COMMUNICATIONS QUARTERLY

THE JOURNAL OF COMMUNICATIONS TECHNOLOGY

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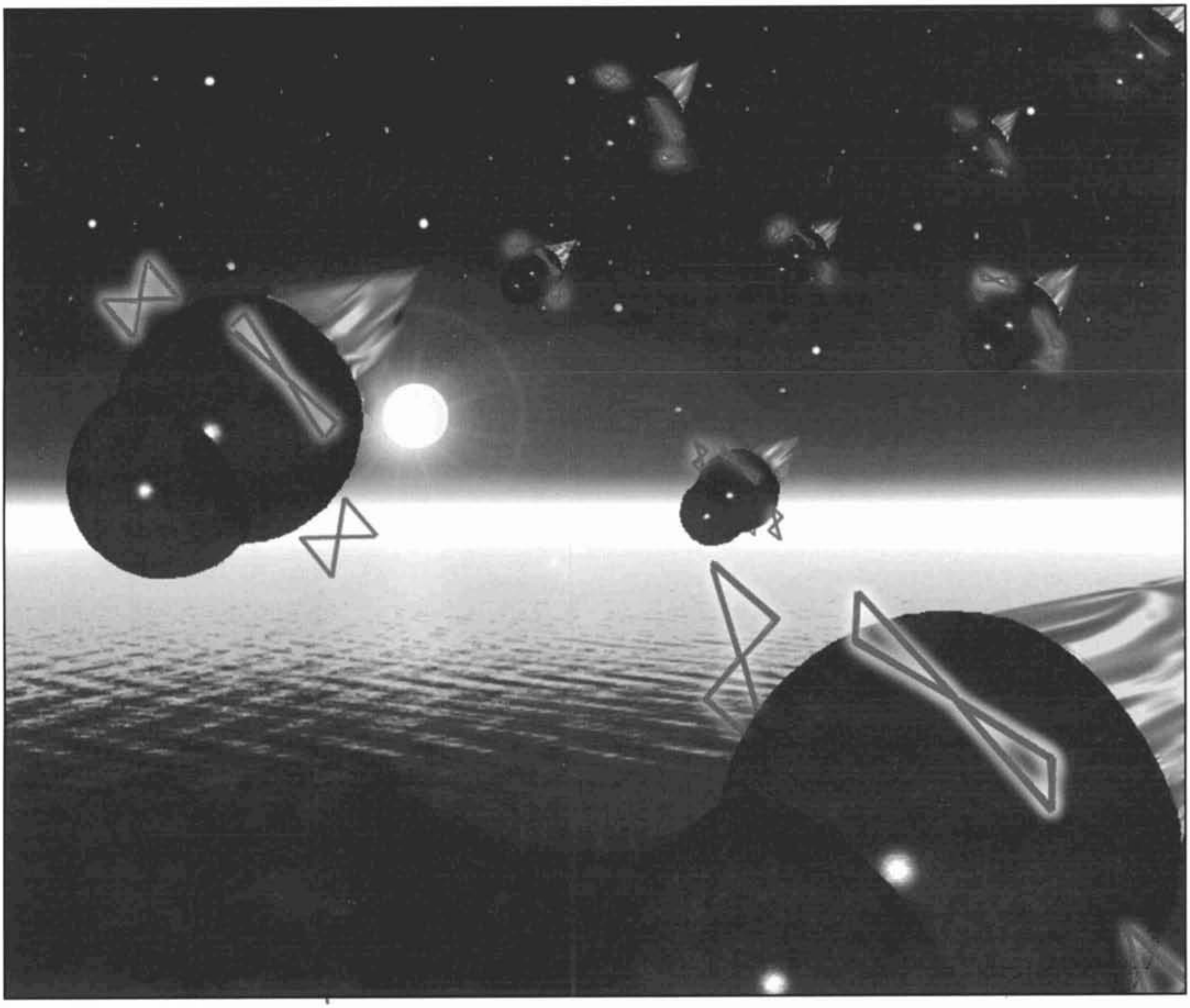

- Are Quarter-wave Radials Really Optimal for Vertical Antennas?
- Announcing the Marriage of EZNEC and NEC-2 in EZNEC for DOS
- Need a Versatile Antenna for DX Work? Try the Lazy-H Vertical
- A Summary of Various Approaches for Validating an F-Layer Algorithm for the Ionosphere
- Upgrade for Boonton Models 92/42 RF Voltmeters Provides a Reliable Replacement for Old Chopper Circuit and Retains Programmability
- A Legendary Radio Service Returns as Globe Wireless Rides the Airwaves Again
- Modeling and Understanding Small Beams—Learn More About Fan, Bowtie, Butterfly, and Dragonfly Antennas

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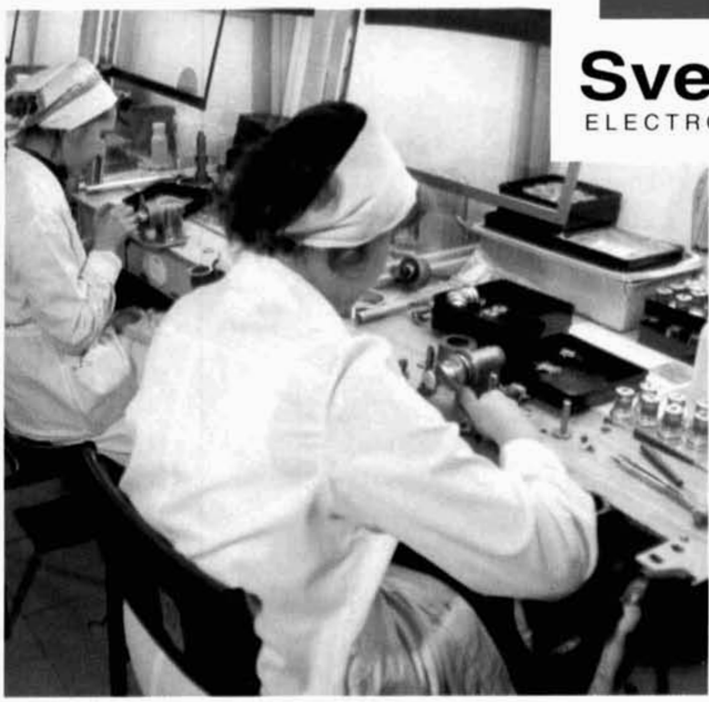

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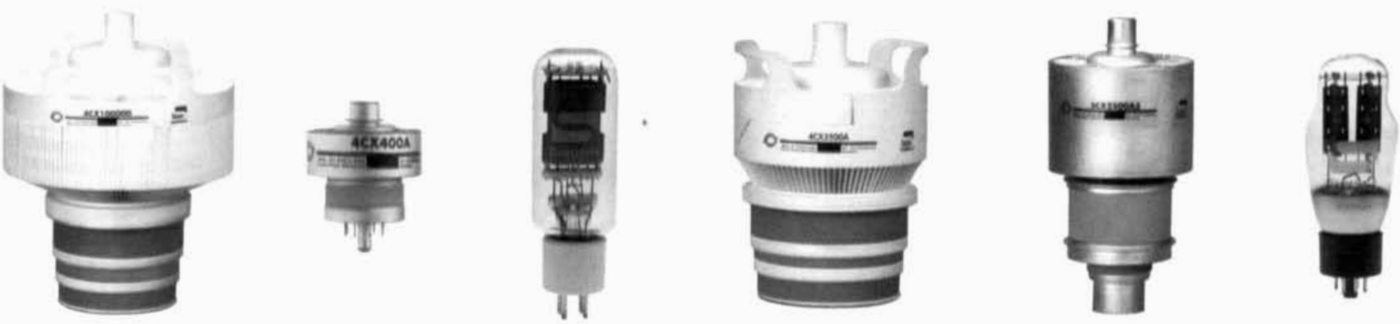

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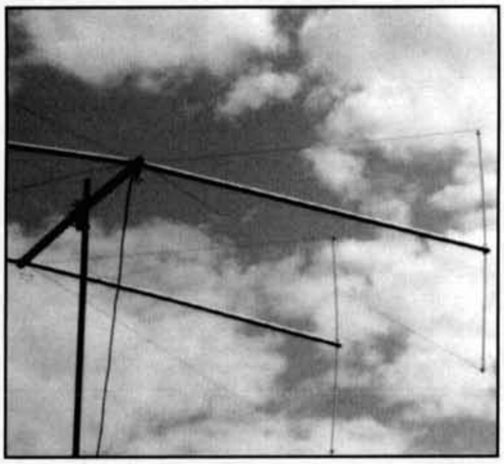

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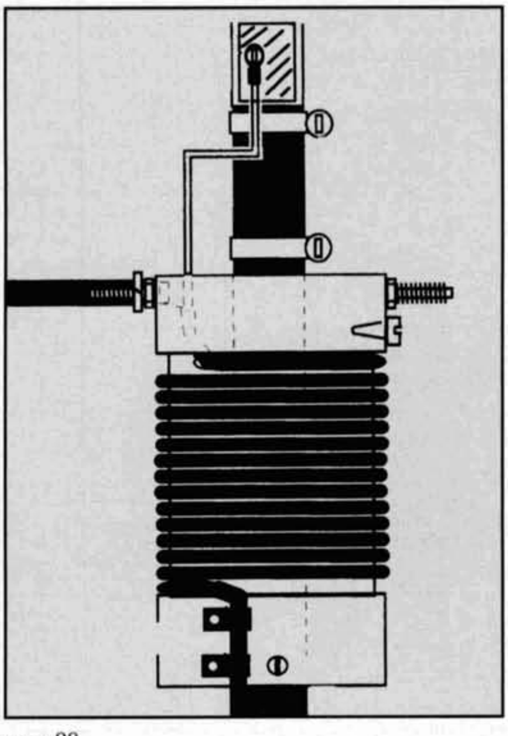

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On the Cover: Igloos topped with fan and bowtie antennas rocket past by the light of the Aurora in Bryan Bergeron, NU1N's version of the popular Flying Toasters™ screen saver. For more on fan and bowtie antennas, see "Modeling and Understanding Small Beams: Part 6," by W4RNL. The article begins on page 81.

EDITORIAL

Is Your Station COOL?

One afternoon, as I was busily trying to finish a manuscript for *CommQuart*, my computer shut off. No warning; it just shut off. Why? The OM had just flipped on his linear amplifier—socking radio frequency interference (RFI) into my feeble computer and shutting it down. Before this, I had experienced RF-induced color changes on my monitor. Sometimes, the letters even performed a snake dance across the screen. That I could live with, but losing an hour's painstaking work (okay, I know I should have saved often) was a setback that left me sniveling. Rick's station was definitely NOT cool! In fact, it was hot with RF. Okay, so turning off my computer was the worst the amp had ever done; but, I could still hear complete QSOs over the television or my "walkperson" radio during contests.

These days, RF-vulnerable equipment is no longer confined to stereos, televisions, and telephones (although the old coil-based phones were fairly numb). Now that our homes are filled with microprocessor-controlled devices such as fax machines, home security systems, specialized home lighting systems, electronic telephones, and computers, anything could happen. Say you're in a contest, everything is up and running, and your linear is in the loop. You hear a rare one, push to transmit, and your computer (and logging program) goes down, the lights go out, and your home security system starts whooping away. Talk about a Murphy strike! Given the possibilities, there's a real need to make our ham shacks RFI COOL.

Reducing RFI

There are some easy ways to reduce RFI. First, try to mount the radiating portion of your antennas as far above—or as far away from—the house as you can. This will reduce high-field radiation into the structure and its wiring. Second, eliminate radiation from feedlines by using a balun, ferrite sleeves where the feedline enters the house (if needed), and a good station ground. Use a clamp-on RF ammeter if necessary, and make sure the feedline is COLD. Finally, install ferrite chokes on phone lines, TV cables, long stereo speaker cable runs, security systems, and so on.

RFI problems

Okay, we all know that RF in the house really interferes with microprocessor-based appliances and other gadgets, and that's a good reason to RFI-proof our shacks. However, it's also important to remember that RF energy can also cause severe burns. There have even been studies of chronic health hazards caused by EMR exposure—although the jury is still out on this claim.

Although it isn't often mentioned, interference is a two-way street. If your feedline isn't cold, it will efficiently pick up signals from all the RFI-emitting stuff in your house—computer hash and timing oscillator carriers, dimmer buzz, motor noises, etc.

Neighborhood etiquette

Our main concern is cleaning up our own nest. Still, there are times when ham signals can get into the neighbor's house and cause problems. We've all heard stories where local radio amateurs and their stations were blamed for neighborhood interference problems, and many of us will try to make adjustments to our equipment (or theirs) in an attempt to eliminate the disturbance.

Of course, these days it's hard to determine what's causing the problem—our signals or the neighbor's assortment of electronic gadgets. Consumer electronics are notorious for being poorly shielded against incoming and outgoing signals. But if you have a big antenna in your yard, certain prejudices come into play—and your station will become a target. But despite the fact that your neighbor's own equipment could be causing the disturbance, it's possible that if your station is disrupting things around your home, it might it also be causing problems to those who live nearby. It's a Catch 22 situation and good public relations are in order.

Becoming COOL

The best way to ensure your station is in tip-top shape, is to perform a thorough spring housecleaning. After all, springtime is traditional the time for hams to check out towers, masts, antennas, and feedlines and get everything in shape for another year of operating. While you're at it, take the time to houseclean for sources of RFI. Turn on your amplifier and then see if it's getting into any of your other electronics. Add filters, chokes, baluns, etc. to your ham gear until you eliminate any RFI. Check other electronics like computers and test equipment you have in your shack to see what type of interference they might be creating. Once you've finished your troubleshooting and installed any needed fixes, you can confidently say your station is in the clear. Your station can be COOL, and your family (and your neighbors) will thank you.

Terry Littlefield, KA1STC
Editor

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"a BRIDGE a la française"

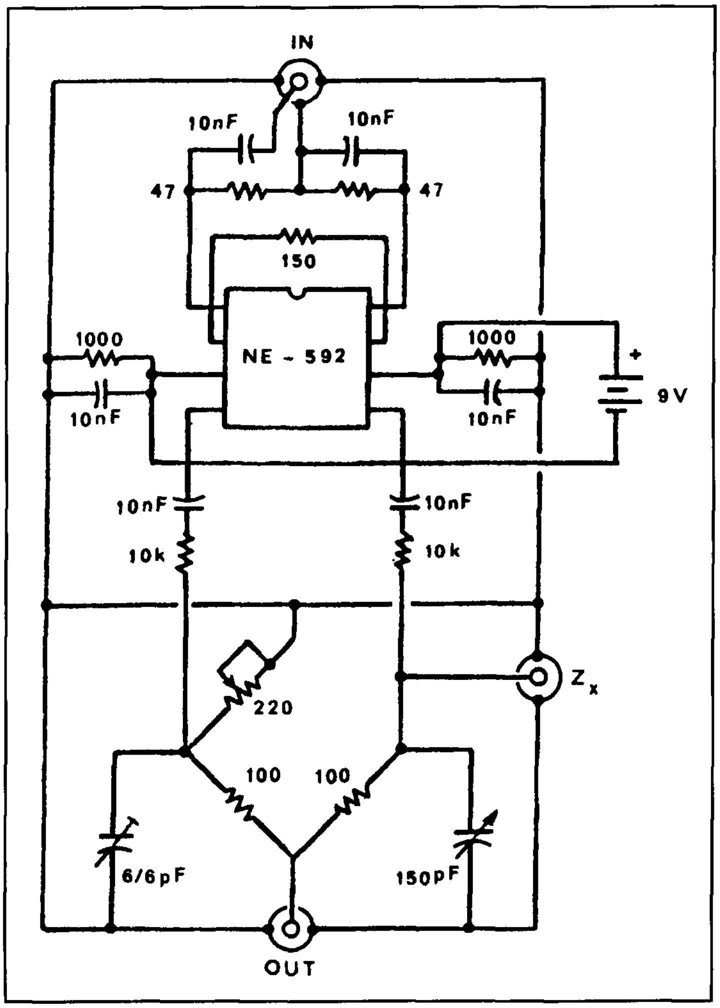
Dear Editor:

Several articles have been published for years in *Communications Quarterly* about "simple and accurate,[1] perfect,[2] and ultimate[3]" bridges.

Here you are "**a BRIDGE A LA FRANÇAISE,**" simple, accurate, and maybe you will find perfect and ultimate!

The principle is to transform the asymmetrical signal coming from a generator into a symmetrical one by means of an active device instead of using a transformer or a floating source. A video amplifier is able to do that function up to tenths of megahertz.

The diagram shows an example working well up to 30 MHz. The signal source may be a signal generator or a noise generator, both modulated or not. The detector will be an HF receiver like a CB one or a scanner. The power input could be in the range -20 dBm up to 0 dBm and the detector able to show -70 dBm to -20 dBm signal; that is not critical at all.



A BRIDGE "a la française."

Finding the minimum is easier with a sine wave signal modulated at 1000 Hz rather with a noise generator, modulated or not. But that kind of source is much cheaper and very common in the OM shacks.

Series and parallel configuration bridges are both workable using that signal source. With smaller parts mounted on a pc board with screens, the frequency range can easily reach 100 MHz.

I hope this could be published in "Technical Conversations."

Andre Jamet, F9HX
Meyzieu, France

REFERENCES
1. Wilfred N. Caron, "A Simple and Accurate Admittance Bridge," *Communications Quarterly*, Summer 1992, page 44.
2. A.E. Popodi, AA3K/OE2APM, "Building the Perfect Noise Bridge," *Communications Quarterly*, Spring 1993, page 55.
3. A.E. Popodi, AA3K/OE2APM, "The Ultimate Noise Bridge," *Communications Quarterly*, Summer 1996, page 25.

Three small errors

Dear Editor:

I found three small errors in Parker Cope's article "A Stable Oscillator" (*Communications Quarterly*, Fall 1996, page 50).

(1) The second equation in the second column on page 15

now: $e_c = R_c I_c + R_k (I_c I_D)$

should be: $e_c = R_c I_c + R_k (I_c + I_D)$

(2) On page 52, about 3-1/2 inches down is a short single sentence paragraph. The first word on the third line reads

efficient

That word should read

coefficient

(3) On page 52, the right column, exactly three inches from the bottom of the printed material

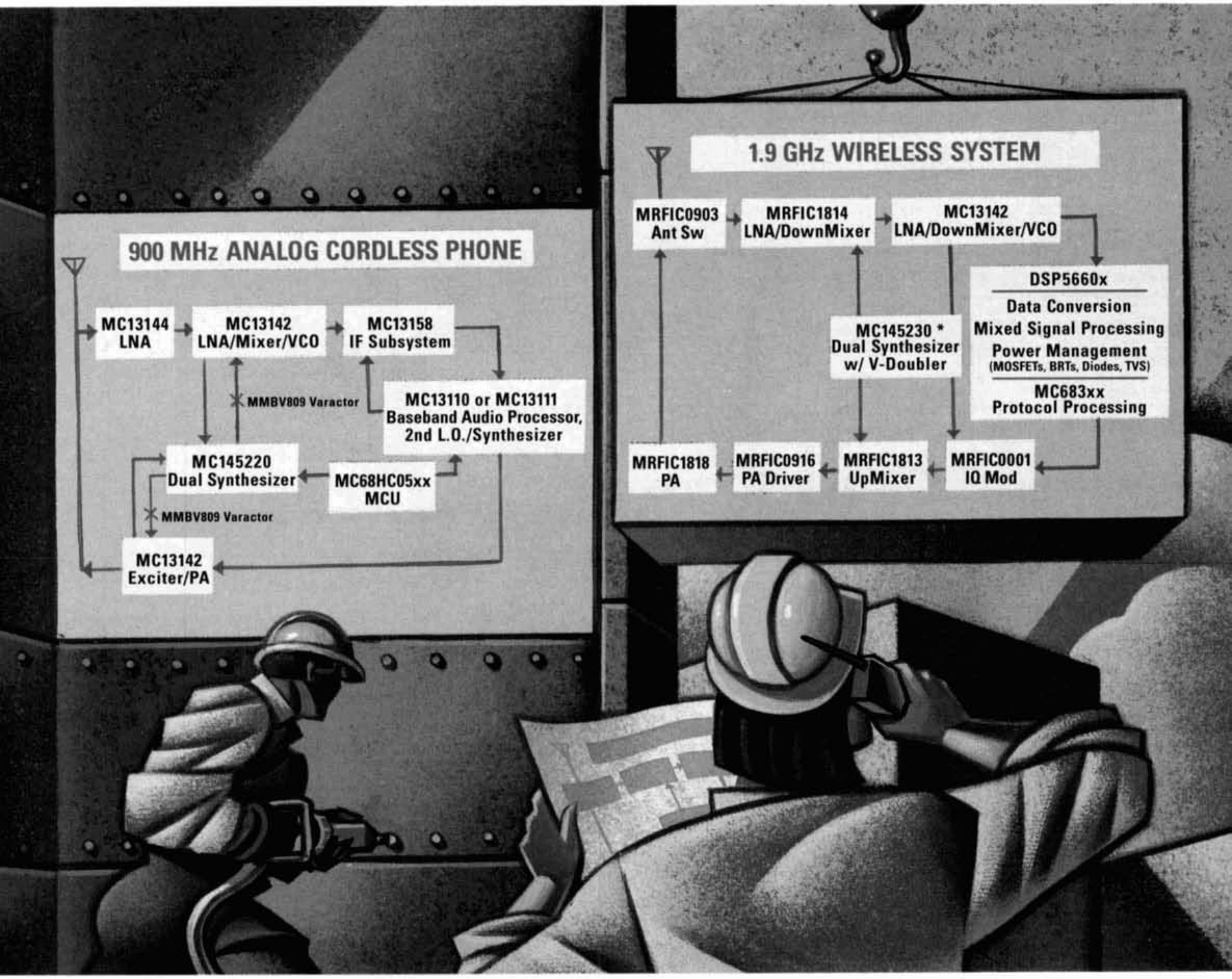
now: For example, a multimeter, a 6-volt battery, and a 22-ohm resistor are...

should be: For example, a multimeter, a 6-volt battery, and a 22-k resistor are...

Charles Cashion

(Continued on page 108)

The Wireless Solution Site



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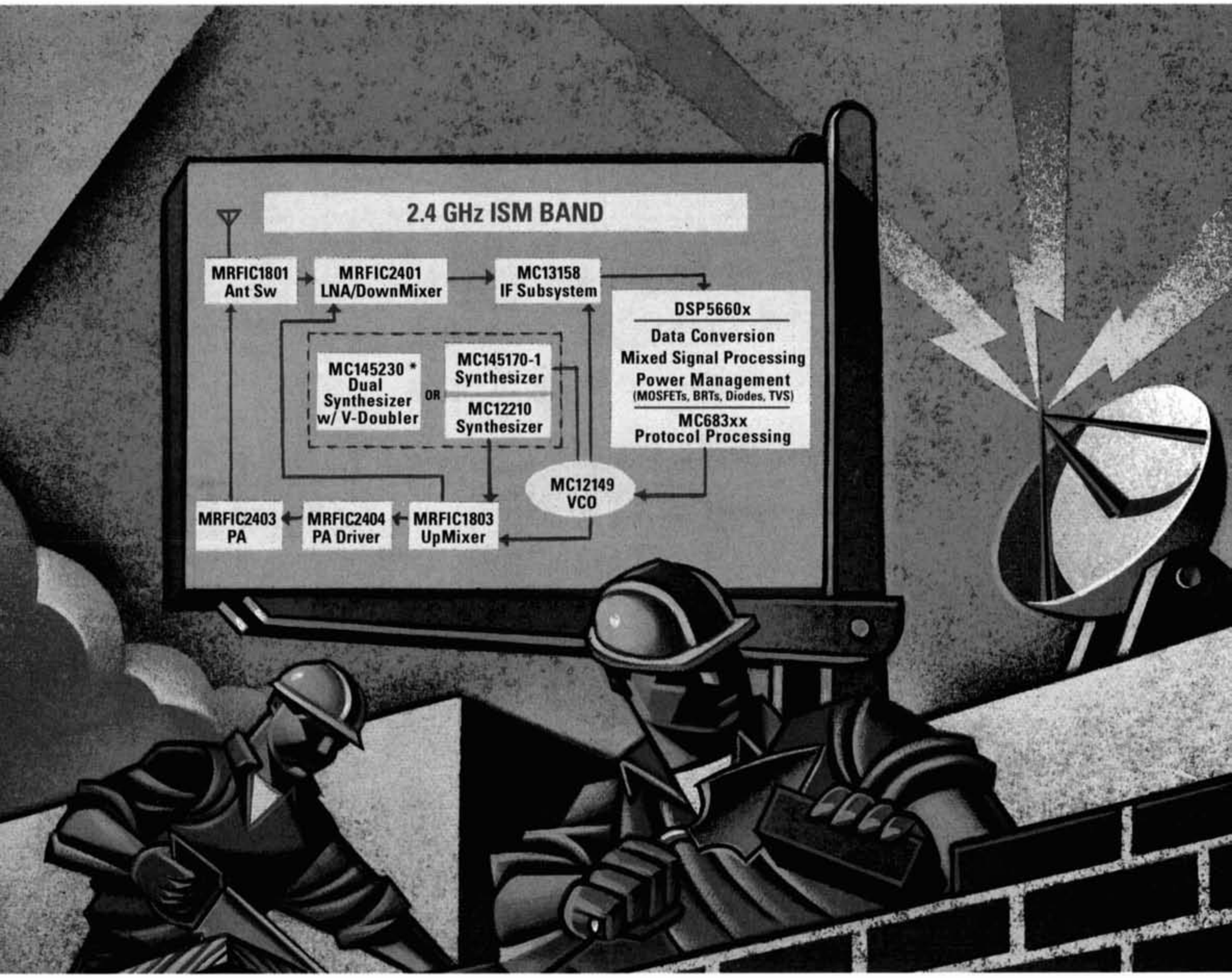
2.4 GHz ISM BAND
MRFIC1801 Ant Sw
MRFIC2401 LNA/DownMixer
MC13158 IF Subsystem
DSP5660x
Data Conversion
Mixed Signal Processing
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(MOSFETs, BRTs, Diodes, TVS)
MC683xx
Protocol Processing
MC145230 * Dual Synthesizer w/ V-Doubler
OR
MC145170-1 Synthesizer
MC12210 Synthesizer
MC12149 VCO
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OPTIMAL ELEVATED RADIAL VERTICAL ANTENNAS

Design minimizes effects of unequal radials' currents

Over the past two years, I've learned much from the measurement of elevated radial currents and computer simulations. One thing I've found is that vertical antennas using quarter-wavelength elevated radials aren't working properly for some fundamental reasons. I'd like to share my findings along with supporting test data and NEC simulations. In addition to discussing quarter-wavelength elevated radials and why they shouldn't be used, I'll present an approach to designing well-behaved, optimized vertical antennas using elevated radials. As part of this approach, the problems encountered with quarter-wavelength elevated radials are circumvented. What I've found is a real eye opener, and I believe the information presented will have an impact on amateur radio operators who use vertical antennas with elevated radials. I also expect that commercial broadcast stations using or anticipating using elevated radials may be affected by my findings.

About 13 years ago, I put up a remotely tuned vertical for 80 meters with four elevated radials. I suspended it from a support wire beginning at the 120-foot level of my 140-foot rotating tower, ending at a 16-foot post as shown in **Figure 1**.

Each radial was a quarter wavelength. I made the vertical wire longer to raise the radiation resistance to 50 ohms—in addition to the increase provided by sloping the radials downward. This arrangement used a remotely controlled motor to drive a capacitor in series with the vertical wire, to tune out the inductive reactance. It had a 1:1 balun to ensure that current didn't flow on the shield of the coaxial feedline. This configuration worked well and allowed me to achieve a very acceptable VSWR across the band, as shown in **Figures 2** and **3**. **Figure 2** shows VSWR curves centered on three frequencies with the capacitor held stationary. I was pleased with these curves because they indicated the antenna wasn't overly broad banded, which would imply excessive losses. Using the remote control feature, I took the lowest VSWR found at 50-kHz steps, and plotted the graph in **Figure 3**. Although the VSWR increased at the upper end of the band, I made no attempt to improve the VSWR at the high end, and I used the antenna this way for the next 11 years.

In the Spring 1995 issue of *Communications Quarterly*, Bill Shanney, KG6GR, published a very interesting article called "Understanding Elevated Vertical Antennas."[1] He discussed the performance of a vertical antenna that used two elevated radials and a potential problem that might occur with such a system.[2] Shanney showed radiation patterns for a 20-meter quar-

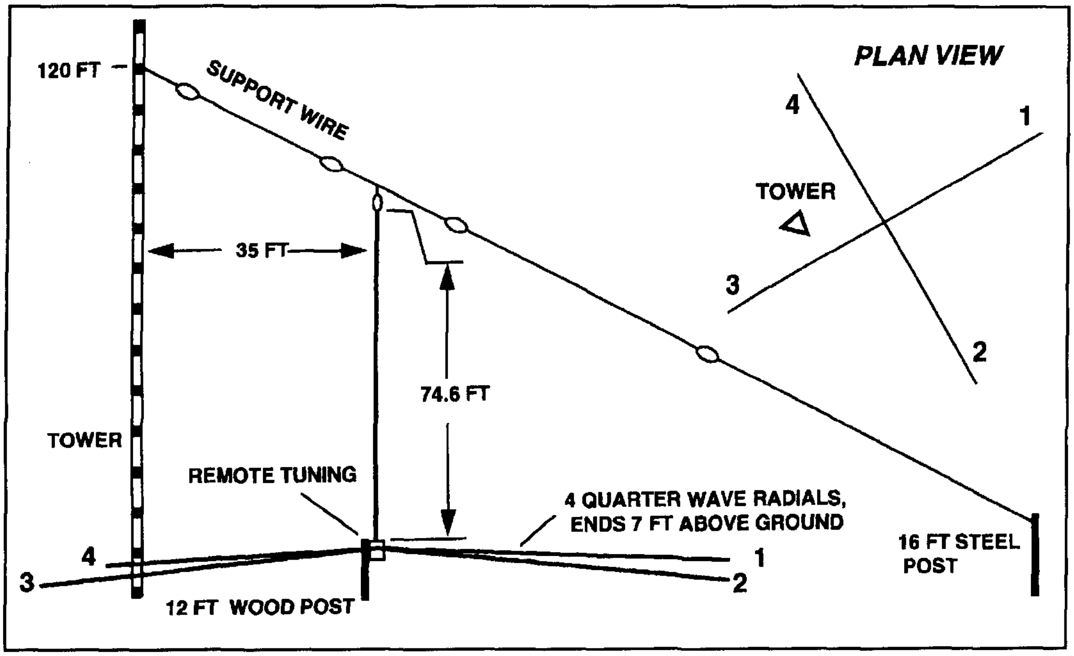

Figure 1. Elevated radial vertical antenna is remotely tuned.

ter-wavelength vertical mounted 20 feet above ground with two quarter-wavelength elevated radials, and another case where one radial was 2.7 degrees shorter than a quarter wavelength and the other was 2.7 degrees longer.[3] Plot comparison revealed noticeable differences. To illustrate this, Shanney's plots are recreated here along with two additional patterns. These are shown in **Figures 4** and **5**, which were generated using NEC-WIN.* "Average ground" was simulated for these models using a conductivity of 0.005 Siemens/meter and a dielectric constant of 13. **Figure 4** shows elevation radiation patterns for the antenna with quarter-wavelength radials; **Figure 5** shows the same plots, where the radials differ by +2.7 and -2.7 degrees from a quarter wavelength. The predicted amplitude and phase of radial currents for both antennas are shown in **Table 1**, where current amplitudes have been normalized to that in the vertical radiator.

Table 1 shows the existence of unequal radial currents in the antenna with the lengthened-shortened radials. These currents cause the patterns to change markedly as seen in **Figure 5**. Because I always had the feeling that my 80-meter antenna worked better in some directions

*Paragon Technologies, Inc. Research Drive, NEC-WIN, suite A-1, State College, Pennsylvania 16801.

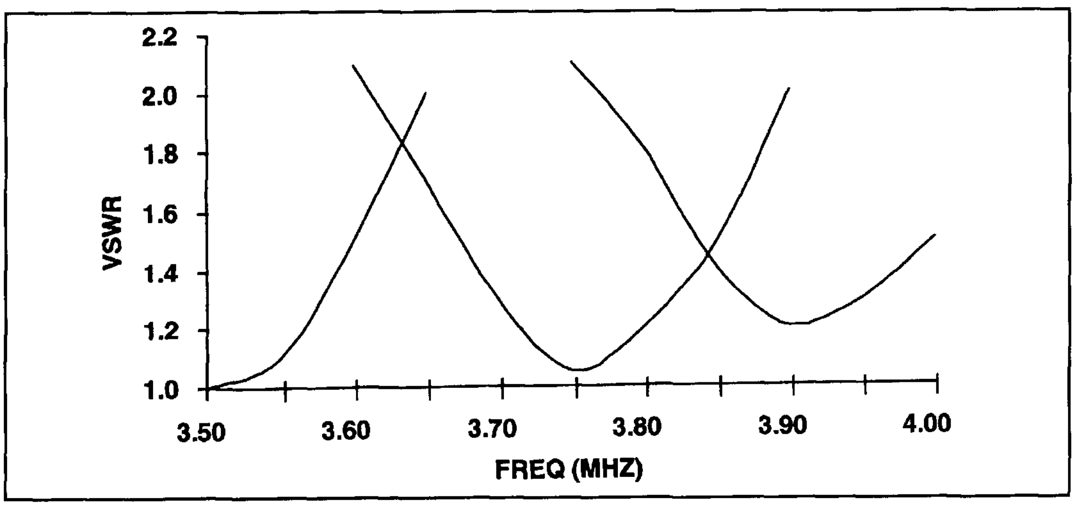

Figure 2. VSWRs with stationary capacitor settings.

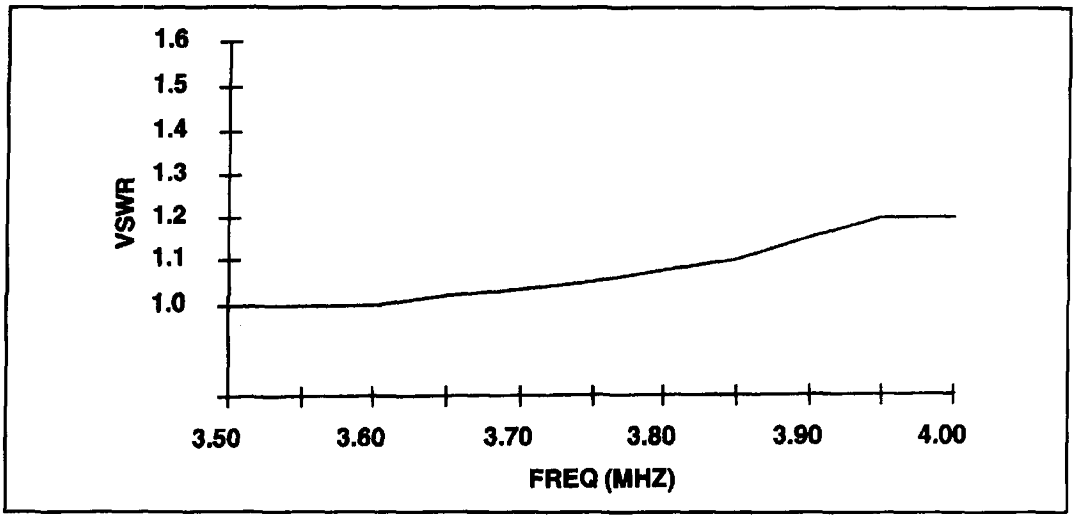

Figure 3. "Best" VSWR obtained using variable capacitor.

than in others, Shanney's original plots prompted me to wonder if my 80-meter vertical had unequal radial currents that might cause its pattern to distort or become asymmetrical in some similar manner. I decided to measure the radial currents.

Radial current measurements

I measured the relative magnitude of currents in the four radials of my 80-meter vertical at 3.528 and 3.816 MHz using a pick-up loop detector as shown in **Figure 6**. I hung the loop from each radial at a convenient distance of 34 feet from the end insulator and read the meter as I keyed the transmitter with a remote line. I measured the relative currents for each radial by moving this setup from radial to radial, without changing the level adjust pot. I normalized the measured currents at each frequency and plotted them, as shown in **Figure 7**, using the radial numbering scheme in **Figure 1**.

I was quite surprised by the results, but they explained why I felt the antenna seemed to work better in one direction. At the time, I didn't have a copy of NEC to help me understand what was happening. Fortunately, Shanney included a discussion of unequal radial currents in his article that referenced two additional sources of information on the topic by Les Moxon, G6XN.[4]

In Moxon's article, "Ground Planes, Radial Systems, and Asymmetric Dipoles," he writes

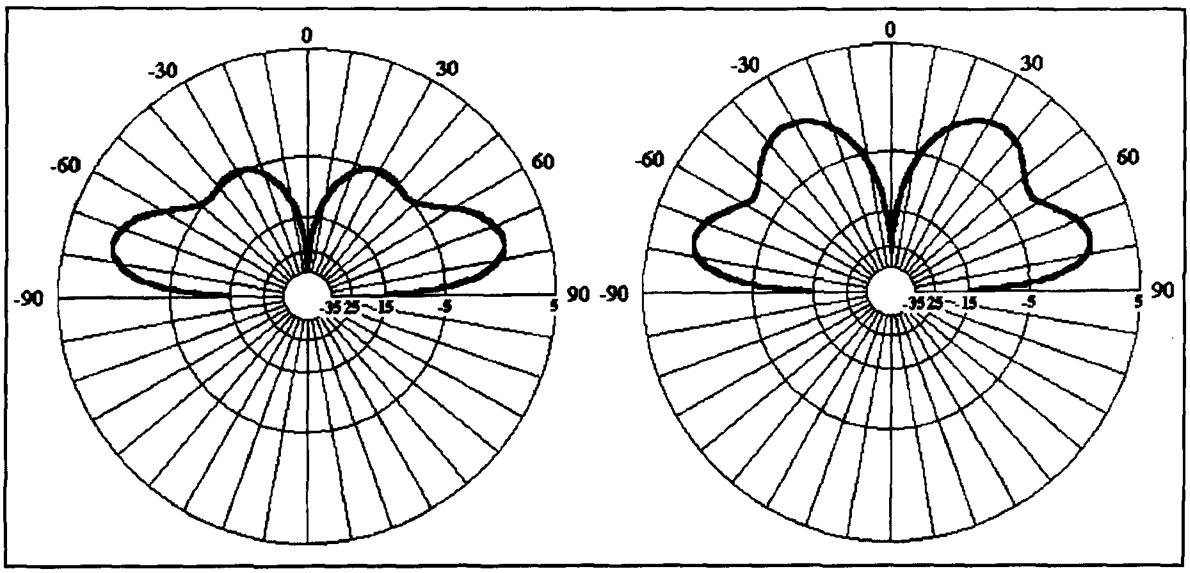

Figure 4. Elevation patterns for 20-meter quarter-wave vertical with two 90-degree wavelength radials 20 feet above ground. (A) Broadside to radials. (B) Into radial end.

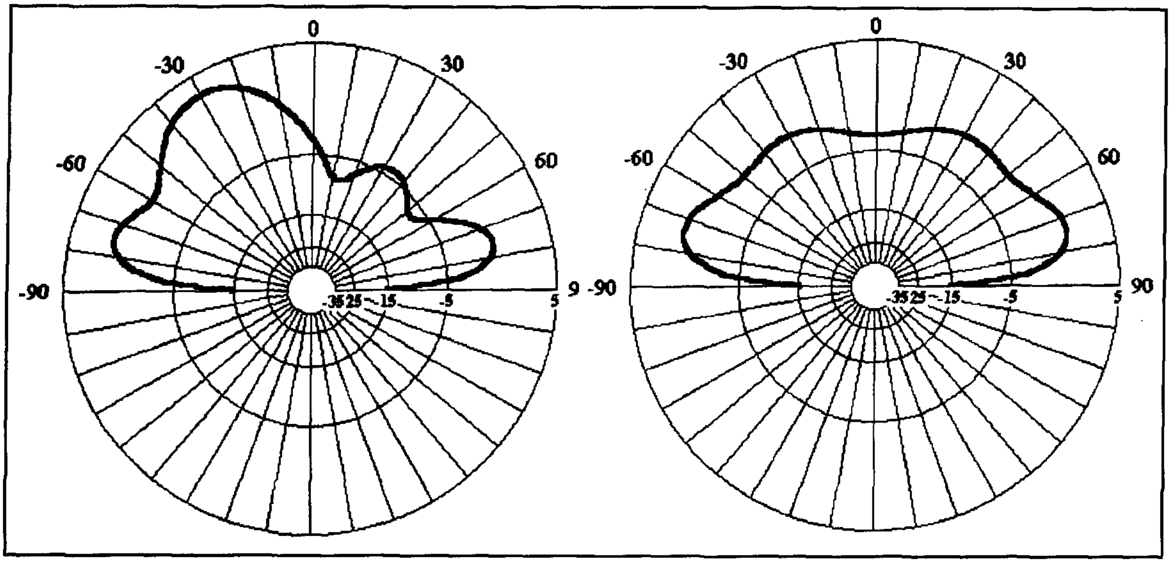

Figure 5. Elevations patterns for a 20-meter quarter-wave vertical with 87.3- and 92.7-degree wavelength radials 20 feet above ground. (A) Broadside to radials (longer radial to right). (B) Into radial end.

that a vertical with elevated radials should not have quarter-wavelength radials, but should use a length more like a 1/8 wavelength with a single inductor to create a resonant system.[5] Moxon argues that reduced length radial systems are much less sensitive to the influences of the surrounding environment, and that equal radial currents will more readily result. After reading these articles, I reduced the length of my radials to 1/8 wavelength and added a radial inductor as shown in **Figure 8**.

After some testing, I found a radial inductor setting that provided excellent VSWR characteristics across 80 meters, as illustrated in **Figures 9** and **10**. It was heartening to see that the plots in **Figures 2** and **9** were about the same, indicating that there was no noticeable change in antenna losses. Even more heartening were the results from radial current measurements using the pickup loop 19.5 feet from the end insulators. It's clear from the plots in **Figure 11** that radial currents were essentially the same at either frequency.

Further experimental measurements

Since the time my antenna was modified, I've been able to obtain radial current test data

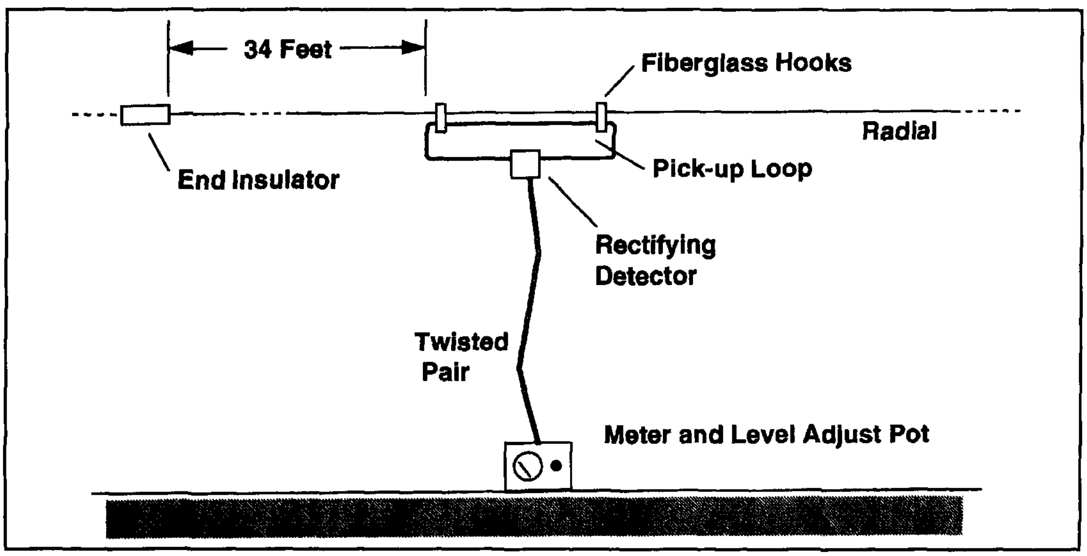

Figure 6. Pickup loop used to measure radial currents.

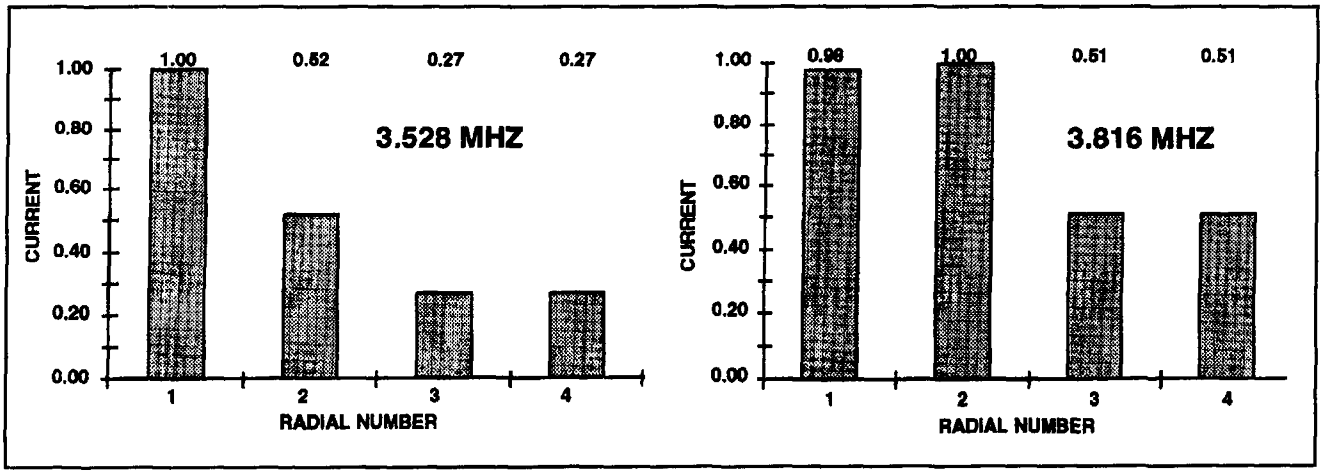

Figure 7. Measured radial currents with quarter-wave radials.

for three other verticals using elevated radials. This provided a welcome chance to see if other antennas with elevated radials had unequal radial currents. First, I had an opportunity to measure the radial currents in the two 160-meter verticals at WXØB (see sidebar). At Jay Terleski's station, each of the antennas had two elevated radials. The feedpoints were 20 feet above ground and the radial ends about five feet high. Each antenna had a coiled line choke at the feedpoint to prevent the flow of current on the feedline shield. Results of our measurements appear in **Figure 12**.

Figure 12 shows that the measured radial currents are quite unequal. Needless to say, Jay and I were both surprised by the behavior of ANT1's east radial. To be sure we weren't making a mistake and that the current pick-up loop wasn't broken, we repeated all of the measurements and checked to make sure the questionable radial had a good electrical connection to the feedpoint. We found that our results were correct.

Jay modified ANT1 to see if the radial currents could be made equal, or at least similar. He shortened the two radials to a 1/8 wavelength, installed a radial inductor, and tuned the antenna for the lowest VSWR at the lower end of the band. After modifications were complete, we measured the radial currents again, and found the current had improved (see **Figure 13**.)

While the currents aren't equal, there has been significant improvement, and the frequency dependency has been greatly reduced. Although ANT1's two shortened radials are the same length, the ends aren't the same height above ground, and they have considerably different objects nearby. These conditions probably contributed to the initial radial current imbalances. They still remain, but their influence is much diminished.

The second set of independent radial current measurements is from Arliss Thompson, W7XU. He has a 160-meter quarter-wave vertical with four elevated radials 15 feet high at the

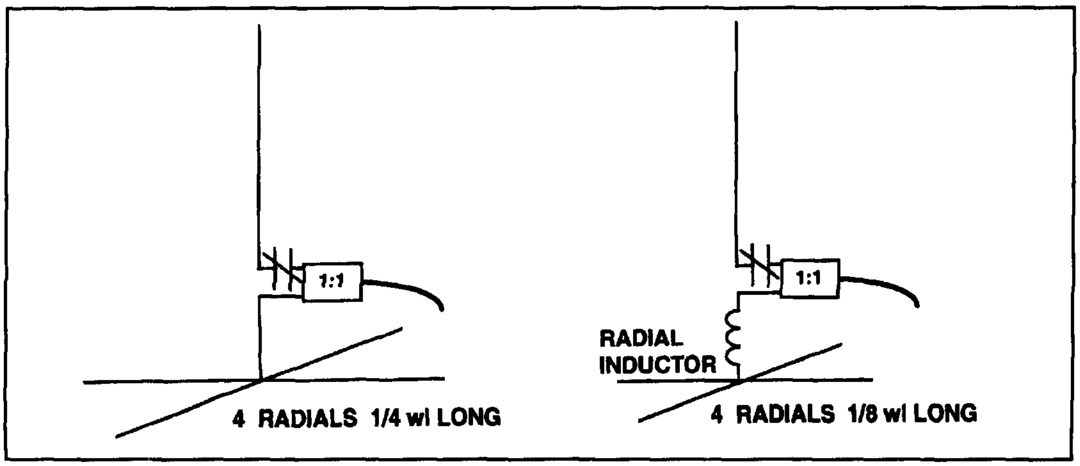

Figure 8. Radial inductor added to use shortened radials.

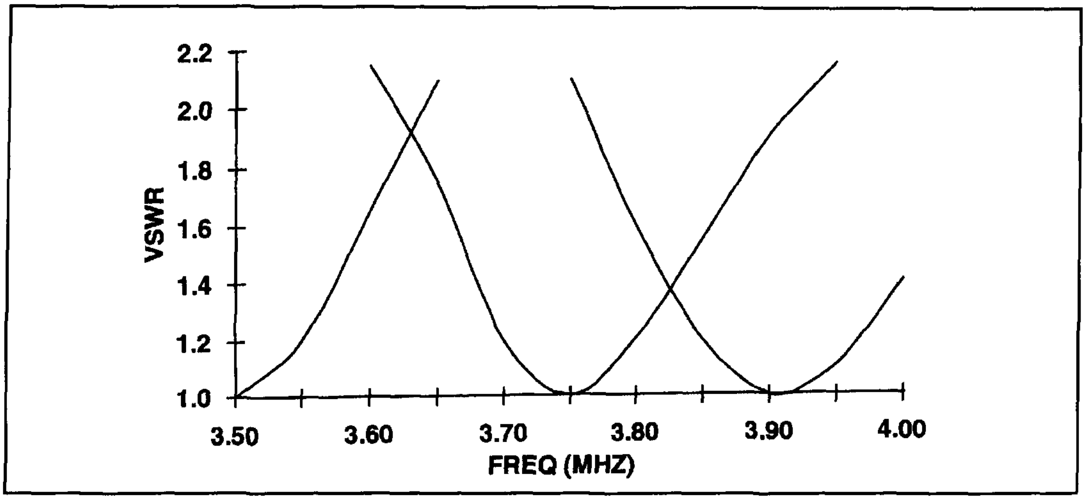

Figure 9. VSWRs with stationary capacitor settings using 1/8-wavelength radials.

tower and five feet at the outer ends. The system is in a field with nothing in the vicinity of the radials. Initial radial current measurements are shown in **Figure 14A**. After initial tests, the radials were measured and found to be of different lengths; however, they only differed by a couple of feet. After making each radial 127 feet long, the currents were measured again as shown in **Figure 14B**. Current partitioning changed, but similar radial currents did not result. This led to numerous unsuccessful attempts to obtain similar radial currents by making small adjustments to the radial lengths. Radial currents for the last attempt are shown in **Figure 15** at two frequencies. Here, too, there is a frequency dependency. Modifications to shorten the radials have been started, but aren't complete at this time.

What kind of radiation pattern results if all or most of the radial system current is carried by one radial? I used NEC-WIN to generate the plots shown in **Figure 16** and to look at this question. The radiation pattern that results when one radial carries all the radial system current is shown in **Figure 16A** for a 160-meter quarter-wave vertical with one elevated radial over average ground. The feedpoint is 15 feet high with the radial end at five feet. The radiation pattern, where there are four radials with equal currents, is shown in **Figure 16B** for comparison. The single radial pattern is most likely what WXØB and W7XU had before modifications. The single radial pattern is also probably close to that which my original 80-meter system produced.

Why are elevated radials a quarter wavelength?

The use of quarter-wavelength elevated radials is an accepted practice that's gone unquestioned over the years by all but a few individuals. It now appears this practice should be carefully reviewed.

