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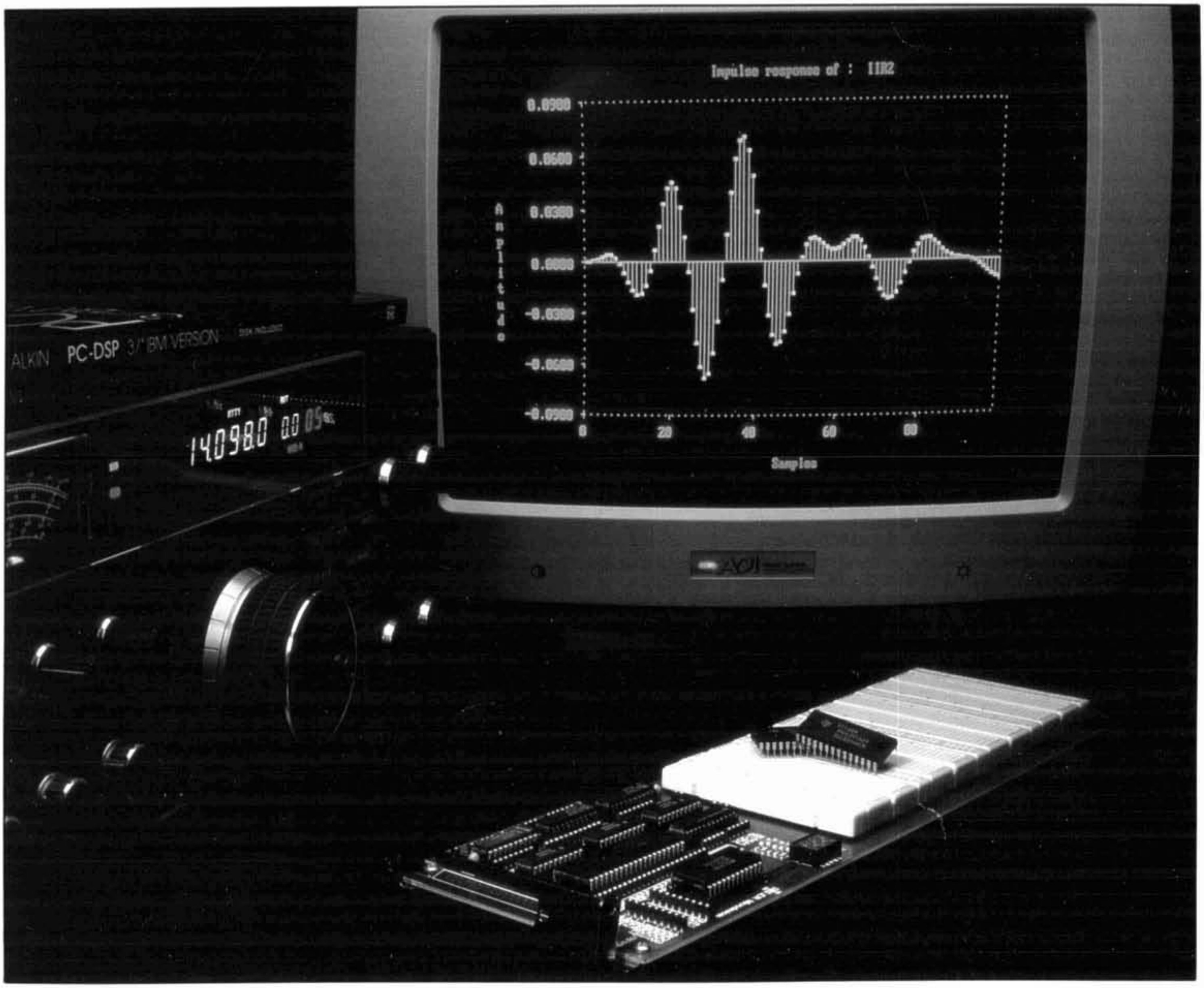

- DSP on the PC
- LORAN-C Frequency Calibrator
- The Triangle Antenna
- Improving Receiver Performance in Modern Transceivers
- Optimizing Amplifier Gain-Bandwidth Product
- How to Design Shunt-Feed Systems for Grounded Vertical Radiators
- Computer-Based Signal Monitoring System
- Antenna-Structure Interaction

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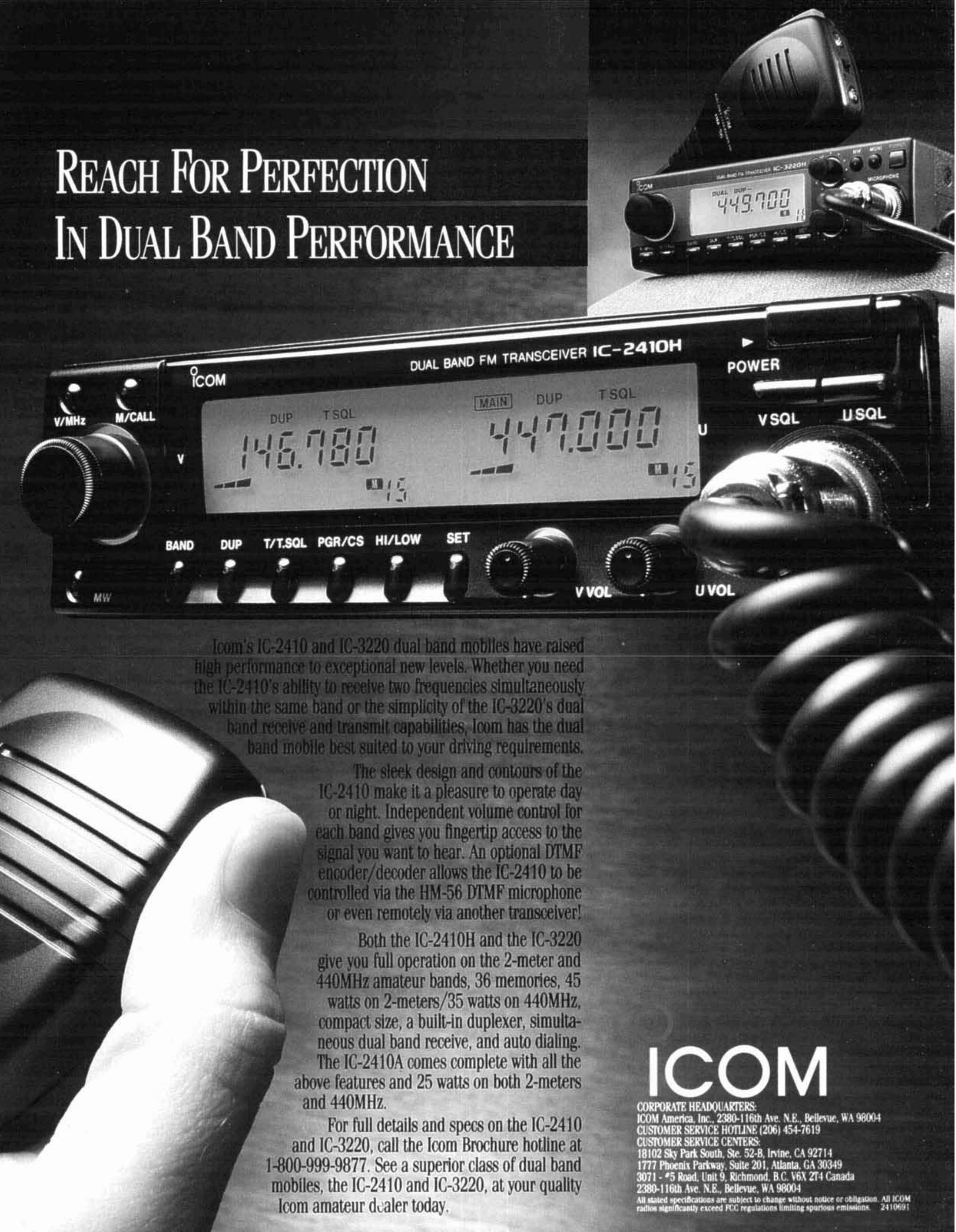
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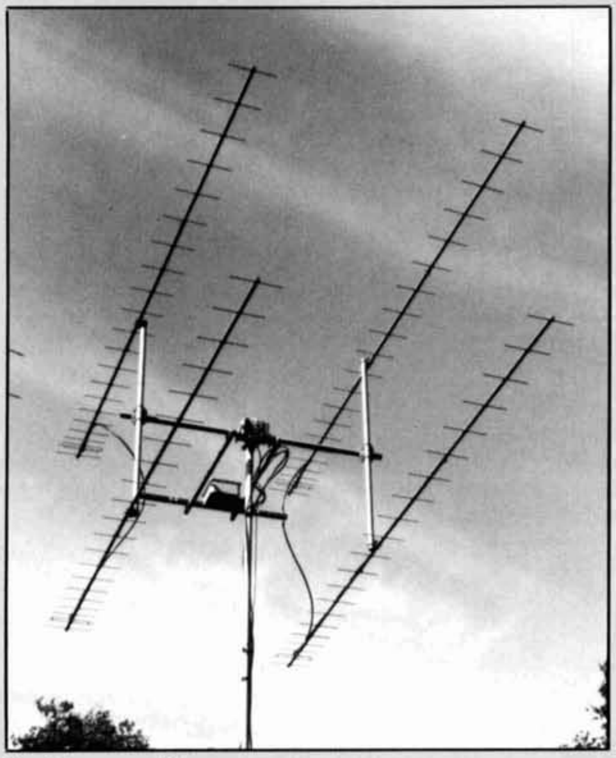

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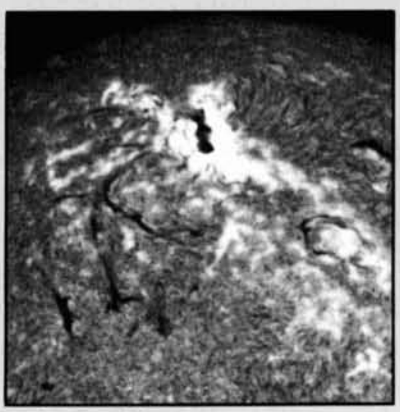
Cover photo: Nick Thomas, KA1YGS. Protoboard courtesy of JDR Microdevices. ICs courtesy of Texas Instruments, designers and manufacturers of the TLC32044. *PC-DSP*, by Oktay Alkin, is published by Prentice Hall.



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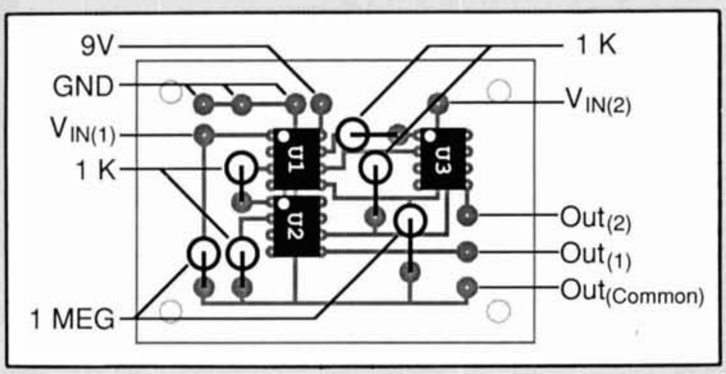

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EDITORIAL

Freedom of Communication: Protecting our right of access to the airwaves.

Imagine you're on the road with your family. You're on vacation, chatting on the radio as you drive along from state to state. You pass through southern Pennsylvania, and cross the border into New Jersey on your way to a week at the Jersey shore. Suddenly, you look in your rearview mirror and see the lights of a police cruiser flashing as it comes up behind you. You check your speedometer, and find you're going the speed limit. "Why's he pulling me over," you wonder. "I haven't done anything wrong." Or have you?

The officer approaches your car and you look up at him in bemusement as he begins asking you questions about your rig. What type of radio are you using, what frequencies is it capable of receiving, on what frequency are you operating, do you have a permit to have this radio in your car? You answer all his questions and tell him that you didn't know a permit was required. You show him your amateur radio operator's license which confirms your right to operate a mobile radio. Or so you believe. The officer asks you to follow him to the local police station. Here you and your family wait while your radio equipment is confiscated. You're told that you are in violation of a New Jersey statute (on the books since the 1930s) which prohibits you from having, in your vehicle, a radio capable of receiving public safety (police, sheriff, fire, EMS, and so on) or government frequencies—an indictable criminal offense. As a matter of fact, you would have been in violation of the law, even if your radio wasn't on when you were stopped. Even if the radio had been disconnected and in your trunk, as long as it was capable of receiving these frequencies, it put you in violation of the law. Mere possession is a crime! The police release you on your own recognizance. You don't know when, or if, you'll get your equipment back, if you'll be required to pay a fine, if there will be other repercussions. Some vacation!

Sounds like the stuff of nightmares, doesn't it? But New Jersey, and a handful of other states (including Virginia, Michigan, Connecticut, and Kansas) have state and local laws which prohibit you from operating mobile radio equipment or carrying mobile scanners that can receive public safety and government frequencies.* In some municipalities, you are allowed to have this type of equipment in your car—provided you obtain a permit from the appropriate government official. However, such permits are hard to come by. Penalties, if you are found in violation of these laws, run the gamut from confiscation of equipment, fines, and, in some extreme cases, imprisonment—all because you have a mobile rig! And even if your radio doesn't receive the frequencies in question, your equipment may still be confiscated because the officer who stops you may not be knowledgeable enough about your particular radio to know if what you say about its operating capabilities is true. Enforcement is arbitrary—at the whim of the public safety official.

The rationale behind these now infamous "scanner laws" (as they have been christened) was probably to prevent criminals—such as burglars—from using police frequencies to determine if officers had been dispatched to the scene of a crime. But if these people are criminals in the first place, will they really worry about carrying "illegal" radio equipment in their vehicles? I think not. In the meantime, amateur radio operators who monitor these public safety bands in order to offer assistance in times of emergency, have their hands tied if they live in an area which prohibits them from listening in.

At the urging of the American Radio Relay League, the FCC has been studying whether to take preemptive action to nullify certain state statutes and local ordinances affecting transceivers used by licensees of the amateur radio service. In response to the ARRL's *Request for a Declaratory Ruling Concerning Possession of Radio Receivers Capable of Reception of Police or Other Public Safety Communications* filed on November 13, 1989, the FCC initiated a *Notice of Inquiry* (PR Docket 91-36). Replies were due by July 8, 1991. Dozens of comments were received from amateur radio operators, General Radio Mobile Service (GRMS) users (explaining why they too should be preempted from the laws), and the Associated Public-Safety Communications Officers, Inc. (APCO). All support the ARRL request to exempt hams from the scanner laws.

Besides the obvious solution of simply granting the exemption, the FCC has been looking to amateur radio equipment manufacturers to eliminate the out-of-band capability in mobile equipment. Other opponents of the laws propose that those who don't want their broadcasts overheard, should simply encrypt their transmissions—as government agencies and commercial operations often do.

There's no way of knowing what the outcome of this debate will be. But we must remain vigilant; this is an issue that affects us all. Hams continue to loose their equipment, pay fines, and face harassment in the few states which choose to enforce these arbitrary laws. Granted there are times when radio security *must* be maintained in the interest of public safety, or to protect the rights of individuals. But why are these agencies trying to take away our freedom of access to the airwaves, our freedom of communication, by insisting that we be the ones to take responsibility for their needs to privacy? The responsibility to protect the integrity of public safety frequency transmissions lies not with the private sector, but with those charged to serve us.

Terry Northup, KA1STC
Editor

*In 1991, a bill was introduced into the New Jersey Senate which would make mobile scanners illegal only if they were used in the commission of a crime; that is, for an improper purpose. The bill has passed the Senate and is now being considered by the Assembly (*W5YI Report*, September 15, 1991).

ON THE ROAD...

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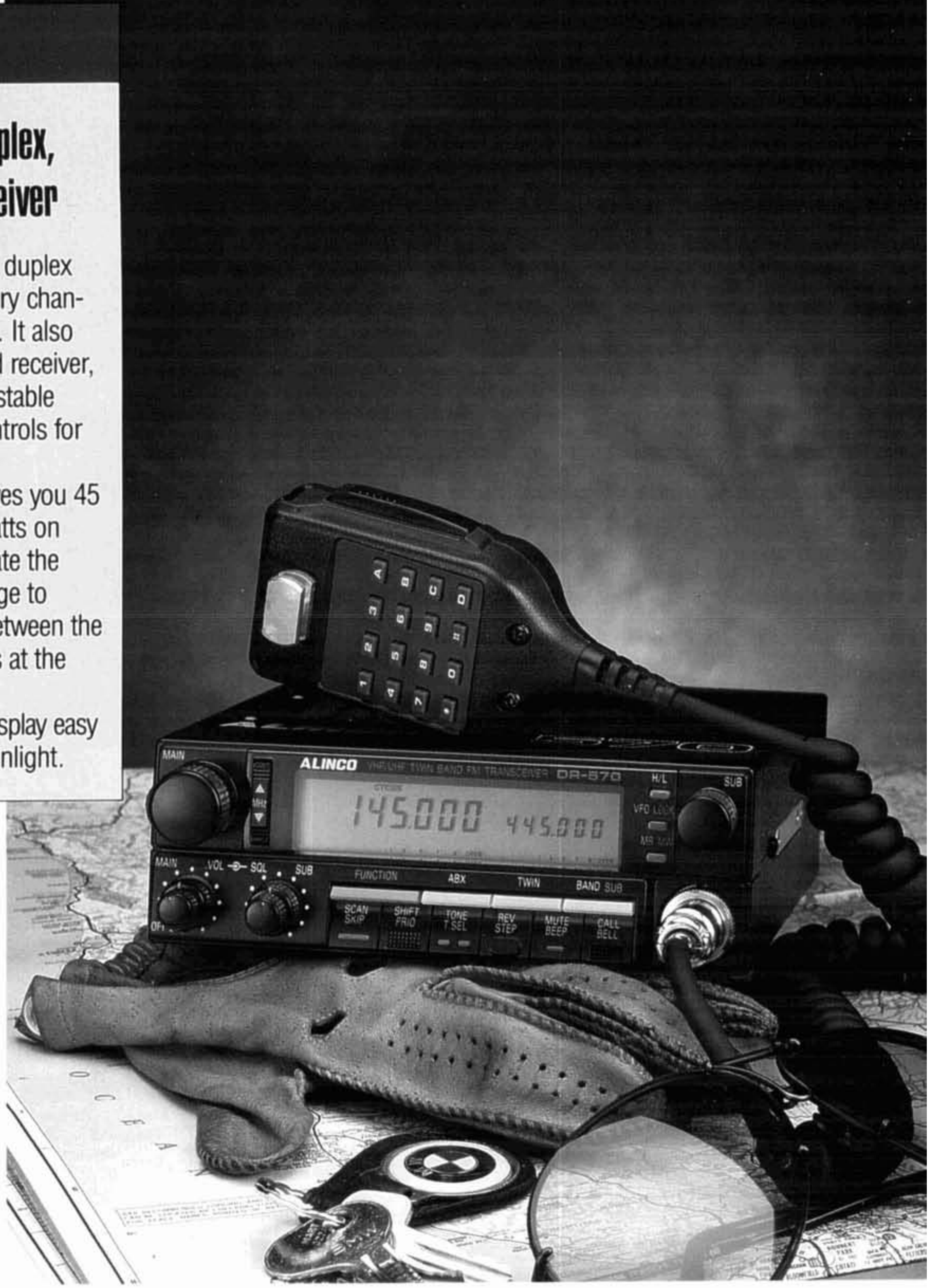

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LETTERS

Comments on "Antenna Angle of Radiation Considerations"

Congrats on another great issue (Summer 1991). I feel compelled to comment on the article by Carl Luetzelschwab, K9LA, as follows:

K9LA's treatment of the tired quad versus Yagi controversy is concise, accurate, and incontrovertible, for which he deserves praise—and here's mine. Unfortunately, though, his article continues to propagate the argument rather than laying it to rest, as someone should.

To do that, the term "radiation angle" must be purged from our vocabulary. The concept implies that propagation occurs at a certain vertical angle, the one at which an *antenna's pattern peaks*, as if the antenna controlled that angle. It doesn't; the geometry of the path selects the angle. Energy radiates at all angles in the antenna's pattern, of course, but the energy at the wrong angle for a path goes somewhere else. *Wave angle* is the name of the vertical angle that supports a particular path, a path which depends on the effective height of the ionospheric "reflection" and on the distance between stations (or remote ground-reflection points). Yes, multipath propagation happens; but when, for example, a two-hop and three-hop between both occur between a pair of stations, each path has its own wave angle. The antenna pattern, affecting how radiated power varies with angle, does affect the relative strength of signals taking each path.

The distinction between maximizing radiation at low *wave angles* and needing a low "radiation angle" is not merely pedantic. The "radiation angle," meaning the angle at which the lobe peaks, only affects the power radiated at any specific wave angle indirectly; sure, a low radiation angle *tends* to radiate more power at low angles, but that's only a tendency, one which misses the point.

To appreciate where such misplaced emphasis leads, let's consider K9LA's article. Generally, DX paths on 20 meters have wave angles between 2 to 10 degrees. Yet K9LA's **Figure 2** (dipole versus loop) stops at about 15 degrees. The two curves are indistinguishable below 20 degrees, yet he concludes the loop is better. Similarly, the Yagi versus quad graph, **Figure 3**, shows the quad to have the lower lobe peak; that's nice, but irrelevant. The quad does exhibit an edge of roughly 1 dB in the range from 25 degrees down to 10 degrees, but no clear advantage at 5 degrees—perhaps because the graph scale can't show it. Thus, even when the author's assertion has merit, the concept of radiation angle prevents him from making his case.

My second point concerns comparisons on the basis of boom length or element count. If a five-element 30-foot boom Yagi is easier to erect than a two-element quad, why compare the quad with a two-element Yagi? Those who insist on that comparison should consider that a few extra feet of mast would boost the Yagi's low-angle signal enough to match that of the quad; surely that small Yagi, even with a taller mast, is easier to erect than a quad, and its lower wind resistance makes the taller mast structurally allowable.

Quads do have advantages; they are cheap, broad-band and, some say, quieter in rain static. But the focus of K9LA's comparisons was limited to radiated signal strength. On that basis, equal element-count quads may perform better at very low angles; but, when antennas of similar bulk, wind loading, or ease of erection are compared, the result may favor Yagis—the article, like most, doesn't tell us. In future technical articles, let's abandon the tired and worthless "radiation angle" and element-count comparisons and compare performance in ways that matter.

David M. Barton, AF6S
San Jose, California

Part 2 of K9LA's article, "Antenna Angle of Radiation Considerations," appears on page 61 of this issue. Ed.

I am convinced!

Yes . . . I am convinced that *Communications Quarterly* is an outstanding contribution to the field of communications.

After reading the first four issues from cover to cover, I have determined that *Communications Quarterly* must be added to my library of outstanding technical publications.

I really had no idea that *Communications Quarterly* was so outstanding in quality and breadth of topics serving the communications enthusiasts.

Kenneth M. Miller, K6IR
Rockville, Maryland

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Cedar Rapids, Iowa 52402

DSP ON THE PC

Explore DSP techniques using live signals from the ham bands and inexpensive PC analysis software

With digital signal processing (DSP) modems and TNCs under development and soon to be available, amateurs must learn the concepts behind the new technology they will be using. But short of enrolling in college, where do you go to learn DSP?* Traditional university textbooks on signal processing are available, but their formal treatments often act as a barrier to all but the math experts among us. Page after page of sigmas and integrals, while essential for mathematical rigor, tend to discourage those just starting out in this most nonintuitive subject. If a supplement to the formal approach existed; if we could be shown, in an approachable way, DSP theory in practical use; then more amateurs would tackle the concepts behind this powerful new technology.

PC-DSP: visual and intuitive

Fortunately, a complement to the traditional methods of learning DSP is available to PC owners. *PC-DSP* by Oktay Alkin, published by Prentice Hall, and priced at $20 is quite a bargain. For this low price, you get DSP analysis software on diskette and a 173-page user's manual.

PC-DSP lets you explore DSP techniques visually and intuitively. Menu driven and interactive, the program's power and utility lie in the superior use of graphics. Usually, a properly scaled graph of the results of a DSP operation is available with a single keystroke. Time is spent learning DSP concepts, not graph plotting—the grunt work of signal analysis.

Signals are created using the program's waveform synthesis capability or they can be imported from ASCII files generated external to the program. Once the signal is input, you can start to analyze or "process" the data by simply selecting among menu items. The author suggests the concept of a DSP calculator. My approach is more casual.

If I don't know what a DSP operation will do for me, if I've never heard, for example, of the Blackman-Tukey method of spectral estimation, I just try it and look at the results. Did the operation do what I suspected it might? For FSK signals, would the operation make it easier to make mark and space decisions? Has the interference been reduced? All this can be estimated by simply inspecting the plot PC-DSP provides. You can use PC-DSP with this nondirected (or possibly, pell-mell) approach toward investigating DSP concepts; the learning takes place as you try to interpret the results you see on the screen. Of course, no study of DSP techniques is complete without a grasp of the underlying theory behind the selected operation. But when used with a good introductory textbook, or classroom instruction, the program will enhance your understanding of the concepts. The program's ease of use and seamless graphics are gripping. The first night I used PC-DSP, I was up until the early hours of the morning experimenting. And I haven't done much in the way of late-night hacking since the early days, when PCs (and I) were much younger.

To run PC-DSP you need a 100-percent PC-compatible computer operating under

*To learn more about DSP, see the articles by Brian Bergeron, NU1N in the Fall 1990, Winter 1991, and Spring 1991 issues of *Communications Quarterly*. Ed.

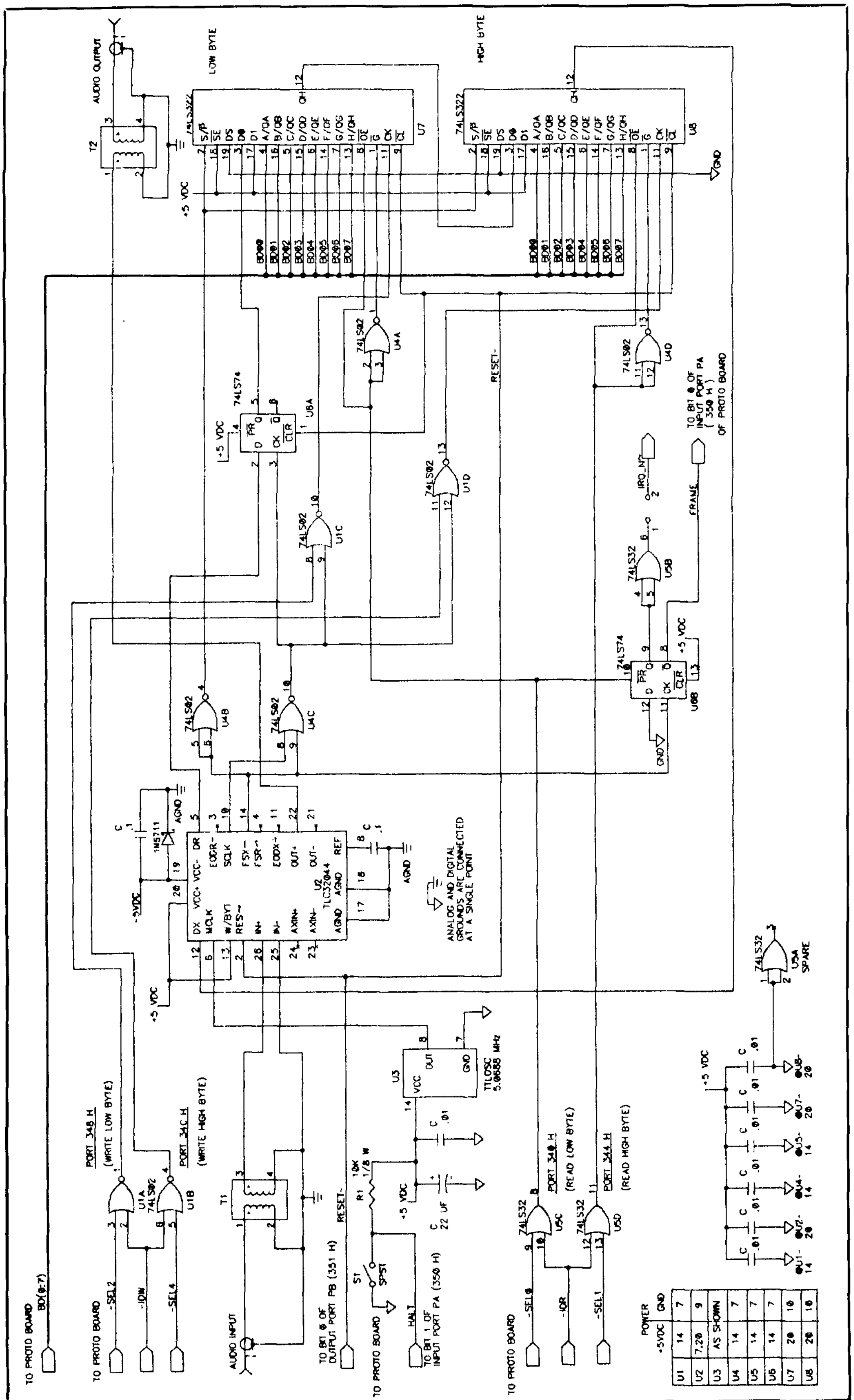

Figure 1. Complete analog interface (synchronous) for the PC using the JDR prototype board (PDS-601). T1 and T2 are 1:1 600-ohm isolation transformers.

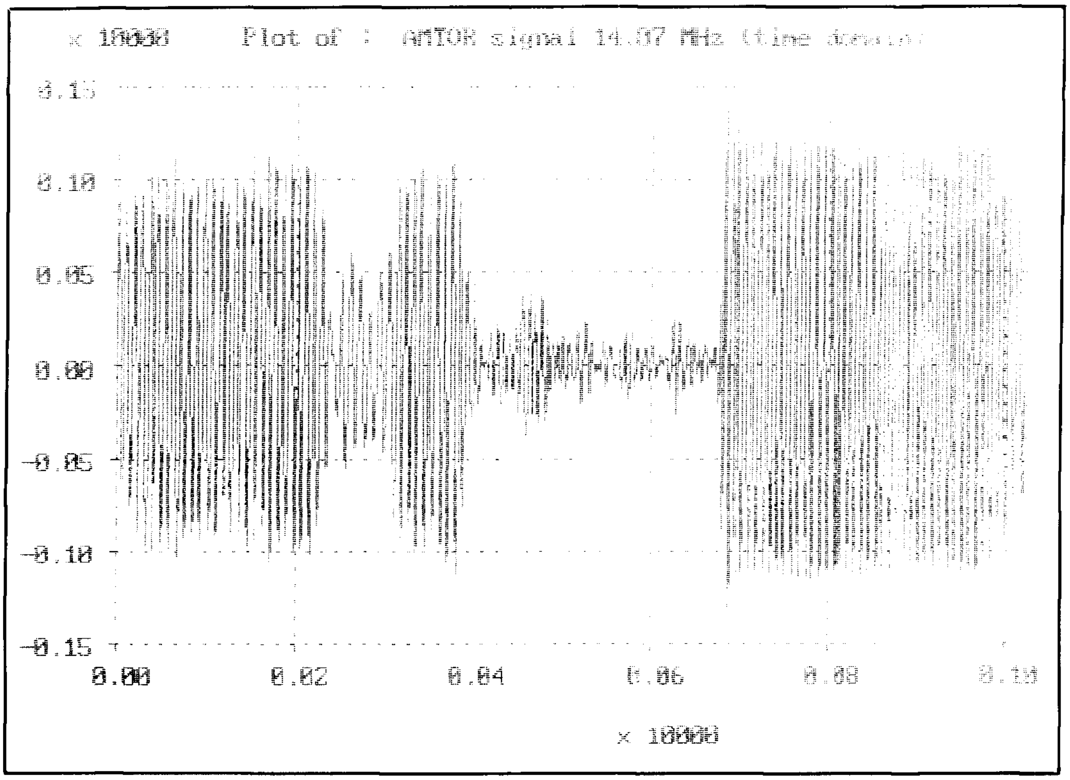

Figure 2. Plot of AMTOR signal 14.07 MHz (time domain).

MS-DOS version 3.0 or higher and at least 512 k-bytes of RAM. A CGA or EGA monitor is essential. The program will run with a nongraphics text display, but you'll miss out on its best features. If you have a math coprocessor, PC-DSP will detect its presence and use it to speed up the processing.

Operation

Waveform generation using the program's synthesis and manipulation operators soon becomes tedious. As I've mentioned, complex and more realistic signals are generated using a high-level language like BASIC or C and imported to PC-DSP. You use their math and file functions to compute the signal samples and write them to a DOS ASCII file. PC-DSP will input your file, assign it to a variable, and treat the data as its own.

Once data is input, you can start number crunching. FFTs, DFTs, correlation, inverse FFT-DFTs, and spectrum analysis techniques are a snap for PC-DSP. Convolution is computed and graphed in an instant.

Finite-impulse-response (FIR) and infinite-impulse-response (IIR) digital filters can be designed and analyzed within PC-DSP. Specify the characteristics of the filter, and the program provides the coefficients and analyzes the filter displaying (where appropriate) impulse response, poles and zeros, magnitude, log magnitude, phase and group delay plots. FIR filter design uses the Fourier series method (with a choice of windows), Parks-McClellan, or least-squares methods. Butterworth and Chebyshev IIR filters are designed using the bilinear transformation method. PC-DSP's ability to analyze designed filters is a powerful learning device. You can enter a filter design from a textbook or an article and plot its characteristics.

I often want a hard-copy printout of my filter's coefficients and characteristics. PC-DSP is deficient here, as it lacks a graphics print function. The only way to get a hard copy is to use the "print screen" key on the keyboard. Unfortunately, this only works if the proper driver program for your graphics printer is installed. Luckily, I found the correct driver software in the public domain on a BBS. You might try your print screen key with a graphic intensive plot on your display. If your printer prints nonsense, then PC-DSP will probably give the same results. This is the only serious flaw in this otherwise terrific software package.

After designing your filters, you can pass your data through them and plot each filter's output. This provides insights into what a DSP microprocessor chip will do for you when running a floating-point filter algorithm in real time. My first project was to generate 1024 samples of an FSK signal (1600 and 1800 Hz) keyed at 300 bps and sampled at 8

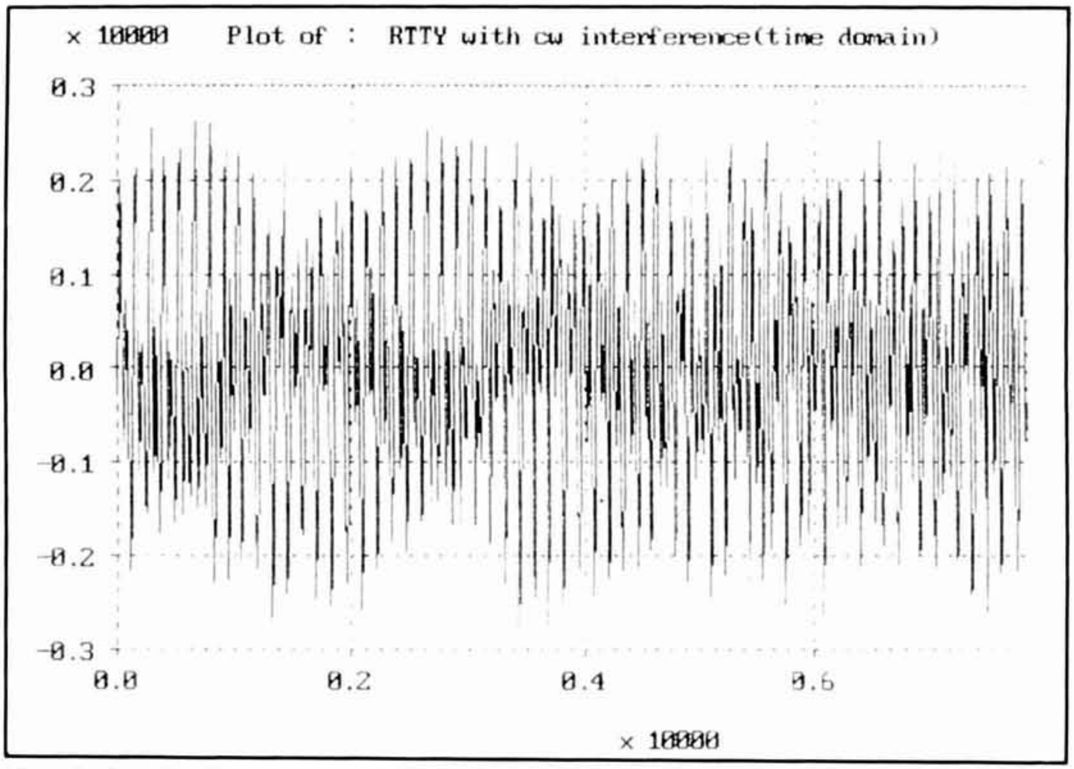

Figure 3. Plot of RTTY with CW interference (time domain).

kHz (0.128 seconds of signal) using a C program. The data was then read into PC-DSP. A plot of the data at this point is similar to a snapshot of an oscilloscope's display. To view the frequency spectrum, simply choose the FFT function followed by the magnitude menu selection and press F2 to view the results. Two spectrum lines are present at 1600 and 1800 Hz in the frequency domain. But you'll observe some smearing of energy into

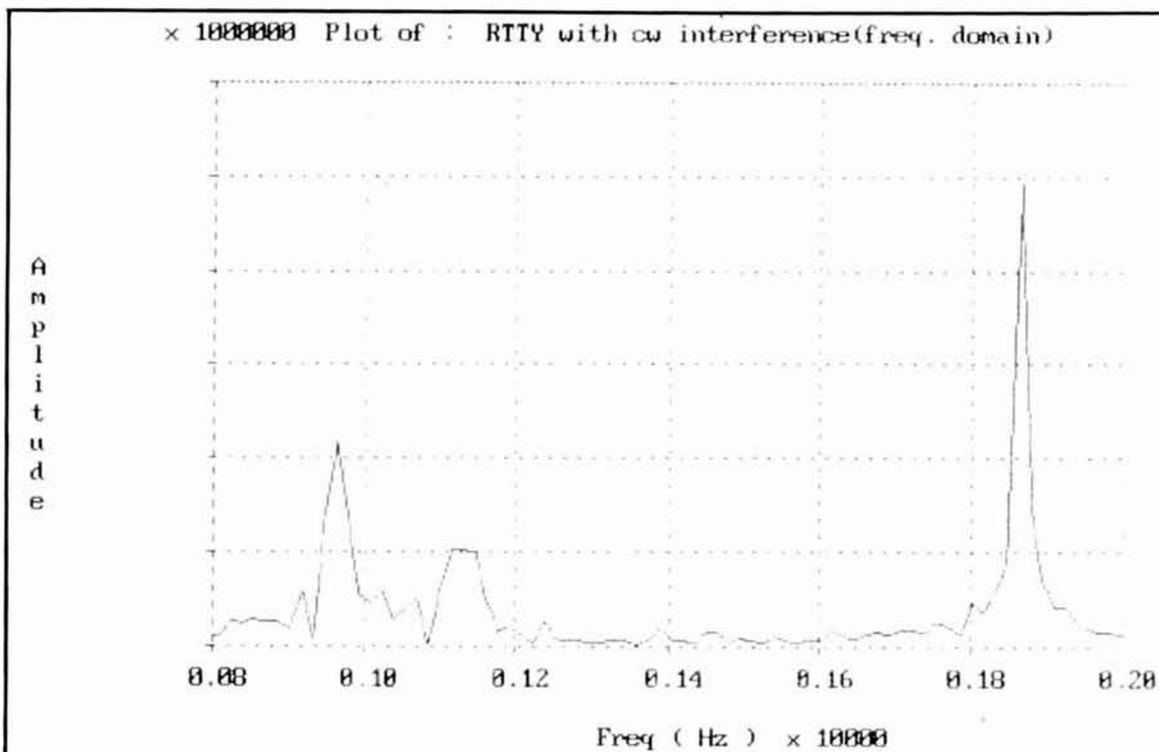

Figure 4. Plot of RTTY with CW interference (frequency domain).

frequency "bins" other than the two you expect. Find the answer to what causes this, and you're on the road to learning the power (and limitations) of DSP.

Very real signals

To paraphrase Mark Twain: "The difference between a real-world signal and an almost real-world signal is the difference between lightning and the lightning bug." One just is *not* a substitute for the other. I realized this after writing several programs that tried to simulate real-world communication signals for use as inputs to PC-DSP. The question was: "Just how real are my simulated signals?"

Clearly, for our hobby, the source for PC-DSP analysis should be live signals right off the ham bands. This would eliminate the time-consuming step of writing a program to generate signals of interest in the first place. Of course, there would be no doubt about the reality of the signals. Packet, RTTY, AMTOR, beacons, and CW signals from the low end of the bands are the ultimate reality.

To copy signals off the air, I needed some data acquisition hardware to interface my PC with my HF rig. Several articles have described such an interface.[1,2] Software drivers for the hardware interface convert the digitized samples to ASCII for input to PC-DSP.

The circuit I chose is shown in **Figure 1** and is built around Texas Instruments' TLC32044 Analog Interface Circuit (AIC). Besides the analog-to-digital (A-to-D) converter I needed to digitize the signals from the radio, the AIC includes a sample/hold circuit and an anti-aliasing filter for use in front of the 14-bit A-to-D. A digital-to-analog (D-to-A) converter and reconstruction filter are thrown in free. The D-to-A and reconstruction filter can be used to generate audio signals (not to the radio, but to a speaker). The AIC is a complete analog interface; only a few additional parts are required to interface with the PC.

I built the circuit on a JDR Microdevices** protoboard (catalog no. PDS-601). The JDR board has all the circuitry necessary to interface with the PC, including address decoding and data buffering. JDR also includes a solderless protostrip area for building up your custom circuits. The board I chose was rather expensive but a lower-priced 8-bit version is available. As most of the signals are low-speed digital, layout isn't critical. Also, because there's no need to solder, the interface makes a simple weekend project.

A few words about the TLC32044 AIC's signals are in order. This chip has an unusual serial interface. The PC doesn't understand this format, so U7 and U8 convert the data to parallel. The AIC has many operational

**JDR Microdevices, 2233 Samaritan Drive, San Jose, California 95124, (800)538-5000.

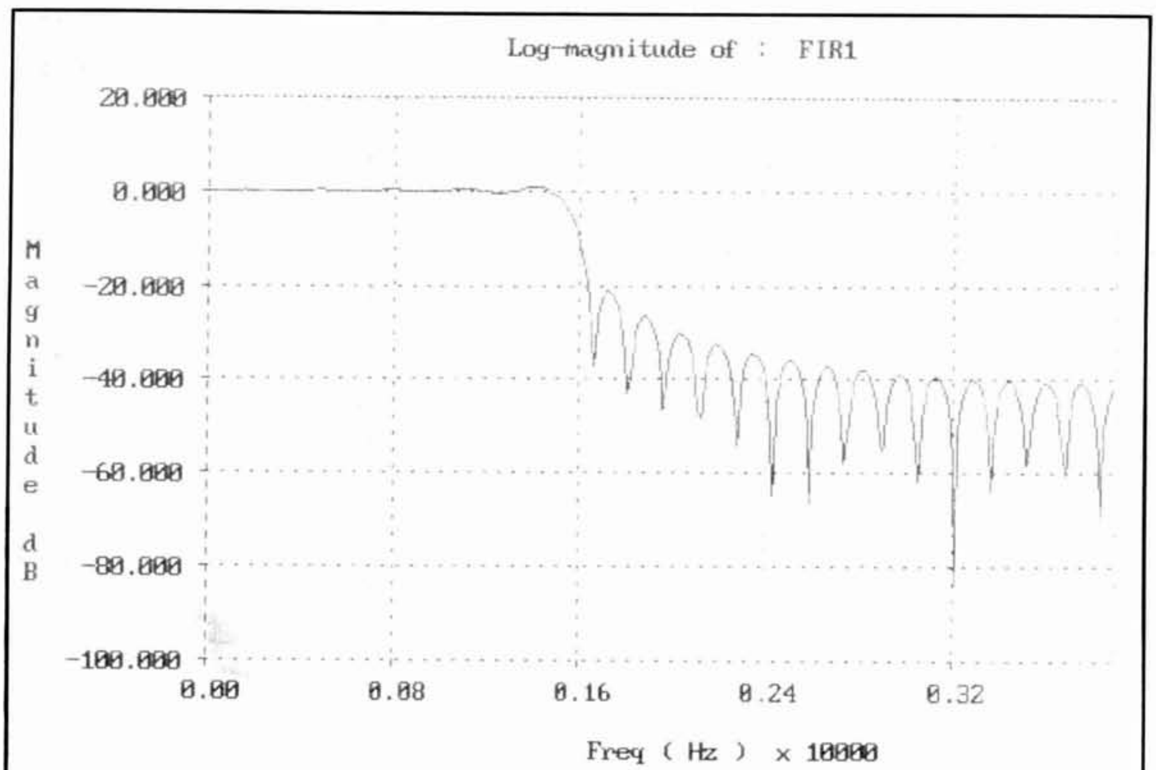

Figure 5. Log magnitude of FIR1.

modes, but the mode it comes up in after reset is all that you'll need. This is fine because the 4.77-MHz PCs are too slow to program it anyway. (The chip is normally used with the super-fast DSP chips. The slower PCs, even when programmed in assembly language, can't meet the chip's demand for command mode timing.) But even the slowest PCs can read the samples that occur at the frame rate of 128μs. The AIC runs in what TI calls the synchronous mode; that is, the A-to-D and D-to-A converters operate on the same timebase—which, again, is fine for this application. This mode has a handy feature. The A-to-D samples are read into U7 and U8 in one frame and (if the PC writes no new data) shifted out the D-to-A on the next. The signal you put on the analog input is looped-back to the analog output and can be heard on a speaker/amplifier. This is a quick check that everything (except the PC interface) is working.

The TLC32044 is available through Texas Instruments distributors, but they often require a sizeable minimum order. The TLC-32041, an earlier version of the AIC, is available from Newark Electronics*** ($25 minimum, TLC32041CN, $28.86, catalog no. 111). This AIC requires an external zener diode voltage reference and has different filter characteristics. Be sure to request the data sheet from TI and read it carefully before ordering the TLC32041. I also used a standard crystal oscillator frequency from JDR because the 5.1845-MHz crystal TI recommends wasn't available. I had no problems here. If the clock isn't correct, the internal switched-capacitor filter response is scaled ratiometrically. The filter's transfer function will be frequency-scaled by the ratio of the A-to-D conversion rate to 8 kHz. Here the conversion rate is 7.833 kHz, and after counting down the signal for the filter, an error of only about 2 percent results. This is not critical.

All that remains is to write some simple software to obtain the samples, convert them to ASCII, and write them to a file. My program starts up and waits for a keystroke before collecting the samples. This lets me lie in ambush waiting for certain signals or band conditions to occur.

***Newark Electronics (213) 638-4411.

PC-DSP at work: A case study in signal processing

I've included some plots of signals copied off the radio and processed by PC-DSP. **Figures 2** and **3** are AMTOR and RTTY time domain signals, respectively. **Figure 3** has been altered just a little. I wanted some practice designing digital filters, so I used PC-DSP to add some CW QRM to the signal as it was received. The interference is difficult to see in the time domain plot, but in **Figure 4**, a frequency domain or spectrum plot, the interference is unmistakable at about 1850 Hz.

After using PC-DSP to design a FIR filter

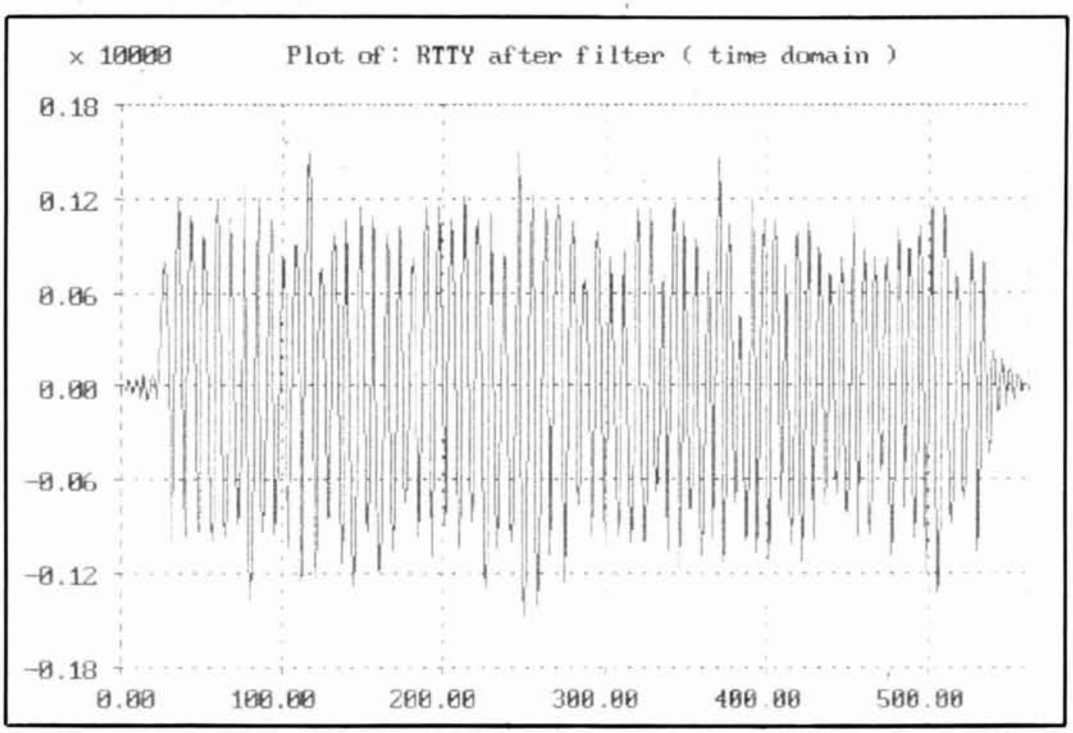

Figure 6. Plot of RTTY after filter (time domain).

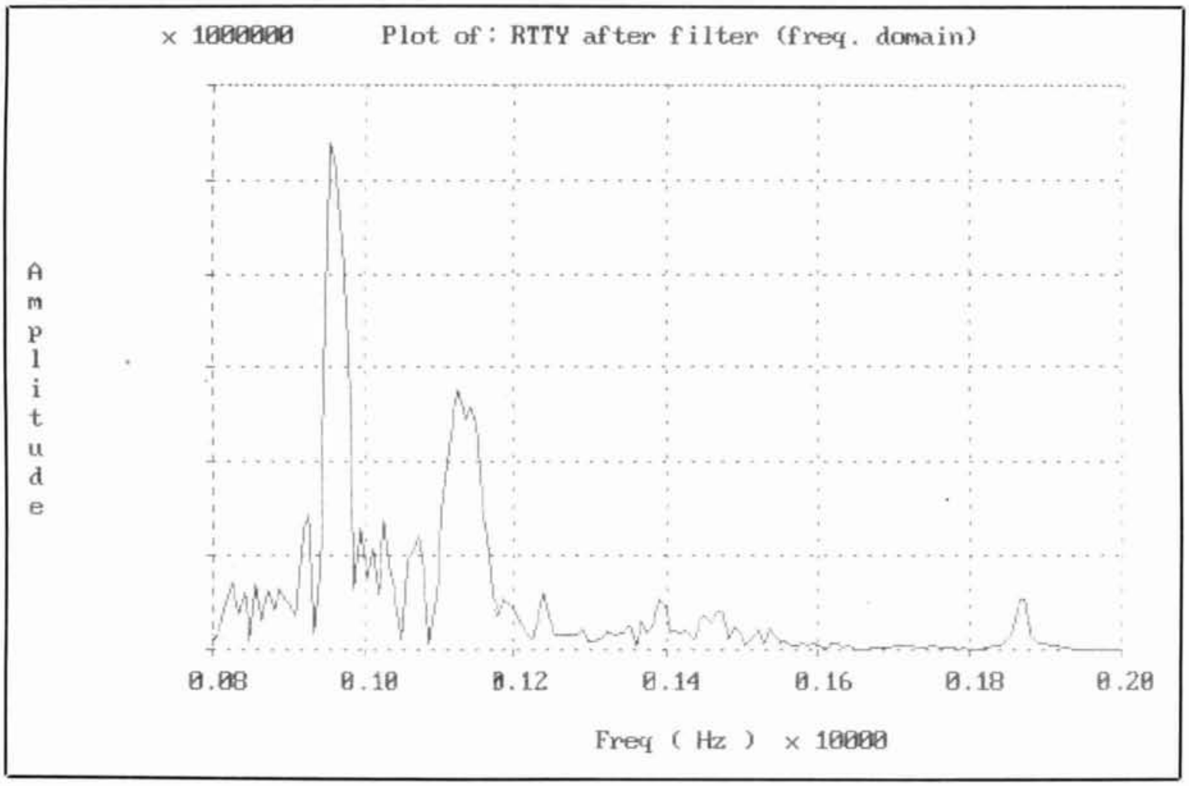

Figure 7. Plot of RTTY after filter (frequency domain).

with the gain-frequency characteristics in **Figure 5** and passing the RTTY signal with added QRM through it, I obtained the time-domain plot of **Figure 6. Figure 7** shows the same filtered signal in the frequency domain, with the 1850-Hz CW interference tone successfully knocked down.

