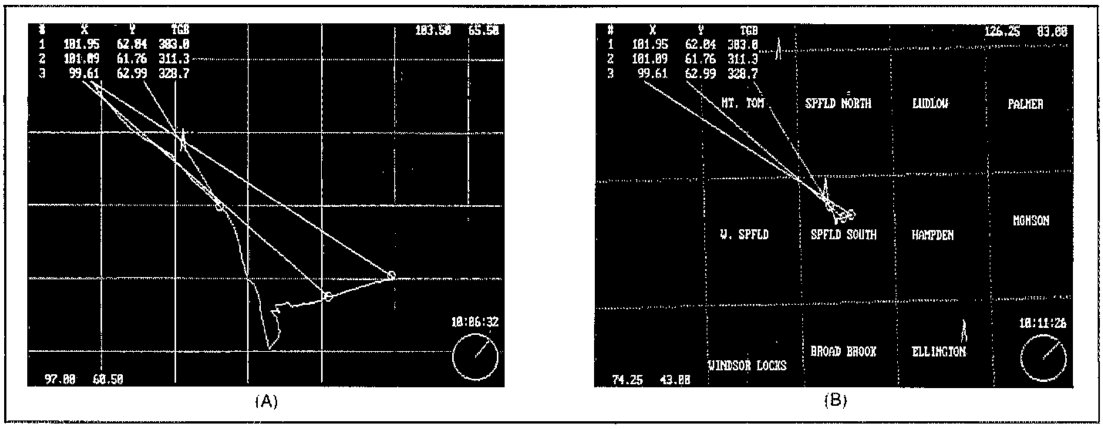


Fig 6—The NAV 5000D Global Positioning System receiver. Using signals from a network of Earth-orbiting satellites, this unit is capable of determining terrestrial position to within 15 to 100 meters RMS. *(Photo courtesy Magellan Systems)*



```
#   X       Y      TGB                  103.50  65.50
1   101.95  62.84  303.0
2   101.09  61.76  311.3
3    99.61  62.99  328.7
                                        10:06:32
97.00  60.50
```

(A)

```
#   X       Y      TGB                  126.25  83.00
1   101.95  62.84  303.0
2   101.09  61.76  311.3
3    99.61  62.99  328.7
                                        10:11:26
74.25  43.00
```

(B)

Fig 7—A typical *RDF* screen plot. (A) RDF bearings are plotted on superimposed grid squares that correspond to topographic maps. Vehicle path and compass heading are displayed in real time. (B) A macroscopic view indicating which topographic maps to reference.

course over ground, and receiver antenna height. A GPS receiver can therefore also serve as an altimeter, a speedometer, a compass, and a clock all in one. (This prompted us to add to our software the ability to display this information in a separate GPS data page.)

If the Global Positioning System can provide compass-heading information, why use a flux-gate compass? One reason is that GPS can provide heading information only while the GPS receiver is in motion. A further GPS limitation is that its information is available to the computer at a maximum rate of one reading every two seconds. Successful mobile foxhunting requires heading reports more often than this, because a car may change heading quickly. The flux-gate compass we use can provide 10 readings per second.

Using the System

Our RDF system, like any system that relies on computer-processed data, operates according to the GIGO (garbage in, garbage out) principle. Provided with perfect position and bearing information, the computer can determine the transmitter location within the limits of its computational accuracy. Just two bearings would be enough to locate a transmitter! But this is never the case. Reflections degrade real-world bearings, a compass's accuracy is limited, and the Global Positioning System is only so accurate. With this in mind, let's work through a typical mobile-RDF foxhunt. We'll assume that the fox may be within a ten-mile radius of our starting location. (We'll also assume that, unknown to us, the fox is located five miles east of our starting point.)

Let's attempt to get a reasonable estimate of overall performance for the complete system. Assuming that our relative bearings to the hidden transmitter are accurate to within ±5° and that our compass is accurate to ±0.5°, we'll be able to determine the true bearing to the transmitter with ±5.5° accuracy. For a target ten miles away, that corresponds to a spread of roughly one mile.

We take our first bearing at our starting location. See Fig 8A. The circle represents the 10-mile-radius search area. Each bearing appears as a straight line, with lines ±5.5° on either side of this representing the bearing's accuracy. Taking just this one bearing narrows our target area down to a sector of approximately 9.60 square miles.

This first bearing tells us in which direction the transmitter lies, but not how far away it is. To get an idea of that, we need a bearing from another angle. So we drive at a 45° angle to our first bearing for two miles. (This is a reasonable strategy. If we were to drive in the direction of the first bearing, our next bearing would be the same as the previous one, and we'd learn nothing about the distance to the transmitter.) Driving at an angle to the initial bearing also allows us to reduce the target area still further, as we'll soon illustrate.