Why are quarter-wavelength elevated radials the current custom? It seems plausible that their use is an extension of the practices followed for ground-mounted verticals. With ground-mounted verticals, resonance is obtained with a quarter-wavelength vertical radiator where the ground appears to provide the other quarter-wavelength portion of the antenna. The efficiency of this type of system is improved using ground radials or a ground screen to lower the ohmic losses incurred by currents flowing back to the antenna feedpoint in the ground portion of the system.

Elevated radials have been proposed as a means to further enhance efficiency. Because efforts have mainly focused on improving the efficiency of quarter-wave verticals, it's easy to see that quarter-wavelength elevated radials are the logical choice, as they provide an easy way to resonate the antenna. This reasoning seems more than plausible because articles in professional journals investigating the use of elevated radials use quarter-wave verticals. Also, there are a large number of articles in amateur radio publications that talk about improving the efficiency of the quarter-wave vertical, some of which discuss using elevated radials. It seems that quarter-wavelength elevated radials have come in the backdoor as part of efforts to improve the efficiency of quarter-wave verticals.

For elevated radials to create an "efficient ground" they must provide two functions. First, elevated radials should be part of an overall antenna system design that is resonant or has an acceptable VSWR across a band of operation. Second, and most important, the radials as a system shouldn't radiate any horizontally polarized energy. If they do, the

	Vertical Radiator		Radial #1		Radial #2	
Configuration	Amp	Ang (deg)	Amp	Ang (deg)	Amp	Ang (deg)
.25 wl Radials	1.00	0.0	0.50	-180.0	0.50	-180.0
#1 +2.7deg / #2 -2.7deg	1.00	9.8	0.41	161.3	0.67	-153.0

Table 1. Quarterwave antenna currents.

NUMBER OF RADIALS		RADIAL LENGTH					
		45 DEG	60 DEG	90 DEG	120 DEG	135 DEG	150 DEG
4	GAIN (dBi)	-0.45	-0.32	-0.06	-0.01	-0.38	-1.18
	ANGLE (DEG)	23	23	23	22	22	22
8	GAIN (dBi)	-0.39	-0.26	0.00	0.19	0.15	-0.3
	ANGLE (DEG)	23	23	23	22	22	22

Table 2. Maximum gain of an 80-meter elevated quarter-wavelength vertical with different length radials 10 feet above ground.

desired pattern produced by the vertical radiator won't be realized. For a set of radials not to radiate horizontally polarized energy, the currents in each radial must be the same, and the radials must be symmetrical in pattern. If not, patterns more like those in **Figure 16A** will result, rather than the ones in **Figure 16B**. The need for equal, or reasonably similar, radial currents has been overlooked.

When a vertical antenna is installed, care is usually taken to ensure the system provides a reasonable match to the feedline. This is done either by adjusting antenna component lengths, or by adjusting a matching network until there's an acceptable VSWR. These steps provide a match to the transmission line, but do nothing to ensure the antenna is working as desired. I'm not aware of anyone who's made radial current measurements to see if they were similar. Without doing so, there's no assurance the antenna is operating properly.

What's wrong with quarter-wavelength elevated radials?

When a radial is compared to a single wire transmission line over ground, there's an indication that radials shouldn't be a quarter wavelength. Looking at a radial, you'll note that the far end of the line is terminated with an open. When the line is 90 degrees long, the open at the far end reflects to the input end as a short with an impedance of 0 ohms. If all radials within a system are exactly 90 degrees, they'll each have an impedance of 0 ohms, and displacement currents will divide equally among the radials. But, will all radials within a set

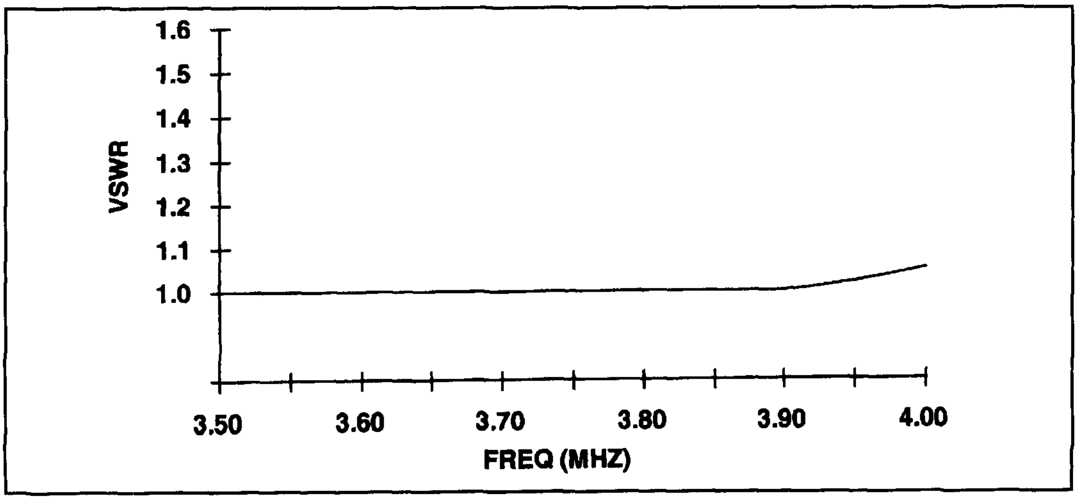

Figure 10. "Best" VSWR obtained using variable capacitor with 1/8-wavelength radials.

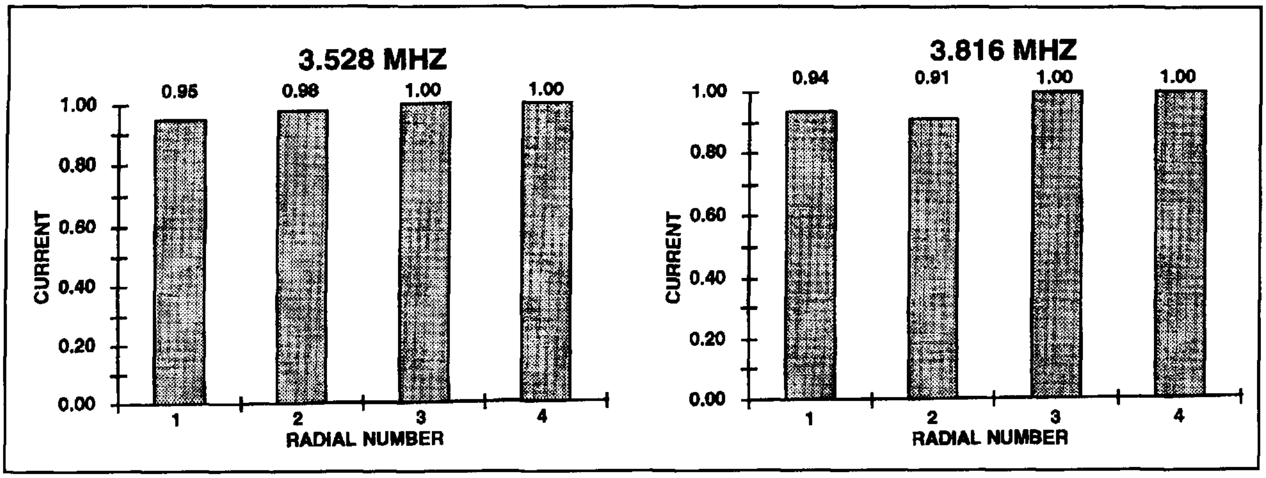

Figure 11. Radial currents with 1/8-wavelength radials.

behave identically? Can they have the exact same impedance at one frequency and across a frequency band?

The first problem with quarter-wavelength radials is readily apparent. A radial can only be 90 degrees at one frequency; therefore, there is a frequency dependency. Can radials within a system all have exactly 0 ohms impedance at the same frequency to begin with? No, because two or more radials can't be built exactly the same. The root issue is the sensitivity of radial impedance to variations, and the effect on current division. This is addressed by finding the impedance of a radial within an elevated radial system.

I set up a NEC-WIN model at 3.5 MHz to determine the impedance of a radial within an elevated radial system of four radials. It used a quarter-wavelength vertical antenna with four radials perpendicular to the radiator. I first set the radial lengths at 90 degrees. Then I lengthened and shortened them while using the reactance of a radial reactive component, like the radial inductor in **Figure 8**, at the feedpoint to null out the reactance of the radials.[6] The impedance of a single radial was taken as four times the conjugate of the reactive component's reactance. Radial impedances were found in free space, 10 feet above perfect ground, and 10 feet over average ground using a ground conductivity of 0.005 Siemens/meter and a dielectric constant of 13. The predicted radial impedances are shown in **Figure 17**. The main point of interest is the sensitivity near 90 degrees, which appears in more detail in **Figure 18** for the 89-91 degree range.

Figure 18 shows several ohms of variance for very small changes in radial length. Now several ohms isn't much, but it is relative to 0 ohms. The effect of small variances is demonstrated using Figure 19, which shows a parallel combination of four "ideal" radials 90 degrees

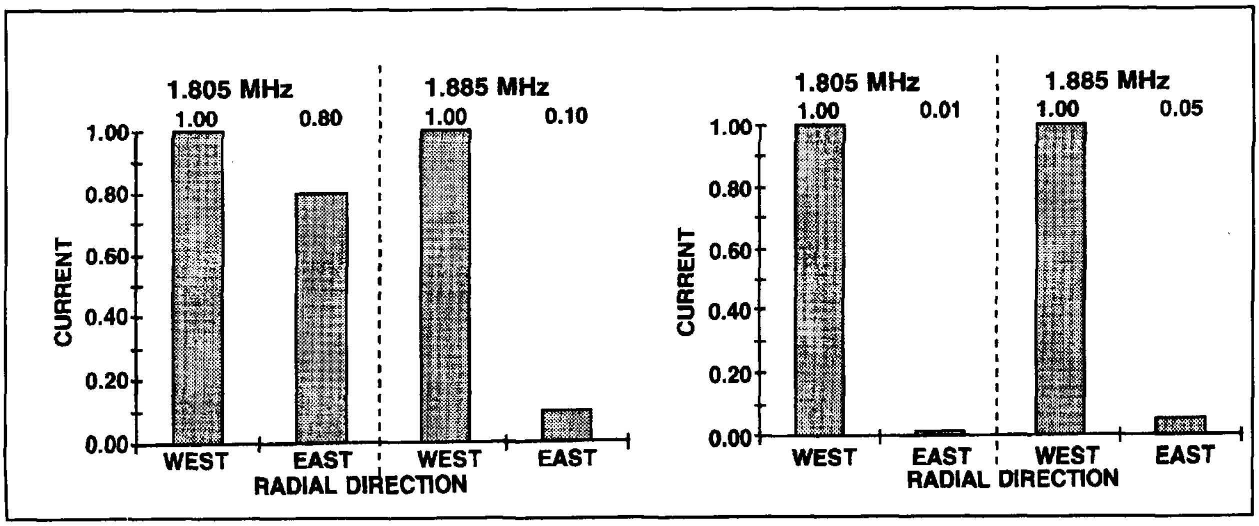

Figure 12. Measured elevated radial currents. (A) 160-meter ANT1. (B) 160-meter ANT2.

long and four that differ by 0.2 degrees. In **Figure 19A**, the four radials are exactly the same and theoretically have equal current. Realistically, it is impossible for four radials, or any other number, to each have an impedance of exactly 0 ohms. There will always be some slight variations in their construction that make their impedances different, as will nearby objects.

The effect of minor variations is illustrated in **Figure 19B**. In this diagram, it's easy to see why currents are unequal with radials set up to be a quarter wavelength. Any "non-zero" impedance is high compared to 0 ohms. Also, current division is "unstable"—switching from one radial to another with changing frequency. Test data show these effects, although at this time the full range of radial sensitivities due to different causes isn't known. Regardless, **Figure 19B** and test data indicate that a quarter wavelength isn't the right choice for elevated radial lengths. This length is too sensitive to minor variations that cause frequency dependent, unequal radial currents.

Why shortened radials?

The use of shortened radials results in a system less sensitive to the impedance variations of individual radials. This is illustrated by **Figures 20** and **21**, which show radials in the vicinity of 45 and 60 degrees where radial lengths vary by 1.0 degree. Although radial impedances in these figures differ, radial currents will be almost the same because impedance differences are small on a percentage basis. For all practical purposes, current division will be essentially equal with shortened radials because they are much less sensitive to variations that cause impedance differences. Although 45- or 60-degree radials don't form a

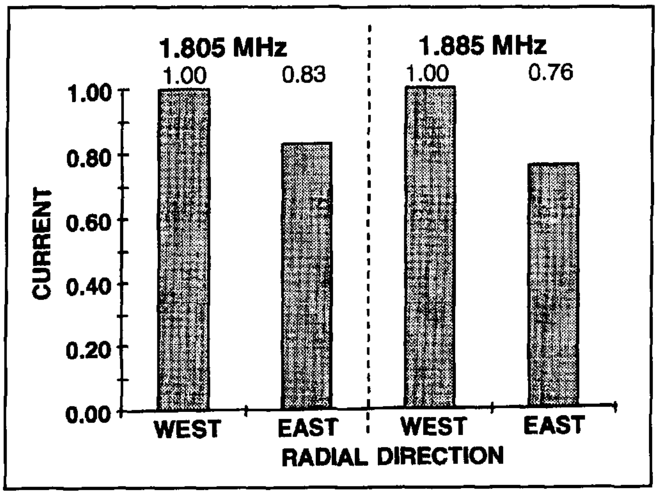

Figure 13. ANT1 radial currents after modification.

resonant system when used with a quarter-wavelength radiator, resonance may be achieved using a single radial inductor as shown in **Figures 20** and **21**. The key to making this system work is to use a single radial inductor as taught by Moxon.[5] The single inductor is used to null out the reactance of the parallel combination of radials. Because the radials are shortened, they will have similar impedances of much higher magnitude relative to 0 ohms.

There may be another way to ensure similar radial currents other than using shortened radials. **Figure 17** shows that 120-degree length radials have an impedance of approximately 330 ohms. It seems logical that lengthened radials should work as well as shortened ones. Of course, a radial capacitor would be needed instead of a radial inductor. There are potential applications for lengthened radials other than as part of a single-band system. This is illustrated

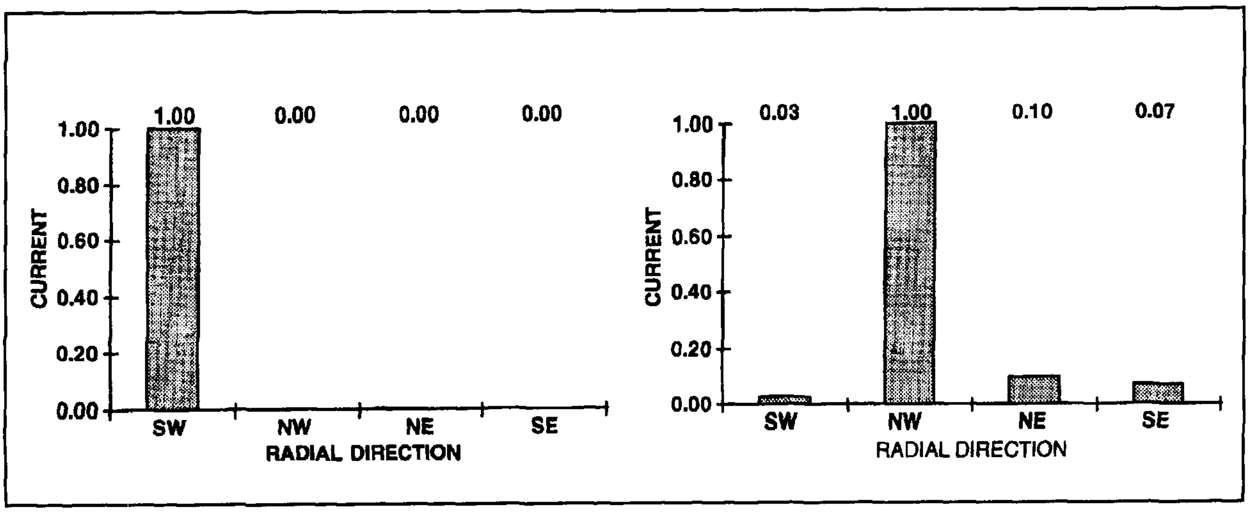

Figure 14. W7XU's radial currents at 1.805 MHz. (A) Initial radial lengths. (B) Equal length radials.

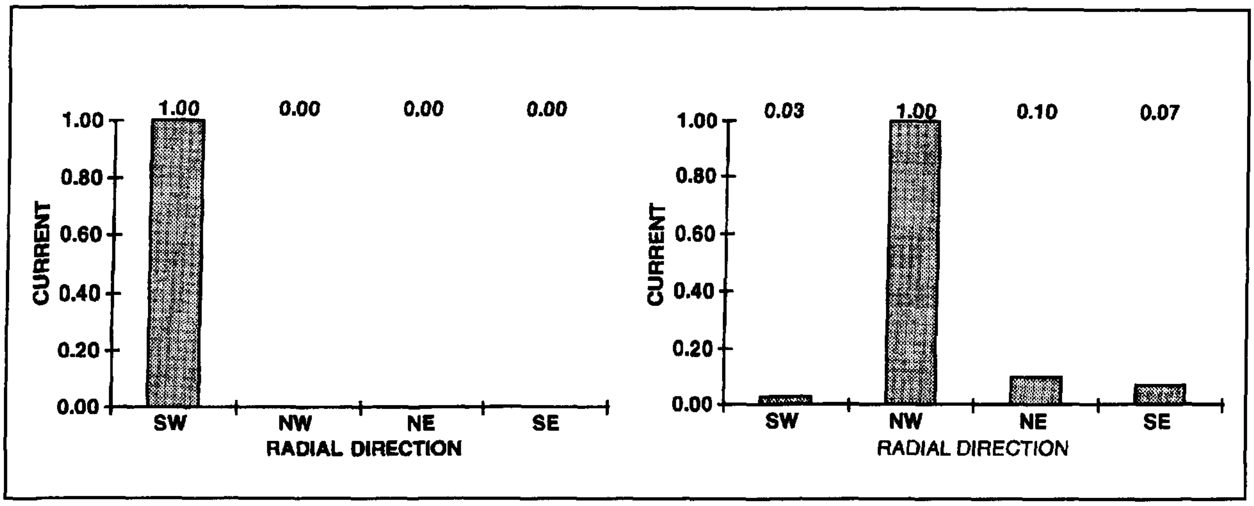

Figure 15. "Last attempt" radial current measurements at two frequencies. (A) Radial currents at 1.8 MHz. (B) Radial currents at 1.9 MHz.

in **Figure 22**, which shows a 160-/80-meter vertical with elevated radials approximately 60 degrees on 160 meters and 120 degrees on 80 meters. Because the radials are not a quarter wavelength on either band, equal radial currents may be realized on both 80 and 160 meters. For this to be a viable system, lengthened radials must be shown to work. In addition, 60- and 120-degree radials need to be shown to be far enough from 90 degrees for radial currents to be reasonably similar.

Additional radial testing

To answer my own question about lengthened and 60-degree radials, I tested radials of 135, 120, and 60 degrees. I used new radials with my elevated vertical in **Figure 1**. In the process, I measured radial currents for 90- and 45-degree radials as a check on tests made 18 months earlier. I didn't use the pick up loop shown in **Figure 6**. Instead, I chose a Palomar PCM-1 snap on current meter to make current readings at 3.528 and 3.816 MHz.* As I had done earlier, I normalized current readings to the largest within the set of readings for each frequency and each radial length. For 135-, 120-,90-, and 60-degree length radials, I positioned the current meter 32 feet from the feedpoint. For the 45-degree radials, I placed the meter 16 feet from the feedpoint. **Figures 23** through **27** show normalized radial currents at the two test frequencies.

*Palomar Engineers, Box 462222, Escondido, California 92046; (619) 747-3343.

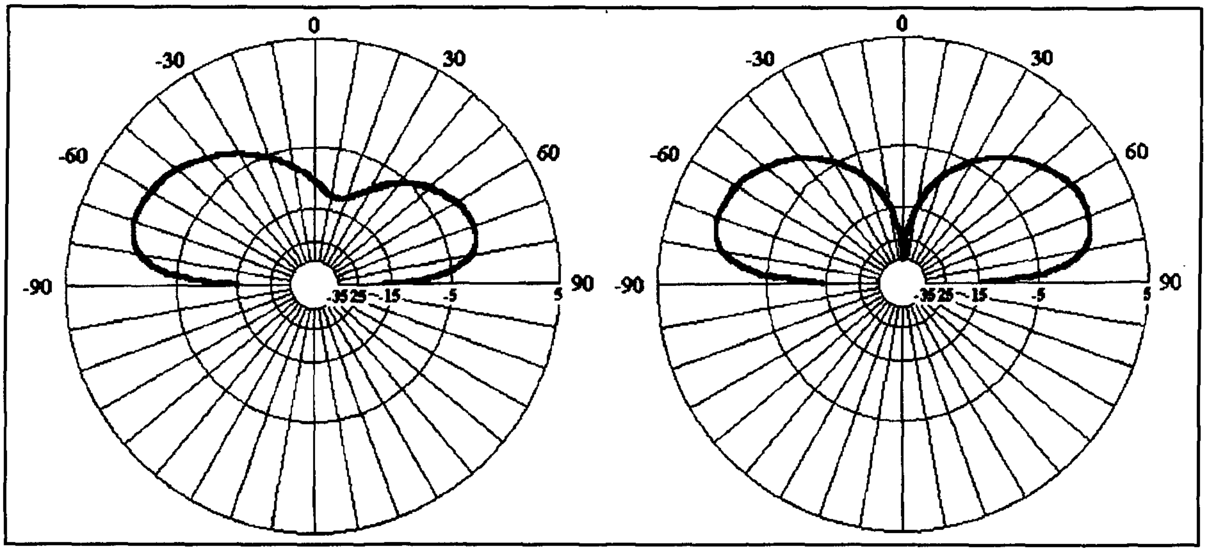

Figure 16. One radial compared to four radials with equal currents. (A) Broadside to single radial (radial to left). (B) Broadside to one radial pair (four equal current radials).

	"F"(FORWARD)	"S" (SIDE)	"B" (BACK)
GAIN (dBi)	-0.3	-1.64	-4.15
ANGLE (DEG)	25	24	20

Table 3. Single radial gain at 10 feet.

RADIATOR LENGTH (DEG)		RADIAL LENGTH (DEG) 45	60	90	120	135	150
45	GAIN (dBi)	-0.84	-0.57	-0.34	-1.49	-3.37	-6.25
	ANGLE (DEG)	25	25	24	23	22	24
60	GAIN (dBi)	-0.7	-0.48	-0.17	-0.59	-1.79	-4.38
	ANGLE (DEG)	25	25	24	24	22	22
90	GAIN (dBi)	-0.45	-0.32	-0.06	-0.01	-0.38	-1.18
	ANGLE (DEG)	23	23	23	22	22	21
120	GAIN (dBi)	-0.22	-0.18	0.01	0.12	0.05	-0.49
	ANGLE (DEG)	21	21	21	20	20	20
135	GAIN (dBi)	-0.13	-0.05	0.05	0.15	0.13	-0.16
	ANGLE (DEG)	20	20	20	19	19	19
150	GAIN (dBi)	0.04	0.01	0.12	0.17	0.17	0.01
	ANGLE (DEG)	19	21	19	18	18	20
180	GAIN (dBi)	0.35	0.33	0.28	0.22	0.17	0.07
	ANGLE (DEG)	16	16	16	16	16	16

Table 4. Maximum gain for elevated radial vertical antennas using four elevated radials 10 feet above ground at 3.75 MHz.

Figure 25 shows again that 90-degree radials have a strong propensity toward unequal radial currents. **Figure 23** illustrates that 135-degree radials work well—providing almost identical radial currents, as do the 45-degree radials (see **Figure 27**). The 120-degree radials have radial currents reasonably close to each other, while 60-degree radials have radial currents a bit farther apart. These data indicate that lengthened radials are another way to construct a system with similar radial currents. Also, 120- and 60-degree radials appear on the verge of having noticeably dissimilar radial currents. For design purposes, I suggest that lengths between 60 and 120 degrees not be used until more testing is done to better define this range.

Causes of unequal radial currents other than length variations

The sensitivity of radials to minor changes in length is demonstrated in **Figures 18** and **19B**. However this, in itself, doesn't totally explain the measured unequal radial currents. For example, you might expect quarter-wavelength elevated radials close to the same physical length to be near 0 ohms at some in-band frequency. At this frequency, small impedance variations around zero would cause very unequal radial currents (**Figure 19B**). Let's suppose this frequency is 3.5 MHz. When the system is used at 3.8 MHz, the radials are 7.7 degrees longer, or approximately six feet. Let's assume that when the radials were originally installed, they were within three inches of each other. Because a frequency change from 3.5 to 3.8 MHz corresponds to an electrical lengthening of six feet, we'd expect that length variations of three inches wouldn't affect radial current partitioning at the higher frequency because radial electrical lengths are now reasonably removed from 90 degrees. The same situation occurs if the radials were close to 90 degrees at 3.8 MHz. Operation at 3.5 MHz would result in a shortening of about 7.1 degrees which should, again, be far enough from 90 degrees for the effect of three-inch length variations to be insignificant. This suggests there may be causes of unequal radial currents other than minor length differences.

Highly variable ground conditions under the radial system may be a strong influence on radio current partitioning.[7] In **Reference 7**, Doty, Frey, and Mills measured radial currents in a 200-by-300-foot rectangular counterpoise six to eight feet above ground. The counterpoise had radial wires that tied together with a

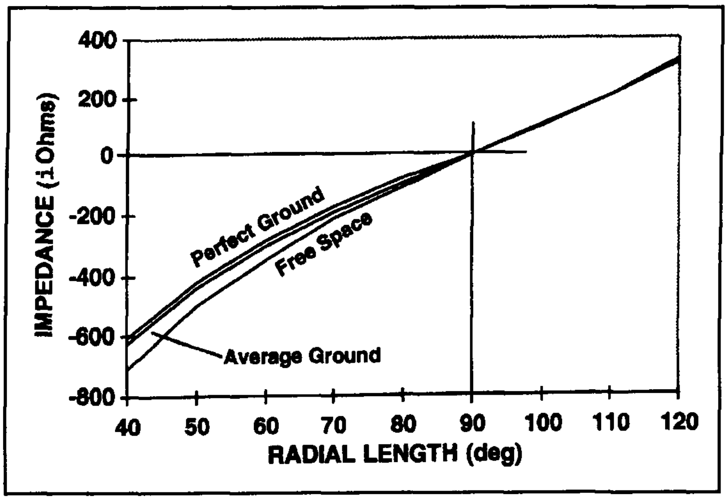

Figure 17. Impedance of a single radial.

vated radial systems. Further testing with equal length elevated radials in a symmetrical pattern is required to verify this theory. In time, we may find there are a number of contributing factors. Ground conductivity variations, length variations, and coupling with nearby conductors may all be significant contributors to radial current imbalances.

With the test data available at this time, and a basic understanding of an elevated radial's impedance sensitivity, I've concluded that a quarter wavelength is not an optimum size for elevated radials. Neither should we use elevated radial lengths between 60 and 120 degrees. I've also shown that shortened and lengthened radials can provide essentially similar radial currents when shorter than 60 degrees or longer than 120 degrees.

perimeter wire, as shown in **Figure 28**. The authors used a top-loaded 1/8-wave vertical on 160 meters set over the inner radial tie point, which was slightly offset from the center point of the rectangular perimeter. Tests performed using 48 radial wires indicated currents were much higher in certain areas of the counterpoise—by 15 to 1 as compared to similar areas elsewhere within the counterpoise. This led the authors to test the ground conductivity under the counterpoise. They found ground conductivity was higher under those radials that carried more current. From this test data, the authors concluded that the higher ground conductivity caused the higher radial currents.[8]

Based on the work performed by Doty, Frey, and Mills, it's safe to conclude that ground conductivity variations play a significant role in creating unequal currents in elevated radial systems.

Elevated radial vertical antenna design considerations

Since 1988, several articles have been published on elevated radials in professional journals and amateur radio magazines.[9–12] The most recent article, "Elevated Vertical Antennas for the Low Bands: Varying the Height and Number of Radials," appears in *The ARRL Antenna Compendium, Volume 5*.[13] Here Christman, KB8I, shows the results of numerous computer simulations used to predict the performance of quarter-wavelength verticals with quarter-wavelength elevated radials on 160 and 80 meters—where the height of the antenna above ground and the number of quarter-wave radials are varied.

On 80 meters, the author used heights of 5,

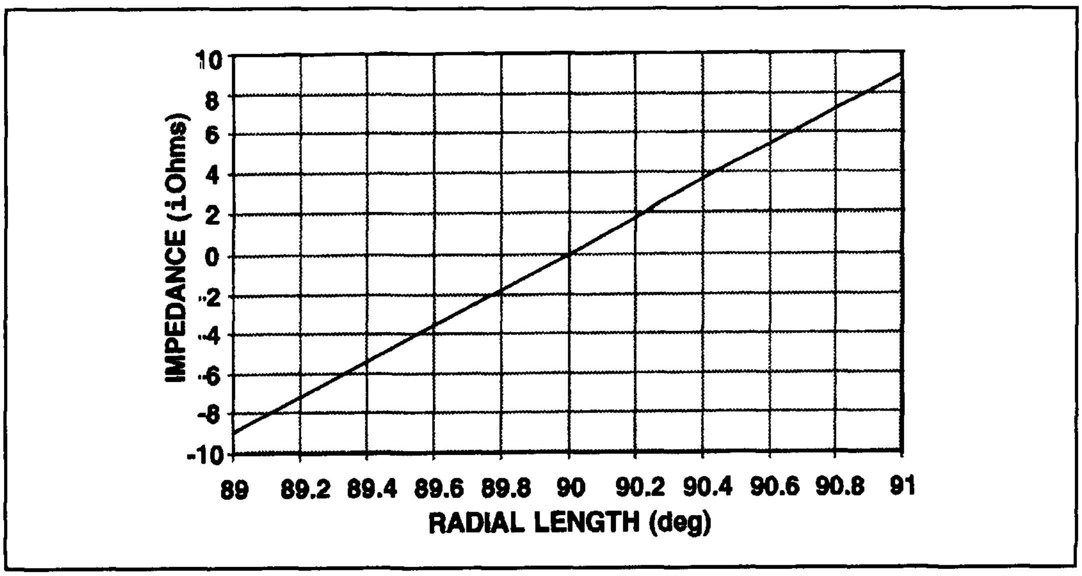

Figure 18. Radial impedance 10 feet above average ground.

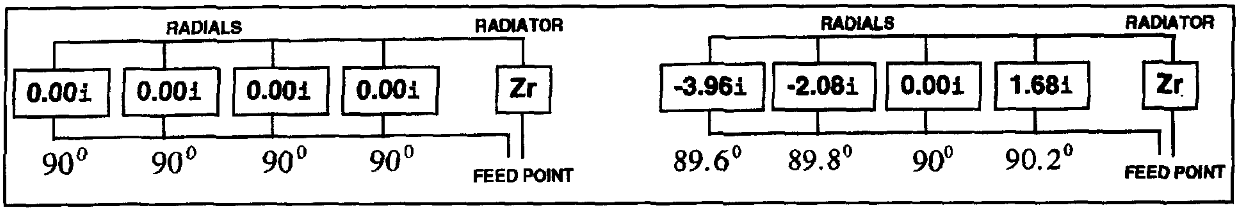

Figure 19. Radial system comparison. (A) Idealized radials. (B) Example radials.

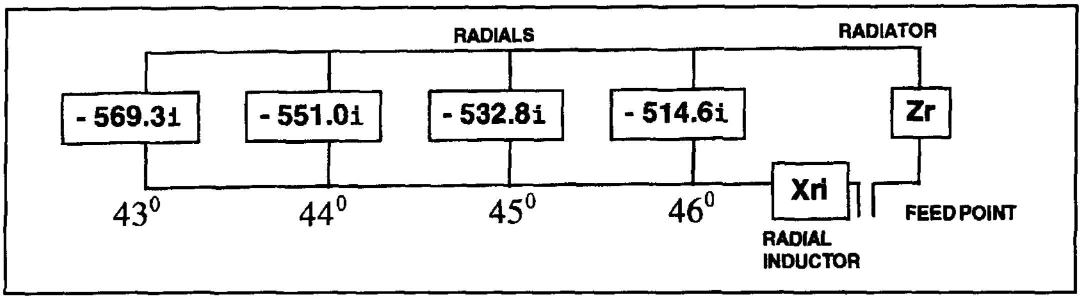

Figure 20. Radials in the 45-degree vicinity.

10, 15, 20, 25, and 30 feet, varying the number of radials at each height between four and 36. His results show that the simplest system, using four radials at five feet, has a maximum gain of -0.145 dBi. Christman's results also show a gain of 0.130 dBi for 36 radials at 30 feet, which is the most aggressive system evaluated. This is a net increase of 0.275 dB going from a very simple installation to one much more difficult to implement to obtain a seemingly small increase in gain of 6.5 percent. Based on the valuable insight provided by this article, I evaluated the gain of elevated verticals using shortened and lengthened radials subject to some practical considerations.

From a practical standpoint, elevated radials at a typical installation would likely be seven to 14 feet high to allow adequate clearance for people passing underneath. In addition, probably four radials, and certainly no more than eight, would be used. To look at elevated radial systems of this type, I used NEC-WIN to compute the gain of an elevated quarter-wave vertical 10 feet off the ground, with four and eight radials that varied in length from 45 to 150 degrees. This was done at 3.75 MHz using the same ground conditions and wire used to generate the data in the Antenna Compendium article. Christman used #12 aluminum wire, a ground dielectric constant of 13, and a ground conductivity of 0.004 Siemens/meter. My results assume the radial inductor used with the shortened radials has a Q of 200, and that the radial capacitor used with lengthened radials has no loss.

Results are plotted in Figure 29 and listed in **Table 2**. These show the maximum gain and the elevation angle where it occurs for different radial lengths. Included with the gains for a quarter-wave vertical using a range of radial lengths are gains for a system using a single 90-degree radial. For the single-radial system, a pattern similar to **Figure 16A** would result. This is representative of an elevated vertical where most of the radial system current flows in one radial. The gains shown for the single

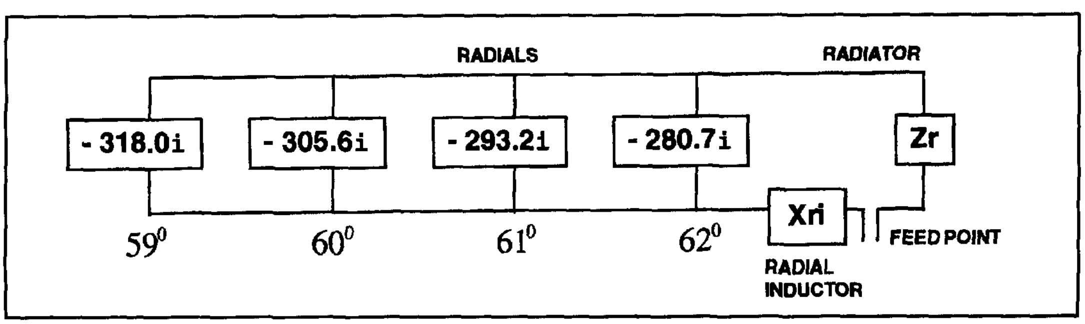

Figure 21. Radials in the 60-degree vicinity.

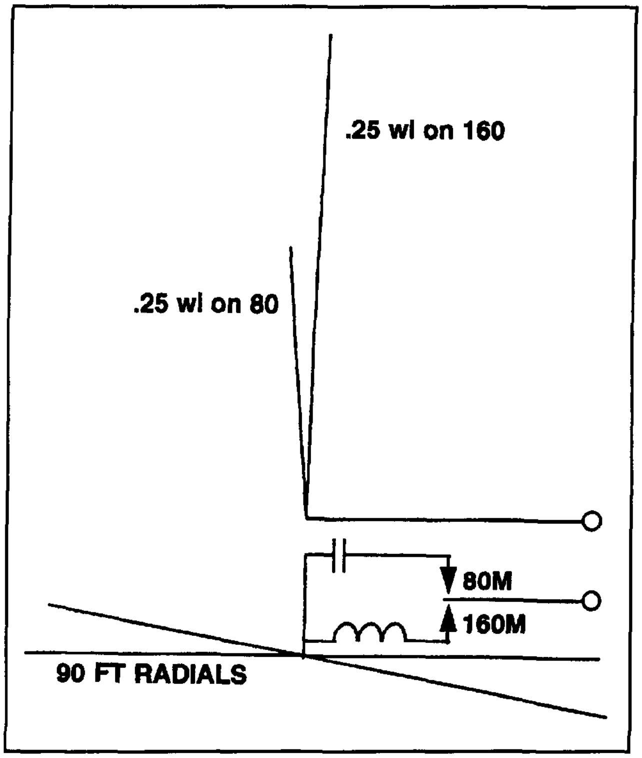

Figure 22. 160-/80-meter dualband vertical with elevated radials.

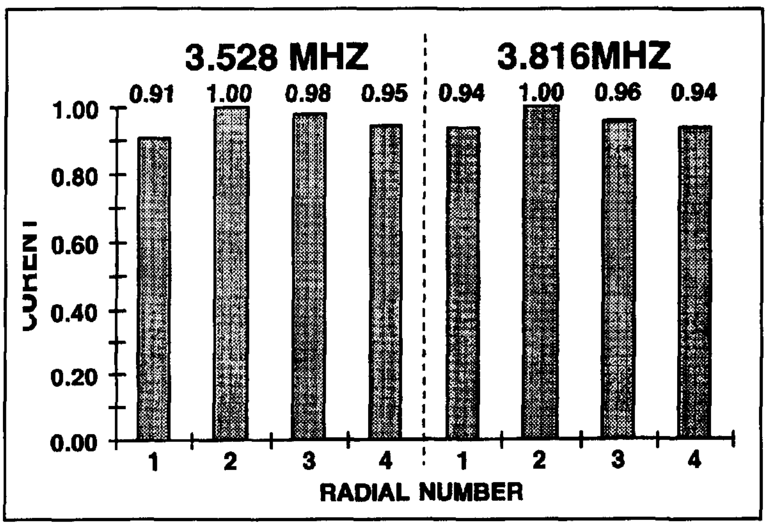

Figure 23. Radial currents with 135-degree radials.

radial system are the maximum forward gain "F," which is in the direction of the radial, "S," which is the maximum gain to either side, and "B," which is the maximum gain of the back lobe. Based on the nature of radial currents measured to date, it's prudent to consider the single radial system gains listed in **Table 3** as typical when quarter-wavelength radials are used. In addition, because the behavior of radials between 60 and 120 degrees is essentially unknown (except at 90 degrees), it's also prudent to consider single radial predictions as very conservative, worst-case estimates of performance using radials in this length range.

Modifications may be made to an existing elevated quarter-wavelength vertical using **Figure 29** as a design guide, with the constraint that radials should be less than 60 degrees or longer than 120 degrees with an upper limit of 135 degrees. (Although I didn't make radial current measurements with 150 degree radials, I feel their use should be discouraged based on the gains shown in **Figure 29**.) If an existing elevated vertical system with 90-degree radials is converted to 45-degree radials, a noticeable difference in performance should result. Omnidirectional coverage with a gain of -0.45 dBi will likely be of significant improvement compared to the gains shown for "S" and "B" in **Table 3**.

Figure 29 shows that eight lengthened 120-degree radials provide the most gain from a quarter-wavelength elevated vertical. Going from four radials 45 degrees long to eight radials 120 degrees long results in a 16 percent improvement of 0.64 dB. To achieve this increase in gain, one needs seven times the area to accommodate the longer radials. From a practical viewpoint, a gain improvement of 16 percent may not be worth the extra expense or effort. But, eliminating your quarter-wave radials and going to four 1/8-wave radials will improve overall performance significantly and take less room.

Elevated radial vertical antenna design for optimum performance

If you accept the idea that quarter-wavelength elevated radials are to be avoided, and that radials longer than 120 degrees and shorter than 60 provide a means to promote similar radial currents, then a new approach to elevated radial vertical antenna design comes to light. This approach involves predicting gains for a range of elevated radiator lengths in combination with a range of elevated radial lengths. The only rule is to use radials of less than 60 degrees or more than 120 degrees with an upper limit of 150 degrees. This rule eliminates 90-degree radials and those within 30 degrees of 90 and 180 degrees. To evaluate these design combinations, I used a NEC-WIN model with 45-, 60-, 90-, 120-, 135-, 150-, and 180-degree radiators 10 feet above ground in combination with four and eight elevated radials 45, 60, 90, 120, 135, and 150 degrees long at 3.75 MHz. Here, too, I used #12 aluminum wire, a ground dielectric constant of 13, and a ground conductivity of 0.004 Siemens/meter. Results appear in **Tables 4** and **5**, which list the maximum gain and the elevation angle where it occurs. Even though I don't recommend the use of 90-degree radials, they are included for comparison.

I used data from **Tables 4** and **5** to create the plots in **Figure 30**. These show the maximum gain for different given vertical radiator lengths over a range of elevated radial lengths. I did this using four and eight radials with lengths between 45 and 150 degrees in 15-degree increments. Predicted gains using 90-degree radials were discounted. At the top of the graph, a table lists the radial lengths used. The radiator length shown directly beneath that, yielded the gains used to make the maximum gain plots. For example, the maximum gain for a 60-degree radiator occurs when four radials are 60 degrees long and when eight radials are 120 degrees long. **Figure 30** may be used as a guide to build an elevated radial vertical antenna with the highest gain for a given radiator length.

Figure 31 also uses data from **Tables 4** and **5**. These plots show the predicted performance for different verticals using an elevated radial system that takes the smallest area to install. This uses data for four and eight radials that are 45 degrees long. The result is a design guide for elevated radial vertical antennas with limited area for the radials. The gain shown in **Figure 31** for a half-wave vertical with 45-degree long elevated radial brings up an interesting question. How short can elevated radials be with a half-wave vertical before gain starts to roll off appreciably? Hopefully someone will be interested in this question and provide the answer. Regardless of which radiator-radial combination you build, using lengthened or shortened radials will improve omnidirectional performance compared to using quarter-wavelength radials.

I haven't included data using different radial lengths with a 5/8-wavelength vertical in the tables and figures showing predicted performance for combinations of radiators and radials. The reason is illustrated in **Figures 32A** and **32B**, which provide elevation patterns for a 5/8-wave vertical over perfect ground and over average ground with an extensive ground radial system. The pattern over average ground shows significant energy in a higher angle lobe, as ground loss effects take their toll. No improvement is seen when elevated radials are used. This is illustrated in **Figure 33**, which shows elevation patterns using eight radials 45 degrees long and eight radials 120 degrees long. On 160 and 80 meters, it's desirable to have a lot of gain at low angles because most verticals on these bands are used for DX. It turns out a half-wave vertical is better here. Its gain using four or eight 1/8-wavelength radials is 0.35 and 0.40 dBi, respectively, at 16 degrees

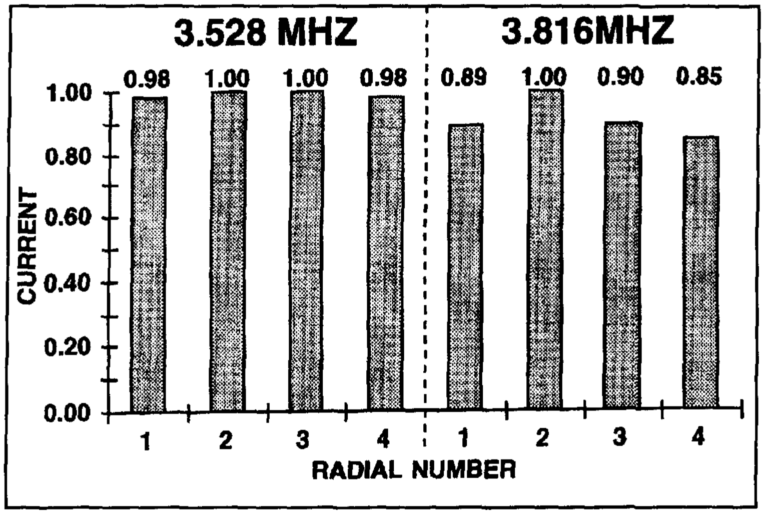

Figure 24. Radial currents with 120-degree radials.

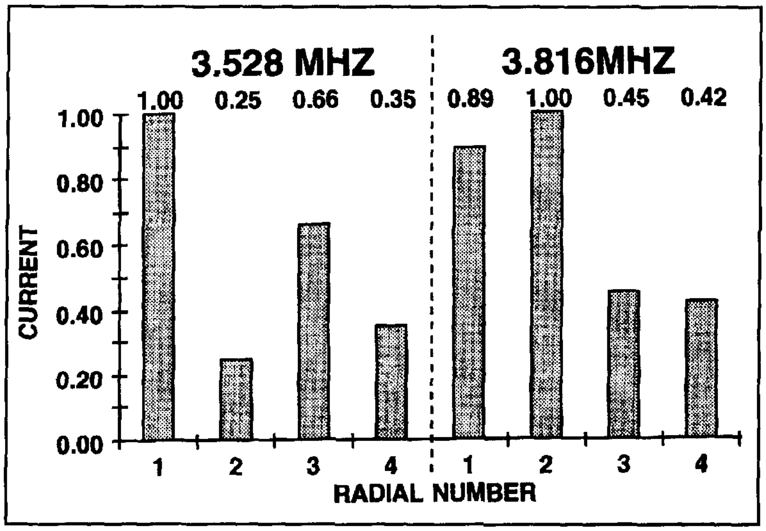

Figure 25. Radial currents with 90-degree radials.

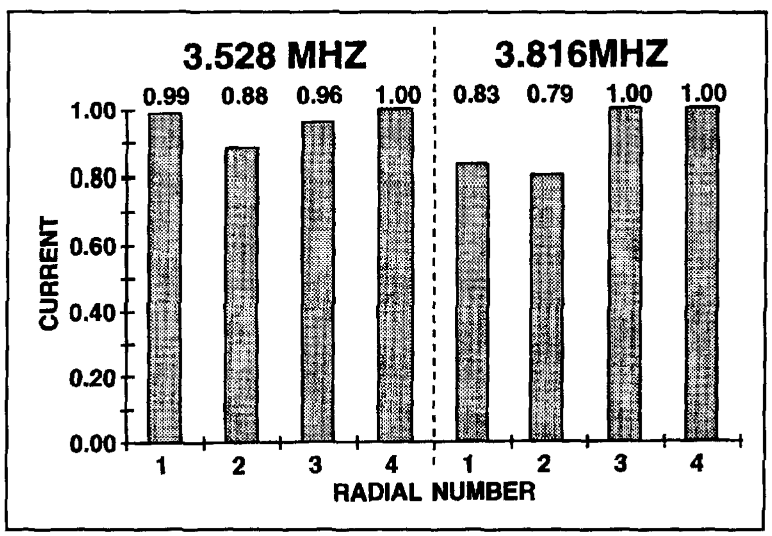

Figure 26. Radial currents with 60-degree radials.

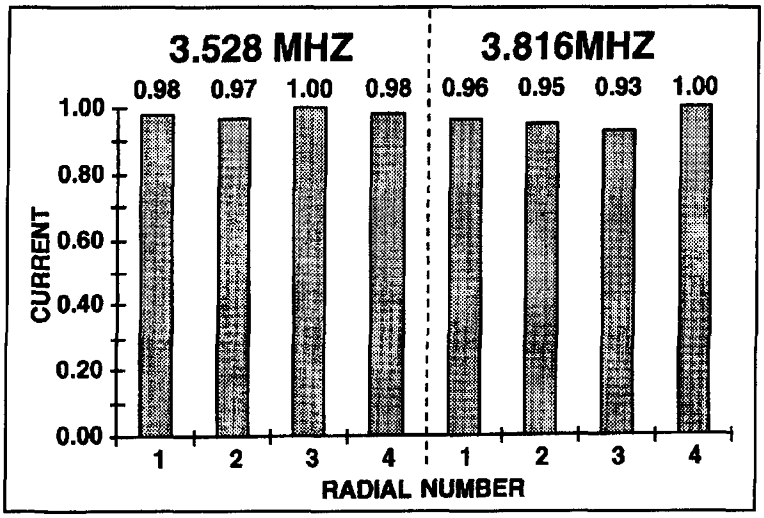

Figure 27. Radial currents with 45-degree radials.

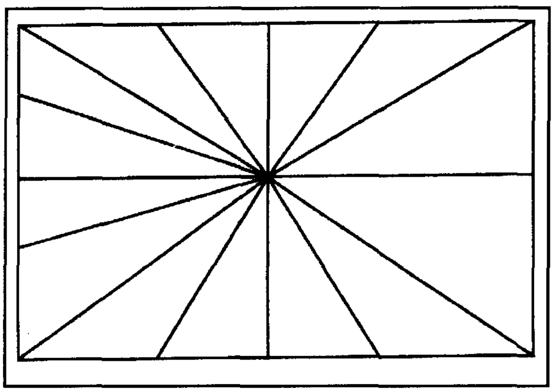

Figure 28. Counterpoise general outline.

elevation. This compares favorably with gains of 0.17 dBi at 17 degrees for the system shown in **Figure 33A** and 0.01 dBi at 17 degrees for **Figure 33B**.

Patterns for half-wave verticals offer better low-angle performance because there isn't a higher angle lobe to rob energy from a lower angle lobe. This is demonstrated in **Figures 34A** and **34B**, which show elevation radiation patterns for a 180-degree radiator ground mounted over average ground with an extensive ground radial system, and for a 180-degree vertical with four elevated radials 45 degrees long. From a practical standpoint, a half-wave vertical offers better gain and is easier to install.

Summary

What started out as curiosity about radial currents in my 80-meter antenna, has grown into a set of issues concerning the design of verticals using elevated radials. Based on the radial current measurements and analysis done so far, I believe antennas built with quarter-wave elevated radials aren't working as well as people believe. Also, it's my opinion that the root issues guiding design procedures to build optimal elevated radial vertical systems have been identified and that the design recommendations offered here are accurate. Although performance studies were performed on 80 meters, the results should apply to 160-meter antennas as well. Hopefully, the design guidelines presented here will be improved and expanded upon, as others investigate the design and use of elevated radials.

For now, I believe that elevated radial systems should be designed following one cardinal rule: Elevated radials should not be a quarter

RADIATOR LENGTH (DEG)		RADIAL LENGTH (DEG)					
		45	60	90	120	135	150
45	GAIN (dBi)	-0.77	-0.5	-0.11	-0.33	-1.17	-2.87
	ANGLE (DEG)	26	25	24	23	22	23
60	GAIN (dBi)	-0.64	-0.42	-0.05	0	-0.37	-1.58
	ANGLE (DEG)	26	25	24	25	22	23
90	GAIN (dBi)	-0.39	-0.26	0	0.19	0.15	-0.3
	ANGLE (DEG)	23	23	23	22	22	22
120	GAIN (dBi)	-0.17	-0.09	0.06	0.2	0.24	0.13
	ANGLE (DEG)	21	21	21	20	20	20
135	GAIN (dBi)	-0.05	0	0.1	0.21	0.24	0.2
	ANGLE (DEG)	20	20	20	19	19	19
150	GAIN (dBi)	0.09	0.11	0.17	0.22	0.25	0.23
	ANGLE (DEG)	19	18	19	18	18	18
180	GAIN (dBi)	0.4	0.38	0.33	0.28	0.25	0.19
	ANGLE (DEG)	16	16	16	16	16	16

Table 5. Maximum gain for elevated radial vertical antennas using eight elevated radials 10 feet above ground at 3.75 MHz.

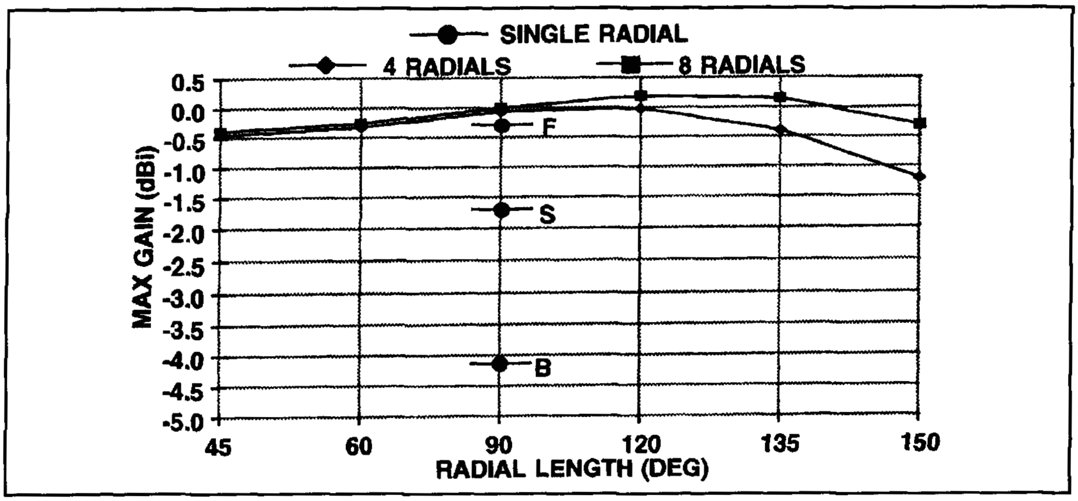

Figure 29. Maximum gain of an 80-meter elevated quarter-wavelength vertical with different length radials 10 feet above ground.