The great thing about the work which I've presented here, is that I was able to perform DSP without making one calculation. It was all conceptual on my part; PC-DSP did all the math! My job was to string DSP operations together in a way I thought might improve the signal, or might let me examine the signal in a new way. PC-DSP was there to do the hard work and to keep me in the realm of the possible when I got carried away. There are some limitations to the program, but for the money, I don't see how it can be beat for teaching the fundamentals of digital signal processing. Add an interface to your radio to pick up signals off the air, and peek into the future of amateur radio. ■

REFERENCES

1. A. Kesteloot, "Practical Digital Signal Processing: A Simple, High-performance A/D Board," *QEX*, April 1990, pages 3 to 6.
2. A. Overcast, "Circuit Transforms PC Into Data Analyzer," *EDN*, May 25, 1989, pages 200 to 202.

PRODUCT INFORMATION

Bipolar Power Transistor Selector Guide

Motorola has revised its Bipolar Power Transistor Selector Guide and Cross Reference. The updated edition covers all applications categories including application specific devices. To obtain a free copy of the Bipolar Power Transistor Selector Guide and Cross Reference, call Motorola Literature Distribution at (800)441-2447, or write to: Motorola Inc., Literature Distribution Center, P.O. Box 20924, Phoenix, Arizona 85063. Ask for SG48/D.

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LORAN-C FREQUENCY CALIBRATOR

Calibrate frequency sources to better than 1 part in 10 easily and inexpensively

A common problem many hobbyists and experimenters encounter is the need to calibrate a frequency source accurately. With the proliferation of inexpensive frequency counters and synthesized signal generators and receivers, it's becoming essential to verify the accuracy of the internal frequency standards used in such instruments.

A simple technique for calibrating a frequency standard uses signals from radio station WWV. Unfortunately, this method provides accuracies of no better than 1 part in 10^7, due to ionospheric distortions. Also, signals from WWV can't be received at all locations at all times as a result of varying propagation conditions.

Another less common calibration technique uses the television network broadcast colorburst signal. When the colorburst signal is sent live from network studios, it's often generated using an atomic standard typically accurate to 5 parts in 10^{12}. This colorburst signal can be used to calibrate a frequency standard.[1] Unfortunately, we are located on the West Coast and seldom receive live broadcasts from the network studios because of the time difference and the use of delayed tapes for most broadcasts. Consequently, the colorburst technique is of marginal use to us. We thought it would be desirable to be able to calibrate frequency standards when we wanted to, not just when a signal was available.

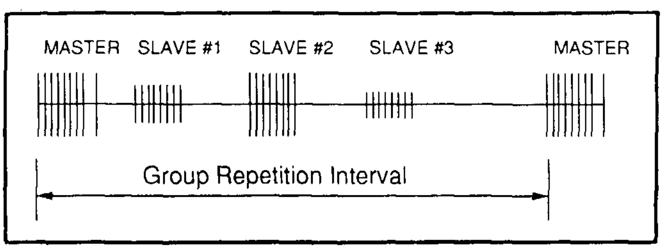

Figure 1. Typical LORAN-C transmit sequence.

We investigated and breadboarded a number of techniques—including the colorburst TV broadcast signals, and 20 and 60 kHz WWVB receivers—and found that the simplest technique used LORAN-C signals. These signals are received almost worldwide and give accuracies equal to atomic standards. LORAN-C transmits cesium standard derived signals at 100 kHz. The accuracy of the transmitted signal is on the order of 1 part in 10^{12}. Because the primary means of propagation is via ground wave rather than sky wave or the ionosphere, the propagation path is very stable and accuracies on the order of 1 part in 10^{11} can be achieved if you are within 1000 miles of a transmitting station.

We found an oven-controlled precision 10-MHz frequency oscillator standard at a local flea market that we could set to better than 1 part in 10^{10}. We designed and built a battery back-up system for the oscillator, with a frequency distribution system that lets us clock all of our frequency counters and synthesizers off this one precision reference source. Our problem has been determining how to set the frequency reliably and moni-

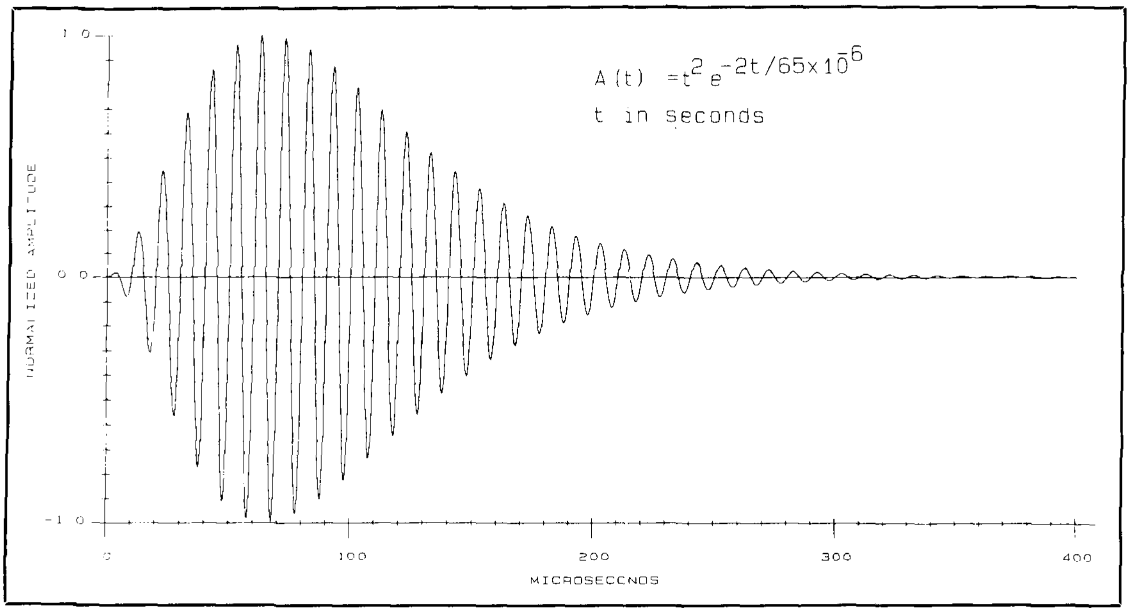

Figure 2. LORAN-C pulse waveform.

tor it for long-term accuracy to ensure that the frequency standard hasn't drifted excessively off frequency.

We'd like to present a technique which will allow you to calibrate frequency sources to better than 1 part in 10^{11} easily and inexpensively. While not many home frequency standards can even achieve and hold tolerances of this magnitude, it's nice to know that your frequency reference—LORAN-C in this case—is typically several orders of magnitude greater than you need.

Our frequency calibrator consists of a LORAN-C active antenna which is easily mounted outside, usually near the roof of a house, and a separate chassis which houses the actual calibrator. The calibrator has inputs for the active antenna, as well as the local frequency standard. An oscilloscope compares the signal received from the local

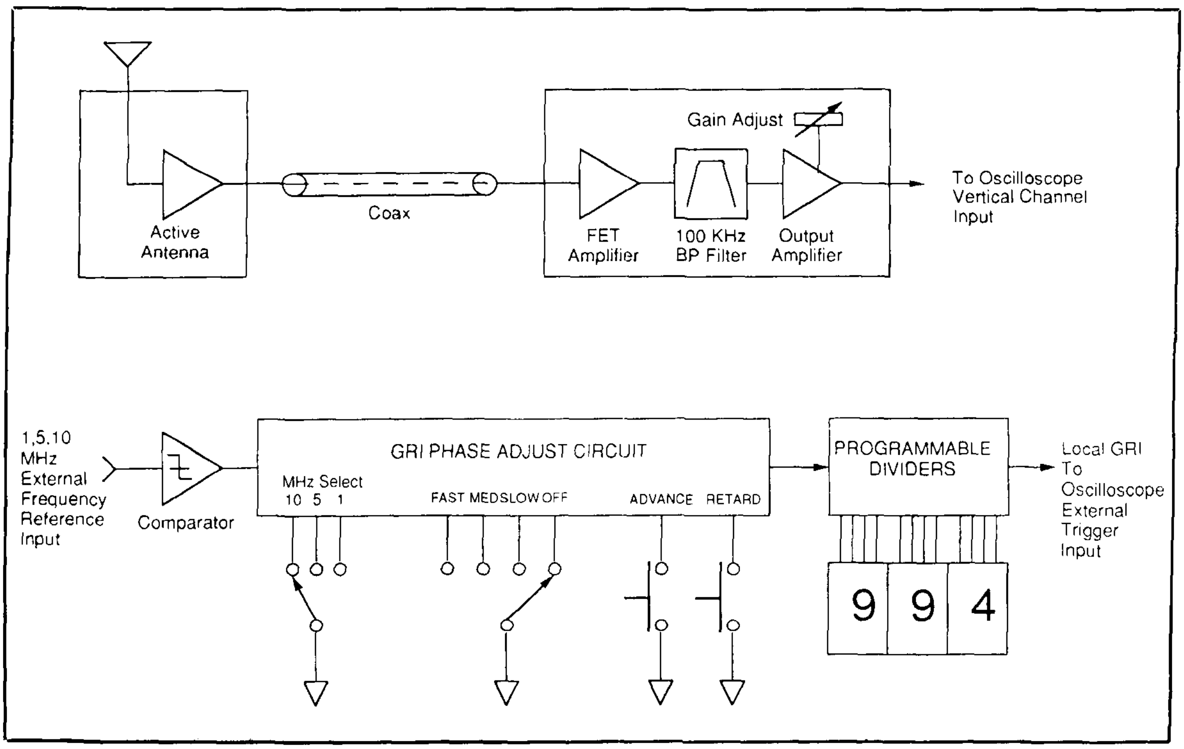

Figure 3. LORAN-C frequency calibrator block diagram.

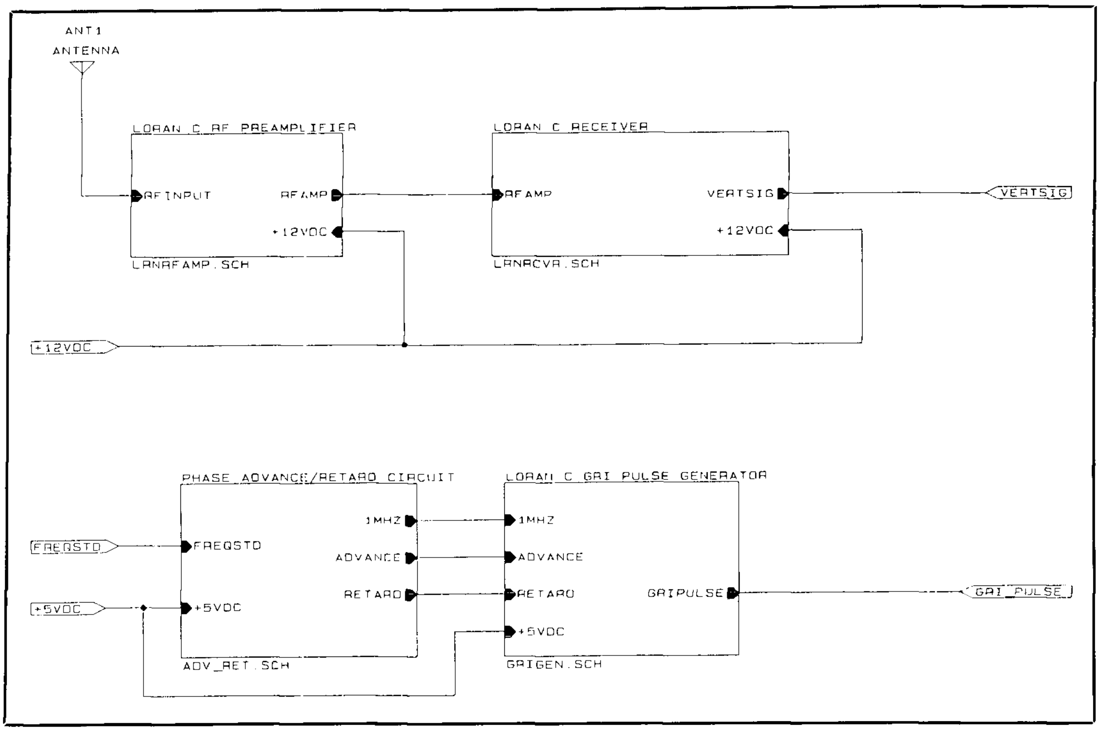

Figure 4. Interconnect diagram for the four modules which make up the LORAN-C calibrator.

frequency standard with the received LORAN-C signal. The calibrator accepts local frequency standard input signals at 1, 5, and 10 MHz at signal levels from 0.05 volts p-p up to 3 volts p-p, with an input impedance of 50 ohms.

Theory of operation

LORAN-C is a navigation system maintained by the United States Coast Guard that is based upon the measurement of the time difference between received signals from two different LORAN-C transmitter locations. The time differences are used to determine the location of the receiver site. Because the measurement of the time difference is critical to the accuracy of the navigation system, the LORAN-C system is designed to ensure that very stable signals arrive at the receiver. It is this inherent time stability that makes LORAN-C an ideal frequency reference.

The excellent time stability of LORAN-C signals results from the use of a 100-kHz carrier frequency. Because of this low frequency, the signals tend to travel primarily by means of ground wave instead of sky wave. The ground wave has a constant signal path from the transmitter to the receiver site because it follows the curvature of the earth; the sky wave has a variable path due to the fluctuations in height of the ionosphere during a 24-hour period.

Each LORAN-C geographical region consists of a master transmitting station and up to four slave transmitter stations. The master and slave stations transmit in sequence; each station transmits a pulse group of eight or nine pulses. The sequence repeats approximately every 100 ms. Each regional area has its own sequence length, called the group repetition interval (GRI). The master station starts the sequence with eight 100-kHz pulses spaced exactly 1 ms apart, followed by a ninth pulse sent 2 ms after the eighth pulse. Each pulse is approximately 250μs in duration. The slave stations transmit only eight pulses, also spaced exactly 1 ms apart. Each slave station transmits at a fixed but unique time after the master station transmits. The intervals between pulse groups are designed to ensure that the groups won't overlap within the reception range of the receive stations. A typical transmitted group sequence is shown in as received in **Figure 1**.

The transmitted LORAN-C signal is a carefully controlled shaped pulse sequence which uses a 100-kHz carrier and timing derived from an atomic frequency standard at each LORAN-C transmitter site. The pulse shape, shown in **Figure 2**, is designed to allow relative ease of differentiation between the stable ground-wave signal and the varying sky-wave signal. This is done by ensuring that the sky-wave pulse from a given pulse has decayed to zero before the ground-wave pulse arrives from the next one. Because the sky-wave pulse bounces off of the

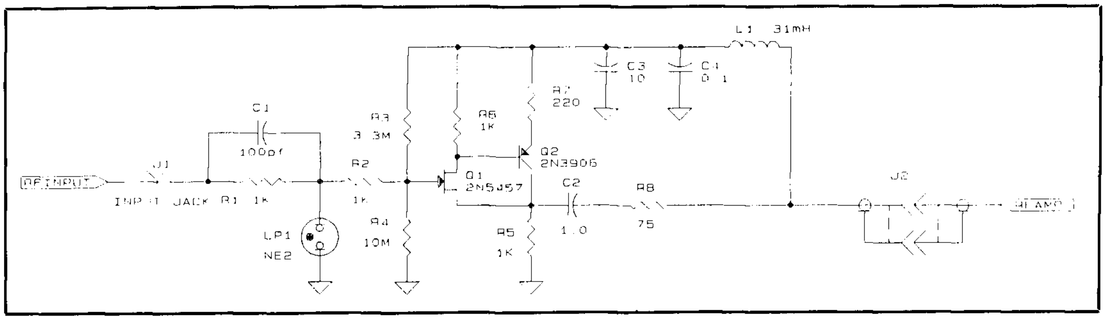

Figure 5. LORAN-C RF preamplifier circuit diagram.

ionosphere before it reaches the receive location, the sky wave always arrives after the ground-wave pulse begins. Consequently, the first few cycles of any received LORAN-C pulse are virtually assured to be composed of only a stable ground-wave component—assuming that the receive location is no more than 1000 miles from the transmitter location. Generally, the optimum signal point for tracking each pulse is the third-cycle zero crossing of each transmitted pulse.

Because the third-cycle zero crossing of any received LORAN-C pulse group is an extremely accurate and stable event, it can form the basis of a very accurate frequency *calibrator. The local frequency standard is* used to generate its own signal at the GRI of the regional LORAN-C chain. By comparing the LORAN-C GRI with the locally generated GRI, it's easy to make a precise measurement of the accuracy of the local frequency standard. You do it by triggering an oscilloscope from the local GRI and observing the received LORAN-C pulses on a vertical channel of the scope. By measuring the drift rate of the third-zero crossing of any received pulse at the GRI rate, you can easily adjust and calculate the accuracy of the local standard to several parts in 10^{12}.

Calibrator description

Figure 3 shows a block diagram of the calibrator system. Refer to this figure during the unit description.

LORAN-C signals are picked up by the active antenna, amplified, and sent to the LORAN-C receiver. The LORAN-C receiver filters and amplifies the received signal. The receive filter is centered at 100 kHz and has a bandwidth of 20 kHz to accurately reproduce the LORAN-C pulse on the oscilloscope. The receive bandwidth must be wide enough to pass the pulse without distortion, but narrow enough to remove any interfering signals. If high-level interfering signals are observed it may be necessary to adjust notch filters to suppress them.

The receiver uses a manual gain control, rather than an automatic gain control (AGC), to adjust the gain of the receiver. Since the receive signals at a given location are relatively stable for the LORAN-C ground-wave signals, we didn't feel it was necessary to use automatic gain. Also, an AGC would have required a complex circuit to prevent it from being controlled by extraneous signals that are likely to be present along with the desired LORAN-C signals.

The calibrator takes the local frequency standard input at 1, 5, or 10 MHz and divides it down to the regional GRI by means of fixed and programmable frequency dividers. To allow adjustment of the transition point of the local GRI signal so it corresponds with the third-zero crossing of any pulse in the LORAN-C pulse train, the frequency divider circuit lets pulses be momentarily added to or deleted from the frequency reference clock signal. By adding or deleting pulses from the clock stream, you can advance or retard the phase of the local GRI signal to the oscilloscope trigger input, and position any pulse in the LORAN-C pulse chain at any point on the oscilloscope trace. By monitoring the drift rate of the third-zero crossing as measured by the sweep speed of the oscilloscope, you may determine the accuracy of the local frequency standard using the equation: Accuracy = Delta t/ Measurement time, where Delta t is the drift amount measured on the oscilloscope and the measurement time is the time duration over which the drift measurement was made.

Circuit description

Figure 4 is an interconnect diagram for the four modules (excluding the power supply) which make up the LORAN-C calibrator. The active antenna/RF preamplifier is shown in **Figure 5**. It consists of an FET input amplifier followed by an emitter follower amplifier which drives the connecting cable to the calibrator circuit. Power is supplied, via the connecting coaxial cable, from the calibrator to the active antenna.

The FET amplifier presents a high input

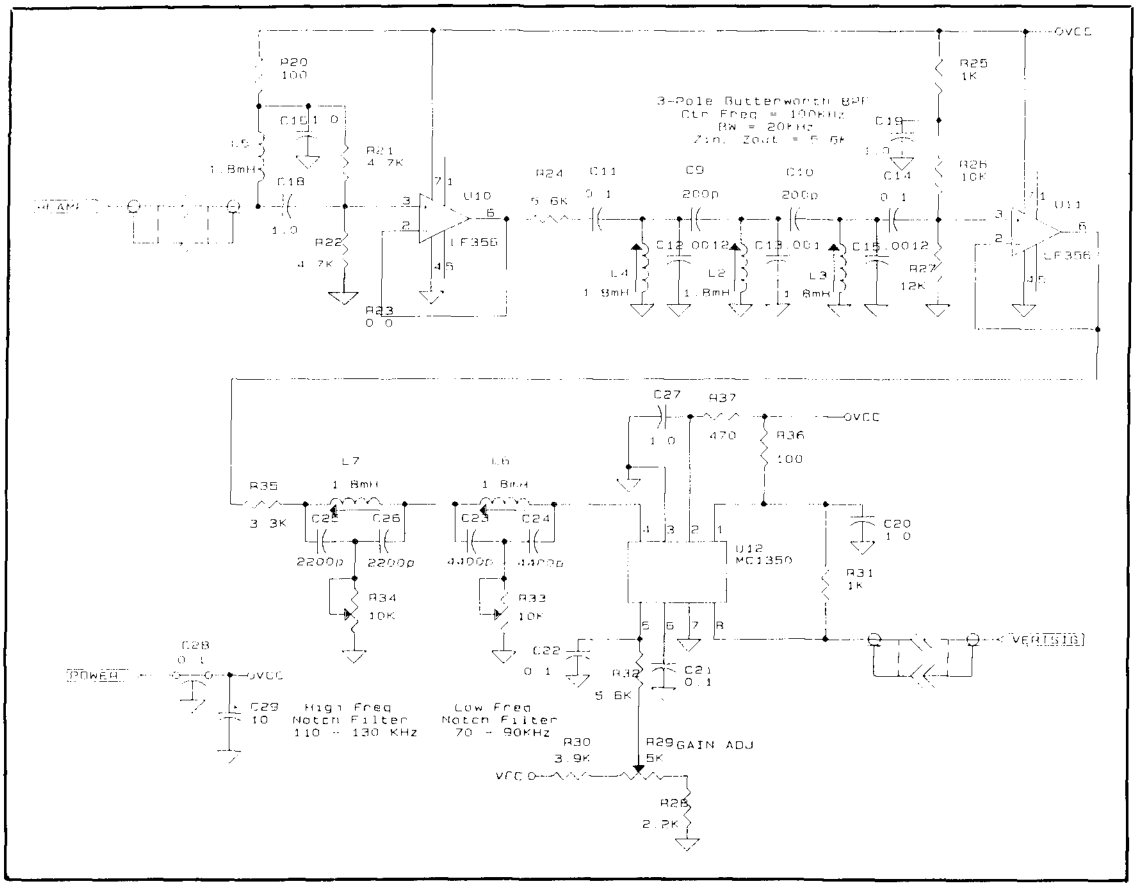

Figure 6. LORAN-C receiver.

impedance to the whip antenna, which may be a 6 to 20-foot vertical antenna. A high input impedance is required because the antenna length is much shorter than the wavelength of the 100-kHz LORAN-C signal, and such antennas characteristically have very high impedances. A low input impedance amplifier would load down the signal to the point where a usable signal could not be received.

Because the input amplifier is relatively broadband, it will pick up many high-level signals—including-60 Hz signals radiated from power lines. Therefore, the amplifier is designed to be able to amplify high-level signals without distorting the desired low-level LORAN-C signals. In this amplifier, signal levels of several volts may be coming into the LORAN-C receiver from the active amplifier. A low-impedance output is used from the active amplifier to minimize the pickup of any other extraneous signals.

The LORAN-C receiver is shown in **Figure 6**. The receiver is mounted in a shielded enclosure, and the input is an amplifier driving a three-pole Butterworth filter with a center frequency of 100 kHz and a bandwidth of 20 kHz. This filter is used to attenuate all but the desired LORAN-C signal from the many signals coming from the active amplifier.

The output of the receive filter goes to an integrated circuit amplifier, a readily available and inexpensive type MC1350, with a manual gain control. The MC1350 has a gain of better than 60 dB and a gain control range of better than 50 dB. The amplifier output is the filtered LORAN-C signals for connection to the vertical input channel of an external oscilloscope.

The GRI pulse generator consists of two circuits: the GRI phase-adjust circuit and the GRI programmable pulse-generator circuit. The GRI phase-adjust circuit is shown in **Figure 7**; the GRI programmable pulse generator circuit is shown in **Figure 8**.

The local reference-frequency input circuit on the GRI phase adjust circuit is a comparator that squares up the input signal for driving the digital circuitry. The use of a comparator allows for the accommodation of a wide range of input signal levels—from 50 mV p-p to 3 volts p-p. The phase-adjust circuit consists of a circuit to advance or retard the phase of the local 10-kHz clock that drives the GRI programmable pulse-generator circuit in precise 4-μs steps. The phase may be advanced or retarded continuously at a fast or medium rate by pushing the advance or retard phase buttons. It may be adjusted or disabled in single steps by positioning the phase-adjust mode switch.

The phase-adjust circuit is a digital circuit that allows a single pulse to be added to or deleted from the clock stream upon generation of a single advance or retard command pulse. The sequence of events is shown in **Figure 9**. As illustrated in the timing diagram, a single pulse is either removed from or added to the clock stream for every phase advance or retard pulse input. This translates into a GRI phase advance or retard change of exactly 4 μs because the pulses are added to or deleted from a 250-kHz clock stream.

A 555 timer chip generates a continuous stream of pulses for "fast" or "medium" phase-adjust mode switch selection whenever the advance or retard phase buttons are pushed. Fast mode moves the LORAN-C pulse chains rapidly across the oscilloscope trace; medium mode allows for a finer trace

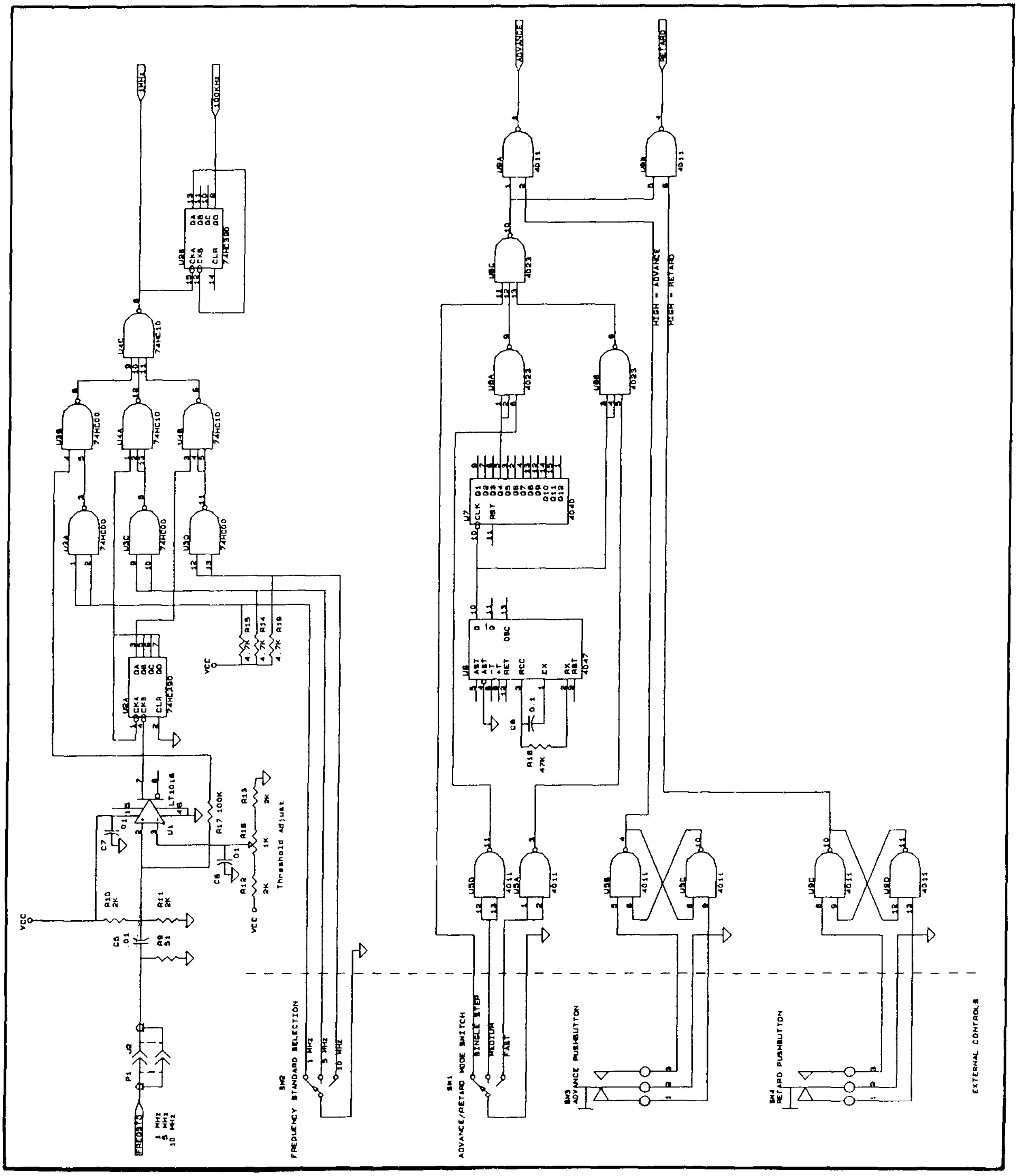

Figure 7. LORAN-C GRI pulse generator phase-adjust circuit.

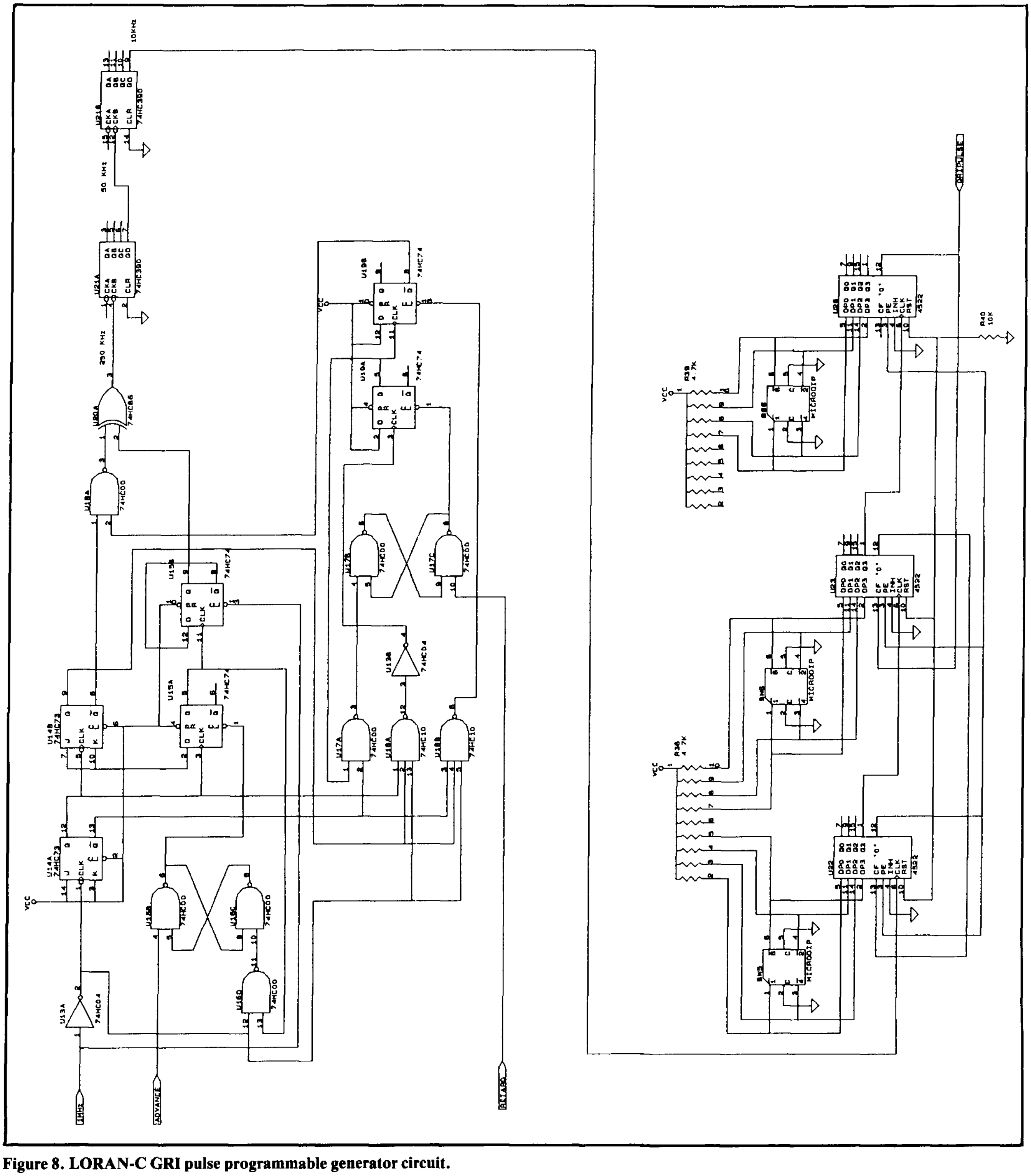

Figure 8. LORAN-C GRI pulse programmable generator circuit.

movement. The single-step mode generates a single pulse whenever the advance or retard phase button is pushed and allows a very fine adjustment of the GRI phase.

The parts list contains complete information on components for all four modules. We didn't list a specific overall enclosure; the box you use to house the unit isn't critical and almost any chassis will work. Nor is the actual antenna shown, because a variety of possibilities exist for implementation. We have used a 20-foot CB vertical, as well as ham VHF whips for the antennas. We do recommend, however, that you place the active antenna RF preamplifier and LORAN-C receiver circuits in separate shielded enclosures, as described in the section on construction.

Construction

Mount the active antenna in a shielded enclosure suitable for outdoor use. We chose a diecast aluminum box with an internal circuit board. Parts placement isn't particularly critical due to the low frequencies involved. Also, it's not particularly crucial that you use the FET and transistors specified. However, to assure maximum signal handling capability of the active amplifier, you should select a FET which provides a voltage of around 4 to 5 volts on the emitter of the output transistor.

You'll also need to mount the LORAN-C receiver circuit in a shielded enclosure to minimize pickup of external signals—particularly those of the digital circuits in the calibrator, if they are mounted in the same chassis. Parts placement is more critical in this circuit due to the high gain of the amplifier and the need for good input/output isolation of the 100-kHz filter. We mounted the circuit in an aluminum chassis which contained a double-sided pc board with a ground plane. Because the entire circuit works at a single frequency, we used an in-line layout to put maximum distance between the input and the output. There's been no sign of oscillation or instability, and the signals presented to the oscilloscope are very clean. Note that the gain control adjustment was mounted on the pc board rather than externally, as we have not found a need to adjust the gain of the unit after the initial setting.

To ensure signal stability, mount the comparator circuit for the reference frequency on a pc board with a ground plane. The last thing you need is an oscillating comparator driving the calibrator circuits!

You can wire wrap the digital circuits of the calibrator or mount them on a pc board. We mounted the switches for the GRI rate selection on the board, rather than locating them externally. The GRI is fixed for a given region and there's no need to change the setting unless you move from one region to another. For that matter, you can use inexpensive wire jumpers, in place of the more costly switches, for GRI selection.

You'll want to mount the GRI phase advance and retard push buttons and the phase-adjust mode switch on the front panel of the calibrator, as they must be adjusted each time the calibrator is used. The reference-frequency select switch should also be mounted on the front panel if different frequency standards are to be calibrated on a regular basis.

Adjustments

While the 100-kHz filter might appear to be difficult to adjust without a sweep-generator test setup, you can obtain a perfectly acceptable passband shape without sweeping the circuit at all. The secret is to preset the filter capacitors and inductors to the required values and then solder them into the circuit. With the availability of accurate and low-cost capacitance measurement meters, it's relatively easy to preset the capacitors (1 or 2 percent capacitors may be purchased and paralleled as required). Measure the inductors at the operating frequency of 100 kHz because the value of ferrite tuned inductors is sensitive to frequency. Do not use any of

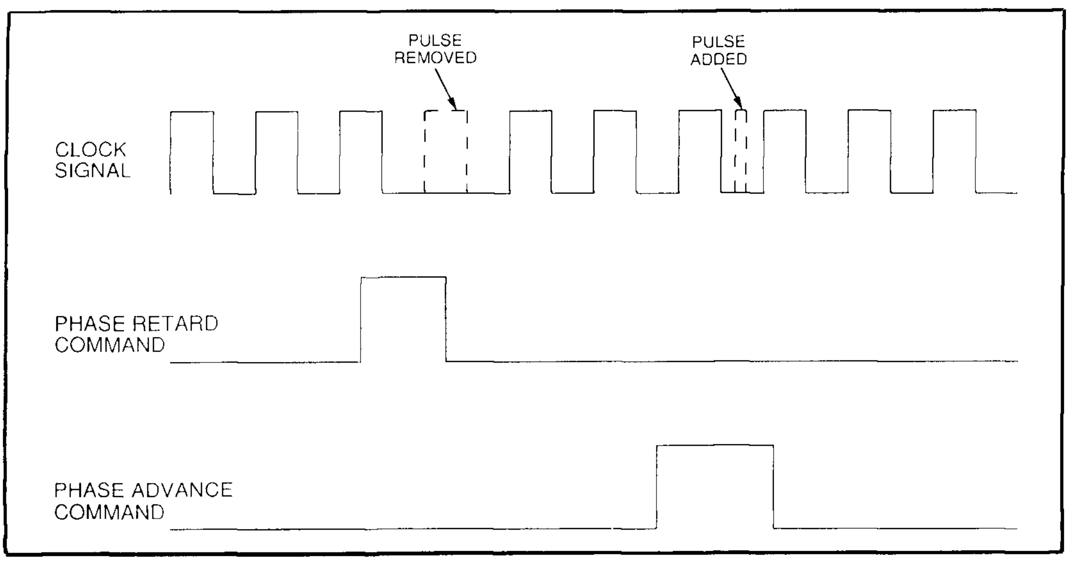

Figure 9. GRI phase advance/retard circuit timing diagram.

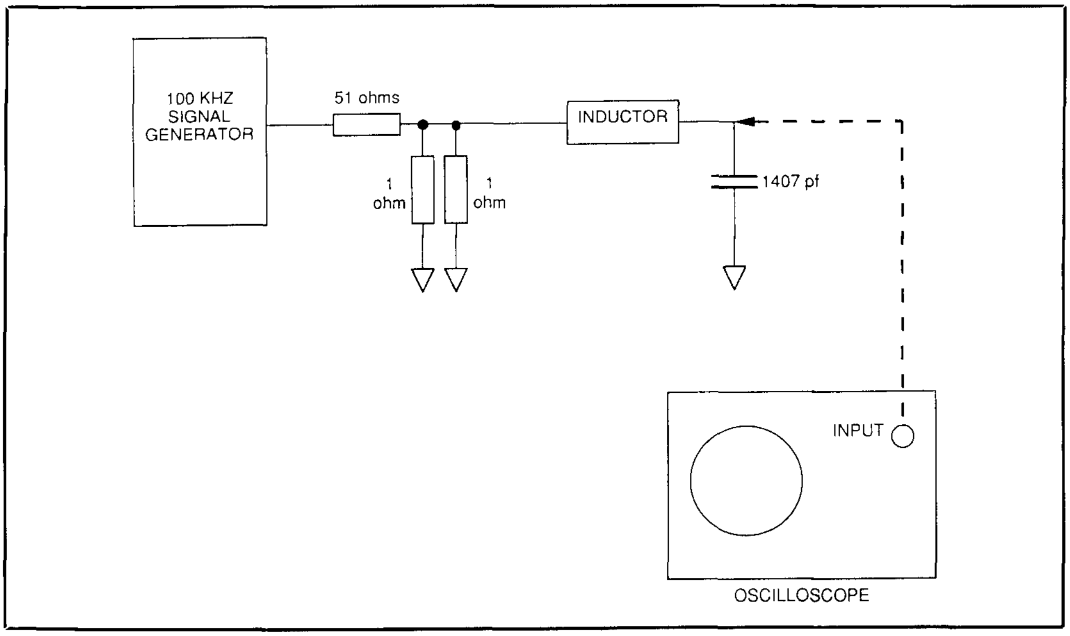

Figure 10. Bandpass filter inductor measurement setup.

the inexpensive inductance meters. Many of them use very low frequencies to measure inductance; for example, 60 Hz.

Figure 10 shows the method used to set the inductors—a classic inductance Q-meter measurement technique. The resistor divider at the input establishes a low drive impedance for the Q-meter circuit. (Note: To see a pronounced signal peak, the source impedance must be less than 1 ohm). The 51-ohm resistor matches the output impedance of the signal generator. It may be desirable to replace this resistor with the proper load value if the signal generator has a different output impedance, say, 600 ohms. With a precise capacitance value of 1407 pF (including stray capacitance) for C1, adjust the inductor for peak signal as seen on the oscilloscope. In practice, the capacitor can be a 1400-pF mica type, plus or minus 14 pF.

After selecting the filter components and installing them into the circuit, verify the filter's frequency response. Adjust the signal generator for a 100-kHz signal at 0.1 volts p-p output level to the input to the active amplifier. Make sure the output of the active

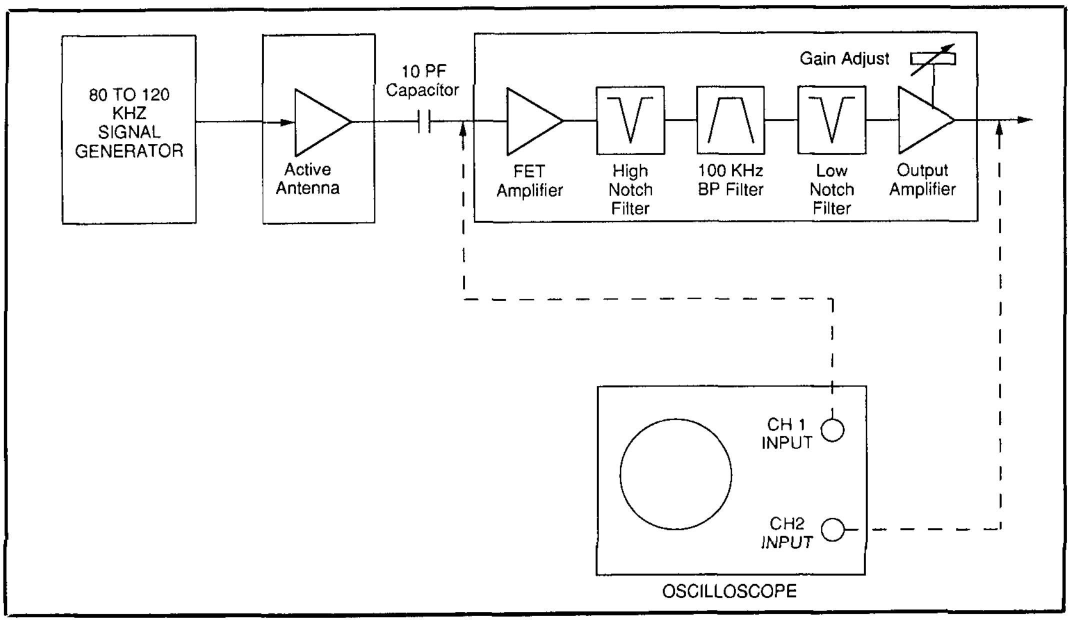

Figure 11. The 100-kHz bandpass filter adjustment setup.

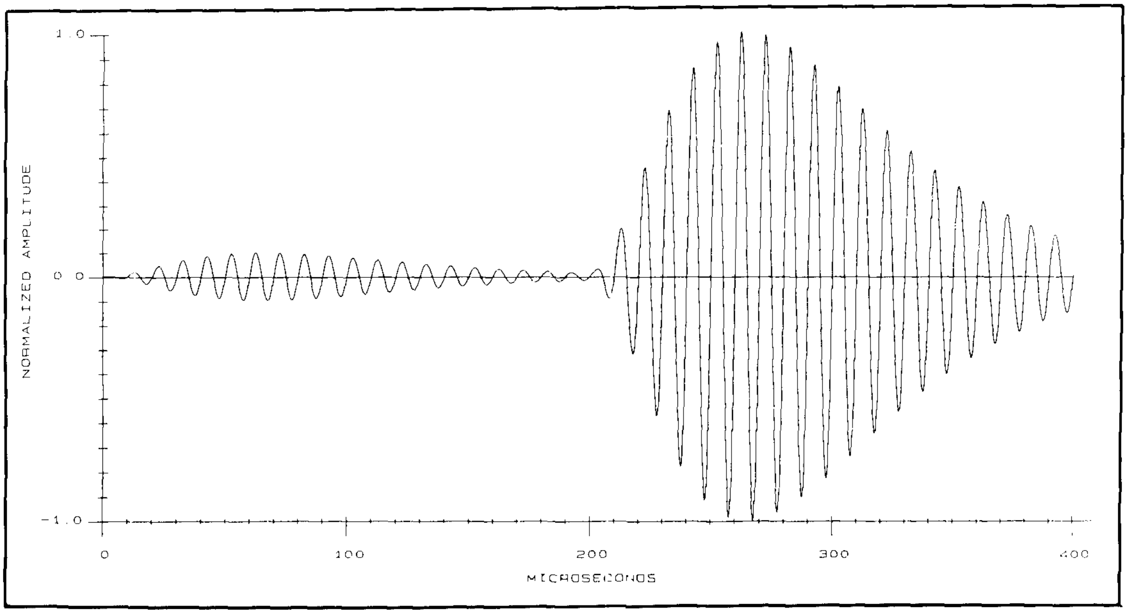

Figure 12. Sky-wave contaminated LORAN-C pulse waveform. Sky wave arrives 200 μs after ground wave. Sky- wave amplitude is 10 times the ground-wave amplitude.

amplifier is approximately 0.05 volts p-p into the LORAN-C receiver. Next, connect the oscilloscope to the input of the MC1350 RF amplifier using a 10X probe to minimize loading effects. Adjust the two notch filters for maximum inductance for the low-frequency notch filter and minimum inductance for the high-frequency notch filter. This effectively places the notch frequencies well away from the 90 to 110-kHz bandpass range.

Now verify that the filter has the proper frequency response shape using the test setup shown in **Figure 11**. Sweep the signal generator from approximately 80 to 120 kHz and confirm that a flat response exists across the 93 to 107 kHz range, with about a 3 dB roll off (0.707 times the maximum in band voltage at 100 kHz) at approximately 90 and 110 kHz. The preset values should require only minor tweaking, if any at all, to obtain the desired response.

Set the comparator threshold adjustment by sending a low-level signal of approximately 0.05 Vp-p at 1, 5, or 10 MHz into the comparator input. Set the threshold adjustment for a square-wave output signal from the comparator. Varying the input signal level up to 2 Vp-p shouldn't change the duty cycle of the square wave significantly.

After you've made these adjustments, connect the active antenna and mount it outdoors—preferably near a rooftop (disconnect the antenna whenever lightning is likely). Set the gain control adjustment by monitoring the LORAN-C receiver output with an oscilloscope set for a sweep speed of 10 ms per division. Adjust the gain control until you observe a number of signals on the oscilloscope. Now adjust the gain control until you see clipping, and readjust the control until the largest signal you see is approximately half the amplitude at which clipping occurred. This should provide an amplifier output which is clean and free from any distortion products that would be generated by an overloaded amplifier.

You can adjust the notch filters to suppress any high-level signals that may be getting through the filter and are clearly unrelated to the desired LORAN-C signals. In the absence of a spectrum analyzer, this is best done by monitoring the output of the LORAN-C receiver on the oscilloscope and adjusting the high and/or low-frequency notch filter inductor and null-depth resistor to minimize the interfering signal. Be sure your adjustments don't affect the LORAN-C signals adversely.

Operation

Frequency calibrator operation is very simple. Just set the phase of the local GRI signal so you can observe the LORAN-C pulse train drift rate on the oscilloscope trace. Do this by first connecting the local frequency reference to the calibrator and selecting the correct switch setting (1, 5, or 10 MHz input) for the reference-frequency select switch. Then, set the local GRI rate to that of the geographical region closest to your location. **Table 1** lists the GRIs for most LORAN-C regions. The GRI switches use only the first three digits of the GRI rate listed in the table. Now, connect the LORAN-C receiver output to the vertical input of the

LORAN-C Chain	GRI
Northeast USA	99400 μs
Southeast USA	79800 μs
Great Lakes USA	89700 μs
West Coast USA	99400 μs
Northwest Pacific	99700 μs
North Pacific	99900 μs
Canadian West Coast	59900 μs
Canadian East Coast	59300 μs
Labrador	79300 μs
Icelandic	99800 μs
Gulf of Alaska	79600 μs
Norwegian Sea	79700 μs
Mediterranean Sea	79900 μs
Central Pacific	49900 μs
East Asia	59700 μs
Western USSR	60000 μs
Eastern USSR	79500 μs

Table 1. Regional LORAN-C group repetition intervals.

oscilloscope and set the oscilloscope sweep speed for 10 ms per division. Adjust the vertical sensitivity so the largest signal you see on the oscilloscope is seven to eight divisions in height. If everything is working correctly, you should observe stationary LORAN-C signal pulse trains across the oscilloscope trace. The display should look similar to that shown in **Figure 1**. However, your display may differ because your LORAN-C region may have a different number of slave stations, and/or the relative amplitudes may vary due to different received signal levels from each LORAN-C transmitter.

There's a good chance that you'll see other LORAN-C signals drifting across the oscilloscope trace. These are LORAN-C signals from other regions which are using a different GRI rate and being received at your location. If you don't see any stationary LORAN-C signals, verify that you have set the GRI rate for your region correctly.

If you observe stationary LORAN-C pulse trains on the oscilloscope, turn the phase-adjust mode switch to fast. Push either the GRI phase advance or retard push button to move the largest LORAN-C pulse train to the left-hand edge of the oscilloscope trace. Increase the oscilloscope sweep speed to about 1 ms per division. Then turn the phase-adjust mode switch to medium, and push the GRI phase advance or retard push button again, to move the largest LORAN-C pulse train to the left-hand edge of the oscilloscope trace. Increase the oscilloscope sweep speed to about 100 μs per division. Turn the phase-adjust mode switch to single step, and push the GRI phase advance or retard push buttons to move the largest LORAN-C pulse train to the middle of the oscilloscope trace. Increase the oscilloscope sweep speed to about 10 μs per division. Adjust the phase advance or retard push buttons to center the third-cycle zero crossing of the LORAN-C pulse on the oscilloscope trace.

At this point, you'll probably see the third-cycle zero crossing drifting either to the right or left of the oscilloscope trace. The drift rate is a measure of the amount of frequency error in your local frequency standard. If you want to calibrate your frequency stand-

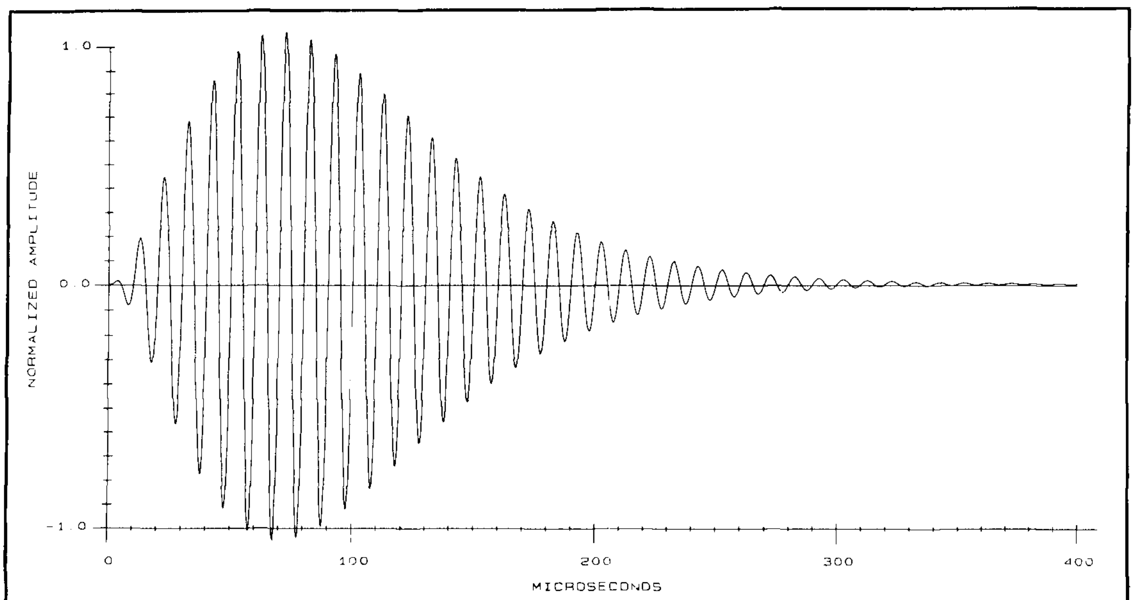

Figure 13. Sky-wave contaminated LORAN-C pulse waveform. Sky wave arrives 40 μs after ground wave. Skywave amplitude is 0.1 times the ground-wave amplitude.

ard, adjust the local frequency standard until you observe no drift on the oscilloscope trace.