After covering two miles, we take a second bearing—again, at ±5.5° accuracy. We've now determined that the transmitter lies in a quadrangular target area of 2.62 square mile (Fig 8B). Note that the angle and position of our second bearing relative to our first determines the size and shape of the quadrangular target area.

After reviewing the data from our first two bearings, we decide to drive two miles east and take a third bearing. This reduces the target area even further—to 0.636 square mile (Fig 8C). We're applying a definite strategy: that of choosing an optimum position for taking each successive bearing to ongoingly reduce the size of the target area. (In practice, the available road choices don't always make implementing this strategy easy!)

Let's review our progress so far. Initially, our search area consisted of a circle with a radius of 10 miles and an area of 314 square miles. After the first bearing, we reduced the target area to 9.60 square miles—roughly 3% of the original area. Taking a second bearing allowed us to reduce the target area to 2.62 square miles, or roughly 0.8% of the original area. The third bearing reduced the target area to 0.636 square mile, or 0.2% of the original area of 314 square miles. We've reduced our target area by a factor of 500! We continue this process until we locate the transmitter by sight, or until we must continue the hunt on foot using hand-held DF equipment. As we close in on the target, we use input-signal attenuation to provide additional information in determining how close we are to the signal source.

We've found that this system does an excellent job, allowing us to quickly narrow our search down to a small target area after only a few bearings. As the drawings show, bearing accuracy is greatly important in quickly reducing the search area. And a bearing is only as accurate as its three components: vehicle location, relative transmitter bearing and vehicle compass bearing.

Our sample foxhunt looks good on paper. Unfortunately, however, unless the hunt takes place in the middle of, say, the Atlantic Ocean or the Bonneville salt flats, reflections will occur. This is where operator experience plays a key role in separating good bearings from reflections.

If possible, minimize reflections by taking bearings from clear locations high above average terrain. If you use a Doppler RDF system, keep moving to improve the results of the averaging algorithm. If you're in an area where multipath propagation is common, careful use of a directional antenna and signal-strength meter may allow you to out pick the direct signal. If terrain makes getting an unambiguous bearing impossible, move to a different spot. Remember that reflections don't knock you out of the race; they just require you to take more bearings to reliably minimize the search area.

Our Conclusions So Far

Using this system in several foxhunts, we've concluded that computers are great tools for plotting RDF bearings because they're fast, accurate, and allow the operator(s) to concentrate on other activities, such as driving or picking the next location for a bearing. Although they cannot replace skilled RDF operators, computers can amplify human RDF skills by providing us with the right information at the right time.

Future Improvements

The ability to store and rapidly display highly detailed color maps pertinent to the search area would add considerable value to this system. The RDF operator could then determine the best path to the transmitter from the screen, rather than having to refer to cumbersome paper maps.

And what about using packet radio to transfer program data between foxhunt participants? We've been thinking about the possibilities of transferring our GPS vehicle position and bearing data via packet radio to other hams running the same software. Working as a group, we could watch our bearings zeroing in as our screen displays updated automatically. Even the fox could get in on the action and watch our progress!

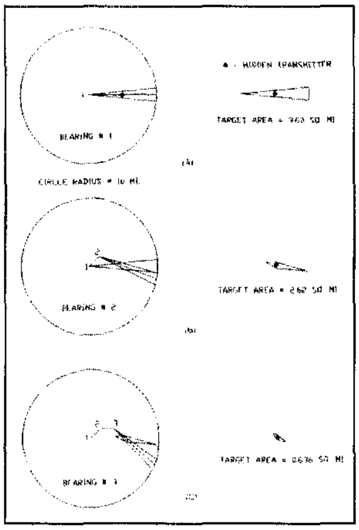


Fig 8—Narrowing down the search area. Starting with a 10-mile-radius circle, successive bearings reduce the search area for the hidden transmitter.

Notes

[1] The program, *RDF*, is available in Microsoft QuickBASIC and compiled executable forms with full instructions, on 3.5" or 5.25" diskettes (please specify) for $20 (includes shipping) from RDF Program, KA1RBH, 48 Manchester Terr, Springfield, MA 01108.

[2] Doppler Systems Inc, PO Box 2780, Carefree, AZ 85377, tel 602-488-9755.

[3] KVH Industries Inc, 110 Enterprise Center, Middletown, RI 02840, tel 401-847-3327.

[4] Magellan Systems Corp, 960 Overland Ct, San Dimas, CA 91773, tel 909-394-5000.