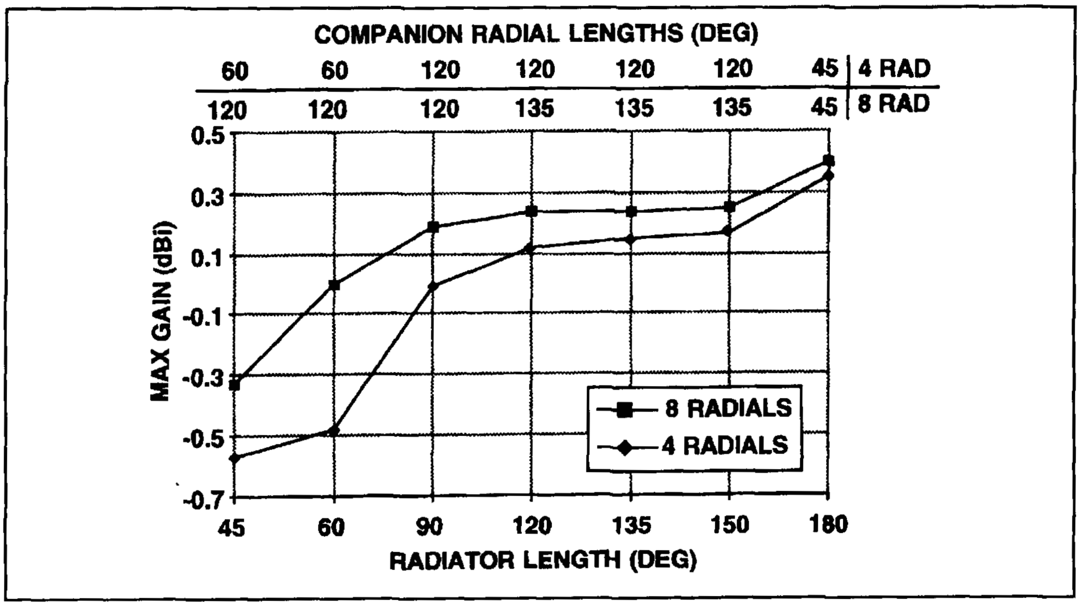

Figure 30. Maximum vertical antenna gain using four or eight elevated radials 10 feet above average ground at 3.75 MHz.

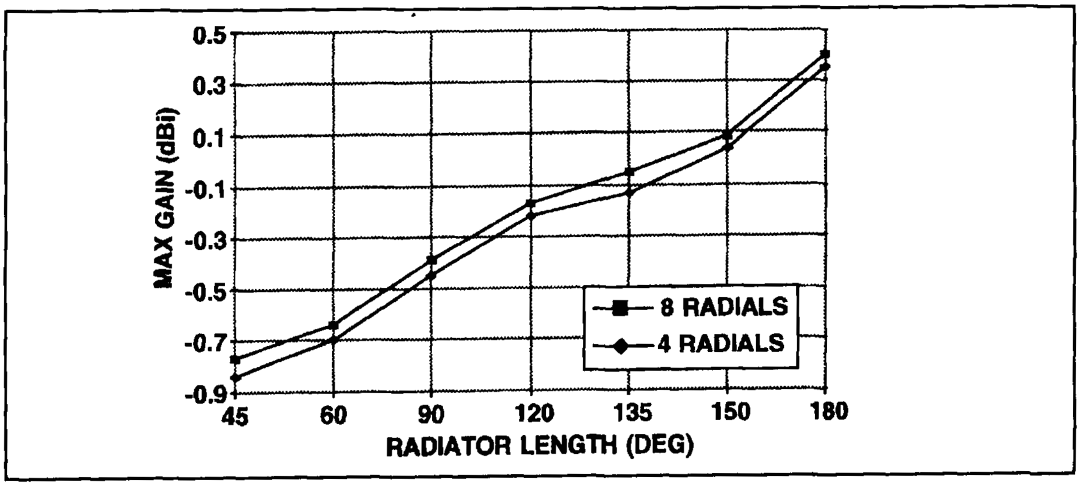

Figure 31. Vertical antenna gain using four or eight elevated eighth-wave radials 10 feet above average ground at 3.75 MHz.

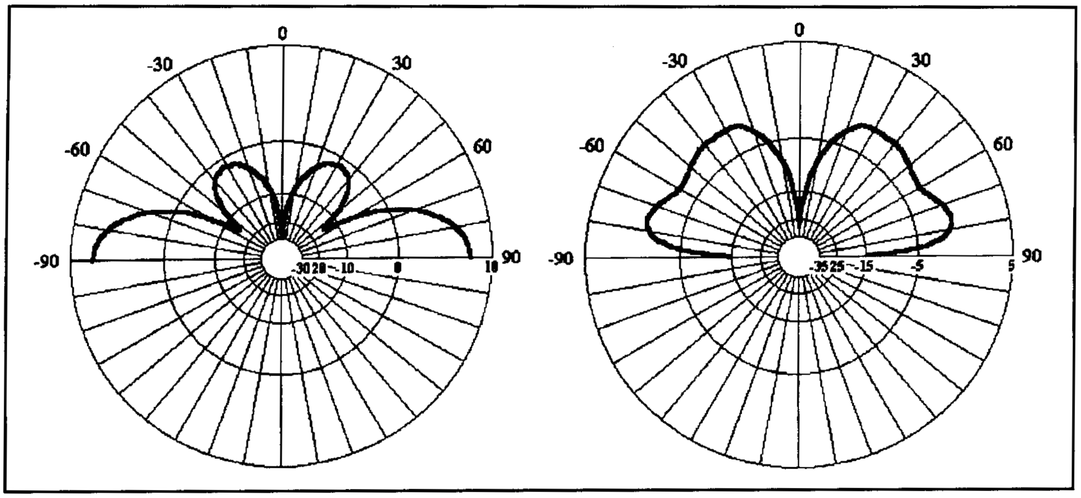

Figure 32. Elevation patterns for ground-mounted 5/8-wavelength verticals at 3.75 MHz. (A) Perfect ground. (B) 120 quarter-wave radials on average ground.

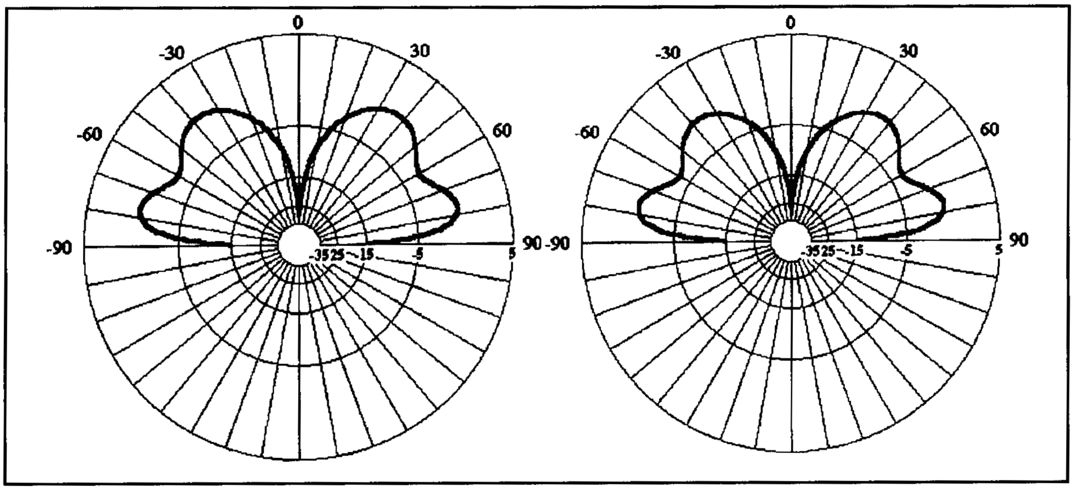

Figure 33. Elevation patterns for 5/8-wavelength verticals at 3.75 MHz with elevated radials 10 feet above average ground. (A) Eight elevated 45-degree radials. (B) Eight elevated 120-degree radials.

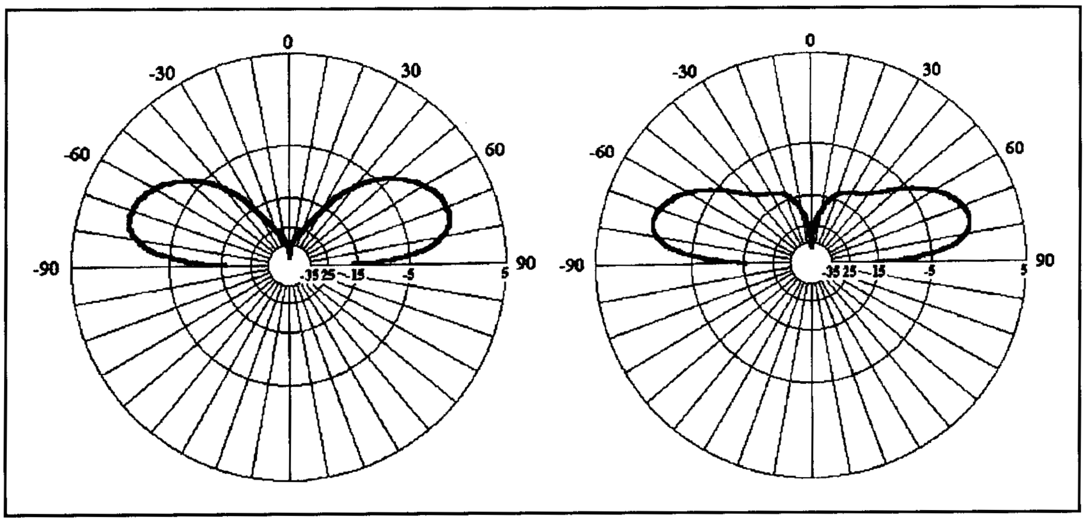

Figure 34. Elevation patterns for 1/2-wavelength verticals at 3.75 MHz. (A) 120 quarter-wave radials on real ground. (B) Four elevated 45-degree radials 10 feet high.

wavelength. Elevated radials should be less than 60 degrees or longer than 120 degrees, with an upper limit of 135 degrees. **Figures 30** and **31** along, with **Tables 4** and **5** are provided as design guides. Using these, you should be able to install an elevated radial vertical that works up to its potential and is optimized for your area and height restrictions. If you have an elevated radial system, I encourage you to make current measurements. You may be amazed at what you find. ■

REFERENCES

1. Bill Shanney, KJ6GR, "Understanding Elevated Vertical Antennas," *Communications Quarterly*, Spring 1995, pages 71–76.
2. Bill Shanney, KJ6GR, "Understanding Elevated Vertical Antennas," *Communications Quarterly*, Spring 1995, page 72.
3. Bill Shanney, KJ6GR, "Understanding Elevated Vertical Antennas," *Communications Quarterly*, Spring 1997, page 75.
4. Bill Shanney, KJ6GR, "Understanding Elevated Vertical Antennas," *Communications Quarterly*, Spring 1995, page 72.
5. Les Moxon, G6XN, "Ground Planes, Radial Systems, and Asymmetric Dipoles," *The ARRL Antenna Compendium, Volume 3*, 1992, pages 19–27.
6. Personal Correspondence, Mr. Warren Bruene, W5OLY, December 6, 1996, (Mr. Bruene suggested this method.)
7. Arch Doty, K8CFU, John Frey, W3ESU, and Harry Mills, K4HU, "Efficient Ground Systems for Vertical antennas," *QST*, February 1983, pages 20–25.
8. Arch Doty, K8CFU, John Frewy, W3ESU, and Harry Mills, K4HU, "Efficient Ground Systems for Vertical Antennas," *QST*, February 1983, page 23.
9. A.M. Christman, R.D. Radcliff, et al., "AM Broadcast Antennas with Elevated Radial Ground Systems," *IEEE Transactions on Broadcasting, Volume 37, No. 2*, September 1991.
10. Al Christman, KB8I,"Elevated Vertical Antenna Systems," *QST*, August 1988, pages 35–42.
11. A.M. Christman and R.D. Radcliff, "Using Elevated Radiuals with Ground-Mounted Towers," *IEEE Transactions on Broadcasting, Volume 37, No. 2*, September 1991.
12. A.M. Christman and R.D. Radcliff, "AM Broadcast Antennas with Elevated: Varying the Number of Radials and Their Height Above Ground," *IEEE Transactions on Broadcasting, Volume 42, No. 1*, March 1996, September 1996.
13. A.M. Christman, KB8I, "Elevated Vertical Antennas for the Low Bands: Varying the Height and Number of Radials," *The ARRL Antenna Compendium, Volume 5*, pages 11–18.

Comments by WXØB

I invited Dick over to measure my 160-meter system's radial currents when I learned that he had shortened the elevated radials on his 80-meter vertical antenna. I'd suspected that my radials weren't performing properly, even though I had "tuned" them as low dipoles, as is commonly done. My suspicion was confirmed when I built a second 160-meter wire vertical suspended from another tower. Like the first one, the second vertical also had two radials. I found it to be somewhat directional. Having read everything I could about verticals, I decided I really needed to know what was going on with the antennas. Measurements were needed.

Dick jumped at the chance to make measurements. He showed me that, indeed, only one radial was working with one antenna, and the currents were out of balance with each other. I might as well be using only one radial per antenna. We actually had to check the feedpoint to verify that one of the radials was connected. I was certain that it must have a bad solder joint, but it didn't. The currents were terribly unbalanced.

Weeks later I had time to shorten the radials of one antenna to a 1/8-wavelength. I loaded them with a common 20-µH inductor and tuned the system to resonance by moving a coil tap. I calculated the inductance needed by modeling the antenna in NEC2 with shortened radials. The model turned out to be very accurate. I needed to change the inductance by only a small amount. With modifications made, Dick returned to make another set of measurements. Bingo! The currents were now close to being equal—even though my radials aren't very symmetrical. Although they are identical in length, they droop at different angles from the feedpoint. But this proved to me that the technique of shortening radials works and probably once I got them level, things would be even better. Another benefit of shortening the radials is that I can now fit four of them into my yard instead of being limited to two per antenna.

During a recent 160-meter CW contest, the single modified vertical seemed to perform much better toward Europe now that the radial system was working as it should. Before, with only the west radial working, I could tell I had a problem with the patter. A computer simulation showed a much reduced lobe when modeling one radial "pointing" in the wrong direction.

I have now finalized my dual vertical system design with confidence. It will use two quarter-wavelength verticals spaced a quarter wavelength apart. Each will have four shortened radials 45 degrees long. The feed is a Hybrid coupler I made and modeled in PSPICE. It will result in a switchable cardiode pattern to the East and West with gain, now that the radials work. Dick and I plan to make signal strength measurements this spring to verify it performance.

I encourage anyone with an elevated radial system, be it one or more verticals, to measure your radial currents. You may be amazed at what you find.

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EZNEC FOR DOS

The marriage of ELNEC and NEC-2

Computer antenna modeling programs provide antenna designers, builders, and users with a wealth of predictive information on antenna performance. The most familiar antenna modeling program outputs are azimuth and elevation plots of far field patterns regularly seen in radio publications and illustrated by **Figure 1**. Typical programs also provide detailed tables of antennas currents and feedpoint data, including impedance, current, voltage, and SWR referenced to a baseline value (50 ohms for example).

Once the program user learns how to model an antenna structure as a set of straight wires, segmented to levels that provide accurate results, and to place the source or feedpoint and any loads at correct points along the modeled structure, the method-of-moments calculation system generates mutual impedance and current data from which other outputs, such as far field antenna patterns, are generated.

For a number of years, only MININEC was available to commercial developers of PC programs. Among standard MININEC weaknesses have been difficulties in modeling large Yagis, antennas whose geometry involved sharp

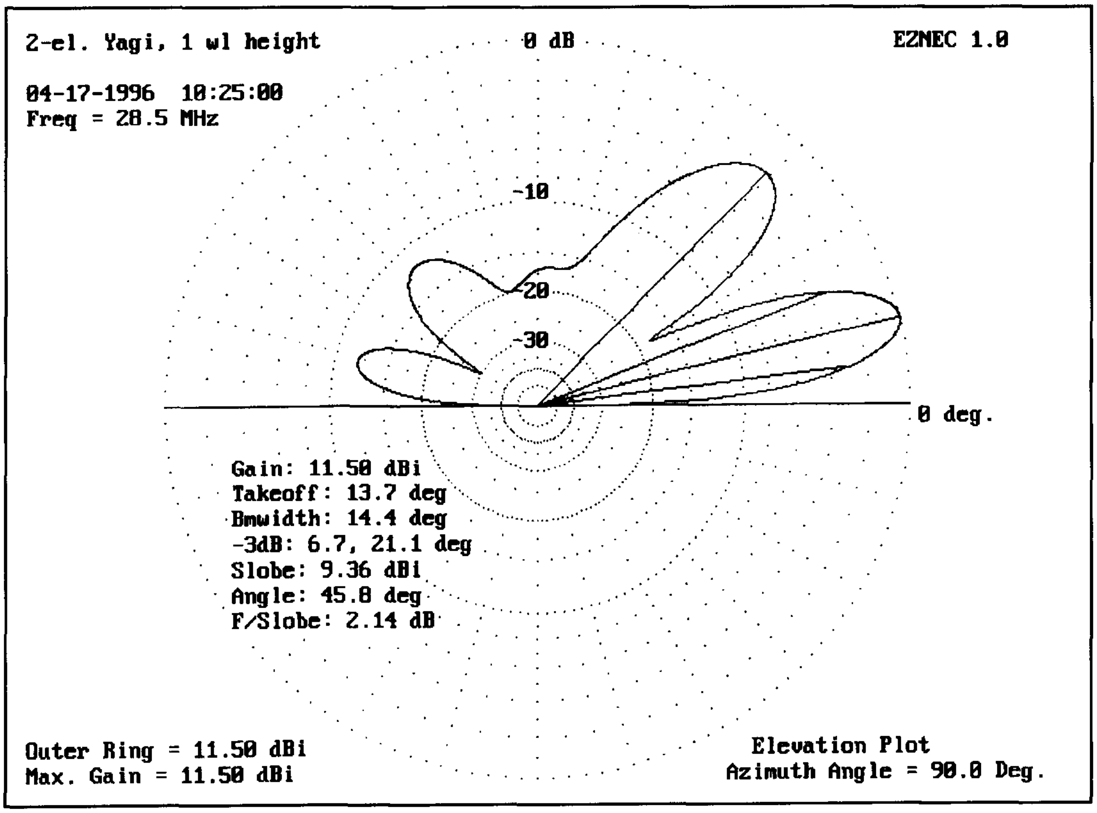

Figure 1. A typical antenna modeling program (in this case, EZNEC) elevation pattern derived from NEC-2 calculations of a Yagi antenna 1 wavelength above medium ground conditions. Note: antenna modeling programs do not themselves take into account the many detailed variables of terrain, but are limited to flat or stepped terrain and to objects that can be modeled as wire frames.

```
                   --------------- WIRES ---------------

 Wire Conn. --- End 1 (x,y,z : in)  Conn. --- End 2 (x,y,z : in)  Dia(in) Segs

   2  W3E1   0.000,  0.000,631.200        102.000,  0.000,631.200   # 28      3
   3  W4E1   0.000,  0.000,631.200         88.335, 51.000,631.200   # 28      3
   4  W5E1   0.000,  0.000,631.200         51.000, 88.335,631.200   # 28      3
   5  W6E1   0.000,  0.000,631.200          0.000,102.000,631.200   # 28      3
   6  W7E1   0.000,  0.000,631.200        -51.000, 88.335,631.200   # 28      3
   7  W8E1   0.000,  0.000,631.200        -88.335, 51.000,631.200   # 28      3
   8  W9E1   0.000,  0.000,631.200        -102.00,  0.000,631.200   # 28      3
   9 W10E1   0.000,  0.000,631.200        -88.335,-51.000,631.200   # 28      3
  10 W11E1   0.000,  0.000,631.200        -51.000,-88.335,631.200   # 28      3
  11 W12E1   0.000,  0.000,631.200          0.000,-102.00,631.200   # 28      3
  12 W13E1   0.000,  0.000,631.200         51.000,-88.335,631.200   # 28      3
  13  W1E2   0.000,  0.000,631.200         88.335,-51.000,631.200   # 28      3
                         ---- (13 total) ----                     Tot segs:  61
    End 1 (x, y, z in in.; conn.; or length)?

     [x],[y],[z] = Coordinates (leave coord blank for no change)
     W#E! = Conn. to end ! of wire #       RA# = Rotate azimuth # deg. (+ccw)
     L# = Change length to #               RE# = Rotate elev. # deg. (+up)
     L+#, L-# = Change length by #
   Use left and right arrow keys to select column; press <ESC> when done
                Preserve Conn = ON
```

Figure 2. A screen print of the group wires modification table, with its list of modification aids. W#E! permits connection by wire end selection. RA# and RE# permit wire rotation around the unselected wire end. L#, L+#, and L-# permit one or many wires to be lengthened or shortened. In this 12-spoke capacity hat model (vertical element not shown), the length (or wire diameter, or the number of segments per spoke) can be changed for all 12 spokes with a single entry.

angles, long multi-wavelength antennas, and all antennas at very low heights. Recently, the larger FORTRAN predecessor of MININEC, known as NEC, became available for development, along with compilers that made NEC-2 usable on PCs. NEC-2 handles larger antenna systems with a more accurate ground calculation system than MININEC, but it has its own limits (at least until NEC-4 is made available for commercial development).

A marriage between ELNEC and NEC-2

EZNEC is a marriage between the popular user interface that Roy Lewallen, W7EL, developed for MININEC (ELNEC) and the more powerful NEC-2 antenna modeling calculation engine. The result is a seamless compiled antenna modeling program for DOS (or a DOS application under Windows). Those familiar with ELNEC will feel instantly at home with EZNEC, but susceptible to all the initial dangers of ignoring the differences between NEC and MININEC.

Both ELNEC and EZNEC are designed to assist beginning modelers in mastering the art of wire antenna model creation, as well as to speed model creation and revision for experienced modelers. The EZNEC antenna file creation system always begins with a past or default antenna file, which is then modified to the intended configuration. The system uses an extensive set of menu selections and "fill-in" tables, guaranteeing a model file that will run. To the ELNEC array of choices, EZNEC has added NEC's transmission line capabilities, although these lines are mathematical rather than physical models. Also added is SOMNEC, the Norton-Sommerfeld ground calculation system for accurate modeling of antennas at low heights. The ELNEC ground is retained as an option for certain types of modeling problems, most notably with vertical antennas.

From the main menu, the user can select the antenna material, units of measure, plot type, reference level, ground type and description, plot range, plot step size, and outer ring of the plot (as a specific level or automatically keyed to the maximum gain of the plot). Separate subscreens provide for specification of source position and type and transmission lines. An "options" screen allows alteration of longer-term settings, such as the subdirectory for model files, plot annotation, or plot border information. Although the default screen output is a far-field pattern plot, a variety of ASCII

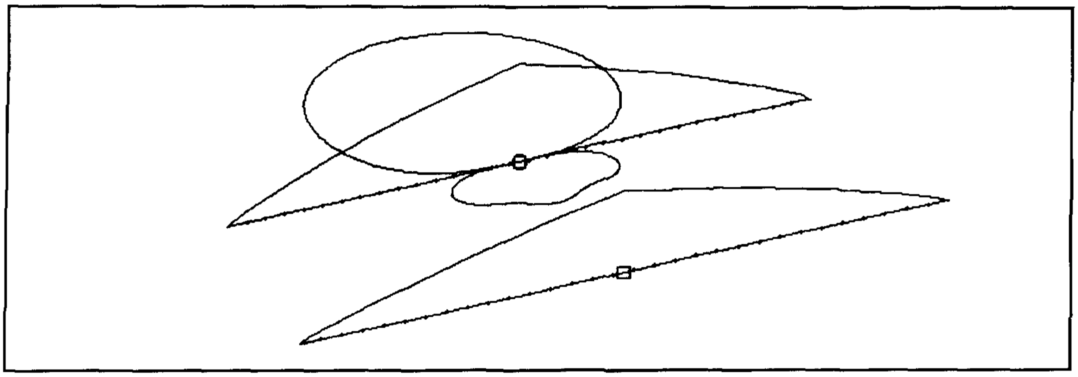

Figure 3. A three-dimensional view of a two-element Yagi showing the wire segmentation, the source and load positions, the relative current amplitude along the wire, and a reference azimuth pattern. Other information may also be displayed.

output reports are also available, including a basic analysis (gains, bandwidth, sidelobes, ratios), source or feedpoint information, and currents. Frequency sweep output data is stored in a user-specified file for screen browsing, with printing outside the main program.

The "Wires" table

The "Wires" table is notable because of the many mechanisms available for modifying wire specifications, either individually or in groups. **Figure 2** shows an EZNEC group wire modification screen, with an array of means to alter antenna wire connections, dimensions, segmentation, and geometric orientation along any of the X, Y, or Z axes. For example, a capacity hat may be constructed by knowing only the length of one radial and then replicating that radial and rotating it to a new angle. The "maintain connection" feature permits modification of an entire antenna structure with a minimum number of entries, an advantage to repetitive or incremental modeling sequences.

Upon leaving the wires table, EZNEC performs a useful "guidelines check" to see that the model meets program requirements for such measures as wire length-to-diameter ratio. It checks at both conservative and minimum levels and offers automated or manual wire revisions.

The compiled FORTRAN computation engine has been designed to use the maximum RAM available, and to use hard disk space as virtual RAM only if actual RAM is insufficient. This measure can speed up runs by reducing the amount of hard disk access during the NEC-2 calculations. EZNEC has a 500-wire-segment limit—more than enough for arrays such as log-periodics or five-band quads, or for simpler antennas with modeled surrounding objects. For truly complex antenna arrays, elevated grounds, and terrain objects, there's EZNEC-M, which has been tested at over 1,000 wires and 2,800 segments with reliability. For speed of operation, EZNEC saves and stores SOMNEC ground solutions, which are reused if within 7 percent of the current model demands. The three-dimensional antenna view feature, if selected after a NEC run, can show a variety of information, including antenna currents and a simplified view of the selected pattern, as demonstrated in **Figure 3**.

Viewing the results

Far-field patterns calculated from NEC results may appear in graphical or tabular form. Graphical patterns can be saved, recalled, and overlaid on other patterns. When used as a Windows application, patterns can be saved to the clipboard and imported to other documents. Plots are printable on HP LaserJet, DeskJet (color or monochrome), and Epson-type printers. Minimum computer requirements for EZNEC are a 386 CPU, coprocessor, 2 MB RAM, and EGA or better graphics.

Although more powerful than MININEC, NEC-2 has its own inherent drawbacks. For example, changes of antenna wire diameter, especially when enclosing an angle, can create unreliable output data (although MININEC handles these cases with reasonable accuracy). Moreover, EZNEC does not implement some of the more rarely used NEC-2 possibilities; for example, near field analysis, ground wave analysis, patches, wire arcs, helices, and networks. Because of program integration, model files are not accessible as ASCII files, although full model data can be printed from within the program.

EZNEC (along with ELNEC and EZNEC-M) is available from Roy Lewallen, W7EL, P.O. Box 6658, Beaverton, Oregon 97007. The current price of the program is $89 postpaid within the U.S. and Canada, with a $3 airmail fee for other countries. ■

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THE LAZY-H VERTICAL

A versatile antenna for DX work

Verticals can be effective DX antennas on 80 and 160 meters. There are, however, some practical problems involved in building such antennas. A quarter-wavelength vertical, for example, will be ≈ 68 feet high at 3.510 MHz and 131 feet at 1.840 MHz. If you use buried radials, you'll need an extensive ground system of radials >0.2 wavelength for efficient operation. Both the height and the ground system can make such a project formidable and put this kind of antenna out of reach for many hams.

What's needed are designs whose performance approaches these ideals, but don't require the height, ground area, and/or complexity of ground system. Wire antennas that may be hung between a tower and a tree or two trees would be quite useful. It's also important that the designs be very flexible in their dimensions, mechanical details, materials, etc., because each situation is different and the antenna must be crafted to fit the available site and resources. This may sound like a tall order, but you can come surprisingly close to filling it.

Al Christman, KB8I,[1,2] has shown that a relatively simple elevated radial system,*isolated* from ground, can provide performance comparable to large buried radial systems. Also, it has long been known that the height of a vertical may be significantly reduced while maintaining good efficiency, by using top loading.[3] Shortening the top-loaded antenna reduces the bandwidth, even if it doesn't significantly reduce the efficiency. This isn't necessarily a problem for DX work on the low bands, because DX operation is highly localized in the "DX windows." With 80 meters, there are two windows—3.510 (CW) and 3.790 (SSB) MHz. Even using a relatively short antenna, it's possible to get 50 to 100 kHz of 2:1 SWR bandwidth. Because the two DX windows are almost 300 kHz apart, some trickery is needed to accommodate both windows with a single antenna. As I'll show you later, both of these

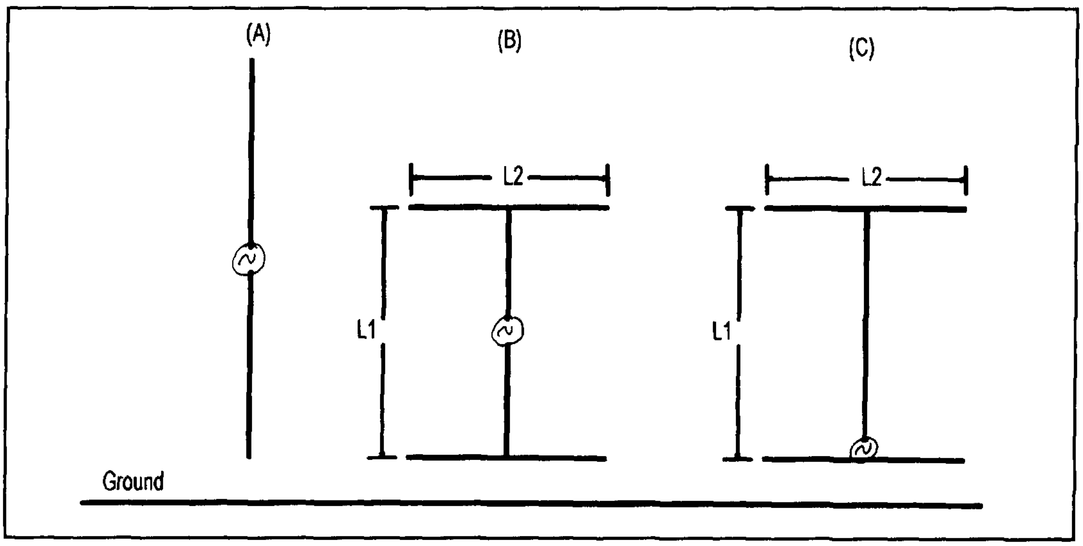

Figure 1. (A) A half-wave vertical dipole. The antenna can be shortened by adding perpendicular wires at the ends (B) and (C).

Table 1. Antenna Comparison at 3.510 MHz

ant	L1	L2	Z_{middle} Ω	Z_{end} Ω	peak gain, dB	peak angle °	wire loss -dB	2:1 SWR Bw kHz
λ/2	137'	0	91	>5000	+.30	16	.08	270
lazy-H	120'	4.4'	96	1096	+.28	17	.02	280
"	100'	10.4'	94	384	+.12	19	.07	280
"	80'	17.4'	81.3	180	-.06	20	.08	260
"	69.8'	21.6'	71.2	127	-.07	21	.09	240
"	60'	26.3'	59.7	90.9	-.15	22	.10	200
"	40'	38.3'	33.7	40.8	-.38	24	.16	140
"	30'	45.6'	21.5	23.8	-.59	25	.23	100
λ/4 2 radials	69.8'	——	——	38.8	.11/-.39	22	.15	200
λ/4, 4 radials	69.8'	——	——	35.7	+.21	22	.13	175

windows may be accommodated in a single antenna by simply switching in a capacitor in series with the input for 3.790-kHz operation.

Top-loaded verticals with elevated radials can take many forms. I'll explore a particularly useful form that looks like an H turned on its side. I call it the lazy-H vertical, for its resemblance to the classic lazy-H antenna. This antenna is functionally the same as the Discpole that appeared in the summer 1996, *Communications Quarterly*.[4] The Discpole antenna was designed for 2 meters and uses solid disks at each end. At low HF frequencies, it's generally impractical to use solid disks. Instead of a disk, two or more wire radials are used at each end. The 160-meter example given later does use a solid rectangular "disk" on the bottom end. The disk is actually the metal roof of my house, which was pressed into service. In general, at low frequencies, wire radials will be used. Keep in mind that the Discpole antenna may also be used with conical, as well as flat disks. As I'll show later in the context of sloping lower radials, the angle of the conical disk allows another degree of freedom in adjusting the driving point impedance. This is more useful in short antennas than long, however. The Discpole article has many useful things to say that are relevant to the lazy-H, and I recommend reading it in conjunction with this article. Moxon, G6ZN,[5,6] has also presented antennas that are closely related to the lazy-H. In fact, a lazy-H vertical appears on page 121 of his book. His articles make interesting reading. There's really nothing new in the idea behind the lazy-H antenna. A recent article in *QST* discussed the first trans-Atlantic QSOs made by hams in 1921. The antenna they used was essentially identical to the lazy-H, except that rather than using two elevated radials at the bottom, they used a fan of 30 elevated verticals.

The paragraphs below include the results of extensive modeling using NEC2 software and full-scale testing of three antennas—two for 80 meters and one for 160. For all the modeling, average ground ($\varepsilon = 13$, $\sigma = 0.005$ S/m) was assumed. The lower ends of the antennas are at 10 feet, and the antennas were modeled using #12 copper wire. A check was made on the effect of varying the height above ground from 3 to 15 feet. The effect was quite small and the information for 10 feet is representative. Wire losses are included in the gain comparisons. All of the modeling comparisons are made on 80 meters, but very similar results would be found for 160 meters when scaled appropriately for wavelength.

Most of the following discussion assumes the lazy-H version with two radials at the top and two at the bottom, all in the same plane. More radials, arranged symmetrically, may be used at both top and bottom and may improve performance. In particular, the SWR bandwidth will increase when more radials are used.

The half-wave vertical

A half-wave vertical dipole (**Figure 1A**) is a very effective DX antenna. However, it's too tall (137 feet on 80 meters, 260 feet on 160 meters) to be practical for most of us. You can shorten the antenna by adding perpendicular wires at the ends as shown in **Figures 1B** and **C**. The end wires provide capacitive loading. For a given height (L1), the length of the end wires (L2) may be adjusted to resonate the antenna. By adding the end wires, you can feed the

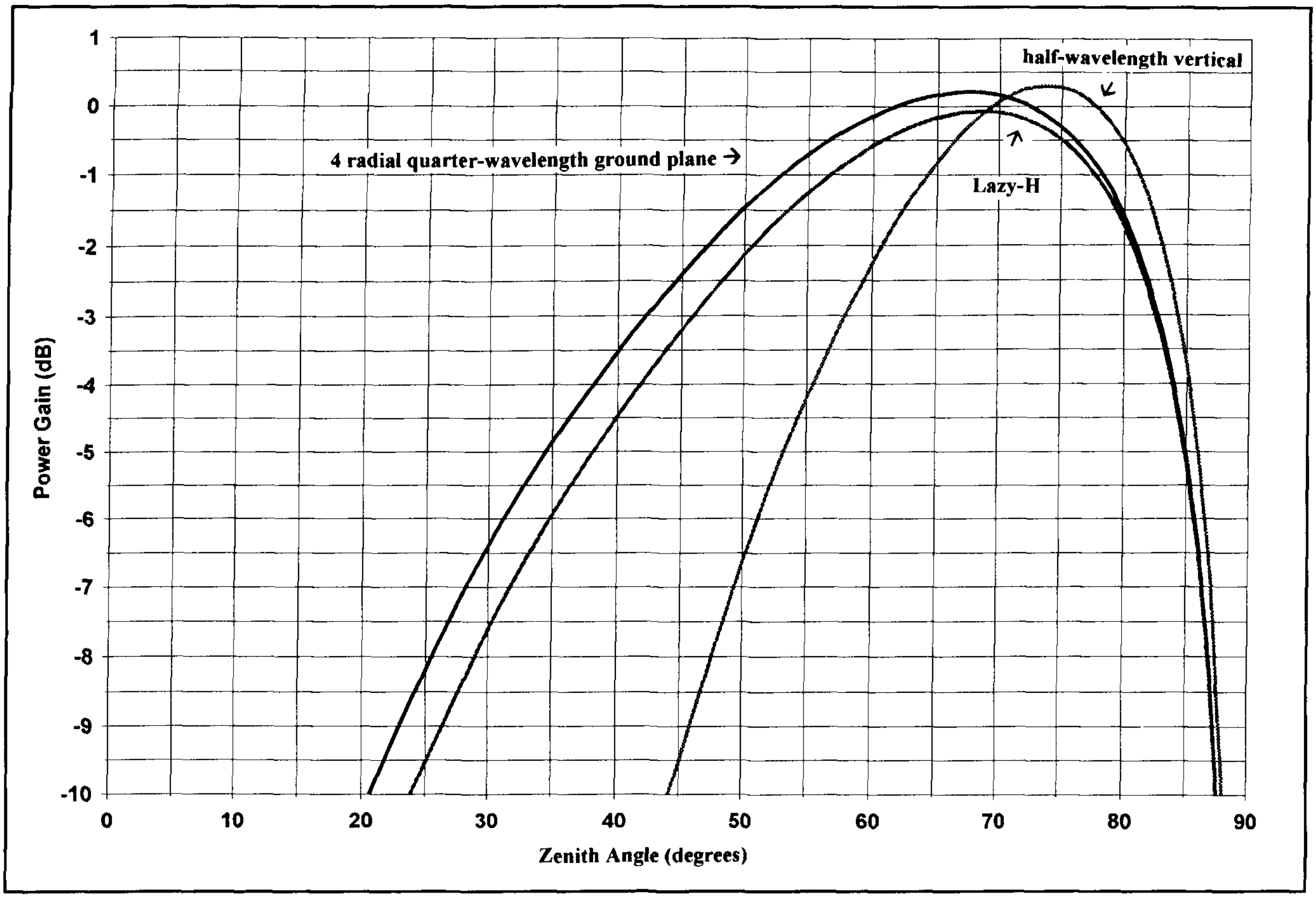

Figure 2. Elevation pattern comparison between 80-meter versions of a half-wave vertical dipole, a quarter-wave antenna with 4 elevated radials, and a lazy-H with L1 equal to a quarter wavelength.

antenna either at the center (B) or, more conveniently, at the lower end (C), which may be near ground level.

How good is this antenna compared to the half-wave vertical dipole or the quarter-wave antenna with a multi-wire elevated ground system? **Figure 2** provides an elevation pattern comparison between 80-meter versions of the half-wave and quarter-wave with 4 elevated radials (a ground plane antenna) and a lazy-H with L1 = $\lambda/4$ (69.8 feet).

The difference between the lazy-H and the quarter-wave ground plane is less than 0.3 dB. You won't notice that on the air. The gain difference between the half wave and the lazy-H is slightly larger, 0.37 dB, but there's an important difference in the peak gain angle. The peak angle is higher in the shorter antennas.

Table 1 provides a more detailed comparison of the lazy-H with values of L1 from 30 to 120 feet, the quarter-wave ground plane with 2 and 4 radials, and the half-wave antenna.

There's some interesting information presented in this table:

1. The peak gain difference between a full-length half wave and L1 reduced to 30 feet is less than 0.9 dB. This difference could be reduced to <0.7 dB if the vertical 30-foot section were made from larger wire or aluminum tubing to reduce the loss.

2. The peak radiation angle is increased from 16 to 25 degrees when L1 is reduced to 30 feet. This is due to the reduced length of the vertical radiator and there's no magic which will change that except to make L1 longer, or to raise the height of the entire antenna. The gain reduction at low angles for L1=30 feet compared with the half wave is shown in **Figure 3**. Even at the lowest angles the short lazy-H is within 2 dB, which is only a fraction of an S unit. The short antenna is still in the game! Because of the symmetrical end loading, the radiation resistance at the current maximum will be higher than other configurations for the same L1. The shortened antenna efficiency can be quite high if care is taken.

3. Compared to the quarter-wave ground plane, the 30-foot lazy-H is very close in peak radiation angle (25 degrees versus 22) and the peak gain is down by less than 0.8 dB, which could be reduced further. Even at 30 feet this antenna is competitive.

4. The gain, bandwidth, and efficiency of the lazy-H are very competitive with the half-wave

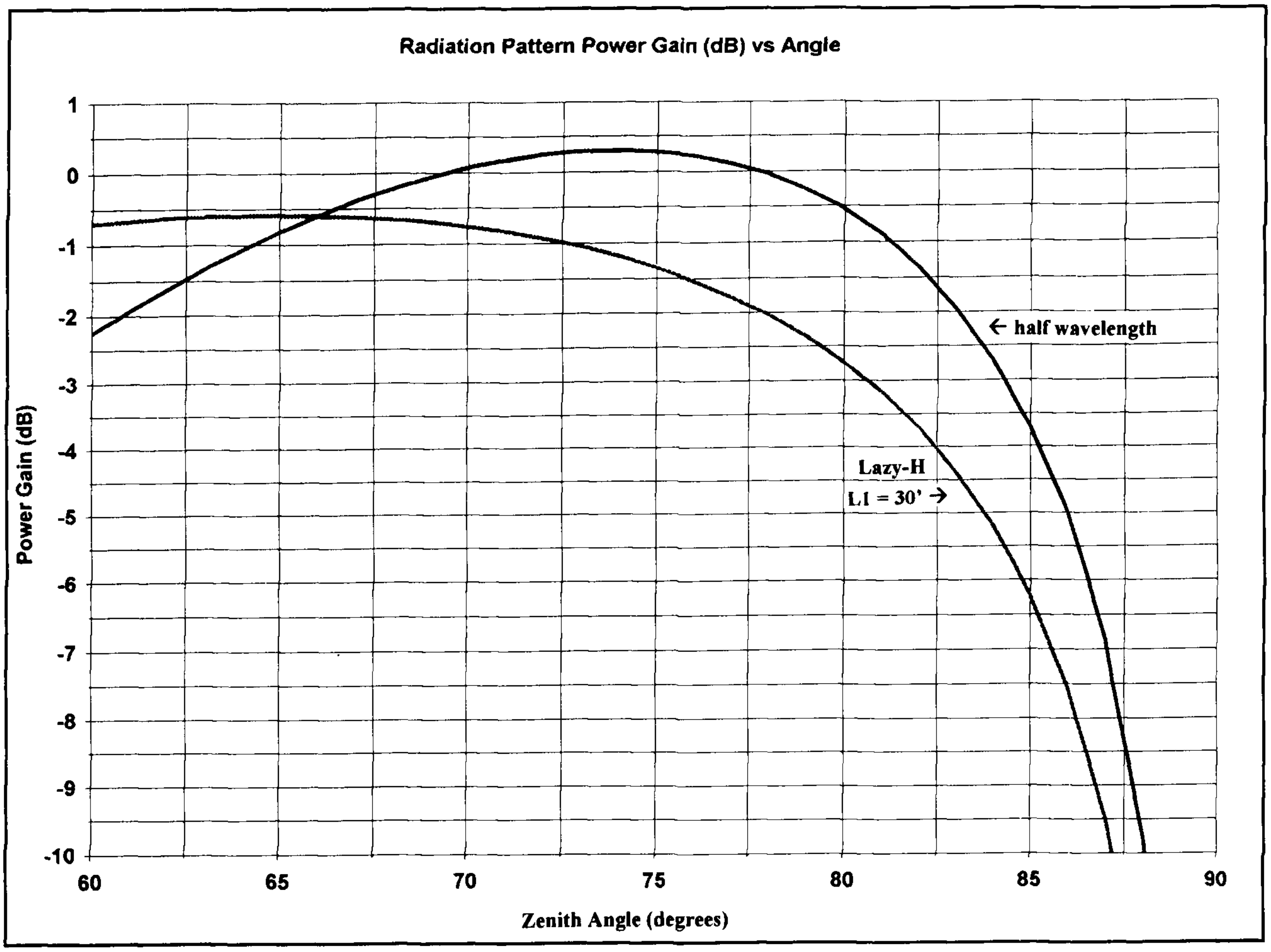

Figure 3. Gain reduction at low angles for L1 = 30 feet, compared with the half wave.

dipole for L1 > 50 feet (≈ 0.18 λ).

5. When compared to the quarter-wave ground plane with 4 radials, the quarter-wave lazy-H has a slightly lower peak gain (-0.28 dB), but also has a lower radiation angle (21 degrees). Performance-wise it's very close. However, the radials on the lazy-H are only 21.6 feet as opposed to 70 feet, and there are only two of them near ground. The lazy-H takes up much less real estate.

6. **Table 1** also lists a two-radial version of the quarter-wave antenna. Two values of peak gain are provided because the azimuth radiation pattern is slightly oval (about 0.5 dB). The maximum gain is broadside to the radials. As the height of the antenna is increased beyond a quarter wave, or if top loading is added to the quarter-wave antenna, the asymmetry in the pattern decreases very quickly.

Asymmetric lazy-H antennas

While the lowest loss is usually obtained when the upper and lower radials are equal and the current maximum is at the center of the vertical section, it's possible to have the lower radials longer than the upper or vice versa. The two-radial, quarter-wave antenna in **Table 1** could be viewed as an example of a lazy-H with zero-length upper radials. The difference in performance is quite small. One interesting feature of the two-radial ground plane is that the lower radials shorten very rapidly when even a small amount of top loading is used. One of the examples given later shows this clearly: where you have 3.5-foot radials at the top, reduce the length of the bottom radials from 70 to 54 feet. Adding top radials also reduces the asymmetry in the azimuth pattern, causing it to become very small in the symmetrical lazy-H.

With only small differences in performance, for a given length L1, the antenna can have a variety of proportions (see **Figure 4**). The length of the vertical section is also a variable. Usually the structure will be adjusted to be resonant inside the band, but even that is unnecessary. There are times when it may be advantageous to make the antenna resonant below the lower band edge to achieve a more convenient input impedance. The accompanying inductive reactance can be tuned out with a very low-loss series capacitor. These variations in shape and/or resonant frequency may be used to accommodate the requirements of a given site, or to manipulate the driving-point impedance or both.

When the antenna is suspended between two supports, the top radials won't be exactly parallel to the ground. They'll need to have some droop toward the center as shown in **Figure 5A**. This doesn't greatly affect the performance. The droop will reduce the length of the vertical section (L1), but this is offset to a degree by the vertical current component in the sloping radials.

The bottom radials may also droop as shown in **Figure 5B**; this can be exploited to vary the input impedance. It's well known that varying the angles for the 4 radials in a ground plane antenna provides a means for adjusting the feedpoint impedance.[7] The same thing happens in the lazy-H antenna.

If the antenna is suspended from a single support, the top radials may droop downward as shown in **Figure 5C**. A small amount of droop (<20 degrees) has very little effect, but a droop of 45 degrees or more will have the same effect as reducing L1. Where L1 is self-supporting (aluminum tubing or a tower for example), it's possible to use rigid radials for a portion of the top and then let the ends hang down as shown in **Figure 5E**. To make these variations work well, it's a very good idea to model them using EZNEC[8] or similar software.[9]

Feeding the antenna

There are many ways to feed this family of antennas, but there's one requirement you must keep in mind: these antennas are isolated from ground. This isolation must be maintained if the antennas are to work as advertised. For example, in the Discpole article, the antenna was fed at the junction of the vertical section and the lower disk. The antenna was isolated with a coaxial choke-balun, like that shown in **Figure 6**, with a shunt inductance of $\chi 1$ µH. At 146 MHz, that represents an impedance of 917 ohms, or roughly 20 times the feedpoint impedance. In my work with these antennas, I found that to be good rule of thumb. For the 80-meter asymmetrical antenna with a 50-ohm feedpoint impedance, 20x = 1000 ohms, which corresponds to 45.3 µH. That proved to be the minimum impedance necessary for isolating the feed. In the end, I used 100 µH and obtained good isolation. The balun in **Figure 6** may be scaled up to provide excellent isolation on 80 and 160 meters. For 100 µH and 1500 watts continuous, I use 30 turns of RG-214 wound on an 18-inch section of 8-inch diameter PVC pipe.

A less aggressive, but still perfectly serviceable, choke could be made using RG-8X wound on 4-inch PVC drainpipe. This inexpensive pipe is available from most building supply stores in 10-foot lengths. Some of the small Teflon™ insulated cables would be very good for this purpose.

The ground-plane antenna with four drooping radials is an old-time example of a floating antenna that benefits from isolation. A number of articles have mentioned the need to decouple the feedline and support structure from the antenna. The AEA isopole antenna is a good example. The antenna uses two conical skirts, the first represents the "radials" and the second is for decoupling.

I've used a 1:1 balun wound on toroidal ferrite cores a la Jerry Sevick.[10] These can work well, but you need 2- to 3-inch diameter cores with perhaps two or three cores stacked, to obtain sufficient inductance for low-band use. This is especially true if you're trying to isolate an antenna where Z_{end} is substantially greater than 50 ohms.

It's easy to tell if you don't have sufficient

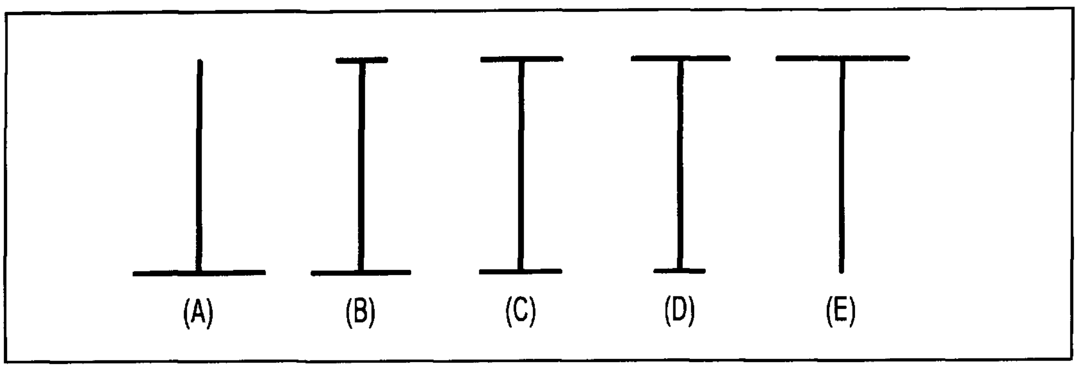

Figure 4. The aysmmetric lazy-H can have a variety of proportions with only a small difference in performance for a given length, L1.

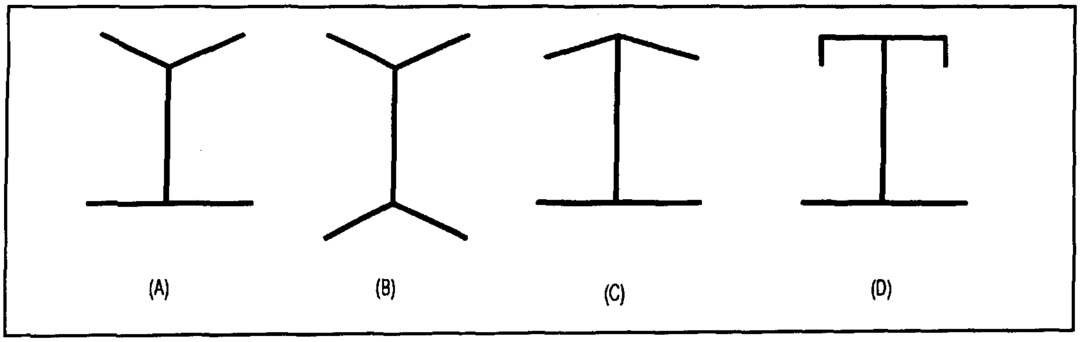

Figure 5. (A)When antenna is suspended between two supports, the top radials will have some droop towards the center. (B)The bottom radials may also droop. (C)If the antenna is suspended from a single support, the top radials may droop down. (D) It is possible to use rigid radials for a portion of the top and let the ends hang down.

isolation. When making measurements with an isolated instrument like the MFJ-249 or the AEA HF analyst, the SWR measurements will change as the instrument is touched. You'll see an even stronger reaction if the transmission line to the shack is touched to the instrument. Another strong indication of insufficient isolation occurs when the resonant frequency is quite different from expected. I noticed this effect in a symmetrical 80-meter lazy-H with a 200-ohm feedpoint impedance when feeding it with a 4:1 balun. The shunt impedance of the balun wasn't nearly high enough, and attaching the coax shifted the resonant frequency from 3.510 MHz down to 3.340 MHz. The resonant frequency was very sensitive to the position of the feedline.