To measure the amount of frequency error present in your local frequency standard, monitor the drift rate of the third zero crossing as measured by the sweep speed of the oscilloscope. For example, suppose the sweep speed is 5 μs per division and the zero crossing moves 2.6 divisions in 10 minutes. The total drift is 13 μs (5 μs per division $\times$ 2.6 divisions) in 600 seconds (10 minutes $\times$ 60 seconds per minute). As stated earlier, the accuracy of the local frequency standard may be determined by the equation: Accuracy = Delta t/Measurement time, where Delta t is the drift amount measured on the oscilloscope and measurement time is the time duration over which the drift measurement was made. For the example above, if a total drift amount of 13 μs was measured on the oscilloscope over a 10 minute period (or 600 seconds), the accuracy is $13 \times 10^{-6}/600$, or 2.167 parts in 10^8.

If you are interested in obtaining the best possible accuracy from the Loran-C calibrator system, you must take precautions to ensure that you're tracking the third-cycle zero crossing of the ground-wave component of the Loran-C signal and not the sky-wave component. The ground-wave component suffers high attenuation levels as it travels over the Earth's surface, while the sky-wave component can maintain very high levels at great distances from the transmitter site. Consequently, the Loran-C sky-wave signal can be much larger than the ground-wave signal at distances greater than 1000 miles from the transmitter, as shown in **Figure 12**. If you are within 500 miles of the transmitting site, the ground-wave signal will tend to predominate as shown in **Figure 13**. Note that even though the waveform is distorted by the sky-wave component, it's not visually apparent, and there's no impact on the third-cycle zero crossing. Hence, the rationale for always trying to track a zero crossing as early as possible after the start of the waveform.

In all cases, however, the ground-wave component will arrive at the receiver site first due to its shorter propagation path. Therefore, it's important to track off of the third-cycle zero crossing of the first LORAN-C signal pulse you observe in a burst sequence. Note that for distances greater than 1500 miles from the transmitter site, the ground-wave component is attenuated to virtually zero and you are forced to depend upon only the sky-wave component. Due to the shifts in the height of the ionosphere, particularly during the day/night transitions which occur around sunrise and sunset, accuracy is de-

PARTS LIST

Quantity	Reference	Part
1	ANT1	antenna
1	C1	100 pF
7	C2,C16,C17,C18, C19,C20,C27	1.0 μF
2	C3,C29	10 μF
7	C4,C8,C11,C14,C21,C22,C28	0.1 μF
3	C5,C6,C7	0.01 μF
2	C9,C10	200 pF
2	C12,C15	0.0012 μF
1	C13	0.001 μF
2	C23,C24	4400 pF
2	C25,C26	2200 pF
1	J1	input jack
1	J2	BNC, female
1	LP1	NE2
1	L1	31 mH
6	L2,L3,L4,L5, L6,L7	1.8 mH
1	P1	BNC, male
1	Q1	2N5457
1	Q2	2N3906
7	R1,R2,R5,R6, R16,R25,R31	1 k
1	R3	3.3 meg
1	R4	10 meg
1	R7	220 ohm
1	R8	75 ohm
1	R9	51 ohm
4	R10,R11,R12, R13	2 k
7	R14,R15,R19, R21,R22,R38, R39	4.7 k
1	R17	100 k
1	R18	47 k
2	R20,R36	100 ohm
2	R24,R32	5.6 k
4	R26,R33,R34, R40	10 k
1	R27	12 k
1	R28	2.2 k
1	R29	5 k
1	R30	3.9 k
1	R35	3.3 k
1	R37	470 ohm
1	SW1	advance/retard mode switch
1	SW2	frequency standard selection
1	SW3	advance push button
1	SW4	retard push button
4	SW5,SW6,SW7,U26	microdip
1	U1	LT1016
2	U2,U21	74HC390
3	U3,U16,U17	74HC00
2	U4,U18	74HC10
2	U5,U9	4011
1	U6	4047
1	U7	4040
1	U8	4023
2	U10,U11	LF356
1	U12	MC1350
1	U13	74HC04
1	U14	74HC73
2	U15,U19	74HC74
1	U20	74HC86
3	U22,U23,U24	4522

graded. If you're forced to use just the sky-wave component, try to make your measurements at the same time each day, but avoid times around sunrise and sunset. Fortunately, very few individuals should have to deal with this problem because of the world-wide distribution of LORAN-C transmitter sites.

We have tried to give you as much information as possible on using LORAN-C as a highly accurate frequency reference, but it's impossible to cover everything in one article. We have included a bibliography of additional readings for those who want more information on the subject of LORAN-C, as well as information on other techniques for calibrating frequency standards. ■

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PRODUCT INFORMATION

Flukes and Philips ScopeMeter®

John Fluke Mfg. Co. Inc. (Everett, Washington) and Philips Test and Measurement (the Netherlands) have introduced a new series of handheld service instruments which combines two test instruments into one. The Fluke 93, 95, and 97 ScopeMeters join a dual-channel digital storage oscilloscope (DSO) with a digital multimeter and include such features as AUTOSET, waveform and set-up memory, combined display of meter results and waveforms, an optically isolated RS-232C interface for instrument calibration, menus, and softkeys. The top-of-the-line model also features a built-in signal generator, component tester, optically isolated RS-232 remote control and printer interface, and backlit display.

Fluke ScopeMeters provide a 50-MHz bandwidth and a 25 MS/s sampling rate. You also get a 40-ns glitch capture time to catch intermittent failures, and storage capabilities of up to eight waveforms and 10 setups. The AUTOSET feature automatically sets volts per division, time per division, and position and triggering controls for any scope input signal. In meter mode, AUTOSET automatically tracks the input signal for the proper range, time per division display, and triggering.

The units have extensive multimeter capabilities along with: diode test; Min Max recording, which provides simultaneous display of maximum, minimum, average, and present readings; relative and percent-relative modes; dBm, dBV, and dBW readings; and autoranging.

The 4-pound ScopeMeter is enclosed in a sealed industrial package for use in the field service environment. The battery-operated instrument runs on either rechargeable NiCad batteries (included) or common alkaline C-cell batteries. Each ScopeMeter comes with probes which have high voltage and high frequency tips, an AC line adapter, a built-in battery charger, a holster, and a tilt stand which can be adjusted to hang over a panel or door.

For more information on the Fluke 90 series of ScopeMeters contact: John Fluke Mfg. Co., Inc., P.O. Box 9090, Everett, Washington 98206. In Europe, contact: Philips Test and Measurement, Building TQIII-4 5600 MD, Eindhoven, the Netherlands.

New HP Catalog Offers Selection of System-Power Products

Hewlett-Packard Company has released the HP-IB System DC Power Supplies and Electronic Loads Catalog. The catalog contains information on HP's entire line of DC power supplies and electronic loads, in addition to 42 recently introduced HP products.

To obtain a free copy, write to Hewlett-Packard Company Inquiries, 19310 Pruneridge Avenue, Cupertino, California 95014. Ask for Lit. 5091-2525EUS.

Steve Powlishen, *K1FO*
816 Summer Hill Road
Madison, Connecticut 06443

432-MHz EME 1990s STYLE

Part 2: Assembling your first EME system

Part 1 of this article detailed the rapid change in UHF and EME technology during the 1980s. This technological revolution succeeded in moving EME operation on the lower-frequency amateur UHF bands out of an experimental enclave of advanced equipment builders, to the homes of regular operators who use primarily commercial equipment. In Part 2 of this article, I'll continue the process of demystifying EME as I show you how to assemble your own 432-MHz EME system.

Your first EME array

As I noted in Part 1, EME QSOs are now being made regularly with single Yagi antennas. For your first array, however, I would recommend something a bit more substantial. Single Yagi operators thus far have been either the leading-edge crazies, who want to prove that it's possible to work EME in this fashion, or those passing through on their way to a larger array. A larger array allows for a greater margin of error in case a piece of the system isn't working properly. This extra system performance will give you a better chance for success, the ability to work many more stations than just the "big guns," and the power to work EME when conditions are less than ideal.

Today, a four-Yagi system is the best compromise in cost, complexity, and performance. The four-Yagi system has a manageable amount of complexity. It uses only a single power divider and four phasing lines. When the Yagis are arranged in a 2 × 2 box, the array can be mounted on a simple stacking frame that lends itself naturally to elevation movement. The low parts count translates into both reasonable cost and easy assembly. Easy assembly, in turn, means less construction time and more time to operate. Four modestly sized Yagis will have high enough performance capabilities to allow you to work many different stations and make random QSOs, even when conditions are less than optimum. Such an array will also let you hear your own echoes when Faraday rotation cooperates. The success of modest four-Yagi arrays is borne out by two of the many currently active four-Yagi stations. K2OS worked over 50 different EME stations in about a year using four 17-foot long Yagis; SM0PYP worked over 100 different stations in 5 years using four similarly sized Yagis (approximately 15 feet long). If you start with four Yagis of reasonable size, you'll have the building blocks for array expansion should the EME bug really bite you, and you decide to expand the array and join the EME big guns.

Just how small can a good 432-MHz EME array be? **Figure 1** shows a typical small four-Yagi 2-meter array compared to a small four-Yagi 432-MHz array. The 2-meter array uses 2.2-wavelength NBS Yagis with 15-foot long booms. This type of array has been widely used on 2-meter EME. The 432-MHz array uses the FO-22 6.1-wavelength Yagi, which has a 14-foot long boom. When combined with the proper station electronics, both of these arrays will give similar EME performance on their respective bands.

Now let's make a closer comparison of the two arrays. You'll note that while the boom lengths are almost the same, the stacking distance for the 432-MHz Yagis is about half that of the 144-MHz antennas, making the overall array size a diminutive one-fourth that of the 144-MHz array. It's interesting to

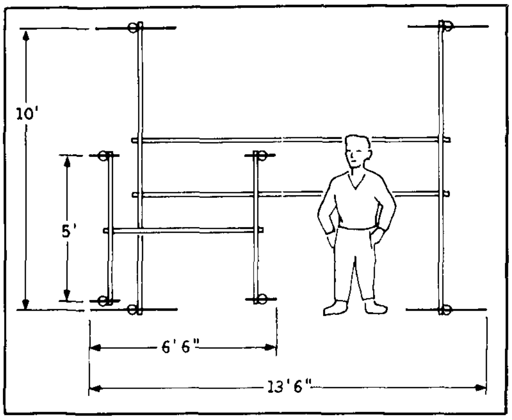

Figure 1. The size of 4 × 2.2-wavelength 144-MHz Yagis dwarfs the 4 × 6.1-wavelength 432-MHz array. Performance of both arrays is about the same on their respective bands—despite the size difference.

note that this 2:1 spacing differential (4:1 array size difference) between 144-MHz and 432-MHz Yagis of the same physical boom length, holds over a fairly wide range of boom lengths. The compact size of the 432-MHz array opens the EME door to a raft of operators who never thought they could manage an EME array. In fact, this EME array is easier to handle (and smaller) than most HF triband beams!

Recommended Yagis

Nowadays, there's a large selection of good Yagi designs. Your major decision is to choose the size you wish. I prefer smaller Yagis; they will give a much better wind load and weight-to-gain ratio than long Yagis. Long Yagis can be used to give some pretty impressive results with just four Yagis, but at the expense of the high mechanical stress long beams create due to their torque loading on the frame, mount, and tower. Long Yagis must also be mounted a fair distance above any guy wires when the array is elevated. Conversely, if at a later date you decide to expand your small Yagi array, you'll be faced with the time and expense of building more power dividers and phasing lines, plus managing that phasing-line loss in order to maintain top array performance. **Table 1** lists some recommended Yagis for 432-MHz EME, along with the expected array gain for four Yagis. The boom length and proper stacking distances are included. I've also listed Yagi dimensions. Although there are other suitable Yagi designs, I've included only models with dimensions that have been covered in the amateur literature or can be purchased from a commercial manufacturer.

My objective here is to describe a small,

Designs for home construction

Yagi	Gain dBi	Length	Stacking E x H "	Array Temp.	Where Documented
FO-22	17.9	14'	66 x 62	26K	ARRL Handbook 1990 or later
FO-24	18.5	17'	70 x 66	25K	Ham Radio, July 1987
FO-32, MK 4	19.9	24'	82 x 78	24K	Ham Radio, May 1989
FO-33	19.9	24'	83 x 81	25K	ARRL Handbook 1990 or later
FO-36	20.4	28'	87 x 85	24K	Ham Radio, May 1989
DJ9BV Yagis	lengths from 17' to 30'			26K	DUBUS No. 2, 1991
DL6WU Yagis	lengths from 5' to >30'			27K	ARRL UHF/Microwave Manual

Suitable Commercial Models

Yagi	Gain dBi	Length	Stacking E x H "	Array Temp.	Notes
Rutland FO-22	17.9	14'	66 x 62	26K	Rutland Arrays
Rutland FO-25	18.6	17'	71 x 68	26K	Rutland Arrays
KLM-432-30LBX	19.2	20'	78 x 74	35K	Single DE mod. recommended
Hy-Gain 70-31DX	19.7	24'	80 x 77	32K	Temp raised due to feed losses
Rutland FO-33	19.9	24'	83 x 81	25K	Rutland Arrays
M^2 432-13WL	20.4	33'	87 x 85	33K	

Notes:

Array temperature is the effective noise pickup of a 4 Yagi array pointed at at cold sky. Phasing line losses are not included. Figures are based on calculations by DJ9BV.

The KLM 432-30LBX uses a 3 element log feed. Unwanted feed radiation raises its temperature. When converted to a dipole driven element its temperature would be about 27K.

The Hy-Gain 70-31DX uses an RG-303 jumper from the driven element to feedline connector. By redesigning the feed without the jumper the array temperature be reduced to about 27K.

Both the KLM-432-30LBX and the Hy-Gain 70-31DX are based upon the DL6WU design.

Table 1. Recommended 432 MHz Yagi Designs for EME

simple 432-MHz EME array which virtually anyone can put up and use successfully. I've chosen the FO-22 Yagi* as the array building block. A 4 × FO-22 Yagi array is small and light enough that commercial elevation and azimuth rotators can be used with the array; but, its performance is high enough that you'll be able work a large number of different EME stations easily and make random EME QSOs regularly. The FO-22 is also a great starting point for a larger array should you later decide to expand your array to 6, 8, 12, or 16 Yagis.

Many hams have had good luck working four Yagis on EME. Bob, N0IS, put up an array of 4 × 23-element Yagis (FO-22s with an additional director) and worked over 20 different EME stations in his first two months of operation. What's even more impressive is that Bob's contacts have been made without the benefit of any schedules. Another good example of the ability of the 4 × FO-22 element Yagis to perform on EME was demonstrated by Per, VS6BI. In only two months of operation with an array of 4 × FO-22s mounted on top of an apartment building, Per managed QSOs with 32 different stations. If one devotes a little effort and time, this array is capable of making contact with more than 100 different EME stations.

Array construction

There is a major problem inherent in EME arrays. When all the components of the array are put together (the Yagis, stacking frame, phasing lines, power dividers, preamplifier/relay box, and so on), they quickly exceed the weight and wind load capability of inexpensive amateur commercial rotors and small towers. The FO-22 Yagi has a wind load of 0.8 square feet, which is a small load. However, if you increase the number of Yagis to four, the array surface area increases to 3.2 square feet without the rest of the components. As I implied earlier, using more small Yagis will usually add weight and wind load at a slower rate than will arrays with large Yagis. In the sample array, when all the pieces are factored in (stacking frame, elevation rotor, phasing lines, power divider, and preamplifier box) the total array is approximately 7 square feet, or almost ten times the wind load of a single Yagi! However, this isn't a fearsome amount of wind load. It's about the same as what I calculate for a typical full-size three-element HF triband beam.

Let's take another look at this size escalation problem. Suppose you want to double the array's gain (add 3 dB). You could come close to this figure by going to 8 × 22-element Yagis. However, this array would occupy three times the physical area as the four-Yagi array. It would also have a wind-load area of about 18 square feet—over twice that of the smaller array. Due to stacking losses and increased phasing line losses, its gain would most likely be only 2.7 or 2.8 dB greater than the four-Yagi array.

Another approach is to use four long Yagis. Due to the nature of the Yagi characteristics, you would obtain around 2.5 to 2.6 dB of additional gain if you made a 28-foot long Yagi—double the boom length of the 22-element Yagi. To obtain 3 dB higher gain, you'd need a Yagi at least 33 feet long! The wind load of an individual Yagi of this length, made rough enough to survive the rigors of a northern climate, may have over 4 square feet of wind-load area per Yagi and a total array wind load exceeding 22 square feet. This very large array would have to be mounted at least 12 feet above any guy wires to allow for elevation clearance; thus, it would require a much stronger tower.

With these numbers in mind, reflect back on the original purpose of this article: to describe a manageable array that will make you successful on 432-MHz EME, before you worry about becoming a big gun. The 4 × FO-22 array's 7 square feet of wind load is also distributed over a relatively small area, so simple rotors can handle the load. This is not the case when using long-boom Yagi arrays.

The stacking frame for the array is straightforward. The main frame uses 1-1/2 O.D. × 0.125-inch wall 6061-T651 extruded aluminum tubing.** The main horizontal piece is 6 feet long and runs through the KR-500 elevation rotor. The vertical pieces are 5 feet, 6 inches long and are held to the cross member by standard 1½ inch TV U bolts and ¼-inch aluminum plates. The plates are 4 × 5 inches, but the actual size isn't critical. The 0.125-inch wall tubing is more rugged and heavier than necessary. I chose it because the tubing is made of extruded material, which is usually less expensive than the thinner wall drawn sizes. It also lets you tighten up the antenna and frame U bolts enough to hold the array in place, without collapsing the tubing. If you are careful about how you align the frame and tighten the U bolts, any tubing with a wall thickness of 0.058 inch or

*Rutland Arrays, Tom Rutland, K3IPW, 1703 Warren Street, New Cumberland, Pennsylvania, (717)774-5298, 7 to 10 p.m. Eastern Standard Time. Rutland Arrays can also supply stacking frames and a wide assortment of antenna parts for those who wish to build their own antennas.

**The Dillsburg Aeroplane Works, 114 Sawmill Road, Dillsburg, Pennsylvania 17019, (717)432-4589, 8:00 a.m. to 5:30 p.m. weekdays, 8:00 a.m. to 1:00 p.m. Saturday. Dillsburg can supply a large assortment of aluminum tubing, bar and plate, in small or large quantities. They also carry high-strength steel alloy tubing and plate.

larger will be adequate. You can also use TV mast or try 1¼ inch schedule 40 6063-T651 aluminum pipe (1.6-inch O.D. × 0.141 wall)—the largest size which will fit through the KR-500 elevation rotor.

Two pieces of 1.25-inch O.D. × 0.058-inch wall × 48-inch long aluminum tubes hold the power divider and preamplifier/relay box behind the array (see **Photo A**). Good array balance is important if you're using the KR-500 elevation rotor. I selected the mast position of the FO-22 Yagis so the array was balanced with the phasing lines in place. If your array doesn't balance, add counterweights. The power divider and preamplifier are mounted at the back of the array in order to minimize the phasing-line losses. At 432-MHz EME, any phasing-line losses can act as if they are three or four times as great on receive due to the low sky temperatures.

My phasing lines are 42-inch lengths of Times FM-8. This is a foam-type RG-8 cable with a solid center conductor. The total line loss with new cable is 0.13 dB. Many VHF operators use Belden 9913 or similar cables. Be very careful when using 9913 or similar cables from other manufacturers. Because of the air volume inside the cable, it can very easily draw water inside when the temperature or barometric pressure changes significantly over a short period of time. Make sure the connectors and the cable-to-connector joints are sealed carefully prevent water seepage. If you do use foam-type RG-8 or 9913-type cables, you must think of these braided shield coaxial cables as expendable parts. Sooner or later, they will deteriorate due to moisture migration and plasticizer contamination (even in noncontaminating jacket cables). Plan on replacing such cables every 2 or 3 years for optimum performance.

You can use RG-213/U or RG-214/U, but you'll experience an insertion loss of 0.18 dB for the 42-inch phasing lines. This will degrade receive performance by 0.2 dB over the new FM-8. The solid polyethylene dielectric of RG-213 and RG-214 is much less sensitive to moisture than foam dielectric cables and will usually last 10 years or more.

If you can find it, aluminum jacketed hardline, like RG-331/U (also called Alumifoam® or Foamflex®), will lower losses further. The best coaxial cable to use for the phasing lines is Andrew LDF4-50A (or a similar copper jacket cable like Cablewave Wellflex® FLC). It will have even lower loss (0.05 dB for a 42-inch piece at 432 MHz). Believe it or not, the 0.08 dB lower loss of LDF4-50A cable will sound like a 0.35 dB difference on EME receive! So be sure to keep the phasing line losses as low as possible; they make a huge difference. Both Alumifoam and Heliax® -type cables will last much longer than braided shield cables. Their solid metal shields are effective vapor barriers, so the useful life for many of these

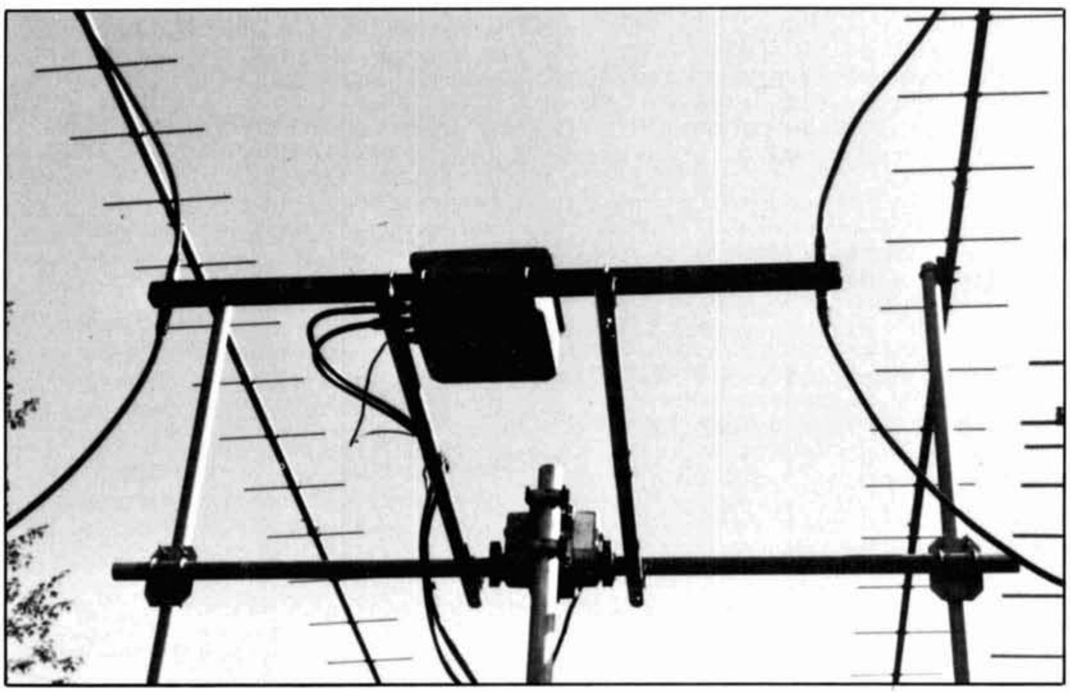

Photo A. The power divider is supported about 4 feet behind the array center to let very short phasing lines be used to reduce phasing-line losses. The preamplifier/relay box is mounted off center on the 3/2-wavelength power divider to allow for better use of the space inside the box.

cables is over 20 years. All-copper cables have the longest useful life of any of the different cable types. In our current polluted environment, rain water is quite acidic because of its sulfuric acid content. Sooner or later, moisture will migrate into the connectors. The results will be much more devastating in cables made with dissimilar metals (i.e., aluminum jacket and copper center conductor) than with all-copper cables.

Much has been said about selecting the length of the phasing lines. Some argue in favor of half-wave multiple phasing lines; others feel that odd quarter-wavelength multiples are best. Theoretically, odd quarter-wavelength cables will minimize currents on the outside of the phasing line shields. This, in turn, will lower the array's temperature and improve receive performance. But if the individual Yagis aren't well matched, the impedance transforming characteristics of odd quarter-wavelength multiple transmission lines can make any mismatch worse. On the other hand, half-wavelength multiple lines should mirror the individual Yagi's feed impedance and reduce SWR problems. I've found that if the driven elements of the individual Yagis are matched and balanced, there will be little difference between half and odd quarter-wavelength multiples. The most important factor is simply to make sure all phasing lines are the same electrical length. I recommend that you measure cables electrically using a return-loss bridge. If you can only measure the cable physically, you may be better off using either a solid dielectric cable like RG-213/U or a high-quality solid-jacket cable like Andrew LDF4-50A, which will have a very consistent velocity factor if the lines are cut from the same cable run.

The power divider used on the array is a 3/2-wavelength unit, which is a ½-wavelength 144-MHz power divider made by WA6MGZ.* The longer unit is used to both reduce phasing-line loss and to allow room for mounting the power divider to the support tubes by getting the coaxial connectors out of the way. The large size (1¼ inch) air dielectric coaxial line which makes up the power divider is very low loss. These types of power dividers can have the same dissimilar metal problems as aluminum jacketed cables should moisture get into them. If you're using an aluminum outer conductor power divider, I recommend that you silver plate the center conductor or coat the entire inside of the power divider (after assembly) with a good clear finish like Rustoleum® Clear Coat. Do this by removing the power divider end caps and pouring the clear enamel through.

Some EME operators have eliminated the connectors and cut their costs by making power dividers out of copper tubing. The copper-shielded Heliax phasing lines are then soldered directly to the power divider.

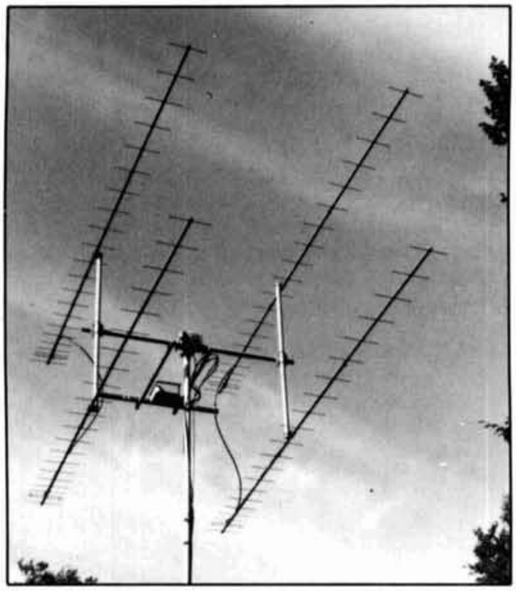

Photo B. The completed 4 × 22-element Yagi 432-MHz array is set up on a mast for portable operation. Because of its small size, it can be handled easily by one person. The array will fit into a small yard or on a roof top, letting those who never thought they could manage EME join in the fun.

The completed 4 × 22-element Yagi array is shown in **Photo B**. The array has been set up so the four Yagis are arranged as mirror images. I used this arrangement because it doesn't necessitate running some of the phasing lines across the boom. Whether you use a mirror-image arrangement or not, it's important to have the array phased correctly. The center conductor from the feed cable must run to the same side of the driven element on all four Yagis (see **Figures 2A** and **2B**).

Azimuth and elevation rotors

For simplicity, I used commercial azimuth and elevation rotors. The Yaesu G-500A (formerly the KenPro KR-500) has been used successfully by a number of four-Yagi EME operators. If the array is carefully balanced, generally good results are possible with four Yagis up to 20 feet long. Several operators have tried using the G-500A with Yagis longer than 24 feet, but have usually experienced repeated rotor failure. The G-500A should work fine with the 4 × 14-foot Yagi array

*Z Comm Custom Engineering, Chuck Smallhouse, WA6MGZ, 803 Mora Drive, Los Altos, California 94022, (415)948-1778, 5:00 to 10:00 p.m. Pacific Time. Z Comm will also supply electrically matched phasing lines.

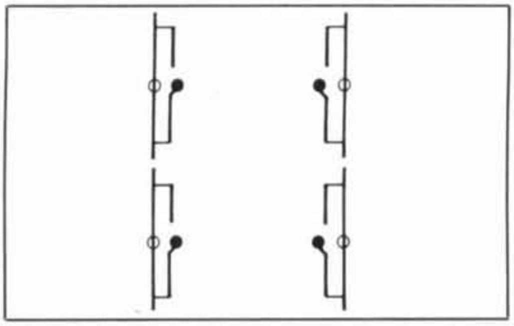

Figure 2A. If the Yagis are mirror imaged, the feedline center conductor connects to the same side of the driven element in the array, thus requiring mirror-image driven-element construction.

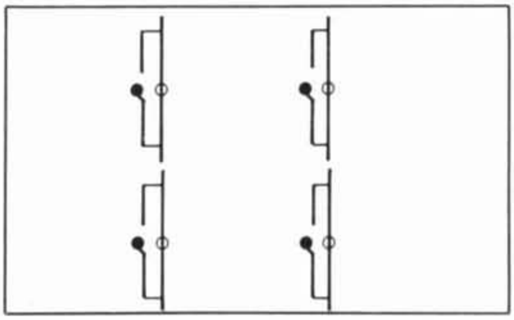

Figure 2B. If the Yagis are not mirror imaged, all driven elements are constructed in the same way, and the feedline center conductor still connects to the same side of the driven element when they are positioned in the array.

shown here. EME operation only requires 0 to 90-degree elevation motion. The KR-500 has the ability to rotate over 180 degrees. To improve the indicator accuracy, I recalibrated my KR-500 control box (see **Photo C**) in 1-degree increments covering 0 to 72 degrees elevation because, at my latitude, the moon never gets above 76 degrees elevation. This recalibrated scale is a lot more convenient than the original 0 to 180-degree scale with its 2.5-degree increments. You could also calibrate the scale in 2-degree increments. This would allow full elevation readout to 90 degrees, plus leave scale room below the horizon. It's nice to be able to point the array into the ground. This lets you measure Earth noise, which is very helpful when verifying your system receive performance.

Larger arrays require a more substantial elevation actuator than the KR-500. Some operators have ganged multiple KR-500s together in order to increase the array load capacity. Others have used the KR-500 (or the Alliance U-100 type rotors) to drive elevation motion through chain or gear reduction drives. TVRO-type linear actuators are popular, but will require the fabrication of some type of hinge mount.

The Hy-Gain CD-45 or HAM-IV are suitable azimuth rotors. The Hy-Gain T2 X rotor could present a problem because it has 6-degree braking increments, and the -3 dB beam width of the array will be around 10 degrees while the -1 dB beam width is only 5 degrees. These type rotors have a bad reputation for indicator accuracy—one of the prime reasons for failure on EME. Even an array this small will require both azimuth and elevation aiming to ± 2.5 degrees for best results. A very easy way to confirm your aim is to use Sun noise. There are several Sun-tracking computer programs available. You simply peak your array for maximum Sun noise by pointing at the array at the Sun and comparing your indicator's heading to the real Sun azimuth and elevation. Compare headings over several directions to make sure of the calibration at different points. It's possible to make optical sightings, but they won't be correct for pattern skewing (should your individual Yagis not be perfectly aligned) as will calibrations made to the Sun's position.

There are other rotors which are also suitable for this array. The Diawa rotor is very popular in Europe and Japan. It has a very good indicator, but the unit is hard to obtain in the United States. The Hy-Gain HDR-300 is popular with some EME operators with large arrays because of its accurate readout and fine resolution brake. However, the HDR-300 is expensive and somewhat of an overkill for this small array. The Alliance HD-73 was recently made available again at a very reasonable price. The HD-73 is popu-

Photo C. The KR-500 control box is shown with the meter scale recalibrated in 1-degree increments. This allows easy and precise elevation readout of the array. An angle readout dial is used to set the KR-500's calibration pot to the new scale.

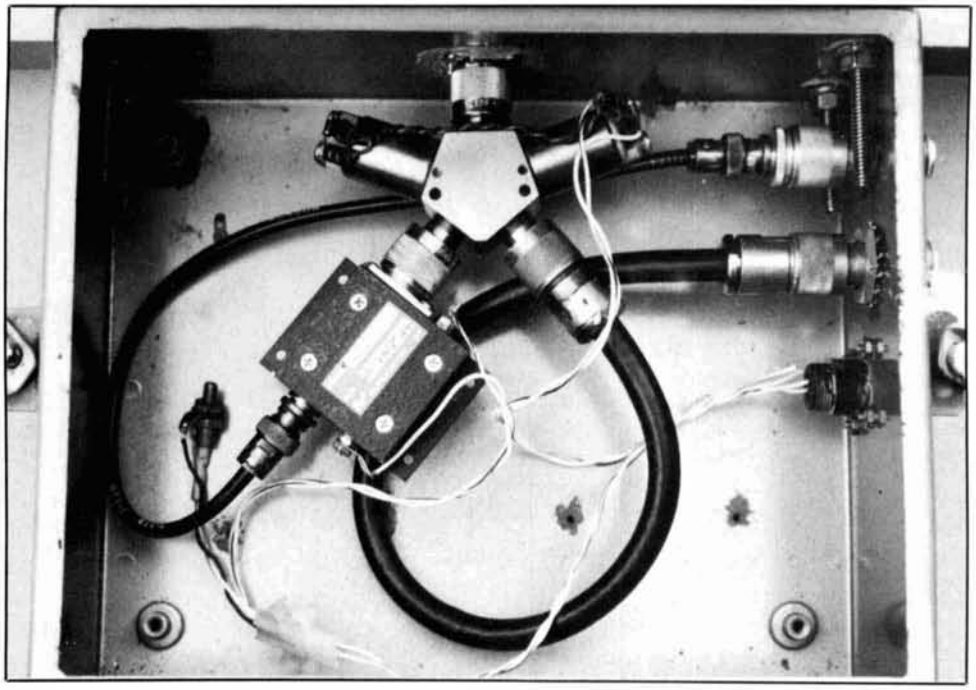

Photo D. The inside of the preamplifier/relay box shows the attention paid to minimize connector and feedline losses. The surplus Transco Y relay mounts directly on the mail N connector of the power divider and another male connector is used on the modified AAR GaAsFET preamplifier to eliminate all jumpers and adaptors from the receive line.

lar with some VHF tropo operators; they like its dual-speed rotation control. Its slow speed is handy for peaking signals or making sure your array is aimed correctly.

Relays

Photo D shows a preamplifier/relay box. It's constructed from a standard steel 8 × 10 × 4-inch oil-tight JIC-type junction box. Vent holes are drilled in the bottom of the box to prevent moisture from building up inside. Since this photo was taken, I've built a new preamplifier/relay enclosure using a plastic junction box (shown in **Photos A** and **B**). The PVC box is a better choice; it weighs less and has less of a tendency to hold moisture. It's also easier to cut connector holes out of the plastic. These boxes are available from local electrical supply houses. If your box isn't light in color, I recommend that you paint it white so it will reflect the sunlight better and reduce heat buildup during the day. Temperature elevation will increase the noise figure of a preamplifier during the day. Greater temperature differentials will also make any condensation problems worse. Many operators put desiccant (like bags of silica gel) in the preamplifier box to protect the preamplifier and relays from moisture.

To reduce losses and eliminate additional coaxial connectors (always a potential problem), mount the power divider directly to the preamplifier/relay so the input connector to the power divider protrudes into the box. A male N chassis mount connector (Pamona model 2454) is mounted on the power divider. This allows the surplus Transco Y relay (model 11300) to connect directly to the power divider without any adaptors. The preamplifier is a modified AAR (model P32VDG) which also has a male N chassis mount connector on its input to eliminate any adaptors, and their associated losses and potential problems.

I chose the Transco Y-type relay because they're available through surplus channels for a reasonable price and their physical configuration allows the connections to the power divider and preamplifier to be made without extra jumpers and adaptors. Be careful when using surplus relays like the Transco Y type. The relay contact's plating can oxidize, causing poor connection on receive. Check the relay if in doubt. Insertion loss for a good Y-type relay at 432 MHz is about 0.035 dB. Most of the surplus relays will have 28 volts DC coils. I prefer DC coils because they switch faster and run cooler than AC types. The 28-volt rating may be a minor inconvenience to some operators because a separate power supply is required. However, having a separate power supply is an advantage as it will prevent relay switching spikes from harming other equipment. Note that the relay coils have diodes across

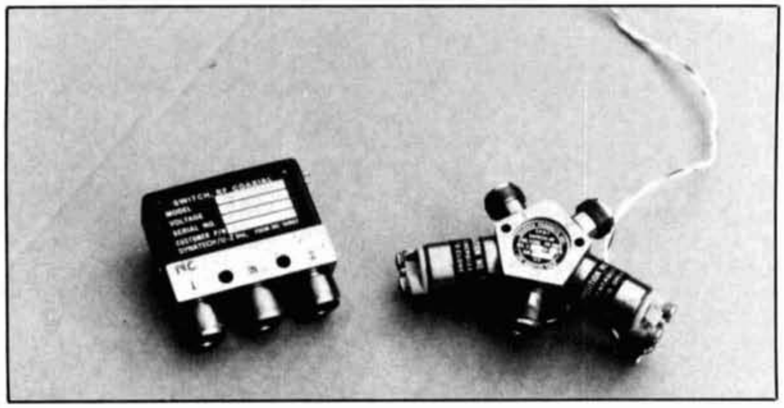

Photo E. Two common types of low-loss, low-SWR microwave relays. A Dynatech D4-P relay, which is a special high-power (1500 watts at 500 MHz) model is shown on the left. A standard Transco Y-type relay, which can handle over 1000 watts at 432 MHz in intermittent amateur keyed CW and SSB service, is on the right.

them to minimize this spike problem.

There are many other suitable relays. The Transco D-type relay (part no. 810C00100) is also a good unit. The D-type relay is made in a rectangular box about 2.5 × 2 × 1-inch, and all three connectors exit at one end. These relays have very high isolation (typically over 80 dB at 432 MHz) and an insertion loss of around 0.05 dB at 432 MHz. The one disadvantage of the D-type relay is that the connectors are close together and all on the same side. This arrangement requires some additional coaxial jumpers or adaptors in order to facilitate connections to the power divider, preamplifier, and transmitter feed line. Relays similar to the Transco D type are made by many other manufacturers like Microwave Devices, Joslyn Defense Systems (formerly made by Amphenol), Dynatech, and Dow Key. **Photo E** shows a Dynatech type D4 relay. Some of these relays are available in both failsafe (energize to switch and keep energized to hold in) and latching models (pulse to switch each way). I prefer the failsafe models used in the energize-to-receive mode. They offer additional preamplifier protection as the array is disconnected when the system is not in use. You will also be able to operate in a "wounded" mode should the antenna preamplifier fail, by substituting a relay and preamplifier in the shack. These relays often show up at amateur radio flea markets and in surplus stores.*

A relay similar to the D type is make in Germany by EME Electronics (model HP 2000/6). EME Electronics also makes a model HF-400, which is very popular with the European EME operators. The HF-400 has lower isolation (about 55 to 60 dB at 432 MHz), but has very low insertion loss (less than 0.2 dB at 432 MHz). This means it can handle 1500 watts with ease. The EME Electronics relays have standard 12-volt coils. Unfortunately, I don't know of a United States distributor for the company.**

I like to use a two transmission-line arrangement for connection to the preamplifier. Because I use a transverter to generate my 432-MHz RF, I can avoid using a second relay in the system. This method also works well if you're using a 432-MHz transceiver. With a two transmission-line system, you use a low-power relay at the output of the transceiver and bypass the high-power amplifier. Alternatively, the transceiver could be modified to eliminate its RF switching and provide separate 432-MHz transmitter output and receiver input. This method eliminates the need for the second relay—just like my transverter arrangement. **Figure 3** shows connection diagrams for the two transmis-

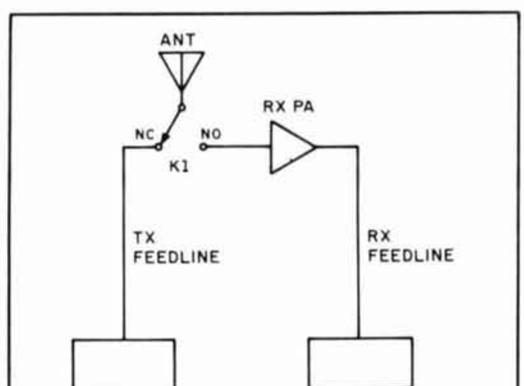

Figure 3. The two-feedline system offers the elegant simplicity of one relay.

*Tucker Surplus Store, 1717 Reserve Street, Garland, Texas 75042, (800) 527-4642, ext. 135 (Hazen Smith), is one of many surplus stores which often have relays and test equipment. Check *QST* Ham-Ads and the Ham Trader Yellow Sheets, too.

**EME Eliktronik, Karl Muller, Benediktstr. 6. 8021 Hohenschaftlarm FRB.

sion-line preamplifier connection.

Some operators prefer to use only one transmission line. This requires the use of a second high-power relay or a high-power double-pole relay (see **Figure 4**). Another high-power relay is required in the shack to bypass the power amplifier. This type of preamplifier connection must have proper relay sequencing to avoid both front and rear-end burnout.

Much has been written about using a second relay between the first relay and the preamplifier to provide additional isolation. The theory is that the second relay will reduce the amount of transmit power that leaks through to the preamplifier. This, in turn, reduces the chances of preamplifier burnout. It used to be common to put a 50-ohm load on the unused port of the isolation to terminate the preamplifier during transmit and prevent the preamplifier from oscillating. In fact, a relay with 60 dB of isolation (the Transco Y or EME HF-400) will present only 1.5 mW of power to the preamplifier if there's 1500 watts of transmitter power at the relay. Modern GaAsFET transistors typically can handle 60 times that power level before they are destroyed. As far as preamplifier oscillation during transmit is concerned, if the DC power to the preamplifier is properly current limited through the use of a series resistor with the power supply, the GaAsFET won't suffer any damage—even if it does oscillate on transmit. A final argument for the two-relay system is that the 50-ohm termination can be used as a reference level when making noise measurements. This doesn't always work out well either. Since one's array is never quite 50 ohms, and is also a narrow-band load, the gain of the preamplifier will change when it's switched between the load and the array. On 432 MHz, pointing the array at cold sky will make a much better noise reference. This is not only because errors due to gain variation will be avoided, but also because the low temperature of a cold sky (versus a 50-ohm resistor at ambient temperature) will give a much greater Y factor (difference in noise output) between cold sky and noise sources. This greater Y factor will give better measurement accuracy of noise sources which, in turn, allows the operator to better evaluate his system's performance.

The EME system is less complicated if you use one good relay, because it has one less relay to create problems. Also, without the additional losses from the second relay and its related connectors, the system's EME receive performance will be improved. It will also be less costly to build.

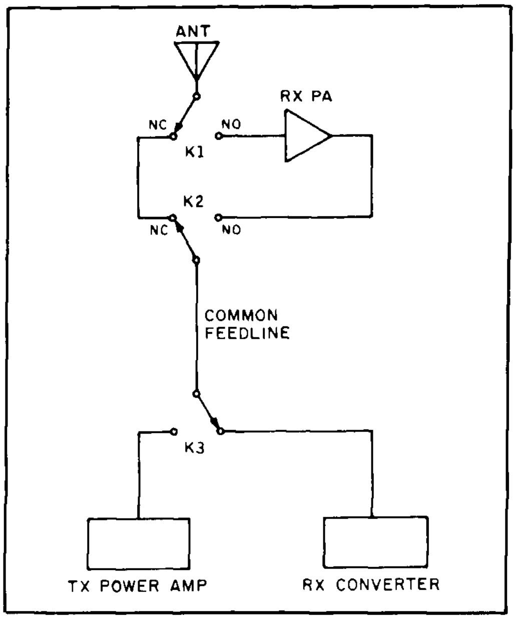

Figure 4. The single-feedline approach requires at least three high-power SPST relays (or one DPDT and one SPST relay).

Electronics

You may find this surprising, but almost all of the latest generation of multimode transceivers are quite EME capable when combined with a good array-mounted preamplifier. Successful 432-MHz EME operators have used the ICOM IC-471A/H and IC-475A/H, the Kenwood TS-790 and TS-811, and the Yaesu FT-726R and FT-736R. I personally prefer to use an HF receiver or transceiver with a good transverter. This type of system will normally have better dynamic range than a multimode transceiver and more selective IF filtering. I'm biased, in part, because of my location in a high-RF density location (line-of-sight from Long Island and only 70 miles from New York City). I also have a preference for very narrow CW IF filters, and normally use an old Drake TR-7A with a good Sherwood 200 Hz IF filter. I recently had the opportunity to try out a Kenwood TS-790A on EME. Once I became accustomed to the sound and tuning of the TS-790A, I found there wasn't much of a difference in receive performance when I used the TS-790A with its 500-Hz CW filter and my HF receiver/transverter arrangement. Many other operators seem to do as well as I do with SSB bandwidth (2.3 kHz) filters and UHF multimode transceivers, or with receivers using 500-Hz CW filters. Always remember that an expensive transceiver with all the "bells and whistles" is not a requirement for EME operation. On his many successful EME DXpedition operations, KL7WE used a simple ICOM IC-402 portable 432-MHz transceiver with a good audio filter. Other EME top guns, like DL9KR and NC1I, use old Drake 4-line tube equipment with transverters. In some cases, it appears that equipment with simple signal paths in the receiver works best.

Those who wish to use their HF equipment with a transverter will find that the Microwave Modules MMT-432S has been the standard 432-MHz transverter worldwide for many years. Although the design is now somewhat dated, they are widely available on the used market at reasonable prices. The MMT-432 has marginal front-end selectivity and limited image rejection. If you live in a populated area, you may need a good filter between the preamplifier and MMT-432 (cavities, striplines, and interdigital filters are suitable). SSB Electronics also makes VHF and UHF transverters. The LT-70S is a complete unit (except for the power supply). The TV28-432 and accessories is a modular system for those who prefer to customize their installations.* Some EME enthusiasts even use 144-MHz multimode transceivers with a transverter which has a 144-MHz IF frequency.

The most important part of EME reception is to try different filter combinations and AGC settings, and to listen to the CW signals at different pitches, until you determine where and how you hear best. The amount of total system gain is also very important. If the preamplifier gain is too low, you'll hear IF and audio noise which may mask the EME signals. Receive performance may also be impaired if there's not sufficient preamplifier gain to override the receiver's RF and IF noise.

If the system gain is too high (usually indicated when the receiver S meter reads up the scale when no signals are present), there will probably be too much AGC action. This will bury the weak EME signals in the background noise. I prefer to adjust the gain so the background noise is just below the point at which the S meter starts to move up the scale. The combination usually masks IF and audio noise and gives some AGC action, but not enough AGC action to mask the weak signals. I prefer a fast AGC recovery, but many operators switch their receiver's AGC off for EME operation. Believe it or not, one of the reasons why I prefer to have the AGC on is that some of the stations are so strong that, if you're wearing headphones, they'll hurt your ears when you tune across them with the AGC off. The amount of system gain is often best adjusted by using an attenuator in the IF line for transverter setups or between the last preamplifier and the UHF transverter. You'll find a good step attenuator is invaluable both for adjusting gain and making noise source measurements.

Everyone's ears are different, but most EME operators (and good HF CW operators) prefer a CW note around 500 Hz or lower. Some prefer notes as low as 300 Hz. For some reason, most current HF and VHF/UHF transceivers are set up for a standard CW pitch of 800 Hz. Receivers or transceivers with some form of passband tuning, IF shift, or adjustable CW offset frequency that allows you to move the filter response to your preferred CW note, are the most desirable models. Some operators use audio filters. I haven't found any audio filter that makes a spectacular difference in EME copy. I do normally use an audio filter on EME to eliminate wideband noise, hum, and rumble. I feel that the audio filter lets me operate for longer periods of time without fatigue by getting rid of unwanted noise—especially high-frequency hiss from the audio stages. I use a relatively simple analog CW peaking filter (using operational amplifiers), while many other EME operators speak highly of the SCAF (switched capacitance) or "digital" filters like the Super Scaf or the Palomar models.

Simply stated, almost any 432-MHz multimode transceiver made in the last seven years, and almost any HF transceiver or receiver made in the last 15 years, is probably good enough for EME—as long as you have good preamplifiers (and a converter, if required). Save your dollars for preamplifiers, power amplifiers, and antennas!

Power amplifiers

EME contacts on 432 MHz have been made by stations with relatively small arrays and as little as 100 watts. WB0GGM has made about a dozen EME QSOs by mounting his 100-watt "brick" at the power divider of his 4 × 17-foot long Yagi array. In keeping with the theme of allowing a margin for error, as well as having regular random QSOs, the minimum recommended power at the array is 500 watts. Unfortunately, a wide variety of amplifiers above the 100-watt solid-state brick units don't exist for 432 MHz. For years, Fred Murray, W2GN, made copies of the K2RIW parallel 4CX250 class tube amplifiers. With a good pair of 8930s or 4CX250Rs, these amplifiers will deliver over 1000 watts of power when run class C. Although the amplifiers haven't been made for

*SSB Electronics USA, K3MKZ, 124 Cherrywood Drive, Mountaintop, Pennsylvania 18707, (717)868-5643, weekdays after 6:30 Eastern Standard Time, or on weekends.

several years (W2GN passed away several years ago), second-hand Arcos amplifiers (as well as homebrew RIW amplifier copies) are frequently available. Keep your eyes open at flea markets and read the classified ads in the ham publications.

There are presently two readily available high-power commercial 432-MHz amplifiers. The Henry Radio 2004-A uses a single 3CX800A7 triode, and has a linear RF output of over 700 watts. Russ Miller, N7ART, is manufacturing an amplifier which uses two 3CX800A7 triodes and will make the 1500-watt power output limit.*

I suggest that those who wish to build their own amplifiers read my article in QEX on power tubes at 432 MHz.[1] The article's bibliography contains references to several 432-MHz amplifier construction projects.

Preamplifiers

Part 1 described how the improvements in receiver technology which occurred in the 1980s were responsible for moving EME operation into the hands of operators with small arrays. Success with a four-Yagi array requires excellent receive performance. Again, keep in mind that every tenth of a dB in system capability will result in a 2 to 4 times improvement in receive signal-to-noise ratio. The target to shoot for is a receive system noise temperature of less than 35 degrees K (less than 0.5 dB noise figure). Few commercial preamplifiers truly meet this requirement. Performance is affected by more than the preamplifier. The following stages can add to the system noise temperature, if the first stage doesn't have high enough gain to overcome the following stages (especially if you use a long feed line run). Simply adding a second stage can affect performance, as amateur preamplifiers are not typically impedance matched or unconditionally stable. This can cause preamplifier oscillation when one stage doesn't like the next one it's connected to.

The GaAsFET preamplifier has been the standard for VHF and UHF weak signal work for several years. There are two basic types of GaAsFET devices. The dual-gate GaAsFET is very popular because it's inexpensive. It can provide a high amount of gain and still be very stable. Unfortunately, dual-gate noise figures are usually over 1.0 dB. The commonly available RF-switched preamplifiers use lossy low-power relays which limit transmit power capability, as well as reduce receive performance. Consequently, most commercial amateur-type switched preamplifiers have noise temperatures in the 120 to 170 degrees K range (noise figures in the area of 1.5 to 2.0 dB). This is not EME capable performance!

The second type of GaAsFET is the single-gate device. This device is the choice for weak signal work. Several commercial preamplifiers are available which will get you under the desired 35 degree K temperature mark. The most popular devices are the Mitsubishi MGF-1302, MGF-1402, and MGF-1412. I feel that the '1412 is the best device, as individual units are very consistent in their electrical parameters. The '1402 is almost as good, but some lots of devices are better than others. While the '1302 will give acceptable performance, it is, on average, measurably inferior to the other FETs. (The lot-to-lot variations in the '1302 are even greater than the '1402—as may be expected because of its lower price.) "Hot" MGF1302s do exist, as evidenced in some of the noise figure contests. Often these extra low noise units are selected devices, not random production units. Other suitable devices are made by Avantek (ATF-10136) and NEC (NE21889).

Preamplifier input circuits can be grouped into three basic categories: coil, stripline, and cavity. Coil circuits can be made with noise temperatures under 35 degrees K (0.5 dB nf). Stripline circuits are often under 29 degrees K (0.4 dB nf). The lowest measured 432-MHz noise temperatures have been from preamplifiers using very low-loss cavity input circuits with noise temperatures as low as 21 degrees K (0.3 dB nf). The Advanced Receiver Research P432VDG uses a traditional coil input circuit. If you remove the GaAsFET gate lead, input coil, and input capacitor connections from their printed circuit board pads and connect them together by suspending them in air, the AAR preamplifier can obtain a noise temperature of about 35 degrees K (0.5 dB noise figure). The AAR preamplifier is small, stable, and moderate in cost.** Although its noise temperature isn't the lowest available, several stations have reported success with the AAR preamplifier in cases where more sophisticated units haven't worked properly. A commercial stripline preamplifier is made by Microwave Components of Michigan. The 432-MHz model typically has about a 29 degrees K noise temperature (0.4 dB noise figure)† and is used by several EME stations.