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Louis A. Calabrese Jr is currently self-employed in the retail/automotive field. He was previously employed by Manufacturer's Technologies of West Springfield, MA, as a hardware/software engineer in their device integration group. Mr. Calabrese also spent two years as a Customer Support Representative for Tandy Corporation servicing commercial computer customers.

New Books

LENK'S DIGITAL HANDBOOK: DESIGN AND TROUBLESHOOTING

By John D. Lenk

McGraw-Hill Inc, 13311 Monterey Ave, Blue Ridge Summit, PA 17294; tel 800-233-1128, fax 800-932-0183. First Edition, 1993. Hardcover, 7½ × 9¼ inches, 724 pp, b&w illus. $39.50. (McGraw-Hill's Tab Books imprint publishes the same book in paperback for $22.95.)

Reviewed By Larry Wolfgang, WR1B
ARRL Senior Assistant Technical Editor

I have several John D. Lenk books in my library, and have learned quite a bit of electronics by reading them. His clear, concise style has made him a best-selling author in the field of consumer electronics. Lenk's books generally provide practical information and step-by-step guidance to help an electronics technician or hobbyist. His books offer plenty of "rules of thumb" and simplified explanations to help the reader understand the material.

I knew Lenk was a prolific writer, but I was amazed to discover that *Lenk's Digital Handbook* is his 75th book! The book has 10 chapters, starting with Digital Numbers and Codes and Basic Digital Logic, and then Typical Digital IC and Discrete Circuits and Microprocessors, Computers and Controllers. Next, the book moves into Digital Test Equipment and three troubleshooting chapters: Digital Troubleshooting, Classical Examples of Digital Troubleshooting and Advanced Digital Troubleshooting Techniques. The book ends with two chapters aimed at specific types of circuits: Troubleshooting Microprocessor-Based Devices and Troubleshooting Digital TV/Video Circuits.

The first chapter begins with the statement that "This chapter is for readers unfamiliar with the number systems and alphanumeric codes used in digital electronics." I doubt that many readers totally unfamiliar with the binary, octal or hexadecimal number systems would feel comfortable working with these systems after reading the seven pages given to this topic. It's a good refresher for those who have studied these systems at some time in the past, however, before jumping into the world of digital electronics.

I ran into some problems reading the Basic Digital Logic chapter. Most of it is good information, but a few (apparently) typographical errors tend to be confusing. For example, the paragraph about DeMorgan's Theorem (although not identified by this name in the text) left my head spinning. Lenk states four identities here, with some overscores apparently missing: $A+B = AB$, $AB = A + B$, $A + B = AB$ and $AB = A+B$. The supporting text is also misleading, because it says, "If the term subjected to the NOT operation contains an OR or an AND operation, in addition to variables, the Or and AND operation is also complemented (or inverted)." While this statement *can* lead to a correct representation of the two identities of DeMorgan's Theorem, Lenk's representation is not correct and his last two identities here are simply not true.

I've never seen a book that didn't contain a few typographical errors, at least until several printings have provided the opportunity to catch and correct them. Overscores and equations in general are particularly troublesome. Still, such typos can be unsettling, because the reader who notices them may feel uncomfortable about what other errors might occur in the book. There's also the problem of that some readers won't recognize the typos and assume these statements to be correct. A few errors don't have to ruin an entire book, however, and there's plenty of good information in *Lenk's Digital Handbook*.

The troubleshooting chapters offer plenty of useful techniques and tips. For example, a technique is described to help locate a power-bus short. Supply a relatively high current (3 to 5 A) to the bus at the normal operating voltage. Then use a digital voltmeter with 0.01-µV or better resolution to measure voltage drops along the power bus. By measuring across two points along the same bus, you can trace the short-circuit current because there will be no voltage drop unless there is current through the trace. Another tip that I found interesting was to freeze the entire PC board, allow moisture to condense on the board, and then power the board with a 3- to 5-A supply at the normal operating voltage. As the board warms up and defrosts, the current path becomes visible, often pinpointing the short.

Lots of charts and diagrams illustrate the book. Many are block diagrams, but there are partial circuit diagrams where appropriate. There are also illustrations showing how to connect test equipment and make the measurements described in the text.

I may not be ready to tear into my computer to do component-level troubleshooting and replacement after reading this book, and I'm not likely to tackle problems with my digital TV even with *Lenk's Digital Handbook* at my side! I do have a better understanding of how these circuits work, however.

If you're interested in learning more about the fascinating world of digital electronics, I recommend *Lenk's Digital Handbook*. Radio amateurs will find the material interesting and useful, especially as we depend on digital computers to assist us in our Amateur Radio activities and as more equipment has microprocessor control circuitry. This book will help you become conversant with the terminology of digital electronics.