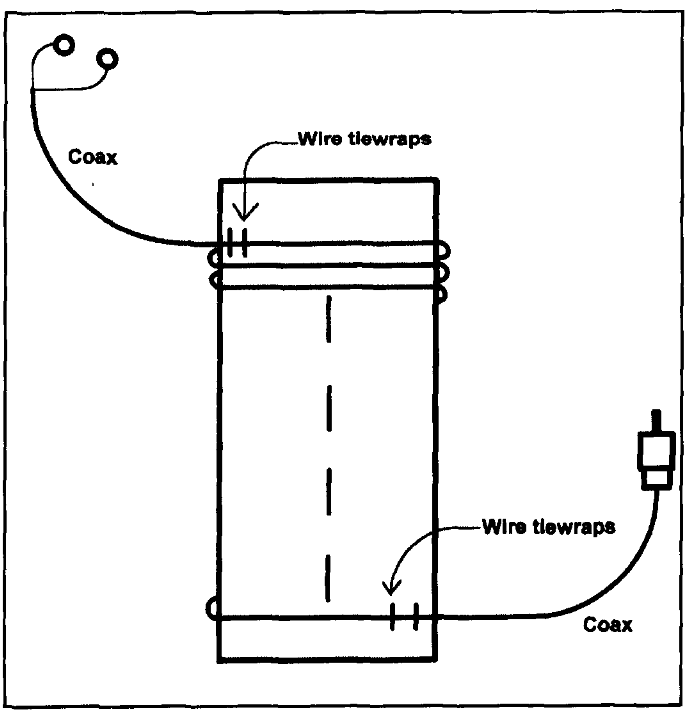

Figure 6. Coaxial choke balun.

These antennas can be fed at any point on the vertical section (L1) by the simple expedient shown in **Figure 7**. The coax shield is connected to the radials at the bottom of the antenna. The coax ends at the desired feedpoint and the rest of the vertical section is formed by a wire connected to the center conductor as shown. Of course, the end of the coax must be carefully sealed to keep moisture out of the cable. The minimum impedance is found at the current maximum. In a symmetrical lazy-H that is at the center of the vertical section. As shown in **Table 1**, the impedance at the center (Z_{middle}) and the bottom end (Z_{end})depend on the length of the center section. As you move away from the current maximum, the impedance rises. For L1 = 50 feet Z_{middle} is very close to 50 ohms. In fact, any length between 45 and 60 feet will give a good match to 50-ohm line. As L1 is shortened further, the feedpoint may be moved from the center towards the end—although for lengths as short as 30 feet, the difference between the center and the end is quite small because there's little difference in the current amplitude. For short antennas, where Z_{end} is low, you can use shunt feed (gamma, delta, omega matches). An example of this for a 160-meter antenna is given later.

The length of L1 that is made up by the coax cable can simply be an extension of the coax in the choke.

For lengths of L1 > 60 feet on 80 meters, there's no point (in a symmetrical lazy-H) on the antenna that's close to 50 ohms and, consequently, other schemes must be used. There are several possibilities:

1. For lengths longer than 65 feet, a point can be found between the center and the end where Z = 112 ohms. A quarter-wave length of 75-ohms coax will transform the 112 ohms to 50 ohms. At 3.510 MHz, using V = 0.66 coax, the length of the coax will be about 46 feet. Only a portion of this length will be needed to form the lower part of L1. The rest can simply be incorporated into the choke-balun.

2. For L1 of 80 feet or more, a 200-ohm point can be found and fed with a 4:1 balun. Be careful, however, most 4:1 commercial baluns don't have sufficient isolation for 80- or 160-meter operation. A coaxial choke will still be needed to provide the isolation on the 50-ohm side of the 4:1 balun. The shunt impedance of the matching balun will provide some isolation, and can reduce the size of the choke.

3. For a given L1, the radiation resistance will increase as longer radials are added at top and bottom. This technique may be used to increase the feedpoint impedance, but will of course introduce a series inductive reactance as the antenna resonance is lowered. This can be tuned out with a series capacitor.

4. The feedpoint impedance can be manipulated by use of asymmetrical radials top and bottom. An example of this is given later.

5. It's possible to adjust the length of the radials to make the feedpoint impedance complex then use a transmission line section to transform this to 50 ohms. A more detailed explanation of this idea will be presented in a later article.

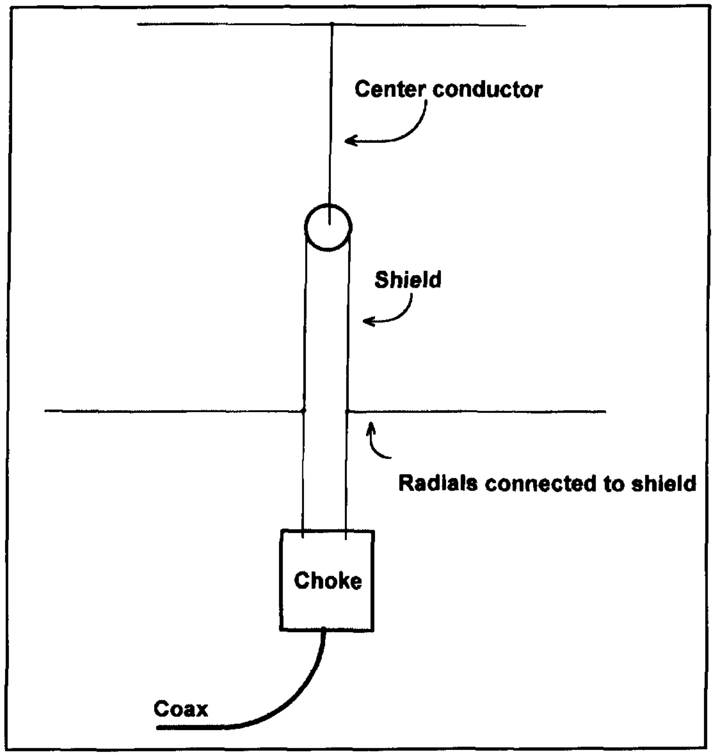

Figure 7. These antennas can be fed at any point on the vertical section (L1) using the method shown here.

A 160-meter lazy-H

Figure 8 shows one of several early versions of my lazy-H. The vertical section is 54 feet high, and there are two 55 foot radials at the top. My house has a metal roof, so I connected all of the panels together with copper strapping (soldered and screwed to the metal) to form large ground plane (≈ 35 x 60 feet). I used this as my lower "disk." Even with such a large area, it was still necessary to use an isolation choke. The gutter system is plastic, so the outer edges of the roof are isolated from ground.

For this short vertical (0.11 wavelength), the highest impedance point is only ≈ 18 ohms. As shown in **Figure 8A**, I used a shunt-feed variation to match the antenna. Textbook pictures of the shunt feed show a wire attached part way up a tower and sloping back to near ground. The inductive reactance introduced by the loop formed by the shunt wire and the tower is tuned out with a series capacitor. When you try this with a wire vertical the "tower" bends, and the bottom of the antenna looks more like a triangle. In fact, I did some modeling to determine the shape and dimensions for the match. I found out that there are any number of proportions which will provide a match. The equilateral triangle offers the best match bandwidth, although deviations aren't greatly different.

The final experimental dimensions for the match are provided in **Figure 8B**. The geometry isn't quite equilateral, but the bandwidth is good. Adjustment is straightforward. I began by fixing the distance along the base of the triangle (the attachment points on the roof), then moved the tap point along the vertical wire with an alligator clip on the shunt wire. I adjusted the length of the shunt wire to keep it approximately equal to the length of wire from the tap point down. The series capacitor is adjusted for minimum SWR at each tap point. It only took a short time to find a good match. When adjusted for minimum SWR at 1.840 MHz, at 1.8 MHz SWR = 1.5 and SWR = 2 at 1.950 MHz. I noted one interesting thing while I was adjusting the match. If I didn't try to get the SWR down to 1.0 at 1.840 MHz, but instead tried to extend the SWR < 2 bandwidth, I could obtain a double humped SWR curve with the maximum SWR < 2 over the entire band. The minimum SWR points were about 1.4, and the hump near mid-band about 1.7. Also, adjustment of the resonant frequency of the antenna (by changing the length of the top radials) could be used to improve the bandwidth.

When the radiation resistance gets this low, wire antennas start to get lossy. To reduce losses, I made the vertical section and a portion of the top radials from some 0.5-inch copper strap I had on hand. I could have done even better if I had made L1 from aluminum pipe, with guys, but the difference wouldn't have been worth the trouble.

I put this antenna up just before the 1996 ARRL 160-meter CW DX Contest. In a few

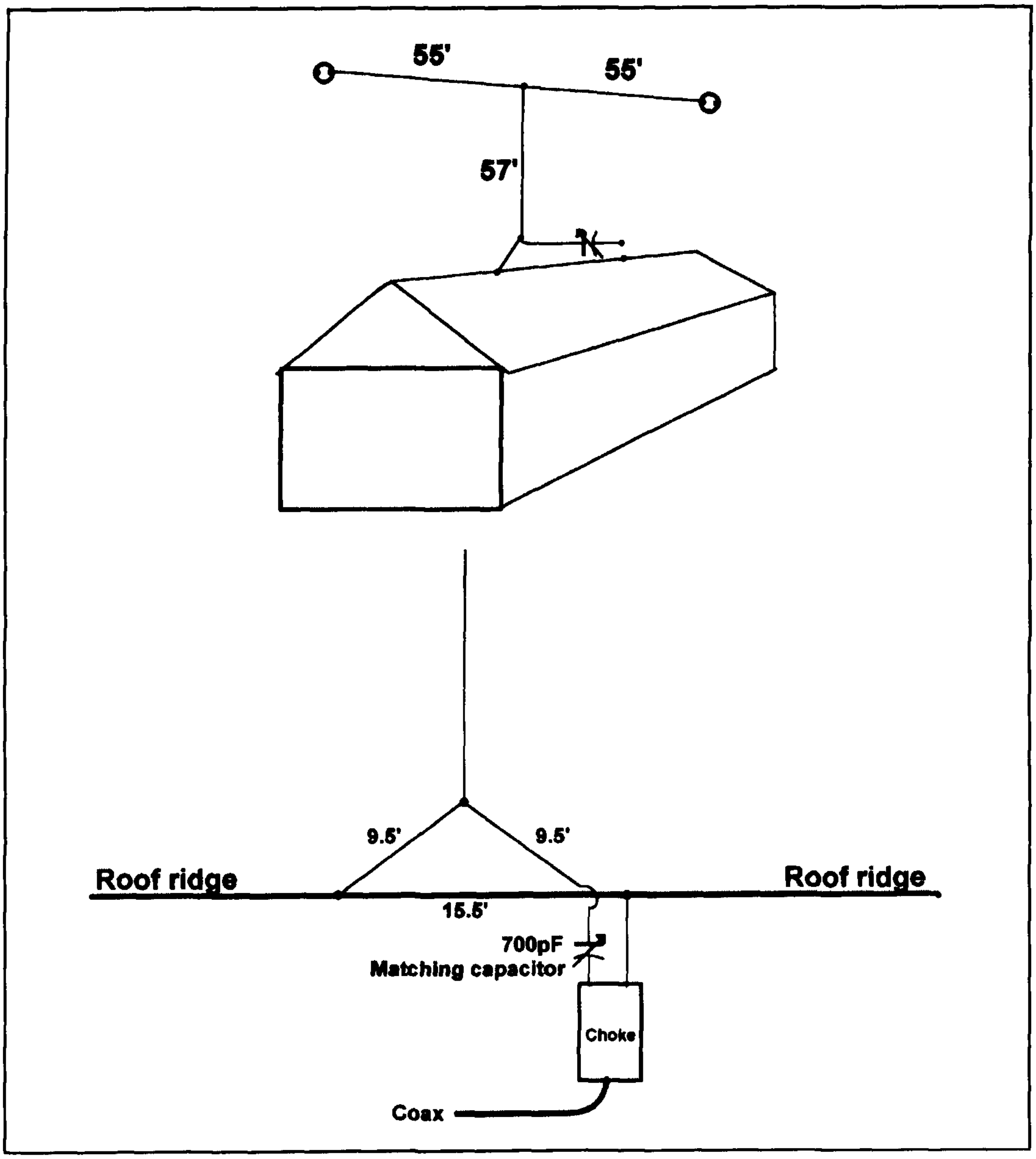

Figure 8. (A) Antenna is matched with a variation of the shunt feed. (B) Final experimental dimensions.

hours of casual operating, I was able to work 45 states—including KL7 and KH6, several VE provinces, XE, and some other DX. (I would have liked to work all 50 states at one sitting, but much of the northeast was QRT due to power losses caused by ice and snow.) I accomplished all this while running only 800 watts. I've been very pleased with this antenna; it's clearly effective. I'm now scheming how to get L1 up to 100 feet, or so!

An 80-meter symmetrical lazy-H

I built a symmetrical lazy-H for 80 meters like the one shown in **Figure 9**. The radials were about 22 feet and the feedpoint impedance, when resonant at 3.510 MHz, was about 130 ohms. This wasn't a very convenient value, but I noticed during modeling that the impedance increased above resonance. By making the radials longer (27 feet), I could move the resonant point down and increase the feedpoint impedance. **Figure 10** shows the feedpoint resistance and reactance from 3.5 to 3.8 MHz, with a 130 pF series capacitor to reresonate the antenna within the band. The resistive component varies from 155 to 250 ohms.

This provides a reasonable match to 200 ohms. However, a single capacitor doesn't provide sufficient 2:1 bandwidth to allow operation at both 3.510 and 3.790 MHz, so I resonated the antenna at 3.550 MHz with a 140 pF capacitor as indicated in the figure. I could have used two capacitors and a relay or switch to change between the CW and SSB DX windows, but I moved on to the next antenna instead.

The antenna worked very well with SWR = 1.2 at 3.550 MHz and a 2:1 SWR band width of 220 kHz.

An 80-meter asymmetrical lazy-H

The final version, which I'm using now, appears in **Figure 11**. The gain of this version is slightly lower than the symmetrical design, but the feedpoint impedance is a convenient 50 ohms when resonant at 3.510 MHz. The SWR = 1.1 at 3.510 MHz, rising to 2 at 3.625 MHz.

To accommodate 3.790 MHz operation, I inserted a 400 pF capacitor in series at the feedpoint, which is shorted out for 3.510 MHz operation as shown in **Figure 12**. Most of my operation is at the CW end of the band, so I chose to use the normally closed (NC) relay contacts. That way, if the relay failed to activate, I would only lose the SSB window.

I also chose to use a separate pair of wires for the relay power, but I could have used the coax itself with an isolation choke; however, I wasn't feeling very clever that day. The relay is one of the old-fashioned types designed to switch

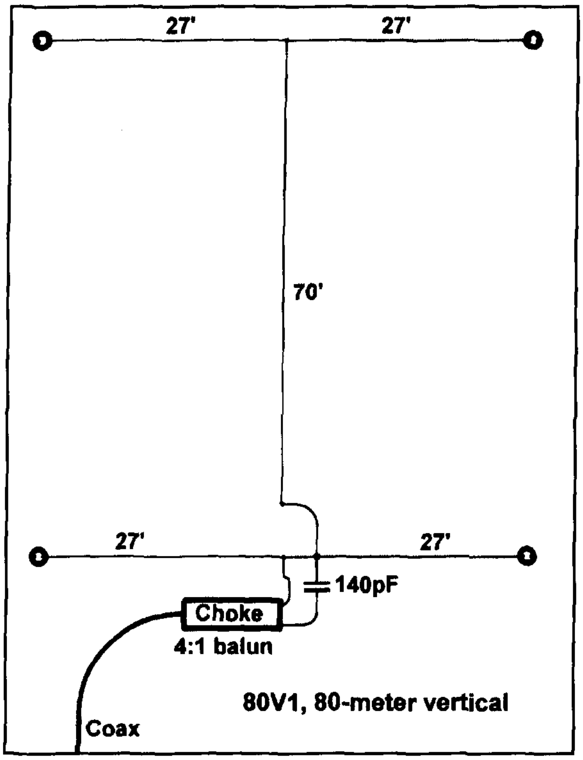

Figure 9. Eighty-meter symmetrical lazy-H.

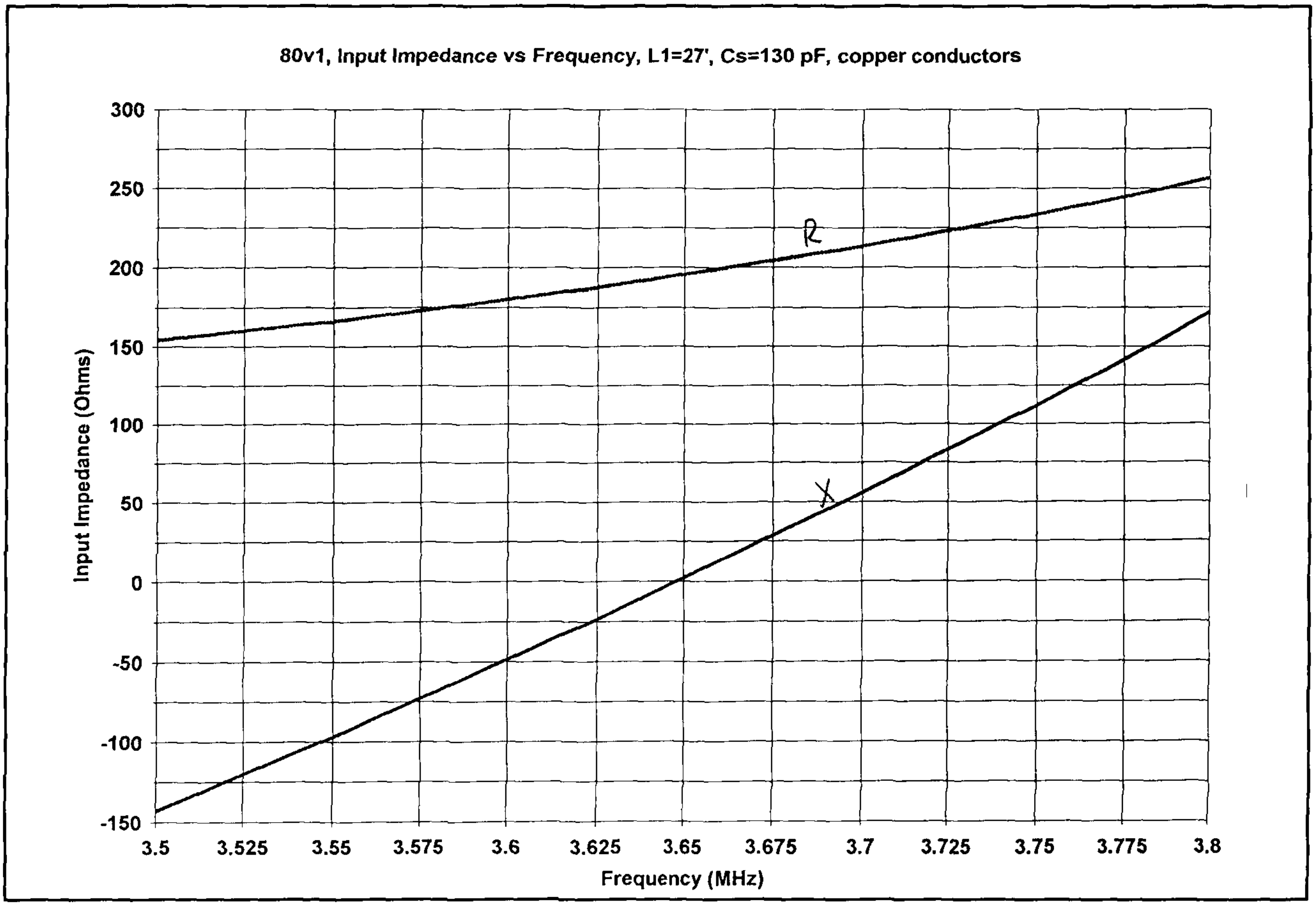

Figure 10. Feedpoint resistance and reactance from 3.5 to 3.8 MHz.

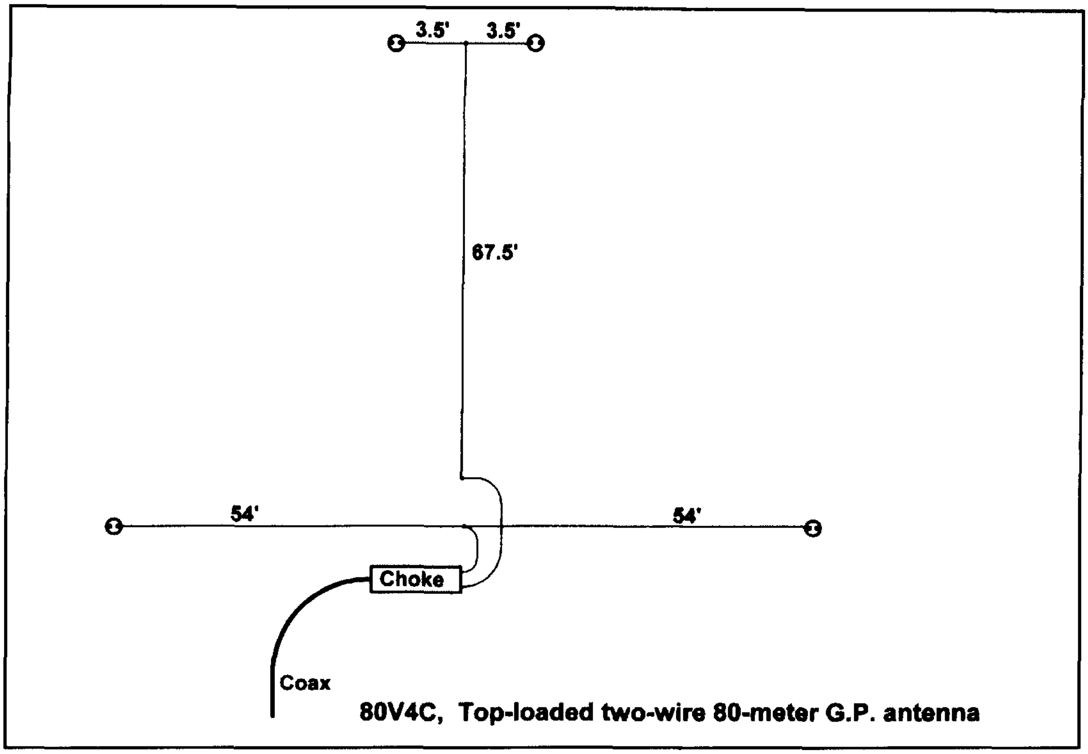

Figure 11. The final version of the 80-meter asymmetrical lazy-H.

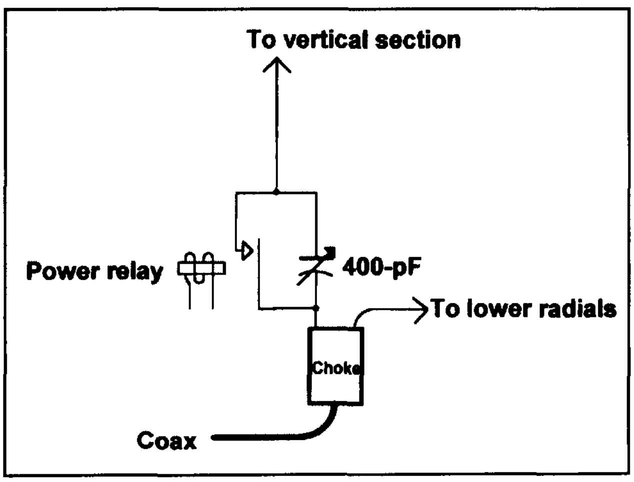

Figure 12. A 400-pF capacitor inserted in series at the feedpoint shorts out the circuit for 3.510-MHz operation, allowing operation on 3.790 MHz.

open-wire transmission lines, and has very little capacitance between the contacts and the coil. It also has very good voltage isolation, but even so, I had to be very careful with lead dress to prevent RF coupling RF back into the relay power lines. If you don't want to monkey with the complexity of a relay, you can use a simple alligator clip and a piece of wire as shown in the figure. If you opt for this method, you'll have to run outside to change frequencies.

When I first fired up this antenna, I was immediately able to work South American stations despite the high noise levels. I'm looking forward to using it under better conditions; I expect it will be very effective.

Conclusion

This family of antennas offers performance comparable to a quarter-wave vertical with an extensive ground system, but is much simpler and less expensive to build. The antennas may be varied greatly in dimensions and materials to accommodate a wide variety of situations and requirements. They are effective DX antennas and worth your consideration. ■

REFERENCES

1. Al Christman, "Elevated Vertical Antenna Systems, *QST*, August 1988, pages 35-42.
2. Al Christman, "Elevated Vertical Antennas for the Low Bands: Varying the Height and Number Of Radials," ARRL Antenna Compendium, Volume 5, 1996, pages 11-18.
3. Smith and Johnson, "Performance of Short Antennas," Proceeding of the IRE, October 1947, pages 1026-1038.
4. R. Littlefield, "Tech Notes: The 2-Meter Discpole Antenna," *Communications Quarterly*, Summer 1996, pages 77-81.
5. L. Moxon, *HF Antennas for All Locations*, 2nd Edition, RSGB, 1993.
6. L. Moxon, "Ground Planes, Radial Systems and Asymmetric Dipoles, *ARRL Antenna Compendium*, Volume 3, 1992, pages 19-27.
7. *ARRL Antenna Handbook*, 10th Edition (and others), pages 210.
8. EZNEC is available from Roy Lewallen, W7EL, P.O. Box 6658, Beaverton, Oregon, 97007.
9. NEC-Win Basic and NEC-Win Professional are available from Paragon Technology, 200 Innovation Blvd., Suite 240, State College, Pennsylvania, 16803.
10. J. Sevick, Transmission Line Transformers, ARRL, 1990, 2nd Edition.

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VALIDATION OF A F-LAYER ALGORITHM FOR THE IONOSPHERE

A summary of various approaches in analytical terms

Propagation prediction as it relates to HF radio communication has been a matter of continuing interest for both professional and amateur radio operators. With the development of vertical-incidence ionospheric sounding and the recognition of solar control of propagation, a number of methods have come forth—from simple to complex and from manual calculations to computer-aided ones—to make propagation predictions. This article will summarize briefly the various approaches in analytical terms and use one F-region algorithm, MAXIMUF, to illustrate how the methods can be improved further.

Basic to all discussions of propagation prediction is the concept of a database that describes the condition of the ionosphere across its various regions. Shortly after World War II, the knowledge gained during the war years appeared, first, as graphical representations of the worldwide distribution of critical frequency information of the E- and F-regions, and also as information on field strength and atmospheric noise (U.S. Department of Commerce, N.B.S. Circular 462, June 25, 1948).

In the years that followed, handbooks on ionospheric predictions and monthly circulars were published, enabling radio operators to calculate the maximum usable frequencies and optimum working frequencies for skywave transmission over any path for any time of day in a month. The circulars contained global contour charts for critical frequencies, foE and foF2, as well as overlays for the charts, nomograms, and detailed instructions needed for calculating MUFs from the data provided. Amateur radio operators took advantage of those opportunities and made their own predictions of propagation conditions for more than two decades.

The method in the early calculations involved the use of control points, the peaks of the first and last hops on a path. The idea was that if skywave propagation failed on a path, it most likely occurred at one end of the path or the other. As a result, foF2 critical frequencies for vertical incidence at the two ends of a path were increased by an M-factor in the range of 3-4, depending on the length of the hop, to make them applicable for oblique propagation of signals within the lower ionosphere. The lowest of the two was deemed to be the limiting frequency and termed the maximum usable frequency, or MUF, for the path.

The graphical displays used in the '60s were the result of numerical mapping methods from computer programs developed at the Central Radio Propagation Laboratory (CRPL) in Boulder, Colorado. Those efforts were part of the development program that gave rise to IONCAP, still in use after first being distributed in 1978. Considering the sophisticated techniques

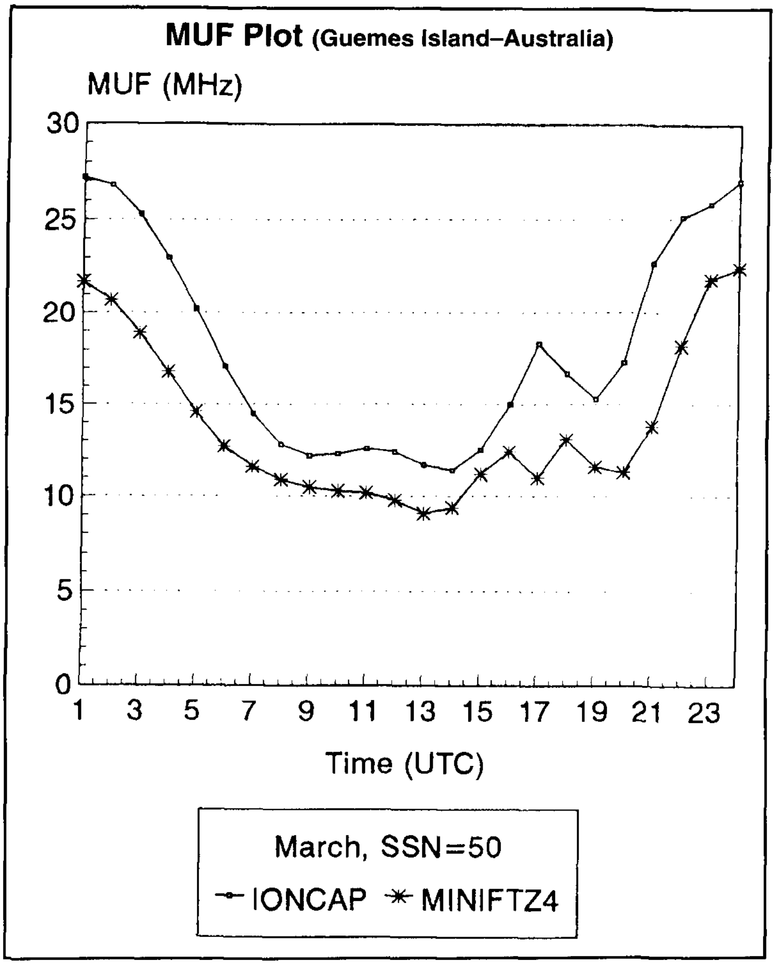

Figure 1. A comparison of MUF predictions by the IONCAP and MINIFTZ4 programs for a path from the Pacific Northwest to Australia (March, SSN=50).

and resources that went into IONCAP, it came to be considered the standard against which all other programs were compared.

PC predictions of critical frequencies

With the advent of the personal computer, efforts were made to bring propagation predictions more directly to radio operators, ending the manual calculations with maps provided by the National Bureau of Standards. The first computer program to reach the amateur radio community was MINIMUF 3.5; its source code was published in the December 1982 issue of *QST*. That program got its start in 1978, was developed by the Naval Oceans System Center (NOSC), and was based on oblique sounder data instead of a worldwide net of vertical sounders, as was the case with IONCAP.

MINIMUF used the control point method, but foF2 values were derived from a single mathematical function using the cosine of the solar zenith angle, adjusted to an "effective" value that took ionospheric relaxation processes into account. The model contained six parameters and their final values were adjusted to optimize its fit with the oblique sounder data. With those parameters, MINIMUF predictions[1] for the oblique sounder paths showed RMS errors ranging from 2.0 to 5.2 MHz, with an average error of 3.8 MHz. A comparison of MINIMUF predictions, and those of large-scale prediction programs for the same paths, suggested that the MINIMUF predictions were as likely to be correct as those from the mainframe computers.

While MINIMUF gained popularity as a rapid way to make MUF predictions, questions[2] arose as to the accuracy of its predictions when compared to measured MUFs. In that connection, Damboldt and Suessmann[3] then carried out a different appraisal of the predictions of MINIMUF, using data from the CCIR Atlas of Ionospheric Characteristics.[4] That atlas came from wide-scale vertical incidence sounding studies, and was considered by the ITU as the appropriate body of data for use in studies of HF propagation.

While making the analysis of MINIMUF predictions, Damboldt and Suessmann included the predictions from a new program, FTZ-MUF2, developed by the Deutsche Bundespost. That program was based on the CCIR Atlas and used a grid-point method, foF2 values obtained by an interpolation scheme instead of any mathematical representation of ionospheric properties, to obtain critical oblique propagation frequencies. In making the comparisons with the CCIR Atlas, MUF(3000) values were calculated at 19 latitudes (+90 to -90 in 10 degree steps), 12 longitudes (0 to 360 east longitude in 30 degree steps), for 24 hours of a day, 12 months of a year, and two sunspot numbers (0 and 100).

In total, 131,328 values were calculated for both MINIMUF and FTZMUF2. The results showed the average difference between the MINIMUF program and the Atlas was +1.0 MHz, while that of FTZMUF2 program was -0.09 MHz. In addition, standard deviations with the Atlas values were 4.4 MHz and 2.3 MHz, respectively, and histograms of the distribution of differences with the CCIR data showed the MINIMUF data with a spread which was twice as wide as that for FTZMUF2. On that basis, Damboldt and Suessmann felt they had demonstrated that MINIMUF lacked accuracy and their program, FTZMUF2, was superior for ionospheric predictions.

Other computer prediction programs appeared on the scene in 1985, with Radio Netherlands publishing source code for programs developed by Raymond Fricker,[5] a propagation engineer in the BBC External Services Division. Those programs included F-layer algorithms in the form of a series of mathematical functions, of time as well as latitude, longitude, and sunspot number. One program, MICROMUF 2+, used a total of 13 functions while another, MAXIMUF, used 26. The parameters of the algorithms were adjusted with the aid of foF2 values from the CCIR Atlas (Report 340), but with an interesting feature—the foF2 value at a geographic location was associated with its latitude relative to the magnetic dip equator at its longitude.

In adjusting the parameters of his algorithms, Fricker used CCIR data for the months of March, April, June, and December as well as four sunspot numbers from 0 to 50, 100 and 150. A seasonal parameter was introduced to interpolate predictions for the other months of the year. Only CCIR data for the Northern Hemisphere were used in constructing the models, and the values for locations in the Southern Hemisphere below the magnetic dip equator were obtained by changing the sign of their latitude relative to the dip equator and adding 6 months to the date of the calculation. When done, Fricker indicated that CCIR foF2 data and values derived from the calculations were largely in agreement, to within 0.75 MHz. The only exceptions noted were in the Southern oceans, where experimental observations were lacking in making the CCIR Atlas.

Path comparisons

With three different PC prediction programs on the scene in the mid-80s, it was natural that comparisons be made—particularly by amateur radio operators with their contacts frequently ranging worldwide. The IONCAP program continued to be considered the standard when it came to propagation, and comparisons were made with respect to its predictions. But the methods of comparison varied, using differences in MUF predictions for paths, both of common or special interest. The simplest types of comparisons used numerical differences and root-mean-square differences, while the more complex comparisons involved linear regression methods.

For example, consider the predictions for a path from here in the Northwest to Australia for the month of March and a sunspot number of 50, shown in **Figure 1**. The prediction programs used in calculating that figure were IONCAP and MINIFTZ4, a program using the FTZMUF2 F-layer algorithm. MINIFTZ4 points fall well below those from the IONCAP program but show the same trend, except for a brief period around 1700 UTC. Analysis shows the mean difference of MINIFTZ4 from IONCAP is -4.2 MHz and the RMS difference to be 4.7 MHz.

Another example, with MUF data for a path from the Northwest to London, is shown in **Figure 2** and uses data from the IONCAP and the MAXIMUF program mentioned earlier. The two curves are quite similar and analysis shows that MAXIMUF has a mean difference of -0.28 MHz from IONCAP and a RMS difference of 1.01 MHz.

A linear regression analysis of those two examples shows that the correlation coefficients were 0.90 and 0.94, respectively, and that the slopes of the regression lines were 0.72 and 0.79. While the two correlation coefficients show a good relation between each of the two data sets, the slopes of the regression lines indicate that changes in the MAXIMUF predictions for the path to London more closely matched in magnitude those for IONCAP than changes in MINIFTZ4 matched those of IONCAP for the path to Australia.

A larger study of the type discussed above was undertaken early in 1988 by the NOAA PBBS Sysop—making use of four paths, four

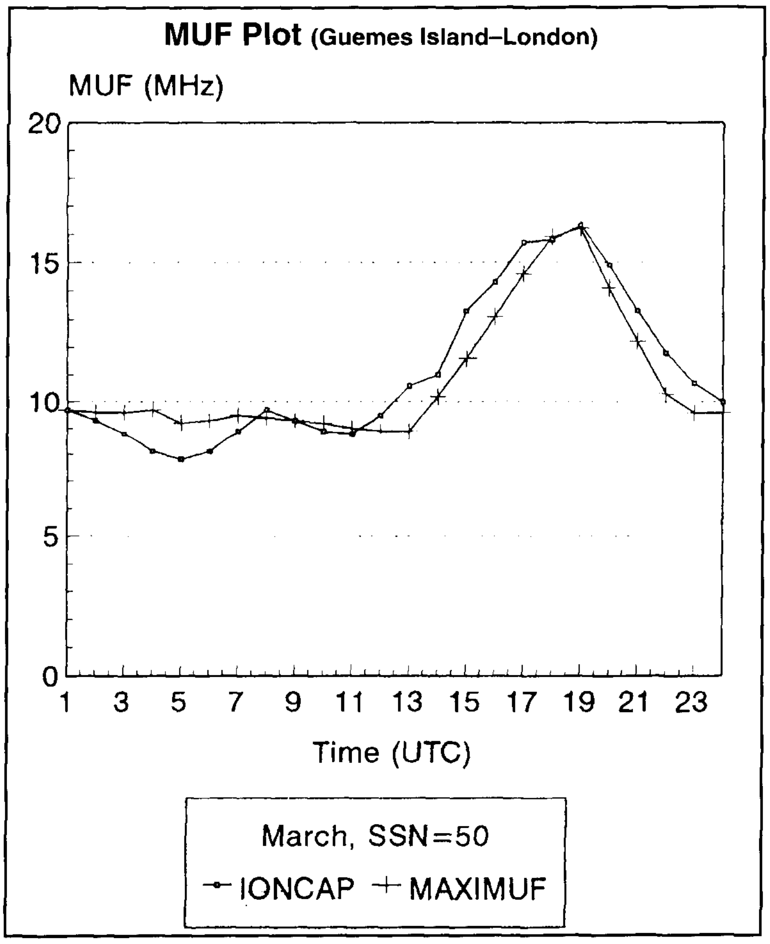

Figure 2. A comparison of MUF predictions by the IONCAP and MAXIMUF programs for a path from the Pacific Northwest to London, England (March, SSN=50).

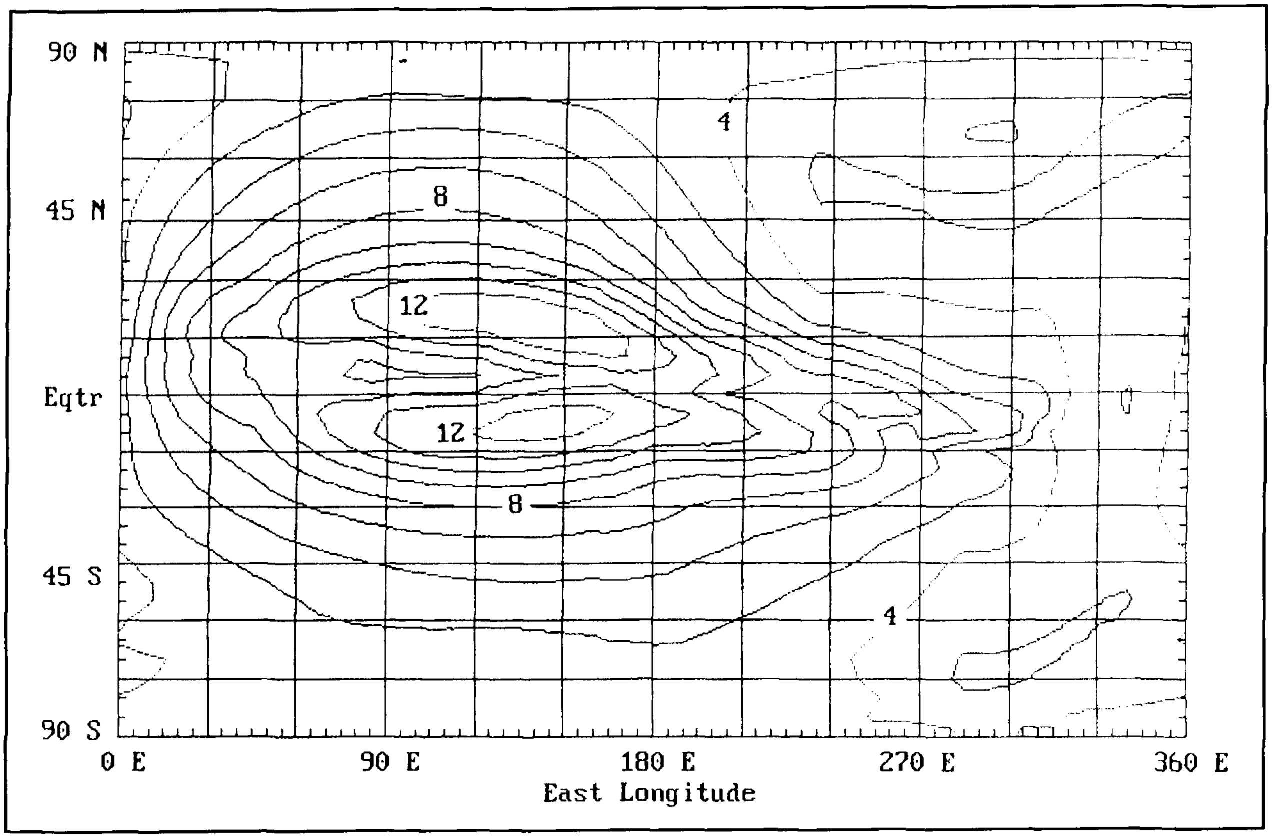

Figure 3. A global contour map of foF2 values (in 1-MHz steps) from the MAXIMUF algorithm for March, 0600 UTC, and a SSN of 50.

months, and two sunspot numbers. The paths were from Boulder, Colorado to Midway Island, Easter Island, London, and Mexico City, and the SSN were 60 and 120 while the months were January, April, July, and October. The programs used in the study included the three mentioned here and others based on Fricker's methods as well as that of NOSC.

For the paths indicated above, the NOAA PBBS study showed that the mean differences with IONCAP were -1.4 MHz, -2.6 MHz and +4.2 MHz for MAXIMUF, MINIFTZ4, and MINIMUF while RMS differences were 2.7 MHz, 3.7 MHz, and 5.6 MHz, respectively. The correlation coefficients were 0.84, 0.79 and 0.82, respectively, but the slopes of the regression lines were not given.

Comparisons using ionospheric maps

Earlier, ionospheric maps were produced centrally by the National Bureau of Standards and distributed on a monthly basis for use in propagation predictions. With the manual methods replaced by computer programs using algorithms to mimic the worldwide distribution of ionization, computer predictions flourished and comparisons were made between programs, as noted above.

Nowadays, personal computers have advanced to the point that ionospheric maps may be produced using the databases from international ionospheric models. Thus, the PROPLAB PRO program from Canada offers the capability of making ionospheric maps for any date or sunspot number using the CCIR or the URSI data files in the International Reference Ionosphere.[6] That program was developed by the Committee on Space Research (COSPAR) and the International Union for Radio Science (URSI), and IRI 90 contains monthly values for magnetically-quiet conditions at non-auroral latitudes from 50 to 2,000 km altitude.

With these advances, it has been possible to examine the degree to which foF2 ionospheric maps from various algorithms resemble those from reference ionospheres, making use of contour mapping between data points across the globe. For example, **Figure 3** shows the foF2 map from the MAXIMUF algorithm for 0600 UTC in March when the sunspot number is taken to be 50. For comparison, **Figure 4** shows the foF2 map from the CCIR reference ionosphere for the same conditions. Visual comparison, while qualitative, is rather favorable as the magnitudes and principal features of the CCIR maps are found on the map derived from the MAXIMUF algorithm.

Now the discussion can go one step further by using the ionospheric mapping technique to make quantitative comparisons, point by point, between the F-layer algorithms and reference ionospheres under various conditions. Indeed, it is capable of showing where significant departures occur between predictions of an algorithm and any reference ionosphere, thus indicating where improvement in the algorithm might be needed, even possible.

Of the three algorithms discussed above, MAXIMUF was chosen for the present discussion as it is better known and provided the best results when compared with IONCAP. In addition, it is of some interest as it has served as the foundation, after some modest modifications and improvements, for two propagation prediction programs presently in use—MINIPROP PLUS and IONSOUND. However, the discussion here uses the original MAXIMUF algorithm (J. Ruys, private communication) obtained from Radio Netherlands in 1985. This was done in order to illustrate the method rather than give any specific data for improvement of the algorithm. That being the case, observations made here should not be construed to reflect on the results of its use in any proprietary setting at the present time.

In this method, essentially a way of validating MAXIMUF, a comparison is made of its predictions at non-auroral latitudes with the CCIR reference ionosphere. This was done using foF2 data obtained from the PROPLAB PRO program, data points taken every 10 degrees in latitude from 80N to 80S and every 12 degrees of longitude from 0 to 360W. With that resolution, predictions of the MAXIMUF algorithm were compared at 527 points with foF2 values from the CCIR ionosphere database in the PROPLAB PRO program. The URSI ionosphere database was also available in that program, but the CCIR ionosphere was used as it seemed the more appropriate—given that MAXIMUF was based on the CCIR Atlas (Report 340).

The difference between MAXIMUF foF2 predictions and those from the CCIR database was determined, point by point, as was the square of the difference. Those were added up

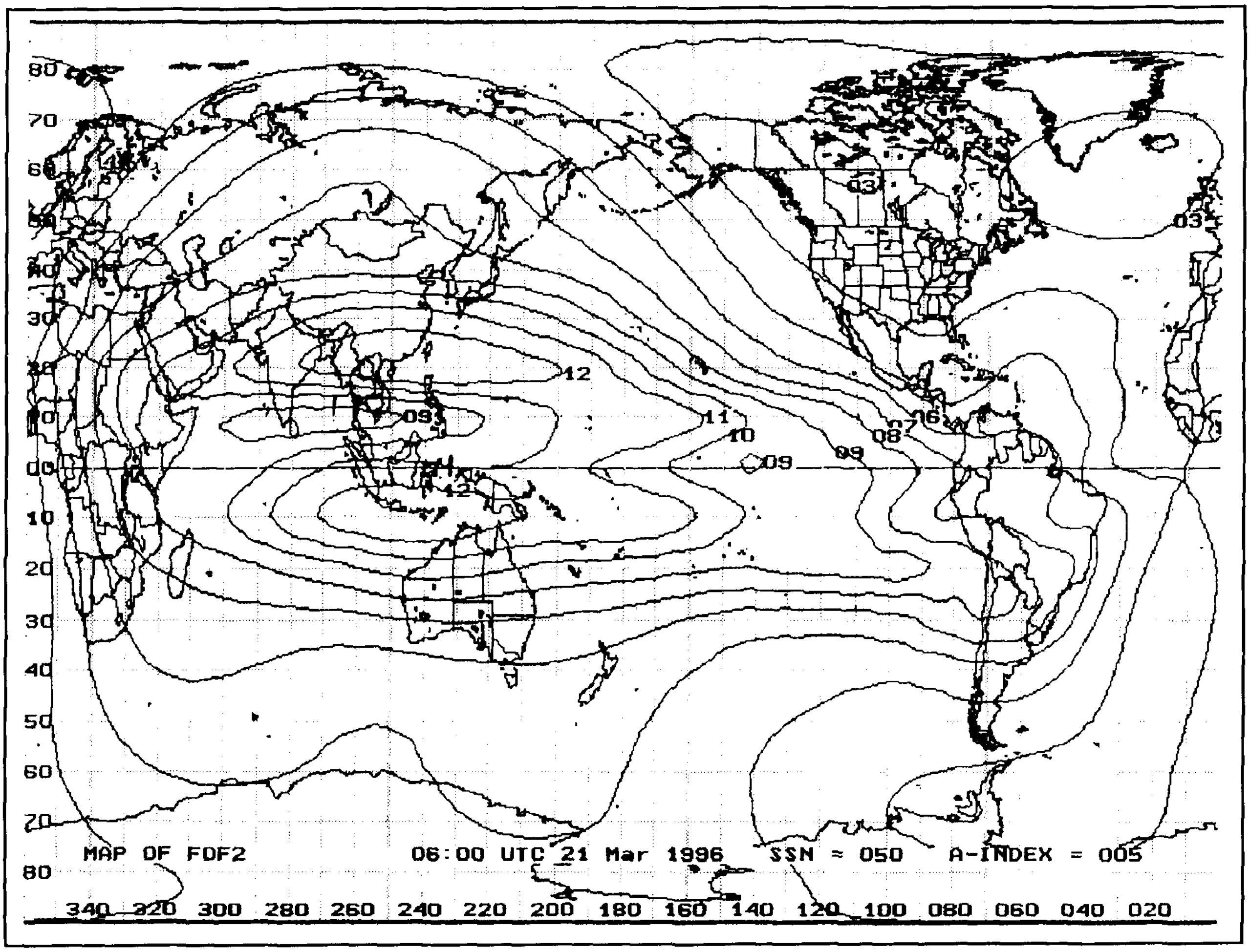

Figure 4. A global foF2 map from the CCIR database for March, 0600 UTC, and a SSN of 50.

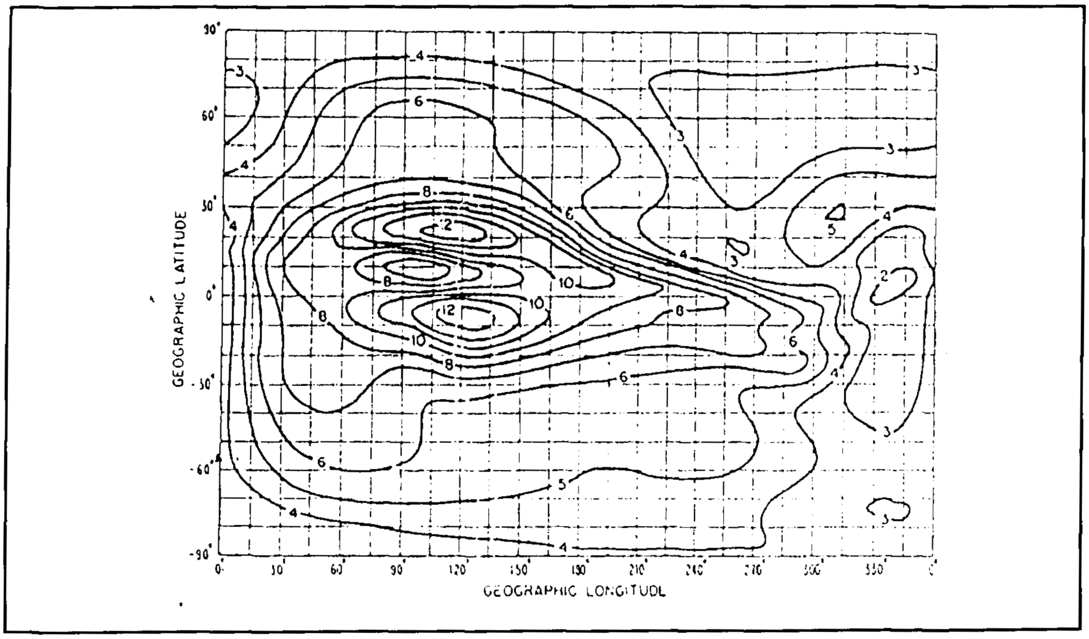

Figure 5. A global foF2 map for a solar minimum (March 1976, 0600 UTC, and SSN=12). From Rush and Davies.[7]

across the 527 data points and printed out as both as a mean difference and a root-mean-squared value for various conditions, say hour, month and sunspot number. In that regard, calculations were made for three sunspot numbers, 25, 50 and 100, and the twelve months of the year—but the main discussion will revolve around details of the results for 0600 UTC in March and for a SSN of 50. That choice of date and time is particularly valuable to the discussion of ionospheric matters as the sub-solar point is at 90 East longitude and, given the symmetry of the solar illumination, it illustrates the main features of the F-region. Two such examples are given in **Figures 5** and **6**, for solar minimum and maximum, respectively.