The cavity preamplifier has been popularized by WA7CJO.[2] As mentioned before, it

*Russ Miller, N7ART, 12041 SW Penninsula Drive, Crooked River Ranch, Oregon 97760. A construction project writeup on the amplifier appeared in "VHF/UHF and Above, March 1988.

**Advanced Receiver Research, Box 1242, Burlington, Connecticut 06013, (203)582-9409.

† Microwave Components of Michigan, P.O. Box 1697, Taylor, Michigan, 48180, (313) 753-4581 (Norman Alred, WA8EUU).

offers the lowest available noise temperature of 21 degrees K (0.3 dB noise figure). Unfortunately, the machining work required to manufacture such a preamplifier has prevented its sale as a commercial item because the price of such an amplifier would be several hundred dollars.

Some of the modern converters and transverters have noise figures under 2 dB and may only need one preamplifier stage for good performance. Most of the 432-MHz multimode transceivers use pin diode switching and fairly hefty front-end filtering, which results in an effective noise figure of over 5 dB. This high a noise figure, combined with transmission line loss, will demand two stages of preamplification to achieve adequate system noise temperature. Remember that the overall system noise temperature is dependent not only upon the first stage noise figure, but its gain, and the noise figure and gain of the succeeding stages.

If a second-stage preamplifier is required, I recommend that you use a device which has impedance-matched input and output circuits (input and output impedances are close to 50 ohms). As I mentioned before, cascading noise optimized preamplifiers typically have very poor input SWRs (when tuned for best noise performance their input impedance will not be 50 ohms). Also, these units are often only marginally stable under this condition. I recommend you use the impedance-matched unit designed by N6CA as a second-stage preamplifier.[3] The N6CA preamplifier has also been successfully used as a first stage by stations experiencing oscillation problems with other preamplifier types. The impedance-matched type preamplifier is also very useful where extreme interference problems exist. A matched preamplifier will work well after a low-loss high-Q cavity filter, while unmatched preamplifiers will often exhibit problems when connected to filters. The one disadvantage of the N6CA preamplifier is that its noise temperature is about 35 degrees K (0.5 dB noise figure) which, due to its input circuit losses, is somewhat higher than can be obtained without high-Q circuits.

In Europe, a two-stage preamplifier design by DL9KR,[4] which uses a high-Q single-gate input stage and a high dynamic range dual-gate second stage, has become popular. The stages are tuned together using a two-stage preamplifier, solving the problem of cascading unrelated stages. The preamplifier is also designed to have enough selectivity to be able to survive reasonably high RF interference locations.

Protection is important if you want to avoid the inconvenience and cost of blown preamplifiers. RF burnout is rarely a problem if you use proper relay sequencing. Proper sequencing occurs when the RF relays actuate before a transmit signal can be generated. This is similar to the sequencing used in QSK systems. A sequencer is described in *The ARRL Handbook*.[5] A commercial sequencer is available from Advanced Receiver Research. Array-mounted preamplifiers are most often destroyed by static electricity pickup in the DC power lines during electrical storms. I strongly recommend that you use shielded cable to run the DC power up to the preamplifier box. In **Photo D**, the picture of the preamplifier box, you can see a 15-volt, 10-watt zener diode. It's connected across the preamplifier power line. MOVs and large-value bypass capacitors will also help to protect against lightning-induced damage. However, you should always remember to disconnect and ground the leads when thunderstorms are in the area. After one severe thunderstorm, I found that the diode across my antenna relay was blown, but the relay and preamplifiers were unharmed thanks to the zener diode.

In general, 432 MHz is free from significant out-of-band interference. But if you live in a high-density RF location, especially if there are UHF and FM broadcast stations nearby, you may experience interference. This interference usually occurs when the broadcast stations are so strong that they overload the preamplifier and cause IMD mixing products in the preamplifier. Solutions include using as selective as possible a first-stage preamplifier (like a cavity-input circuit model), using a trap for the offending FM station, or using a cavity filter in front of the preamplifier. If you appear to have trouble hearing, although everything else seems normal, try tuning around with the receiver set in its widest bandwidth with either an AM or FM detector turned on. You may be able to detect the modulation of either the FM or TV station, if it's causing a problem. My own receiver arrangement uses a very low noise figure cavity preamplifier connected to a four-pole interdigital filter and then to an impedance-matched second stage (similar to the N6CA preamplifier). Without the filter between the stages, I would be wiped out from UHF TV and FM broadcast interference. My attempts to use an unmatched second-stage preamplifier were unsuccessful.

Feedlines

Questions are often raised regarding the type of transmission line to use. There's a simple answer for the transmit line. You want the largest, lowest-loss line you can afford or scrounge and still be able to put up your tower. The run should be the shortest possible length. Just as with phasing lines, the all-copper "hardlines" like Andrew Heliax® or Cablewave Wellflex® (low density foam or air dielectric) are best. Cable in ½ and ⅞-inch sizes often shows up at flea markets. The ⅞-inch sizes are often more available, as many amateurs don't want to hassle with the large cable. You can usually find copper rigid line as low-cost surplus (typically around the scrap copper cost). Rigid line comes in 1⅝ inch and larger sizes (up to 8 inches!). It's heavy and cumbersome to handle, but very low loss. Aluminum jacket hardline is the next best choice. Seventy-five ohm CATV cable is better than braided shield coaxial cables and can have very low loss. You can either make the whole system 75 ohms, or use matching transformers. Don't use 75-ohm CATV cable for phasing lines unless you can measure the lengths electrically. I've seen some large velocity factor variations in CATV cable—even from the same reel.

Choosing the flexible line which goes around the rotors is more problematical. Here your choice is limited. Andrew ½-inch Superflex will be the lowest loss, but it's a very hard cable to keep water free. RG-213/U and RG-214/U are marginal in terms of power handling and are high loss. Larger solid dielectric cables like RG-14/U or RG-17/U aren't all that flexible, and can pull the connector center pins apart in cold weather. Some operators have used semiflexible cables like LDF4-50A and simply made very large loops. Using a foam-type RG-8 cable and replacing it every couple years may be the best tradeoff. I've never had any problems with N connectors handling 1500 watts at 432 MHz, provided they were properly installed. However, I've completely destroyed N connectors that didn't have good shield contact or had badly misaligned center pins.

The requirements for the receive line depend on how long your run is and how much preamplifier gain you have. If you have a long run of cable, you'll need either a two-stage preamplifier system or very low loss cable (like 7/8-inch O.D.). Unless you can find a good deal on cable, the best price tradeoff may be to use a two-stage preamplifier system and smaller cable (foam RG-8 type, for instance). If you are using a single-stage preamplifier, you don't want more than 2 dB of feedline loss. For two-stage systems, line losses of up to 10 dB can be tolerated.

Tracking the Moon

One of the major reasons new EME stations fail is that they simply may be unable to reliably point their array at the Moon. Some stations may start out visually sighting the moon. This quickly becomes tiresome because it requires many trips out to the array. Cloud cover can also make visual sighting impossible, causing you to miss operation on good EME days. The advent of the personal computer has made Moon tracking a simple task. For those with IBM MS-DOS compatible computers, there's a wide variety of shareware, public domain, and commercial software.

One of my favorite programs is W9IP RealTrak. This program will not only print out antenna azimuth and elevation headings for almost any celestial object, but if used with an analog input/digital output board, it allows you to automatically track any of the objects with one or two arrays in real time. Objects you can track include: the Moon, the Sun, celestial noise sources and cold sky areas, and most amateur satellites. RealTrak is available in complied form only for IBM DOS-compatible systems, with or without a math coprocessor.* RealTrak can use the IBM 6451502 Data Acquisition Board,** the Real Time Devices AD-200 or AD-1100 A/D boards,† and can also be interfaced with the Kansas City Tracker system. A similar program, VHF PAK, is sold by WA1OUB.†† VHF PAK has real-time Moon and celestial object position information, but doesn't have automatic array tracking capability. It does include other programs for meteor shower prediction and distance calculation.

Another automatic array tracking program has been introduced recently by Gary Meyers, K9RX.* You can order the program and, if you wish, the hardware to interface the computer to the array indicators and rotors. One of the available hardware interfaces allows several hours of tracking information to be downloaded to the interface. This means you can use your computer for other tasks, or turn it off while the array still tracks in real time.

Still another real-time tracking program is available from Doug McArthur, VK3UM.** The VK3UM EME Planner features a large database of DX locations which allows for easy common Moon window calculations

*Michael R. Owen, W9IP/2, 21 Maple Street, Canton, New York 13167. Cost is $40. Purchasers can receive free updates by sending formatted floppies.
**This IBM A/D interface has been out of production for several years. They have been, and still may be, available through surplus channels.
†Real Time Devices, State College, Pennsylvania.
††VHF PAK ($35) and VHF87 PAK ($45 for computers with math coprocessors) are available from Bob Mobile, WA1OUB, RFD #2, Box 442, Hillsboro, New Hampshire 03244, (603)464-3187.
*EME Tracker Program, Gary Meyers, K9RX, 1753 Elmwood Drive, Rockhill, South Carolina 29730, (803)327-6024.
**Doug McArthur, VK3UM, 30 Rolloway Rise, Chirnside Park 3116, Australia.

from virtually any place on the Earth. Automatic array tracking capability is also built into the program for use with an interface circuit which connects to an RS-232C serial port. Separate versions are available for use with and without math coprocessors. The program is available in compiled form only, but is free for a blank disk and return postage. Doug also has available for sale a limited number of printed circuit boards for the construction of the computer-to-array interface circuits.

Before the widespread use of real-time programs and auto-tracking, the WA1JXN MOON program was one of the most popular.[6] The program will display on the screen or print out the Moon's azimuth, elevation, GHA, declination, and sky temperature. Provision is included to calculate predicted signal loss from ideal conditions (due to the Moon's distance) and the sky temperature behind the Moon (due to celestial noise sources). Other programs available from WA1JXN include: SUN, which calculates the Sun's position, and STARS, which determines the position for several celestial noise objects. BASIC source code is available, as well as compiled versions.*

Those with other home computer systems don't have quite as large a selection of EME software. However, several programs are available. One source for programs for the Apple, TRS-80, and Commodore VIC-20 is *Computer Programs for Amateur Radio.*[7]

Those who don't own computers can still find the Moon when it's not visible. *The Astronomical Almanac*[8] (printed annually) contains data on the position of celestial objects in GHA and declination format. Several programs are available to translate this information into local azimuth and declination data.[9]

Once you have a computer program which gives you the moon's azimuth and elevation, you need to calibrate your azimuth and elevation indicators. The -3 dB beamwidth of this array will be approximately 11 degrees in both the E and H planes. However, array aiming accuracy must be much greater than ± 5.5 degrees, as a 3-dB loss (6 dB on your own echoes) will be unacceptable. Minimum acceptable aiming accuracy is that necessary to keep the array pointed at the Moon, within its -1 dB beamwidth points; that is, 7 degrees or ± 3.5 degrees. For best results, the array should be aimed within its -0.5 dB points; this corresponds to ± 2.5 degrees.

Your first step is to make sure all the Yagis are aligned parallel to each and are perpendicular to the stacking frame. If all Yagis are steered off to one direction, the pattern of the array will be skewed in the same direction—usually by twice the physical misalignment. Gain deterioration will be minimal if the alignment isn't too bad. However, this misalignment will make a boresight useless for confirming the array's direction.

The best way to calibrate the indicators is to use the Sun. Because most programs give headings for the Sun, you simply peak the array on the Sun by rocking it back and forth (in both azimuth and elevation) until you obtain the maximum S-meter reading. Then adjust your indicator calibration controls so the indicators agree with the correct position. Many rotors let you do this from the shack. In some cases, the rotor indicator may not track correctly over the rotor's full rotation. It's advisable to check your azimuth and elevation readouts at several positions. If calibration cannot be maintained, it's advisable to make up a calibration chart listing indicator readings versus actual position. Having good position indicators will make EME operation much more enjoyable and possible when you aren't actually able to see the Moon.

Testing your system

The simplest system test you can perform involves pointing your array at the Moon and listening for EME signals. However, you may want to use a more quantitative approach to find out how well the system is working. Sun noise is a very useful test. The method recommended to measure Sun noise uses a step attenuator in the IF line. First point the array at cold sky. Then using the widest available receiver bandwidth, turn the product detector on and the AGC on slow speed, and adjust the attenuator for an up scale S-meter reading (S3 to S5 is usually good). The two major cold sky spots (Aquarius and Leo) are listed in many of the tracking programs I've described. Now turn the array towards the Sun and move it around until you find the highest noise level. Then switch the attenuation in until the S meter returns to its original cold sky reading. The number of dB of Sun noise is the difference between the current and original number of dB attenuation inserted. The Sun is the strongest, but also the most variable, noise source; this limits the effectiveness of Solar noise readings. For the 4 × 22-element Yagi EME array, you should see between 10 and

*Distribution of the WA1JXN, and in North America, the VK3UM programs has been a problem. WA2TIF has been handling the freeware distribution. To receive copies of the programs, send formatted disks with a postpaid self-addressed mailer. The VK3UM EME planner and data files fill a 360-KB disk. Specify standard or coprocessor version. The WA1JXN MOON, STARS, and SUN programs with both source and object files will all fit on a single 360-KB disk.

14 dB of Sun noise—depending upon the Sun's activity level. The best absolute check is to compare Sun noise readings with a few of the more experienced EME stations, when measured at the same time. This information is often exchanged on the weekend 20-meter EME nets. Sun noise is a combination measurement; that is, it's dependent upon the array gain, noise pickup, phasing-line and relay losses, and system noise temperature.

Other noise sources that should be detectable are the galactic core (Sagittarius), Cygnus-A, and Cassiopeia. Sagittarius should give about 4.5 dB to cold sky, Cygnus-A about 2.0 to 2.5 dB, and Cassiopeia about 2.0 to 2.2 dB. Cygnus-A and Cassiopeia behave the same as the Sun; that is, they are composite system indices of performance. Because the galactic core is such a large noise source (in terms of sky area covered), it's less dependent upon array gain than the other noise sources. Use the same measurement procedure to measure these noise sources that you'd use to measure Sun noise. Establish cold sky reference level and then move the array to the noise source. These noise sources are very stable with time; however Cassiopeia is slowly fading in strength.

You can use Earth noise as a final system check. To measure Earth noise accurately, your array must be fairly high and able to be pointed about 10 to 15 degrees below the horizon. Because the Earth is larger than the array's beamwidth, Earth noise is independent of array gain and is a measurement of the unwanted array noise pickup, along with phasing-line losses and receive system noise temperature. You measure Earth noise using a technique similar to those for other noise sources. Once again, establish a cold sky reference level. Than lower the array elevation until you observe a noise peak (typically 12 degrees below the horizon).

A comparison of these noise sources can help you diagnose array problems. For example, if Sun, star, and Earth noise are all low, your problem is most likely a poor (or overloaded) preamplifier, bad T/R relay, or lossy phasing lines. It could also be due to poor Yagi design, if you're using one of the older designs not included in **Table 1**. Good Earth noise, but low Sun and star noise readings, indicate an array gain problem. It could be due to an out-of-phase Yagi, array misalignment, or phasing lines of unequal electrical length.

EME operation

The real excitement begins when your array is completed and the electronics are connected. EME operation on 432-MHz is highest on the one weekend per month designated as the activity weekend. This is the time when the Moon's position during its 28-day cycle is most favorable to making EME QSOs. Favorable conditions are a tradeoff of several factors: the Moon's declination, distance from the earth, offset from the Sun, and sky temperature (that is, background noise level). The weekends are usually scheduled several months in advance, so you have plenty of notice. You can find out about these optimum operating times by checking into the 432-MHz and above EME net held on 14.345 MHz at 1066Z Saturdays and Sundays. Active EME operators obtain operating and schedule information by reading the "432-MHz and Above Newsletter" published by K2UYH.[10] Those with a more casual interest in EME news can read reprints from the "432-MHz and Above Newsletter" and other EME news by subscribing to the "EME Newsletter" published by K0IFL.*

On activity weekends, there's plenty of operation between 432.000 to 432.025 MHz. Stations generally transmit and listen in 2.5-minute sequences to avoid confusion about when a station has finished transmitting. More specific information on how to operate can be found in *The ARRL Handbook*.[2] Many stations often operate without any particular sequencing when signals are strong enough. Just as they do on the HF bands, a station will simply find a clear frequency, call a CQ, and see who answers. For myself, the ultimate thrill of all amateur activity is to pound out a CQ and have a new and distant EME station answer.

During the activity weekends, there are scheduled QSOs on 5-kHz frequency increments starting at 432.025 MHz (432.025, 432.030, 432.035, and so on). Scheduled operation is normally held to line up stations that haven't been able to catch each other randomly and to help facilitate QSOs with new stations or between weaker stations. Because the Moon is up as many as 14 hours on a high-declination day, it helps to know the best times to operate if you only plan to put in an hour or two that day. The greatest number of 432-MHz EME stations are in Europe; the next highest number are in the United States. Consequently, peak activity times occur when the Moon is up over both Europe and the United States. If you are a North American station and have any options as to where you locate your array, try to get a good shot to the northeast so you

*"EME Newsletter," John M. Carter, K0IFL, P.O. Box 554, Union, Missouri 63084. Stations not currently active should also obtain subscriptions to K2UYH's through K0IFL in order to assure that active operators obtain schedule and activity information from K2UYH in a timely manner.

can operate as close to moonrise as possible during high-declination day. The Moon's movement through the sky (rising in the east and setting in the west) is actually very similar to the Sun's perceived movement. The Moon simply moves through its primary cycle (maximum to minimun declination and distance) in 28 days, instead of the 365.25-day Earth-Sun orbit.

You'll often find EME activity outside the designated weekends—especially when the Moon is up in the evening. The only times you're really unlikely to find someone on the air are those when the Moon is at extreme southern declinations (low in the sky) and behind the galactic core (highest sky noise). The peak activity each year occurs during the ARRL International EME Competition held on one weekend during October and a second during November. Last year more than one station managed close to 150 QSOs with different 432-MHz stations during the contest.

Conclusion

Assembling a 432-MHz EME station is certainly more complex than getting on 75 meters with a dipole. However, your EME station can be just an assembly of commercial or surplus equipment. No design work is required, and you needn't build anything from scratch. If you isolate each of the complexities which EME operation adds to your system requirements, view them as separate steps, and handle them one at a time, you'll find that EME suddenly isn't all that intimidating. ■

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10. *The 1990 ARRL Handbook*, American Radio Relay League, Newington, Connecticut, pages 23-30 to 23-38.

PRODUCT INFORMATION

HP Introduces PC-Card Version of the HP 4957A Protocol Analyzer

Hewlett-Packard Company has expanded its family of wide-area-network (WAN) field-service protocol analyzers to include a personal computer (PC) card version of the HP4957A protocol analyzer. The HP4957PC fits into a number of portable and desktop PCs including the HP Vectra PC and other IBM-compatible PCs.

Like other HP WAN-testing solutions, the new protocol analyzer is designed to help network-service organizations of major computer manufacturers, communications-equipment manufacturers, and network service providers isolate network problems quickly. The HP4957PC has a 3/4-Mbyte capture random-access memory (RAM) and an optional 256-Kbps fast-capture mode to support high-speed links. In addition to performing all of the test functions of a full-sized protocol analyzer, the PC-card solution performs PC functions like terminal emulation and report generation. It also provides solutions for ISDN, T1, and frame-relay protocols.

For more information on the HP4957PC contact Hewlett-Packard Company Inquiries, 19310 Pruneridge Avenue, Cupertino, California 95014.

Brian Beezley, *K6STI*
507-1/2 Taylor Street
Vista, California 92084

THE TRIANGLE ANTENNA

An alternative horizontal omnidirectional antenna

Although they offer less gain than directional antennas, antennas which radiate equally in all compass directions have many applications. It's easy to get omnidirectional coverage from an antenna system if you're willing to accept vertical polarization. However, you may need horizontal polarization for communications compatibility; when you're doing weak-signal VHF/UHF work, for instance. And for ionospheric propagation, significant signal enhancement is often possible with horizontal polarization.

I'd like to present an alternative configuration for a horizontal omni. Comparison is made with conventional omnis.

Antenna modeling

I modeled all the antennas on an IBM PC compatible computer using software based on the MININEC algorithm.* This antenna-analysis method was developed at the United States Naval Ocean Systems Center in the 1980s. Antenna geometry is defined by the user. You specify the antenna conductor dimensions, their orientation in space, feed-point location, operating frequency, and other information. Ground conductivity and dielectric constant may be specified as well.

MININEC applies the fundamental equations of electromagnetism to the antenna system and solves for the current along each conductor. The input impedance, forward gain, front-to-back ratio, conductor losses, SWR, and radiation patterns are calculated using the currents. Patterns are displayed on the computer screen, and you have the option of printing a hard copy for later reference.

Using a computer to compare antenna designs offers many advantages over performing field experiments with real antennas. First, it's possible to investigate a great many trial designs on the computer in the time it would take to construct and evaluate just one physical model. Computer experiments may be performed under ideal, free-space conditions without measurement inaccuracies caused by ground terrain or nearby objects. You may observe small but significant differences in antenna performance, which might be easily lost in experimental measurement errors. Periodic recalibration of the antenna range and test equipment isn't necessary. Finally, and perhaps best of all, you can accurately and easily determine absolute antenna gain.

Conventional horizontal omnis

Amateurs use several types of horizontal omni antennas. Here are some of the most popular designs.

•The Turnstile

The Turnstile is shown in **Figure 1**. This antenna consists of two perpendicular dipoles with equal currents and 90-degree phase shift. When the dipoles are placed in the horizontal plane, the azimuth radiation pattern is substantially omnidirectional. **Figure 2** shows the azimuth pattern using rectangular coordinates. For the antennas considered here, pattern detail is much easier

*MN 4.0 Antenna-Analysis Software was used to analyze all antennas. This software includes corrections for the following problems in the original MININEC algorithm: frequency-offset errors, errors for large-diameter wires, bent-wire anomalies, and errors from dissimilar segment lengths. The NEC-2 Numerical Electromagnetics Code was used to verify results for Triangle antennas.

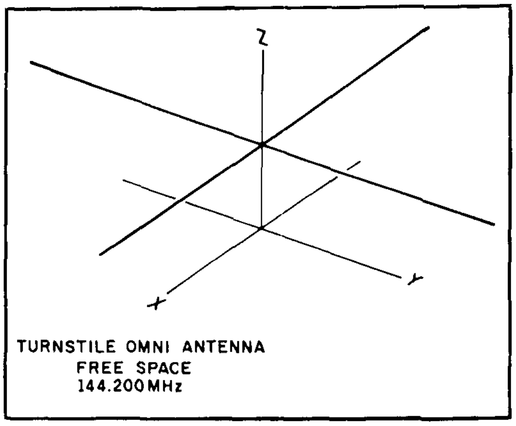


Figure 1. Turnstile antenna. The half-wave dipoles are fed at their centers with a phase shift of 90 degrees.

to see when rectangular coordinates are used instead of polar coordinates.

On average, the Turnstile is 3.6 dB down from a single dipole in free space (− 3.6 dBd). The pattern also has about 1 dB of ripple in the azimuth plane. This is good enough for most omni applications.

The Turnstile requires a feed system with 90-degree phase shift. The phase shift can be achieved by feeding one of the dipoles directly and feeding the other dipole through a phasing line. The most straightforward feed system uses a phasing-line impedance equal to the dipole input impedance and a quarter-wave electrical phasing-line length. The correct phasing-line impedance is a function of the diameter and length of the dipole element. The parallel impedance combination of the phasing line and remaining dipole is then matched to the transmission line. This can be accomplished with a second quarter-wave transformer section. The impedance of this line won't be the same as that of the first phasing line. The design complexity of the feed arrangement is one of the drawbacks of the Turnstile. On the other hand, the turnstile still gives reasonable omnidirectional coverage with a less than perfect feed system.

•The Half-Wave Loop

The Half-Wave Loop is simply a dipole bent into a square (ends close but not touching). On average, the gain of this antenna is − 2.7 dBd in the horizontal plane. The Half-Wave Loop has an average gain of 0.9 dB over the Turnstile. However, the omnidirectional characteristics aren't nearly as uniform. The Half-Wave Loop shows a 3.5-dB variation in azimuth response. This amount of pattern ripple may be too large for many applications. The Half-Wave Loop has an input impedance of about 10 ohms.

•The Halo

The Halo antenna is similar to a Half-Wave Loop, except that the dipole is bent into a circle rather than a square. Halos often use capacitive loading between the open ends of the circular dipole. For reference, I modeled an unloaded Halo on my computer using a 40-sided polygon. Its performance was somewhat better than that of the Half-Wave Loop. The unloaded Halo showed 3.1 dB of pattern ripple in azimuth and an average gain of − 2.5 dBd.

•The Squalo

When a capacitively loaded Halo is bent into a square, the antenna is called a Squalo (**Figure 3**). Capacitive loading reduces the resonant size of a loop and improves performance. For a Squalo with 8-inch sides at 144 MHz (39 percent area reduction), pattern ripple drops to 2.1 dB and average gain rises to − 2.0 dBd (**Figure 4**). However, the input impedance drops to less than 4 ohms. This reduces SWR bandwidth. The Squalo can be gamma matched. It also may be excited with

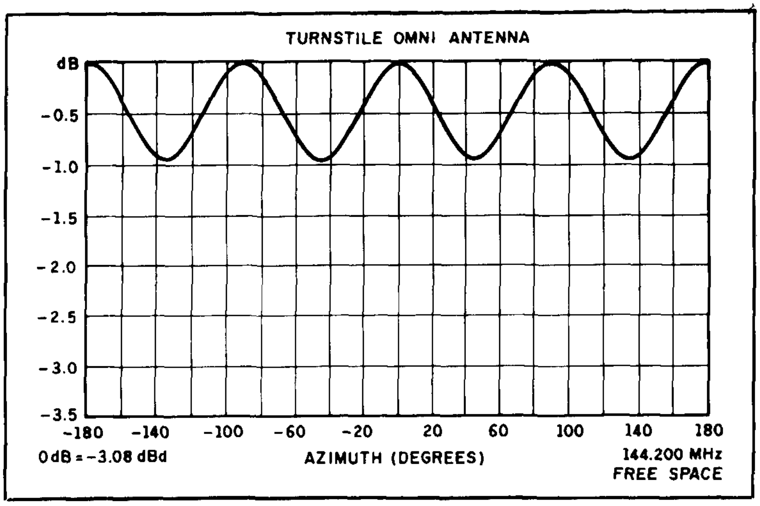


Figure 2. Turnstile azimuth pattern.

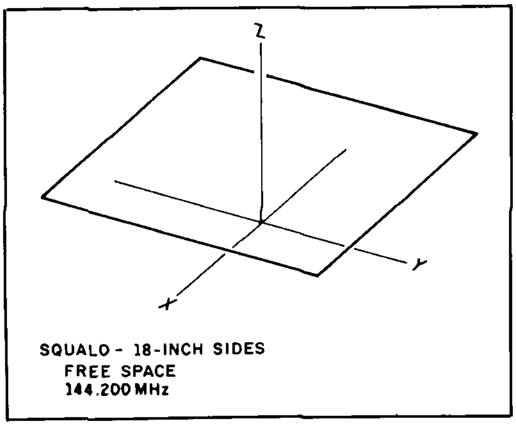


Figure 3. Squalo Antenna. The antenna is fed at the center of one side. Capacitive loading is placed at the center of the opposite side.

a small, auxiliary loop inductively coupled to the main loop.

The Squalo is very compact when compared with a Turnstile. It's often used as a mobile antenna for 2-meter SSB work.

•The Small Loop

When a capacitively loaded loop is greatly reduced in size, a Small Loop antenna is formed. You can distinguish Small Loops from larger loops by their uniform conductor current. The Small Loop has a very uniform azimuth pattern when the plane of the loop is horizontal. Multi-turn, vertically polarized Small Loops have been used as medium-frequency receiving antennas since the earliest days of radio.

Small Loops have one great disadvantage. The input impedance is extremely low (well under an ohm for most designs). This results in low efficiency and very narrow bandwidth. Special low-loss construction techniques must be used for transmitting applications. The gain of a typical Small Loop designed for transmitting is usually several dB below that of a dipole.

I won't go into any more detail on the Small Loop because its losses depend critically on the specific construction techniques used.

•The Two-Dipole Vee

A simple omnidirectional radiator may be formed from a pair of half-wave dipoles arranged in a right-angle vee (**Figure 5**). A balanced feed is used at the high-impedance point where the dipoles join. The pattern of this antenna isn't symmetrical (**Figure 6**). It has a unidirectional peak along the line bisecting the vee, in the direction away from the feedpoint. The average response of this antenna is about -3.0 dBd and the azimuth ripple is about 3.1 dB.

•The Eggbeater

The Eggbeater (**Figure 7**) is similar to a Turnstile, except that full-wave loops are used instead of dipoles. The Eggbeater's name comes from its appearance when circular loops are used. I modeled square loops here for convenience. The Eggbeater outperforms the Turnstile, showing only 0.5 dB ripple in azimuth and an average gain of -2.1 dBd. This antenna is quite compact. Its pattern is shown in **Figure 8**.

The input impedance of one loop of an Eggbeater is about 120 ohms. This antenna can be fed in the same way as a Turnstile, though the required phasing-line impedances are different.

•The Big Wheel

The layout of the Big Wheel is shown in **Figure 9**. This design evolved from three dipoles spaced 120 degrees and fed in phase.[1] No curves were involved, so I modeled the three-dipole system on my computer (**Figure 10**). The response of the curved antenna structure should be similar because the high-

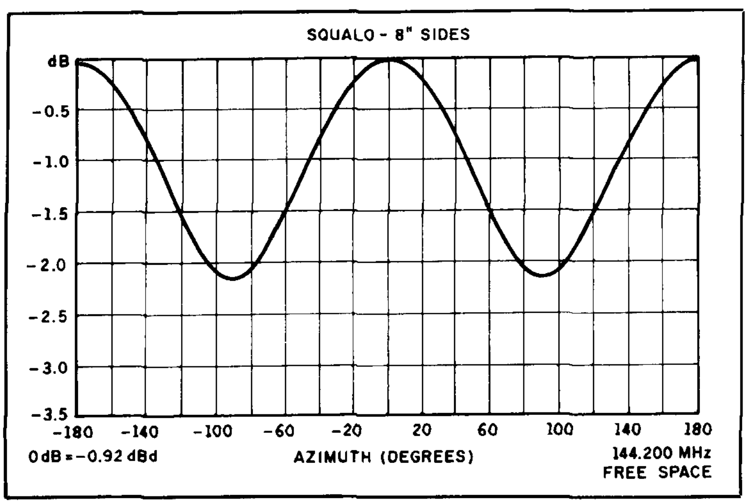


Figure 4. Squalo azimuth pattern.

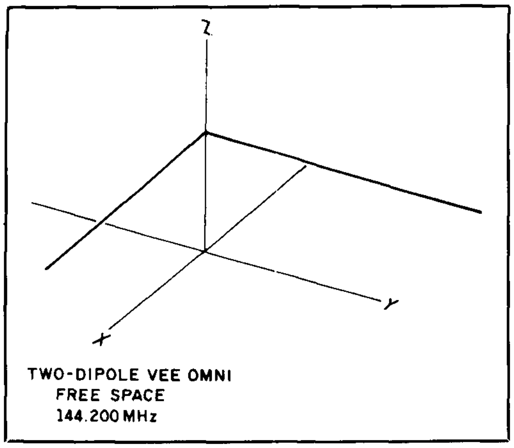


Figure 5. Two-Dipole Vee antenna. The antenna is fed at the high-impedance junction of the two half-wave dipoles.

current portions of the two antennas coincide.

At −1.1 dBd, the Big Wheel shows the highest gain of any of the designs surveyed. It has 0.6 dB pattern ripple in azimuth (**Figure 11**). This antenna is the largest of the group. Although three feeders are required, the elements are simply fed in phase. Folded dipoles fed with 300-ohm twin-lead can be used, with the feeders connected together to form a balanced 100-ohm feedpoint. A 1:1 balun and a quarter-wave matching transformer of 75-ohm coax will yield a 50-ohm unbalanced feedpoint. Big Wheels made of curved conductors use a single compound feedpoint.

The Triangle antenna

The Triangle antenna resulted when I tried to determine if a horizontal omni could have fewer sides than a Squalo. Looking at the layout of the antenna in **Figure 12**, you wouldn't have any reason to suspect that this particular shape might produce an omnidirectional radiation pattern.

The Triangle antenna measures 0.16 wavelength across the base and 0.20 to 0.22 wavelength from base to apex. At the apex, the conductors are separated by a 0.01-wavelength gap. For 0.003-wavelength diameter conductors, use a height of 0.20 wavelength. When using thinner conductors, like wire at HF, increase the height slightly. For example, a height of 0.214 wavelength is just right for no. 12 wire at 7 MHz. The dimensions of a Triangle aren't particularly critical.

The Triangle is fed at the center of the base, where the free-space input impedance is 8 to 9 ohms in series with an inductive reactance of several hundred ohms. The im-

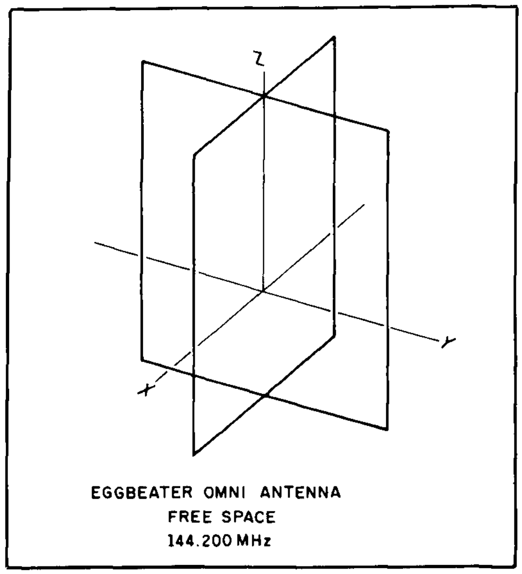


Figure 7. Eggbeater antenna. The full-wave loops are fed with a 90-degree phase shift.

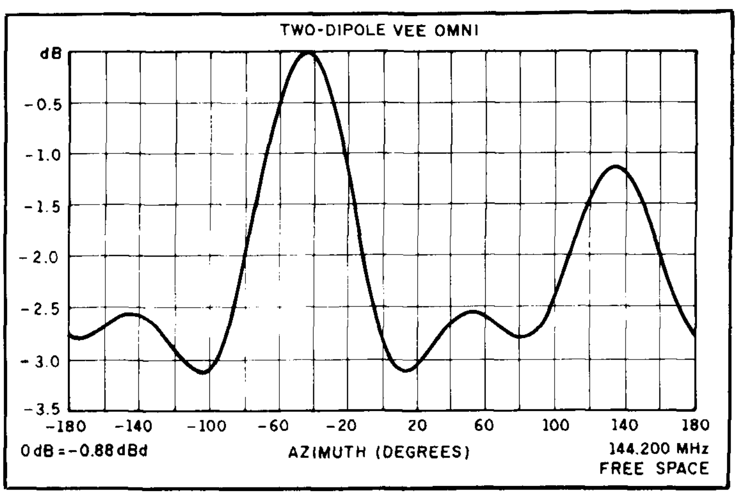


Figure 6. Two-Dipole Vee azimuth pattern. The pattern is symmetrical only about the line bisecting the vee.

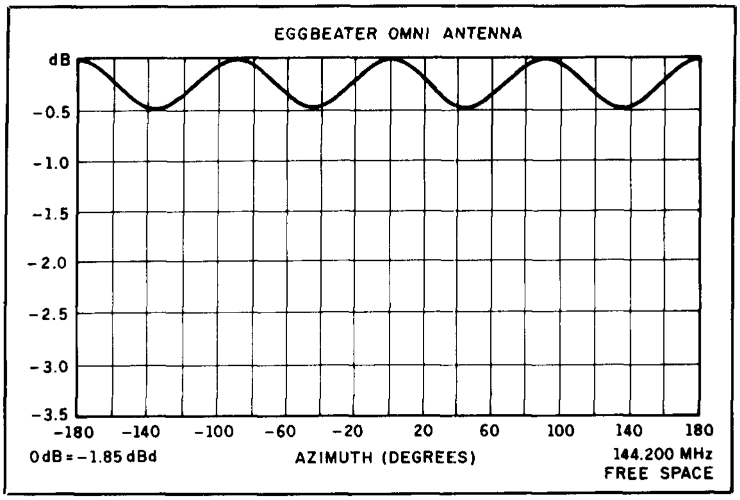


Figure 8. Eggbeater azimuth pattern.

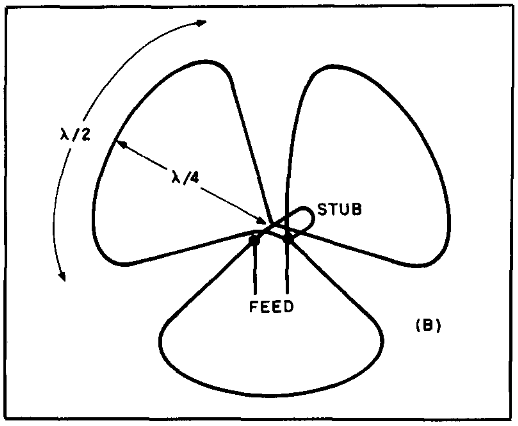


Figure 9. Big Wheel.

pedance varies with conductor diameter and apex height. The large ratio of reactance to resistance suggests that simple matching networks will be effective only over a narrow bandwidth.

The following dimensions at 144.2 MHz produced the most uniform azimuth pattern in my computer model for a conductor diameter of 0.25 inch: base, 13 inches; base-to-apex, 16.25 inches; apex gap, 1 inch. The calculated input reactance is approximately 180 ohms. At 7.1 MHz, the following free-space dimensions produced uniform azimuth response using no. 12 wire: base, 22.2 feet; base-to-apex, 29.6 feet; gap, 17 inches. The input reactance is about 390 ohms.

The azimuth response of the Triangle antenna is shown in **Figure 13**. Surprisingly, the pattern is, in essence, perfectly omnidirectional. I quit tweaking the dimensions when the azimuth ripple dropped below 0.1 dB. The Triangle has a gain of about –1.4 dBd.

The Triangle antenna is nonresonant. This doesn't mean that it won't radiate, or that it's necessarily inefficient. It simply means that the matching network between antenna and feedline must compensate for reactance as well as transform impedance.

The Triangle antenna may be matched to 50-ohm coax by an L-network **(Figure 14)**. Due to the inductive reactance at the feedpoint, matching-network inductance isn't required. Two series capacitors are used rather than one to help maintain physical symmetry, though this probably isn't necessary at HF. At VHF, the capacitors may be constructed from small aluminum plates. When the Triangle is made using flat material, the parallel capacitor can be formed by overlapping the conductors at the feedpoint. A thin piece of Teflon® or other dielectric material may be used for mechanical stability. The series capacitors can be fabricated in the

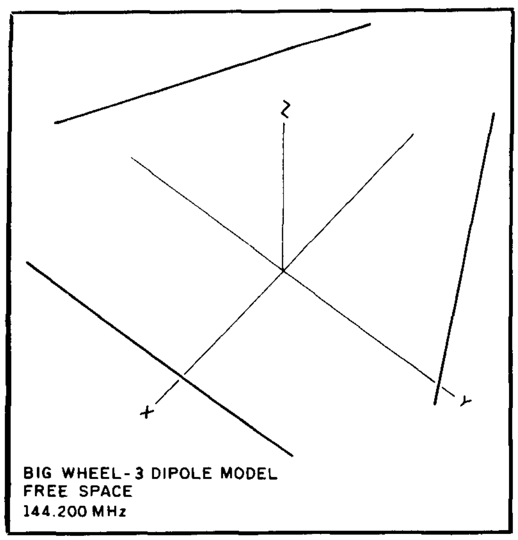


Figure 10. Three-dipole model of the Big Wheel antenna. The half-wave dipoles are fed in phase at their centers.

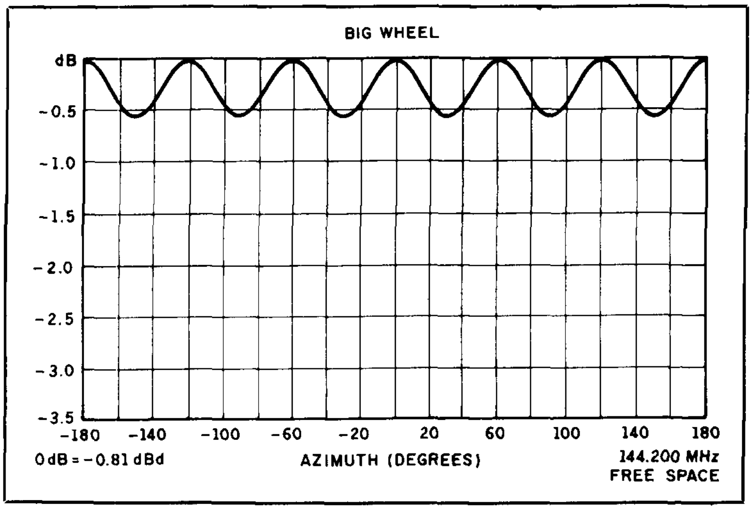


Figure 11. Big Wheel azimuth pattern (three-dipole model).

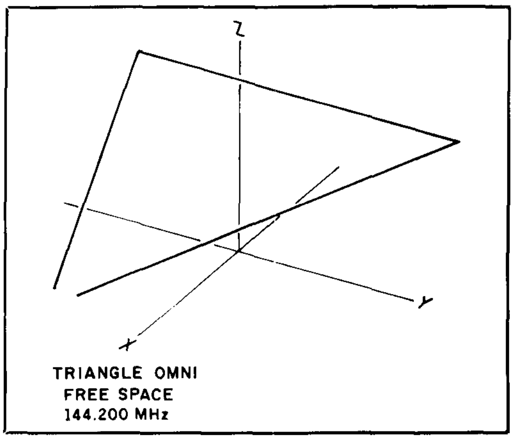


Figure 12. Triangle antenna. The antenna is fed at the center of the triangle base.

same way to form a three-capacitor sandwich at the feedpoint. Use a balun to maintain current symmetry in the antenna conductors and to prevent feedline radiation. The current distribution of the Triangle is shown in **Figure 15**.

The Triangle's main disadvantage is its narrow SWR bandwidth. A Triangle designed for 144 MHz has a SWR of less than 2 over a 1.3-MHz band (the excellent omnidirectional pattern characteristics hold over a much wider bandwidth). A 7-MHz Triangle built of wire has a SWR of less than 2 over a 40-kHz bandwidth. Except for the WARC bands, an antenna tuner is needed to cover an entire HF amateur band, unless a complex reactance-compensating matching network is used.

The Larger Triangle

Another form of the Triangle omni is shown in **Figure 16**. The base of this antenna is more than twice as long as that of the smaller Triangle, although the base-to-apex height is a little less. Calculated 144.2-MHz dimensions for 0.1 dB azimuth ripple using 0.25-inch diameter conductors are: base, 28 inches; base-to-apex, 13 inches; apex gap, 1 inch. At − 1.3 dBd, the gain of the larger Triangle is about the same as that of the smaller Triangle.

The interesting thing about the larger Triangle is its input impedance. For the 144.2-MHz dimensions, the impedance is just above 50 ohms with a series reactance of around 600 ohms. This means that a simple series capacitor and a balun are all you need to feed the larger Triangle with 50-ohm coax. In addition, the larger Triangle has a wider SWR bandwidth. At 144 MHz, the SWR is less than 2 over a 2.2-MHz band. This bandwidth is about 70 percent greater than that of the smaller Triangle. The current distribution of the larger Triangle is shown in **Figure 17**.

At 7 MHz, a larger Triangle made of no. 12 wire has conductivity losses about half those of the smaller Triangle (more on this later). The input impedance is near 60 ohms with a series inductive reactance of about 1300 ohms. The bandwidth is about 75 kHz for a SWR less than 2.

The larger Triangle offers performance advantages in applications where its size isn't a drawback.

Antenna Comparisons

Table 1 compares essential characteristics of the various antennas.

The Turnstile and Squalo are the most commonly used horizontal omni antennas. However, the less well-known antennas offer improved performance in many applications.

The Eggbeater has more gain than the

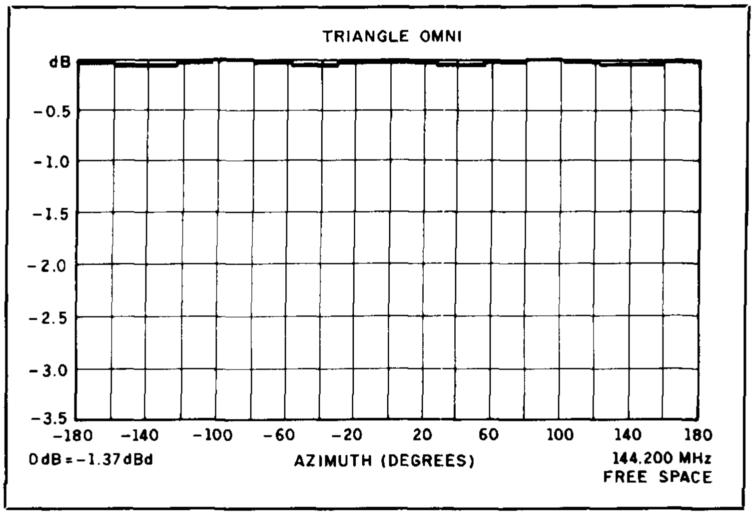


Figure 13. Triangle antenna azimuth pattern.

Turnstile, better omnidirectional performance than the Squalo, small size, and wide bandwidth. Like the Turnstile, it requires a phased feed system.

The Big Wheel offers the highest gain of any of the antennas surveyed, very good omnidirectional performance, and wide bandwidth. While the feed system is somewhat elaborate, the antenna's omnidirectional performance doesn't depend on a good impedance match because all dipoles operate in phase. The large dimensions of the Big Wheel and its physical complexity are its main drawbacks.

The Triangle antenna has perfect omnidirectional performance, unmatched by any of the other antennas surveyed. The smaller version is the second smallest in size, and comes very close to the highest gain. It has 0.6-dB gain over the average response of a Squalo, 2.2-dB gain over a Turnstile, and is only 0.3 dB down from the Big Wheel. The main drawback of the Triangle antenna is its narrow SWR bandwidth. The larger Triangle has a somewhat wider SWR bandwidth.

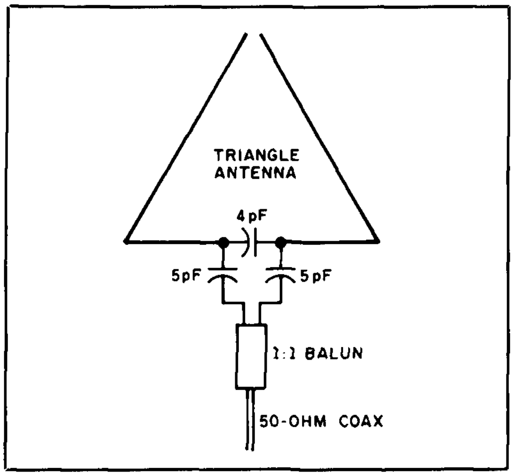


Figure 14. L-network feed for Triangle antenna. Approximate component values for 144 MHz are shown.

Polarization and ionospheric propagation

The magnitude and phase of electromagnetic waves reflected off ground vary with wave polarization. This effects the relative performance of horizontal and vertical antennas because direct and reflected waves combine to form the propagated wave.

For ionospheric propagation over random paths, matching of transmit and receive antenna polarization isn't necessary. Wave polarization changes upon reflection from the ionosphere, with returning waves exhibiting varying polarization[2]. Because

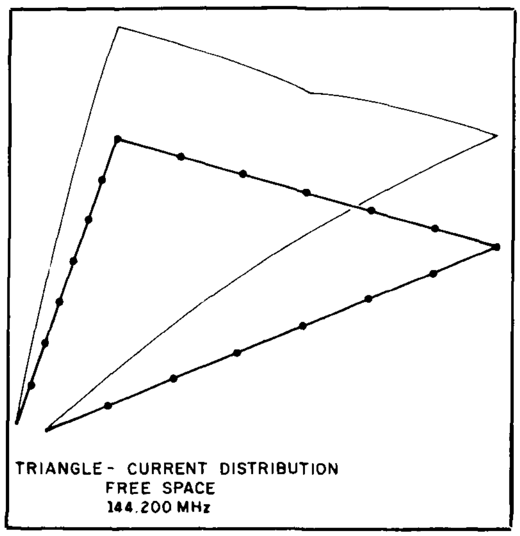


Figure 15. Triangle antenna current distribution. The dots along the antenna conductors indicate the wire segmentation used in the computer model. Only current magnitude is shown.

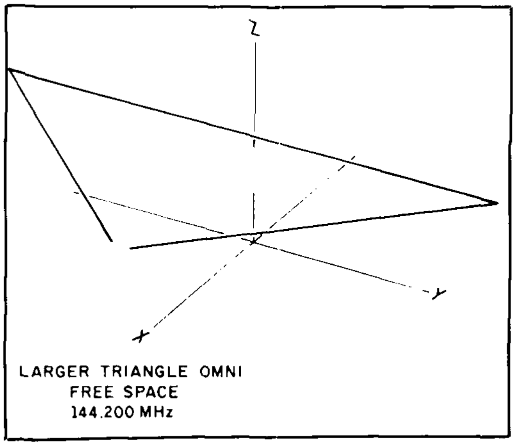


Figure 16. Larger Triangle antenna. The antenna is fed at the center of the triangle base.

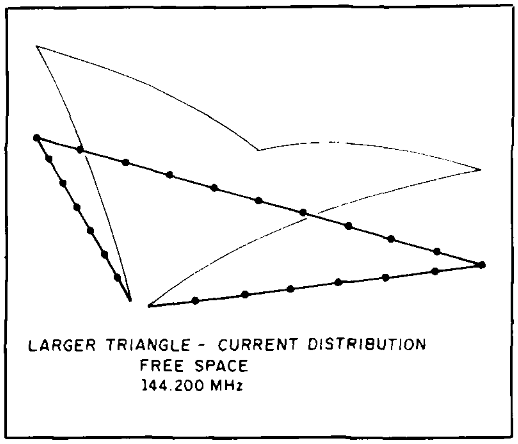


Figure 17. Larger Triangle antenna current distribution (magnitude only).

single vertical radiators are inherently omnidirectional, it might seem that a simple groundplane or vertical dipole would be the perfect antenna for omnidirectional ionospheric coverage.

However, over most kinds of earth, vertical antennas incur ground-reflection losses of several dB when compared with similar antennas oriented horizontally.[3] (This doesn't hold true for low antenna heights or for reflection off seawater.) For example, a 40-meter dipole at 60 feet has about 6.5-dB gain over a ground-mounted, quarter-wave vertical at a take-off angle of 32 degrees, assuming average earth conductivity and dielectric constant. Even at a radiation angle of 10 degrees, the dipole has 2.4-dB gain over the vertical. These are large differences. This comparison assumes no ground-current losses for the vertical, a condition that isn't easy to achieve in practice.

The theoretical low-angle radiation advantages of vertical antennas over perfect ground really don't materialize for most types of real earth. Because their ground-reflection characteristics are so much more favorable, horizontal antennas generally produce better signals than similar antennas oriented vertically, unless radiation is directed over seawater or the horizontal antenna must be located at a low height (in terms of wavelength).

Conductivity losses

Antenna gains quoted so far are for

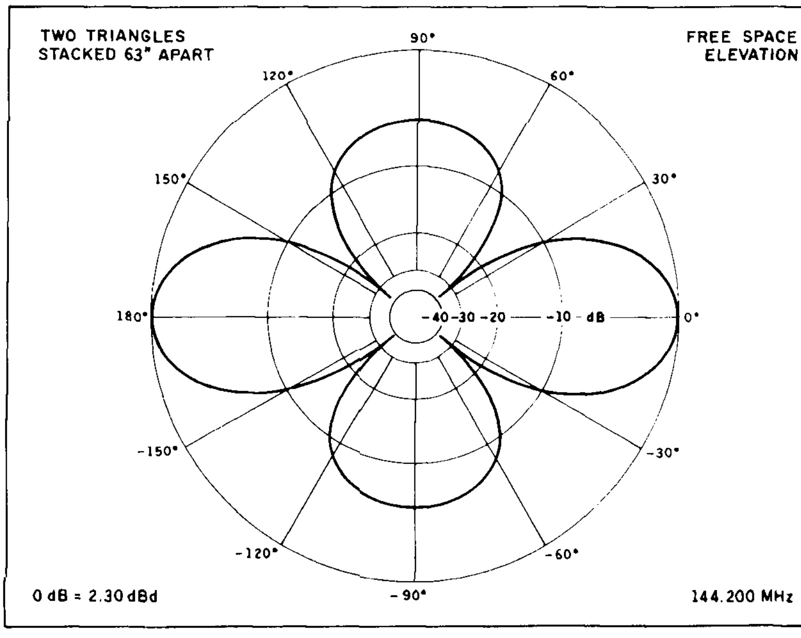


Figure 18. Free-space elevation pattern of a pair of Triangle antennas studied in the H-plane.

Antenna	Gain (dBd)	Ripple (dB)	SWR Bandwidth Compared With a Dipole
Turnstile	−3.6	1.0	Similar
Two-Dipole Vee	−3.0	3.1	Similar
Half-Wave Loop	−2.7	3.5	Much less
Unloaded Halo	−2.5	3.1	Much less
Eggbeater	−2.1	0.5	Less
Loaded Squalo	−2.0	2.1	Much less
Triangle	−1.4	0.1	Much less
Larger Triangle	−1.3	0.1	Much less
Big Wheel (three-dipole model)	−1.1	0.6	Similar

Table 1. Antenna comparisons.

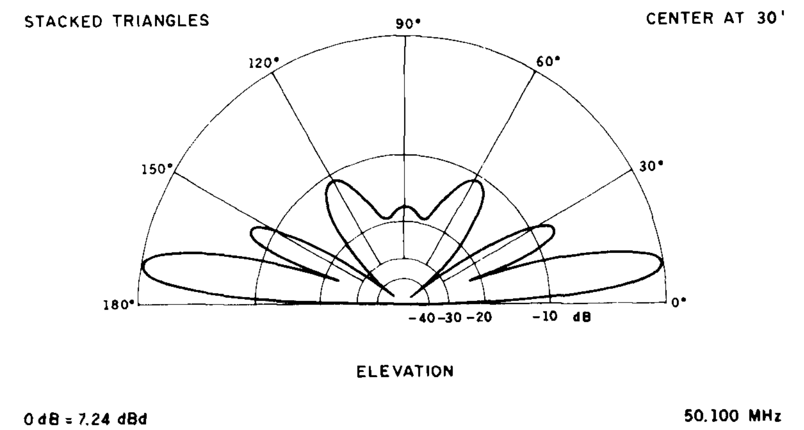


Figure 19. Elevation pattern of stacked Triangle antennas mounted 30 feet above average ground.

lossless conductors. Low-impedance designs incur additional losses due to conductor resistivity and skin-effect. These losses are a function of conductor material, conductor diameter, current distribution, proximity to ground, and frequency.

For example, a free-space Triangle antenna for 7.1 MHz has a calculated gain of − 1.36 dBd using lossless conductors. For no. 12 copper wire, the gain drops to − 2.25 dBd and the efficiency is about 82 percent. A larger Triangle of lossless wire has a calculated gain of − 1.22 dB. The gain drops to − 1.67 dBd for no. 12 copper wire and the efficiency is about 90 percent.

For a 144-MHz Triangle with 0.25-inch diameter conductors of 6061-T6 aluminum alloy, the conductor loss is 0.12 dB and the efficiency is greater than 97 percent.

Applications

•VHF mobile

For horizontally polarized VHF SSB mobile work, the Squalo has been the preferred amateur antenna. Some amateurs use a stacked pair. The Squalo is the most compact of the antennas surveyed. Although the Eggbeater is somewhat larger, it still protrudes less than 10 inches from the support mast at 144 MHz. The Eggbeater has about the same gain as a Squalo, but a more uniform azimuth pattern. It really makes a nice mobile antenna if you don't mind its appearance.

The Triangle is even smaller than the Eggbeater. It's planar, has 0.7 dB more gain, and has better omnidirectional performance. The narrow bandwidth of the Triangle isn't a problem in this application. The Triangle does have pointed corners, unlike a circular-loop Eggbeater. This may be important when the antenna can't be mounted above face level on a vehicle. However, it should be possible to bend the corners of the Triangle into smooth curves with little loss of performance.

•Six-meter beacon

Hundreds of unattended, low-power, automatic beacon transmitters are active worldwide in the 50-MHz amateur band. The beacons allow amateurs to detect and track the irregular and unpredictable ionospheric band openings that occur in this frequency range.

For beacon use, a vertical groundplane or

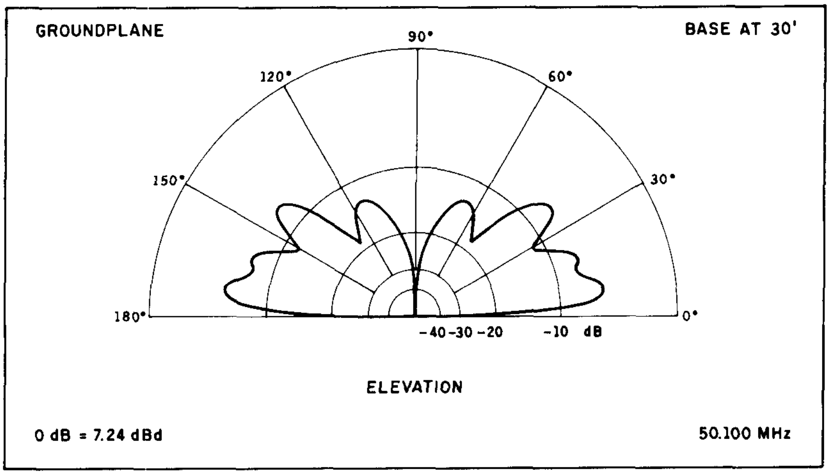


Figure 20. Elevation pattern of a groundplane antenna mounted 30 feet above average ground. This pattern uses the same scale as *Figure 15*d so relative gains may be compared easily.

vertical dipole is a great low-angle radiator at locations where reflection can take place off seawater. Turnstiles also are commonly used for beacon antennas. However, unlike a groundplane, a Turnstile radiates much of its power overhead. Because of this, a groundplane outperforms a Turnstile by 1 to 2 dB at take-off angles of less than 5 degrees—even over lossy earth. But a Triangle antenna has 2.2 dB gain over a Turnstile, better omnidirectional performance, and is more compact. Its narrow VSWR bandwidth isn't a drawback in this application. For additional gain, a pair of Triangles can be stacked in the vertical plane.

A stacked pair of Triangles has 3.7-dB gain over a single Triangle and 2.3-dB gain over a dipole. For maximum gain, a pair of Triangle antennas should be stacked 0.77 wavelength apart. This distance isn't critical. The stacking gain is within 0.1 dB of maximum for stacking distances between 0.70 and 0.84 wavelength. **Figure 18** shows the elevation pattern for a pair of stacked Triangles in free space. Note the compression of the pattern on the horizon.

Figures 19 and **20** compare the elevation patterns of a groundplane and a stacked pair of Triangles at 50.1 MHz. The plots are normalized to the same absolute gain scale so that gain differences are apparent. The center of each antenna is 30 feet above ground. The stacked Triangles have 3 to 4 dB gain over the groundplane at useful wave angles (below 5 degrees). Low-angle details of the patterns are shown in rectangular coordinates in **Figures 21** and **22**. Average earth characteristics were used for these models (conductivity, 5 millisiemens per meter; dielectric constant, 13).

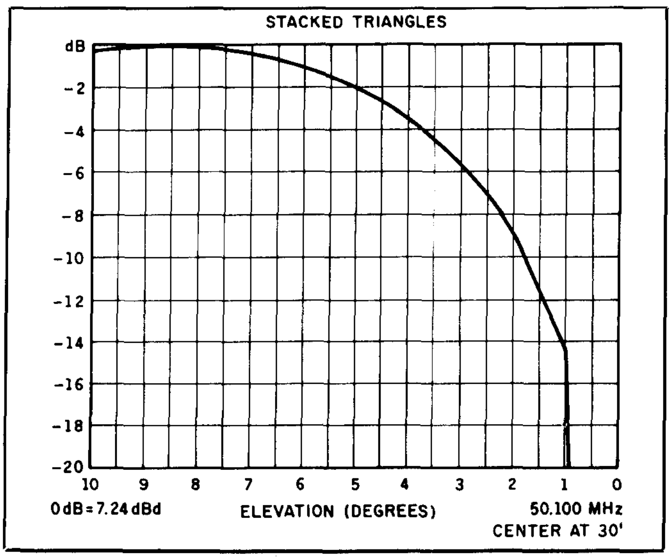


Figure 21. Low-angle details of the elevation pattern of the stacked Triangles of *Figure 19*.

•VHF beacon

Four Triangle antennas may be stacked with 0.77 wavelength between adjacent antennas to provide 5.6 dBd gain. This antenna system is about 16 feet tall at 2 meters. **Figure 23** shows the free-space elevation pattern for four stacked Triangles. the compression of the lobe on the horizon is very pronounced.

A stack of four Triangles would make a nice antenna for a 2-meter tropospheric beacon. This system also could be used as a high-gain, omnidirectional antenna for a horizontally polarized, packet-radio network. Orthogonal polarization might be used to accommodate additional packet-radio traffic in congested metropolitan areas by creating new virtual channels on (or sandwiched in between) frequencies with existing vertically polarized activity.

•ATV repeater

The polarization used for amateur television varies across the United States. For horizontal polarization at 432 MHz and above, four or more stacked Big Wheels could be used for an ATV repeater antenna. The bandwidth of the Triangle antenna (even the larger version) is probably too narrow for this application. Wide antenna bandwidth is necessary for good ATV picture fidelity. The size of the Big Wheel isn't a drawback in this frequency range. If you choose the three-dipole version, you may use a single piece of rod for each folded-dipole element and associated feedline to simplify construction.

•HF

A Triangle antenna is usable at HF, if you can live with restricted SWR bandwidth or are willing to retune an antenna tuner. When the antenna is made of wire, three supports are required and there will be some conductor loss. On the plus side, the Triangle has much more uniform coverage than a dipole or inverted vee. The antenna can be used on more than one band, although perfect omnidirectional performance will exist only at the design frequency. Even so, at twice the design frequency the Triangle has just 4.3-dB ripple in azimuth. This still provides much more uniform coverage than a dipole, inverted vee, G5RV, or random wire. The Triangle shouldn't be used much below its design frequency. At half-frequency the input impedance falls to 1 ohm, a value much too low to be practical with wire conductors.

A Triangle for 10 or 12 meters can be made of aluminum tubing and mounted on a single mast. This solves both the three-support and conductor-loss problems. A Triangle can easily cover all of 12 meters.

At HF, the lower losses and wider SWR bandwidth of the larger Triangle may offset the inconvenience of its larger size. At half-frequency, the larger Triangle has an input impedance of about 7 ohms in series with a capacitive reactance of about 350 ohms. It has half-frequency losses of 0.6 dB for no. 12 wire and a dipole-like pattern with 12 dB of azimuth ripple. While not omnidirectional, the antenna will be reasonably effi-

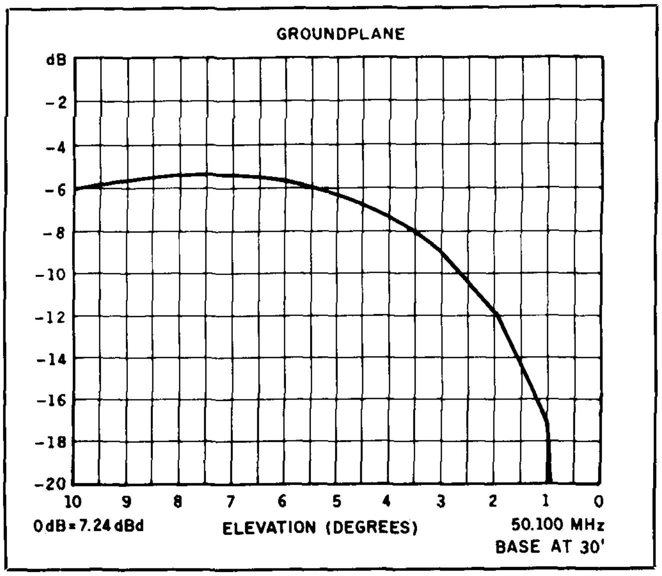


Figure 22. Low-angle details of the elevation pattern of the groundplane of *Figure 20*. Same scale as in *Figure 21*.

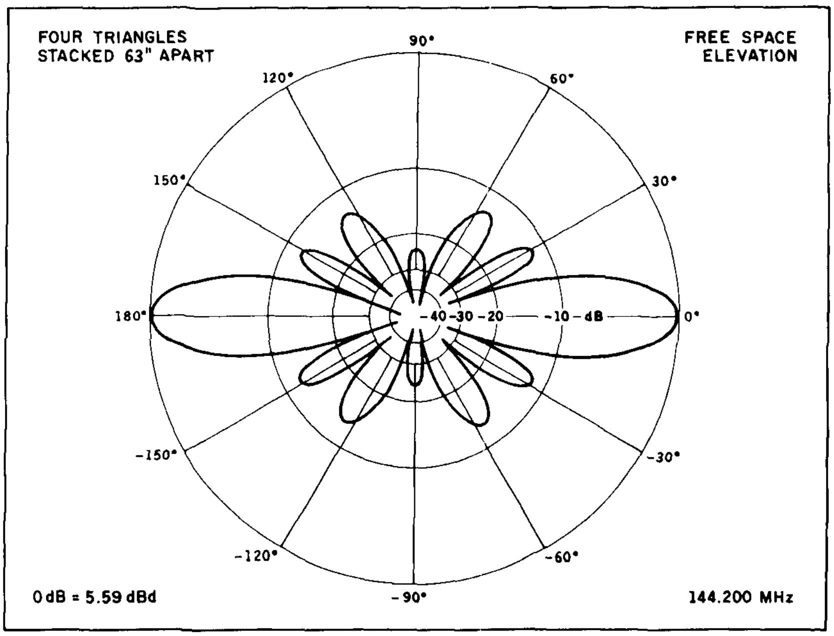


Figure 23. Free-space elevation pattern of four Triangle antennas stacked on the H-plane.

cient at half-frequency when matched with an antenna tuner.

A Half-Wave Loop has wider bandwidth than a Triangle and is easier to feed. Although it has 1.3 dB less gain, you give up some of the Triangle's gain advantage when wire conductors are used. Wire losses for a Half-Wave Loop are about half those for a smaller Triangle and are comparable to those for a larger Triangle. The Half-Wave Loop's 3.5 dB of azimuth ripple isn't bad when compared with typical HF antennas.

A Two-Dipole Vee is much less compact than a Half-Wave Loop and has slightly less gain, but it has better pattern ripple and requires one less support. It may be used easily on other bands (where it will be directional), including at half-frequency.

An auxiliary, horizontally polarized omnidirectional antenna is useful in conjunction with a high-gain, directive array, such as a large Yagi. Weak signals are easy to miss when a narrow beam is pointed in another direction. If you switch to an omni periodically while tuning around, you may be able to detect stations missed on the Yagi. The omni also can be used to quickly determine whether your beam is pointed in the right direction without having to rotate it. If you switch to the omni and the signal strength doesn't drop dramatically, the station is being received on a sidelobe.

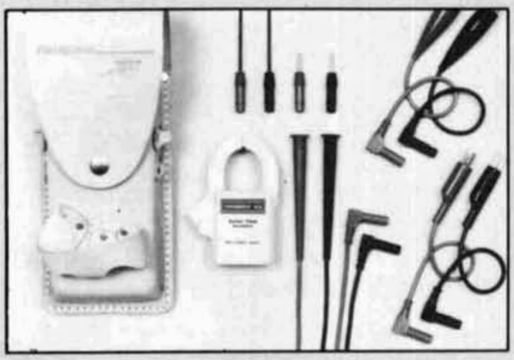
REFERENCES

1. *The ARRL Antenna Book*, 12th Edition, American Radio Relay League, 1970.
2. B. Sykes, G2HCG, "The Enhancement of HF Signals by Polarization Control," *Communications Quarterly*, Fall 1990, page 23.
3. C.J. Michaels, W7XC, "Some Reflections on Vertical Antennas," *QST*, July 1987, page 15.

PRODUCT INFORMATION

New DMM Accessories

Fieldpiece Instruments has a new line of multimeter accessories. The accessories, which include a modular test-lead design, an AC current clamp, and a leather carrying case for the tool belt, are designed to fit all Fieldpiece meters, plus the Fluke 70 Series and the Beckman 220/150 series meters.



The test-lead probe "handle" houses a standard female banana jack into which all Fieldpiece accessories (including replaceable probe tips) can be plugged. The probe tips come in two versions: short (1/2 inch), and insulated long (3 inch). Large and small alligator clips with 9-inch pigtails and sleeved banana plugs fit either directly into a meter or into the test-lead probe handle converting the test lead into a ground clip with a 4-foot lead. Two deluxe test leads can be connected to double the length of the lead.

Kit model ADK10 includes a pair (red and black) of each of the following: 42-inch deluxe test leads with female banana jack in the "handles" and right-angle sleeved banana plugs; small and large alligator clips with pigtails and sleeved banana; short (1/2 inch) and long (3 inch) insulated probe tips with integral sleeved banana plugs; and a plastic case.

The model ACH current clamp head, used with a pair of deluxe test leads, lets you use professional grade meters with "Fluke" style jacks and ACV ranges to measure up to 300 amps of AC current without breaking the circuit. The ACH plugs directly onto the Fieldpiece HS20 "Stick" series DMM.

Fieldpiece's model ALC1 full-grain leather case holds a meter and accessories. It has a leather loop on the back for a tool belt. Leather loops riveted to the front of the case hold two extra probe tips (long or short) and the ACH current clamp head. The case will hold the HS20 "Stick" series meter (with or without the ACH current clamp head attached), the Fluke 70 series (without the boot), or the Beckman 220/150 (without the tilt stand) with test leads connected and wrapped the meter. The ALC1 leather case is also available as a kit (without the multimeter) containing deluxe test leads, three small probe tips, one long probe tip, and the ACH current clamp head (model ALCK8).

For more information write to Fieldpiece Instruments, Inc. at 8322B Artesia Blvd., Buena Park, California 90621 or call (714)992-1239.

Carl Luetzelschwab, *K9LA*
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ANTENNA ANGLE OF RADIATION CONSIDERATIONS

Part 2: Performance Comparisons of quads and Yagis mounted at low heights

In the last issue of *Communications Quarterly*,[1] I discussed why antennas at the same height can have different angles of radiation. I then focused on a two-element quad and a two-element Yagi tuned to the same broadband-performance goal. These results showed that the quad has a 2.5-degree lower angle of radiation when both are mounted at low height (0.432 wave-length). The question I posed was whether the lower angle mattered in the real world. Here are my on-the-air comparisons of the two-element quad and the two-element Yagi at 30 feet on 20 meters and 15 feet on 10 meters.

Site considerations

On the surface, an A-B comparison of two antennas is deceptively simple; you put them up and see which one gives stronger signals.

Unfortunately, several key items need to be considered in order to achieve *valid* results. In addition to the obvious factors (equal length feed lines of equal loss, equal impedances so identical power is delivered to each antenna, a balun on each antenna to preserve the true radiation pattern, a calibrated S-meter, and so on), the site should be void of obstructions, of flat terrain, and of similar composition under both antennas. There's one other critical factor that came out of the 20-meter results. I will discuss it briefly.

Twenty-meter antenna construction, site, and preliminaries

The construction of the two antennas was very straightforward, as were the designs they were based on (see Part 1).

I used the TAPER program in K4VX's YAGINEC (see **References 2** and **3**) to determine the actual Yagi element lengths for the tubing I had available. The feedpoint resistance of the Yagi was around 25 ohms, so the driven element was shortened and a hairpin match was used to get to 50 ohms.

The quad was a square type fed at the bottom center. Its feedpoint impedance came out to about 100 + j10 ohms, so I used a quarter-wave of RG-59 coaxial cable to get to 50 ohms.

Both antennas used an eight-turn choke balun and were fed with equal lengths of RG-8 coax.

Figure 1 gives a top view of the antennas. The antennas are shown pointed toward Australia. Note that the house sits between the two towers and that the Yagi is looking "through" the quad.

Before making signal comparisons, I calibrated the S-meter of my TS-180 so signal differences could be measured accurately. The calibration indicated that an S-unit was generally 6 dB, except for weak signals when an S-unit was more on the order of 3 dB (which was typical of several receivers I measured). This is good information to know when making antenna comparisons.

If the signals from the two antennas were so weak that they didn't move the S-meter needle, I made an aural judgement. Most people can discern a difference in signal strength beginning at about 3 dB. I'll admit that this isn't very scientific, but I believe it's adequate for my purposes. Remember I'm

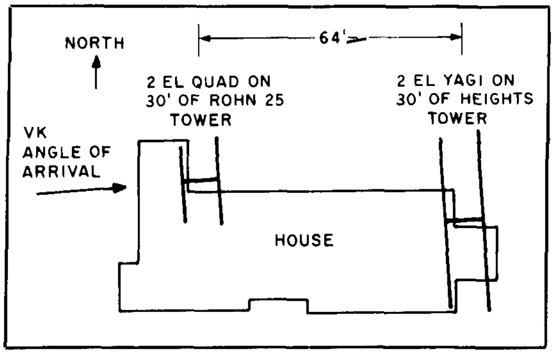


Figure 1. K9LA site, 20 meters (1 inch = 30 feet).

trying to determine if the 2.5 degree lower angle of the quad results in fewer hops which, in most cases, should be at least several dB greater than my measurement uncertainty. Ten dB is one common value I've seen in other literature, but this depends on several factors.[4]

Twenty-meter results

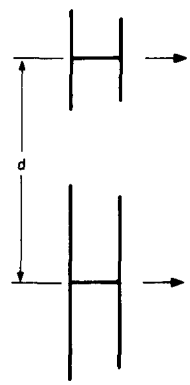
I evaluated the quad and Yagi on four paths: Europe short path (4000 miles to the northeast); Caribbean, YV, short path (3000 miles to the southeast); Asia, JA/BV/HL, short path (7000 miles to the northwest); and Australia, VK, short path (9500 miles to the west). Emphasizing the longer distances, I evaluated a total of 147 stations during the 1990 CQ World-Wide Phone DX Contest and for two weeks thereafter. I didn't have an opportunity to evaluate any long path stations.

After completing my on-the-air comparisons, I did a brief analysis of the results. For the most part, the quad and Yagi ran neck-and-neck to the northwest (Asia) and to the southeast (Caribbean). The Yagi had a definite advantage of a couple dB to the northeast (Europe), while the quad had a definite advantage of a couple dB to the west (Australia).* These results were contradictory, and indicated either a site or antenna interaction problem. Although the house lies between the two antennas, and may be part of the problem, I felt that antenna interaction was the major contributor.

Antenna interaction

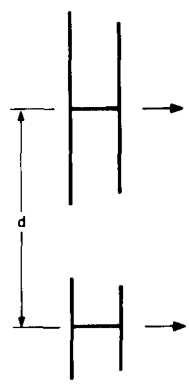
In 1967, Lew McCoy[5], W1ICP wrote an interesting article about some antenna stacking experiments he had performed. His basic premise was that if antenna A's VSWR didn't change when antenna B was brought near, then there was no interaction between two antennas. In a later article[6], he did some modeling on MN, and discovered significant interaction—even though VSWRs were unaffected. Based on these two articles, I modeled my two-element quad and two-element Yagi in various orientations: side-by-side (a line between the two antennas is perpendicular to the direction in which they are pointed), the quad looking through the Yagi (both pointed in the same direction with the Yagi in front), and the Yagi looking through the quad (again, both pointed in the same direction, but with the quad in front). The results of this exercise are shown in **Figures 2A** through **2C**.**

QUAD WITH YAGI OFF TO THE SIDE



d in wavelengths	Quad gain at peak of main lobe
1.0	8.60 dBd
1.5	8.29 dBd
2.0	8.37 dBd
no Yagi	8.36 dBd

YAGI WITH QUAD OFF TO THE SIDE



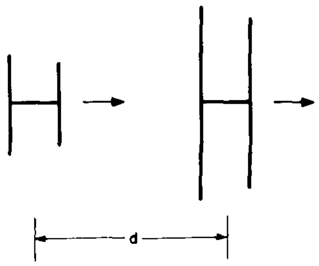
d in wavelengths	Yagi gain at peak of main lobe
1.0	8.02 dBd
1.5	7.97 dBd
2.0	7.85 dBd
no quad	7.90 dBd

Figure 2A. Side-by-side interaction.

*It's relatively easy to measure small dB differences on local signals, but to do so with any degree of certainty on long-distance signals experiencing QSB is a difficult task, at best. This is why I state my results as "a couple of dB" rather than giving a specific numerical amount.

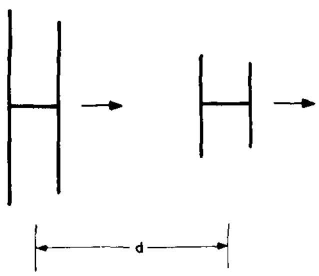
**These results, in which the gain of the antenna in back is degraded by the antenna in front, may be unique to the physical dimensions and spacings of the antennas that I modeled. I believe that the gain of an antenna could be enhanced by an antenna in front if the dimensions of front antenna and the spacing were appropriate. I also briefly looked at different terminations on the antenna not being fed. The variation with an open, short, or 50-ohm load was insignificant.

QUAD LOOKING THROUGH YAGI



d in wavelengths	Quad gain at peak of main lobe
1.0	6.17 dBd
1.5	7.03 dBd
2.0	7.47 dBd
no Yagi	8.36 dBd

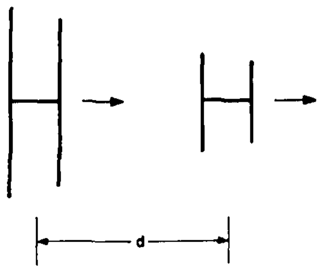
QUAD WITH YAGI IN BACK



d in wavelengths	Quad gain at peak of main lobe
1.0	8.49 dBd
1.5	8.32 dBd
no Yagi	8.36 dBd

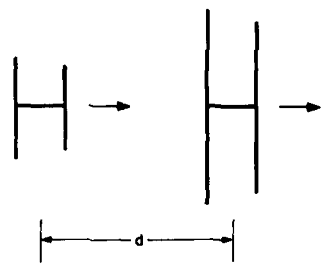
Figure 2B. Quad interaction.

YAGI LOOKING THROUGH QUAD



d in wavelengths	Yagi gain at peak of main lobe
1.0	6.09 dBd
1.5	6.46 dBd
2.0	6.91 dBd
no quad	7.90 dBd

YAGI WITH QUAD IN BACK



d in wavelengths	Yagi gain at peak of main lobe
1.0	7.82 dBd
1.5	7.84 dBd
no quad	7.90 dBd

Figure 2C. Yagi interaction.

It's obvious from these results that there is significant interaction to the back antenna when it's looking through the front antenna—even when the spacing is 2 wavelengths. I have only presented gain data. In reality, the entire pattern is changed (affecting gain, front-to-back (F/B) ratio, θe, θh, and angle of radiation). The theoretical results at 1 wavelength are in general agreement with the results I observed. There isn't much interaction when the antennas are side-by-side (even at close tip-to-tip spacing). There also isn't much interaction to the front antenna with the other antenna in back (presumably due to the F/B ratio of the front antenna).

Based on these results and the fact that the Yagi was looking almost directly through the quad when pointed to Australia (**Figure 1**), it was with great reluctance and disappointment that I declared my 20-meter results to the west invalid. Even the results to the northeast, northwest, and southeast were suspect. Although neither antenna was looking directly through the other antenna in these directions, I felt it better to invalidate these results, too, rather than report questionable data.

It became obvious that I would achieve valid results only if I could evaluate directions in which the two antennas were side-by-side and separated by at least 1.5 wavelengths. Ten meters was the only band that allowed me to do this at my location.

Ten-meter antenna construction, site, and preliminaries

I scaled the 20-meter designs down to 10 meters and reoptimized for the new length-to-diameter ratios. The construction and impedance matching of the 10-meter antennas was identical to the 20-meter versions. Equal length RG-8 feedlines fed each antenna, and each had an eight-turn choke balun.

Both antennas were supported on three-legged 15-foot wooden tower frames that could be moved easily to maintain side-by-side orientation on the paths evaluated. Rope guys were used to stabilize the tower/antenna assemblies, as I began my evaluation during the spring thunderstorm season. **Photo A** shows the two tower/antenna assemblies.

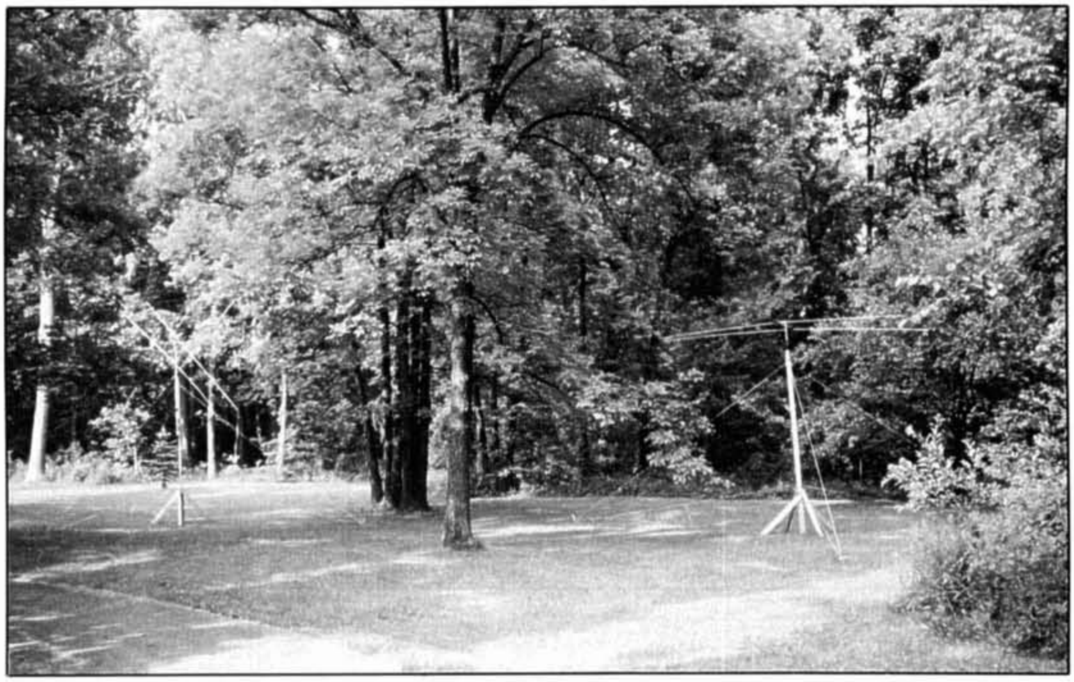


Photo A. Ten-meter quad and Yagi tower/antenna assemblies oriented toward Australia.

Figure 3 shows the site from overhead; the tower assemblies are pointed toward the southeast (South America, short path and Japan, long path).

Center-to-center tower separation was 1.5 wavelengths (50 feet) on all paths. The paths evaluated were restricted to those that put the house either off the back or off the side of the two antennas, with the distance to the house at least 2 wavelengths. I checked the TS-180 S-meter on 10 meters and obtained results similar to those obtained on 20 meters.

With the tower/antenna assemblies pointed southeast, I took a relative gain measurement on the two antennas with a local station 1.4 miles away. The results showed the quad to have about a 1-dB advantage as measured with a 0.5-dB step attenuator. This indicated that the antennas were performing as expected, with no significant site or antenna interaction problems. This tended to confirm that my quad did have a 2.5-degree lower angle than my Yagi (more on this later).

Ten-meter results

During the month of April 1991, the tower/antenna assemblies were pointed southeast to evaluate the antennas on the 4500-mile Brazilian (PY) short path. I monitored several Brazilian beacons: PT7BCN on 28.213 MHz, PT8AA on 28.218 MHz, and PY2AMI on 28.225 MHz. I made my measurements when the band was just opening, when the band was open, and when the band was just closing. Of the 28 days I

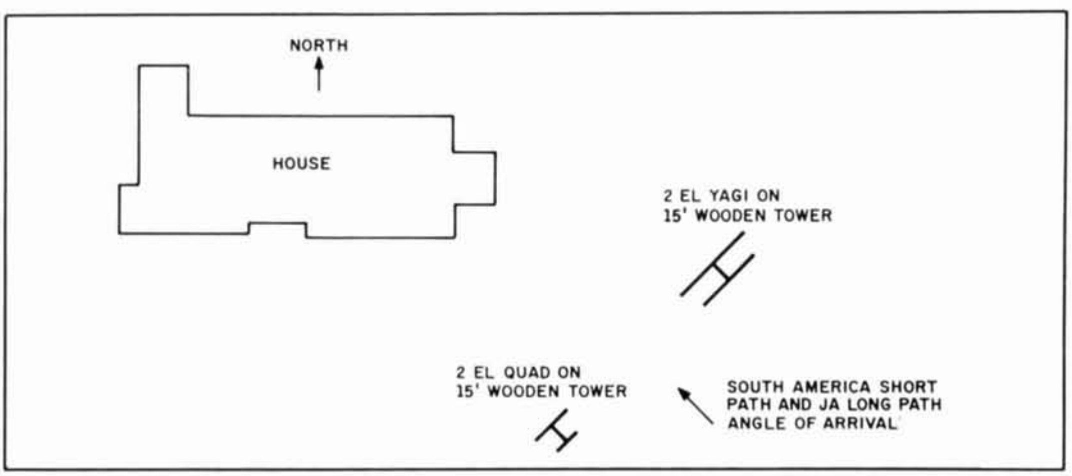


Figure 3. K9LA site, 10 meters (1 inch = 30 feet).

was able to listen, I heard the beacons on 26 days. As best as I could measure, the signal strengths from the quad and Yagi were identical. At no time did the quad open the band earlier or close it later than the Yagi.

During the month of May, the tower/antenna assemblies were pointed to the east to evaluate the antennas on the 8500 mile South African short path. Once again, I monitored several beacons: ZS5VHF on 28.202 MHz, Z21ANB on 28.250 MHz, and ZS1LA on 28.274 MHz. As before, the evaluations were done when the band was just opening, when the band was open, and when the band was just closing. Of the 30 days I was able to listen, I heard the beacons on 25 days. The signal strengths from the quad and Yagi were again identical when the band was opening, when it was open, and when it was closing.

During the month of June 1991, the tower/antenna assemblies were pointed west to evaluate the antennas on the 9500-mile Australian short path. I monitored two beacons: VK5WI on 28.259 MHz and VK2RSY on 28.261 MHz. Of the 27 days I was able to listen, I heard the beacons on 20 days. Again, the signal strengths from the quad and Yagi were identical throughout the listening periods.

Finally, during the first half of July 1991, the tower/antenna assemblies were oriented back toward the southeast to evaluate the antennas on the 18,000-mile Japanese long path.* If the quad was going to show superiority over the Yagi due to its lower angle, this was the path it would do it on. I maintained a daily schedule with JH3DPB. Of the 16 days we ran the schedule, the long path was open 4 days (there were several solar flares during this period which adversely affected the path). My measurements indicated that the signal strengths from the quad and Yagi were identical. At no time did the quad open the long path earlier or close it later than the Yagi.

Discussion of the concept of angle of radiation

Before I draw any conclusions from the 10-meter data, I'd like to discuss what I really compared in the 10-meter evaluation (and what I would have compared in the 20-meter results, barring antenna interaction). Up until I began analyzing the 10-meter results, I believed that I was evaluating whether the lower angle of radiation of the quad would result in less hops. As I thought about this, and looked at the mechanics of vertical patterns and propagation, I realized that I had fallen into the same trap that has snared many others. Here's why I believe that the concept that "a lower angle of radiation results in less hops" is wrong. I'll also show what I really compared in the 10-meter evaluation.

Refer to **Figure 2** of Part 1. It showed the angle of radiation versus the height of a square loop and dipole over perfect ground. At 0.25 wavelengths, the dipole's angle of radiation is 90 degrees, while the loop's is 49 degrees. From this data it's natural to conclude (as I did initially) that the loop is better at low height than the dipole, because it has a lower angle of radiation. This sounds very convincing. However, as I stated previously, I believe there's a major misconception inherent in this conclusion.

When one makes this conclusion, he assumes that the energy at the angle of radiation is the energy which results in the QSO. This is equivalent to saying that the angle of radiation determines the number of hops between stations—a supposition which is incorrect. The ionosphere and the distance determine the number of hops between stations. (This can be analyzed using ray tracing methods similar to those found on page 22-4 of *The 1988 ARRL Handbook*, done with PROPHET. IONCAP is another program which gives the same type of result.) What's important is the energy at the elevation angle dictated by the ionosphere and distance. In general, for long distance communications, this means low elevation angles, regardless of the height of the antenna.[7] Thus, the true comparison of two antennas is not their angle of radiation difference, but the magnitude differences of their patterns at these low elevation angles. Let's look at this in more detail for the loop and the dipole.

Figure 4 shows the vertical pattern of the loop and dipole at 0.25 and 0.75 wavelengths. As expected, they are quite similar at the high height. At low height, it's easy to see the difference in the angles of radiation. It's also obvious that there's still energy radiated at lower angles for both the loop and the dipole. This is because the vertical radiation patterns of horizontally polarized antennas at low height are very bulbous—this means the peak, which is defined as the angle of radiation, is very broad.

Note the relationship between the loop and dipole at low elevation angles at the high height, and this relationship at similar low

*My limited experience with this 10-meter long path indicates it occurs from about 1130Z to 1215Z during mid-June to mid-July here in the midwest. *I welcome any reports from other mid-westerners who have experience with this long path and/or the similar 10-meter long path to western Australia.* Also see page 20 of the January 1989 issue and page 23 of the September 1989 of *The DX Magazine* for a west-coast view of 10-meter long path.

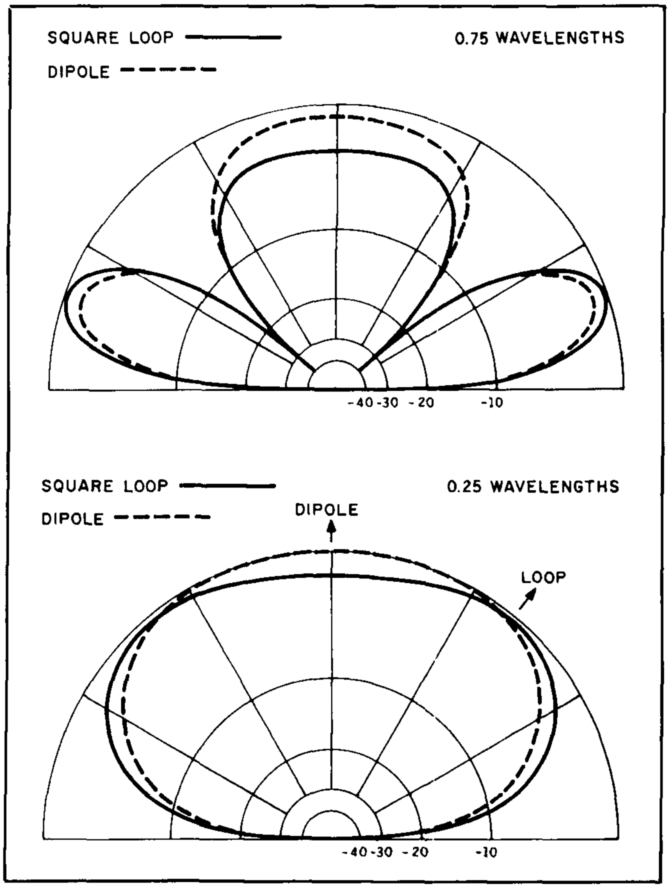


Figure 4. Vertical pattern of loop and dipole at 0.25 and 0.75 wavelength.

elevation angles at the low height; there's not much difference. **Table 1** gives the values (from MN) of the patterns at a 5-degree angle. (I chose 5 degrees as typical from the data in **Reference 7**, but this isn't critical in the following analysis. I could have used 1, 10, or 15 degrees with equal results.)

At the high height, the loop has 1.36 dB gain over the dipole at the 5-degree elevation angle. At the low height, the loop now has 2.02 dB gain over the dipole at the 5-degree elevation angle. Thus, not only is the loop a better performer than the dipole at high height, it's an additional 0.66 dB better at low height (due to the dipole's increased radiation resistance at 0.25 wavelengths). However, this isn't a significant amount. And, I believe it's a more accurate picture of what's happening than making a statement involving the angles of radiation.

*The difference is less than the loop/dipole case because the radiation resistance variation with height of the quad and Yagi is less than the loop and the dipole.

It should be obvious that the differences seen by Parrott (two-element monoband 20-meter quad at 50 feet opened the band earlier and closed it *later than a four-element trapped tribander at 50 feet—see **Reference 3 of* Part 1) and by Fitz (four-element monoband 20-meter quad at 80 feet 1 to 2 dB better than a four-element monoband 20-meter Yagi at 80 feet—see **Reference** 4 of Part 1) were due to gain differences and not angle of radiation differences. Also note that antenna interaction was not addressed in either article.

This analysis of the loop and dipole is the extreme case. If we go through the same analysis for my two-element quad and two-element Yagi, we'll see that the quad has 0.85 dB gain over the Yagi at the high height at low elevation angles, and has 0.90 dB gain over the Yagi at the low height at low elevation angles. (The difference is again due to radiation resistance variation.*) To summarize, my quad maintains the same gain over the Yagi at low elevation angles (which is the free-space gain) *regardless of height.*

So what does the angle of radiation tell us? To reiterate, it doesn't tell us anything about the number of hops to a DX station. All it indicates is that my two-element quad achieves its free-space gain over my two-element Yagi by compressing the vertical pattern. As a result, the quad will show this free-space gain (and no more) at low elevation angles (the angles important for DX) at any height. Again, this simply shows how two functions interact when multiplied together. It should now also be clear why I said that the 1 dB gain I measured to a local station tends to confirm that the quad had a lower angle of radiation. The lower angle exists because the quad has free-space gain over the Yagi.

The preceding information shows that the 10-meter evaluation simply compared the 1 dB or so inherent free-space gain that my quad has over my Yagi. Thus the question, "Does the lower angle matter in the real world," should have been "Does the 1 dB or so gain matter in the real world?"

Conclusions

My work has led me make two conclusions. First, based on the above theoretical analysis, my quad's performance relative to my Yagi will be the same at low height as at high height. The quad will be a better performer at any height only to the extent of its free-space gain over the Yagi.** I believe this conclusion is also applicable to the loop and dipole, in spite of claims to the contrary.

Second, based on experimental results, my 10-meter data did not indicate that the 1 dB or so inherent free-space gain of my quad over my Yagi made a difference. The difference is there, of course, but it's tough to measure (or even notice) a 1-dB difference on long-distance signals. Perhaps I would have noticed it if I had continued my evaluation for many months under varying seasonal and solar conditions.

A final comment

There's one other factor that's worthy of mention. Although the signal strength re-

	Dipole	Loop	Loop with Respect to Dipole
0.75 wavelengths	– 0.04 dBi at 5 degrees	– 1.40 dBi at 5 degrees	1.36 dB
0.25 wavelength	– 9.87 dBi at 5 degrees	– 7.85 dBi at 5 degrees	2.02 dB

Table 1. Vertical pattern values of *Figure 4*.

sults aren't valid, I did notice on 20 meters that the quad was 3 to 4 dB quieter than the Yagi about 20 percent of the time. (I didn't notice this effect to any great extent on 10 meters.) If a signal was at the noise level on the Yagi, it was perfectly readable on the quad. This could be interpreted as opening the band earlier or closing the band later. This 3 to 4 dB noise advantage of the quad (which may be due to the reduced high angle response of the quad) could be much more significant than the 1 dB or so gain.

Acknowledgements

I would like to thank K9MK/5 and KC9LA for reviewing and making suggestions to the rough draft of this article. I'd also like to thank the local gang (AF9Y, N9GWG, KR9U, K9UWA, and KC9LA) for their help in making measurements and building the antennas. Finally, a big thanks to JH3DPB for maintaining our daily schedule to evaluate the antennas on the l0-meter long path. ■

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PRODUCT INFORMATION

New NE32400 GaAs Hetero Junction FET

California Eastern Labs announces the availability of NEC's new NE32400 GaAs Hetero Junction FET. Designed specifically for commercial C to Ka band applications, the NE32400 chip uses a new pseudomorphic design to help reduce noise figures.

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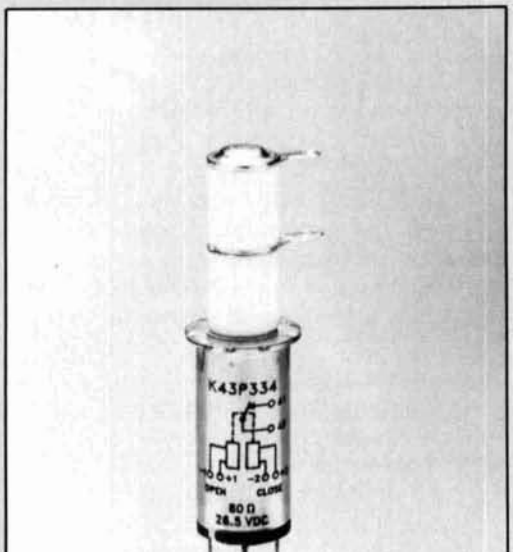
High gain: 11.0 dB typical (10.0 dB min) at 12 GHz

High reliability: $3.7 \times 10E9$ hours at $T_{ch} = 100$ C

For more information on the NE32400 FET write California Eastern Laboratories, Inc., 4500 Patrick Henry Drive, Santa Clara, California 95056-0964 or call (408)988-3500.



High Power Latching Relay

Kilovac Corporation introduces a new high-voltage latching relay. The K43P is a small, lightweight vacuum relay that can carry 24 amperes continuously at kVDC.

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For more information about the K43P write Kilovac Corporation, P.O. Box 4422, Santa Barbara, California 93140 or call (805)684-4560.

Michael E. Gruchalla, P.E.
4816 Palo Duro Avenue, NE
Albuquerque, New Mexico 87110

OPTIMIZING AMPLIFIER GAIN-BANDWIDTH PRODUCT

An analysis of multistage-amplifier behavior

It often seems that no matter what the gain or bandwidth of an amplifier, I find myself needing a little more of one or the other, or both, in my applications. Unfortunately, when I try to add another amplifier stage to improve one of these parameters, the other suffers. For example, I may add a second amplifier stage to increase gain only to find that now I don't have enough bandwidth. With multistage amplifiers, there's an optimum gain for each stage that will maximize the overall gain-bandwidth product of the total amplifier system. But it's quite possible that you'll find this optimum-stage gain a bit surprising.

In the discussion which follows, I show the computations for optimum stage gain. Although this result is relatively well known and the derivation straightforward, it's somewhat difficult to find references describing the work in any detail.[1] In any event, I hope you'll find the information here of interest and of some value in your amplifier design work.

Gain versus bandwidth

Many amplifiers, like operational amplifiers and some RF amplifiers, have a constant gain-bandwidth product. This means that if you increase the gain, you have less bandwidth, and vice versa.

An operational amplifier has this characteristic. The industry-standard 741 operational amplifier has a nominal bandwidth of 10 Hz and a nominal voltage gain of about 100,000—a gain-bandwidth product of 1 MHz. If you lower the gain to 1000 by using a feedback network, the bandwidth will increase to 1 kHz. Suppose that you were to use two 741 amplifiers in a two-stage amplifier circuit with each stage having a gain of 316. The bandwidth of each amplifier would be about 3.2 kHz.

The total gain of the pair of amplifiers would be 100,000 and the combined bandwidth about 2.1 kHz. The gain-bandwidth product will have been increased to 210 MHz—a factor of more than 200 over that of a single 741. Suppose four amplifiers, each with a gain of 17.78 and a bandwidth of 56.2 kHz, are used. The total gain will still be 100,000, but the bandwidth will be 24.4 kHz. The gain-bandwidth product is 2.44 GHz—an increase of more than a factor of 2000 over that of a single amplifier.

You can continue this process, but you'll reach a point where a further reduction in amplifier gain and an increase in the number of stages will decrease the gain-bandwidth product. For example, suppose that the new amplifier gain selected is unity. The bandwidth of a single stage would be 1 MHz. No matter how many of these amplifier stages you use, the gain will remain at unity; however, the bandwidth will be reduced as you add more stages. So the best gain-bandwidth product you can obtain with a stage gain of unity is 1 MHz, using a single stage. That's considerably lower than the results obtained with four stages. Because a reduction in gain per stage from 100,000 with multiple stages results in an increase in gain-bandwidth product, but using a gain of unity results in a lower gain-bandwidth than that of the four-stage case, you'd expect that there's an optimum gain that will provide the best possible gain-bandwidth product.

Amplifier bandwidth and midband gain

Assume an amplifier with a midband voltage gain g_0 and a lower frequency response to DC, or near DC. Let the upper cutoff frequency, the frequency where the gain is 3 dB below the midband value, be f_c. The amplifier bandwidth is defined as the bandwidth between the lower and upper cutoff frequencies. Because the lower cutoff frequency is far below the upper frequency, the amplifier bandwidth (BW) is essentially equal to the upper cutoff frequency, BW $\approx f_c$. Also, let this amplifier have a constant gain-bandwidth product $g_0 \times f_c$ that is a constant k for any gain. The amplifier voltage gain as a function of frequency is given by:

$$A_v(f) = \frac{g_0}{[1 + j(f/f_c)]} \qquad (1)$$

The term g_0 is the midband gain of the amplifier and f_c is the upper cutoff frequency. In the open-loop 741 case, g_0 is 100,000 and f_c is 10 Hz.

Because you're really only interested in the magnitude of the voltage gain, there's no reason to keep the complex form of **Equation 1**. The magnitude of the voltage gain is given by:

$$|A_v(f)| = \frac{g_0}{[1 + (f/f_c)^2]^{1/2}} \qquad (2)$$

Now, if some number of these amplifiers were connected in a multistage amplifier circuit, the midband gain would be increased and the bandwidth reduced as shown in the previous 741 example. Let the number of amplifier stages be n. The magnitude of the voltage gain is then given by:

$$|A_v(f,n)| = \frac{g_0^{\,n}}{[1 + (f/f_c)^2]^{n/2}} \qquad (3)$$

Generally, the overall voltage gain is the gain term in which you are interested. Let the midband gain of the *total* amplifier be G_0. Then $G_0 = g_0^{\,n}$.

$$|A_v(f,n)| = \frac{G_0}{[1 + (f/f_c)^2]^{n/2}} \qquad (4)$$

Upper cutoff frequency

The upper cutoff frequency of the multistage amplifier is that frequency where the total amplifier gain is 3 dB below the midband value. Let the upper cutoff frequency of the total amplifier system be F_c. At that upper cutoff frequency, the voltage gain will be reduced by a factor of $\sqrt{2}$ below the midband value. Thus:

$$|A_v(F_c,n)| = \frac{G_0}{[1 + (F_c/f_c)^2]^{n/2}} = G_0/\sqrt{2}$$

and

$$G_0/\sqrt{2} = \frac{G_0}{[1 + (F_c/f_c)^2]^{n/2}} \qquad (5)$$

You can easily solve **Equation 5** for F_c.

$$F_c = f_c \times \sqrt{2^{1/n} - 1} \qquad (6)$$

Equation 6 now provides the overall bandwidth (actually the upper cutoff frequency) of a total multistage amplifier of n similar stages. For instance, in the previous example using the 741, if a 741 amplifier were designed for a voltage gain of 17.78, the bandwidth would be 56.2 kHz. If four such amplifiers were used in a multistage amplifier, the overall gain would be $(17.78)^4$, or about 100,000. The bandwidth of the total amplifier is given by **Equation 6** as 24.4 kHz.

A multistage amplifier design example

Suppose you wish to build an amplifier system using one or more amplifier stages with a gain and frequency performance described by **Equation 2**. You want the overall amplifier to have a gain G_0 and the maximum possible bandwidth. The gain-bandwidth product of a single stage is $g_0 \times f_c$. The single-stage gain-bandwidth product is k and constant for any gain.