For that choice (0600 UTC, March, SSN=50), calculations showed that, on the

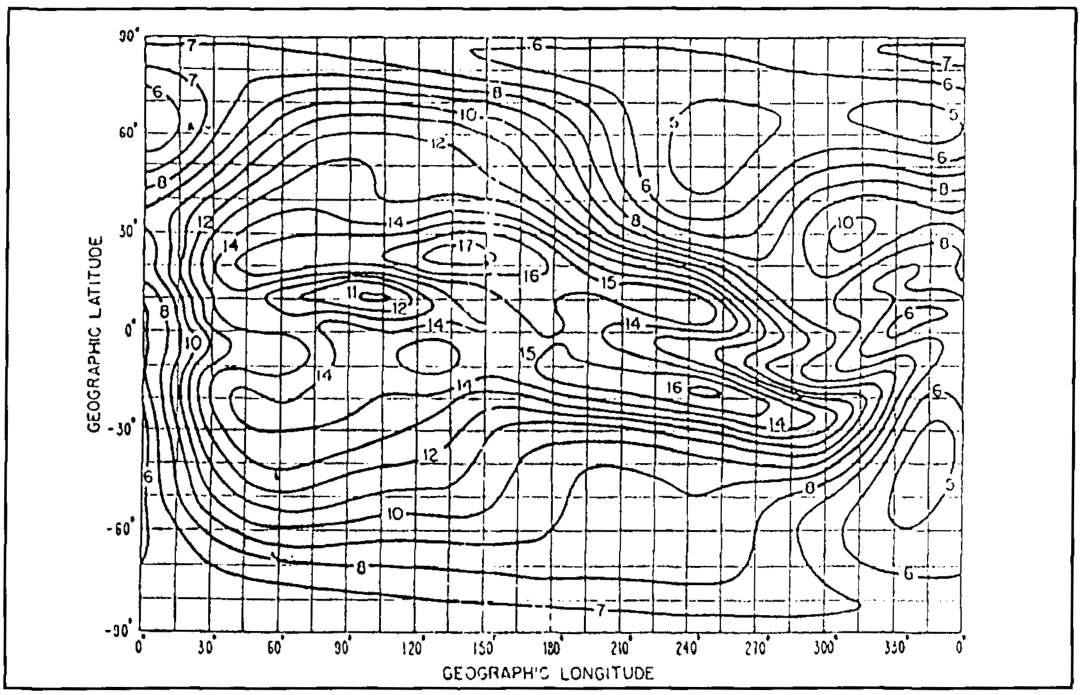

Figure 6. A global foF2 map for a solar maximum (March 1979, 0600 UTC, and SSN=137). From Rush and Davies.[7]

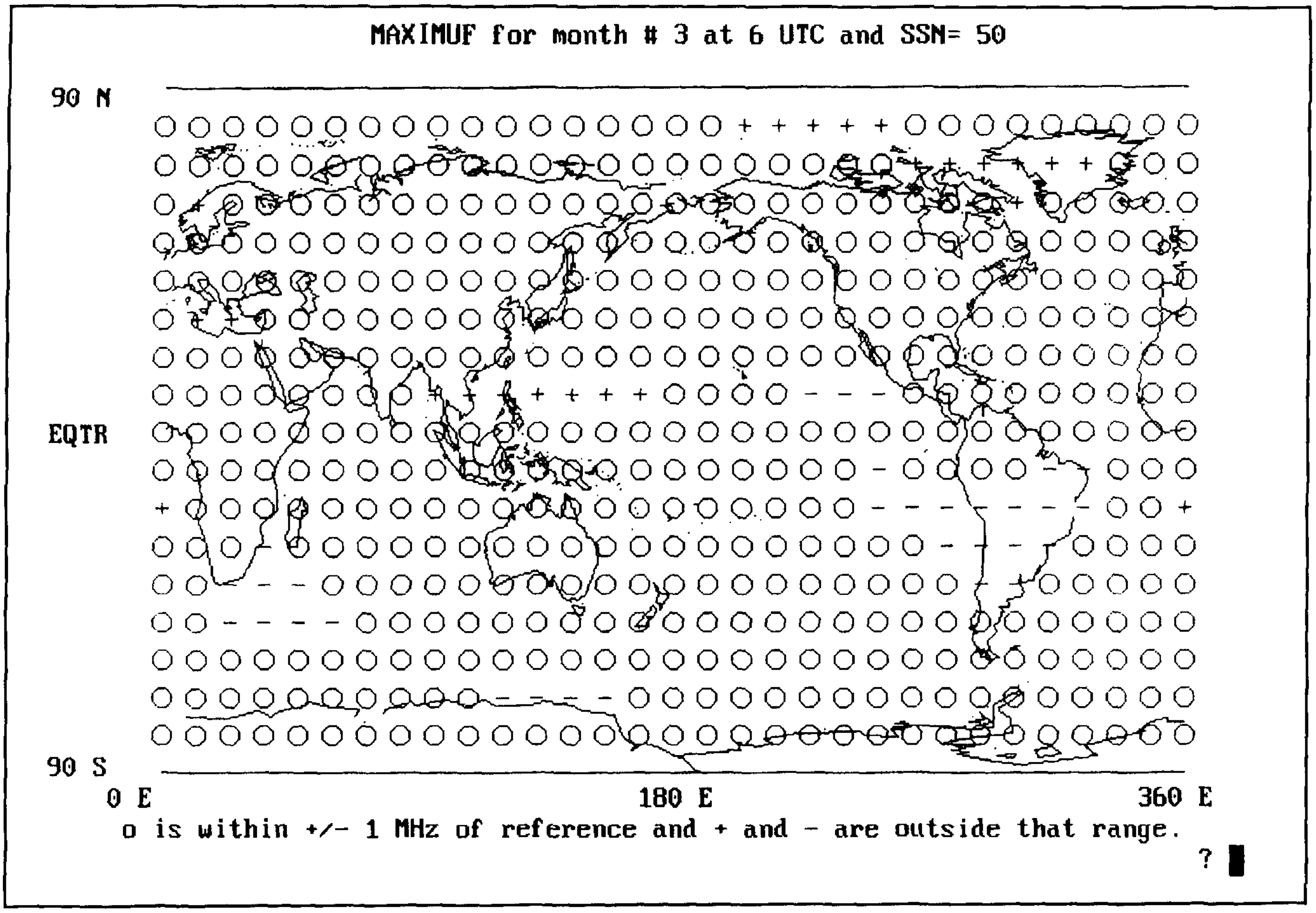

Figure 7. The global distribution of foF2 differences between the MAXIMUF algorithm and the CCIR database for March, 0600 UTC, SSN=50 with a 0.5 MHz offset added.

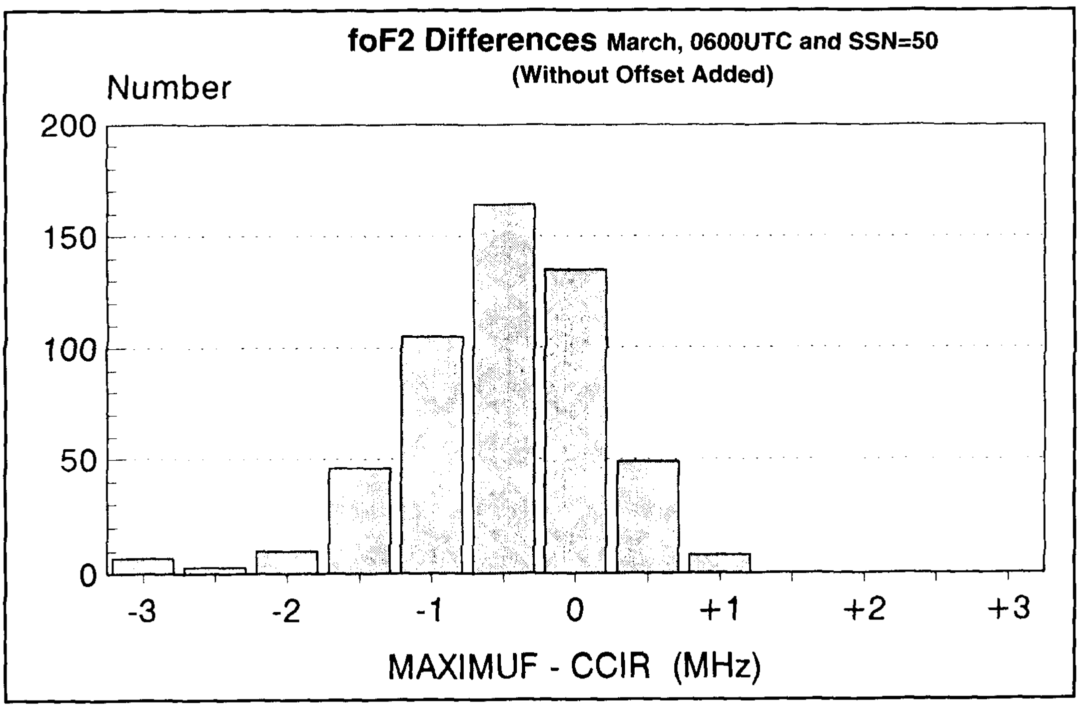

Figure 8. The numerical distribution of foF2 differences between the MAXIMUF algorithm and the CCIR database for March, 0600 UTC, and SSN=50.

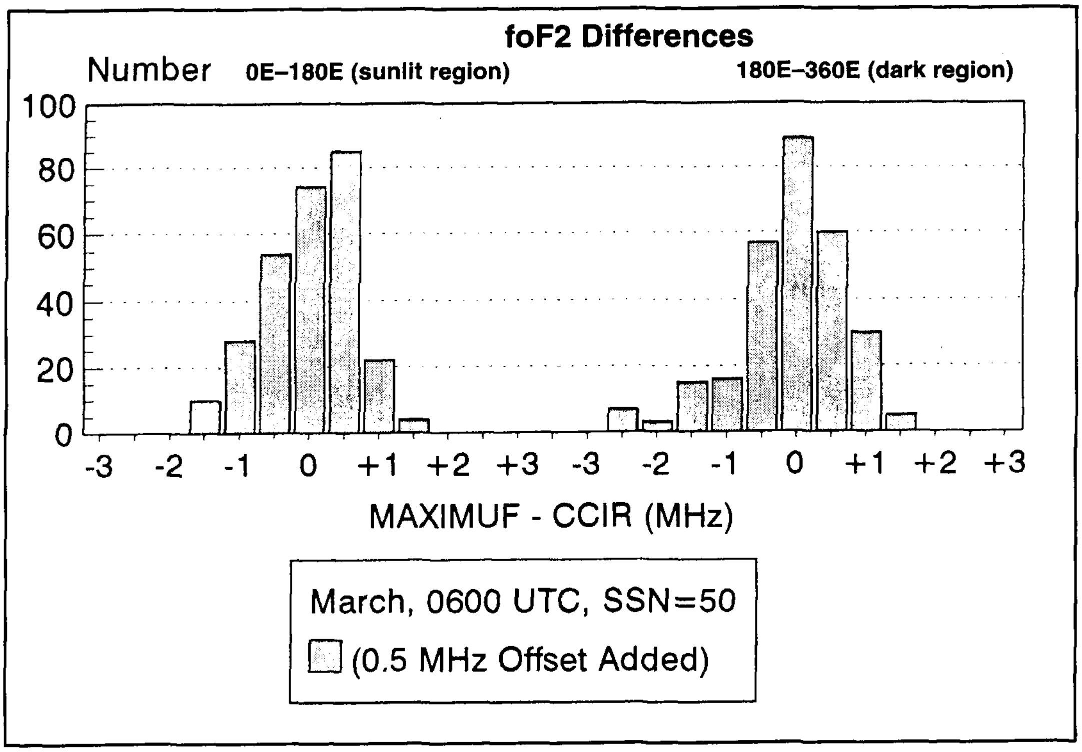

Figure 9. The numerical distribution of foF2 differences in the sunlit and dark regions of Figure 7.

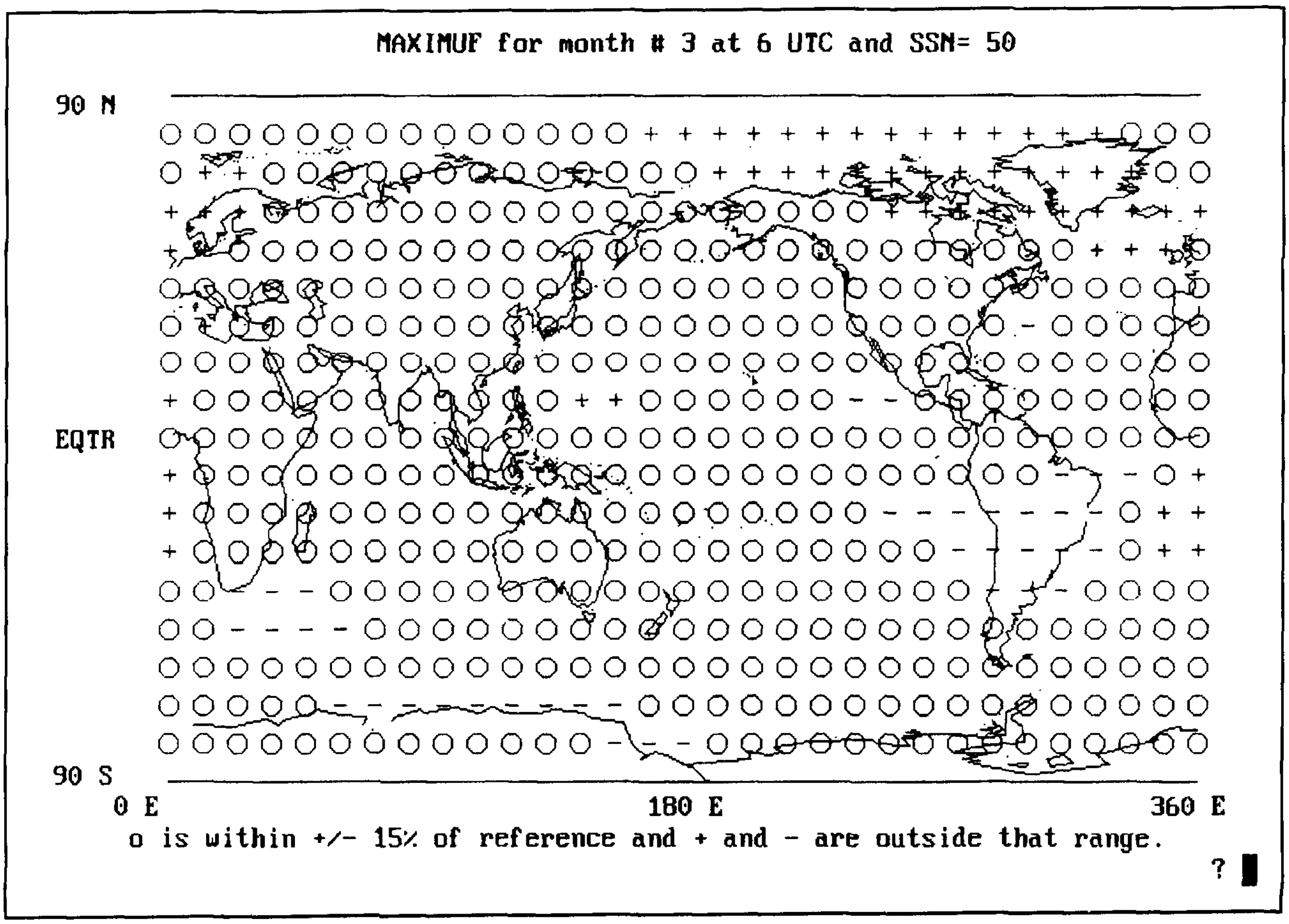

Figure 10. The global distribution of foF2 differences, for the MAXIMUF values within +/-15 percent of the CCIR database, for March, 0600 UTC, SSN=50 with a 0.5 MHz offset added.

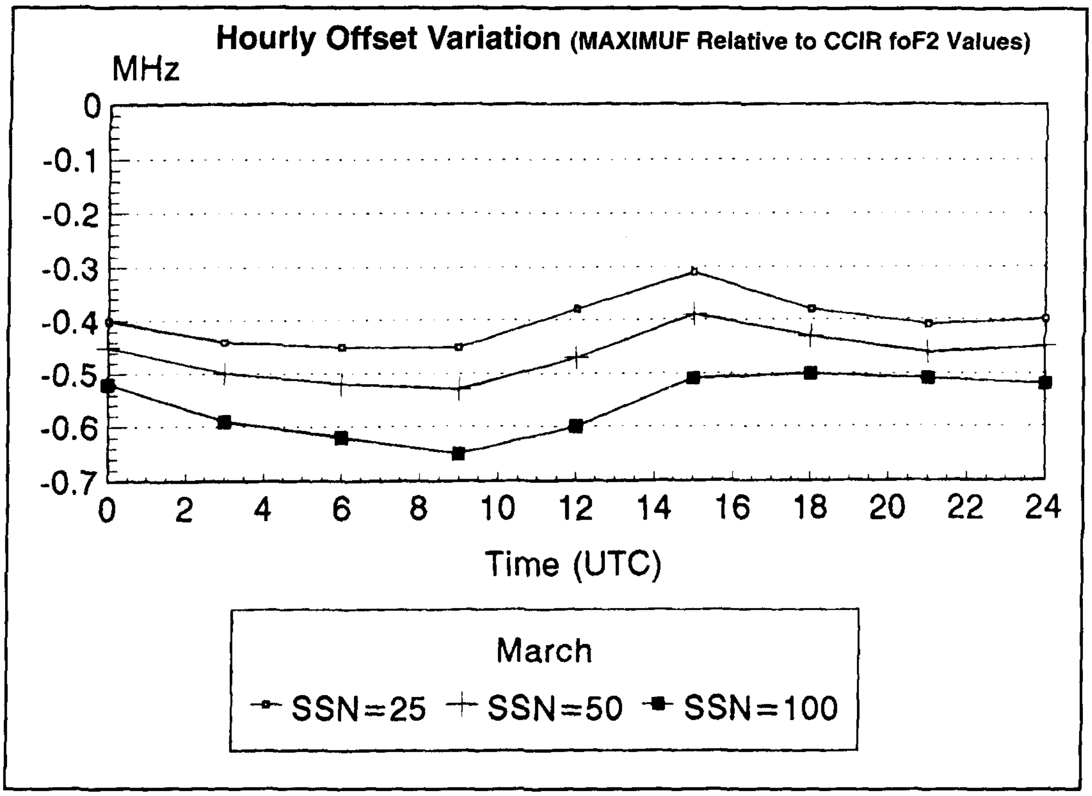

Figure 11. Hourly variation of the mean offset of foF2 values between the MAXIMUF algorithm and the CCIR database for three levels of solar activity in March.

average, MAXIMUF foF2 predictions were low by -0.52 MHz and had a RMS difference of 0.85 MHz with respect to the CCIR reference ionosphere. Of the 527 data points, the difference between MAXIMUF and CCIR data was positive at 116 points and negative at 411 points. When examined for values above or below the CCIR ionosphere, 2 data points were found to exceed the CCIR reference by more than +1 MHz and 106 data points had differences beyond -1 MHz from the reference.

Of the 106 negative data points cited above, the vast majority were in the Southern Hemisphere, in the Indian Ocean area to the east of Africa, and the southern Pacific Ocean area to the west of South America. In addition, there were some negative data points in the mid-Pacific Ocean around Hawaii. In a sense, those negative data points are related to Fricker's observations about the paucity of good data from ocean areas in the CCIR Atlas.

Given that the MAXIMUF predictions were offset from the CCIR values by -0.52 MHz, it was of interest to add that amount to the MAXIMUF predictions, point by point, and evaluate the change that resulted. The results are shown by the map in **Figure 7** where plus (+) and minus (-) symbols are used to designate points on the map where the MAXIMUF foF2 predictions were more than 1 MHz above or below the CCIR values, respectively. Otherwise, circles indicate where the differences were less than 1 MHz. That figure shows a marked change in the number of positive and negative points, now some 22 and 28, respectively, and amount to 4.2 and 5.3 percent of the total data points, respectively—leaving just over 90 percent of the foF2 values from MAXIMUF within +/- 1 MHz of the corresponding CCIR values. It should be noted, however, that negative points are still largely in the Southern Hemisphere, but the number is reduced drastically, from 80 points without the offset correction to 28 with the offset included.

In the above discussion, the choice of 1 MHz as a comparison figure was arbitrary—of more value in a practical sense than a statistical one. Thus, a more appropriate value would be the RMS difference after correction for the offset. For the case above, the RMS departure from the CCIR values was 0.67 MHz; thus, the choice of 1 MHz corresponds to 1.5 times the RMS value.

The same discussion, when applied to the cases of sunspot numbers of 25 and 100, provides RMS values of 0.62 and 0.79 MHz, respectively, after correction for offset. For those cases, 1.5 times the RMS values are 0.93 MHz and 1.2 MHz, respectively, and graphical representations similar to **Figure 7** showed that foF2 values from MAXIMUF were within 1.5 times the RMS value of the CCIR values

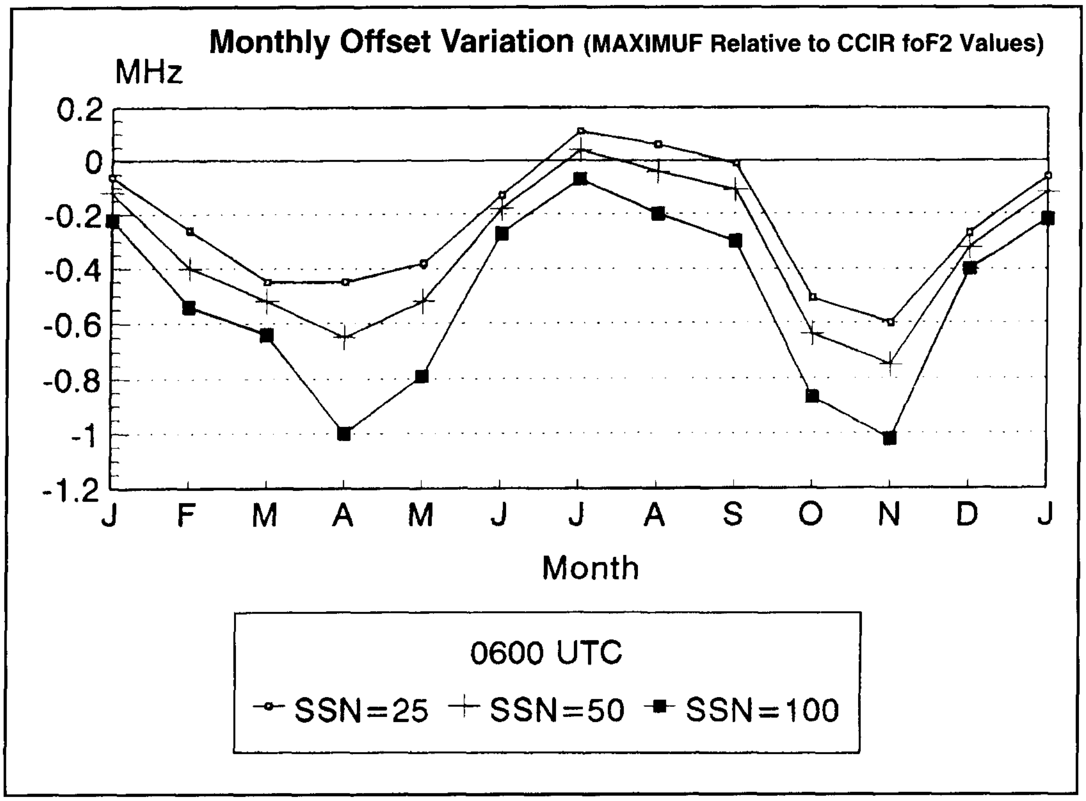

Figure 12. Monthly variation of the mean offset of foF2 values between the MAXIMUF algorithm and the CCIR database for three levels of solar activity at 0600 UTC.

for 90.3 and 87.5 percent of the points in the global maps.

In **Figure 7**, the sunlit hemisphere extends from 0 to 180 East longitude and the dark hemisphere of the earth goes beyond that region to 360 East longitude. In that regard, it is of interest to examine how the departures of MAXIMUF predictions from CCIR values varied in magnitude, across the entire globe as well as in those two hemispheres. Without any offset included, **Figure 8** shows how the foF2 differences across the globe were distributed in magnitude, peaking at -0.5 MHz. With +0.52 MHz added, the distribution in **Figure 8** is displaced to the right by that amount and the distributions in the two hemispheres are as shown in **Figure 9**, more positive in the sunlit region and negative in the dark region. Those last points would be expected from the fact that critical frequencies are higher in a sunlit region than in the dark, where the electron density is slowly declining after sunset.

As mentioned above, an offset of 0.52 MHz was applied to all points and the departures of MAXIMUF foF2 values from CCIR values noted using constant differences across the globe, either 1 MHz or 1.5 time the RMS difference of the entire sample of points. But it would be worthwhile to take note of the fact, mentioned above, that critical frequencies are higher in sunlit regions than in the dark, and vary or use different magnitudes of departure in the regions. One possibility would be a constant percentage of difference, say 15 percent of the CCIR value; that gives a result shown in **Figure 10**.

With that choice, differences of 0.6 MHz would be examined in a region of the dark ionosphere beyond 180E—where the CCIR value is 3-4 MHz and greater than 1.5 MHz in the low-latitude sunlit region where CCIR values are above 10 MHz. That approach gives positive differences of 15 percent or more at 57 points in **Figure 10**, and negative differences of 15 percent or more at 40 points. Thus, the positive and negative points correspond to 10.8 and 7.6 percent of the 527 points, respectively, and for 81.6 percent of the points, the MAXIMUF foF2 value is within +/- 15 percent of the CCIR values across both the sunlit and dark regions.

Turning now to changes with sunspot number, the hourly values for the offset of MAXIMUF foF2 values in March relative to the CCIR ionosphere are shown in **Figure 11**, for sunspot numbers of 25, 50, and 100. That figure shows that MAXIMUF, as formulated by Fricker, has a small daily variation in the offset. **Figure 12** shows that MAXIMUF contains seasonal features as well, displaced a month or so relative to the solstices and equinoxes. If those

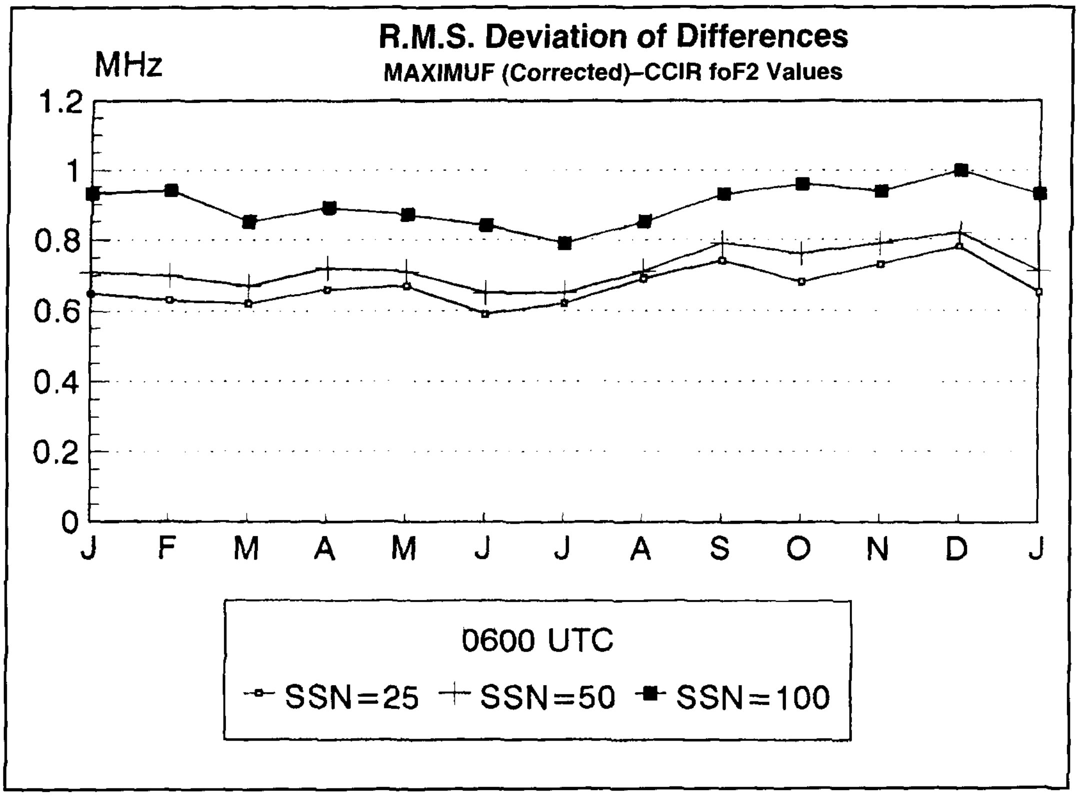

Figure 13. Monthly variation of the RMS differences of foF2 between the MAXIMUF algorithm (when corrected for offset) and the CCIR database for three levels of solar activity at 0600 UTC.

offsets are added to MAXIMUF foF2 values, the RMS differences show smaller seasonal effects, seen in **Figure 13**.

Using the offsets shown in **Figure 12** and the RMS deviations in Figure **13**, difference maps similar to **Figure 7** for the three sunspot numbers were constructed. Examination of the data points for all the months of the year showed that the predictions from MAXIMUF were within 1.5 times the RMS values for at least 85 percent of the data points and within 1.5 MHz of the CCIR values. As with the case in **Figure 7**, negative excursions beyond 1.5 times the RMS values exceeded positive ones by a factor of 1.5.

Discussion

By way of validating the MAXIMUF F-layer algorithm, its global ionospheric maps have been compared with those obtained from the CCIR reference ionosphere. The comparisons were made on a monthly and hourly basis, using a grid of 527 points—17 in latitude and 31 in longitude—and included the effects of solar activity. A different validation method was used earlier when Damboldt and Suessman made comparisons that were summed over latitudes, longitudes, hours of the day, months of the year, and two sunspot numbers. That comparison showed predictions from FTZMUF2 were much closer to the reference ionosphere values than those from the MINIMUF program. But in practice, predictions are not made nor used in such a summary fashion; instead, they are needed for hours of the day in a given month and for a specific sunspot number. Thus, the comparison of hourly ionospheric maps, as performed here, has more meaning in a practical sense.

The results from the comparison of ionospheric mapping points show how the MAXIMUF's F-layer algorithm gives foF2 values that are generally too low at the outset. In its applications making propagation predictions, the critical frequencies for vertical incidence, foF2, are increased by an M-factor—something like a factor of 3—to obtain the critical frequencies for oblique propagation. As a result, from the mean differences presented above, critical frequencies for oblique propagation derived from the original MAXIMUF algorithm could be low by as much as 3 MHz for a sunspot number of 100 or with only a small error, depending on the season, time of day, and geographic location.

How important those differences would be depends on whether the location in question is the one which determines the MUF or not.

However, from this study, it would appear that empirical correction factors, separable with regard to time of day, season, and level of solar activity, could easily be applied to the MAXIMUF algorithm and thus bring its mean predictions into closer harmony with those from the CCIR ionosphere. Variations about the mean values would still remain but be within reasonable bounds.

Another point of interest in the present discussion is the efficiency of using algorithm-derived MUF values as compared to those from more computationally intensive routines that use the full ionospheric database. The use of algorithms allows rapid MUF calculation, and all the adjustments suggested for the MAXIMUF algorithm could easily be put in look-up tables or expressed simply in analytical form, thus not significantly increasing the computing time required for a MUF calculation. Beyond that, with the fact that actual ionospheric conditions at any time are most likely within the limits of the FOT and the HPF, use of adjusted algorithms with modest RMS deviations may be sufficient, falling within that range, and thus more than adequate for prediction purposes and certainly faster than any result derived from a full ionospheric database.

In conclusion, it would be of interest to see this technique applied to other F-layer algorithms, showing how their ionospheric maps compare with those of international reference ionospheres, either CCIR or URSI, and whether any systematic differences exist in their application. That would be particularly the case with regard to the IONCAP program, released first in 1978. But such comparisons are best done by those who created the algorithms, have modified existing ones or have familiarity and experience with the programs. Thus, I must defer to others for any further discussion of the matter.

Acknowledgements

I am deeply indebted to Cary Oler of the Solar Terrestrial Dispatch in Canada for his generosity in providing a file that provided the foF2 data associated with ionospheric maps in the PROPLAB PRO program. Without this able assistance, the study would have been more difficult and taken much longer to complete.

In addition, I would like to express my gratitude to Hans Bakhuizen of Radio Netherlands for making available copies of the work of Raymond Fricker, and to Johannes Ruys, N6ZX/7, for sharing them with me back in 1985. At this late date, we are in debt to Raymond Fricker and can only marvel at his insights and analytical abilities. ■

REFERENCES

1. R.B. Rose and J.N. Martin, "MINIMUF 3.5, Improved version of MINIMUF-3, a simplified HF MUF prediction algorithm," Naval Ocean Systems Center, Technical Document 201, 1978.
2. J. Caruana and M.W. Fox, "A Critical Study of the NOSC HF Prediction Program MINIMUF," Technical Report IPS-TR85-07, Department of Science, IPS Radio and Space Services, Australia, June 1985.
3. Th. Damboldt and A. Suessmann, A Simple Method of Estimating foF2 and M(3000) with the Aid of a Home Computer, Research Department of the Deutsche Bundespost, Darmstadt, Germany, 1988.
4. CCIR, "Atlas of Ionospheric Characteristics," CCIR Report 340, Int. Telecom. Union, Geneva, 1983.
5. R. Fricker, "A Microcomputer Program for the Critical Frequency and Height of the F-layer of the Ionosphere," BBC External Services. Fourth International Conference on Antennas and Propagation (ICAP 85), IEEE-URSI, University of Warwick, 1985.
6. D. Bilitza, The International Reference Ionosphere 1990, National Space Science Data Center, NSSDC-WDC-A-R&S 90-20, Greenbelt, Maryland 1990.
7. K. Davies and C.M. Rush, 'High Frequency Ray Paths in Ionospheric Layers with Horizontal Gradients," Radio Science, Vol. 20, 95-110, 1985.

PRODUCT INFORMATION

Hand-Portable Power

Cutting Edge Enterprises' 9-lb, portable power supply provides 140 watts of 115 volts AC (surges to 200 watts) and up to 20 amps of 12-volt DC power. The compact (4" x 4.5" x 6") Powerport 149 contains a 12-volt, 9-amp gel-cell battery and may be charged via an automobile's cigarette lighter, the unit's fully automatic wall charger, or optional flexible solar panels.

Powerport will run and charge handheld radios, test equipment, soldering irons, emergency lighting, laptop computers, handheld GPS receivers, electric hand tools, video cameras, or fax machines in the field. Prices range from $159.95 plus shipping for the 140-watt, 9-amp model; $136.95 plus shipping for the 140-watt, 7-amp model; and $114.95 plus shipping for the 50-watt Powerport.

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UPGRADING BOONTON MODELS 92/42 RF VOLTMETERS

Replace the mechanical chopper and AC amps with chopper-stabilized, low-offset op amps

The Boonton model 92/42 transistorized RF voltmeters have been industry standards **since** the 1970s. Their wide frequency range, extending above 1 GHz (or 18 GHz depending on the probe used) combined with eight ranges (in the model 92) covering 1 mV to 3 volts full-scale, make them very useful equipment for general RF/microwave work. The model 92 was designed to replace the model 91 which used vacuum tubes. However, all models use a mechanical chopper to convert low-level DC (from the rectified RF) to a low-frequency AC signal (typically 90 Hz). Like all mechanical devices, this chopper is subject to wear and constitutes the weakest link in an otherwise excellent design.

This article describes a redesign of the Boonton model 92 and 42 front-end circuits that replace the mechanical chopper and AC amplifiers with chopper-stabilized, very low offset op amps. I'll also show how to recalibrate the modified instrument and provide results of linearity measurements done on my own instrument.

Original circuit

Refer to the block diagram shown in **Figure 1**. The RF probe generates a bipolar DC output that's converted to a square wave at the chopper frequency. The chopper is like a mechanical SPDT relay that alternately selects the probe "+" and "-" terminals to perform DC to AC conver-

RANGE (mV)	ATTENUATOR GAIN	FIXED GAIN AMP	SECOND AMP GAIN	TOTAL GAIN	NOMINAL INPUT V AT F.S.	MEASURED INPUT V AT F.S.
1	1	100	1000	100000	3.10E 05	3.80E 05
3	1	100	100	10000	3.10E 04	3.76E 04
10	1	100	10	1000	3.10E 03	3.68E 03
30	1	100	1	100	3.10E 02	3.19E 02
100	0.18716	100	1	18.72	0.166	0.1844
300	0.04231	100	1	4.231	0.733	0.748
1000	0.013325	100	1	1.333	2.326	2.610
3000	0.004231	100	1	0.4231	7.327	8.660

Table 1. Boonton RF voltmeter gain distribution before modifications. Rightmost column data is for the author's voltmeter.

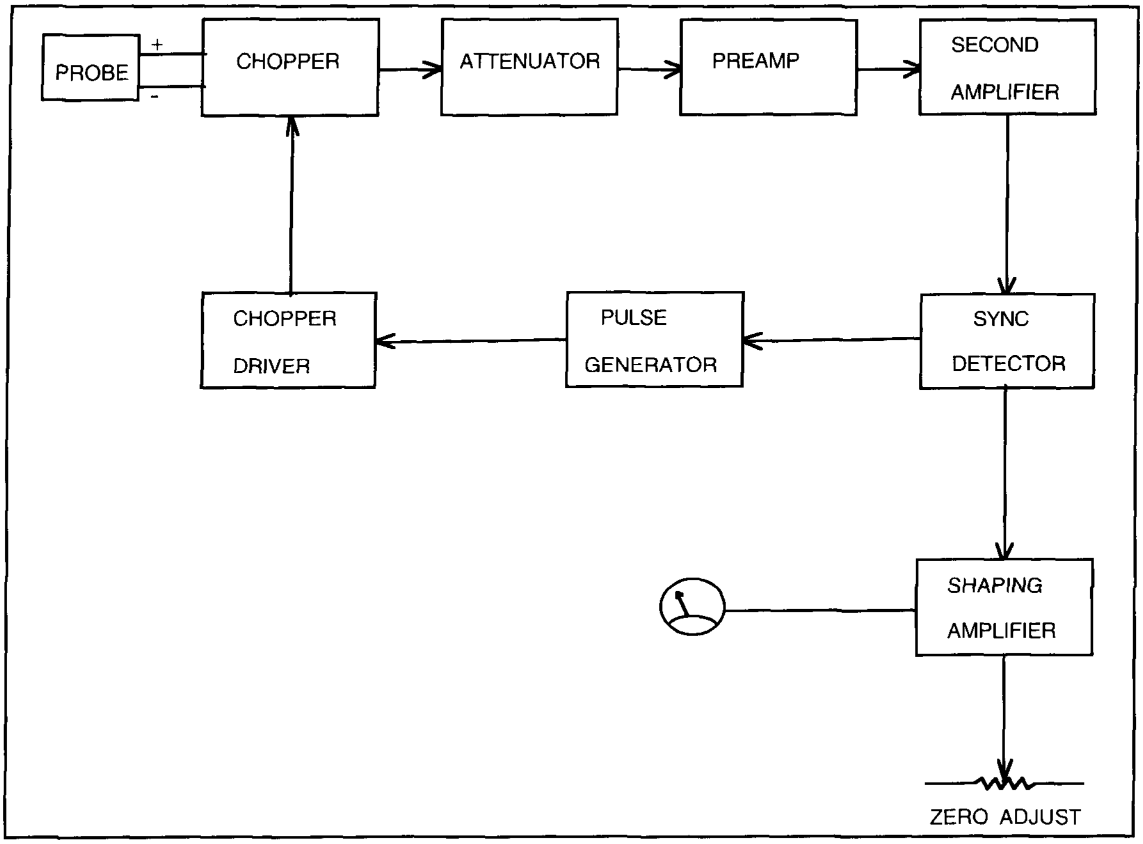


Figure 1. Model 92 series block diagram.

sion. This allows low-level AC signals to be amplified more easily than DC signals, because the amplifier offsets don't affect the AC signals. The attenuator scales the AC input level to a maximum of about 30 mV peak, which allows switching of AC signals with FET switches. Two AC-coupled amplifiers follow the attenuator. The first amplifier has a fixed gain of 100, while the second amplifier has gains of 1, 10, 100, or 1000 in order to bring the probe voltage to -3.1 volts full-scale at the sync detector output. **Table 1** shows the attenuation and gain values for the instrument's eight ranges. The sync detector converts the AC amplified signals back to DC and provides filtering against nonsynchronous signals—such as white noise that's present on the most sensitive ranges.

The amplifiers and attenuators are located on the motherboard; the chopper driver/sync detector resides on a plug-in pc board. A second pc board, called the shaping amplifier, provides the nonlinear compensation required, depending on the input level and the selected range, to give an output proportional to the AC signal voltage present at the probe input.

Proposed circuit

Figure 2 shows the revised block diagram of the new amplifier circuit. The previous chopper driver/sync detector circuits are replaced by two integrated circuit chopper-stabilized preamplifiers with gains of 1 or 100, depending on the selected range. The second amplifier has range-dependent gains that vary from 0.42 to 1000. As shown in **Table 2**, the gain of the second amplifier has been chosen to provide the same overall gain on every range as the unmodified voltmeter. This also means that the shaping amplifier may be reused because the new amplifiers provide the same -3.1 volts at full-scale output.

Figure 3 shows the circuit schematic of the new amplifier that plugs into the chopper driver/sync detector slot of the voltmeter (connector slot *J101*). This slot provides the ±15 volt supplies and the range control lines for switching the amplifier gains. The ICL7650 chopper-stabilized operational amplifiers IC1 and IC2 amplify the differential probe signals. Resistors R4, R5, and R6, and photo-MOS relay IC6 set the amplifier gain. The LO 4 line is derived from diode logic on the voltmeter motherboard. It turns on IC6 output MOSFETs at the 30-mV range and below. This sets the preamplifier gain at 100. The photo-MOS relay IC6 minimizes thermally induced offsets generated at the relay "contacts" (normal relay contacts can easily generate 10 µV offsets, while this device has 1 µV typical offset specification). IC6 also pro-

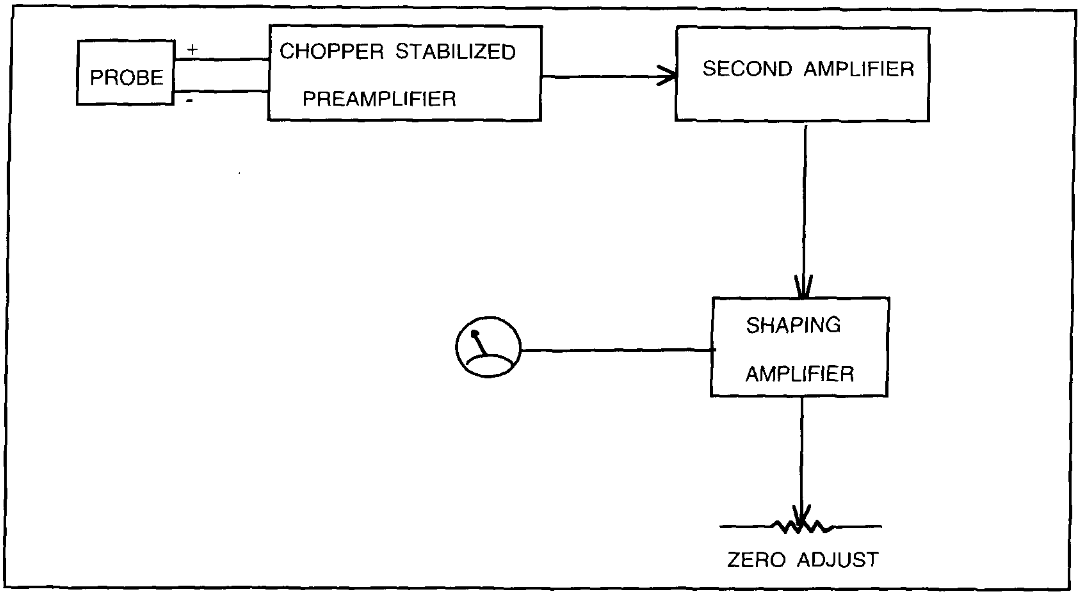


Figure 2. Revised block diagram incorporating new amplifier circuits.

vides very low leakage consistent with the high input impedance (5.6 megs) of the chopper-stabilized amplifiers. IC3 is a unity gain amplifier that provides balanced-to-unbalanced conversion. The second amplifier, IC4, has controlled gains of 0.42 to 1000, as set by the FET switches Q2 to Q7 with their associated drain resistors, or with analog switches IC5A and IC5D. On the 3-volt range, all FET switches are open and IC4 gets its input via resistors R11, R12, and R13. This "T" circuit synthesizes a 2.363-meg resistor by using resistor values of 100 k and below, thus eliminating the need for precise resistors above 1 meg. I also used the same technique for R17, R18, and R19 to synthesize a 1.734-meg resistor. Range selection is accomplished by grounding one of the FET gates either via the front panel push buttons or via the rear remote control connector.

The main concern when designing high-gain amplifiers is to minimize DC offsets. Here, it is mandatory to use chopper-stabilized op amps with input offsets in the microvolt range, as the probe DC output is around 31 μV at 1-mVAC input. With the preamplifier operating at a gain of 100, this translates to a 100-μV offset at the preamplifier output.

The voltage offsets at the second amplifier output may come from two main sources: first from its input offset voltage (typically 30 microvolts), and second from the FETs and analog switch leakage currents (0.1 nA gate-to-source leakage for every FET and 1 nA typically for every switch). Therefore, the revised offsets at the maximum total gain of 100,000 (1-mV range) will be:

Preamplifier offsets: 1 μV * 100 000 = 100 mV

Second amp offsets: 30 μV * 1000 = 30 mV

Second amp switch leakage currents: (0.1 nA * 6 FET's + 1 nA * 3 switches) * 1 Mohms = 3.6 mV.

RANGE mV	PREAMP GAIN	PREAMP OUTPUT VOLTAGE	GAIN SECOND AMP	TOTAL GAIN	FULL SCALE INPUT VOLTAGE	
1	100	0.0031	1000	100000	3.10E	05
3	100	0.031	100	10000	3.10E	04
10	100	0.31	10	1000	3.10E	03
30	100	3.1	1	100	3.10E	02
100	1	0.166	18.716	18.72	0.17	
300	1	0.631	4.909	4.909	0.63	
1000	1	2.326	1.333	1.333	2.33	
3000	1	7.327	0.4231	0.4231	7.33	

Table 2. Boonton RF voltmeter gain distribution after modification.

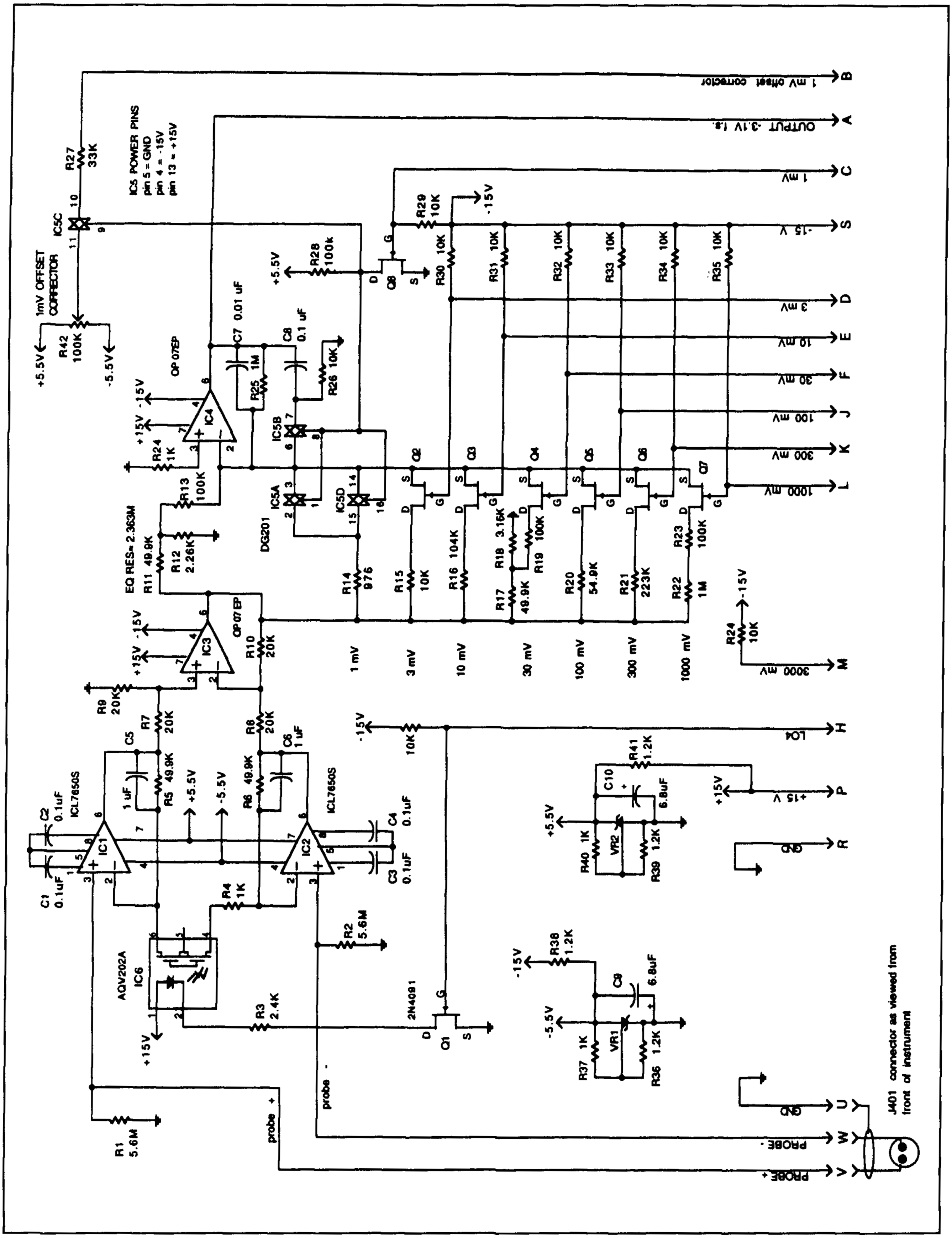


Figure 3. DC amplifier for modified Boonton 92 RF voltmeter.

Reference Designator	Parts Description
DIP plugboard	VECTOR #3662A-6 plugboard Includes ground plane on one side with clearance holes that clear the 0.1 in. X 0.1 in. hole pattern hole size = 0.042 in., 2 x 22 fingers, spaced 0.1 in. Cut to fit inside voltmeter.
IC1, 2	Op amps ICL7650SCPA1 (Harris)
IC3, 4	Op Amps OP-07 EP (Analog Devices)
IC5	Analog switch DG201CJ (Harris)
IC6	Photo MOS relay AQV202A (AROMAT)
Q1-Q8	N Channel FETs 2N4091 or 2N5459 or equivalent
VR1, 2	Shunt Regulators TL431CP (Texas Instruments)
C1-C4	Polycarbonate capacitors 0.1 uF 10%
C5, C6	Ceramic capacitors 1uF matched to 1%
C7	Ceramic capacitor 0.01uF 10%
C8	Ceramic capacitor 0.1uF 10%
C9, C10	Tantalum capacitor 6.8 uF
R1, 2	Resistors metal film 5.6 Mohms 5% (NOTE 1)
R3	Resistor 2.4K 10%
R4, 24, 37, 40	Resistors metal oxyde 1K 1%
R5, 6, 11, 17	Resistors metal oxyde 49.9K 1%, matched 0.1%
R7-10	Resistors metal oxyde 20K 1%, matched 0.1%
R12	Resistors metal oxyde 2.26K 1%
R13, 19,23	Resistors metal oxyde 100K 1%
R14	Resistors metal oxyde 976 ohms 1%
R15	Resistors metal oxyde 10K 1%
R16	Resistors metal oxyde 104K 1%
R18	Resistors metal oxyde 3.16K 1%
R20	Resistors metal oxyde 54.9K 1%
R21	Resistors metal oxyde 223K 1%
R22, 25	Resistors metal oxyde 1M 1%
R24, 26, 29-35, 43	Resistors 10K 10%
R27, 28	Resistors 100K 10%
R36, 38, 39, 41	Resistors metal oxyde 1.2K 1%
R42	Multiturn pot PCB mount 100K

NOTES:
R1, R2 resistors 5.6 Mohms: one already exists on motherboard as R104
Parts may be obtained from NEWARK Electronics tel 1-800-463-9275

Table 3. List of components.