The 741 example showed that a much wider bandwidth may be obtained for any given gain by using several stages. You may use n stages to provide the total desired gain G_0. The gain of each stage, g_0, must then be $G_0^{\frac{1}{n}}$. So,

$$k = g_0 \times f_c \qquad (7)$$

$$G_0 = g_0^{\,n} \quad (8)$$

For your overall amplifier, you've specified a specific gain and the maximum possible bandwidth. In other words, you want the maximum possible gain-bandwidth product. The total amplifier gain-bandwidth product is simply $G_0 \times F_c$. You must find this product in terms of the individual stage parameters. You've found G_0 in terms of g_0 in **Equation 8**. But G_0 is a constant—the overall gain that you want. You also know F_c in terms of f_c by **Equation 6**. Then, the gain-bandwidth product of your overall amplifier in terms of the overall gain and the bandwidth of the individual stage is given by:

$$G_0 \times F_c = G_0 \times f_c \times \sqrt{2^{1/n} - 1} \quad (9)$$

It's convenient to put **Equation 9** into a somewhat different form. The radical term is a bit troublesome, but you can make an approximation to improve it. The natural logarithm of a number x, Ln x, may be expressed in an infinite series by **Equation 10**:[2]

$$Ln\,x = (x-1) - \frac{1}{2}(x-1)^2 + \frac{1}{3}(x-1)^3 - \cdots \quad (10)$$

$$Ln\,2^{1/n} = (2^{1/n}-1) - \frac{1}{2}(2^{1/n}-1)^2 + \frac{1}{3}(2^{1/n}-1)^3 - \cdots \quad (11)$$

If you choose a number of stages for n greater than about 4, the first term of **Equation 11** dominates and the approximation of **Equation 12** is accurate to better than 10 percent. The accuracy improves with more stages.

$$(2^{1/n} - 1) \approx Ln\,2^{1/n} = \frac{1}{n} Ln\,2 \quad (12)$$

Equation 9 may then be written as:

$$G_0 \times F_c = G_0 \times f_c \times \sqrt{\frac{1}{n} Ln\,2} \quad (13)$$

Because you're trying to find the best gain for each stage, you must express f_c and n in terms of g_0. Find this using **Equations 7** and **8**, respectively.

$$f_c = \frac{k}{g_0} \quad (14)$$

$$n = \frac{Ln\,G_0}{Ln\,g_0} \quad (15)$$

By substituting all that information into **Equation 13**, you'll obtain the results of **Equation 17**.

$$G_0 \times F_c = G_0 \left(\frac{k}{g_0}\right)\left[\left(\frac{Ln\,g_0}{Ln\,G_0}\right)(Ln\,2)\right]^{1/2} \quad (16)$$

$$= G_0 \times k \left(\frac{Ln\,2}{Ln\,G_0}\right)^{1/2} \left(\frac{Ln\,g_0}{g_0^2}\right)^{1/2} \quad (17)$$

Gain-bandwidth product

Equation 17 now gives the overall amplifier gain-bandwidth product in terms of the various constants and g_0. The term g_0 is the only variable in **Equation 17**. It might appear that G_0 is a variable because it may be expressed in terms of g_0. However, G_0 is the overall total gain desired in the multistage amplifier and is therefore a defined constant. Actually, g_0 is a function of G_0. If you examine **Equation 17**, you'll see that if g_0 is chosen as unity, the gain-bandwidth product will be zero (a unity-stage gain implies an infinite number of stages needed). Similarly, if a very large g_0 is chosen, the gain-bandwidth product also approaches zero (try a gain of 1,000,000). However, if you choose a value of n such as 10, the gain-bandwidth product will be greater than with either unity gain or very high gain. This implies that there is a specific gain which will provide the highest possible gain-bandwidth product. Let that optimum stage gain be g_{opt}. This optimum gain is relatively easy to compute using calculus.[3] However, because not everyone is interested in all the mathematical details, that work is relegated to a sidebar.

After going through the math, the optimum gain g_{opt} and the optimum number of stages *n* are given in **Equations 18** and **19**.

$$g_{opt} = e^{1/2} = 1.649 = 4.34dB \quad (18)$$

$$n = 2\,Ln\,G_0 \quad (19)$$

$$F_c = f_c \sqrt{\frac{1}{n} Ln\,2}$$

[from Equations (6) and (12)] **(20)**

Equations 18 and **19** give the optimum stage gain for maximizing gain-bandwidth product and the number of stages needed for any desired gain of a multistage amplifier.

There's a very important but subtle point you should observe in **Equation 18**. Note that the optimum gain of the individual stage is totally independent of everything! It's not a function of the overall gain desired, or the stage bandwidth, or the stage gain-bandwidth product, or anything else except the constant e—the base of the natural logarithm = 2.718. Therefore, if you wish to build an amplifier with the best possible gain-bandwidth product, you'll need a multistage amplifier with individual stage gains of about 1.65—no matter what overall gain you may desire.

Consider the 741 again

If you want to use 741s to build an amplifier with a gain of 100,000 and maximum bandwidth, **Equation 18** shows that the individual stage gain should be 1.649. With that gain, a 741 would provide a bandwidth of about 607 kHz. **Equation 19** indicates that a total of 23 stages would be required. Finally, **Equation 20** shows that the overall bandwidth would be 105 kHz. The overall gain-bandwidth product would be 10.5 GHz. This is considerably better than the 1 MHz gain-bandwidth product of a single 741, although the same voltage gain of 100,000 is provided. I know this may seem a bit peculiar, but just try to obtain a 100-kHz bandwidth with a gain of 100,000 using 741s in any other configuration.

RF amplifier

Even though the 741 example is accurate, it's not a particularly practical or a commonly needed configuration. Consider an RF amplifier. Suppose you have a basic 20-dB gain (voltage gain of 10) amplifier with a 1-GHz bandwidth, 10 Ghz gain-bandwidth product, with the gain adjustable down from 20 dB such that the gain-bandwidth product remains constant. Say you wish to build a multistage amplifier with a gain of 40 dB (voltage gain of 100), but with the best possible bandwidth. If you simply use two of the amplifiers in a two-stage amplifier configuration, you'll have a 40-dB gain and a bandwidth of 644 MHz. However, if you were to use nine stages (nearest integer number to the results of **Equation 19**), each designed for a voltage gain of 1.668, or about 4.44 dB, the overall voltage gain would also be 100. But the bandwidth would be 1.7 GHz, almost a factor of three greater than a simple two-stage unit.

Even a simple 20-dB amplifier can be improved. A single stage of the amplifier above provides a bandwidth of 1 GHz. If five stages, each with a gain of about 1.585 (4.0 dB) were used in a multistage amplifier, the gain would still be 20 dB, but the bandwidth would be slightly greater than 2.3 GHz. That's more than an octave increase in bandwidth.

Conclusions

It's often desirable to optimize the gain and bandwidth of a multistage amplifier to achieve the maximum possible gain-bandwidth product. The optimum voltage gain of each stage of such an amplifier, with at least four stages or more, is $\sqrt{e}$, or approximately 1.65. That stage gain is totally independent of any of the parameters of the individual amplifier stages, or those desired of the multistage unit. Therefore, when designing a multistage amplifier, you should design each stage for a voltage gain of about 1.6 to 1.7 with sufficient stages to provide the desired gain.

You may wish to use prefabricated amplifier building blocks for the individual stages. Typical examples of such building blocks are the MAR devices from MiniCircuits, the GPD and MSA devices from Avantek, and there are numerous others. Although you typically can't modify the gain of these building blocks effectively, there's generally a selection of gain and bandwidths

CALCULATING THE OPTIMUM GAIN

$$G_0 \times F_c = G_0 \times k \left(\frac{Ln\,2}{Ln\,G_0}\right)^{1/2} \left(\frac{Ln\,g_0}{g_0^{\,2}}\right)^{1/2} \quad \textbf{(17)}$$

G_0 = Desired overall gain = constant

$n > \approx 4$

Limit $G_0 \times F_c = 0$ and Limit $G_0 \times F_c = 0$

$g_0 \to 0 \qquad\qquad g_0 \to \infty$

$$For\ 0 \le g_0 \le \infty,\ G_0 \times F_c > 0$$

Therefore, a maximum must exist. To find the maximum, take the first derivative of $G_0 \times F_c$ with respect to g_0, equate to zero, and solve for $g_0 \equiv g_{opt}$.

$$\frac{d}{dg_0}(G_0 \times F_c) =$$

$$G_0 \times k \left(\frac{Ln\,2}{Ln\,G_0}\right)^{1/2} \frac{d}{dg_0}\left(\frac{Ln\,g_0}{g_0}\right)^{1/2} \equiv 0$$

$$(Ln\,g_0)^{-1/2}(g_0 - 2g_0\,Ln\,g_0) = 0$$

$$(Ln\,g_0)^{-1/2}(1 - 2Ln\,g_0) = 0 \quad [for\ g_0 \neq 0]$$

$$1 - 2Ln\,g_0 = 0 \quad [for\ Ln\,g_0 \neq 0]$$

$$Ln\,g_0 = \frac{1}{2}$$

$$g_0 \equiv g_{opt} = e^{1/2}$$

$$G_0 = g_{opt}^{\,n} \quad \textbf{(18)}$$

$$n = \frac{Ln\,G_0}{Ln\,g_{opt}} = 2\,Ln\,G_0 \quad \textbf{(19)}$$

offered in each series of devices. The lower-gain devices usually provide the wider bandwidth. If you wish to maximize the bandwidth for any desired gain in a multistage amplifier, it's important to select the individual amplifier device of the group with the *lowest* gain, and presumably, widest bandwidth.

Although the optimum gain-bandwidth product of a multistage amplifier is obtained with a stage gain of $\sqrt{e}$, there may be other circumstances where you may prefer a higher gain in a specific stage. One situation where this is true is in low-noise designs. In these cases, it's very important that the first stage have a relatively high gain to prevent compromise of the amplifier noise figure by the noise of succeeding stages. You may also prefer higher gain in a specific stage in high-output power amplifiers. A final-stage gain of only 4 dB may be insufficient to allow the maximum output of the preceding stage to drive the final stage to its maximum output capability. In this case, a higher-gain final would be desirable. Therefore, in specific cases, a gain higher than the optimum for best gain-bandwidth product may be required. However, when such a higher gain is used, it should be understood that it's a tradeoff of gain-bandwidth product for improvement in some other parameters. ■

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1. P.M. Chirlian, *Analysis of Design of Electronic Circuits*, McGraw Hill, New York, 1965.
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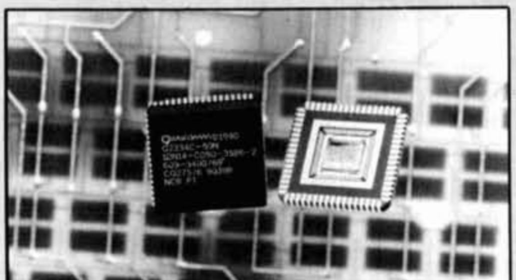
3. E.J. Johnson and F.L. Kiokemeister, *Calculus*, Allyn and Bacon, Inc., Boston, 1959.

PRODUCT INFORMATION

New 20-MHz Direct Digital Synthesizer

QUALCOMM, Inc. announces a 20-MHz dual Direct Digital Synthesizer (DDS), the Q2334M-20L, which provides two independent synthesizers on one integrated circuit for military applications. The Q2334 provides output over a wide bandwidth and generates two independent signals for completely separate circuit functions (i.e., complex signal generation, I/Q channels). The device includes two patented features: a noise reduction circuit, which lets the user specify less expensive DACs without the expected increase in spur levels; and an algorithmic sine lookup.



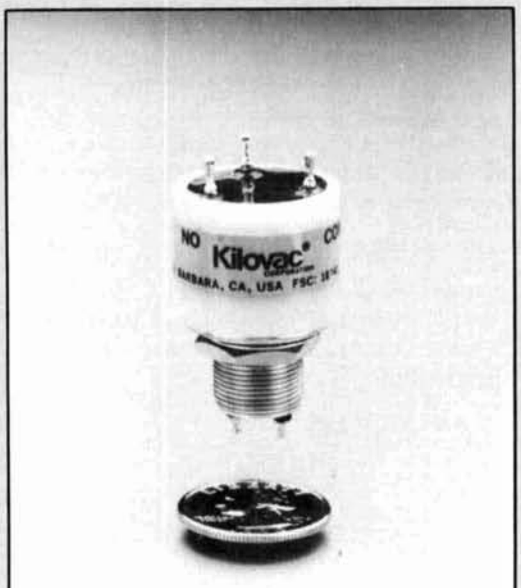
Packaged in a 68-pin hermetically sealed ceramic leaded chip carrier (CLDCC), the Q2334M-20L is screened to the requirements of MIL-STD-883, Level B techniques of Methods 5004. QUALCOMM has also released a commercial ceramic version of the 20-MHz DDS (part no. Q2334I-20L) and the 30-MHz DDS (part no. Q2334I-30L).

For further information contact QUALCOMM, Inc., VSLI Products Division, 10555 Sorrento Valley Road, San Diego, California 92121-1617.

New High Voltage Relay Available

Kilovac Corporation introduces a new high voltage relay. The HC-6 is a new pressurized gas-filled relay with tungsten and molybdenum contacts, and a rated operating voltage of 8 kV. The relay's continuous current carry is 15 amps and, under certain circumstances, it can make up to 150 amps.

For more information on the HC-6, contact Kilovac Corporation, P.O. Box 4422, Santa Barbara, California 93140.



R.P. Haviland, *W4MB*
1035 Green Acres Circle North
Daytona Beach, Florida 32119

ANTENNA-STRUCTURE INTERACTION

Modeling with MININEC

One of the problems of the real world is that antennas must be mounted on some type of a structure, like a boom, a mast, or a tower. Often, even larger structures are involved, such as an auto or an airplane.

Over the years, a number of rules of thumb have developed regarding good mounting practices. These guidelines have always been suspect, sometimes without reason, but other times as a result of flaws like poor performance in some direction, or an unusually high SWR. All too often, a nagging uncertainty remained: Is this the best that can be done?

Antenna modeling by computer provided a way to look at these antenna-to-structure interactions in considerable detail. I'd like to introduce techniques for solving amateur type antenna problems, using equipment, techniques, and the MININEC software now readily available to individuals.

I recommend that you use one of the versions of MININEC 3 (see ads in the amateur magazines). Those with change 12 incorporated are the best, because this change eliminated a restriction on angle between wires. However, any version after change 6 should be satisfactory.

There's a version of NEC available for the PC family. Unfortunately, it's relatively expensive and available only to members of The Applied Computational Electromagnetics Society. However, those interested in Yagi antenna design might want to investigate NEC for Yagis 1.0 by Brian Beezley, K6STI. This IBM-compatible program provides analysis of Yagi designs using the NEC code. You'll find advertisements for NEC for Yagis, and Brian's other programs (MN 4.0 and YO 4.0), in most amateur radio publications.

Concepts of structure modeling

When doing this type of work, it's convenient to consider that there are two kinds of structures. The first can be called "antenna-like" structures. These include many items commonly found close to the antenna. Booms, towers, and guys are structures of this class. Because their dimensions aren't greatly different than antenna dimensions, they are easy to model using essentially the same rules followed for modeling the antennas themselves.

The second class of structure doesn't really have a good name. Relatively large areas of metal are involved, usually curved surfaces like those found in the body of an auto or the skin of an aircraft. Because of the large amount of area involved, the standard antenna modeling techniques using straight, thin wires can't be used directly. Instead, the structure must be approximated by a model composed of the thin, straight wires which the antenna models can handle. The usual name for these are "wire-frame" models. You can find examples of these models in most electronic and computer journals—

often in ads for drafting software. The precomputer name would have been "Tinker-Toy" models, after the child's toy.

Always remember that the performance output of a computer antenna model is only an approximation of real antenna performance. This is true for most antenna calculations, because the equations representing conditions are so complex that they can only be solved by approximation. The approximations are even less accurate when large areas are analyzed. This doesn't mean you must throw out the model results, however. These results are useful guides to design and operation. At least your starting point will be in the right ballpark, if not exactly at the right plate. Use the model results for what they are: guides to the real situation which are accurate enough to give relative performance and a good starting point for final adjustment.

Antenna-like structures

The class of structures made of long thin elements are easily modeled using standard antenna rules. To review, these rules are:

- Neglect undriven elements if they are less than about 0.1 wavelength long.
- Use four segments per half-wave for an indication of performance, eight for reasonable accuracy, sixteen or more if high accuracy is important.
- Check the geometry, especially the end points, to be certain the elements are properly positioned.

A boom is treated exactly like a dipole element, except that it normally isn't driven, but parasitic. Guys are also simple, because they are simply inclined long wires. One end of grounded guys must end at the earth's surface, z-dimension of zero. Poor earth can be simulated by introducing a loss—a load resistance at the junction point.

Over most common frequencies, the tower is simple—just a grounded vertical antenna. But you must watch the tower size-frequency relationship. For example, a common tower leg width is 12 inches, or about 0.3 meter. At 14 MHz, this represents 0.014 wavelength, well within the "wire size" limit for good accuracy. However, at 144 MHz the tower width is about 1/8 wavelength, and accuracy will be poor because the thin-wire approximation isn't valid.

However, this doesn't mean that the results are worthless. In a typical situation the tower currents will be low if the antenna is symmetrically mounted on the tower. A typical value of maximum tower current will be one percent of the antenna driving point current. Because power varies as the square of current, the tower radiation will be around 1/100th of a percent of the antenna radiation, or 40 dB below the drive power. About the only effect of the tower current will be to fill in the nulls somewhat.

Suppose the calculation of this current is off by a factor of ten, so the actual current is 10 percent of the antenna drive. Now the power radiated will be 20 dB down. Such an error won't affect the main-lobe calculations at all, but may mean that the secondary lobes and nulls are very much in error. In addition, there may be some error in the driving point impedance. However, the calculated results will still give an overall estimate of performance.

There are several ways to check the importance of such stray currents. One is to compare the pattern and drive resistance of the antenna alone and with the tower present. A second is to introduce a large resistance at the point of maximum stray current, and make the same comparison. Still another is to move the drive point from the antenna to the stray current point. If the antenna-alone pattern appears to be badly degraded, try another structure, or consider introducing insulators—the equivalent to the large resistance trial. In this connection, it's no longer necessary to physically cut a guy or even a tower to stop the current flow. Several ferrite beads, or even pieces of old TV H-oscillator or deflection ferrite cores, at the point of maximum current will reduce the current to a negligible value. See the book, *Reflections* by Walt Maxwell, W2DU and *The ARRL Antenna Handbook* for methods of estimating and measuring the effects.

If the stray currents on a thick tower do seem to be important, analyze the tower interaction as a wire-frame model (see the following section).

Example of antenna-like structure analysis

Suppose you've been asked to advise a friend on the use of his sailboat rigging as an antenna. The common method of doing this is to insulate the backstay, and use this as a sloping vertical fed against the ballast, motor, and other metal hull parts. Is this the best way to go?

For simplicity's sake, assume that the boat is 12 meters LOA, 4 meters beam, with a 10-meter high mast, or about the dimensions of common 35 footers. Also for simplicity, assume that the mast is stepped on the ballast, and that the stay and shroud chain-

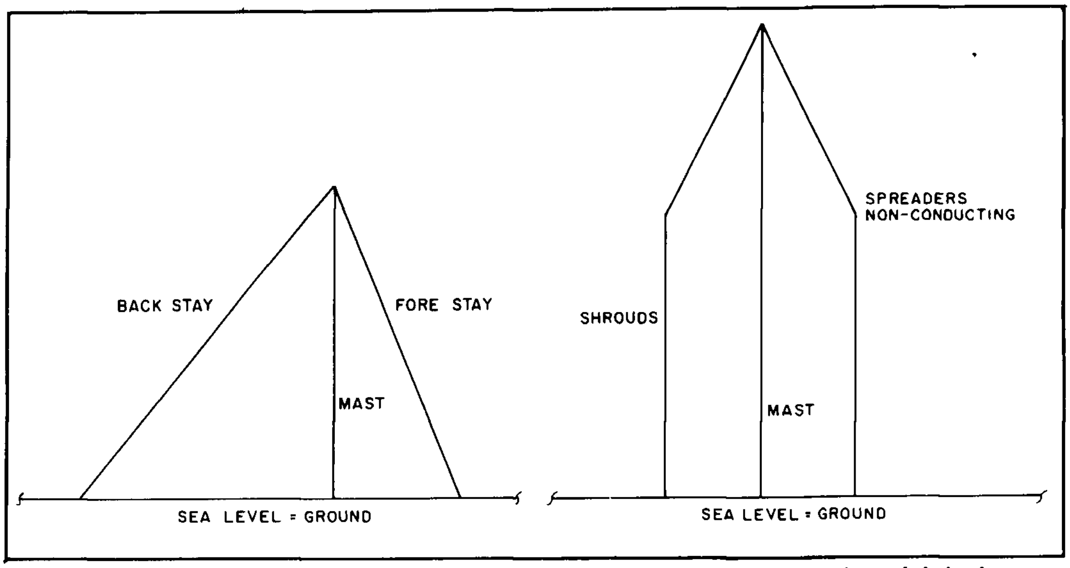


Figure 1. RF model of a typical sailboat. Only the key metallic elements are shown. Keel, engine and chain plates are assumed to be at water level or below. (A) View from port or starboard. (B) View from bow or stern.

plates make a straight line with their wires, effectively connecting to the earth at water level. The spreaders are at 5 meters, and are non-conducting. The ship model is as shown in the two views of **Figure 1**. Assume that the shroud/stay diameter is 0.01 meter, and the mast diameter is 0.2 meters. Use sixteen segments for the longer elements, and eight for the shorter.

Solution time depends on the computer. Fifteen to 20 minutes per frequency is typical. Making a printout of the results will add to the time, but it's worthwhile for future reference.

Assuming that the feed is at the base of the backstay, with other stay/shroud/mast connections intact, you should obtain the following results:

	Drive (Ohms)		Radiation	Ratios, dB
Band	**R**	**X**	**F/Back**	**F/Side**
160	0.4	− 260	0	6.6
80	80	1240	− 0.5	1.5
40	335	− 450	− 1.6	−0.8
20	30	80	3.3	6.5
15	170	− 270	10.7	0.6

With a parallel resonance at 5.24 MHz, and a series resonance at 13.28 MHz, this will be a good antenna from 80 to 15 meters with a good antenna match box, and a usable but inefficient one on 160. There is some signal variation as heading changes, but this is less than the normal fading range.

You can make additional runs with the feed at other locations, and with various insulation trials, at the top and/or bottom of shrouds and stays. You can introduce other antennas, like the often used "Hustler Whip" on the quarter. Because this is an example rather than a real problem, I won't explore these here.

Large-surface structures

To illustrate the approach to using thin-wire antenna analysis in modeling surfaces, assume that an antenna is to be mounted on a square metal plate—a ground plane. The plate can be represented by a grid of crossed wires. Two forms are shown in **Figure 2**.

One limiting condition shows up if it's necessary to mount the antenna at the center of the plate. Because there's no wire junction at the center on **Figure 2A**, there is no "ground" to feed the antenna against. This point does exist in **Figure 2B**. It's evident that the planning of the wire grids must include antenna location as a factor.

The usual "rule of thumb" is that the wire surface area should equal the area of the surface it is to represent. But there are limits to this. One is that the wire radius must be small compared to a wavelength; a radius greater than 1/30 wavelength is likely to give errors. The segmentation of wires is also important. A segment shouldn't be longer than 1/5 wavelength.

The usual arrangement is a square mesh; that is, four 90-degree angles. However, there are situations where other angles are necessary to model the surface geometry, and even places where different angles give better results. For example, the flat plates of **Figure 2** can also be represented by the radial wires of **Figure 3**. These give at least as good a representation as the square mesh, and require fewer wires and therefore less computation time. Probably the best rule is to

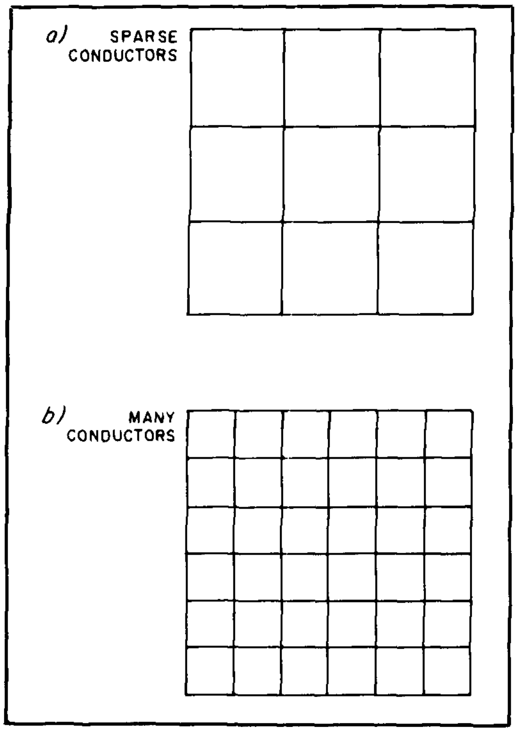


Figure 2. Wire-grid models of a square ground-plane plate. (A) A coarse grid, which requires large conductor size and limits antenna locations. However, it conserves computer memory and requires less time for solution. (B) Fine grid, suitable for use at higher frequencies with less restriction on antenna location. Wire size can be one-half that of (A) for the same accuracy. Grids are shown as squares, the preferred arrangement, but may be 3, 5, or more sides, with angles other than 90 degrees.

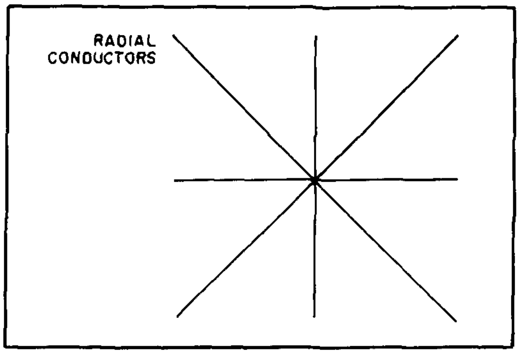


Figure 3. An alternative to an equal-sided grid, especially useful at the antenna mounting location. Such radials may be used as diagonals to ensure good distribution of current and better accuracy. However, such refinements increase memory demand and run time.

use the square mesh for the general body, and radial connections at and close to the antenna mounting point.

One problem with the models of **Figures 2** and **3** is that the current is necessarily the same on the top and bottom of the plate. This may not be accurate in some cases; for instance, with a ground-plane antenna mounted atop a mast. In such situations two plates separated by a small distance are necessary, as shown in cross-section in **Figure 4**.

It's evident from these examples that there's a considerable amount of art in the design of the model for analysis. Any attempt at accurate modeling is likely to lead to a large number of wires, which, in turn, requires a large computer and long run times. Models for amateur use which are modeled on a PC, will have to be crude, and run times will be long.

It's very easy to make an error in laying out and entering the model data. Use a consistent approach to numbering wires, and use symmetry to help keep the numbers and dimensions correct. A program which assembles antennas from wires or elements is a big help, especially if it will allow you to mirror-image the elements—an easy way to get symmetry. In a complex model, the procedure is to check and recheck your design.

Example analysis: a car antenna

Suppose you want to use a diplexer with the factory-mounted whip antenna of a typical small two-door automobile. The basic questions you must ask are: what is the drive characteristic, and what kind of a pattern is obtained?

Figures 5, 6, and **7** show three views of a wire-frame model of such a car. Note that the top is supported at three points along the sides, with no other top-body connection, to represent windows. The top and body sec-

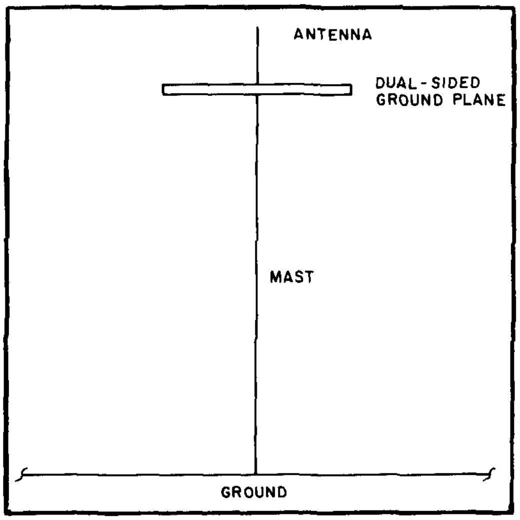


Figure 4. Accurate models sometimes require a closed surface to allow currents on the top and bottom to be independent. The mast-mounted ground plane is an example. The ground plane may be a grid, as in *Figure 2*, or radial, as in *Figure 3*.

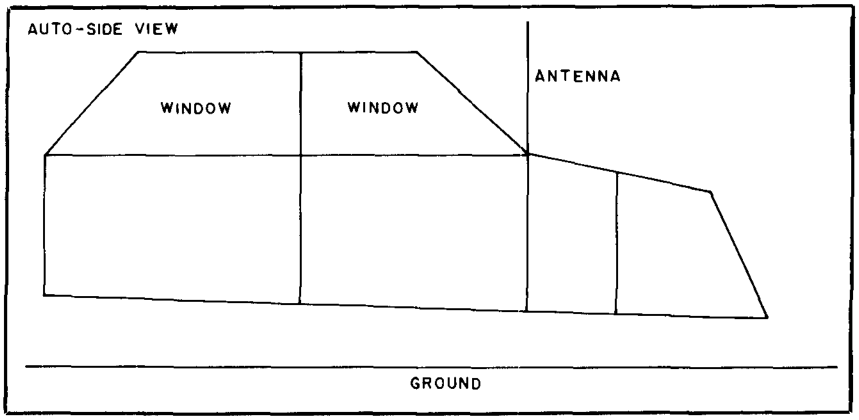


Figure 5. Modeling flat or curved surfaces requires many wire sections. This side view of a typical small automobile ignores many features, like wheels and bumpers. The relatively coarse grid is adequate for reasonable accuracy at 28 MHz, but the number of wires could be increased for better accuracy at 144 MHz.

tions do have a wire along the center line, and the body does have three side-to-side side straps across the bottom.

Despite it's relative crudity, the model approaches the limit of typical PCs. Only one current pulse per wire segment can be used with the standard MININEC program.

You can obtain worthwhile and interesting results even with this crude model and the limits of the computer. At 146 MHz, the factory mounted 30-inch whip on the right front fender at the windshield front shows a drive impedance of 95 – 18 ohms, an easy match. The pattern is surprisingly good. As seen by the car driver, the front-to-back ratio to the horizon is 0.8 dB. The right side radiation is 0.6 dB below the forward, and the left side is 0.9 dB lower. The pattern is essentially circular. Diplexer design should be relatively easy.

The built-in antenna could be used at 28.5 MHz. As with all short antennas, the drive impedance is poor: 2.4 ohms resistive and –415 ohm reactive. The pattern remains quite good, with a F/B ratio of –0.9 dB. The front-to-right ratio is –2.2 dB and the front left is –3dB. The poorest radiation is toward the front.

Once you have the basic car model, it's easy to move the antenna location and to change length. The absolute values calculated from this coarse model are likely to be in error. But the relative change will be better, and it will be easy to see if a "relatively optimum" location has been found. (However, it will take some time to run the range of possibilities.)

More wires and a greater degree of accuracy would be possible with a late-model computer with more memory, provided the arrays used in the program are redimensioned. Compiling the program (as in Quick Basic) is a good time saver. If much work is to be done, a numeric co-processor saves time.

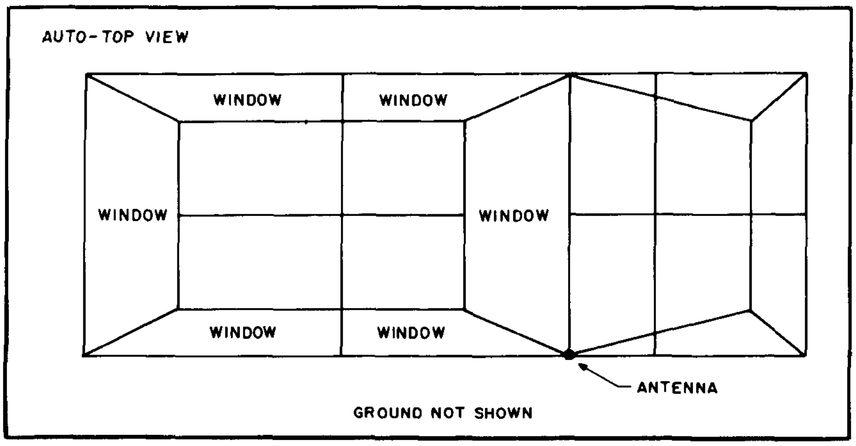


Figure 6. Top view of the automobile in *Figure 5*. In numbering wires for computer entry, time can be saved by making use of the symmetry of the structure. It's also easier to avoid errors if the wires are numbered in mirror-image pairs.

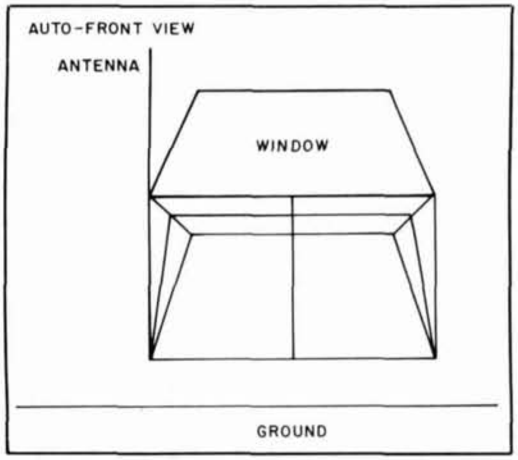


Figure 7. Front view of automobile in *Figure 5*. This and additional views aren't usually necessary to see all of the model wires, but are helpful in avoiding numbering, dimension, and entry errors. Some CAD programs produce this third view automatically, and also output a table of line (wire) end points. If much work is to be done, a small program can be written to transfer the table to RF analysis format.

Data on modeling

There's considerable data to be found on the relationship between model design and accuracy in the paper, "Verifying Wire-Grid Model Integrity with Program 'Check'," by C.W.Trueman and S.J.Kubina in the *Applied Computational Electromagnetics Journal* (vol. 5 no. 2, Winter 1990). This reference includes a table of warning/error conditions, and many sketches showing probem situations. You might also want to read "Selecting Wire Radius for Grid/Mesh Models," by L.A.Oyckanmi and J. Watins, in the same issue of the journal, or "Modeling Electrically Small, Thin Surfaces with Wire Grids," by T.H. Hubing and J.F. Kauffman in the previous one.

You may also find it helpful to read "Wire Grid Modeling of Surfaces," by A.C. Ludwig, in the *IEEE Transactions Antennas and Propagation* (vol. AP-25, September 1987). All the references in the papers listed here will lead you to other papers on this subject.

A caveat

Computer modeling of large area surfaces is a relatively new field, and one in which there is little or no amateur experience to draw from.

The reference articles warn that the field is more of an art than an exact method of analysis. Until we get some experience, including comparisons between computer model results and actual performance measurements, the results of any simulation run should be used carefully. In particular, expect that the results are really no more than general guidelines, and a good starting point for work in the "real world" of antennas and their mountings. ■

PRODUCT INFORMATION

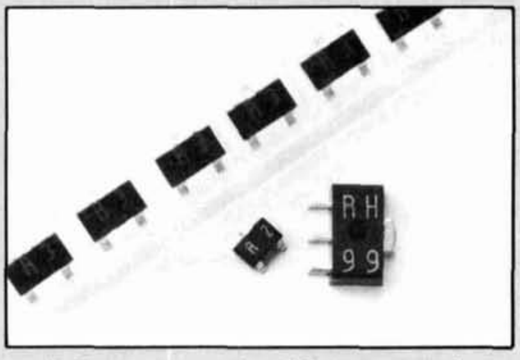
New Medium-power Silicon Bipolar Device

California Eastern Laboratories announces the arrival of NEC's new medium-power NE46134 silicon bipolar device. This tiny surface-mount device delivers high dynamic range and low noise figures.

Key performance features include:

High P_{out}
1/2 watt P_{out} at 1 dB compression
(12.5-volt power supply)
400 mW P_{out} at 1 dB compression
(10-volt power supply; IP3 > 37 dBm)
170 mW P_{out} at 1 dB compression
(5-volt power supply; IP3 > 32.5 dBm)
1 watt saturated P_{out}
(10 to 12-volt power supply)
Low noise figure
1.5 dB at 500 MHz
2.0 dB at 1 GHz
Low IMD
– 40 dBc two-tone at 1 GHz
(20 dBm total P_{out})

The surface mount version, NE46134, is



available on tape and reel for automated assembly, and is recommended for amplifier applications to 1.5 GHz. The chip version, NE46100, is for TO-8 and medium-power hybrid amplifier designs to 3 GHz.

Both the NE46134 and NE46100 are available from California Eastern Laboratories. For more information write California Eastern Laboratories, Inc., 4590 Patrick Henry Drive, Santa Clara, California 95056-0964, or call (408) 988-3500.

Cornell Drentea, *WB3JZO*

IMPROVING RECEIVER PERFORMANCE IN MODERN TRANSCEIVERS

Simple front end PIN diode modifications improve dynamic range

This article investigates the design criteria for receiver front-end diode switching. It explains why some diodes are better than others at switching RF currents, and makes recommendations regarding circuit changes in present and future equipment. The information here applies to most general coverage transceivers and receivers on the market today. I hope that Amateurs, as well as manufacturers, will benefit from this discussion.

The problem

Modern shortwave receivers have an incredible number of input signals entering on the antenna lead. The dynamic range requirement created by adjacent signals, as well as far-removed signals, can be very stringent. For this reason, highly selective filters and good linearity are essential to limit the creation of in-band spurious signals.

These extraneous signals arise from the intermodulation of multiple signals in nonlinear circuit elements. Some older designs used massive gang switching or miniature relays to preserve linearity, but that was costly, unreliable, and compromised shielding. Nowadays, electronic switching is preferred. Unfortunately, many top-of-the-line receivers use simple nonlinear diodes in these switches. The result is a noticeable loss of performance under heavy traffic conditions. However, you can use PIN diodes as a replacement for these inferior diodes and obtain substantial improvement.

A new point of view

The replacement of diodes in transceivers isn't a new technique. If you listen to the various equipment nets on 20 meters, you'll hear about the "radical" improvements obtained by changing all front-end filter switching diodes in one radio or another. However, one misconception aired on such nets concerns the replacement of the original PN-junction diodes, like the 1N914, 1N4148, or their Japanese counterparts, with Schottky (hot carrier) diodes, like the Hewlett Packard HP 5082-2800. Although the performance after such changes has been billed as "better," the HP 5082-2800 Schottky diodes are actually the wrong diodes to use in these applications. Apparently, many experimenters don't know that while Schottky diodes may appear to improve the low end signal handling of receivers, their overall perform-

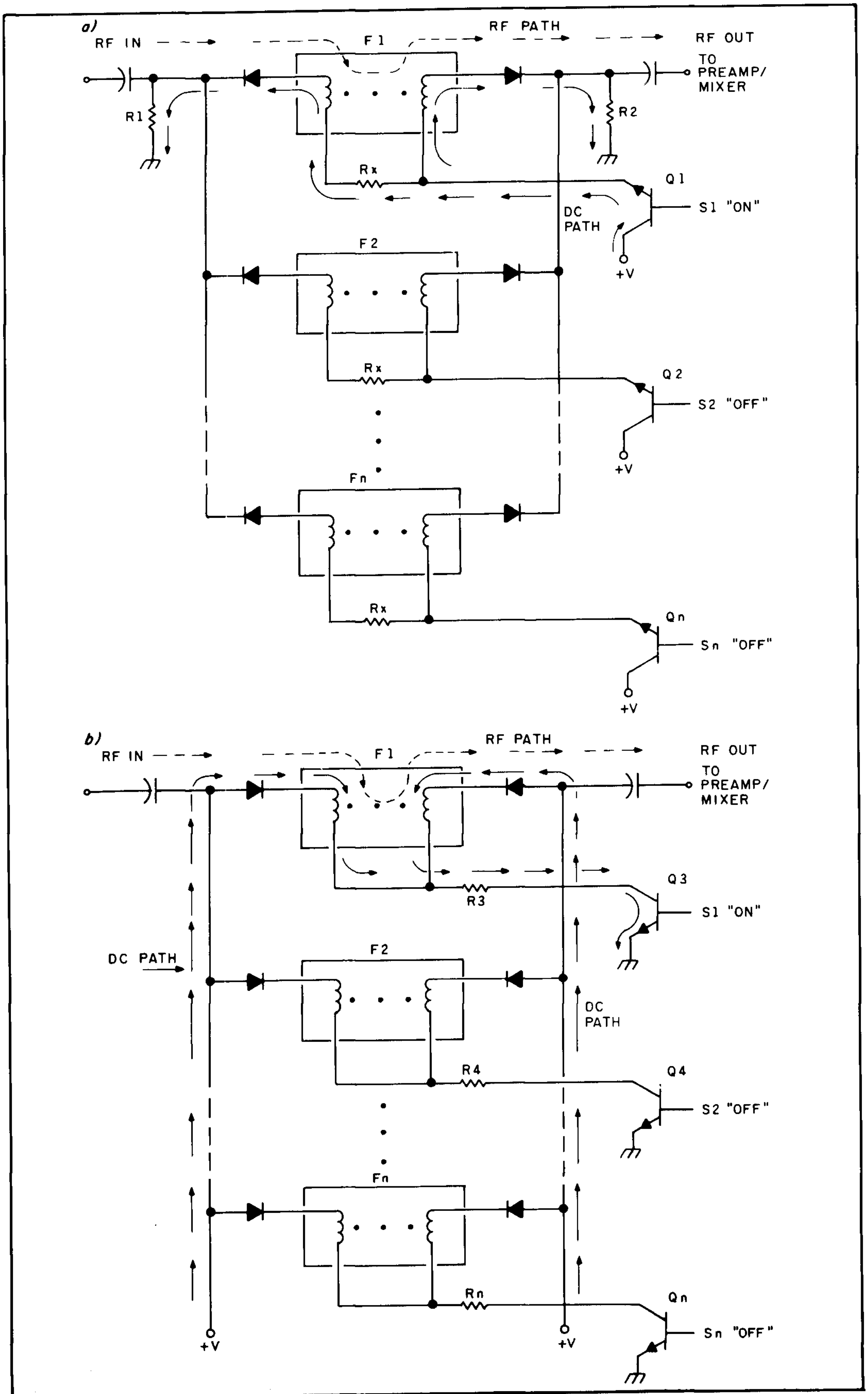


Figure 1. Typical implementation of automatically switched filters for the front end of a general coverage transceiver. (A)Standard forward biased. (B)Forward biased with open collector. Inexpensive silicone PN junction diodes like the 1N914, 1N4148, or 1SS53 are used routinely today.

ance as basic RF switches may be marginal at best when multiple strong signals are present at the receiver's inputs. Simply put, Schottky diodes act as mixers versus perfect RF switches, like the relays or mechanical switches, for the incoming signals, creating unwanted characteristics for the receivers. Consequently, what is perceived as a "hotter" receiver could actually be a "noisier" one.

To understand how this happens, let's look a little closer at the various types of diodes available to the receiver designer, and their characteristics.

Simple PN-junction diodes (the 1N914 and the like) and Schottky diodes (the HP 5082-2800, for instance) are intended to generate products when presented with different frequency signals. This nonlinear behavior is a desirable feature usually exploited in controlled mixer applications. However, this property is undesirable when these diodes are used to switch bandpass filters at the input of a radio. Ideally, such a function requires hard, mechanical-type switching, like that provided by relays. This kind of switching would be preferable from an electrical point of view, because relays don't generate intermodulation distortion (IMD). On the other hand, using several relays in front-end switching would quickly prove a cumbersome and expensive proposition (miniature RF relays are available from several vendors).

PIN diodes, because of their physics, act closer to a perfect distortionless mechanical switch—defined in this application as one that should not generate mixing products. If PIN diodes are used to switch in bandpass filters, receiver performance maintains a certain integrity in the third-order intercept point when compared with either PN-junction diodes or Schottky diodes. To the human ear, this translates into crisper, more intelligible weak signals in the presence of strong ones.

Before delving further into the physics of diodes, let's look at how they are configured as filter switches in the front end of typical general coverage receivers.

Some design considerations

Modern general coverage transceivers employ not just one or two receiver filters, but a complex bank of automatically switched suboctave filters intended to reject out-of-band interference over a wide frequency range. These filters overlap and provide automatically switched continuous coverage.[1] One, out of up to eight such filters, is usually selected by the synthesizer with the help of the inexpensive silicon PN-junction diodes discussed earlier. This is implemented in the simplified diagram of **Figure 1**.

A digital decoder senses the operating frequency of the rig from its synthesizer control circuits (**Figure 1A**). One of several control lines selects a filter by turning on NPN drive transistor Q1 or Q2. DC current then flows through the Q1 or Q2 transistors, and through both ends of the selected filter circuit into the diodes, as shown. The DC path is completed to ground through current limiting resistors R1 or R2. A similar process is shown in **Figure 1B**. The difference here is that all diodes have voltage applied constantly to their anode sides with current limiting resistors R3 and R4 placed in series. The circuit is completed to ground through open collector transistors Q3 or Q4. When properly biased, a filter is selected in much the same manner as at A. In either case, just one filter is selected at a time. This means that only the pair of diodes corresponding to the selected filter is being biased at any given moment. The diodes are said to be forward biased, a state which allows RF to flow through the selected filter.

Now let's discuss the occurrence of IMD in these circuits.

In a simplistic way, when in-band RF signals coming from the antenna cross the I-V characteristics of the activated diodes, unwanted signal products result. These products translate into audible whistles and increased noise floor, which can be further aggravated by the synthesizer's phase-noise performance and by the receiver's AGC. When unmatched PN-junction diodes are used, the various signals going through the biased filter circuit cross the uneven I-V characteristics of the input and output diodes at different points on their curves. This creates additional IMD. Schottky diodes eliminate some of these problems by providing guaranteed matched I-V curves. However, Schottky diodes have been designed to generate products which are useful in mixer applications. Consequently, they will also be undesirable parts for these applications (see **Figure 2**). In contrast, the switching characteristics of PIN diodes work on a totally different principle. PINs act as variable RF resistors whose resistance depends on DC excitation. That is, RF energy will pass through the diodes when a small amount of DC current flows through them. In a front-end filter switching system, RF signals coming from the antenna would be superimposed over the DC current. Their incident energy would cause an additional rectified current to be generated in the diodes. This, in turn, would further lower the diode resistance in such a

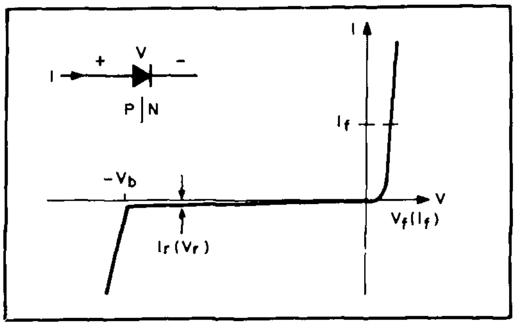


Figure 2. Current versus Voltage (I-V) characteristic of a typical PN-junction signal diode. In standard, unmatched parts, this curve will vary from part to part. In-band RF signals will cross the curves of input and output diodes at different points producing unpredictable distortion. In Schottky diodes, the position of these I-V curves is guaranteed to be the same for all parts over a wide frequency range. Better signal handling results from this, especially at the lower end on the curves—giving an impression of a more sensitive receiver. PIN diodes, on the other hand, minimize these problems by acting as current controlled resistors or switches, without exhibiting the severe intermodulation distortion problem of the PN-junction or Schottky diodes. They are recommended for receiver front end filter switching.

way that the RF would flow through them and consequently the filters they're switching. When properly biased, you would find that, for the most part, only the fundamental frequencies of the signals would pass through the filters, much as they do in mechanical switches.

A little bit of physics

A close look at the physics of a PIN diode like the HP 5082-3080 reveals that it is constructed very differently from a PN-junction or a Schottky diode. While signal diodes are manufactured on the principle of a metal deposited on a P or N semiconductor, the PIN diode is constructed with a thick high-resistance nearly intrinsic silicon layer (I) sandwiched between the P and the N semiconducting materials, hence the name PIN diode. The resistance of the I layer with no DC applied is on the order of about 10 k. This property is known as isolation. The PIN diode's depletion capacitance is also reduced due to the thick intrinsic layer between the P and the N layers, allowing the diode to operate as a good attenuator or switch at much higher frequencies than the regular PN-junction diodes (see **Figure 3**).

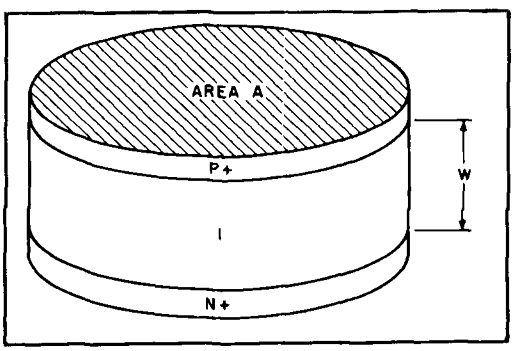


Figure 3. Construction of a PIN diode. Unlike a signal diode, the PIN diode uses a thick near intrinsic silicon material which is sandwiched between the P and the N materials. When properly DC biased, RF energy will propagate through the diode with minimum internal distortion.

When the diode is forward-biased at a given current, holes and electrons are injected into the I region from the P and N regions. These holes and electrons don't immediately recombine in the I region as may be expected, but rather coexist for a period of time. The average time before recombination is called the carrier lifetime. The mean distance a charge travels before recombination is the diffusion length, which, in turn, is related to the carrier lifetime. This injection-recombination process is continuous when the diode is forward biased. This results in a steady-state charge in the I layer which depends on the forward current and the carrier lifetime. The final effect is to lower the resistance of the I region slowly making it a better conductor.

Looking at it another way, it can be said that PIN diodes use their dynamic intrinsic properties to make RF signals remain in an on state longer than PN-junction diodes, much like relays would, but without the distortion rich crossings of the I-V curves so characteristic of the other two types of diodes.

We have seen that the RF conductance of PIN diodes depends on the DC current pumped through them. Within reason, the more current, the better RF switches they make, and the better their intermodulation distortion performance. This is why PINs have been classified as power-hungry devices when used in RF switching circuits. In general, current specifications for PIN diodes vary upwards to about 200 mA for a hard turned on diode. Low current PIN diodes have also been manufactured. For example, the HP 5082-3080 requires only 10 mA for an on-resistance of 8 ohms. The HP 5082-3081 exhibits 10 ohms for the same forward current, while M/A-COM's MA4P1200 is designed for less than 1 ohm at the same current. These later currents are compatible with most existing drive circuits in today's general coverage transceivers (see **Figure 4**).

Performing tests

Several laboratory tests were performed

after the regular PN-junction switching diodes in a Kenwood 430, an ICOM 725, and a Yaesu FRG 7700 were replaced with HP 5082-3080 PIN diodes. HP 5082-2800 Schottky diodes were also tested. No modifications were necessary for the biasing characteristics of the existing receiver circuits. Our tests produced consistent results in all rigs. As expected, the PIN's performance was superior to the PN-junction and Schottky diodes. In addition, the PIN HP 5082-3080 was found to perform better than the 3081. This was attributed to the 3080's lower on-resistance for the same current (10 mA), as was discussed earlier. A two-tone receiver laboratory test was performed for all three radios in the 20-meter band with 20 kHz spacing. It showed typical spurious free dynamic range (SFDR) improvements of 6 to 8 dB when using the PIN over the PN-junction diodes. Although these numbers may not sound significant, they constitute substantial improvements which can be heard in the receiver outputs. The changes made the difference between copy and no copy for some weak signals in the presence of strong ones. Casual listening observations proved rewarding and showed considerably cleaner audio with the PINs installed. Schottky diodes were also tried in one of the radios. Although the receiver sounded somewhat "hotter," that is, more sensitive, the radio's attenuator needed to be used to reduce intermodulation distortion on strong signals. Conversely, with PINs no attenuator was needed to receive weak signals in the presence of strong ones.

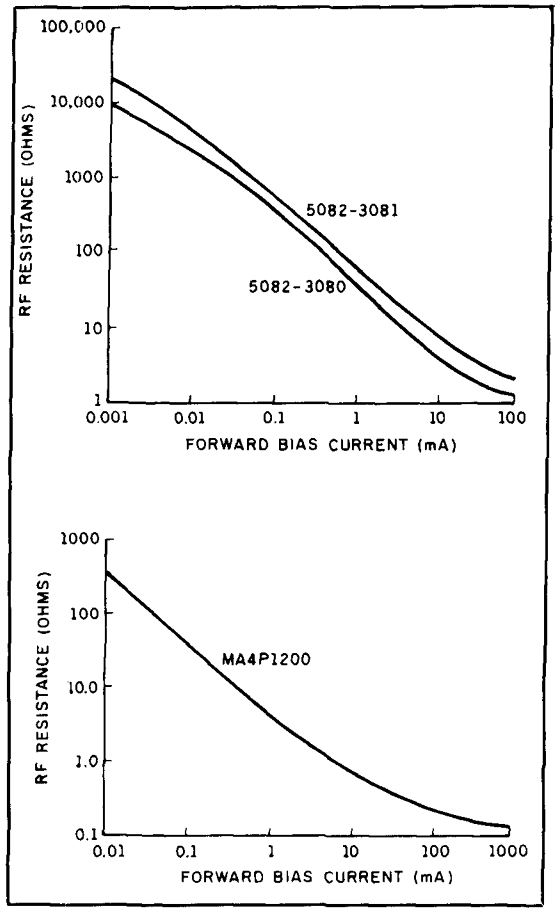


Figure 4. Forward bias current versus RF-on resistance for three types of PIN diodes. Only 10 mA is required to provide 8 ohms on-resistance for the HP 5082-3080. This PIN diode is ideal for front end applications in general coverage transceivers.

Making the changes

Figure 5 shows the areas affected in a typical general coverage transceiver.

Changing diodes in equipment is a relatively simple operation providing that care is exercised in removing and reinstalling the boards, and that proper tools are used for removing the old and installing the new diodes. Because of the mechanical variations in rigs, I can't assume responsibility for your implementation. Once you obtain the proper PIN diodes, use the general steps in the following paragraphs when performing the modifications.

First, study your receiver schematic diagram carefully to identify the diodes of interest before beginning. Typically, in Japanese radios these will be 1SS53 PN-junction, silicon switching diodes. A service manual will usually help in understanding the circuits.

You don't have to replace all diodes in the radio to realize some improvement. If you wish, you can modify just the bands of interest. Additional diodes can be added later if you find the modification useful. Some radios use two different banks of filters in the front end—one for the ham bands and one for the general coverage.

You can replace all of the diodes in both banks, or only those for the ham bands, depending on your requirements. If you aren't sure of what you want, just replace one filter, preferably in a busy band, so you can tell the difference. You can change the rest later.

Don't confuse the front-end filters with the high power output lowpass filters which are usually switched with heavy duty relays, or with the IF filters which usually follow the mixers and are also diode switched.

When replacing the diodes, I recommend that you use a 12 to 30 watt chisel-tip soldering iron equipped with a three-wire cord and a grounded tip. You can also use a soldering iron isolated through a transformer. Older irons may have 117 volts on the tip and may cause electrical damage on contact.

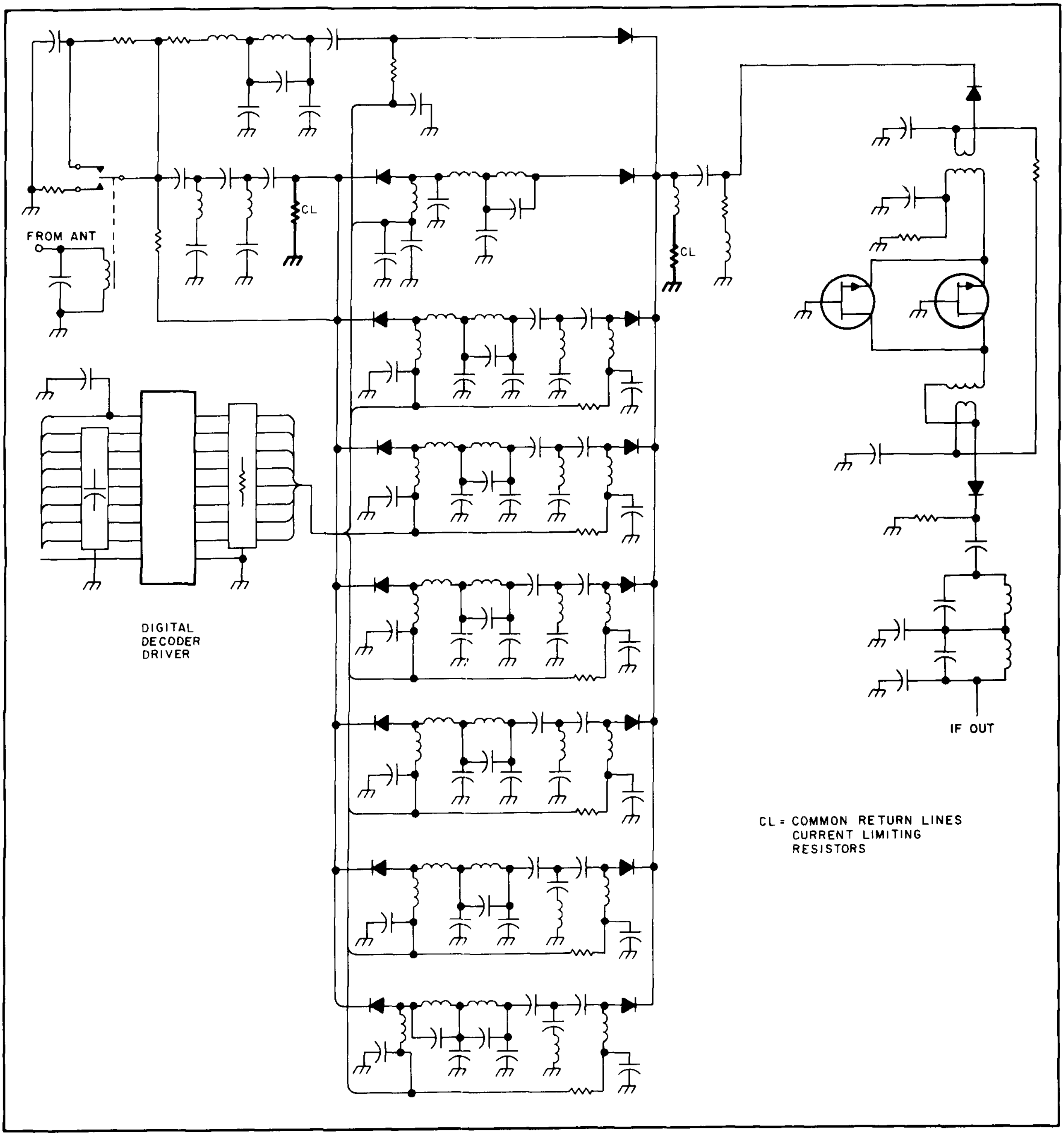


Figure 5. Suboctave filters and diode switching implementation in the ICOM IC 725. This is typical of most general coverage transceivers.

Use only 60/40 rosin core solder, never acid core solder. Also, use solder wick rather than a solder sucker when performing the changes. The latter tends to destroy circuits.

Locate the board which contains the front-end filters. To do so, look for several repetitive circuits switched by similar diodes electrically located in front of the first mixer (typically a balanced JFET circuit). Refer to the schematic in **Figure 5** as an aid in finding the board. Once you've identified the board, locate the filters and diodes of interest.

Carefully remove all of the connectors from the board. Make sure each connector is labeled, so you can restore the wiring to its original path after the change.

Next, unscrew all holding screws and carefully remove the board from the rig. Move the board to a static free area, making sure you've taken the grounding precautions mentioned earlier. If possible ground your body and all of the tools that you'll be using.

Turn board with the solder side exposed to you. With the help of a light, locate and mark all the solder diode points to be desoldered. You can use a permanent magic marker to write on the solder points for easier identification.

Carefully apply the hot soldering iron through the solder wick to the points to be desoldered. The solder should be absorbed by the wick within seconds. Make sure the leads of the old diode have been detached from the foil and will play freely in the holes.

Repeat this process for all other points. When you're finished, turn the component side of the board toward you and carefully remove desoldered diodes with a small pair of pliers or tweezers. Repeat the above steps if you can't remove the diodes easily.

Noting polarity, insert the replacement HP 5082-3080 PIN diodes. Bend the leads a little on the opposite side. Cut any excess leads.

Turning to the trace side of the board once again, solder the new diodes in place. Cut any excess wire.

Check all the new solder connections, and reinsert the board into the rig. Replace all the screws which hold the board in position. Finally, reinsert all the connectors. This should complete the change. If you had steady hands, and replaced the right diodes, you'll be rewarded with a considerable improvement in the dynamic range of your rig.

Conclusion

This article discussed, in some detail, why PIN diodes perform better than PN-junction and Schottky diodes for switching front-end filters in general coverage transceivers and receivers. I did not present the mathematics describing the processes of intermodulation distortion in order to keep the work as practical as possible. For those interested in the theory, several references which go into additional detail about the process of using diodes in RF switches and attenuators are listed in the bibliography.

I have also discussed procedures for implementing PIN diode switching in present equipment.

Acknowledgements

I would like to express my appreciation for the inputs and test materials provided by Hewlett Packard and M/A-COM Corporations. Particular thanks go to Jack Lepoff of Hewlett Packard, Gerald Hiller of M/A-COM Semiconductor Products, and Professor Robert Caverly of the Southern Massachusetts University. Also, many thanks to Rick Whiting (WØTN) for his input, and for reviewing the entire article before publication. ■

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PRODUCT INFORMATION

Optoelectronics Inc. Releases Inexpensive GHz Frequency Counter

Optoelectronics has released a new frequency/detector counter for use in two-way radio, ham radio, frequency monitoring, and other applications. The Handi-Counter Model 2300 features full eight-place readout resolution, 10-mV sensitivity for signal detection at maximum distance from the transmitter, and a display-hold switch.

The Model 2300 Handi-Counter is priced at $99 each in unit qualities. An optional NiCad battery pack is available for $29.

For details contact Optoelectronics Inc., 5821 NE 14th Avenue, Fort Lauderdale, Florida 33334.

Peter O. Taylor
AAVSO Solar Division
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SPECIAL COLUMN

THE SOLAR SPECTRUM

So many of you enjoyed our special sunspot report—"Outlining June's Strong Solar Flare Activity," and its companion article, "Recording Solar Flares Indirectly,"—in the summer 1991 issue that we asked author Peter Taylor, chairman of AAVSO's Solar Division, to give us a quarterly update on solar activity. Here's his first installment. Ed.

This is the first of my columns for *Communications Quarterly*, a format which provides a welcome opportunity to join in the exploration of our mutual interests. My intent is to share information about various kinds of solar activities with readers, emphasizing those which are the source of the terrestrial effects which influence us all. Hopefully in the process, I'll learn more about the ways that these conditions affect radio amateurs, while describing some of the Sun's features that I find so fascinating.

Variations in solar phenomena

A majority of the solar phenomena which disrupt the Earth's environment vary with the solar cycle, so let's take a quick look at where we are in that respect. Then we can cast a predictive eye towards what might be expected to occur in this regard during the next several months. I hasten to add that "might" is an appropriate qualifier in this case. Precise predictions of future solar activity are difficult at best, and frequently return to haunt the forecaster. I'll try to stay away from these.

There are many ways to assess the progress of a solar cycle. Among the more popular are the relative sunspot number, the solar 10.7 centimeter radio flux, the Sun's x-ray index, and the numbers of solar flares. These activities are usually monitored on a daily basis and the results averaged or summed over some convenient interval, like a month or year. These data, in turn, are often subjected to a mathematical smoothing process which allows the trend of the index to be more clearly defined.

Unfortunately, the indices don't necessarily rise and fall at exactly the same times, and it's not uncommon for one to reach its greatest amplitude much earlier or later than another. For example, the radio flux curve for cycles 18 and 19 peaked at about the same time as the spot-maximum.[1] However, flux rates for cycles 20 and 21 peaked a year or so afterwards, and the more recently devised solar x-ray index attained its highest value two years after the sunspot maximum of cycle 21.[2]

How can this apparent contradiction be explained? One explanation may be that while all of these indices measure the solar cycle, each monitors a somewhat different aspect of it. Thus, while it's true that a cycle is generally described in terms of its sunspot number, no one index should be thought to have more value than another. **Figure 1** shows how several indices have performed thus far during cycle 22. It's extremely likely that this cycle reached its peak smoothed sunspot number in July 1989, marking the shortest rise from minimum for any recorded cycle—less than three years—but also contributing to its reputation as a rather unusual cycle.

One interesting characteristic of the sunspot cycle is that even-numbered cycles tend to have shallower maxima than the cycles which bracket them. More importantly for our purposes, even cycles almost always exhibit extended maxima—a property which may provide us with a clue of things to come.

The greatest flares usually erupt several years after a cycle's spot-maximum when the more magnetically complex spot groups also emerge. Long-lived coronal holes, which are strongly related to recurrent geomagnetic storms, also form at this time. As a result,

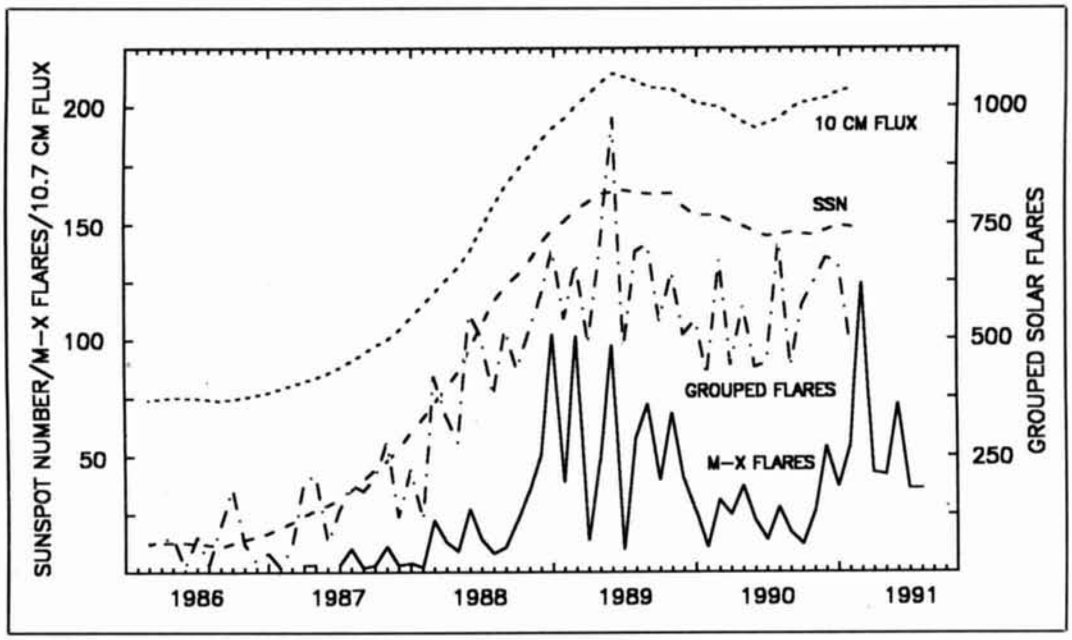

Figure 1. The performance of several solar activity indices thus far during solar cycle 22. Solar radio flux and sunspot numbers (SSN) are smoothed monthly-mean values. Grouped and M-X flare indices are monthly totals. (Grouped flares are the actual number of flares recorded during each month. Class M-X flares are strong X-ray events.)

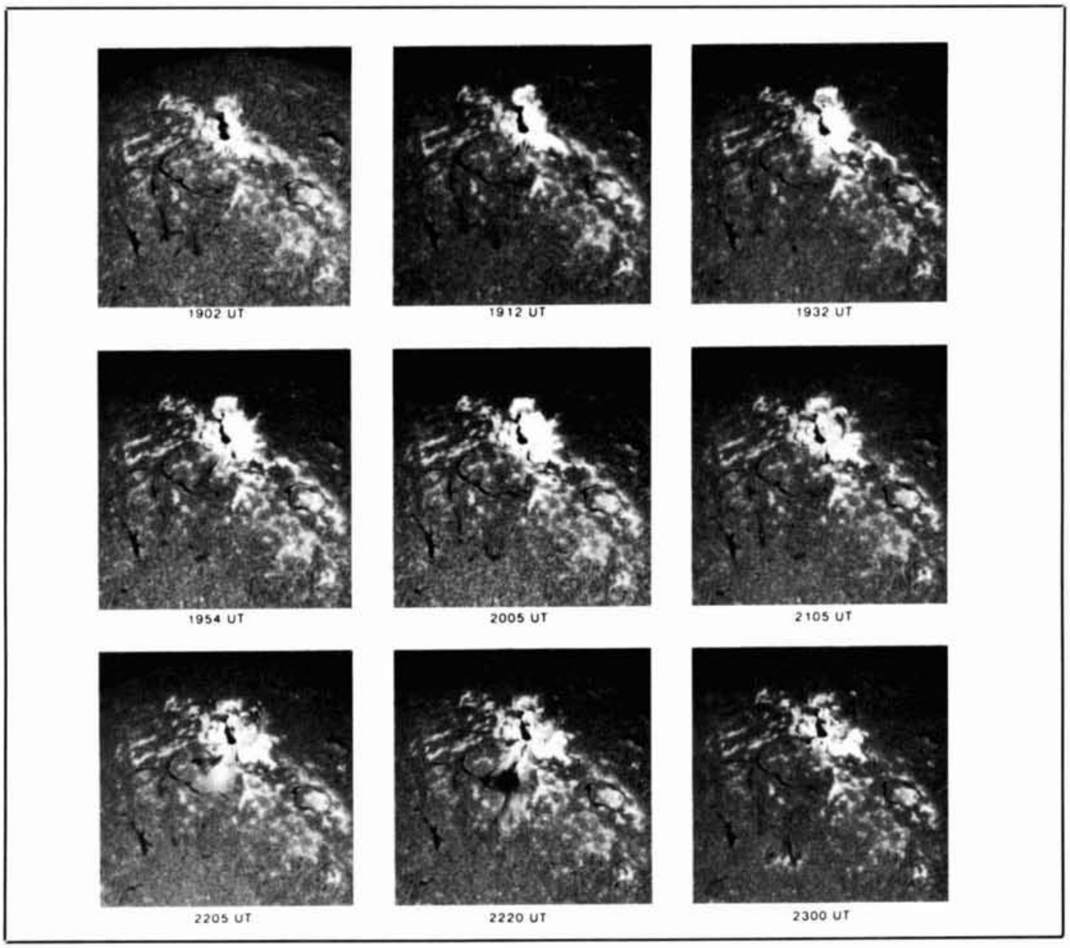

Photo A. The intense flare shown in this series of photographs taken at the National Solar Observatory—Sacramento Peak, reached maximum at about 19:12 UT. Note the enormous ejection of dark material into the solar atmosphere in the frame taken at 22:20 UT.

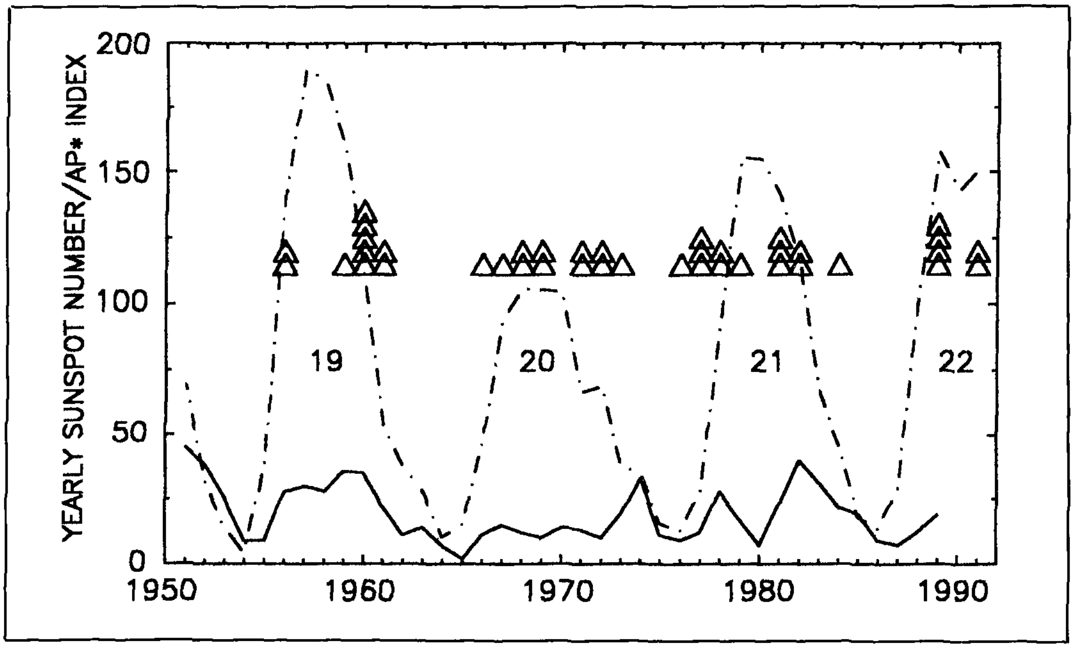

Figure 2. The yearly sunspot number and Ap* index from 1951 to 1991.5 and 1989, respectively. Note that these large geomagnetic storms tend to reach a peak sometime after the maximum sunspot number. Triangles represent ground-level proton events (data may be incomplete after 1986).

it's reasonable to expect intervals of enhanced activity to occur through 1992 and even into 1993. A slow decline to a cycle minimum sometime in late 1996 or 1997 should begin shortly thereafter.

Recent occurrences

The last several months have produced some of the most intense flares and highest daily sunspot numbers of this cycle, although some periods of relatively low activity have also taken place. As would be expected, the terrestrial environment has not always escaped the consequences of these events. In fact, the level of geomagnetic activity between late March and this fall—the number of days at major storm levels or above—has been the greatest in over thirty years.[3] A smattering of flares have been ex-

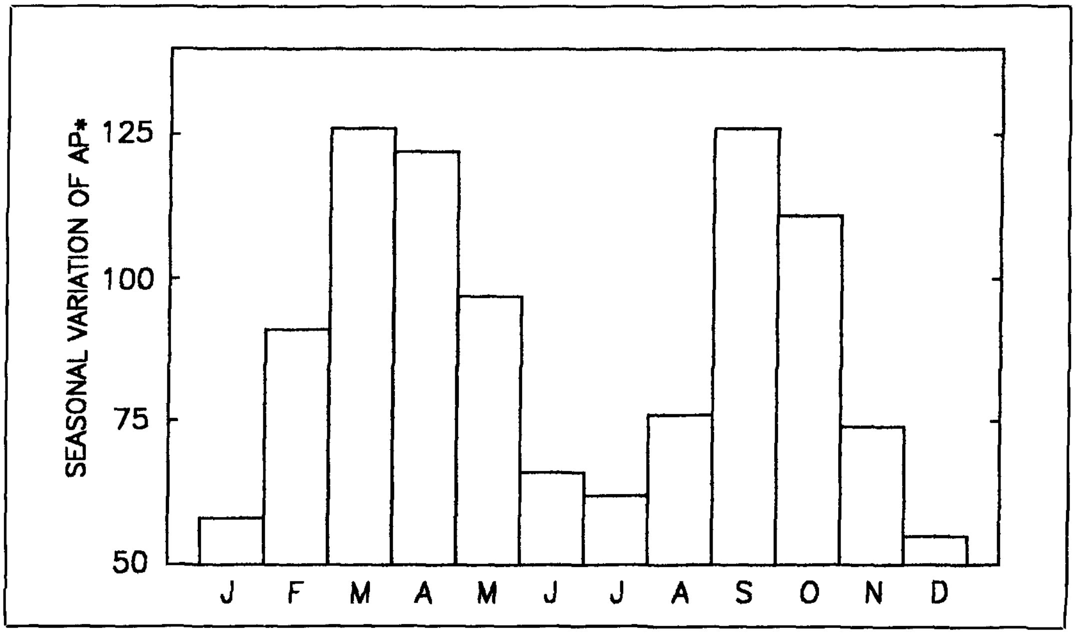

Figure 3. Large geomagnetic disturbances are strongly correlated with the spring and fall equinoxes when the alignment between the Sun and Earth is most favorable.

tremely powerful; a few actually spawned bursts of energy which equaled the energy consumed by humans throughout their history.

Photo A is a portion of a remarkable series of photographs that detail the eruption of an intense flare and huge ejection of material into the solar atmosphere. Amazingly, the cloud of matter released in the frame taken at 22:20 Universal Time dwarfs the Earth several times over. Small wonder that the combination of flare-generated radiation and shock-wave in the solar wind which accompanies many of these events causes such dramatic effects on the Earth's magnetic field and atmosphere!

The large magnetic storms which frequently occur as a result of eruptions on the Sun and variations in the density of its atmosphere (that is, coronal holes) often attain their maximum amplitudes over intervals which span more than one day. As a result, it can be difficult to tell if mid-range values on adjacent days actually reflect a single storm with a much higher disturbance level just by examining the standard three-hourly or daily magnetic index.

The Ap* index

In 1978, NOAA's Chief of Solar-Terrestrial Physics, Joe H. Allen, sought to minimize this obstacle by devising a new way to gauge these events. He called it the Ap* index. This measurement, which is derived by smoothing the "equivalent amplitude" three-hourly ap values, attempts to define each storm separately. Judging by this standard, more than 1000 such storms occurred between 1932 and 1989 (see **Figure 2**).

According to researchers at the National Geophysical Data Center in Boulder, Colorado, applications of the Ap* index include comparisons with spacecraft anomalies, communications problems, weather and climate relations, and possible biophysical interactions. Furthermore, the strong tendency for geomagnetic storms to occur around the spring and fall equinoxes may provide a partial explanation for the relatively soft terrestrial effects of June's extraordinarily intense flare activity. The seasonal nature of this relationship is shown in **Figure 3**.

The genesis of many large magnetic storms is often signaled by a phenomenon which can play havoc with ground and satellite communications. Since shortly after the end of World War II, it has been known that the Sun has the ability to accelerate particles with an energy sufficient to penetrate to the Earth's atmosphere. However, because they are relatively rare at the Earth's surface, these events have only been routinely observed since the advent of a strong satellite technology in the 1960s.

The arrival of these particles at the Earth was originally called a solar cosmic ray event, because their presence was first detected with ionization chambers. However, the means to detect them has changed and become far more sophisticated in the ensuing years, requiring a more definitive terminology. Thus, satellite-level proton events are those which are detected by the extremely sensitive monitors aboard spacecraft, and ground-level events are those recorded at the Earth's surface. The definition for a polar cap absorption was extended in the 1950s to include proton events which are felt particularly in the polar ionosphere.

Studies have shown[4] that more than 200 such incidents have reached event threshold (energies greater than 10 MeV) during the last thirty five or forty years. However, only a comparative handful (about 15 percent) have had a degree of energy sufficient to allow their penetration to ground level.

This same research indicates that no real pattern of occurrence can be determined in these data other than a very general correlation with the solar cycle itself (**Figure 2**). Consequently, ground level and other proton events are one more offshoot of the Sun's activity which continues to be difficult to foresee, and both inconvenient and potentially costly to experience. Perhaps Einstein had complications such as these in mind when he remarked, "I never think of the future. It comes soon enough." ■

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John R. True, *N4BA*
Reprinted from *Ham Radio*, May 1975

HOW TO DESIGN SHUNT-FEED SYSTEMS FOR GROUNDED VERTICAL RADIATORS

A graphical design system for using your tower system as a shunt-fed vertical antenna

Fifteen years ago, this article triggered a useful round of vertical antenna development. At this point in the solar cycle (nominally 22 years), the Maximum Usable Frequency (MUF) is going down again. This means renewed interest in lower-band DX. DX requires low angle radiation. Therefore, the economically limited amateur usually ends up experimenting with vertical antennas. We hope you enjoy this look back and the implementation notes by Hunter Harris, W1SI. Ed.

Vertical antennas have several advantages over horizontal dipoles on the lower amateur bands, especially in those cases where the dipole cannot be raised at least one-half wavelength above ground. An earlier article showed how to use a 54-foot tower, top loaded with a quad or Yagi, as a grounded vertical radiator on 40 and 80 meters.[1] However, to properly design the shunt-feed matching system for these two lower bands a good quality impedance bridge was required. Once the complex input impedance had been determined, a graphical method was used to calculate the components required to match that impedance to a 50-ohm transmission line.[2]

This antenna system generated a great deal of interest, but since few amateurs have access to an RX impedance bridge, they were unable to use this technique to adapt their own towers for use on the lower amateur bands. For this reason, I decided to make a series of measurements which would be used to generate a set of graphs which would simplify the design of shunt-fed vertical radiators. These graphs are presented in this article.

First, a series of antenna tests were conducted by scale modeling to determine the electrical height of towers which are capacitance loaded by a typical Yagi beam or cubical quad. Further tests were conducted to determine how long the gamma-type shunt feed rod had to be to permit the use of a practical L-network for matching to 50-ohm coaxial cable.

All tests were made with an aluminum-tubing gamma rod about 1 inch outside diameter, spaced 10 ± 2 inches from one leg

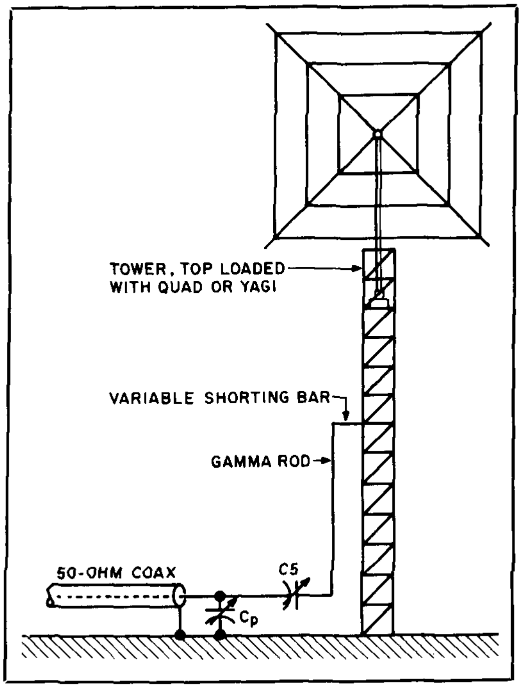

Figure 1. System for using a tower as a shunt-fed grounded vertical radiator on 40, 80, or 160 meters.

of the tower. This size was chosen for maximum physical and electrical stability, as well as high conductivity. The resultant design curves show the electrical height of the tower, required gamma rod length, and series capacitance, C_s required to cancel the inductive reactance of the gamma rod. The *parallel matching capacitance*, C_p, is also given (**Figure 1**). The series and parallel capacitors should both be air variables, so the matching system can be adjusted to provide as low VSWR as possible.

Using the curves

The graph of **Figure 2** shows the relationship of physical height to electrical height of a thin wire (calculated from $234/f_{MHz}$), measured electrical height of a 1-1/2 inch diameter conductor (which coincides very closely with the predicted electrical height of a thin conductor), and a tower 1 to 2 feet in cross section, top loaded with a Yagi or cubical quad. If you wish to use your present tower as a vertical antenna for the lower bands, you can determine its electrical height from the data presented in **Figure 2**.

The data in **Figure 3** is for use on the amateur 7-MHz band and shows the length of the gamma rod and required series capacitance for towers up to 90 feet high (about 3/4 wavelength on 40 meters). Towers which are taller than this will produce a large lobe of high-angle radiation that reduces the radiation at lower vertical angles. Some shorter towers may actually be shorter, physically, than the recommended gamma rod; in that case more parallel capacitance will be required to match the system to 50 ohms. **Figure 4** shows the same type of data for the 80-meter band (towers *higher than 180 feet* exhibit the large, high-angle lobe).

The data in **Figure 5** is for use on the 160-meter band. Note that a tower which has an electrical height of 90 feet, requires a

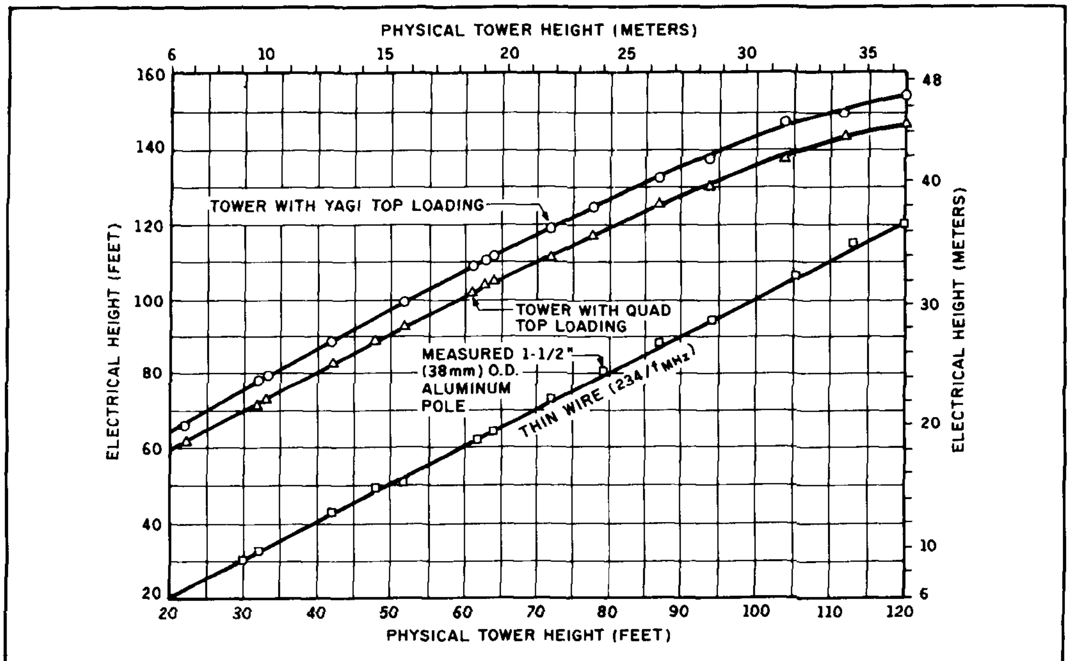

Figure 2. Physical versus electrical height of towers top loaded with Yagi beams or cubical quads.

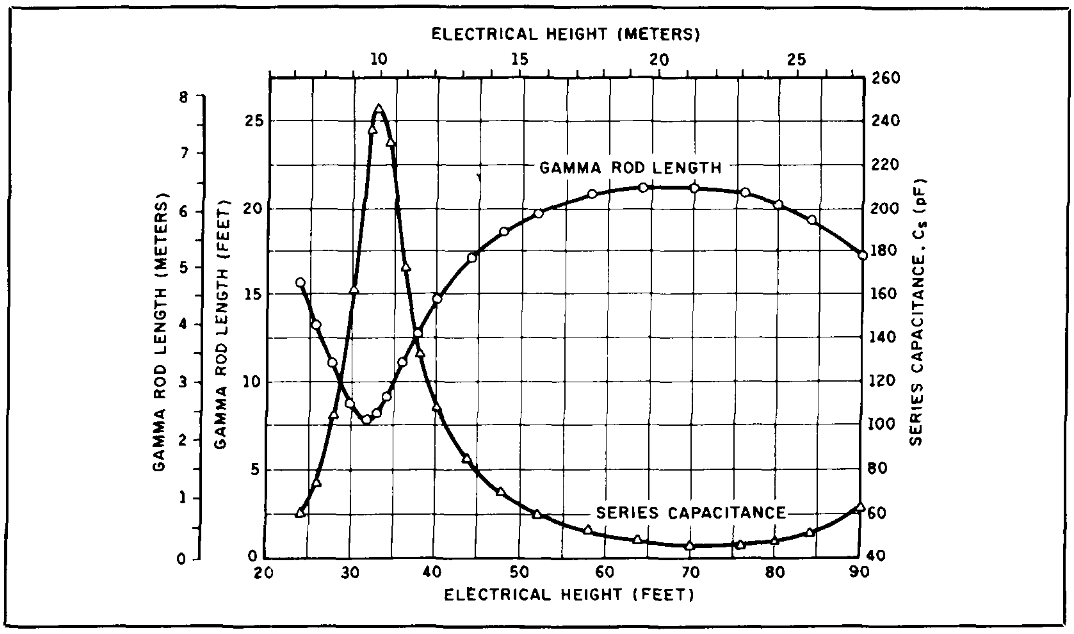

Figure 3. A 40-meter vertical. Gamma rod length and series capacitance versus electrical height of tower. Recommended parallel capacitance to match 50-ohm transmission lines is 320 pF (at least 100 pF of which should be variable).

gamma rod which is 60 feet long. Since a 43-foot tower with a Yagi represents an electrical height near 90 feet, a 60-foot gamma rod is obviously an impossibility. The use of a shorter gamma rod and more parallel capacitance *may* provide a match to 50 ohms, but in this case an RF bridge and graphical solution will save a lot of time.[3]

Note that for towers with electrical heights near 53 and 70 feet, a gamma rod approximately 20 feet long will provide operation on both 80 *and* 40 meters (the rod is about a quarter-wavelength long on 80 meters, one-half wavelength long on 40). For operation on both 80 and 160 meters, a similar coincidence occurs for towers which are electrically near 110 and 135 feet high. In this case, a gamma rod approximately 40 feet long will provide operation on both bands. In either of these dual-band systems, adjustments of the parallel tuning capacitor, C_p, will compensate for differences from the

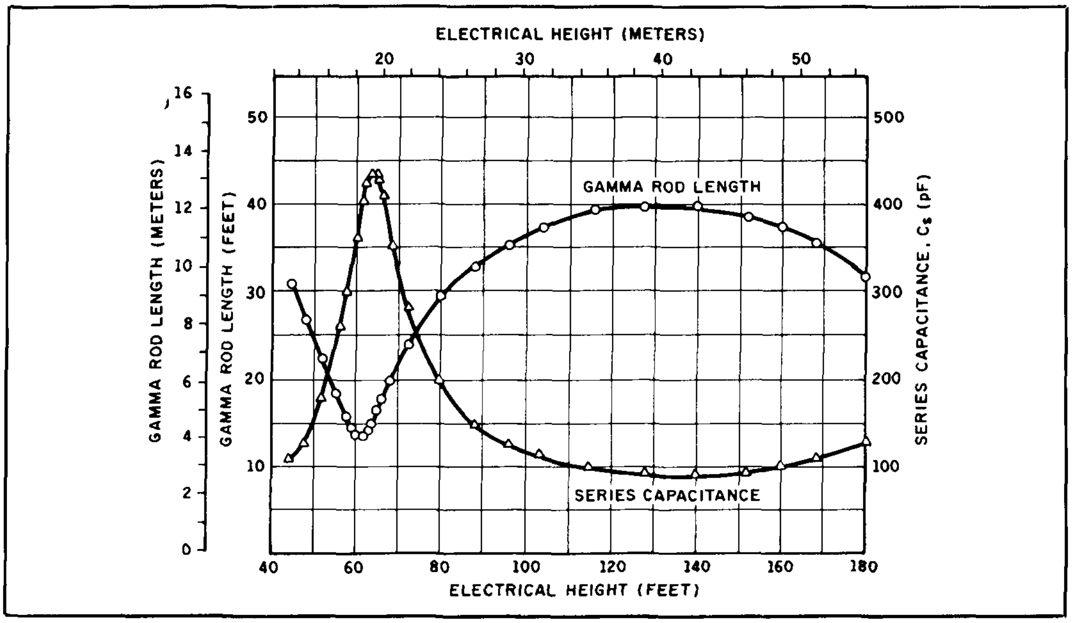

Figure 4. An 80-meter vertical. Gamma rod length and series capacitance versus electrical height of tower. Recommended parallel capacitance to match 50-ohm transmission lines is 650 pF (at least half should be variable).

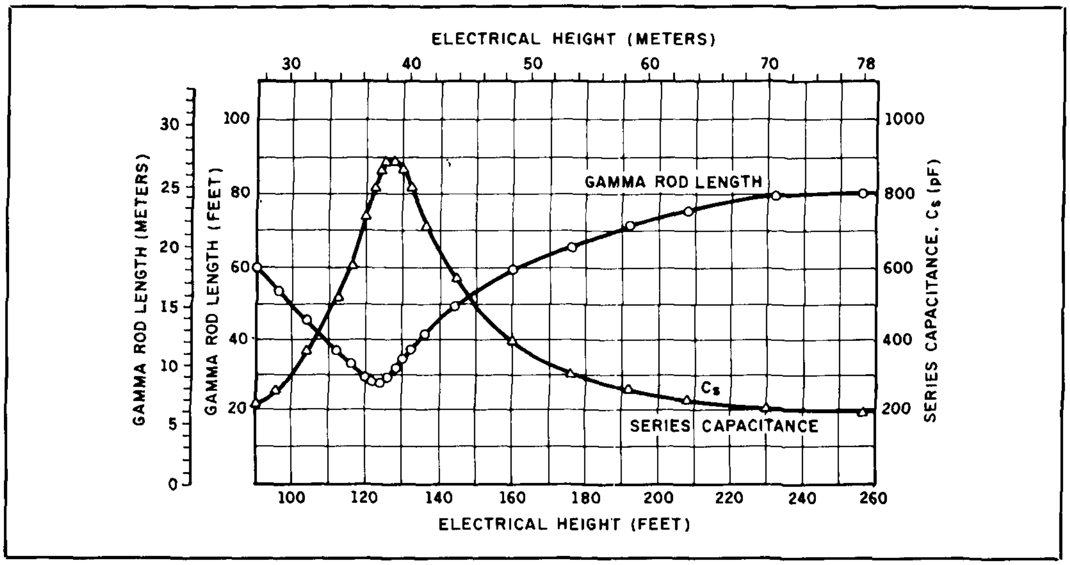

Figure 5. A 160-meter vertical. ***Gamma rod length and series capacitance versus electrical height of tower. Parallel capacitance required to match 50-ohm transmission lines is approximately 1300 pF.***

specified gamma rod length.

The electrical height of towers higher than 120 feet can be extrapolated by adding about 35 feet for a three-element 20-meter Yagi with a quarter-wavelength boom. Add about 45 feet of electrical height for a multi-element beam like the Hy-Gain TH6DXX. A two-element 40-meter beam adds 50 to 60 feet. Although cubical quads add about 25 feet, multi-element quad designs add little more because the elements are well away from the top of the tower and insulated from it.

Matching capacitors

Because the reactance of the series capacitor, C_s, is quite large except in those cases where the tower is approximately a quarter-wavelength high, this capacitor should have a breakdown rating of about 1000 volts for transmitters up to about 200 watts output. For transmitter powers of 2000 watts, this capacitor should have a breakdown rating of 5000 volts or more.

The parallel matching capacitor, C_p, does not require such a high voltage rating unless excessively high VSWR is expected at full power. For a 200-watt transmitter, an old style BC capacitor with 700 to 1000 pF maximum should work nicely. For 2000 watts PEP, the parallel capacitor should have a rating of 1500 volts minimum with current-carrying capacity of seven amperes.

Where large capacitance values are recommended, it is suggested that at least half be variable with the rest made up with fixed padding. Note that *both* the stator and rotor of the series capacitor must be isolated from ground.

The ideal matching network for this antenna system would use two vacuum-variable capacitors. These capacitors are not seriously affected by humidity or changes in barometric pressure, and they can be connected to small geared motors so they can be controlled remotely from the operating position. A 300-pF vacuum variable rated at 7500 volts, and a 1000-pF vacuum variable with a 2000 volt rating should handle practically any legal amateur transmitter with low VSWR.

A remote-control system that I have used for several years is shown in **Reference 1.** It's obviously a lot easier to remotely control the matching system from your hamshack than it is to traipse out to the backyard in snow,

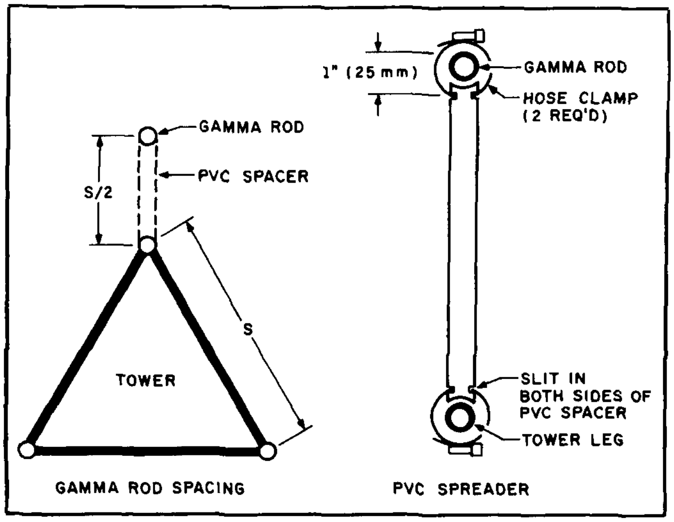

Figure 6. Construction of the shunt-feed system for grounded vertical radiators. The spacers are made from PVC water pipe.

sleet, and rain each time you want to shift your operating frequency.

Construction

A typical gamma rod installation is shown in **Figure 6**. On my vertical antenna, the gamma rod is mounted with PVC insulators spaced about 10 feet apart. The insulators are made from 1-inch diameter PVC water pipe. The movable shorting bar is made from the same material as the gamma rod.

To attach the PVC insulators to the gamma rod, first notch the ends so one end fits around one leg of your tower, the other end around the gamma rod. Then cut half-inch long slits on each side of the PVC pipe, about 1-inch in from each end (see **Figure 6**). Stainless-steel hose clamps are run through the slits in the PVC pipe and around the vertical member.

If you wish, the same tower may be used on more than one lower-frequency band—simply install gamma rods on more than one leg of the tower. You can use separate capacitance matching systems or remotely controlled vacuum-variables, depending on your operating requirements. A vertical tower antenna system which I use successfully on both 40 and 80 meters is described in **Reference 1**.

Ground requirements

Remember that the vertical element is only one-half of a vertical antenna system—the vertical element must operate against a good ground plane or the ground losses will be so high that the antenna performs poorly. The so-called ideal ground system consists of 120 equally spaced, quarter-wavelength radials, but even such an elaborate ground plane as this still introduces 2 ohms of series loss resistance into the total radiation resistance.

SOME COMMENTS ON IMPLEMENTATION

In terms of theory, there's little to add to John True's article. He discusses 160, 80, and 40-meter operation. Many amateur tower mounted Yagis aren't easily converted into efficient 40-meter verticals because they are electrically too long. However, they can be made very effective on 160 and 80 meters. I hope the implementation hints and conceptual observations which follow will permit a better appreciation of True's article.

An efficient vertical radiator for 80 and 160 meters

Photo A shows a fifty-foot tower with a Yagi and the two feeds that transform it into an efficient vertical radiator for 80 and 160 meters. The shorter feed design for 80 meters is faithful to True's article. For 160 meters, a larger spacing performed better and warrants further discussion.

The theory for shunt-fed antennas is complicated by practical geometries, so the True method of measuring the driving point impedance is a practical and semi-empirical approach. True visualized the antenna as a gamma-matched radiator. Viewing it as a folded monopole permits a better appreciation of its potential for increased driving-point impedance and broader bandwidth.

The traditional vertical is a half dipole where the missing half is an image below ground. The image doesn't really exist, but it is an analytical device that solves the electromagnetic field boundary conditions along the surface. Field computations above the surface give the right answers. The driving-point impedance is half that of a full dipole. A folded monopole is one side of a folded dipole mounted vertically over a good ground. The image concept still applies. The

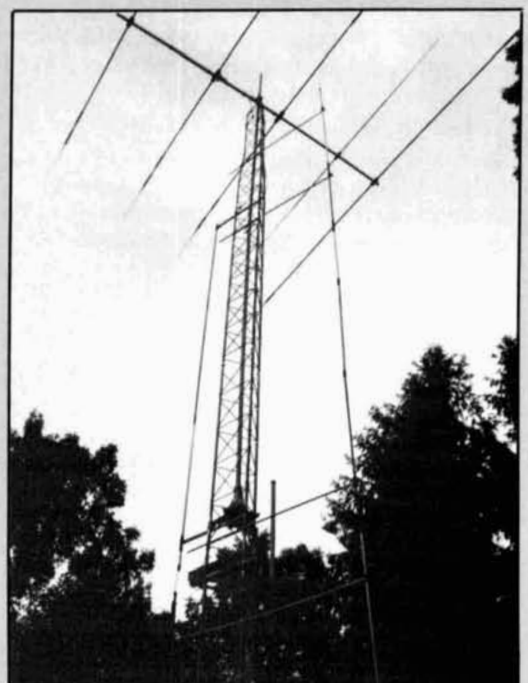

Photo A. A 50-foot tower with a Yagi and the two feeds that transform it into an efficient vertical radiator for 80 and 160 meters.

Since short vertical antennas are characterized by relatively low radiation resistance, ground resistance is higher, proportionately, than it is with vertical elements which are a quarter-wavelength or more. A complete discussion of ground system requirements is contained in **Reference 4**. For more information on short vertical antennas, consult the excellent series of articles by W2FMI.[5,6,7,8]

The tower which you use to support your higher frequency antennas can easily be used as a practical antenna system for 40, 80, and 160 meters. The graphs presented here will help you to design the necessary shunt-matching system, but note that since conditions vary from one location to another, some adjustments will be necessary to obtain a low VSWR. However, with an SWR bridge installed near the base of the vertical (very short leads), alternately adjust the series and parallel tuning capacitors until the reflected power approaches zero. If the amount of parallel capacitance for low VSWR seems excessive, make the gamma rod slightly longer.

The setting of the series capacitor is rather critical because reactance changes sharply near zero so it may take several tries before you can get the capacitor set exactly right. However, with a good ground system, the shunt-fed grounded tower can provide a very efficient antenna system for relatively little cost. ■

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intrinsically higher driving-point impedance of the folded dipole is halved in the folded monopole configuration, but it's still much higher than that of the traditional vertical. This is a good beginning for more efficient vertical antennas. A folded monopole is nothing more than a shunt-fed vertical.

In **Photo A**, the vertical folded monopole concept is more evident with the larger 160-meter feed. The side-mounted conductor extends virtually the entire length of the tower. It is electrically connected to the tower and the Yagi ground at the top. It's insulator isolated at all other points. Because this particular tower folds down, there's a copper braid hinge for the feed at the fold point.

I decided on the larger spacing for 160 meters after reading a review of the folded dipole and monopole.[1] The published literature suggested possibilities for further increasing the driving-point impedance via spacing and the relative size of the two vertical members. To realize this, I used a 1-inch feed line and the varied the spacing around four to five feet. The best spacing was not a strong function, but five feet was definitely better than the smaller spacings seen in many publications, including the True article.

Superimposed measured plots of the 160 and 80-meter verticals and a horizontal dipole's SWR versus frequency plot are given in Figures 1A and B. The folded monopole bandwidth is much broader than that of a simple horizontal dipole. Please note that these folded monopole SWR curves are for a single setting of the tuning capacitor. The low SWR portion of the curve can be moved easily over the entire band using the remote tuning feature. I tuned the folded monopole to the center frequency of existing station dipoles, which were at fixed frequencies, and then made comparisons.

True's implementation used two motor-driven capacitors for the matching network connected to a single feed. My implementation needed only one motor-driven capacitor per feed. During integration it's wise to use two variable capacitors for experimentation. If the power is kept low, simple receiver-grade components can be used temporarily.

Photo B shows the homebrew 80-meter matching unit. **Figure 2** is the circuit sche-

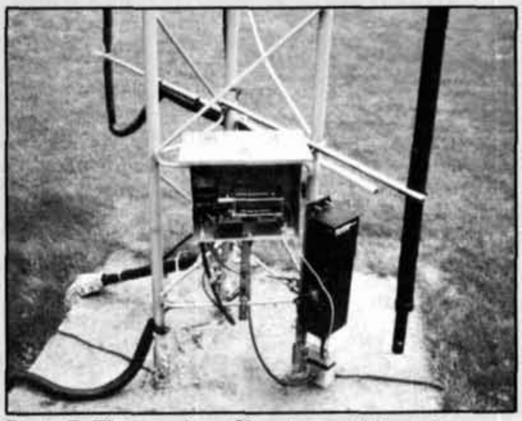

Photo B. The homebrew 80-meter matching unit.

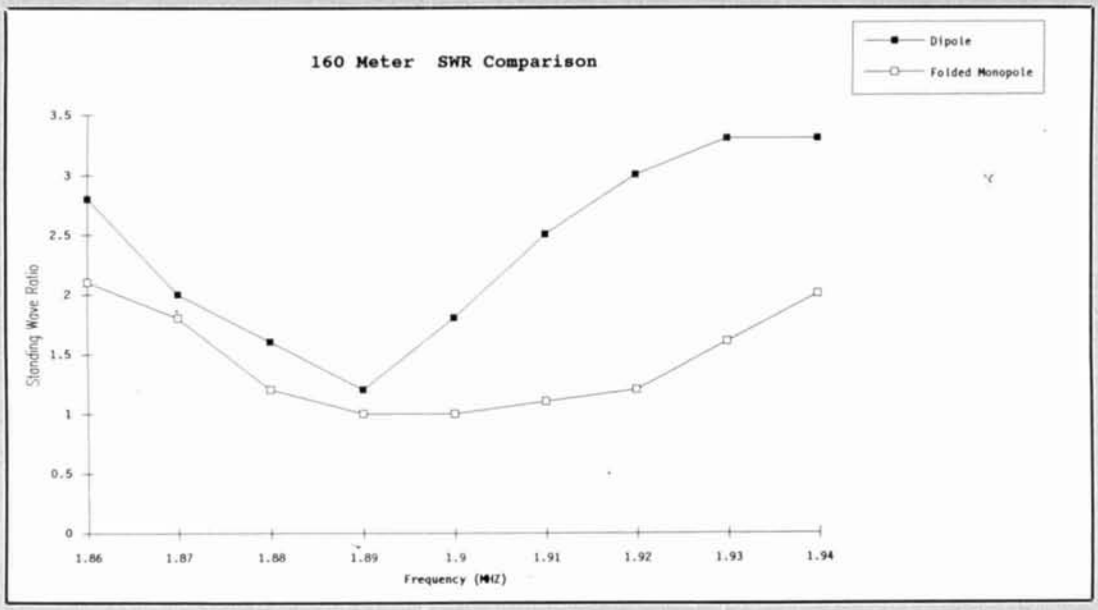

Figure 1A. 160-meter SWR comparison.