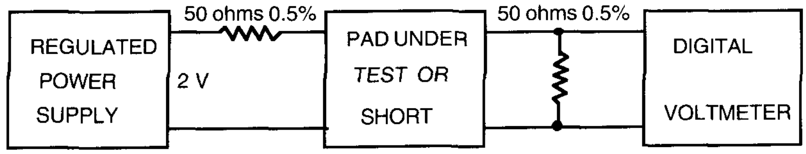
Total offset voltage at the second amplifier output: 133.6 mV.

This corresponds to 4.3 percent (0.133 volt ÷ 3.1 volts) of full-scale, which is quite acceptable.

These simple calculations show that the combined offsets of the two amplifiers are dominated by the offsets of the preamplifier. The front panel zero adjustment cancels amplifier offsets at the shaping amplifier input on the three most sensitive ranges. Note that its offset canceling range increases by a factor of 10 on the 3-mV range and by a factor of 100 on the 1-mV range.

On the 1-mV range, two analog switches are used in parallel to increase the ratio of gain setting resistance R14 to the IC5A and IC5D switch resistance, as IC5 resistance will vary somewhat with temperature. On the 3-mV and 10-mV ranges, the voltmeter zero adjustment is the same for both ranges. IC5C brings in the offset corrector pot R42 on the 1-mV range. Pot R42 (connected in parallel with the front panel zero pot) fine tunes the offset on the 1-mV range, so the front panel zero doesn't need readjustment.

The DC probe signal is filtered by R5*C5 and R6*C6 (50-msec time constant) and R25*C7 (10-msec time constant). This sets a basic 60-msec time constant on all ranges. Note that the 0.1 μF capacitors connected across the probe + and - terminals also contribute additional filtering. On the 1-mV range, IC5B also turns on to add an additional 100-msec time



$$\text{PAD ATTENUATION} = \frac{\text{DC VOLTAGE WITH PAD}}{\text{DC VOLTAGE WITH SHORT}}$$

Figure 4. Pad attenuation measurement.

constant with R25*C8. Note also that R26 provides "hitless" switching because the left side of C8 is kept at ground potential (with a 1-msec time constant) when IC5B is off. When IC5B turns on, it connects the left side of C8 at IC4 virtual ground, preventing the meter from momentarily going off-scale. Resistor R26 has a negligible effect on IC4 open loop gain when IC5B closes.

The two shunt regulators, VR1 and VR2, provide ±5.5 volts to the chopper-stabilized amplifiers IC1 and IC2 because their supply can't exceed ±7 volts. This limits the amplifier outputs to ±5 volts and also sets IC1 and IC2 maximum gain at 100. See **Table 3** for the component list.

Construction

Refer to the detailed component description in **Table 3**. I built the circuit on an epoxy glass Vector plug-in board with a ground plane on the component side. It has gold-plated fingers that fit into the edge connector of the chopper driver/sync detector circuit (J101). The axial components are held in place with wire wrap pins—T42-1 "microclips"—also from Vector Co. I used wire wrap sockets for all ICs. I suggest you cut all pins, so they protrude about 1/8 inch on the wiring side. The DC probe output must be rewired to unused pins V and W of the plug board via a two-conductor shielded cable. Add a short jumper on the voltmeter motherboard between J101 pin B and the shaping amplifier connector J102 pin A. It's a good idea to clean the plug-in board with methanol to eliminate potentially conductive residues produced during the fabrication and to remove finger grease and salts. It's especially important to minimize stray leakage paths around the probe signals inputs. Note that the DVM motherboard is epoxy based and coated with a varnish to keep its insulation resistance up under high humidity conditions.

Testing and calibration

Verify that the ±15-volt and ±5.5-volt supplies appear at the proper pins before inserting the ICs. If necessary, adjust the -15 volts to read -15 ± 0.1 volts using pot R145 located on the motherboard.

Connect the RF probe with a 50-ohm termination, and verify the chopper amplifier offset voltage by connecting a DVM at IC3 output pin 6. Set the range to anywhere between 1 and 30 mV. The DVM should read within 0.0 ± 0.5 mV. With the range set to the 3-mV scale,

VOLTMETER F. S. RANGE mV	PAD NOMINAL ATTENUATION dB	PAD MEASURED ATTENUATION	SET GEN LEVEL mV
1	20+30	______	______
3.16	20+30	______	______
10	20+30	______	______
31.6	20	______	______
100	20	______	______
316	0	1.00	316
1000	0	1.00	1000
3160	0	1.00	3160

GENERATOR LEVEL FOR F. S. = F. S. RANGE / PADATTENUATION

Table 4. Boonton voltmeter calibration.

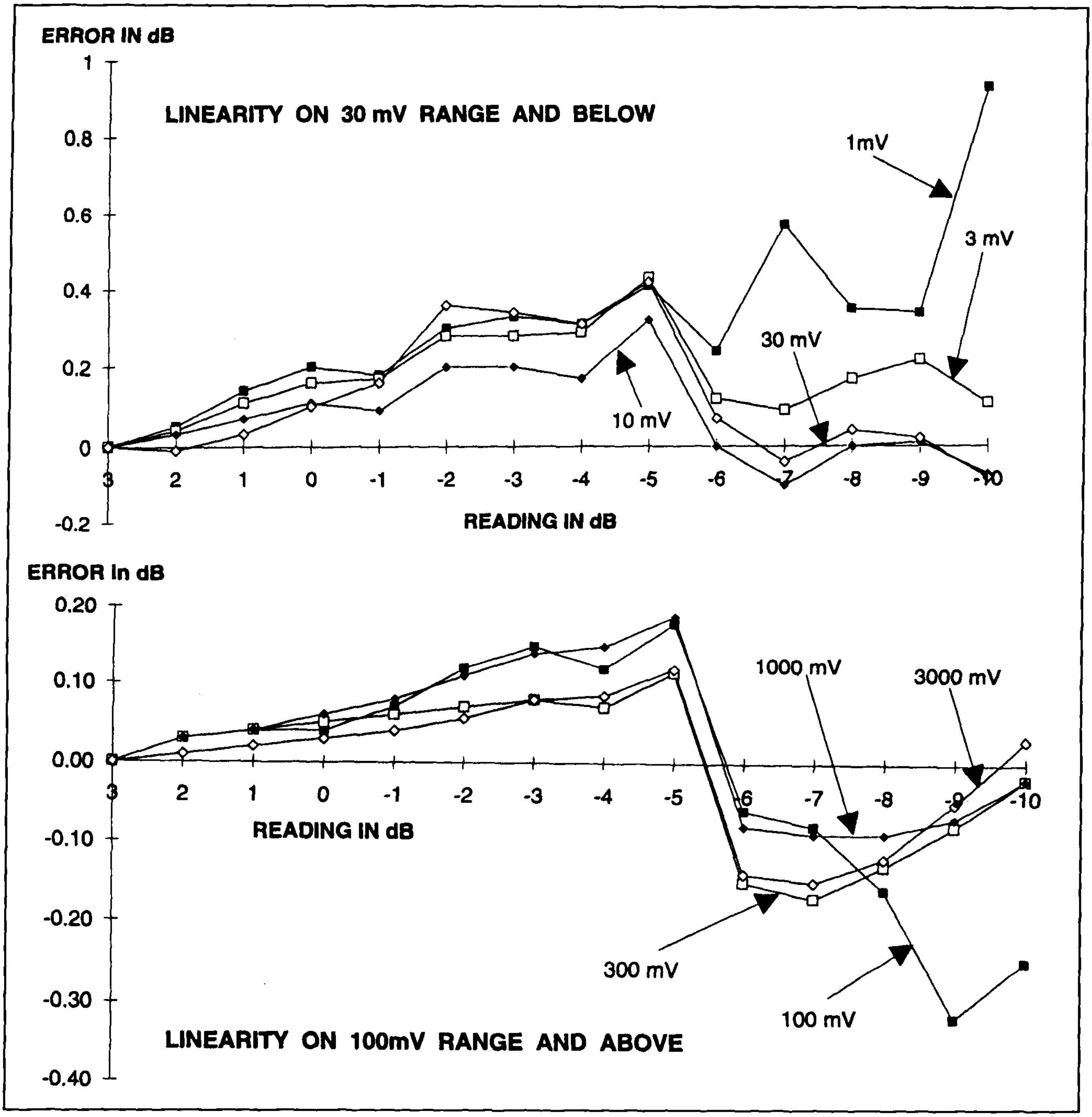


Figure 5. (A) Linearity on 30-mV range and below. (B) Linearity on 100-mV range and above.

adjust the front panel zero control for a zero reading on the Boonton voltmeter. Set the 1-mV range, and adjust R42 on the plug board (1-mV offset corrector) for a zero reading on the voltmeter. The remainder of the adjustments consist of calibrating the full-scale gain pots for every range.

I recommend performing the calibration with the 50-ohm probe termination in place, as this is most likely the way the voltmeter will be used most of the time (note that probe models 41-4x have a built-in 50- or 75-ohm termination). Using the 50-ohm shielded probe termination also minimizes the possibility of picking up stray signals with the high-impedance probe. Be careful if you use an audio signal generator for calibrating the RF voltmeter! I found that my Krohn-Hite 4200 audio generator acts like a VHF antenna and picks up close to 1 mV of broadband VHF signals because it doesn't have RF-type shielding. This occurred even with the oscillator turned off and with the power cord disconnected at the generator side. The stray pickup completely disappeared when I used an

PROBE TYPE	FREQUENCY RESPONSE	NOTES
91-4C	1 KHz - 250 MHz	Originally used on model 91 (vacuum tube)
91-12F	10 KHz - 1.2 GHz	Standard
952001A (silver barrel)	10 KHz - 1.2 GHz	Standard
Power Sensor 41-4E (50 ohms)	200 KHz - 18 GHz	Normally used on model 42xx microwattmeter
Power Sensor 41-4C (75 ohms)	200 KHz - 1 GHz	but works fine on model 92, (+13 dBm max.)

Table 5. RF voltmeter probes usable with Boonton Model 92 voltmeter as verified by the author.

RF signal generator with a shielded output attenuator. It's also a good idea to check the resistance of the probe termination with your ohmmeter; it should read 50 ±1 ohms.

I recommend using a minimum frequency of 100 KHz for calibrating the RF voltmeter for all probes except the 75-ohm power sensor model 41-4C, which should be calibrated at 1 MHz. Use an audio signal generator covering 100 Hz to 100 KHz (or 1 MHz) with less than 1 percent distortion. As shown in **Table 4**, the following calibration voltages are required: 316 mV, 1.000 volt, and 3.16 volts into a 50-ohm load. Measure these levels at 100 Hz using a DVM set on AC ranges. Using a scope, verify the generator flatness up to the calibration frequency of 100 KHz. Use the measured level at 100 Hz as the reference. Note that inexpensive DVMs (like the RadioShack variety) cannot be used at 100 kHz because they typically lose accuracy above 1 kHz and below 300 mV. Consult your DVM manual for frequency response specifications. If necessary, adjust the generator output level to maintain a constant peak-to-peak amplitude up to the calibration frequency.

Adjust the front panel zero and calibrate the lower ranges (1 mV to 100 mV) using previously calibrated 50-ohm, 20-dB and 30-dB coaxial pads or a 10-dB step attenuator. Use coaxial pads here to prevent stray pickup and to attenuate the low-level noise coming from the audio generator. Ensure that the pad reflection coefficient at both ends is below 1 percent: terminate one side of the pad into 50 ohms and verify the DC resistance on the other side. It should read 50 ±1 ohms. Repeat this test for the other side.

For greatest accuracy, first verify the exact attenuation of the pads at DC. See **Figure 4** for the test set-up. As shown in **Table 4**, compute the signal generator levels required for a full-scale voltmeter reading by dividing the full-scale voltage for the range being calibrated by the pad attenuation measured previously. Adjust the full-scale gain pots on every range for a full-scale reading. Note that the RF voltmeter calibration is only valid for the probe used during calibration. Exchanging one probe for another of the same or different model normally requires that a new calibration be performed. Errors up to 5 percent could result if the recalibration is omitted. As shown in **Table 5**, many types of probes may be used with the RF voltmeter. I found that the power sensor probe model 41-4E can be calibrated at 100 kHz, even though it is rated from 200 kHz up.

Linearity tests

Refer to **Figures 5A** and **5B** for voltmeter linearity test results as measured on my unit with probe model 91-12F. The test frequency used was 100 kHz. The reference point is at full-scale. I performed linearity tests from +3 dB (full-scale) down to -10 dB, in 1 dB steps, for all eight ranges of the voltmeter—using a calibrated step attenuator that covers 0 to 13 dB of attenuation. For these tests, I measured the RF voltmeter DC output with a Fluke model 8920A digital voltmeter that converts the DC readings to dB ratio. The estimated accuracy is ±0.02 dB.

Conclusion

The new circuit described above constitutes a reliable replacement for the old chopper circuit. It fits in place of the chopper pc board and is easy to calibrate. This circuit has been verified in my model 92A RF voltmeter. The same chopper/amplifier circuit works on the model 42 with one difference: the +20-dBm range is missing. After looking at the model 42 schematic, it appears that the above DC amplifier circuit could be used in place of the chopper driver/sync detector circuit. I have verified that the probes for both models can be interchanged as long as the power input to the 42-x probes doesn't exceed +13 dBm. The model 42 is called a microwattmeter because it only displays power (into a 50-ohm load), even though the two instruments are basically RF voltmeters. The above modifications also retain the programmability features of the voltmeter. A future article is planned that will describe how it can be easily interfaced to a PC computer.■

Bibliography

1. *Boonton Model 92 RF Voltmeter Service Manual*. Includes circuit diagrams.
2. *Boonton Model 45 RF/Microwave Power Meter Service Manual*. Includes circuit diagrams.
3. Keithly Instruments, Inc., *Low Level Measurements*, 1984 edition.

Hank Olson, *W6GXN*
(Silent Key)

GLOBE WIRELESS RIDES THE AIRWAVES AGAIN

A legendary radio service returns

As of January 1995, a name well remembered by older commercial "brass-pounders," once again came to life—Globe Wireless. Thought to have ended operation in 1960, Globe Wireless' name will again be heard on the marine HF radio bands, doing the kind of radio communications for which it was originally created: ship-to-shore radio.

While Globe Wireless and its antecedent, Dollaradio, were by no means the *first* in the ship-to-shore radio business, they certainly were one of the first companies to conduct their business using high frequency (HF) tube-type equipment. In fact, neither Globe Wireless nor the other pioneering San Francisco Bay area radio communications company, Federal Telegraph Company, *ever* used the older spark-gap technology.[1]

A cooperative venture

Globe Wireless came into being due to the efforts and cooperation of two San Francisco Bay area men: Ralph Morell Heintz and Robert Stanley Dollar. Ralph Heintz was the technical man and R. Stanley Dollar was the businessman. It seems odd, looking at their backgrounds, that these two men should have come together to create one of the world's most innovative radio communications companies. Ralph Heintz studied chemistry at the University of California, Berkeley, and at Stanford University. R. Stanley Dollar was the son of a lumberman whose father got into the shipping business out of the necessity to provide logistical support for his Pacific Coast lumber mills and timber stands.

Heintz's college work towards a chemistry degree at U.C. Berkeley was interrupted by World War I. Because of his interest in ham radio, he was assigned to the Army Air Corps in England and became part of its early efforts to use radio direction-finding equipment aboard airplanes. Following the armistice, he returned to the San Francisco Bay area and completed his degree in chemistry at Stanford University

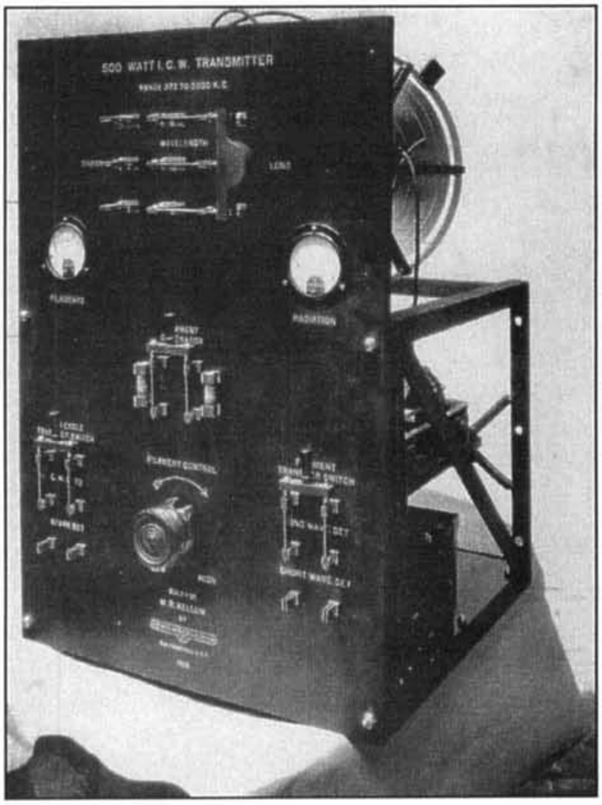


Photo A. KFUH MF transmitter.

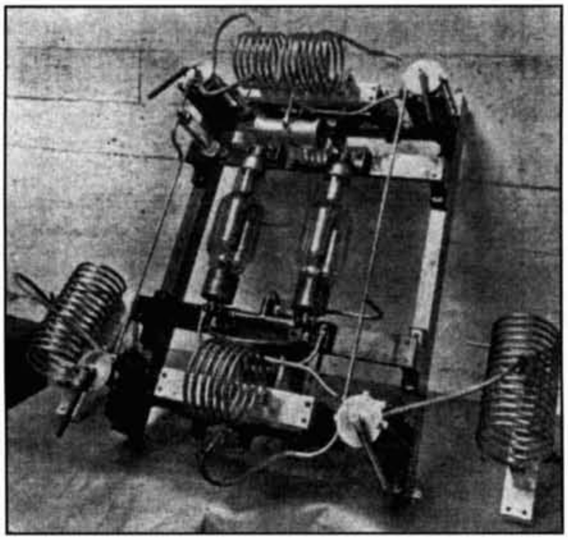


Photo B. KFUH HF transmitter.

in 1920. After graduation, he set up a one-man shop on Mission Street, in San Francisco, called Ralph Heintz, Scientific Apparatus. This small company made all sorts of custom equipment, from special medical apparatus for eye surgery, to individual tube-type radio equipment for experimental use.

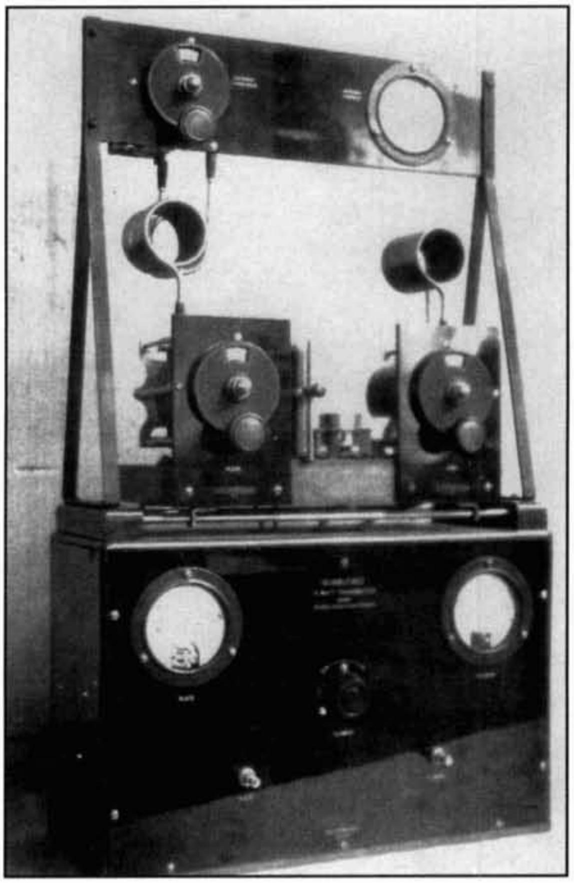


Photo C. H&K Bumblebee transmitter, front view.

In 1921, Ralph Heintz formed a corporation with his brother-in-law, Herman Kohlmoos, the original H&K, Heintz and Kohlmoos, incorporated in the state of California. Kohlmoos was brought into the corporation because he had the capital to expand H&K into its new quarters at 219 Natoma Street, San Francisco. As Heintz and Kohlmoos, the company became well known for designing and constructing radio equipment for expeditions, radio stations, military contracts, and even wealthy hams and yacht owners. An example of the Heintz and Kohlmoos custom-made radio transmitting equipment is shown in **Photos A** and **B**. These are MF and HF transmitters built for KFUH, on the yacht *Kaimiloa*, owned by M.R. Kellum. The HF performance of KFUH, run by legendary radio operator Fred Roebuck (FD), as the *Kaimiloa* sailed around the Pacific for two years, was to have a major impact on the acceptance of HF as a ship-to-shore radio mode.[1]

The *Kaimiloa* was no ordinary yacht; it was a four-masted schooner, with the lines of an extreme clipper; that is, it was 182 feet long, 38 feet in the beam, and 50 tons. It carried a load of lumber and other construction materials that Kellum planned to use to build a home for himself and his new bride, once they found "just the right island." Thus, the two-year voyage around the Pacific.

By 1924, Heintz and Kohlmoos had achieved some degree of fame, especially in the design and construction of HF equipment, but differences in temperament between Heintz and his brother-in-law led to dissolution of the company. By mutual agreement, in the dissolution settlement, Ralph Heintz kept the company and Herman Kohlmoos kept the building at 219 Natoma street.

A new partnership

Heintz and his old U.C. Berkeley friend Jack Kaufman formed a new California-based corporation. Kaufman was to be the business manager, a job for which Herman Kohlmoos was apparently not well suited. The new firm, Heintz and Kaufman, Inc., continued on at 219 Natoma Street for a time.

Heintz and Kaufman turned out all sorts of radio transmitting and receiving equipment during the period of 1926 through 1929 and also made related units, like motor-generator sets. Among H&K's customers were the U.S. Army Signal Corps, U.S. Marine Corps, California Fish and Game Commission, Byrd's Expeditions, Sir George Hubert Wilkins' Expedition to Antarctica, Examiner Publishing Co. (Hearst's mobile press station KUP), broadcast stations KFWM and KFDB, San Francisco Presidio's WVY, and Signal Corps

Communications station and amateur station 6HM (Retired Col. Claire Foster). **Photos C** and **D** show one of the better-known H&K transmitters—the "Bumblebee"—without tubes. It was a self-rectified, two-tube, tuned plate, tuned grid (T.P.T.G.) oscillator using 900 Hz AC, and it produced a rather high-frequency A2 note (2 x 900 Hz = 1800 Hz)—thus the name "Bumblebee." The circuit is shown in **Figure 1**. Standard RCA or Western Electric tubes were used, because all of these units were one-of-a-kind and considered experimental and, therefore, met the "experimental use only" restrictions of the tube manufacturers.[1]

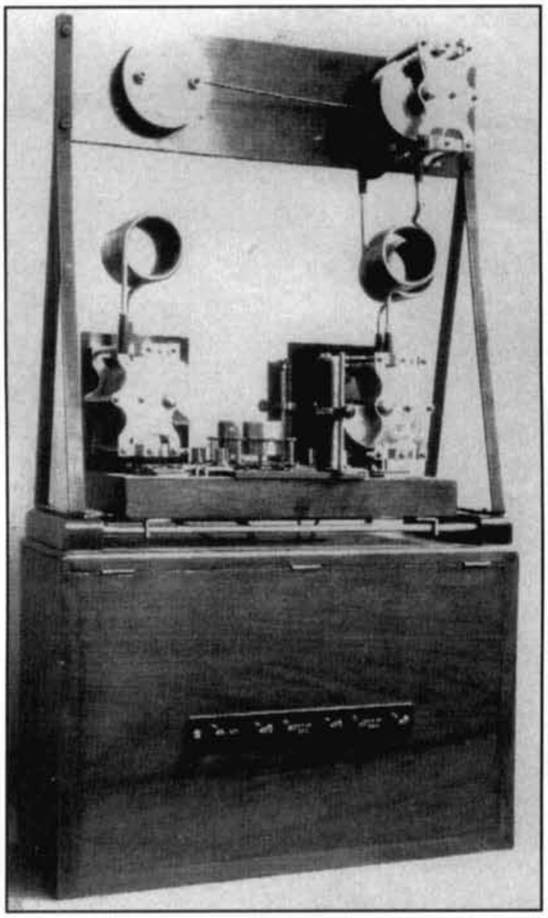


Photo D. H&K Bumblebee transmitter, rear view.

The Dollar Line

Robert Stanley Dollar was born July 6, 1880, in Bracebridge, Ontario, Canada, but moved with his parents to Michigan when he was only two years old. Then, when he was only eight, the family again moved to the U.S. West Coast, to San Rafael, California. Dollar was educated in public schools and, upon graduation from high school, attended Heald's Business College in San Francisco.

Seventeen-year-old R. Stanley Dollar went to work at his father's lumber business office as a bookkeeper on May 2, 1898, at a salary of $26 per month. The Dollar office, at that time, comprised two small rooms at 10 California Street, San Francisco. The Dollar "fleet" consisted of

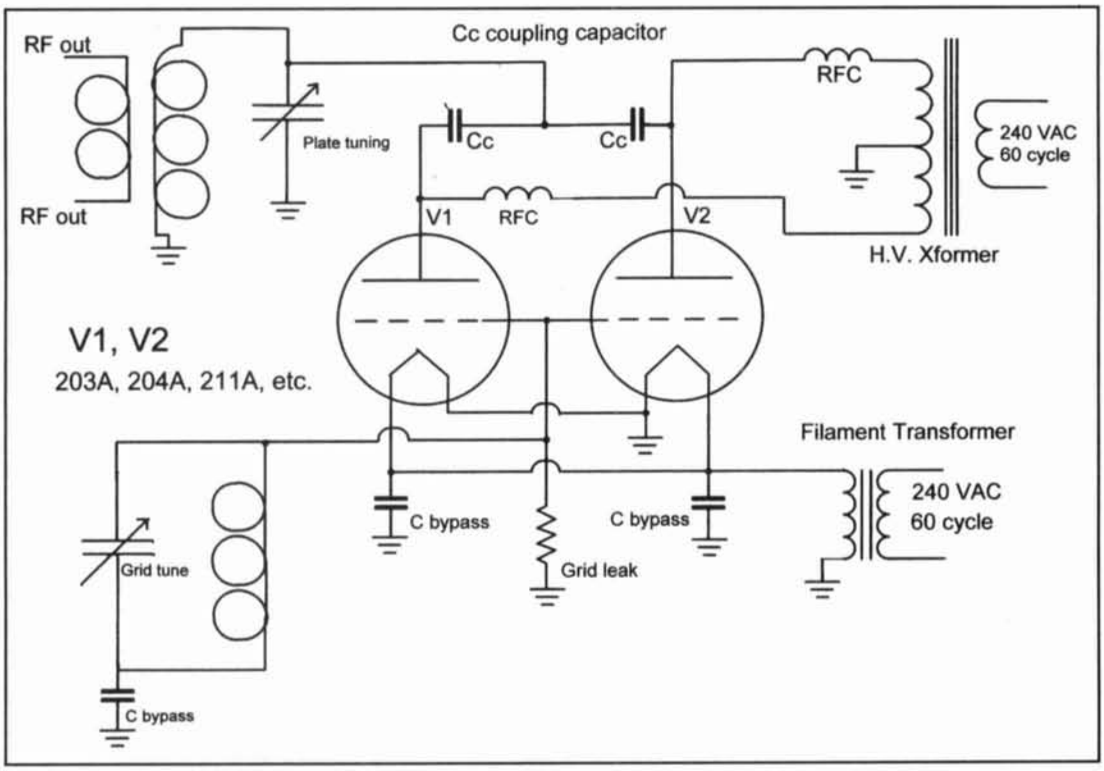


Figure 1. Circuit of early H&K T.P.T.G. transmitters.

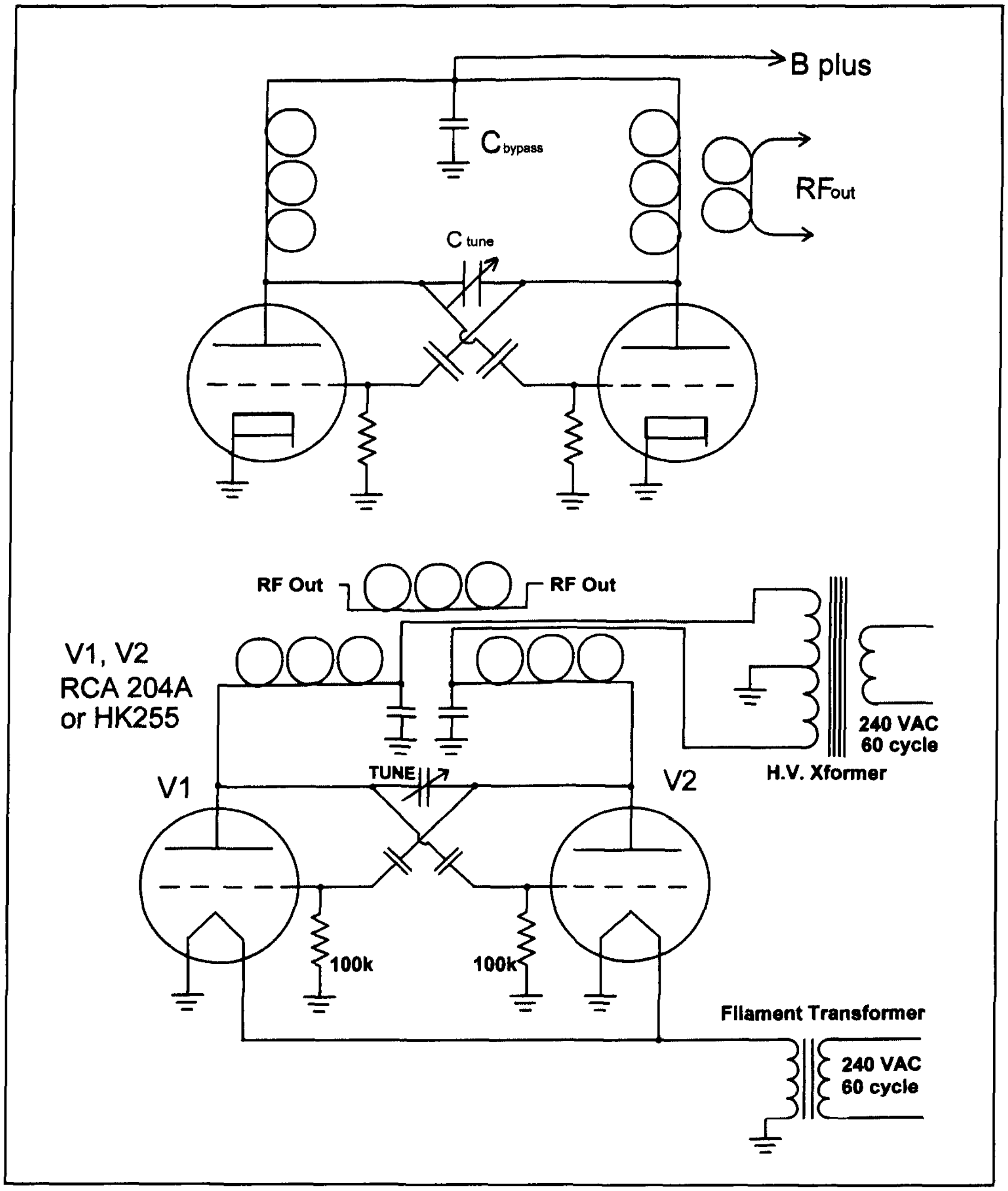


Figure 2. (A) Basic Simpson Oscillator using modern symbols. (B) Approximate MC-201 circuit.

the 200-ton steam schooner *Newsboy* that the senior Dollar had bought to carry lumber from his mill at Usal, in Mendocino County, to markets along the California coast.

Just five days after R. Stanley Dollar began working for his father's lumber company, the Dollar fleet doubled in size, with the launching of the *Grace Dollar* at Fulton Iron Works, in San Francisco. This was the first of a long line of Dollar ships that Captain Robert Dollar and his sons had built (and also bought or leased) over the next 40 years. The first, *Newsboy*, was a small wooden schooner. It was followed by the great steel freighters, and eventually Dollar owned palatial ocean liners that traveled the Pacific and around the world. The famous Dollar Steamship Line "Round the World" service was started in 1924, with ships sailing west every two weeks. These Dollar steamers were the first to carry the U.S. flag around the world on a regularly scheduled basis.

With a fleet of ships, large enough to be required to carry radio equipment and licensed operators, sailing worldwide, the Dollar Line found itself in the radio business. At that time, radios on ships used frequencies of 500 kHz and lower, usually with spark transmitters and crystal detector receivers. The combination of spark transmitter and (passive) crystal receiver usually meant short-distance communications with a shore station and then the relay of messages back to the home office via undersea cable. This was a very expensive way to communicate—between 50 and 80 cents per word for Dollar's San Francisco office to communicate with his ships at various locations in the Pacific. This rate was excessive for the long manifests and other communications of his fleet.

Enter HF radio

From 1924 to 1926, R. Stanley Dollar, who was now in his 40s and taking a major role in decisions concerning operations of the Dollar Line, saw that HF radio *might* be the answer to his shipping communications problems. The voyage of the *Kaimiloa* (KFUH) with its Heintz and Kohlmoos HF equipment, and a similar HF trial in the Pacific by the U.S. Navy (NRRL aboard the *USS Seattle*) were two major influences. Earlier, at R. Stanley's request, one of Dollar's shipboard operators, William H. Phillips, kept in contact with Charles King, a Dollar Line Superintendent at San Francisco, using the amateur bands as a further test of the feasibility of using HF for his ships. These amateur HF tests were performed on several routine Dollar trips across the Pacific to China. It was a sign of the times, and KFUH, NRRL, and the Phillips/King amateur communications all agreed. R. Stanley Dollar could see that HF was probably the answer to his ships' communication problems; and who better to implement such an HF system than Ralph Heintz, his San Francisco neighbor.

In 1926, R. Stanley Dollar went to Washington, D.C. and managed to obtain 17 high-frequency assignments for use by his shipping company and its new communication subsidiary "Dollaradio." The frequencies of 1.5 MHz and higher had been allocated to amateur radio use in 1912 when it was thought they were useless for long-range communication. But, now that HF had been shown to be a useful long distance mode, the Navy and companies like Dollar and Boeing Aircraft moved quickly to have these shortwave channels assigned to them for their exclusive use. The amateurs were reassigned to relatively narrow bands in the HF range at 160, 80, 40, 20, and 10 meters.

Dollar commissioned Heintz and Kaufman to design equipment for an HF ship-to-shore test to start in late 1928. The *President Taft* (KDRW) was to be the "ship" and Mussel Rock (now Pacifica) was to be the "shore." The Mussel Rock site had earlier been used as a "quiet location" by Fred Roebuck (FD) and Ron Martin (RM) of KUP in San Francisco when receiving weak HF press releases from Byrd's Antarctic Expedition. Martin, a close friend of Ralph Heintz, and Fred Roebuck, who roomed at the home of Ralph and Sophie Heintz, suggested the location, and Dollar purchased it to set up the HF test station 6XBB (Ralph Heintz's experimental license).

The required Heintz and Kaufman HF equipment was installed onboard the *President Taft* for its normal "Horseshoe Run" across the Pacific, starting in San Francisco and ending in Seattle. Radio operator Charles Cross ran the ship's HF equipment, with Chester Pelmunder and Roger Bunce operating at Mussel Rock. When the *President Taft* ended this first HF test voyage in February of 1929 at Seattle, the results were rather mixed; sometimes communication had been good and at other times nonexistent. Roger Bunce traveled north from Mussel Rock and joined engineer Neil Brown in Seattle to make changes to improve the coupling of the HF transmitter to the ship's existing MF antenna.

The *President Taft* re-embarked in March, 1929, on the "Reverse Horseshoe Run" from Seattle, across the Pacific, and then back to San Francisco. She carried three radio operators: Bob (Pop) West, Harold Van Wegan, and Roger Bunce. The aim was to make hourly contact with 6XBB, 24 hours per day, on HF frequencies between 6 and 18 MHz. This second test run on the *President Taft* was a resounding success. No hourly schedules were missed, except for a few between Shanghai and Manila, due to high noise levels.

The next Dollar ship to be outfitted with HF radio equipment was the *President Polk* (KDOZ), with operator Fred Roebuck (FD), on a round-the-world cruise. Again, the use of HF proved highly reliable as the *President Polk* communicated at almost all locations with 6XBB and KUP, except during a few days in the Indian Ocean.

By the time the *President Polk* returned from its first round-the-world cruise using HF equipment, the successes of the two "president" ships' tests had induced Dollar and Heintz to go ahead with plans to equip a shore station in Manila to provide better coverage in the Southern Hemisphere. The HF equipment bound for Manila was already at sea before the *President Polk* returned from her first round-the-world HF test.

A fly in the ointment

This flurry of communication activity by Dollar and Heintz and Kaufman did not go unnoticed by the existing radiogram and cable companies. They could easily see that their revenues would soon be adversely affected by the

Dollar move to set up its own HF radio communication company. As a result, RCA and Western Electric refused to sell Heintz and Kaufman any more transmitting tubes, invoking their patent notices on the side of each tube carton "for experimental use only." (Some Western Electric 211As even had the patent numbers printed on the glass tube envelope.)

As Ralph Heintz points out in his U.C. Berkeley interviews with Arthur Norberg, H&K *could* have purchased tubes from England or France, but using them in transmitters in the U.S. would have still infringed on patents held by RCA, Western Electric, and other members of the Radio Trust.[1]

The management of H&K decided the company would fabricate its own tubes for transmitters aboard Dollar ships and at Dollaradio shore stations. The H&K Gammatron would have *no* grid, just another plate (nearer the filament than the output plate, or anode). The control-plate was named by H&K employee Jim Brown (W6AY), probably after the third letter of the Greek alphabet. Having solved the transmitting tube patent problems, the remaining hurdle was the pair of DeForest basic triode oscillator patents, controlled by the Radio Trust.

A Seattle company, Simpson Radio, had a valid patent on another form of triode vacuum-tube oscillator, using two tubes in push-pull. Ralph Heintz was dispatched by airplane to Seattle to talk with Howard Mason of Simpson Radio. At issue was whether the Simpson Oscillator circuit (**Figures 2A** and **B**) would function with the Gammatron tubes that H&K was going to use. Ralph Heintz examined the circuit, determined that it was capable of operating with the tubes, and immediately advised Dollar to buy the financially strapped Simpson Radio and its patents. With the Simpson oscillator circuit, and their own gridless Gammatron tubes, H&K was ready to go into production and equip all Dollar ships and Dollaradio shore stations with an HF transmitter design, the MC-201, using a two-tube power oscillator.

Changes

In 1929, Heintz and Kaufman reincorporated in Nevada as Heintz and Kaufman Ltd., with the Dollar Company as the majority stockholder (two thirds of the common stock). This provided a large injection of capital to allow expansion to meet the needs of Dollaradio and to radio-equip Dollar ships. The H&K plant moved from 219-221 Natoma Street, San Francisco, to a new, much larger facility in South San Francisco at 240 Dollar Avenue (in San Mateo County).

The new H&K plant, with its tube-fabrication section and increased level of HF equipment production, required substantial increases in staff. During the transition period, H&K took on, among others, Bill Eitel and Jack McCullough (W6UF and W6CHE), who became the principal vacuum-tube men. These two hams would later leave H&K to found Eimac, which would become one of the largest power tube manufacturers in the world.

Operators loved the note of the MC-201 operating in A2. It was easy to tune in and could be received on any HF receiver with or without a BFO. Some observant operators noticed that the filaments of the HK255s in their MC-201 transmitters would move back and forth with "key up" and "key down." This movement could not have helped the stability of the self-excited oscillator, and probably added some "chirp" to the signal. But the MC-201 performance should be judged by the standards of the era in which it was used, when frequency tolerances were loose, assigned channels wide, A2 permitted, and crystal-control not required.

By the end of 1929, Dollaradio shore stations were in operation at Mussel Rock (near San Francisco); Cypress (near Los Angeles); Sherwood, Oregon; Edmunds, Washington; Garden City, New York; Haeia Point, Hawaii; Manila, Philippines; and Shanghai, China. With these shore stations able to communicate with each other on point-to-point frequencies and with Dollar ships on marine frequencies, R. Stanley Dollar had the efficient, cost-effective communication system he'd set out to achieve. Managing this far-flung communication system became such a large job that a separate company was formed to handle all communication-related business.

Globe Wireless Ltd. was incorporated under the laws of the state of Nevada on January 24, 1930, with Jack Kaufman of H&K as Vice-President and General Manager. Globe Wireless received, by transfer, the 17 HF frequencies that had been assigned earlier to the Dollar Company and purchased, at cost, the radio equipment built from 1928 through 1930 for the Dollar Company by H&K. Globe also paid the cost of acquiring the Simpson Radio Company and its patents.

Justice Department forces concent decree on the Radio Trust

In 1932, the U.S. Justice Department forced RCA and other members of the Radio Trust (GE, AT&T, Western Electric, Westinghouse, United Fruit, etc.) to sign a consent decree. This consent decree was forced upon the Radio Trust because of numerous complaints that they were in restraint of trade because of their practice of

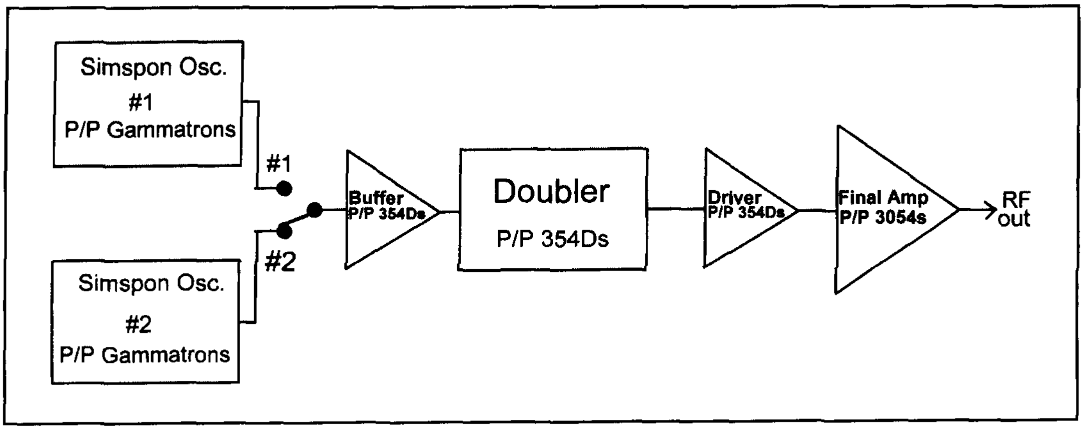


Figure 3. Heintz and Kaufman 10-kW air-cooled transmitter.

sharing their patents exclusively with each other. The consent decree also stipulated that members of the Radio Trust would divest themselves of stocks in each other's companies. As a result of this 1932 consent decree, the tube business suddenly opened up to outsiders like H&K, Taylor, and others.

No longer restricted to the gridless Gammatron-type tube, H&K began to develop more conventional types. The company retained the trademark "Gammatron"—even though the new H&K tubes had grids. The only true H&K *production* tube that was a gridless Gammatron, was the HK255—and these were essentially *all* produced for MC-201 transmitters for Dollar ships or Dollaradio shore stations.

FCC re-evaluates exclusive-use HF assignments

In late 1933, Globe Wireless was advised that the FCC had been re-evaluating its policy of granting exclusive-use HF assignments to privately owned companies like Dollar and Boeing Aircraft. It was no longer possible, in the FCC's view, for companies to own frequencies, as this was not in the public interest. (Legally, such private use of frequencies amounted to selling or giving away public domain.) So, if Dollar and Globe Wireless wanted to continue using their 17 HF assignments, they would have to do so as a common-carrier or public utility. That is, they would have to make their communication system available to the public, as did Western Union and Bell Telephone. There was simply no way around this mandate, so, on February 15, 1934, the marine facilities of Globe Wireless were opened to the public for ship-to-shore messages. On April 20, 1934, the point-to-point Globe Wireless facilities were also opened to the public.

Immediately upon becoming a common-carrier, Globe Wireless ran up against its old rivals. Western Union, AT&T, and RCA resisted helping Globe Wireless deliver its messages. So Globe adopted a mode of business called *Radio Mail*. Messages received in the U.S. by Globe stations were either sent on to their destinations by U.S. Mail or could be picked up at the Globe office. No priorities were given to any messages, and the rates were commensurably low—approximately 25 percent lower than Western Union's Night Letter service.

It seems incredible that Globe's foes couldn't understand that trying to impede the delivery of messages wouldn't hurt Globe Wireless at all. After all, Globe was in the common-carrier business *only* to preserve its Dollar steamship communications system, not to become competitive in the public message business. By adopting the *Radio Mail* method, Globe Wireless actually simplified its operations and could undercut the competition while making a profit. What Globe had first seen as a nuisance, the common-carrier business, was simplified by the *Radio Mail* way of operating and became a profitable sideline. To quote Ralph Heintz, "We started handling words by the million, literally. Because all I had to do was send it on a teletype that was downtown. The girls would tear off the message, we'd paste them up in strips and stick an airmail stamp on it. We were just flooded with business, just flooded. I got $550 for the earnings of just this one month. That just worked like a charm."[1]

Lawsuit threatened

RCA and its Radio Trust members had been threatening to sue H&K from the time they first began to set up Dollaradio with equipment for the Dollar Steamship line in 1929. The attorneys representing RCA had sent H&K letters of intent to bring suit for patent infringement of no less than 22 U.S. patents and five patent reissues, and ordered accounts for all past profits that these alleged infringements had made

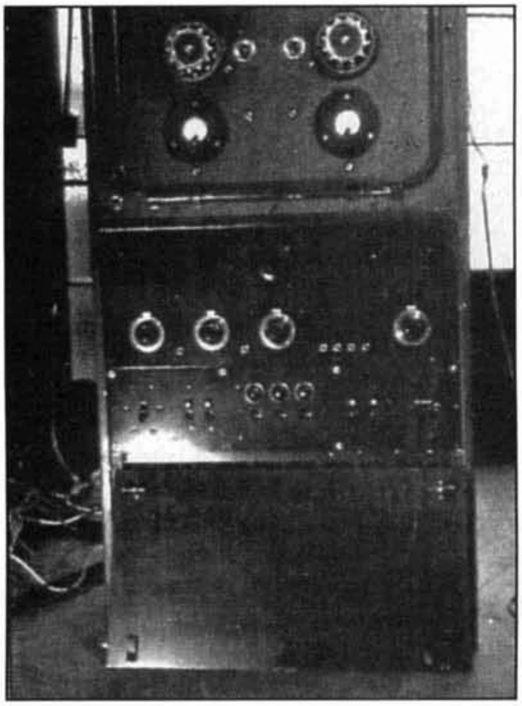


Photo E. Crystal-controlled driver and power supply of H&K A4001 (added section below final amplifier).

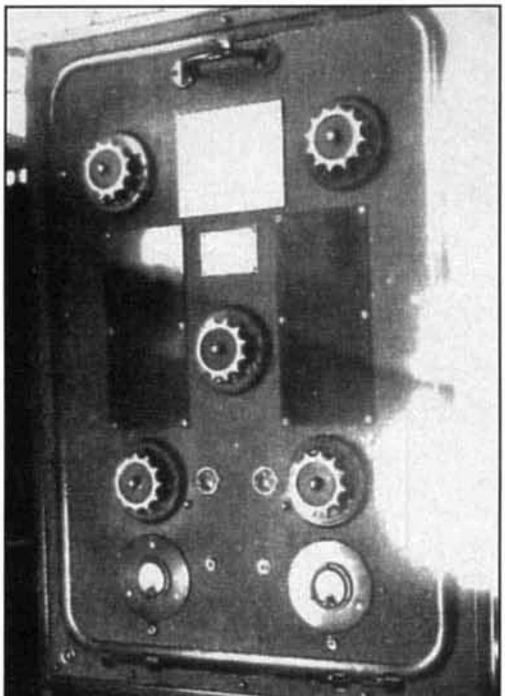


Photo F. Final amplifier section of H&K A4001 (formerly an MC-201 Simpson power oscillator transmitter).

possible. H&K simply ignored the warning letters, answering them (in effect) with "We'll see you in court."

Finally, on October 31, 1935, RCA, AT&T, and Vreeland Apparatus Company filed a Bill of Complaint against H&K. Ralph Heintz sought out Dr. Frederick E. Terman of Stanford University to be his technical witness at the upcoming trial in San Francisco Federal Court. Heintz and Terman's strategy was to attack the validity of the DeForest patents. They planned to show that three-element (triode) vacuum tube oscillators had existed prior to DeForest's stated date of conception (August 6, 1912) and that DeForest was more like the *fourth* inventor to demonstrate a triode oscillator—not the first. RCA had apparently gotten some intelligence on how H&K's defense might proceed and so had included Vreeland Apparatus as one of the complainants because Frederick K. Vreeland's triode oscillator (a two-anode, mercury-vapor magnetron type) patents had been filed for on February 28, 1905 and granted on August 28, 1906.

When court opened, one of the RCA attorneys asked for a brief recess, apparently to consider what H&K had arrayed against them for use in the trial. This included the exhibits, depositions, and expert testimony of F.E. Terman. What I think happened during this recess is that the RCA lawyers and technical experts realized that the H&K "defense" was really a very sophisticated "offense" designed, not just to ward off a patent infringement case, but to call into question the very basis of the DeForest triode oscillator patents. Perhaps H&K could even get the RCA-controlled patents declared invalid, by reason of prior art, after 12 years of possession. Better to license H&K, drop the infringement suit, and enjoy another five years of patent control than to risk the whole ballgame. After the recess, the leading attorney for the prosecution asked that RCA be allowed to withdraw its complaint against H&K—Ralph Heintz had won!

After the trial, RCA Vice President Ralph Beale flew to San Francisco with RCA's attorney, Otto Shirer, to settle the case with Heintz. They offered H&K the same agreement RCA had made with GE, Westinghouse, and others, covering the use of RCA patents for a minor royalty. H&K accepted the agreement.

The HK920

Early in 1935, and *before* the filing of the Bill of Complaint against H&K by RCA, AT&T, and Vreeland Apparatus on October 31, 1935, H&K had completed a design for a new 10-kW transmitter for use in its shore stations—the HK920. It took advantage of the newer conventional triodes H&K had developed after the Consent Decree had been forced on the Radio Trust in 1932. The HK920 was an entirely air-cooled tube system using the largest of the H&K tubes, HK3054s, in push-pull in its final amplifier. The block diagram appears in **Figure 3**.

I have a copy of the HK920 transmitter manual, with all circuit diagrams and blueprints. It seems some deceptive practices were used in printing the manual, probably to catch anyone trying to duplicate it for purposes of the upcoming trial. (Such "mistakes" are often used in circuit diagrams for electronics units made for in-house use, by manufacturers, so unauthorized copies of these circuits can be quickly detected—if the copier makes any unwanted use of the circuits.) Some of the "mistakes" in the HK920 transmitter circuits were even perpetuated in a new series of blueprints made in 1938 at Globe Wireless.

Apparently only about five of the 10-kW type HK920 transmitters were ever produced—enough for most of the Globe Wireless shore stations. One ended up at KUP, Hearst's new transmitting site on the San Francisco Bay mud flats of Redwood City.

The HK920, being all air-cooled and of multikilowatt capability, used a lot of blown-air for cooling its (large) final amplifier tubes. The KUP staff affectionately called it the "Wind-Jammer." The pair of HK3054 tubes in its final amplifier were some of the largest glass tubes ever made in the U.S., measuring 30.75 inches long and 9 inches in maximum diameter. The final amplifier had photocells to detect "color" on either of the HK3054 anodes (caused by excessive plate dissipation), which would trip out the HV power supply circuit-breaker. On occasion, when troubleshooting, the final amplifier rear door would be open and the sunlight would shine in a window at just the right angle to strike one of these photocells, and the HK920 would be tripped off the air. This odd "failure" mode puzzled the engineers, until the cause was ferreted out.

Looking back at the period 1928 to 1936, it seems that H&K, with Ralph Heintz as its technical leader, had spent the better part of a decade not just getting a global communications company up and running, but also exhausting every conceivable way to get around the DeForest triode oscillator patents of the "Radio Trust." In so doing, H&K had reinvented the transmitting tube and created at least two transmitting tube companies on the west coast that would *never* let RCA and AT&T regain their lock on that market. Beating RCA et al. at the trial of 1936 was Ralph Heintz in his ultimate victory with H&K; and, having done so, he began to chafe against the internal restrictions imposed by Dollar Steamship Co., rather than those from outside.