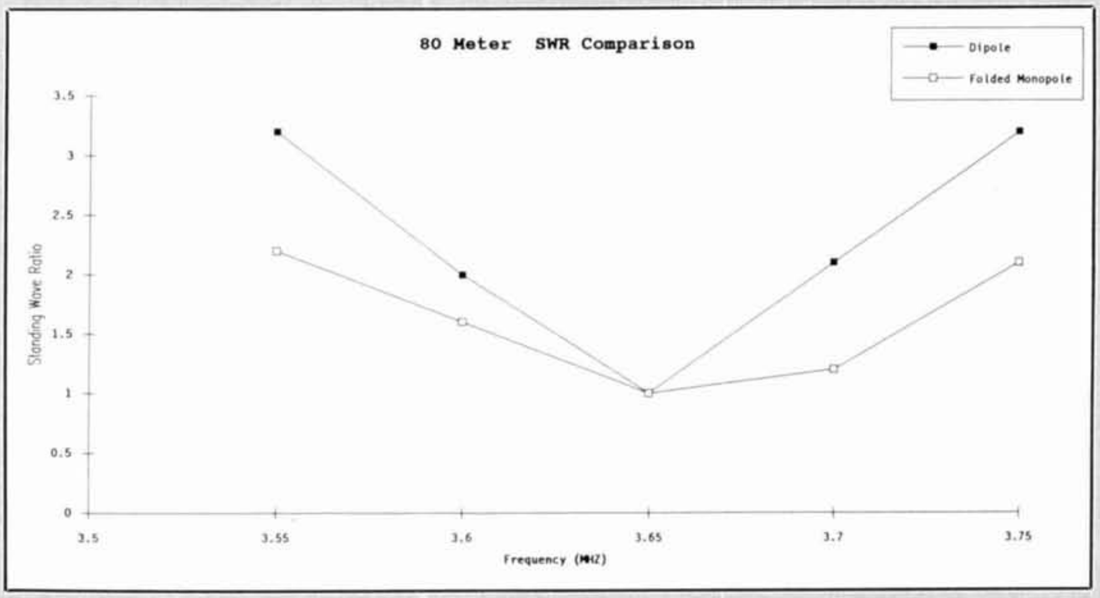

Figure 1B. 80-meter SWR comparison.

matic. This tuning unit could be switched between the two feed lines. I used two separate tuners because parts were available and the tuning ranges could be better optimized for each band. A single switched remote power supply is used selectively for either feed line tuner. A dedicated coax feed for each band goes back to the operating position.

The series capacitor must be motor driven. Both sides of the capacitor must be high voltage isolated from ground. For the 80-meter feed, I reduced an old National 1000 pFd capacitor to a little over 100 pFd by removing plates. The wide spacing is important to withstand the high RF voltages which True describes. It's driven by a reversible 12-volts DC gear motor. The motor is controlled by reversing the polarity of DC superimposed with the RF drive. The control signal is isolated by RF chokes and capacitors. Many popular remote switching units use this methodology. These tuners are simple to build. The Heath unit I used, came after the fact; unfortunately, it is no longer available. The only real construction problem I experienced was in providing some ventilation for drying, draining condensation, and keeping out nesting wasps. The output wire must have several thousand volts insulation.

All isolation and tuning capacitors must be rated for very high RF currents (5 + amps). High voltage transmitter-grade micas or wide-spaced variables are preferred. The 160-meter tuner uses a Heath remotely controlled capacitor that I picked up as a half-price discontinued item. It has an approximately 450-pFd variable capacitor with an acceptable voltage rating. The tuning range from 1.8 MHz to 2.0 MHz requires only about one half of this capacitance range. For 160 meters, a discrete parallel capacitor was required. It is 1000 pFd and is housed in a mini-box that feeds the Heath unit.

No parallel capacitor was required for the 80-meter implementation. At 160 or 80 meters, the feed was 1/6 or 1/3-wavelength of RG-8 coaxial feed. I realize that the complete validity of a transmission-line concept may be questionable for such a short line. However, one could reason it as a discrete circuit equivalent. A 100-foot feed with a shunt capacitance of about 30 pFd per foot represents a sizable parallel capacitance. The required additional parallel capacitance can be found with a little experimentation. For the geometry used, a single fixed-value capacitor sufficed. I chose zero for 80 meters and 1000 pFd at 160 meters. The cable shunt capacitance is the same for both bands.

The feed has high voltage on it during transmission. Physical contact with it could inflict an RF burn. The bottom of both feeds was encased in PVC tubing. This is a lossy dielectric, so only use it where it's necessary. Good electrical connections are important. Just clamping aluminum rods together won't provide sufficient protection from time and weather. Bolt or screw polished pieces together and protect the joints with tape. The top of each feed is connected to the tower with a heavy supplementary copper wire using solder lugs and galvanized wire cable clamps. Use the same technique to connect the Yagi ground to the tower. Don't forget to provide a service loop that allows for Yagi rotation.

As with all vertical antennas, radials are important. Put in as many quarter-wave radials at 80 and 160 meters as you can. Insulated hook-up wire held tight to the ground with short "hairpins" made of welding rod work fine. After a few weeks during the growing season, everything will disappear under the turf.

Tuning

Tuning is easy. It's best to use a noise bridge or one of the newer SWR indicators.

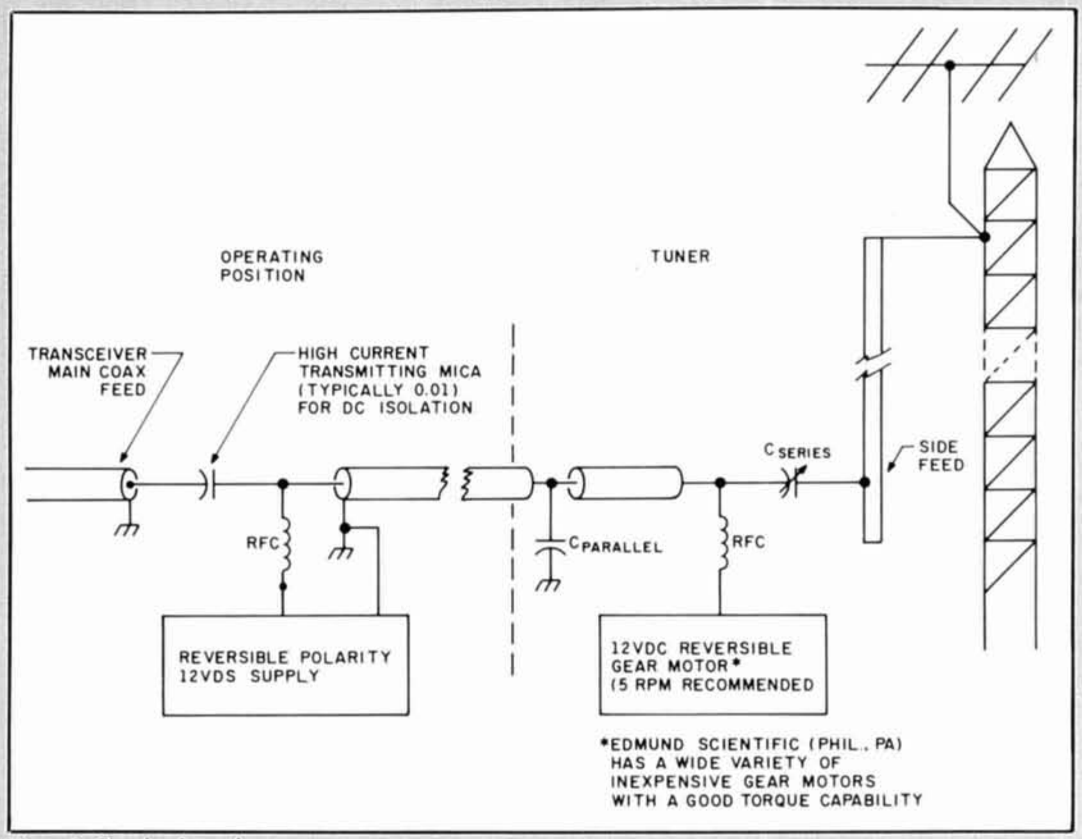

Figure 2. Circuit schematic.

This minimizes interference. As a last resort, key the transmitter and activate the motor control to minimize reflected power. At the 160-meter band edges, the worst SWR is 1.3 after tuning. For most of the 160-meter band, the highest SWR was 1.0. This is a consequence of using a single value parallel capacitor on 160 meters. A motor-driven parallel capacitor would be a small but questionable improvement. For the 80-meter band edges, the maximum SWR was 1.0 after tuning. I used some padding capacitors to exactly center the 80-meter band series capacitor tuning. These are visible in **Photo B**.

A few final words

Operationally, this is a long-range antenna. For local ragchewing, a horizontal dipole close to the ground will work better because of its almost vertical angle of radiation. As you listen comparatively to increasingly longer range stations, the vertical will show its superiority. Turning on your transmitter will make you a convert.

The lower bands have a lot of atmospheric noise, especially in the summer. Supplementary Beverage receive-only antennas can help. At any time, noise is less of a problem for RTTY/AMTOR enthusiasts. It's puzzling why there aren't more of these signals on 160 meters. These particular digital modes have a bit processing advantage over SSB or packet, and they cause less interference to the user community.

Hunter Harris, W1SI
Communications Quarterly
Editorial Review Board

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PRODUCT INFORMATION

Frequency Counter Has 1 mV Sensitivity

Optoelectronics Inc. has a new frequency counter with full-range operation, 1 mV sensitivity, and resolution to display 10 Hz in 3 GHz.

The Model 2810 frequency counter detects frequencies up to 200 MHz and resolves them to 1 Hz in 1 second. Frequencies up to 900 MHz are resolved to 1 Hz and displayed in 4 seconds. Frequencies up to 3 GHz are resolved to 10 Hz and displayed in 1.6 seconds. A front panel selector switch lets you choose four different gate times (0.01 second, 0.1 second, 1.0 second, and 10 seconds).

Sensitivity is specified at better than 1 mV for signals up to 200 MHz, and better than 10 mV for signals for up to 2400 MHz. The unit is operable up to and beyond 3000 MHz.

Input antennas or lab probes connect to either of two pushbutton-selectable BNC connectors, providing either high input impedance (1 meg) for signals up to 50 MHz, or standard 50 ohms input impedance for signals from 1 MHz to 3000 MHz and beyond.

The Model 2810's time base is a 10-MHz crystal oscillator. Options include a ± 2 PPM temperature-compensated crystal oscillator, second NiCad battery pack, and LCD backlight.

For more information on the Model 2810 frequency counter, contact Optoelectronics Inc., 5821 NE 14th Avenue, Fort Lauderdale, Florida, 33334.

Service Equipment Catalog Available

The new 1991 Electrical/HVAC-R Service Equipment Catalog is now available from John Fluke Mfg. Co., Inc. The 15-page catalog features information on Fluke multimeters, thermometers, and accessories for the electrical service industry.

The catalog is available free of charge from electrical wholesalers nationwide. For further information write to John Fluke Mfg. Co., Inc., P.O. Box 9090, Everett, Washington 98206, or call (800)44-FLUKE.

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Brookline, Massachusetts 02146

A COMPUTER-BASED SIGNAL MONITORING SYSTEM

Putting the Commodore-64 to work as an analog-signal monitor and recorder

The measurement, categorization, and logging of the various signals involved in communications are important aspects of amateur radio. Most of us wouldn't think of operating a station without adequate instrumentation for determining received signal strength, transmitter output power and VSWR, or power-supply current and voltage. And all of us, at one time or another, have wished for some means of logging these values automatically. An automatic logging system would certainly come in handy during a contest, in the midst of a complex OSCAR link, or when attempting to diagnose a piece of communications gear. Here's how the analog-to-digital conversion hardware in an inexpensive microcomputer, the Commodore-64, can be used as a general-purpose analog signal monitor and recorder.

Introduction

Amateur radio, at least from an operational perspective, is about signals—signal generation, detection, measurement, manipulation, and logging. In most cases, we work directly with the transceivers, test equipment, and other hardware involved with the signals in question. But in many circumstances, it would be desirable to delegate some of this signal-handling work to an intelligent assistant, like a microcomputer.

When contesting, it would be a great time saver to have a microcomputer measure and record the signal strengths (along with the time, etc.) of stations worked. It would also be helpful to have an automated alarm system to warn of such problems as excessive plate current in the final amplifier, or excessive VSWR in the antenna system. In some situations, microcomputer-based signal monitors may be the only means available for acquiring certain data. For example, tracking radio beacons on a minute-by-minute and day-to-day basis would be an insurmountable task for a human operator.

For these and other reasons, microprocessor-based signal monitoring is a desirable feature to have as part of any amateur station. However, despite the great flexibility of general-purpose, digital microcomputer systems, they are typically ill-equipped to handle analog signals. But the incompatibility between the analog and digital representations of data need not be an insurmountable problem. A wide variety of analog data capture systems for microcomputers are available.

The better microcomputer-based data acquisition systems, sold as multi-channel analog-to-digital (A-to-D) converters or digitizers, can cost far more than a modern HF transceiver. Luckily, you can purchase single-channel, low resolution A-to-D converters for the Apple Macintosh, Tandy, and

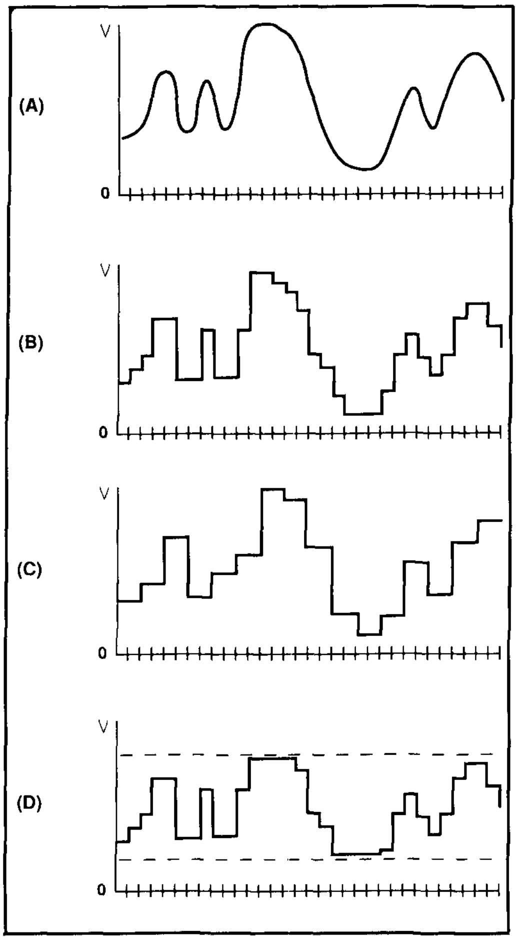


Figure 1. Amplitude (x-axis) versus time (horizontal axis) tracings that illustrate the trade-offs associated with A-to-D conversion. (A)The original analog signal. Note the smooth transitions in signal amplitude. (B)A digitized version of the original signal, with a sample rate of one arbitrary time unit. Note the abrupt changes in amplitude with time. (C)The digitized waveform with a sample rate of two time units. This waveform bears little resemblance to the original analog signal, especially when the higher frequencies are compared. (D)This digitized waveform illustrating the effects of clipping. In this example, the dynamic range of the original analog signal exceeds that of the digitizer (the dynamic range of the digitizer is indicated by the dashed lines). Note the distortion caused by clipping, even though the original waveform was digitized at a sample rate of one arbitrary time unit.

IBM-PC computers.* If you prefer the home-brew approach, you can build your own without much effort.[1] While these relatively inexpensive microcomputer peripherals may be appropriate for digitizing audio signals, there's a better approach for low-frequency audio (less than 100 Hz) and DC. Many of the inexpensive microcomputers, such as the Commodore-64, come factory-equipped with simple but effective A-to-D conversion circuitry. But before I consider the specifics of the C-64, a brief discussion of A-to-D conversion is in order.

A-to-D conversion techniques

Of the wide variety of techniques available for analog-to-digital signal conversion, the most common include parallel encoding, successive approximation, voltage-to-frequency conversion, and single-slope integration.[2] Each of these basic techniques has its own advantages and limitations. For example, the technique of parallel encoding, in which a signal is fed simultaneously to a bank of comparators, each connected to equally spaced reference voltages, is the fastest conversion technique available. It's also the most expensive, at several hundred dollars per converter.

The popular successive approximation technique, while prone to nonlinearities, is relatively fast, accurate, and inexpensive. This technique makes use of a digital-to-analog converter to generate a voltage that's sent to a comparator circuit. When the comparator output is zero, signifying that the D-to-A converter output voltage is the same as the voltage to be digitized, the digital value input to the D-to-A converter is taken to be the digitized value of the unknown voltage.

Voltage-to-frequency techniques rely on the input voltage to control the frequency of a pulse train, with frequency proportional to input voltage. The output frequency is then measured to determine the input voltage. This method, while only moderately accurate, is very inexpensive.

Single-slope integration techniques, while not very accurate, offer good resolution and a very simple design. In this technique, a ramp generator is started at the same time as a timer circuit. When the unknown voltage level is equal to that of the ramp generator, a comparator circuit stops the ramp and timer. The timer value is proportional to the input voltage.

A-to-D conversion limitations

Regardless of the digitization technique used, you should keep three limitations of A-to-D conversion in mind: the sampling frequency of the digitizer, quantization noise, and the dynamic range of the digitizer system. I'll describe these limitations in more detail.

•**Sampling frequency**. A typical analog waveform will change in amplitude, frequency, and shape over time (**Figure 1A**). To store such a signal in a digital computer, dig-

*For the Macintosh try "MacRecorder," Fallon Computing, Berkeley, California. For Tandy and IBM-PC computers contact Covox, Inc., 575-D Conger Street, Eugene, Oregon.

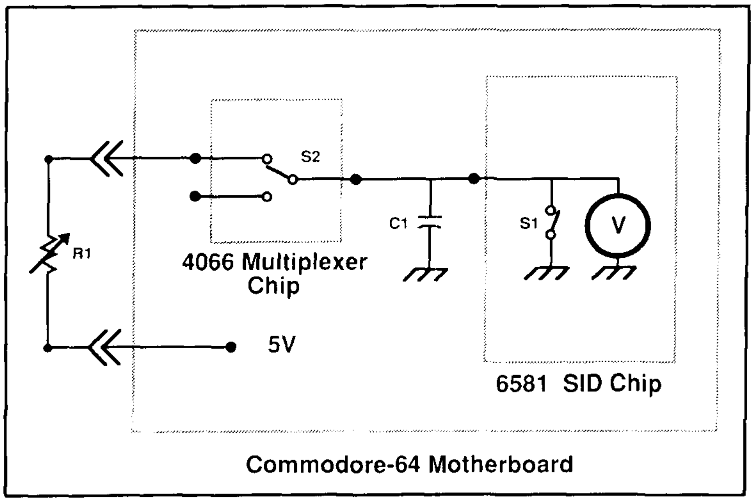


Figure 2. A simplified diagram of a single channel of the basic D-to-A conversion hardware on the C-64 motherboard. A 1000-pF capacitor (C1) is repetitively charged through a variable resistance (R1) and discharged by the equivalent of a switch (S1) in the SID chip. The maximum voltage across capacitor C1 is inversely proportional to the variable resistor (R1) value. The multiplexer chip switches between two external resistance connections to provide an additional channel of data input (switch S2 simulates this switching, which is synchronized with switch S1). A second pin on the SID chip similarly supports two additional data input channels.

itizers take samples of the waveform at evenly spaced intervals and store these amplitude values in memory (**Figure 1B**). If the signal is sampled often enough, the waveform can be reconstructed from the samples (as in a digital-to-analog converter). If the signal is sampled too slowly, it will be impossible to reconstruct the original analog signal (**Figure 1C**). The Nyquist criterion states that the minimum sampling frequency is at least twice that of the signal to be sampled. Note that the Nyquist criterion assumes a perfect system. In reality, the sampling frequency should be three or four times the highest frequency to be digitized.[1]

When the sample rate is less than double the signal frequency, the reconstructed signal will have a lower frequency than the original. The higher frequencies in the original signal appear as lower frequencies in the reconstructed signal—a phenomenon known as aliasing (see **Figure 1C** for an example). Aliasing can be a significant source of noise in A-to-D conversion. One way to avoid aliasing is to use a low-pass filter that cuts out frequencies greater than half of the sampling rate.

•**Quantization noise**. Most of the inexpensive digitizer systems convert analog data to an 8-bit format. That is, the sound amplitude is restricted to a range of integers from 0 to 255. Smooth waves, even when digitized at high sampling rates, become jagged (**Figure 1B**). The effect is to introduce "quantization noise" in the digitized waveform. Quantization noise sounds like the high frequency hiss in a cassette deck without Dolby. It's possible to minimize quantization noise by using a 12 or 16-bit digitizer, coupled with modifications in the computer system's hardware and software.

•**Dynamic range**. Restricting the representation of the signal amplitude to integers between 0 and 255 limits the dynamic range (the difference between the largest and smallest signal that can be recorded) of the digitized signal. With an 8-bit system, the dynamic range approaches 48 dB. Signals that fall outside the dynamic range of a digitizer are simply clipped (see **Figure 1D**). Clipping can be avoided by restricting the dynamic range of the analog signal to be digitized.

A-to-D conversion with the C-64

The Commodore-64 A-to-D conversion circuitry, set up to implement a variant of the single-slope integration technique, is contained within the 6581 sound interface device (SID) chip. While mainly concerned with sound generation and modulation, this chip contains two A-to-D potentiometer interfaces.[3] These interfaces are provided for use with game paddles and as front-panel controls for a music synthesizer. Each potentiometer interface allows the microprocessor to read 8-bit values from 0 (minimum resistance) to 255 (maximum resistance). The value is always valid, and is updated every

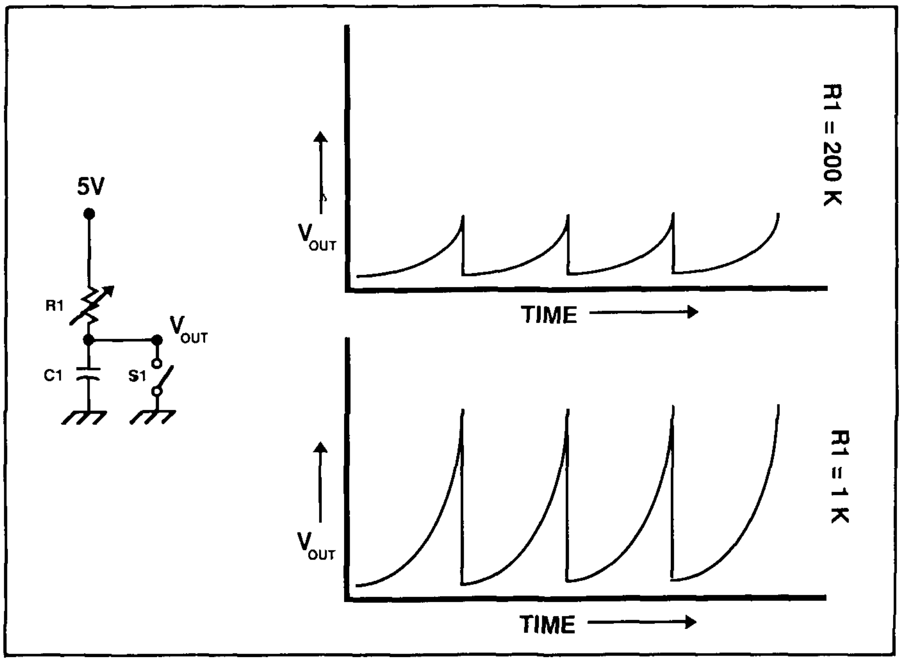


Figure 3. A simplified view of how the C-64 accomplishes D-to-A conversion. Assuming switch S1 closes momentarily every arbitrary time unit, the voltage across C1 (V_{out} will be inversely proportional to the value of variable resistor R1. Circuitry within the SID chip converts the maximum voltage value to an equivalent digital value, presumably through a variant of single-slope integration.

512 system clock cycles (nominally 1.0 MHz).

The A-to-D conversion processes supported by the SID chip are based on the rate of charge of a 1000-pF capacitor tied from a pin on the SID chip to ground (**Figure 2**). In normal operation, the capacitor is charged through a 200-k potentiometer tied to 5 volts DC. Every 512 clock cycles, or about 2000 times per second, the voltage across the capacitor is read and the capacitor is discharged. In this simple RC circuit, the maximum voltage across the capacitor during each cycle is inversely proportional to the resistance of the potentiometer (**Figure 3**). A separate potentiometer and capacitor are required for each of the two A-to-D converters supported by the SID chip.

To provide simultaneous support for two players, or two game-paddle pairs, the A-to-D converter pins of the SID chip are connected to the two control ports of the C-64 through a 4066 multiplexer chip (**Figure 2**). The 4066 rapidly switches connections between the two control ports and the SID chip. The switching rate is precisely timed so that the SID chip knows which of the control ports it's connected to at any given instant. Because each of the A-to-D converters within the SID chip are active every 512 clock cycles, each control port is effectively sampled only every 512 × 2, or 1024 clock cycles, or about 1000 times per second. This limitation, imposed by the C-64 hardware, represents the maximum possible sampling frequency. As described in the next section, the actual sampling frequency is normally limited to about 100 Hz due to non-hardware factors; for example, software execution speed.

The hardware

To use the basic A-to-D conversion circuitry within the C-64, you need some means of transforming the analog voltage to be monitored into a resistive equivalent. The simplest approach would be to use the analog voltage to bias a transistor that's used in place of a game-paddle potentiometer. Although this configuration might be sufficient for some applications, it would be inadequate if equipment isolation were a concern, or if the voltage source couldn't supply the required current. For example, if the signal to be monitored is taken from the back of a panel meter, you wouldn't want the current through the meter to change as a result of shunting current through the transistor circuit.

A better approach is to use an operational amplifier together with an optocoupler (**Figure 4**). The operational amplifier provides a high-input impedance, and therefore requires negligible current from the signal source, while the optocoupler provides equipment-computer isolation. The optocoupler not only minimizes the possibility of intra-equipment interaction due, for example, to the coupling of input and output signals sent to the same digitizer, but it also minimizes the coupling of computer signals into your communications equipment, and vice versa. The optocoupler also protects your computer against voltage spikes, high-voltage signals, and other disruptions that could destroy the computer hardware. It's much easier to replace an inexpensive optocoupler than it is to replace the CPU.

I have included a parts list for the two-channel system described in **Figure 4**. All of the components, with the exception of the custom circuit board, can be purchased from Radio Shack. The individual components of the system are described in more detail in the sections which follow.

Operational amplifier

The op amp used in this project, the TL092 (RS part no. 276-1746), is a dual N-FET device that has a very high input impedance (10^{12} ohms). This inexpensive op amp has an added benefit of requiring only one supply. As **Figure 4** shows, the TL092 is used as a transconductance amplifier, or voltage-to-current converter. Using the top half of **Figure 4** as a reference, the output voltage of each half of U1 (pin 1) and the LED current are related to the input voltage, V_{in}, by the following equation:

$$V_{out} = [V_{in}(R2 + LED\ \Omega + R3)]/R3 \quad \mathbf{(1)}$$

and

$$I_{LED} = V_{out}/(R2 + LED\ \Omega + R3) \quad \mathbf{(2)}$$

Without the op amp, the optocoupler requires considerable current from the signal source for operation (**Figure 5**). With the op amp, the current requirements are for the most part due exclusively to the 1-meg resistor across the non-inverting input of the op amp. Although not strictly required for circuit operation, I found that without the 1-meg resistor, the op amp became unstable with no input signal. For a very good introduction to op amp fundamentals, see **Reference 4**.

Optocoupler

LED/phototransistor pairs, like the TIL-111 and the Monsanto MCA-2, typically provide several kilovolts of isolation, 10^{12} ohms of insulation resistance, and less than 1 pF of coupling between input and output.[5] You can purchase an assortment of optocouplers from Radio Shack (RS part no. 276-1654) for about two dollars. Included in

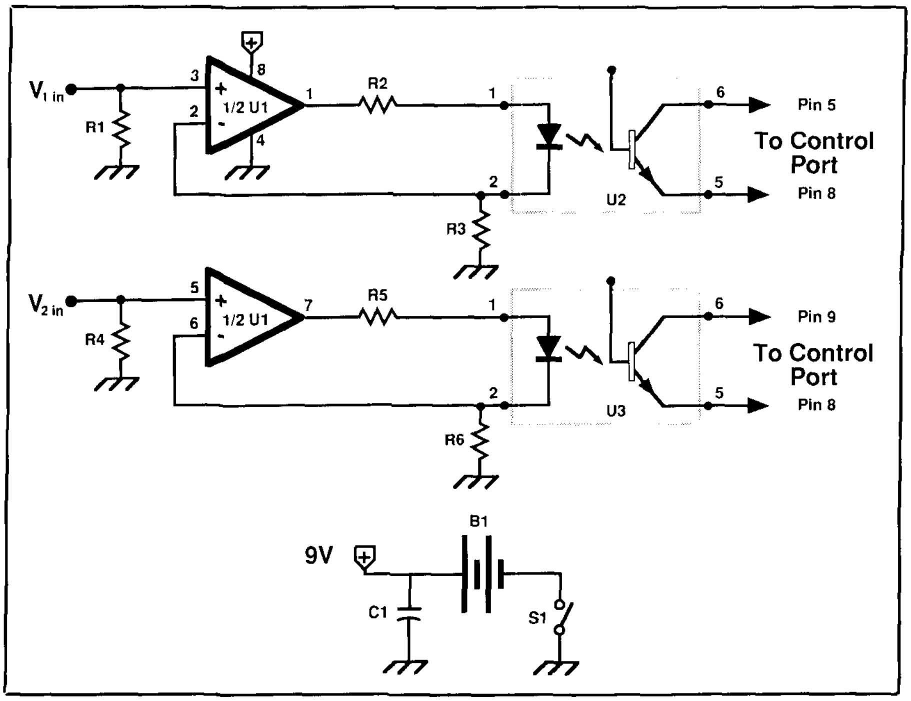


Figure 4. The schematic for a two-channel voltage-to-resistance converter for the C-64 (or other computer system that accepts resistive input devices). The resistors are rated at 1/4 watt, 5 percent tolerance. R1 and R4 are 1 meg, and the remaining resistors are 1 k. U1 is a TL092 op amp, and U2 and U3 are TIL-111 optocouplers. C1 is a 0.1-μF bypass capacitor.

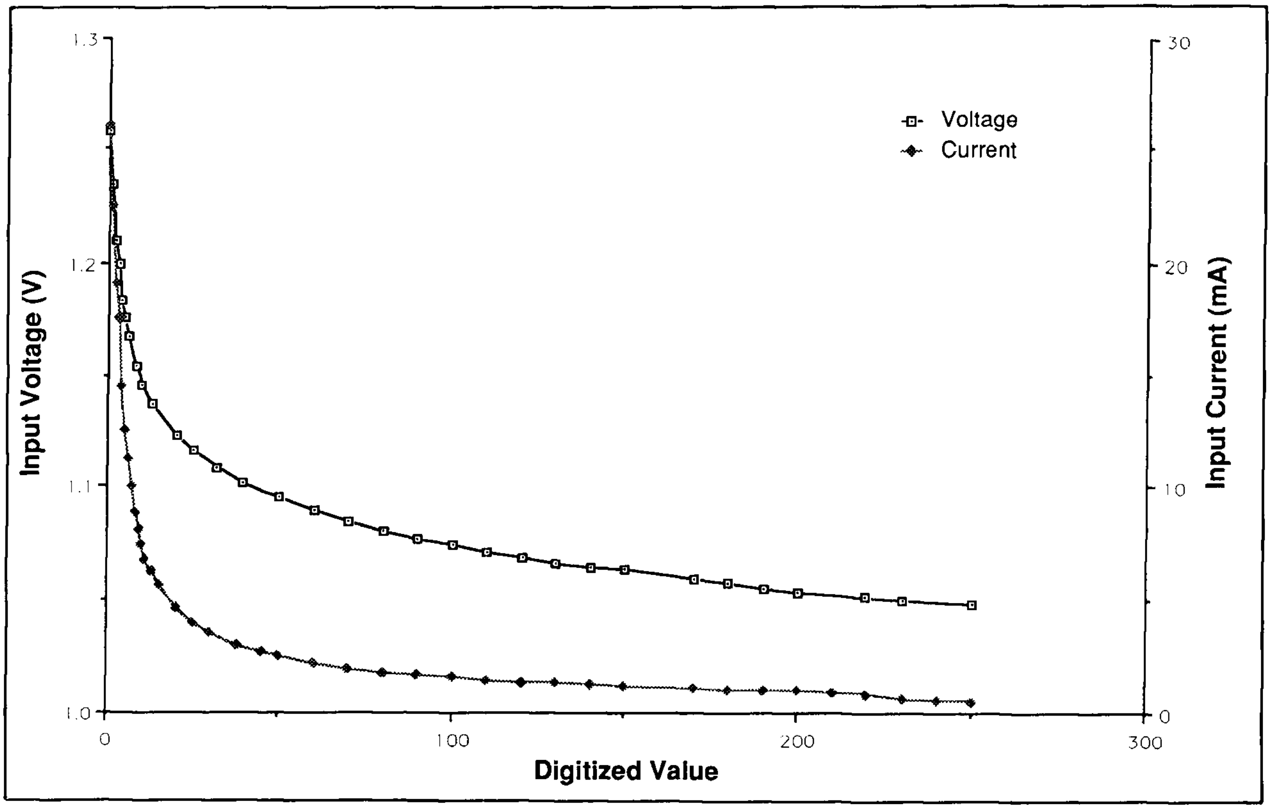


Figure 5. Input voltage and current versus digitized or paddle value for the TIL-111 optocoupler. The data were obtained by bypassing the op-amp input described in *Figure 4* and driving the LED directly. The operational range of the optocoupler is only a few tenths of a volt, and current drain from the source can be considerable.

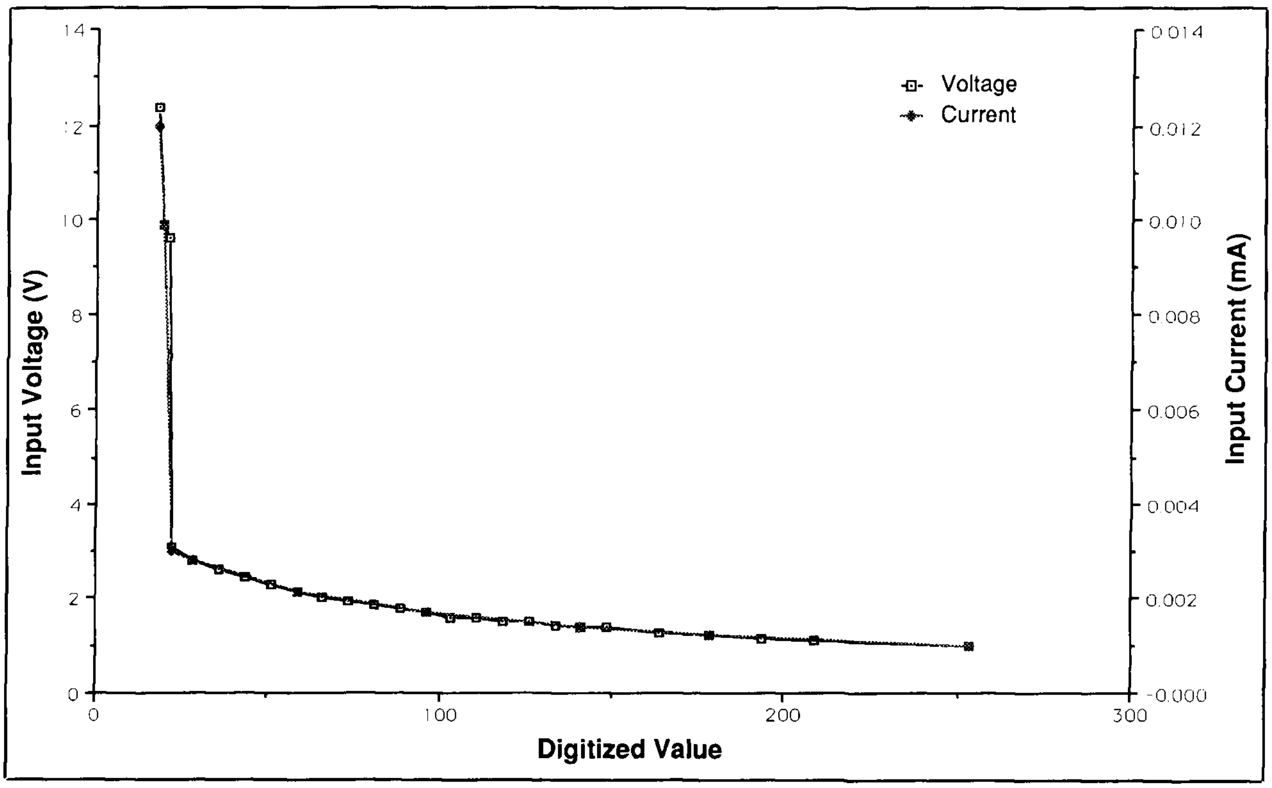


Figure 6. Input voltage and current versus digitized or paddle value for the complete voltage-to-resistance converter described in *Figure 4*. Notice that, compared to *Figure 5*, the operational range of the system has increased dramatically. The best input voltage range extends from about 1 to 3 volts, corresponding to digitized or paddle values from 35 to 250. The input-signal current requirements of the system are minimal, on the order of 0.002 mA.

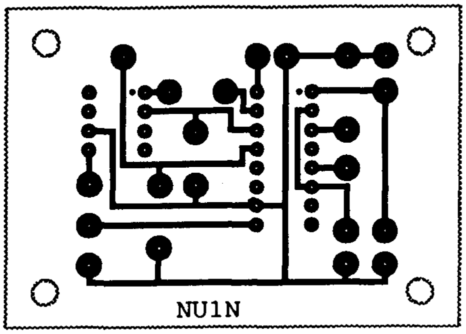


Figure 7. Circuit board artwork (foil side) for the voltage-to-resistance converter.

this assortment is the TIL-111, with an infrared emitting diode rated at 1.7 volts and 20 mA maximum. The isolation voltage is 1500 volts and the typical response time is 20 μs—more than adequate for our purposes.

As you can see in **Figure 4**, the phototransistor output takes the place of the game paddles in the C-64 system. The output of the TL092 drives the LED, which in turn modulates the conductance of the phototransistor. **Figure 6** shows the relationship between LED voltage/current and the paddle values read by the C-64. The base of the phototransistor isn't used in this circuit.

PC board

Although you can use perfboard or generic IC boards available from Radio Shack and other sources, I have included the artwork for a custom printed circuit board for this project **(Figure 7)**. This board makes for a nice, compact unit **(Figure 8)** that can be installed in your communications gear or used as a stand-alone device. Because each board will handle two input signals, you'll need two circuit boards (and two of each item listed in the parts list, with the exception of the power-supply components) to support four channels.

Cable

The easiest way to connect the coupler circuit to your C-64 is to use the six-foot joystick extension cable from Radio Shack. Simply cut off the connector designed to mate with the joystick, and use the wires attached to pins 5, 7, and 9 **(Figure 9)** to connect to the paddle inputs of the C-64 control ports. Although I have only used a handful of these cables, the color code seems to follow this standard: pin 5 (paddle AY or BY) is green, pin 7 (5 volts DC) is grey, and pin 9 (paddle AX or BX) is black.

The software

The BASIC code in **Listing 1** illustrates the procedure for reading the resistive values connected to the control ports. This routine simply reads the values from each of the four paddles and displays them on the screen. Because the resistive values on the C-64 aren't reliable when read from BASIC alone, the best way to use the A-to-D circuitry is to SYS to a machine language routine from BASIC and then PEEK the appropriate memory locations. The PEEK function returns an integer in the range of 0 to 255, which is read from the specified memory location.

In practice, you'd modify **Listing 1** to suit your particular needs. Instead of printing the resistive values, you might want to move a sprite or other graphic element across the screen, with its position a function of two or three resistive values. Power input to an amplifier could be displayed by measuring the input voltage and current, and then computing and displaying their product on the screen. It's also possible to modify the listing to accumulate multiple measurements, per-

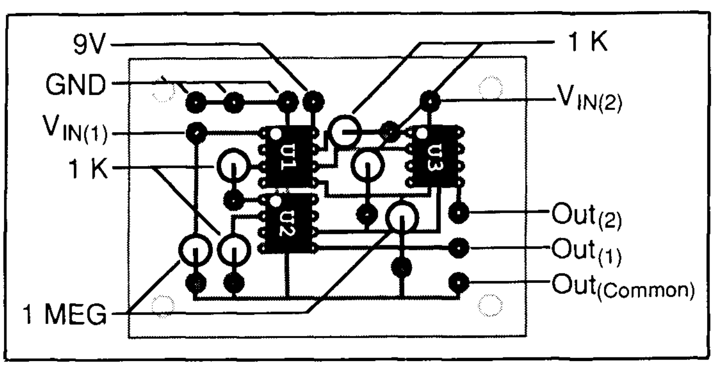
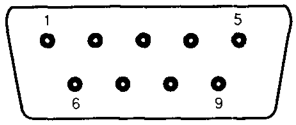


Figure 8. Component placement for the voltage-to-resistance converter. Note that U1 and U2 share a 16-pin DIP socket.



Pin No.	Function
1	Joystick A0
2	Joystick A1
3	Joystick A2
4	Joystick A3
5	Paddle AY
6	Button A
7	5 VDC 50 mA
8	GND
9	Paddle AX

Figure 9. Control port 1 of the Commodore-64. Pins 5 (paddle AY), 9 (paddle AX), and 7 (5 volts DC) are used in the D-to-A converter system. A second port on the Commodore, control port 2, provides two additional input channels, also through pins 5 (paddle BY), 9 (paddle BX), and 7 (5 volts DC).

```
100 REM READ C-64 CONTROL PORT PADDLES
110 REM DEFINE START OF PADDLE ROUTINE
120 C = 12 * 4096
130 REM READ IN MACHINE LANGUAGE DATA
140 FOR I = 0 TO 63
150 READ A
160 POKE C + I, A
170 NEXT
180 REM CALL MACHINE LANGUAGE ROUTINE
190 SYSC
200 REM PRINT PADDLE VALUES
210 PRINT "POT #1= ",PEEK(C+257)
220 PRINT "POT #2= ",PEEK(C+258)
230 PRINT "POT #3= ",PEEK(C+259)
240 PRINT "POT #4= ",PEEK(C+260)
250 REM WAIT A WHILE
260 FOR W = 1 TO 50
270 NEXT
280 REM CLEAR SCREEN AND GO HOME
290 PRINT "♥": PRINT: GOTO 190
300 REM
310 REM MACHINE LANGUAGE DATA
320 DATA 162, 1, 120, 173, 2, 220, 141, 0, 193
330 DATA 169, 192, 141, 2, 220, 169, 128, 141
340 DATA 0, 220, 160, 128, 234, 136, 16, 252
350 DATA 173, 25, 212, 157, 1, 193, 173, 26
360 DATA 212, 157, 3, 193, 173, 0, 220, 9, 128
370 DATA 141, 5, 193, 169, 64, 202, 16, 222
380 DATA 173, 1, 220, 141, 6, 193, 88, 96
```

Listing 1. A BASIC listing for the C-64 to display paddle values in an endless loop. After each resistive value is read (line 190) and displayed (lines 210 to 240) on the screen, the routine waits for about a second (lines 260 to 270) before clearing the screen (line 290) and repeating the process.

haps for averaging or plotting at a later time. For example, if you want to store a series of values, you could use the code in **Listing 2**. In this routine, 100 D-to-A conversions are performed in rapid succession, and the resulting resistive values are stored in Array A. The data in the array could be used to plot a waveform, as the basis for computation, or simply saved to disk for future use.

You might have noticed that in **Listing 2**, unlike **Listing 1**, there's no WAIT state imposed on the routine after each read. Instead, the main read loop, lines 160 to 190, is optimized for speed. (Actually, the main loop could be further optimized, for example, by removing the REM statement and linking the entire loop onto a single line). Even so, **Listing 2** provides only 73 samples per second. Based on the Nyquist criterion, the frequency of the signal to be measured by this routine shouldn't exceed 73/2 or 36 Hz. Extending **Listing 2** to record the remaining three control ports results in each resistive value being sampled at about 105 Hz, with a corresponding maximum frequency of 105/2 or 53 Hz. The seemingly paradoxical increase in sampling frequency, as a function of increasing the number of resistance values sampled, indicates that the overhead imposed by the looping routine is significant.

If your application requires an increased sampling frequency, you can probably realize a two-fold improvement by using a BASIC compiler, like BLITZ. If you want to approach the hardware limit of the C-64 (about 1000 samples per second), you'll have to make extensive use of low-level ASSEMBLER routines.

Use and calibration

After assembling the interface circuit I've described, your first step is to calibrate the system. That is, you will need to determine, for every analog signal value, the corresponding digitized value. Calibration is a simple task requiring only a battery or other low-voltage DC source, a potentiometer, and an accurate, high-impedance volt meter (**Figure 10**). With the voltage-to-resistance interface hardware in place, and the program in **Listing 1** running on your computer, vary the input voltage to the op amp circuit, and record the resulting digitized value displayed on the computer screen. Next, graph the relationship between input voltage and digitized value (as in **Figure 6**). In this manner, you can create a calibration graph for each channel of your system.

You'll notice that the useful dynamic range displayed in the calibration graph of

Figure 6 *is less than that of the basic potentiometer circuit;* that is, from about 35 to 250, instead of 0 to 255. Below a paddle value of 35, large changes in input voltage are required for small changes in digitized values.

With the calibration graph in hand, your next step is to determine what it is you want to measure with the system (signal strength, plate voltage, etc.), and to provide an appropriate sample of the signal to the computer interface. For instance, if you plan to monitor a 24-volts DC signal, you'll need a voltage-divider circuit, much like the circuit used to calibrate your system **(Figure 10)**, to provide an appropriate signal sample.

Adjust the voltage divider (a high-resistance potentiometer) so the range of voltages to be measured falls within a useful area on your calibration graph—from 50 to 150, for example—with the nominal voltage corresponding to a digitized value of 100. Based *on the calibration graph in* **Figure 6**, *this* would correspond to an input voltage of 2.5 *volts at a digitized value of 50,* 1.5 volts at 100, and 1.1 volts at 150. Assuming a linear voltage divider circuit like the one in **Figure 10**, the source voltages corresponding to digitized values 50, 100, and 150 would be 39, 24, and 18 volts, respectively. Each digital increment or change in paddle value would represent roughly 0.2 volt.

Because the arbitrary paddle values are of little merit in and of themselves, you must translate them into meaningful terms. This *task is handled easily in software.* In the preceding illustration, a digitized value of 100 corresponds to a voltage source of 24 volts DC. In your BASIC program, simply add the statement:

$$voltage = (255 - PV) / 6.5 \qquad \textbf{(3)}$$

where PV is the value returned by a PEEK into the appropriate memory location for the paddle value in question **(Listing 1)**. PV is first subtracted from 255, the full range of the digitized value, to invert the normal-voltage/digitized-value relationship. That is, with *increasing* signal voltage, the equivalent resistive value *decreases*. In **Equation 3**, 6.5 was found by solving the following equation:

$$voltage = (255 - \text{paddle value with voltage input}) / k \qquad \textbf{(4)}$$

and

$$24 = (255 - 100) / k$$
$$24 = 155 / k$$
$$k = 6.5$$

```
100  REM READ ONE C-64 CONTROL PORT
110  REM PADDLE INTO ARRAY "A"
120  DIM A(100): C = 12 * 4096
130  REM READ MACHINE LANGUAGE DATA
140  FOR I = 0 TO 63 :READ A :POKE C + I, A :NEXT
150  REM PUT NEXT 100 PADDLE #1 VALUES INTO A
160  FOR I = 1 TO 100
170  SYSC :REM CALL ML ROUTINE
180  A(I) = PEEK(C+257)
190  NEXT
200  REM PRINT ARRAY VALUES
210  FOR I = 1 TO 100
220  PRINT "POSITION ", I, " = ", A(I)
230  PRINT
240  NEXT
250  REM
260  REM MACHINE LANGUAGE DATA
270  DATA 162, 1, 120, 173, 2, 220, 141, 0, 193
280  DATA 169, 192, 141, 2, 220, 169, 128, 141
290  DATA 0, 220, 160, 128, 234, 136, 16, 252
300  DATA 173, 25, 212, 157, 1, 193, 173, 26
310  DATA 212, 157, 3, 193, 173, 0, 220, 9, 128
320  DATA 141, 5, 193, 169, 64, 202, 16, 222
330  DATA 173, 1, 220, 141, 6, 193, 88, 96
```

Listing 2. A BASIC listing for the C-64 to store sequential resistive values of a single paddle in an array. Digitized values, once stored (*line 180*), are printed on the screen (*lines 210 to 240*). The values could have also been manipulated and displayed in other formats, for example, graphed, or simply saved to disk.

Remember that we assigned a paddle value of 100 to an input of 24 volts. Had we centered 24 volts at a resistive value of 150, then K would have been:

$$24 = (255 - 150) / k$$
$$24 = 105/k$$
$$k = 4.4$$

For most applications, this calibration technique should suffice. However, in situations where greater accuracy is required, you should try to use the most linear part of the calibration curve; that is, from 100 to 200 in **Figure 6**. You can also correct for the slight nonlinearity of the system with programming. For instance, with nonlinearity correc-

```
100  REM READ IBM GAME CONTROL ADAPTER
110  CLS
120  LOCATE 1,1
130  PRINT "POT #1 = "; STICK(0)
140  PRINT "POT #2 = "; STICK(1)
150  PRINT "POT #3 = "; STICK(2)
160  PRINT "POT #4 = "; STICK(3)
170  GOTO 120
```

Listing 3. A BASIC listing for the IBM-PC to display resistive values in an endless loop, similar to the C-64 code in *Listing 2*. This code assumes that the game control adapter has been installed in the PC, and that the voltage-to-resistance converter described for the C-64 is in place.

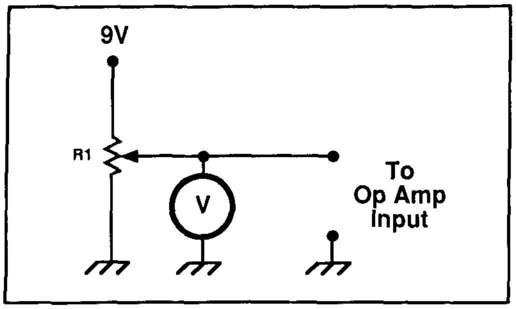


Figure 10. The calibration circuit for use with the voltage-to-resistance converter. R1 is a 10 to 100-k potentiometer. For the voltmeter (V), you should use a DVM or other high-impedance instrument.

tions, the preceding BASIC statement might appear as:

voltage =
*k1 * ((255 - paddle value with voltage input) / k) - k2* **(5)**

where k1 and k2 are appropriately defined constants. For those not mathematically inclined, there are a variety of programs available to help determine these constants. You can also use the trial and error approach.

Alternative computer systems

Computer-based monitoring need not be limited to the C-64. If you have an IBM-PC system, a game control adapter will support four resistive inputs. This card fits into one of the expansion slots, and the game control interface cable attaches to the rear of the adapter.

A BASIC program that reads the game control adapter card on the IBM-PC, equivalent to the listing for the C-64, is provided in **Listing 3**. Note that the STICK(0) retrieves all four paddle-input values, in addition to returning the value for the first paddle. STICK(0) must therefore be called repeatedly, in a manner analogous to the SYSC call in **Listing 1** for the C-64.

The voltage-to-resistance interface hardware described for the C-64 should work, without modification, for the IBM game control adapter. Although your calibration curves will probably vary somewhat from those produced on the C-64 system, operation should be similar in all other respects.

Summary

For about $15 in parts, a half-hour of programming, and a little ingenuity, you can extend your microcomputer's capabilities to include automated signal monitoring. The uses of such a system are limited mainly by your imagination. The signal sources can be as varied as your communications arsenal. Once the signals have been digitized and captured by your computer, you can store, manipulate, and display them in a variety of ways. Enjoy! ■

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3. *Commodore 64 Programmer's Reference Guide*, Commodore Business Machines, Inc. and Howard W. Sams & Co.,Inc., 1982.
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5. P. Horowitz and W. Hill, "Measurement and Signal Processing," *The Art of Electronics*, Cambridge University Press, Cambridge, Massachusetts, 1980.

PARTS LIST

CAPACITORS

C1	*0.1-μF, 30-volt ceramic disk capacitor (for power supply bypass)*

RESISTORS

R1,R4	*1-meg, 1/4-watt resistor, 5 percent tolerance*
R2,R3, R5,R6	*1-k, 1/4-watt resistor, 5 percent tolerance*

INTEGRATED CIRCUITS

U1	*TL092 N-FET dual op amp (RS part no. 276-1746)*
U2,U3	*TIL-111 optoisolator (RS part no. 276-139)*

MISCELLANEOUS

B1	*9-volt alkaline battery*
S1	*SPST micro switch*

Battery clip, 9 volt (RS part no. 270-325)
DIP socket (16 pin)
DIP socket (8 pin)
Dual IC board (RS part no. 276-159) or pc board in Figure 7
Joystick extension cable (RS part no. 270-1705)
LED (optional)
Optional 3-k, 1/4-watt resistor for LED
Plastic project box (RS part no. 270-221)

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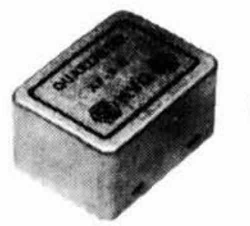

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