Enter the HK354 triode

In 1934, H&K had developed the HK354 triode, once the restrictions of the Radio Trust had been removed by the 1932 Consent Decree. The HK354 seemed to be an ideal 150-watt triode for the amateur radio market. This was no surprise, as nearly everyone at H&K was a ham, including Heintz, Eitel, and McCullough. So H&K began a ham marketing strategy for the HK354, with a ham transmitter construction article in *Radio* magazine (January 1934) and ads in both *Radio* and in the soon-to-be-released 1936 edition of *Radio Handbook.*

Before the ham marketing program got halfway off the ground, R. Stanley Dollar put the kibosh on it, with his edict that H&K's primary and *only* mission was to support Globe Wireless. Because R. Stanley was majority stockholder in H&K, he apparently could effect such micromanagement, but the move caused widespread loss of morale among company employees. Within a few months of Dollar's imposition of the non-ham marketing decision, both Eitel and McCullough quit H&K to start their own tube company in San Bruno called Eitel and McCullough (Eimac). Their first tube, for sale to the ham community, was the 150T, a close copy of the HK354, but in a differently shaped envelope.

Heintz leaves H&K

It is to Ralph Heintz's credit that he was able to keep enough of the H&K group together and put out the tremendous efforts that finally defeated the Radio Trust in the patent infringement trial in late 1936. But after the victory, tired of the restrictive policies of Dollar, Ralph Heintz sought out new venues for technical achievement. Heintz left H&K in 1937 to go east to work for Eclipse-Pioneer, a division of Bendix, designing and building 400-Hz, three-phase aircraft power systems. You must understand that Ralph Heintz was a true renaissance man, technically, and was as comfortable working on motors, refrigerators, searchlights, or surgical equipment as he was at radio and vacuum tubes. After a two-year period with Eclipse-Pioneer, Heintz left that firm to start Jack & Heintz with Bill Jack, an ex-president of Pump Engineering Service Company (PESCO), a division of Borg-Warner in Palo Alto, California. He came up with a new design of aircraft starter that was superior to the ones he'd seen at Eclipse-Pioneer. After some local union troubles, Ralph Heintz and Bill Jack moved to Jack's home town of Cleveland, Ohio, and built a company that eventually would employ thousands, making aircraft-starters, bomb sights, auto pilots, and all sorts of electro-mechanical devices required during World War II.

After Ralph Heintz left H&K in 1937, the company continued its work producing tubes

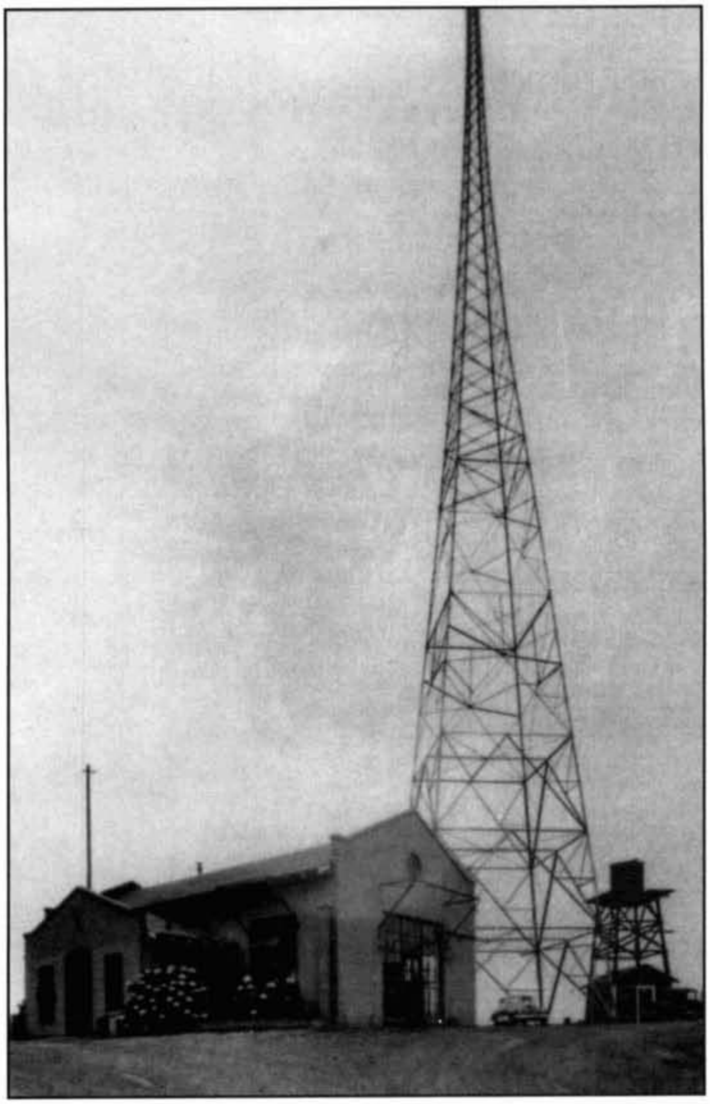


Photo G. Mussel Rock tranmitter site.

and other radio equipment for Globe Wireless and the communication needs of Dollar Steamship Line. Apparently, Dollar Steamship Co. had a change of heart about H&K selling its tubes on the open market, and placed an ad in the 1940 *Radio Handbook*. Note that H&K by this time had a fairly complete line of triodes as well as a pentode transmitting tube (HK257) that required low drive and minimal neutralization even at the high end of the HF band. These tubes were being designed into the newer Globe Wireless transmitters as they became available, along with circuit techniques like crystal-control to update the existing MC-201s or replace them with completely new designs that used similar mechanical configurations. (H&K transmitters were built like battleships, and the special castings and other metal work were better re-used than scrapped.)

The MC-201s eventually were either modified to become, or were replaced by, the new Globe Wireless A-4001 transmitters, the driver of which is shown in **Photos E** and **F**. Note that this transmitter exciter used a crystal-controlled oscillator, several RCA-type tubes, electronic keying, and an HK257 pentode that required no neutralization. The blueprint is dated November 11, 1941, and initialed (H.O.S.) by Hans Otto Storm, an innovative young electrical engineer from Stanford, whose designs at Globe Wireless ran the gamut, from antennas through remote site receiver tuning to transmitter design. His talents also included authoring several books, mostly on social justice themes.

The blueprint of the A-4001 exciter may have been one of the last prints that Hans Otto Storm ever initialed. He was electrocuted while working on one of the Globe transmitters at the Globe laboratory facility on Drumm Street in San Francisco.

Operation of Globe Wireless

Now let's get back to the operation of Globe Wireless, the communication company. The Mussel Rock (MR) facility was the Dollar Steamship's home station. Mussel Rock started as a single 15 by 10 foot wooden shack in 1928—large enough for one operator and his radio equipment. As Globe Wireless expanded, a second shack was added about 100 yards away. One housed the receivers and operating position, and the other housed the transmitters. The 100-yard separation was a first attempt to prevent transmitter energy from causing overload and damage to the receivers if both stations were in operation simultaneously, as when "break-in" operation is desired. Although, the 100-yard distance helped matters, it wasn't sufficient separation for duplex operation, where transmitter and receiver are *intentionally* on at the same time. Also, as Globe's operation became larger, the MR site had multiple operators operating separate transmitters and receivers on different frequencies, from 436 kHz to 23 MHz. These uncorrelated operations could also cause mutual interference. The amount and seriousness of the interference between MF and HF channels depends on many factors: antenna proximity, power level, class of transmitter power amplifier operation, receiver dynamic range, transmit antenna VSWR, frequencies used, ad nauseum. The fixes for this mutual interference are many and varied, but the one *direct* way to solve the problem is to *put some distance between transmitter and receiver sites!*

The second configuration of the MR station was to have the receiver site at the beach (south of the actual landmark Mussel Rock) and the transmitter site up the hill about one quarter mile away—each site with its own building and "antenna farm." **Photos G** and **H** show this configuration of receiving and transmitting sites at MR. The two buildings were of hollow ceramic block construction, designed by Ralph Heintz himself (and replicated at Globe stations around the world). The transmitter site had two steel towers, each over 200 feet tall, to hold up the KTK Marine Band off-center tapped MF antenna so it could be resonant at both 436 and 500 kHz. MF was used for marine communication only, but was a necessity then.

The beach receiver and uphill transmitter arrangement was a great improvement over the original "all-on-the-beach" arrangement, in terms of mutual interference. But when full duplex operation became the rule rather than the exception, the separation had to be increased even further. The answer was to leave the transmitter site at its Mussel Rock uphill location and build a new receiver site at Cahill Ridge at the top of the coast range that separates San Francisco Bay from the Pacific Ocean. The site chosen was near the present junction of State Highways 92 and 35, near Skylawn Cemetery and close to where KUP had a receiving site on San Francisco Water District (Hetch Hetchy) land, just to the east. It was known as SK (for Skyline) by Globe Wireless operators. The SK location gave Globe Wireless a receiving site with a spectacular panorama to the Pacific Ocean, and it put some steep hills between itself and MR. Once again, KUP had become Globe's "pathfinder" in locating a new site. KUP was able to build its receiving site east of the skyline summit on San Francisco Water Company property (because Hearst and the Mayor of San Francisco were political allies), but Globe had to purchase farm land on the west side of the San Francisco Water Company Preserve. As it happened, when Globe sold out in 1960, the new user of the site needed VHF and HF coverage to the west, so SK was ideally situated at the 1100-foot level with a clear view towards Hawaii. The present VHF Curtain array of ARINC (Aeronautical Radio Incorporated) is able to communicate with commercial airplanes hundreds of miles out to sea on the 131.6-MHz aircraft frequency; and, during occasional favorable conditions, all the way to Hawaii, using only 20 watts.

With the final arrangement of locating the transmitters near Mussel Rock (MR) and the receivers on Cahill Ridge (SK) after 1938, the physical setup of KTK (and its point-to-point calls, KGL, etc.) didn't change for the rest of the lifetime of Dollar's Globe Wireless. Both MR and SK continued to handle ship-to-shore and point-to-point traffic as the San Francisco terminus of Globe Wireless' network of MF and HF stations. Although technically a common-carrier, or a public utility that handled messages for the public via the Radio Mail service, Globe Wireless remained, at its core, a communication company for the handling of Dollar Steamship traffic.

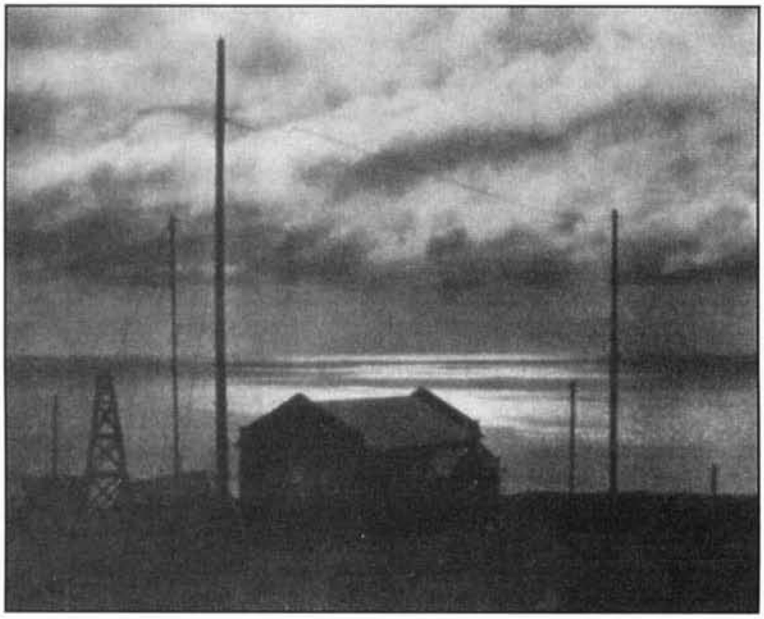


Photo H. Mussel Rock receiving site (at Salada Beach).

War

At 10:50 a.m. on December 7, 1941, KTK, and its competing marine stations KFS and KPH, heard the frantic SSSS and SOS signals from the wooden-hulled steamer Cynthia Olson, 750 miles southwest of Seattle (Submarine Sighted, Help). The SS Lurline (KIEK) answered the call first, asking: "is the sub surfaced?" The Cynthia Olson (call unknown) answered "R (yes), WAIT, SSSS, SOS, SOS, SOS, POSN...." That was the last signal heard from the Cynthia Olson and one of the very first signals that World War II was about to begin.

Both KTK and KPH were ordered off the air that day by the U.S. Navy, as was KFS, temporarily. Globe Wireless (KTK) and RCA Marine Communications (KPH) were shut down for the duration. Station KFS was taken over by the U.S. Coast Guard for military use. The story of KFS and some of its World War II experiences with Philippine Guerrillas are very interesting reading.[1]

The operators from Globe Wireless were of great assistance to the war effort, manning various radio operator positions on Merchant Marine ships. The Globe Wireless equipment facility, on Drumm Street in San Francisco, expanded to perform an increasing load of war-related radio equipment work—just as H&K

The Dixon Relay Station

Yesterday...

The history of the Dixon Relay Station (**Photos I and J**) goes back more than 50 years. When the Office of War Information (OWI) initiated shortwave radio broadcasting during the World War II, it turned to commercial broadcasters to fill the need for facilities and equipment and to contract operators to staff the stations.

In order to broadcast to target audiences in Asia and the Pacific, two identical stations were built in California—one at Dixon, near Sacramento, and the other at Delano, near Bakersfield. Construction was started for the 640-acre radio transmitting facility at Dixon in 1943. The Dixon station first went on the air on December 27, 1944, as the Voice of America and was operated by the National Broadcasting Company (NCB) under contract to the U.S. Government's Office of War Information (OWI). The Delano station was operated in a smiliar fashion by the Columbia broadcasting System (CBS). Federal government operation of Dixon was assumed on November 1, 1963.

The original complement of the station was four RCA 50-kilowatt transmitters and two Federal Telegraph Company modulators. Each modulator fed two of the transmitters. Federal later built a 200-kilowatt transmitter for Dixon. While controlled by NBC, the transmitters at Dixon were operated under the callsigns of KNBA, KNBC, KNBI, and KNBX. The original antennas were rhombics and the target areas were Japan, Australia, and the Philippines.

About 1952, two General Electric G-100C transmitters, 100 kilowatts each, were installed. A major modernization was undertaken in 1965. The building was expanded at the front and three Collins 821A-a transmitters, 250 kilowatts each, were installed.

The station temporarily ceased broadcasting in August, 1979, and remained in a caretaker status until October, 1983, when it resumed operation. During this period of operation the station's primary function was to provide Spanish language programming to Central America. Broadcasts were again suspended in April, 1988, as the result of reductions in operating budgets.

On September 30, 1993, the Voice of America relinquished its interest in the 800 acres of land, the buildings, the antennas, and the skeletons of five high-powered shortwave transmitters. The property was put up for sale by the San Francisco office of the General Service Administration and ultimately sold to a private owner. The decommissioning of Dixon

Photo I. Entrance to the former VOA Dixon Relay Station. The main building is visible in the background, to the right of the sign.

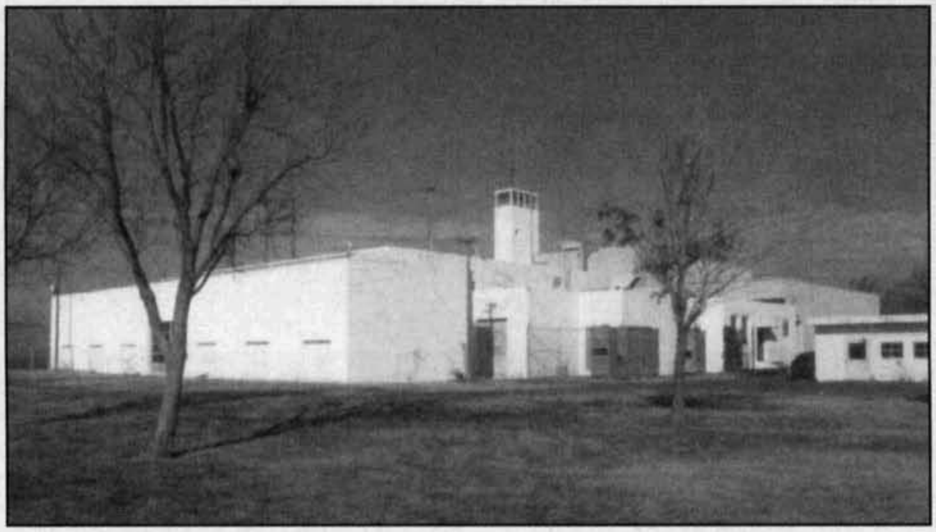


Photo J. Rear side view of the main building of the former VOA station. The tower on the building was apparently a security feature.

brought to a close a 50-year-long chapter in the history of international shortwave broadcasting by the United States Government.

...and today

The story of the Dixon facility itself, however, is still being written. Recently, Globe Wireless, a maritime communications service provider, acquired the former Voice of America transmitting site. The company intends to use the station to connect vessels in the Pacific Ocean with land-based electronic systems, including the Internet.

Still remaining on the site from the days of shortwave broadcasting are two massive dipole curtain arrays (**Photo K**) and 10 rhombic antennas, most still operational. Skeletons of the GE and Collins transmitters also remain. Globe Wireless plans to install transmitters and antennas at Dixon for its maritime public coast station KFS. The current KFS transmitter location, in Palo Alto, California, will be phased out of operation over the next few years.

According to company officials, Globe Wireless may also relocate the transmitters for public coast station KPH to the new Dixon location. Transfer of that station's license to Globe Wireless from MCI International is pending FCC approval. The MCI station currenly transmits from Bolinas, California.

Aeronautical Radio, Inc. (ARINC) will sublease space at the Dixon site from Globe Wireless. ARINC is installing transmitters that will also allow it to communicate with the flight crews of aircraft flying over the Pacific Ocean and South America.

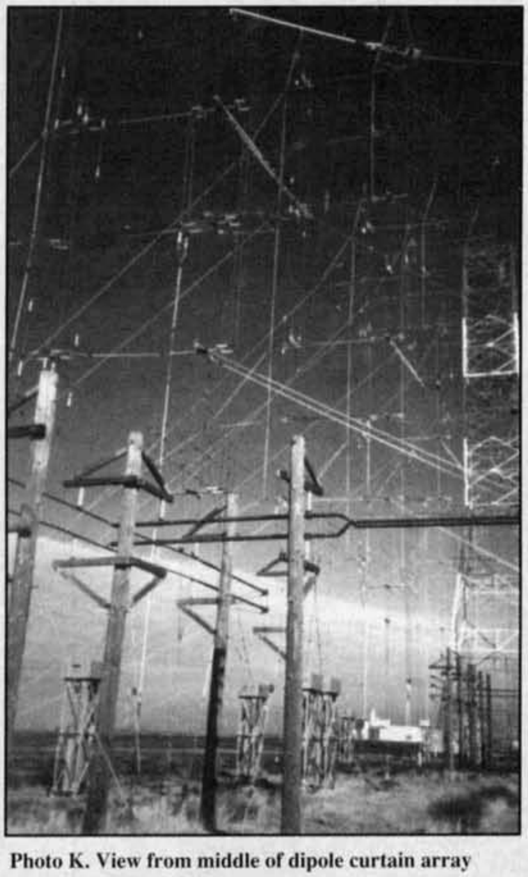


Photo K. View from middle of dipole curtain array showing details of feed system.

Globe Wireless® operates the Global Radio Network™ of public coast stations, including A9M in Bahrain, KEJ in Hawaii, KFS in California, SAB in Sweden, VCT in Newfoundland, WNU in Louisiana and ZLA in New Zealand.

went into wartime tube production, three shifts a day. But during the war years, the signals of Globe Wireless at MR were only a memory, although the site was used by the U.S. Army as an Army shortwave communications site.

Post-war activities

After the war, all three MF/HF marine stations in the San Francisco area were reinstated to their pre-war status. Station KFS reverted from U.S. Coast Guard operation to MacKay Radio, KPH was on the air again as RCA Marine Communications, and KTK resumed operation as Globe Wireless. The return to pre-war status wasn't as automatic as you might think. Except for the fact that HF and MF were still used, nearly everything had to be redone: frequency allocations, modes of transmission, transmitters, receivers, and antennas. Some tasks were easy to accomplish, such as replacing tired old transmitters from the 1930s with newer World War II surplus models. Transmitters made by Press Wireless were installed at the MR site. They included PW15s, a PW 20/40, and a U.S. Navy TEB (similar to a PW15). National HROs became the standard receiver at the SK receiving site. In all the resurrection work, good advantage was taken of World War II surplus at bargain prices.

CW remained the mode of communication (especially with ships), but the war had brought the widespread use of automatic mode transmission, like teletype, using the old five-level Baudot code. Globe also used a nonstandard mode of automatic communication, developed by IBM—the Radiotype. The Radiotype was used throughout the Globe system, in addition to conventional 60 word-per-minute teletype. Although it was capable of functioning at 100 words per minute, Radiotype was different enough from the more standard teletype system that it was always considered "odd-ball" by operators and technicians. Teletype and Radiotype were used extensively for messages between HF stations and offices. Information was transmitted via phone lines, as neither mode was error-correcting. The automatic systems were used on HF only when the circuits were "solid." CW was used at HF when circuits were marginal because of the operators' ability to correct errors. After World War II, frequency shift keying (FSK) became the standard mode of transmission when operating Teletype or (in Globe's case) Radiotype. FSK on HF started out with 850-Hz shifts between the mark and space tones and narrowed over the years as channel-assignments got tighter and better frequency stability requirements were called for by the FCC, down to 170-Hz shift.

Globe Wireless carried on from after World War II until 1960 pretty much the same as it had prior to December 1941; however, its paid point-to-point traffic was on the decrease, as newer modes of point-to-point communication became competitive. Over land, microwave relay systems were the competition to HF point-to-point. Undersea, newer wideband cables were laid with in-line amplifiers that allowed multiple CW signals and voice to be passed. Finally, satellite systems began to really chew into HF communication revenues, with such phenomenal bandwidth that multiple voice, and even television, signals could be carried to any part of the world.

Globe Wireless' competitors in point-to-point even went to multiplex (MUX) systems used directly on HF point-to-point circuits (such as the system made by Siemens Halske); but it was a losing fight, to try to compete with microwave relay, newer undersea cables, and satellite modes. These competitive systems had a real advantage: they didn't depend on the variability of the ionosphere. The militaries of the world realized that HF radio *could* become their only means of global communications should a large-scale war break out (which meant that undersea cables and satellites, and perhaps even the microwave relay systems, could instantly become unusable). However, commercial communications companies, like Globe Wireless, couldn't keep propping up their HF point-to-point systems against the possibility of world war. After all, if war occurred, they'd be out of business anyway.

Globe is sold

In 1960, Globe Wireless was sold to ITT/MacKay Radio—their competitors. The point-to-point business was essentially gone, and the fortunes of the Dollar Steamship line had turned sour. It no longer made any sense to continue operating the marine portion of the business for Dollar's exclusive use. The stations were all closed down. Some of the operators were transferred to ITT/MacKay to continue marine operation, and much of the Globe Wireless equipment was transferred to the ITT/MacKay stations.

At the San Francisco terminus (MR), the transmitter building was emptied of its PW15 transmitters, which were trucked to the KFS transmitter site at Palo Alto. The PW20/40 was donated to a religious broadcaster for use overseas (where it wouldn't have to meet FCC-type approval). The "new" PW15s and TEB (Navy version) replaced the former complement of Federal transmitters at the ITT Palo Alto marsh transmitting site (MX), and all the old cooling towers were pulled down as the MX site became an air-cooled transmitter facility.

The towers at MR were simply felled and sold for scrap, and every trace of the uphill transmitter site was scoured away as a land developer bulldozed the hillside into a site for a housing tract. The lower MR site, formerly the receiver building before it moved to SK, was sold to an architect and was converted into a very nice beach home on Palmetto Drive, Pacifica. It remains a residence.

The Skyline receiving site (SK) was emptied of its National HRO, receivers and other equipment. These were moved to the ITT receiving site near Half Moon Bay (LO), to be set up alongside the Hammarlund SP600JX receivers of KFS. For a few years, KTK marine continued operating out of LO and MX as a marine station—the only revenue-producing activity of either KFS or KTK. The SK site was occupied by ARINC for use in HF/VHF communication with commercial aircraft. It continues as such to this day, on Cahill Ridge not far from cellular telephone and paging system block houses, with their familiar monopole arrays of UHF antennas.

World shipping trade changes

As the marine business decreased due to several factors (fewer U.S. ships in the world shipping trade and the increasing use of direct ship-to-shore satellite communications systems), KTK was discontinued in September 1964 (its call letters were reassigned to an FM broadcast station). ITT/MacKay was downsized to the KFS station (receiving at LO, transmitting at MX). Even the property at MX was sold to the city of Palo Alto for a nature preserve; KFS retained only a right-of-way for operation of its marine transmitters and antennas for what was expected to be a short-term period. The expected life of MF/HF marine radio in 1990 was estimated to be a decade or less, after which satellite systems would take over long-range communications for ships, and VHF voice communications would take over coastal, near-land use. These predictions didn't come true, as world shipping moved away from U.S. flagged ships with their expensive union crews. The new trend was toward foreign-flagged ships, with less stringent safety rules and foreign crews—many of whom had "papers" from countries with diploma-mill-type Maritime Academies. It doesn't take much arithmetic to figure out that it's not as profitable to ship a boatload of copra from the Philippines to some U.S. soap company using a union crew versus sending the same copra on some old "rust-bucket" with an all foreign crew and only simple CW MF/HF equipment aboard. Either way, the copra usually gets to the U.S. port, but in the latter case it can be much cheaper.

The net result is that there continues to be a constant volume of HF/MF ship-to-shore traffic worldwide, as foreign flags-of-convenience increase using older radio-equipped ships and low-paid foreign crews. The pressure to operate shipping via this cheap mode is immense, as shipping companies continually try to find ways to reduce expenses. Every now and again these cost-cutting practices result in a serious shipping disaster, as when a tanker takes an illegal short-cut on its way from the North Sea oil fields to the open water of the Atlantic Ocean to save even more money by conserving fuel. Only then does the general public become aware that the laws of sea-shipping and the practices of sea-shipping are not always the same.

Major world powers like the U.S., Britain, and Japan continue to insist on high standards for ship seaworthiness, high-tech communication equipment, and well-paid crews. To some degree, they are able to do this and still garner a reasonable share of the market for certain kinds of shipping—such as Japan's successes in vertically integrating their auto delivery ship system into the automaker's production and distribution system. But for *most* shipping, the major world countries are fighting a losing battle to keep their ships busy, because they have to compete with an amorphous conglomeration of ships that don't play by the same rules. The remedy for this seemingly unfair shipping competition is for major world powers, like the U.S., to attempt to have more exacting laws of the sea written to level the playing field. Unfortunately, this legislation takes time to pass in a World Court that's politically diverse and basically hostile to regulation.

The upshot is that the world shipping scene will see radio communications laws and practices remain pretty much as they are today, unless there's some powerful incentive to see them change. Such incentives could take the form of refusal of maritime insurers to write policies for ships not meeting international standards. But the traditional carriers of maritime insurance, like Lloyd's of London, are at their weakest state in history due to World Court policies of unlimited liability and mega-catastrophes, like that of the Exxon Valdez. In short, the use of MF/HF maritime communication will probably remain at present levels, or even increase with time, and there will be a continued need for MF/HF shore stations in the foreseeable future.

KFS World Communications resurrects Globe Wireless

To put itself in a position to serve the needs of the world's shipping fleet, KFS World Communications, the surviving single station company of what was the MacKay Radio System, has changed its name to Globe Wireless. Since it was MacKay that bought Globe Wireless in 1960, and continued KTK on the air until 1964, and still uses some of the transmitters from that station, it seems a logical name change—in view of the new expansion plans for this company. In 1993, KFS World Communications bought Slidell Radio/WNU near New Orleans from Tropical Radio, changing its operational mode in order to control it remotely from the KFS receiving site at Half Moon Bay, California.

Next to be added to the KFS World Communication network was VCT in Newfoundland, Canada. At this point, the rather restrictive title of KFS World Communications was thought too parochial, and Globe Wireless was chosen as the new radio system name. This title became official as of January 1995, as announcements were made that not only KFS, WNU, and VCT were part of the new Globe Wireless, but that three new stations, ZLA in New Zealand, KEJ in Hawaii, and SAB in Sweden, would soon be added.

Globe Wireless now offers a broad spectrum of communications services to ships at sea. These include CW on MF and HF, SITOR on HF, Inmarsat and an error-correcting digital system called CLOVER. CLOVER allows ships equipped with such computer-controlled HF equipment to run their digital communication with shore at speeds up to 10 times that of SITOR. Like SITOR, the basic HF transmitter and receiver on board needn't be replaced, but rather a modification is added for digital modulation and demodulation. Globe Wireless intends to explore converting existing ships' HF equipment for CLOVER compatibility.

This sort of "customer development" is very much like what United Wireless did in the period prior to 1912, outfitting ships with spark wireless equipment to develop a market for its shore stations. United Wireless was acquired (after a destructive patent suit) by American Marconi, and its 70 shore stations and 500 shipboard stations were taken over in 1912. That aggregation became part of RCA. How will the present-day remnants of RCA Marine (now MCI), and its competing HF marine system (KPH/WCC), react to the new Globe Wireless expansion? The next decade of ship-to-shore communication history may prove as interesting as was the first decade of this century.

Acknowlegements

The story of Globe Wireless would not have been written without the help and encouragement of many people. I would like to thank Bill Ayers, W6VQ; William A. Brenniman, W6JU; Roger Bunce, W6EFT; Rod Deacon, NR7E; Alan Douglas; Bud Hall, K2LP; Ralph Heintz, Jr.; Carl Hoeck, WA6NQV; Rich Hoeck, W6RZL; Al Jones, K6DIA; Ken E. Jones; Paul Letsinger, W6SYL; Ron Martin, W6ZF; Jim Maxwell, W6CF; Craig McCartney, ; Bill Orr, W6SAI; George Parks, W6AOF; Rex Patterson, W6VJJ; Ed Prather, W6GXF, and George Turner. ■

References

1. H.D. Olson and William Orr, "The KFS-Federal-MacKay Store, from CW Arc to Silicon Valley," *AWA Review 1/8*, September 1993, pages 75–107.
2. R.M. Heintz, KFUH, "Somewhere East of Suez," *QST*, November 15, 1925, pages 15–20.
3. David Atkin, "South Seas Saga," *World Radio*, February 1985.
4. Thorne Mayes, *Wireless Communication in the United States*, Chapter 5, "Operating Companies," Section K, "Globe Wireless Telegraph Co.," pages 84–199. Published by New England Wireless & Steam Museum, Inc., East Greenwich, Rhode Island, 02818.
5. Ralph M. Heintz, *Technical Innovation and Business in the Bay Area*, an interview by Arthur L. Norberg, Bancroft Library, University of California, Berkeley, June 2, 1977.
6. Case File 3945L, National Archives and Records Administration, 1000 Commodore Drive, San Bruno, California.
7. F.E. Terman Papers, Heintz and Kaufman File Box, Series V, 6-1 Patent Interference Case, Stanford University Libraries, Department of Special Collections.
8. Thorne Mayes, "Seven Diagrams Covering the Exhibits in the Foothills College Electronic Museum That Were Prepared by Ralph Heintz for the Trail of RCA vs. Heintz and Kaufman," June 1976.
9. *Klystron Technical Manual*, Sperry Gyroscope Company, 1944.
10. A.H. Halloran, "Farnsworth's Cold-Cathode Electron Multiplier Tube Uses Neither Grid Nor Filament," *Radio Magazine*, October 1934, pages 6, 14, and 15.

In Memorium

Henry "Hank" Olson, W6GXN, a radio physicist, research engineer, and prolific writer of amateur and other technical articles, died March 9, 1996. He was 64 years old.

A resident of Menlo Park, California, for nearly all his life, he earned his undergraduate and master's degrees from Stanford University, where he studied radio engineering under Dr. F.E. Tennan. He also worked as a research assistant for the Radio Propagation Laboratory under Dr. "Mike" Villard, W6QYT, and Dr. Allen Peterson, ex-W6POH, with Dr. Leonard Fuller as laboratory administrator.

After serving two years at the U.S. Army Proving Ground at Ft. Huachuca, Arizona, Olson returned to Stanford to work on HF backscatter radar for two years during the International Geophysical Year (IGY). In 1958, he transferred to Stanford Research Institute where he spent 28 years engaged in work with the Radio Physics Laboratory. Subsequently, he taught night classes in electronics at Foothill College. A dedicated and energetic man, he was also involved in the Boy Scouts of America, the Institute of Radio Engineers, the American Radio Relay League, the Standford Amateur Radio Club, and the Society of Wireless Pioneers.

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WRITING THE AMATEUR RADIO ARTICLE

Are you ready to give it a try?

It's not easy to put out a magazine unless top-quality editorial material is available. This editorial material is most often provided by freelance writers; that is, writers who aren't on the staff of the magazine. That freelance writer could be you. To help you get started, I'm going to share some of what I've learned about technical writing in the more than 29 years since *QST* accepted my first article.

Let's get one thing straight right away. Technical writing isn't magic, it's a skill—and it's a skill that can be learned by almost anyone who has the brains to read an article in this magazine. It's not an arcane art practiced by some specially talented elite; it's a skill that can be developed. You don't even need to be an English major to do a good job!

Slant

Not all articles are suited to all publications. The difference lies in what the trade calls "slant." This term refers to the point of view the author takes in his article in relation to the type of readers who buy the magazine. The same topic can sometimes be sold to several different magazines, provided the slant is different, the timing is different, and, generally, that the magazines aren't in competition with each other.

The slant is merely your effort to aim the article at the readers of a particular magazine. Every editor can tell you something about his readers. You can also get a good feel for the readership by studying recent issues of the magazine. For example, *Communications Quarterly* magazine caters to technically literate ham radio operators, so don't send Terry, KA1STC, an article on DXpeditions or contesting (other magazines, such as *CQ Contest*, use those types of articles). Using the information you've gathered, you can home in on the types of articles the magazine will buy. Although a few editors issue "want lists," most of them will tell you that they don't know what they want, but "I'll recognize it when I see it."

It's difficult to predict what an editor will *never* buy, but there are some guidelines you can deduce from perusing the magazine content for a few months. Of course, even a "normally no-no" article might find a home if it's uniquely presented, has a special focus, or is especially well written.

Another "market survey" test is to make sure that the same magazine hasn't published a similar article in the last couple of years. Some magazines won't revisit a topic for as many as five years, so if you find a piece within that period that's close to your topic, change the focus or slant of your idea to make it unique.

Now that you've identified a topic, you sit down, dash it off on the word processor, and mail it in, right? Not quite. All editors get "over the transom" (unrequested) articles, and will look at them (if they are smart). But editors have one thing in common no matter what the topic or type of magazine: they're very busy people! It takes a lot of time to properly evaluate an article, more time than they may have for months to come. The way to get the editor's attention is to send a query letter or e-mail. (Terry's e-mail address is <ka1stc@aol.com>.) Your letter should be short (one page), to the point, tell the editor what the topic is and how you'll approach it, and then politely ask for a reply. If you use e-mail, make sure the e-mail address is embedded in the message (not all e-mail services print the address header).

Does a positive response to a query letter mean the article is sold? No, not by a long shot. You'll still be submitting the piece on speculation. But it does mean that the editor is interested, and that there's a high probability of sale. If you write the article as specified, then it's likely—but not guaranteed—that you'll sell it.

Article format

There are several types of articles that appear in ham magazines. Some are tutorial pieces, some are experiments or construction projects, some are DX field-trip narratives, and some are "how-to" pieces. Let's take a look at the basics of the how-to article. The one common denominator for all such articles is that they offer practical instruction and advice. This definition covers a lot of territory, including most practical technical articles. There's no fixed universal format for all how-to pieces, and almost any format will work some of the time.

However, there is one format that almost always works, so new writers might want to follow it until they get a little experience. That format, which I learned at a writer's conference, is called the "Tell-'Em-Cubed" method, and follows these steps:

1. Tell them what you're going to tell them.
2. Tell them.
3. Tell them what you told them.

The first "tell them" should take up no more than about three paragraphs, and may sometimes occupy only one short paragraph. This "tell them what you're gonna tell them" segment is the preamble that must grab—and hold—the reader's attention, and convince him to continue reading. The main body of the article is the "tell them" portion, and it should occupy the bulk of the space. Finally, in the "tell them what you told them" section, do a quick (one to three paragraphs) summing up. Use it to highlight the main points—especially those that should be remembered.

For science and technology articles, which are basically how-to pieces, there's a modified "Tell-'EM" format, which I call the "Tech Writer's Eight-Fold Way:"

1. Tell them what you're gonna tell them.
2. Tell what it's gonna do for them.
3. Tell them how it works.
4. Tell them where to get materials and how to build it.
5. Tell them how to test it.
6. Tell them how, when, and where to use it (as appropriate).
7. Tell them how to modify or adapt it for other applications.
8. Tell them what you told them.

Of course, not all of these elements need be included in every article, but this method does represent a stylistic shopping list. Nor is this the only viable format: it's one that usually works well, that's all. If you have another format and want to give it a try, then lots of luck—it could work.

Writing the piece

Most successful authors prepare at least a basic outline for the article. This road map needn't be as formal as one for an English class. It's just a guide to ensure that all bases are covered, and are covered in logical order. The outline keeps you on the right track.

Each major topic in your article deserves at least a paragraph. A major mistake made many by novice writers is to mix several topics in the same paragraph. If your outline is written to the paragraph level, however, then it's unlikely you'll fall into this trap, and the article will flow more naturally.

Another common mistake is to cover too many topics in a single article. A magazine article is a capsule of information on a specific, usually quite narrow, topic. Shortly after my initial success in *QST* in June 1968, I sent a manuscript to Jay W. Phipps, then editor of *Electronic Servicing*. Jay apparently saw something good in that mess of a manuscript, because he took the time to write a four-page bit of fatherly advice (not something one learns to expect from busy editors). He pointed out that there were at least four different articles in that one nine-page manuscript. When I finished rewriting the piece, there were actually five article topics buried there, and Jay bought all five from me just in time for me to get married (being a poor college student, I needed the money).

How long should the article be? The quick-draw response is "long enough to tell the story." While that is true, it's not the practical answer. Take a look at the articles in your target magazine. You'll notice that most of them fit into a relatively narrow range of lengths that follows their format. In general, an article should be five to 15 double-spaced, typewritten pages with two to six illustrations. Some magazines publish longer pieces, and certainly some publish shorter pieces; but, in general, those that are bought usually fall in the middle range of sizes. If you really feel strongly that an article needs a long treatment, write to the editor and make a proposal for either a long article or one with multiple parts. If the topic strikes the editor's fancy, you may get a no-obligation "speculative" go-ahead.

Preparing the manuscript

When you prepare the manuscript for your article keep in mind that a real, live, warm-blooded editor must read and work with your piece. Put the pieces together with an eye to making that job easier. I've seen a lot of potential writers over the years who would get fewer rejection slips if they did a better job of preparing the manuscript. If an author is too sloppy to do the mechanical job correctly, the editor might get the idea that he's a little sloppy with the facts, as well.

Editors require typewritten manuscripts, so don't even *think* about sending in a handwritten piece. Make sure your typewriter or computer printer is in good repair and prints well. Dot matrix submissions are accepted by most publishers, but only if they are easily readable. An editor spends a lot of time every day reading, so a washed out, low-resolution dot matrix submission might just go untouched. A "near letter quality" printer with a fresh ribbon produces an acceptable manuscript.

The manuscript should be double spaced, with one-inch margins all around. Don't attempt to get too much text onto a single sheet. It's false economy, and could cost you the sale. Type the final draft on 8.5- X 11-inch plain white #20 paper. Don't use colored paper, or paper with ruling lines on it. Also, don't use "erasable bond" paper. The erasability that appeals to you is caused by a surface coating that permits the typing to rub off on an editor. Nasty stuff!

Dot matrix printer users need to make sure that they use a high grade of continuous feed paper. The cheap stuff (which costs only a little less than the good "laser trimmed" stuff) leaves a coarse, ragged edge when you tear off the sprocket-hole tracks, and that annoys some editors. Of course, if you use a nice laser printer, the problem is solved, isn't it?

The first page of the manuscript should contain your name, address and telephone numbers (with area codes!) in the upper left-hand corner. If you have a facsimile number, then include it as well. Ditto your e-mail address. The title should be about one-fourth of the way down the page, with your byline beneath it. The byline should be written the way you want to see your name in print.

Don't send in a manuscript containing a lot of hand corrections. In general, most professional writers will retype a page if more than three minor corrections appear on it—and even then only if they use a typewriter instead of a word processor. Most editors don't mind if someone who uses a typewriter makes a few legible hand corrections, but don't overdo it.

When the manuscript is finished, bind the pages together with a single paper clip, not a staple. Also paper clip the illustrations to the text. In a technical article the pictures are as much a part of the manuscript as the text, so don't forget them. Send the manuscript flat in a large manila envelope (don't bend it over and force-fit it into a standard #10 business envelope). Make sure that there's enough postage to carry it all the way to its destination, or it'll come back to you undelivered.

Keep in mind that many articles are rejected (even when they're really good). Many editors will only return a piece to you if you include a separate self-addressed stamped manila envelope.

Illustrations

"A picture is worth a thousand words," says an old cliché. That old saw might be true in some cases, but when you're being paid on a page-rate basis, a picture is worth more like 200 words. The real value of the picture, however, is that it enhances the article and makes it easier to follow. In fact, for technical articles, the picture might make it possible for your work to be accepted in the first place. A picture, in that case, isn't worth a thousand words—it's priceless.

You don't have to submit professionally drawn illustrations. Pencil drawings are acceptable to most magazine editors, but they have to be done in a way that can be interpreted by the magazine's artist. For your sketches, use some sort of coarse grid graph or "quadrilled" paper, or engineering sketch pads. The latter are green or yellow tablets that are blank (with border) on one side and gridded on the other. The grid lines show through to the blank side enough to guide you in making the drawing, but don't appear in the sketch.

If you have a computer drawing software package available, use it. I draw my illustrations with *Visio Technical 4.1*. In many cases, the editor can use computer-drafted artwork as "camera ready." That saves time and money—which makes your article more valuable to them. If you use a drawing package, however, make sure you set all lines to three pixel size, and don't use less than nine point type. Smaller type and thinner lines are hard to reproduce in a magazine.

Provide the editor with the graphics files you create. The native file to *Visio* is suffixed "*.vsd", which is a proprietary format. Ask the editor what types of files they can use—especially if they don't have *Visio*. Most editors can use either encapsulated postscript (*.eps) or tag imaged file (*.tif) files. Visio and most other drawing programs can convert the native format to either *.eps or *.tif, plus other formats. *Communications Quarterly* accepts .pcx, .bmp, .tif, .jpg, and .gif (pc format) and tiff, pict, and eps (MacIntosh format).

The basic requirement for your illustration is that it be understood by the editor and the magazine's artist or draftsman. The line drawing should be neatly done, and contain all of the information the reader needs. For schematics, that requirement includes dimensions, component values, semiconductor device part numbers, and other pertinent data. Keep in mind that there are different drawing practices in effect at different publications. Study the illustrations in the magazines to see which symbols and style elements they elect to use.

Photographs are also very useful for illustrating the technical article. There are some general guidelines for making photos. Don't use the low-cost 110, 126, or disk format films. Use 35 mm or larger (for example, 120 or 220 size) film for your photos, even if you have to borrow a camera. If possible, use black and white film such as T-Max, Verichrome Pan, Panatomic-X, Plus-X, Tri-X, or their equivalents.

Don't use color print film, unless an editor says it's OK. Some magazines can use slides, but check with the editor before you hang your piece on a color transparency. A photo laboratory can make a black and white print from your slide by shooting a black and white internegative from your transparency.

The print should be glossy, with borders (you may need to ask for these), and be either 4 x 5, 5 x 7, or 8 x 10. Keep in mind that 35-mm negatives may not reproduce well at 8 x 10. Place your photo in a celluloid "page protector" (available at office supplies stores for about 50 cents). Tape the photo to the inside paper in a way that keeps the tape off the print. (See why you need to order bordered prints instead of the borderless type that's now standard?)

Captions

If you want to hear Terry scream bloody murder from all the way up there in "live free or die" country, then submit a heavily illustrated article without a list of captions. The editor can't write the captions as well as the author—if only from lack of familiarity—unless he spends a lot of time ferreting the details out of the text. Captions should be typed on a separate sheet of paper and clipped to the illustrations.

Electronic submissions

Personal computers are a fact of life today, and in the magazine business that goes double. In fact, magazines were among the earliest large scale users of computers because of page layout programs and electronic typesetting. When I submit an article, I send an IBM-formatted 3.5-inch 1.44-mbyte diskette with two text files on it. One is the *Word for Windows 6.0* version of the article, and the other is an ASCII version (most word processors recognize ASCII text). If the editor can use graphics files, I also include those.

Conclusion

There are any number of reasons why you might want to write an amateur radio article: it pays money, it brings recognition, it helps others, it helps the amateur radio hobby, and it's a heckuva lot of fun. And, guess what? You can do it? ■

PRODUCT INFORMATION

Svetlana's EF86/6267 Audio Small-Signal Pentode

Svetlana Electron Devices, Inc. has announced the availability of its EF86/6267 audio small-signal pentode—manufactured at the company's facility in Russia. The device features:

- Very high voltage gain in pentode connection.
- Low noise, low microphonics, low heater-cathode hum induction.
- Internal shielding.
- Solid metal shield canister.
- Low distortion in pentode or triode connections.

For details contact Svetlana Electron Devices at 3000 Alpine Road, Portola Valley, California 94028; Phone (415) 233-0429; Fax, (415) 233-0439.

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MODELING AND UNDERSTANDING SMALL BEAMS: PART 6

Fans, bowties, butterflies, and dragonflies

Except for the EDZ and the ZL Special, all of the antennas I've looked at in this series of small beams have averaged about 3/4ths normal side-to-side length—about 12 feet on 10 meters. To achieve this size, I had to either adopt a nonlinear geometry (the X-beam and the Moxon rectangle) or load the linear element (the linear-loaded Yagi). The nonlinear geometries were both in the horizontal plane, adding nothing to the height of the antenna structure.

An alternative, driven by a different goal, was to increase the thickness of the beam's dipoles, automatically shortening them. The goal has been to increase the SWR bandwidth of the basic half-wavelength dipole. A "fat" wire, such as a large diameter conduit, will achieve this within the limits of the added weight, but a virtual fat wire works just as well. A virtual fat wire is a structure of multiple thin wires that simulates, by outlining, the overall wire size desired. Enter the fan and the bowtie.

The terms "fan" and "bowtie" have a checkered history in amateur antenna lore. Some writers have treated any set of spread multiple dipoles as a fan antenna. The most common use of the term in this way is for two or three dipoles—with each end independent—cut for various portions of the 3.5- to 4-MHz band. This multiple antenna exhibits separate SWR minima for each dipole. The spread multiple dipole isn't what I consider a fan. To be a "true" fan (at least for my purposes), the structure must electrically connect the element members at the end. The "true" fan is a single virtual fat element that exhibits a single resonant point. Unlike the folded dipole, the wires of each side of the fanned element are also joined at the feedpoint.

Moreover, almost any version of the spread single element (except for the multiwire cage) has been called either a fan or a bowtie, with no distinction between the two terms. Perhaps the use of the antenna in television reception occasioned the double reference, or perhaps it has been around since the antenna achieved some notice in the late '30s and '40s, evolving into the "Wonderbar" of the '50s. However, there are two distinct ways of widening an antenna element. The first involves a linear fanning of two wires and bringing them back together vertically at the end (with or without a conductive center tube for support): I call this element the fan. The second way involves spreading the wires to a maximum separation somewhere along each half of the dipole, usually the midpoint, and bringing them back together angularly toward the end: for obvious visual reasons, I'll call this the bowtie. Fan and bowtie elements have similar, but not identical, properties.

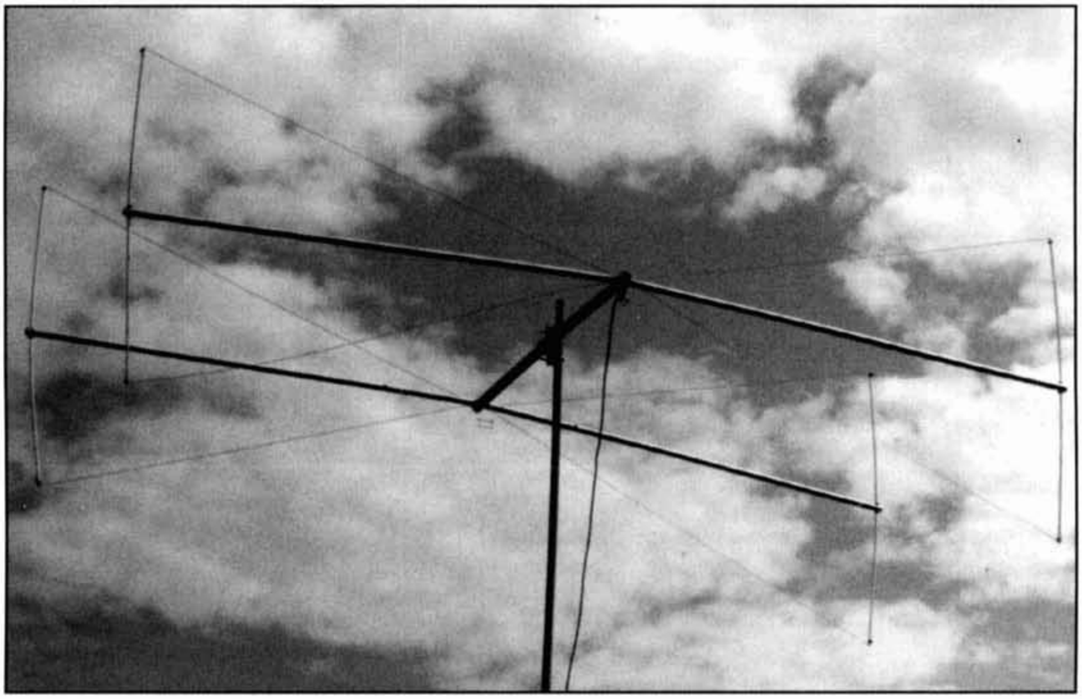

Photo A. The fan dipole.

A look at models of these antenna configurations can lead to the construction of a very practical small beam. Because the center support-conductor of certain fan and bowtie antennas required a large-diameter tube, while the remaining structure was either wire or small tubing, MININEC was the modeling program of choice.[1]

The fan

Center-mounted antennas, like a half-wavelength dipole of aluminum tubing, permit simple mountings. Therefore, the classic fat-wire geometry is the fan, as sketched in **Figure 1**. Ignoring the horizontal member for a moment, the basic fan consists of a triangular loop of wire or tubing on either side of the center point. Some have been made from a continuous piece of 0.25- to 0.5-inch diameter tubing, bent to shape, flattened at the center junction, and mounted to a center insulator-support. After a violent thunderstorm, these wind-grabbers often took on shapes only seen in abstract art.

Use of a horizontal center support made the structure more durable. Sturdier aluminum tubing of standard Yagi diameters could support a thinner vertical tube at the end. In turn, the vertical tube supported wires forming the slanting members of the fan. The resulting fan slipped the wind well, stressed the tubing for strength, and reduced the element's weight.

A fan dipole is still a dipole; in fact, a slightly shortened dipole. Therefore, it displays the typical dipole pattern, as shown in **Figure 2**, but at a slight reduction in gain relative to a standard thin-wire dipole. The reduction is insignificantly fractional (about 0.1 dB) and cannot be detected in operation, but it is determinate.[2]

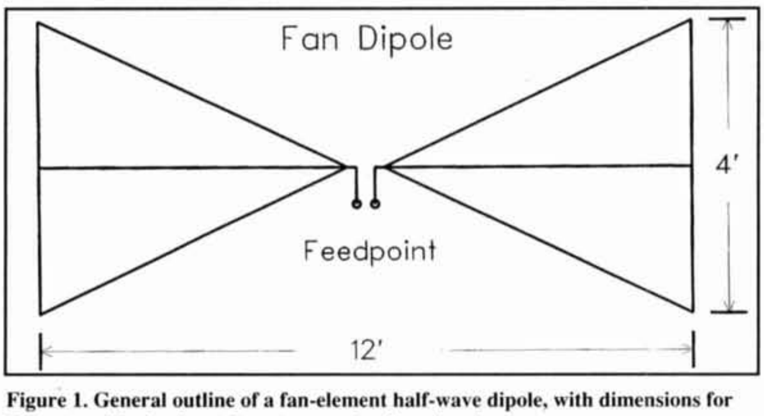

Figure 1. General outline of a fan-element half-wave dipole, with dimensions for 10 meters. The horizontal center member is optional.

Adding the conductive center member to the fan structure tends to increase the resonant frequency of a fan dipole relative to a similarly sized fan without the member. The amount of change is not insignificant. A 10-meter fan without the center member with the dimensions shown in **Figure 1** is resonant at about 28.25 MHz, while adding a center member raises the resonant frequency to about 28.75 MHz. The increase in resonant frequency results from changes in the current distribution, some of which is now flowing on the shorter center member. Conversely, fattening the vertical

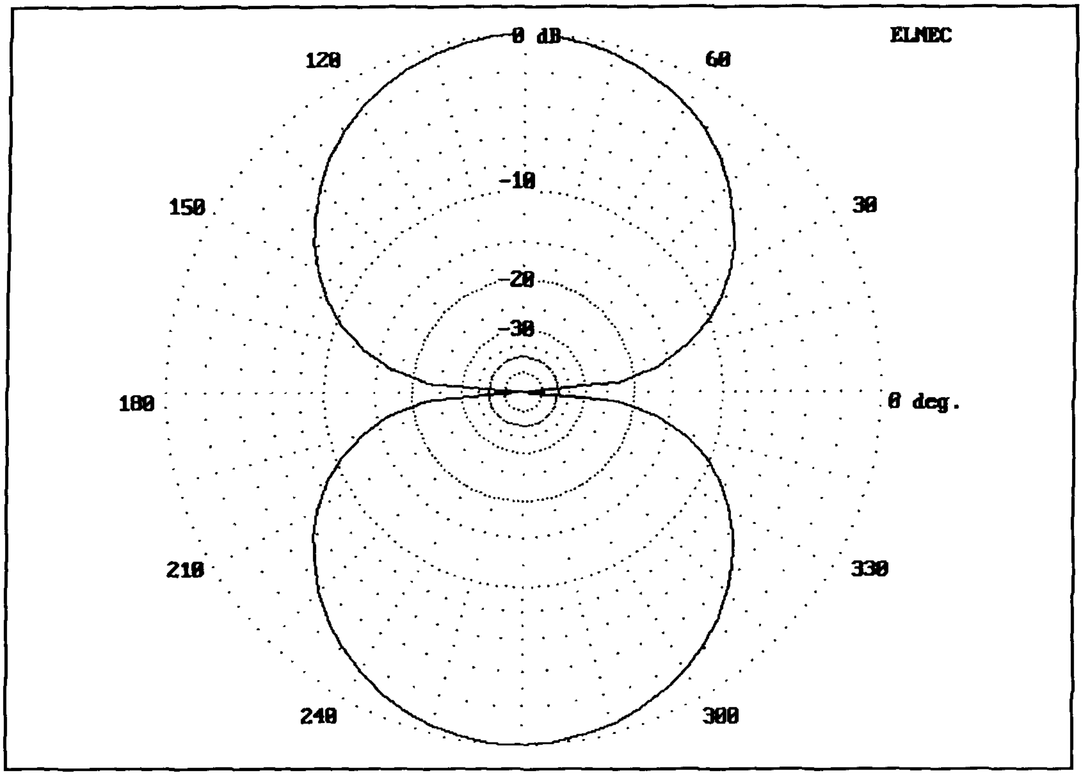

Figure 2. The azimuth far field pattern (in free space) of a fan or bowtie dipole is virtually identical to its linear counterpart.

members at the outer ends of each element lowers the resonant frequency.

How broad-banded is the fan? The somewhat disillusioning answer appears in the graph in **Figure 3**. While the fan with a 3:1 ratio of overall length to vertical height shows an SWR bandwidth that's significantly better than that of a #14 wire dipole, it approximates the bandwidth of a simple horizontal half-wavelength dipole constructed from 1-inch diameter aluminum tubing. With most rigs, all three antennas would operate satisfactorily across the wide reaches of 10 meters.[3] However, the bandwidth advantages of fans begin to appear when one constructs a two-element Yagi from them.

The fan dipole, using the dimensions of **Figure 1**, does have one small additional advantage. The resonant frequency of the structure is close to 50 ohms, a closer match for the standard ham coaxial feedline than the linear dipole.

The bowtie

An alternative to the fan is the bowtie, another form of the vertically fattened dipole. The bowtie stretches the vertical dimension in the middle of each half of the dipole, as shown in **Figure 4**. The vertical supports for the wire perimeter are normally nonconductive. As with the fan, the bowtie may use or omit a conductive horizontal center member.

The dimensions shown for the bowtie will produce a resonant dipole at 10 meters. Without a center member, the assembly will resonate at about 28.5 MHz; with the center member, the resonant point moves to 29.25 MHz.

The bowtie shows a resonant feedpoint impedance a bit lower than a fan, but still close to 50 ohms. Likewise, its SWR bandwidth is quite similar to the fan, as shown in the four curves of **Figure 5**. Despite the offset in resonant points, the curves are all congruent to a high degree.

The bowtie requires a larger structure to achieve resonance at the same frequency as a comparable fan, about a foot longer and several inches higher. Largely for that reason, plus the fact that the element is fatter in the region of higher current, the antenna shows a slight gain advantage over the fan (but still marginally less than a linear dipole). However, **Figure 6** demonstrates that the advantage is nowhere greater than 0.1 dB across the band, an increase that could make no difference in practical amateur operation.

Fans and bowties have no magic formulas for determining the ratio of overall length to height. With the exception of the 12-foot length of the fan, the dimensions are somewhat arbitrary; although the 3:1 length-to-height ratio appears to be a practical limit for home construction. The two antenna configurations do lend themselves to comparisons relative to how

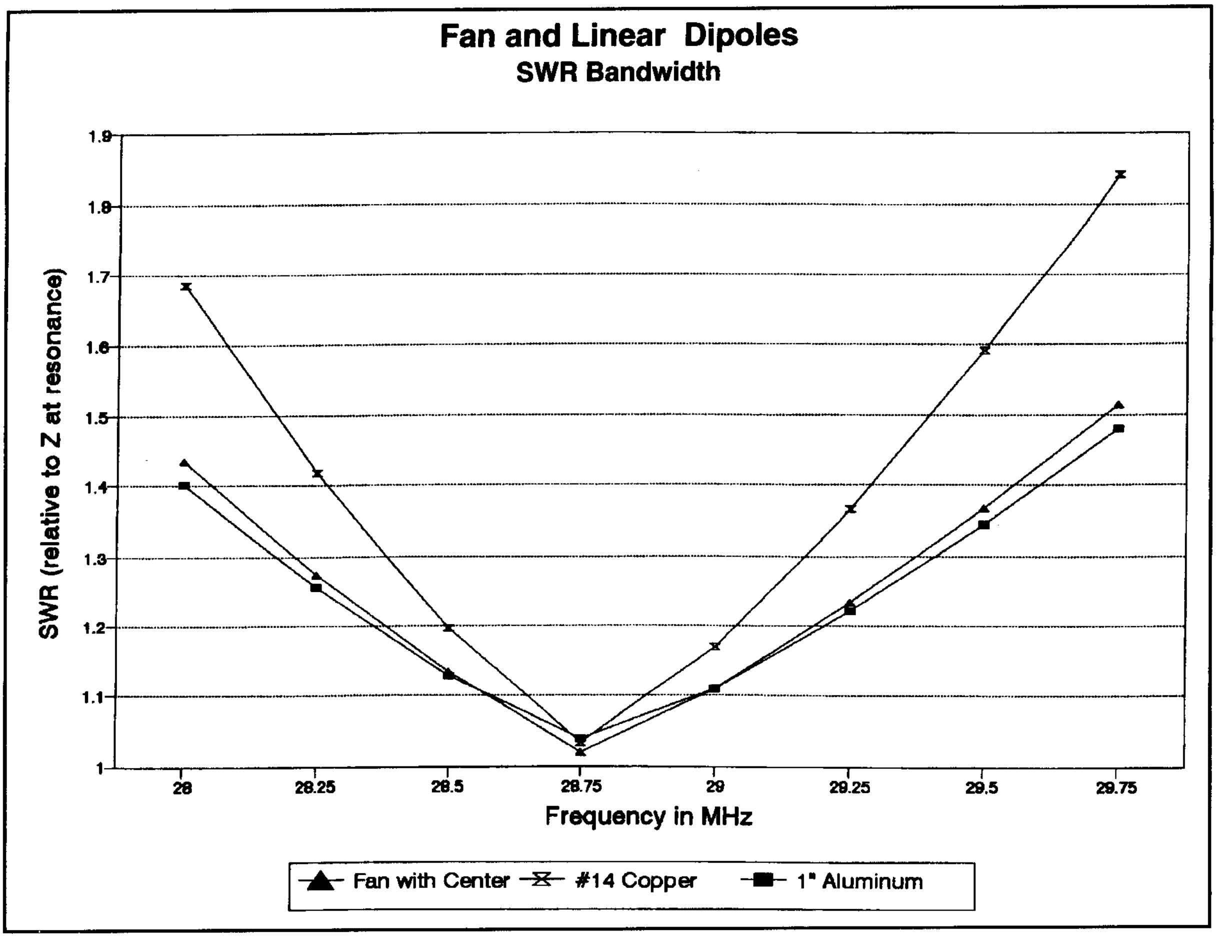

Figure 3. SWR bandwidth curves for two linear dipoles (#14 copper wire and 1-inch diameter aluminum tubing) and for a fan dipole.

fat we make them. The amount of current carried in the fat part of the wire assembly begins to show up in terms of several different factors.

If we reduce the height of the assemblies to 75, 50, and 25 percent of their 10-meter resonant height, and then find the resulting resonant frequency, we obtain the two curves in **Figure 7**. Note that the curve for the fan dipole is relatively linear, while the curve for the bowtie tapers toward its value for a 7.05-foot linear element. A similar difference occurs with respect to the gradual increase of the feedpoint impedance as the antenna approaches the condition of being a linear half-wavelength dipole. As shown in **Figure 8**, the fan dipole impedance increases almost linearly, while the bowtie impedance tapers toward the dipole value.

The effects of increasing the height of the

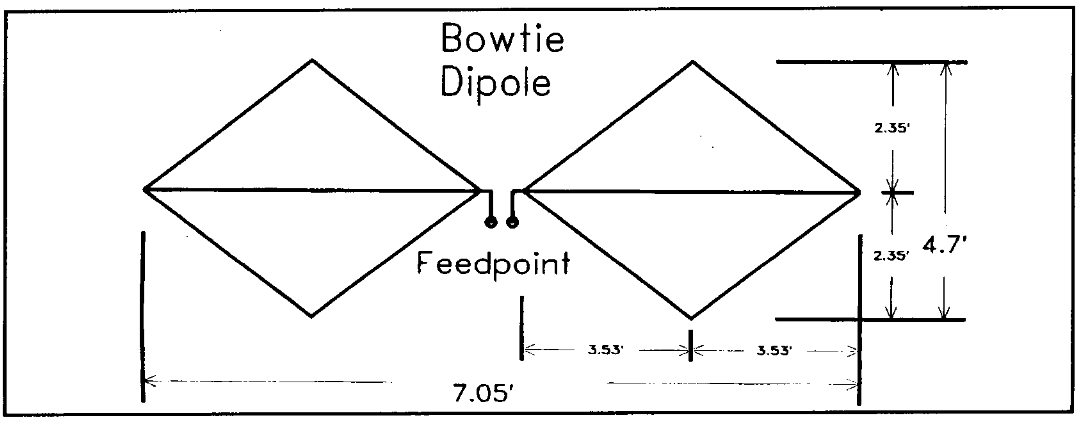

Figure 4. General outline of a bowtie-element half-wave dipole, with dimensions for 10 meters. The horizontal center member is optional.

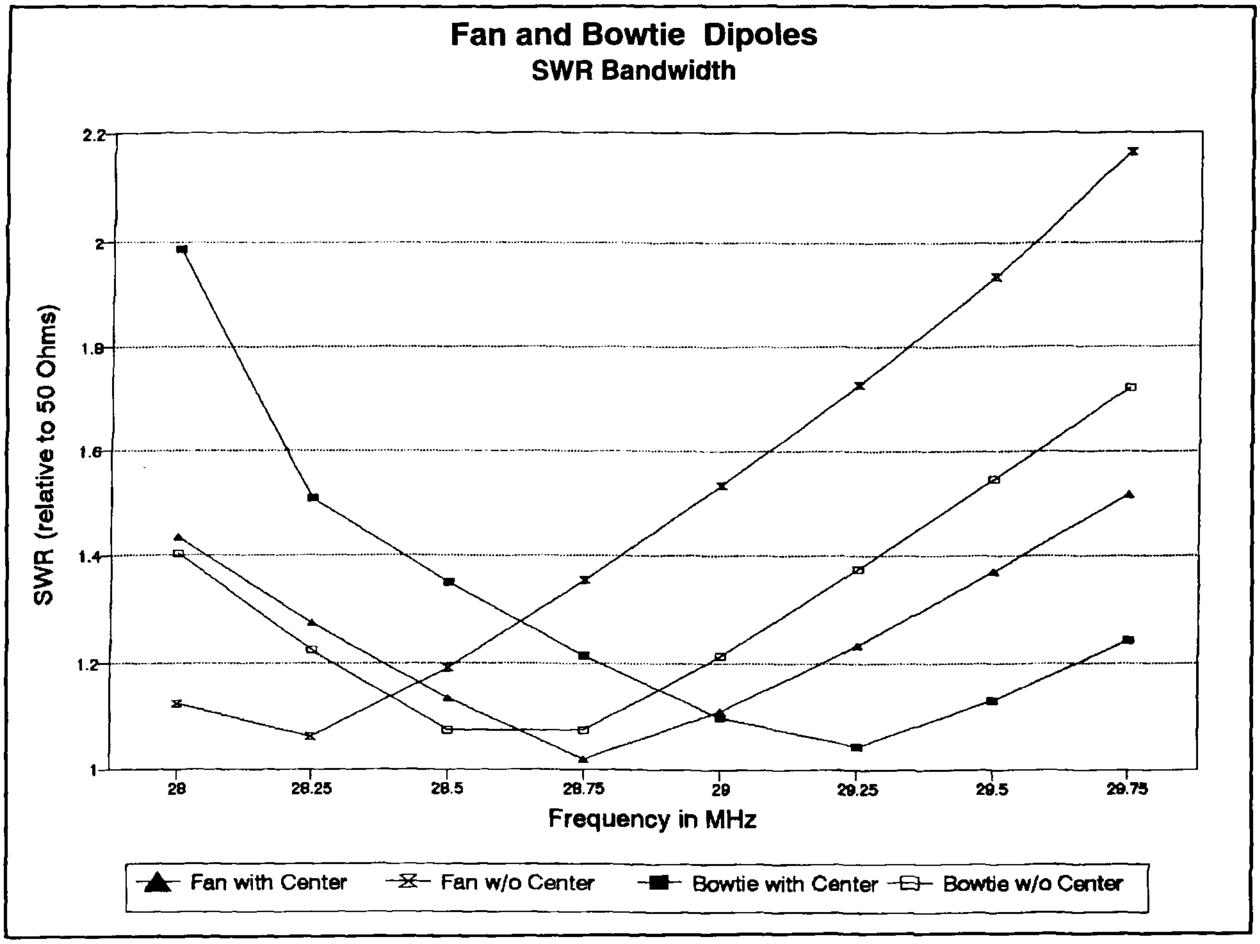

Figure 5. SWR bandwidth curves for both fan and bowtie dipoles, with and without horizontal center members. The displacement of the resonant points of these congruent curves permits the display of values at greater distances from resonance.

bowtie closer to the high-current portion of the antenna structure also show up in a comparison of gain figures. Interestingly, as shown in **Figure 9**, the fan's gain increases as it approaches a resonant linear element, but the bowtie's gain decreases under the same condition.

If one wishes to build a half-wavelength dipole to cover all of 10 meters and wishes to have a reasonably good match to 50-ohm coax, then the fan dipole may be a good choice. Six-foot lengths of aluminum tubing are standard hardware outlet items, and the 4-foot vertical members may be 0.25 or 0.375-inch diameter rods from the same source: a single 8-foot section will suffice for both ends. The wire pieces can be #14 stranded, which is widely available. Construction details appear later when we look at a practical 10-meter two-element Yagi.

Butterflies and dragonflies

Because fans and bowties are fat dipoles, rather than loaded or geometrically altered dipoles, you would expect the performance of comparably spaced two-element Yagis to be similar. For this exercise, I used as a standard the two-element broadband Orr Yagi, with a driven element 16.1 feet long, a reflector 17.6 feet long, and 4.25-foot element spacing. Centered at 28.5 MHz, this antenna exhibits moderate gain and front-to-back ratio, while holding its characteristics across most of the 10-meter band.

For comparison, I modeled Yagis made up of both fan and bowtie elements. A commercial version of a Yagi using fan elements has enjoyed considerable success: the Butternut HF5B Butterfly beam.[4] If a fan Yagi is a butterfly, then perhaps a bowtie Yagi is a dragonfly.

The bowtie and fan models in this study use 1/8th wavelength spacing. It's easier to retain the basic geometry of the fan and bowtie dipole elements and load them for parasitic operation than it is to reshape each element for optimal performance without loading. (The Butternut beam also uses this method, since its two basic element structures are identical.) Therefore, the elements in the models have dimensions the same as those in **Figures 1** and **4**. The models explored used horizontal center members, as it's difficult to build a sturdy element without the center tube.

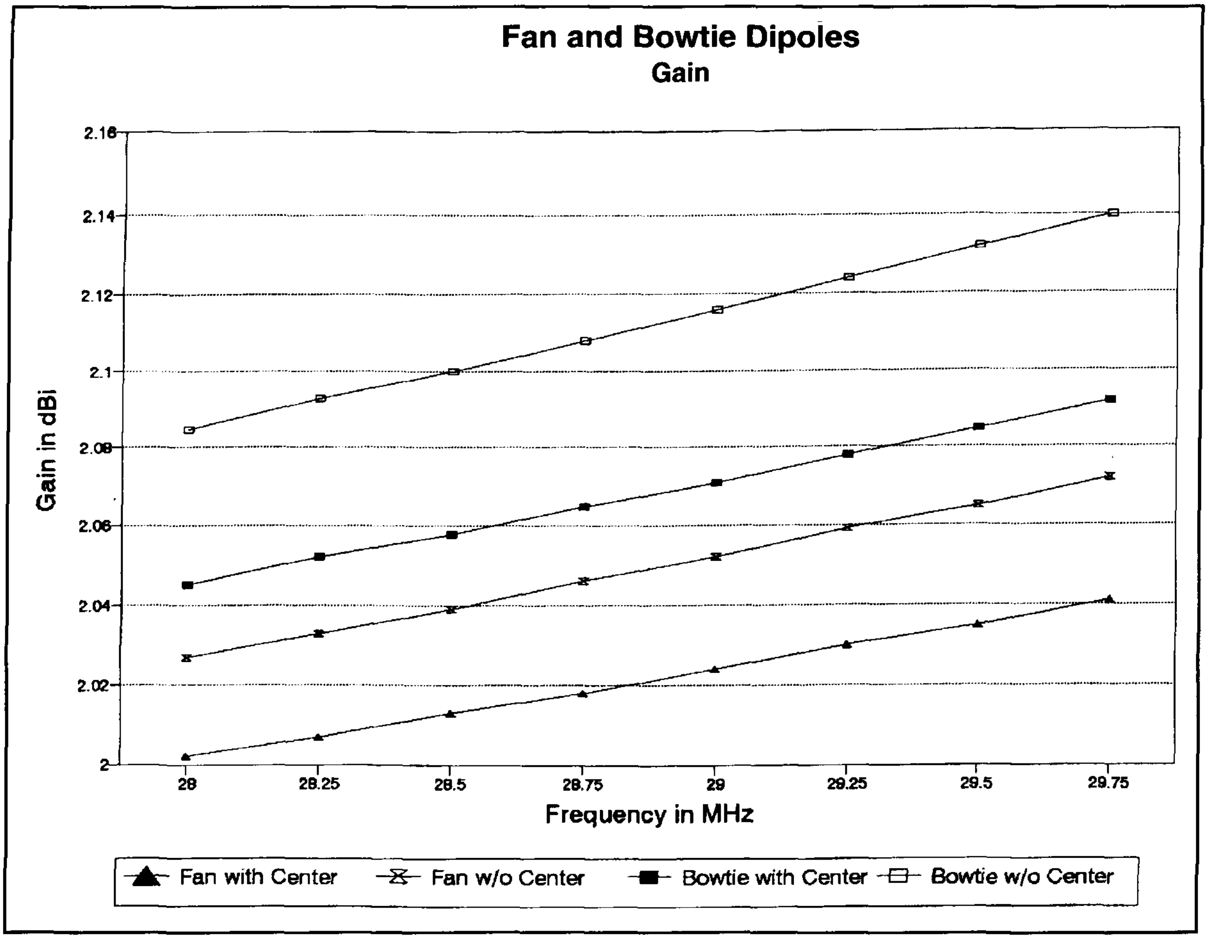

Figure 6. The gain properties of both fan and bowtie dipoles, with and without horizontal center members.

A resonant dragonfly beam requires about 35 ohms load on the reflector for maximum front-to-back ratio. A small inductor or inductive (shorted) transmission line stub across an insulated split in the reflector would provide the required loading.[5] The resulting feedpoint impedance at the design center frequency of 28.5 MHz is close to 20 ohms, with a reactance of +j12. If the inductive reactance is compensated with a pair of series capacitors (about 930 pF each), a 2.5:1 impedance ratio (1.6:1 turns ratio) broadband transformer might provide a good matching system, as would one of W2FMI's baluns. However, one might also consider using a modified beta match, with a capacitor (about 140 pF) across the terminals as the shunt element.[6] The actual values requiring compensation may vary with the exact materials used for the elements.

A resonant fan-element Yagi using 12 by 4 feet with a 1-inch diameter horizontal center member requires similar treatment. A 1/8th wavelength spaced Yagi using fan elements as described earlier requires reflector loading with either an inductor or a shorted transmission line stub. The driven element shows a resistive feedpoint component of about 17 to 25 ohms, with an inductive reactive component that varies with the size of the element members. One may either use a pair of capacitors on each side of the feedpoint to compensate for the reactance and then insert a 2:1 balun, or use a beta match with a capacitor as the parallel element.

Among the three two-element Yagi model antennas—the linear, the dragonfly, and the fan-element version—there's nothing to choose with respect to pattern, gain, or front-to-back ratio. **Figure 10** records the modeled gains of the antennas across the 10-meter band. The maximum difference of 0.2 dB would be undetectable. The dragonfly and the fan Yagi do exhibit something over 2 dB additional front-to-back gain at design center, but this is less than half an S-unit improvement, and that figure deteriorates as one moves to the high end of the band. Moreover, if the reflector is loaded to achieve its gain and front-to-back ratio, the load losses, although quite small, will have an impact on the figures, reducing them very slightly.

Figure 12 shows the composite patterns of any of the three antennas across the band. The innermost rear lobe (almost inverted-bell

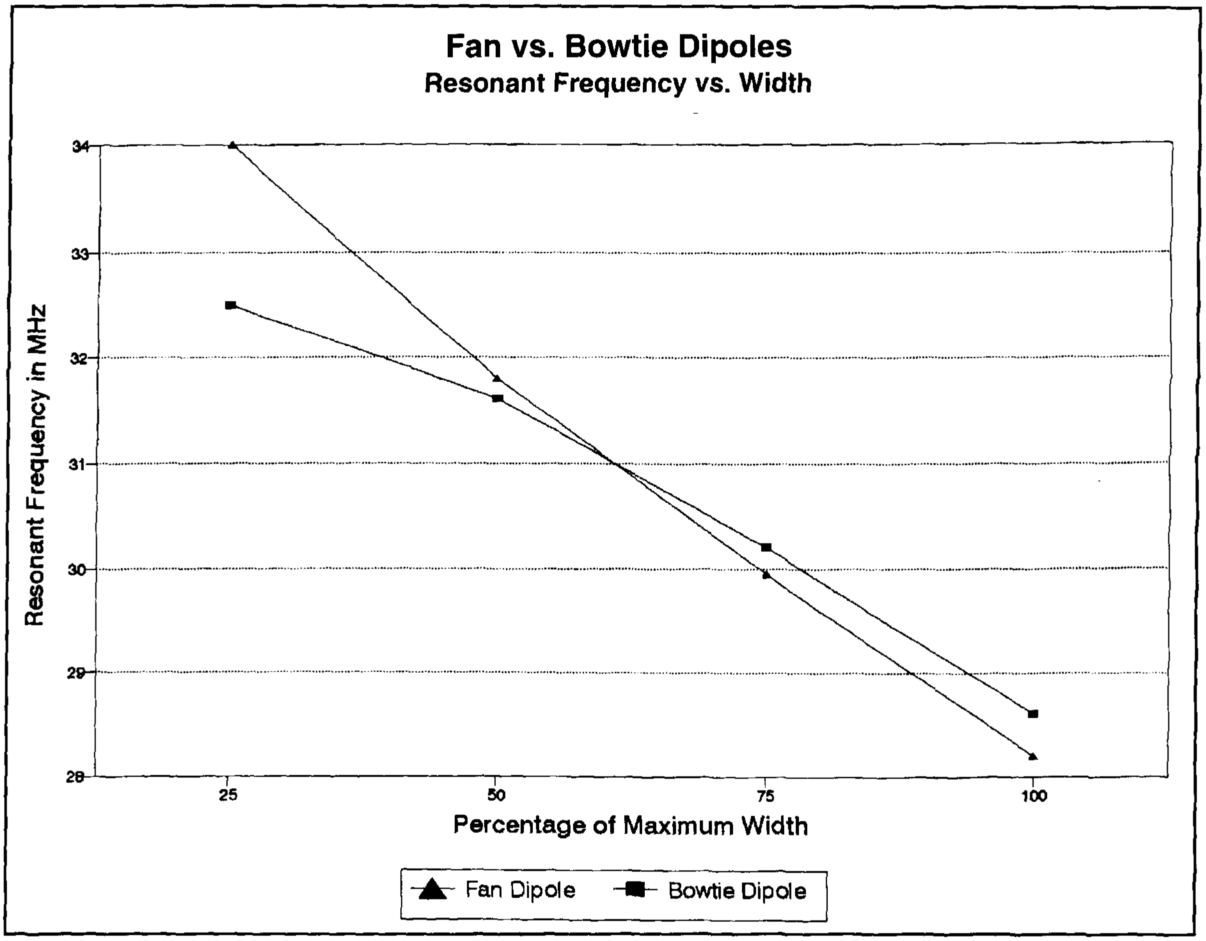

Figure 7. The resonant frequency variation of both fan and bowtie dipoles as the vertical dimension is varied.

shaped) represents the 28-MHz pattern, as does the highest forward gain circle. The other gain rings decrease sequentially as the frequency increases. Likewise, using an angle of 330 degrees as a reference line, the rear lobe patterns are sequential outward on that line as frequency increases. The innermost lobe line at 270 degrees (directly to the rear of the forward lobe) represents the maximum front-to-back ratio, which is at 28.5 MHz, the design center for all three antennas.

All three antennas have the same gain and front-to-back characteristics. Moreover, the dragonfly and the fan Yagi are more complex structures. Why should one build anything but a full size Yagi? There are two potential reasons. First, the full size Yagi is over 17 feet wide, while the dragonfly is a little over 14 feet wide, and the fan Yagi is 12 feet wide. If space is at a premium, then fan and bowtie elements may be very useful for providing full-size performance by trading some vertical space for some horizontal space.

Second, the fan Yagi and the dragonfly have a broader 2:1 SWR bandwidth than a linear two-element Yagi. **Figure 13** traces those curves. The linear Yagi can cover 28.1 to 29.2 at a 2:1 SWR. However, by moving the design center of the dragonfly and the fan Yagi to about 28.75 MHz, either would likely cover the entire 10-meter band. The fan Yagi is slightly superior in this regard to the dragonfly.

For many—if not most—hams, these reasons may not be significant enough to prompt a construction project. Moreover, the remaining bands from 14 MHz upward are narrower and don't require element fanning for whole band coverage. Nonetheless, it's worth the effort to build a trial antenna to test the values suggested by the models.

Building a fan-element Yagi

Of the two element types—the fan and the bowtie—the fan promises to save the most room with full-size performance. Moreover, it appears easier to build. Therefore, I decided to try a fan-element Yagi, even though I already had a Butternut Butterfly on my tower. The commercial antenna is multiband, has numerous matching pieces attached, and is somewhat

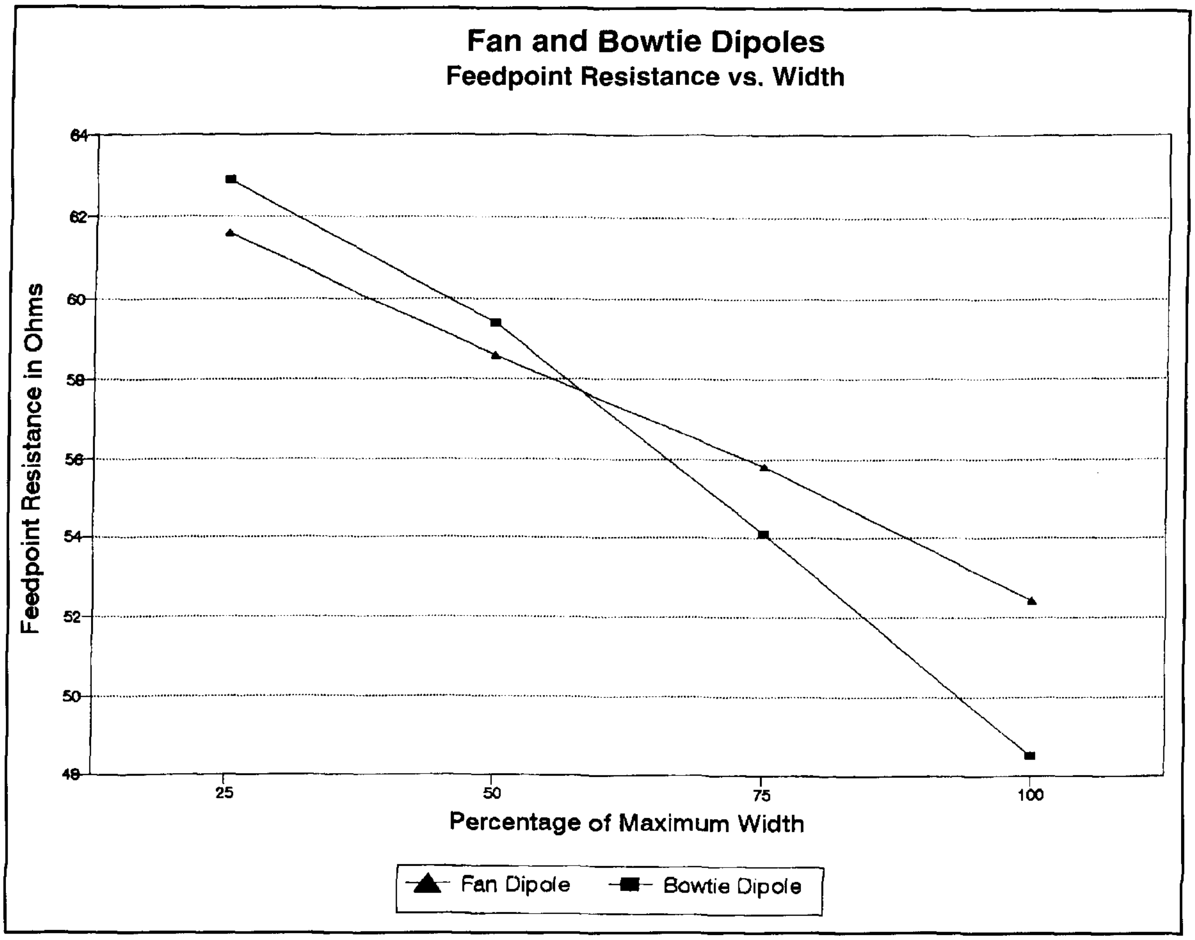

Figure 8. The feedpoint resistance variation of both fan and bowtie dipoles as the vertical dimension is varied.

larger than the 10-meter model developed here for single-band use. (The Butternut elements are about 12.5 feet long, with 6-foot vertical end spreaders on a 6-foot boom. An individual fan element, minus the added metal of the matching system, is resonant above 25 MHz, with the antenna loaded to Yagi performance on each band.)

A Yagi using fan dipole elements presents interesting construction challenges. **Figure 14** shows the general outline of a two-element fan Yagi. Its function is to associate the component names with the physical structures of the antenna. Each element has three major conductive components: the center member, the vertical member, and the fan wire. Everything else, including the boom and mast, are supports. The wire portions of each element come together at the feedpoint or the center of the reflector. The center connection and the boom-to-element support must not interfere with each other. The vertical end pieces must fasten to the center member of the element without slipping. The wires must connect to the vertical member ends with enough stress to stiffen the structure, but not enough to deform or break it. A simple system of epoxy-coated plywood plates wouldn't do here.

Because my standard test boom is a 5-foot length of 1.25-inch diameter nominal Schedule 40 PVC, I decided to put the drill press to use to mount a 0.5-inch diameter nominal Schedule 40 PVC through the boom. The real outside diameter of the half-inch material is closer to 0.875 inch for a good fit with 1 inch outside diameter aluminum (0.55 thickness) center members, with a little tape at two points on each side of center for a tight fit. If the sag of the Schedule 40 PVC, which allows about a 3-degree droop in the fans, causes concern, you may stiffen the center insulated piece. One technique is to thread thin strips of fiber glass cloth through the center of the half-inch PVC. Then seal one end, stand the piece vertically, and fill it with car-repair epoxy resin. Allow extra setting time, as the tube is closed. The resulting rod is easily machined with wood bits and will outlast any other part of the antenna.

Aluminum rods, either 0.25 or 0.375 inch in diameter, form the end members and fit through holes in the center member. Although I initially had some fears of weakening the center mem-

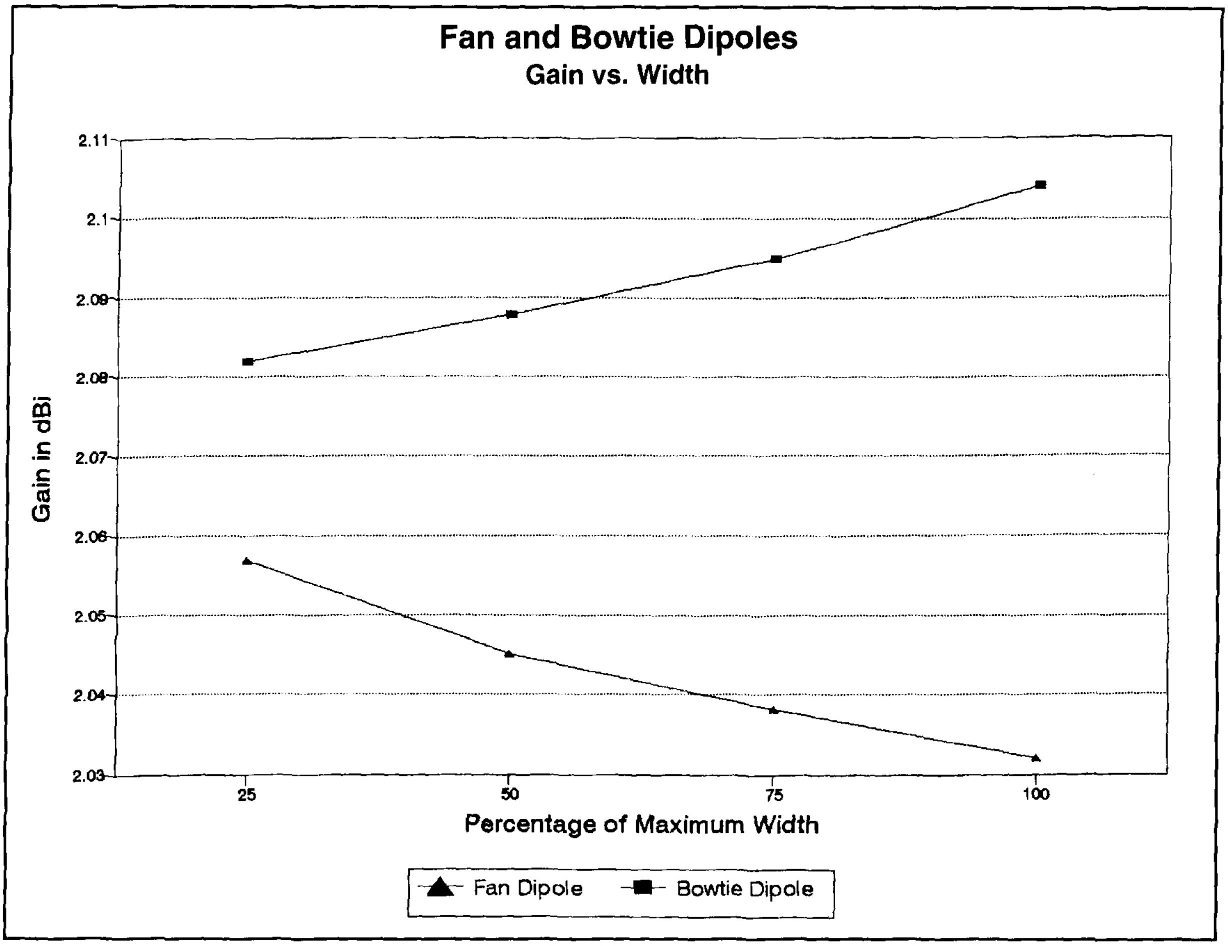

Figure 9. The gain variation of both fan and bowtie dipoles as the vertical dimension is varied.

ber tubing by drilling out holes, the result has stood the test of time and wind—so far. If you have access to 0.5-inch diameter aluminum tubing for the end members, you can save a significant bit of downward force at the ends of the center members. The rods used in the test antenna have created no problems and were used owing to their easy availability at hardware depots.

Actual construction may take any electrically and mechanically sound form. The construction method used in the test antenna began with a 5-foot length of 1.25-inch (nominal) Schedule 40 PVC. At a distance of 4.31 feet apart, I used a 7/8-inch diameter hole cutter (for wood) to form the openings for the center-member supports, two 3-foot lengths of 0.5-inch (nominal) Schedule 40 PVC. Aligning the first hole for the center of the tubing in a drill press or other jig is the first careful task. A little sanding prepares the opening for the center member support, which is bolted in place with a 2-inch long #10 stainless steel (SS) bolt and nut. The first support acts as an alignment tool for setting up the second hole in the boom. The half-element fans will slide over the center-member support during final assembly, as shown in **Figure 15**. For now, remove the supports from the boom.

To prepare the 1-inch diameter 6-foot long aluminum center members, use a drill press or other alignment jig to drill 0.25 inch (or 0.375 inch for 3/8 inch rod) holes within a half inch of the outer end of each tube. Slide the 4-foot sections of aluminum rod through the holes with their centers in the middle of the tubes. Several clamping systems are possible. The simplest might be a pair of the smallest size stainless steel hose clamps available placed above and below the center member. A #6 or #8 SS bolt and nut assembly can act as an electrical contact to a small flat spot filed on the rod.

Alternatively, the U clamp shown in **Figure 16** can be inserted while passing the rod through the tube. The U should be just large enough to press into the tube. Again, a separate bolt and nut system should be used to ensure good electrical contact between the two elements.

Drill a hole in each end of each vertical member for a bolt and nut to secure a clamp to be soldered to the fan wire, as suggested in **Figure 16**. The sketch shows separate wire

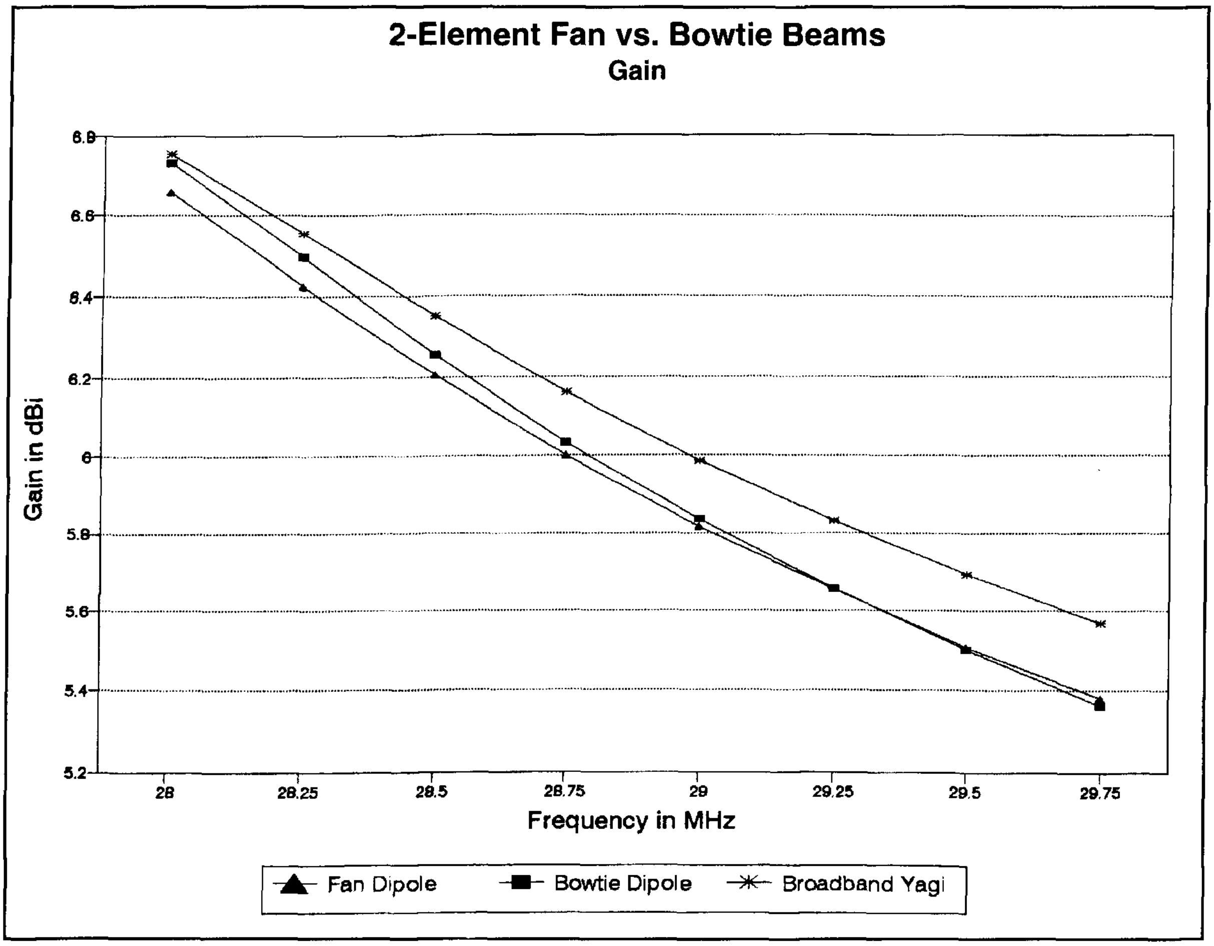

Figure 10. The gain across 10 meters of a fan, a bowtie, and a linear element Yagi.

clamp and wire elements, but wrapping the wire around the nut and using a sufficiently large washer to provide good electrical and mechanical contact should also suffice. Wrap the wire back around itself for mechanical strain relief and solder the wire with a large iron or small torch. (Be certain that the #14 fan wire is long enough—about 12.65 feet plus wrap-around stubs—to make both fan wires on each side from one wire length. See below.)

Prepare the hole in the center member at the antenna element center. At this stage, it's easiest to insert the PVC center-element support to ensure total system alignment. If the fit is too loose, wrap electrical tape around the support at a minimum of two points to create a snug but nonbinding fit. If you fit both halves of the element to the support and lay them on a floor surface, you can be sure that the fan wire bolt holes at the center (see **Figure 15**) will be aligned.

The bolts closest to the boom that secure the center member to the PVC support should be about 2 inches long because they'll go through the center member and support, secure the feed and loading elements, and clamp the fan wire. Extra nuts and washers are useful to isolate these functions. You can place the fan wire on the "outside" of the overall structure, using large washers to clamp the midpoint of the wire and secure the assembly (to this point, without matching or loading elements) with a lock-washer and nut. Later, you can add leads for the antenna feed and for the reflector load.

The fan wire for each half element can be a continuous piece running from the upper tip of a vertical member to its lower tip (12.65 feet plus stubs). Prepare one tip end as described earlier and run the wire around the bolt at the center. Stress the vertical member slightly and clamp the center bolt. Then run the wire to the other tip, choosing a length that will stress that end of the rod by an equal amount. Now you can cut and finish the wire according to your choice of tip clamps. Finally, loosen the clamping washers and nut at the element center. Equalize the stress on the vertical member and reclamp.

While the element is flat on the floor, you may add a second bolt through the center member and its support near the end of the support within the tube, about 15 to 17 inches from the

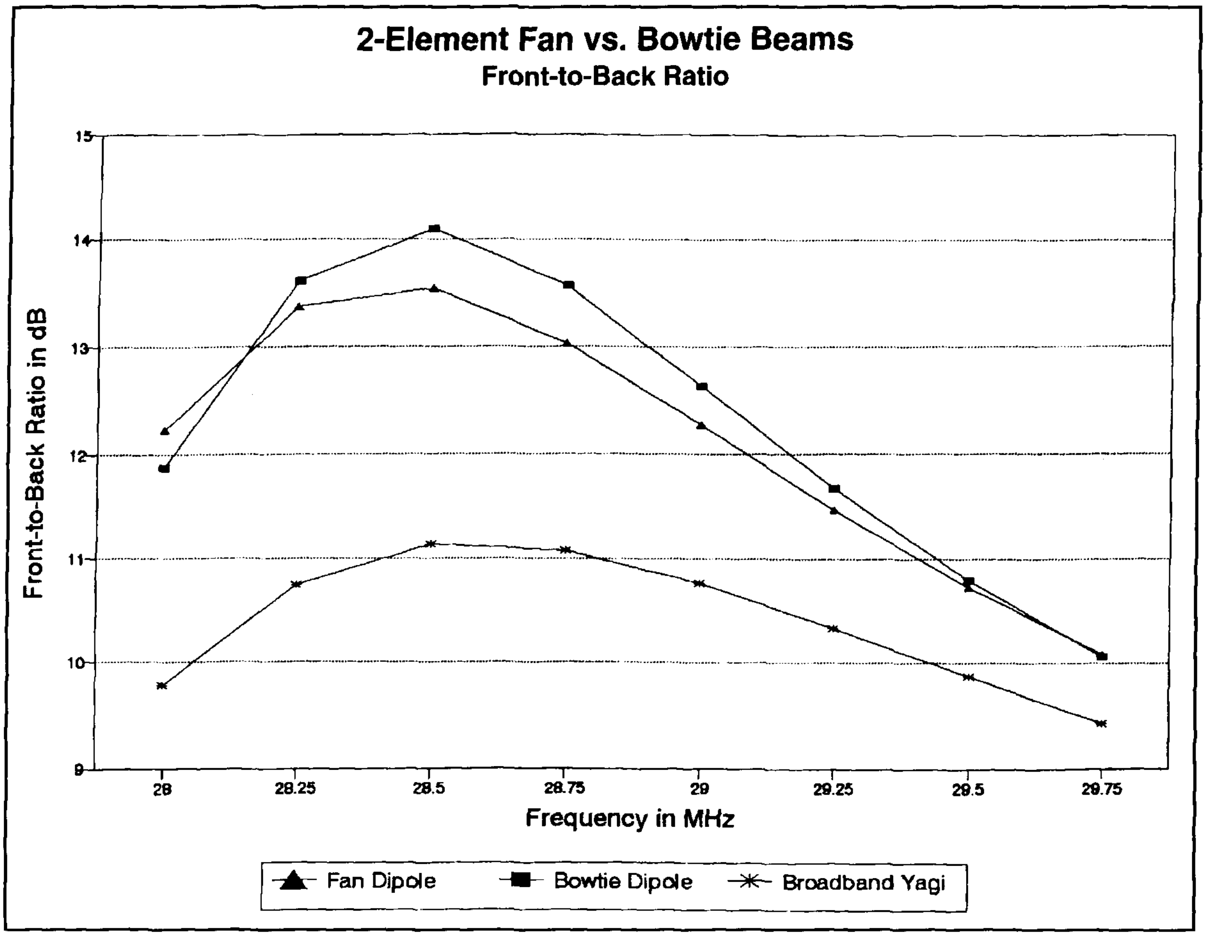

Figure 11. The front-to-back ratio across 10 meters of a fan, a bowtie, and a linear element Yagi.

element center. Choose a point that doesn't go through the tape shim.

Once both antenna elements are ready, insert the center-member support PVC through the boom hole and lock it in place with its bolt. (The boom will likely be vertical if the antenna element is on the floor during this phase of assembly. I taped mine to a plumb house support post to ensure correct element alignment during this step of construction.) Reassemble the other half element to the support, remembering to clamp the fan wire center with equal stress on the vertical member tips. Using a prop to support the completed element, add the second element to complete the basic assembly. Place a boom-to-mast plate at the center of the boom and install a short section of mast. These additions will permit you to balance the assembly in a fixture while installing the feedpoint match components and the reflector load. An old "dish-pack" box with a 1.25-inch hole in the top is my favorite prop for light antennas.

After experimenting with several construction techniques for long-term installation, when it came to the final test antenna I opted for the most rapid assembly and disassembly techniques. **Figure 17** shows some of the variant details of this option. Vertical members were locked in place with stainless steel hose clamps. The wires passed over grooves in the ends of the vertical rods and were clamped with similar hose camps. The resultant structure requires about 15 minutes for field assembly and a like time for disassembly. The antenna pieces and the four sections of 5-foot 1.25-inch diameter masting require a space about 1 by 1 by 6 feet in the back of a pickup truck for transport.

Test results

The antenna is almost ready for testing. The study of the general properties of fan and bowtie elements and their Yagi counterparts employed simplified models using #14 copper wire for all members. The trends and curves that result from these models are accurate, but changes in materials will alter the test results and expectations of performance—especially with respect to matching and feeding the array. The actual antenna is predominantly aluminum,

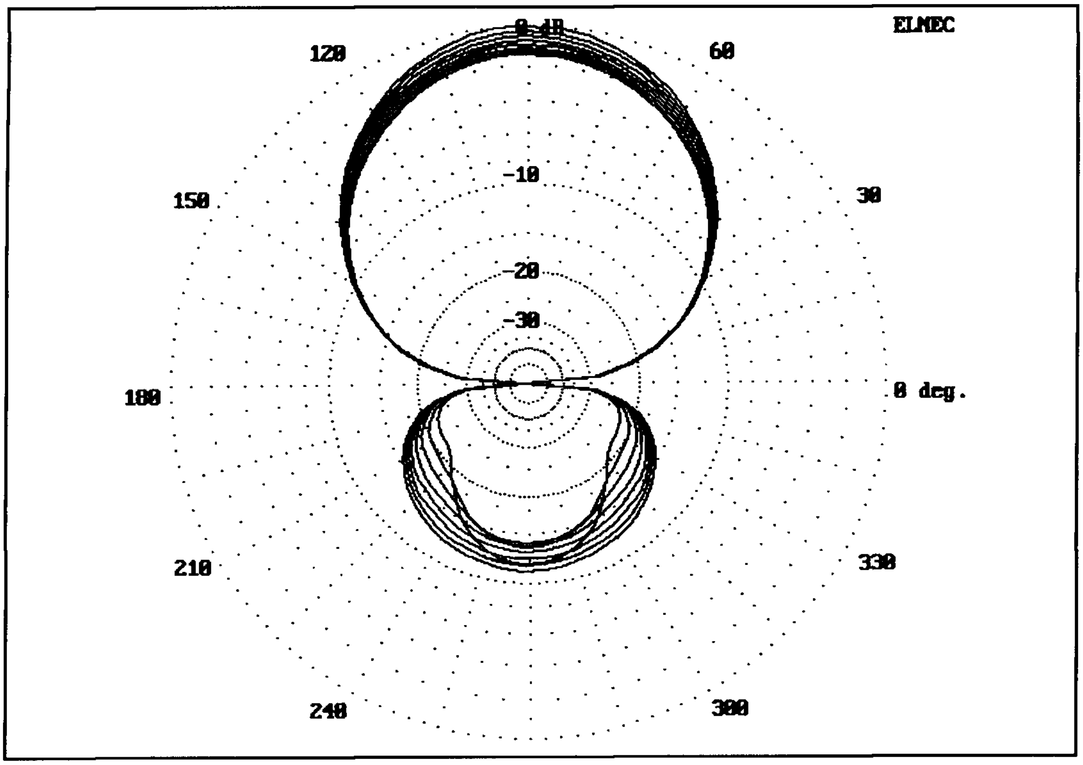

Figure 12. Azimuth patterns (in free space) of a fan, a bowtie, or a linear element Yagi across 10 meters at 0.25-MHz intervals.

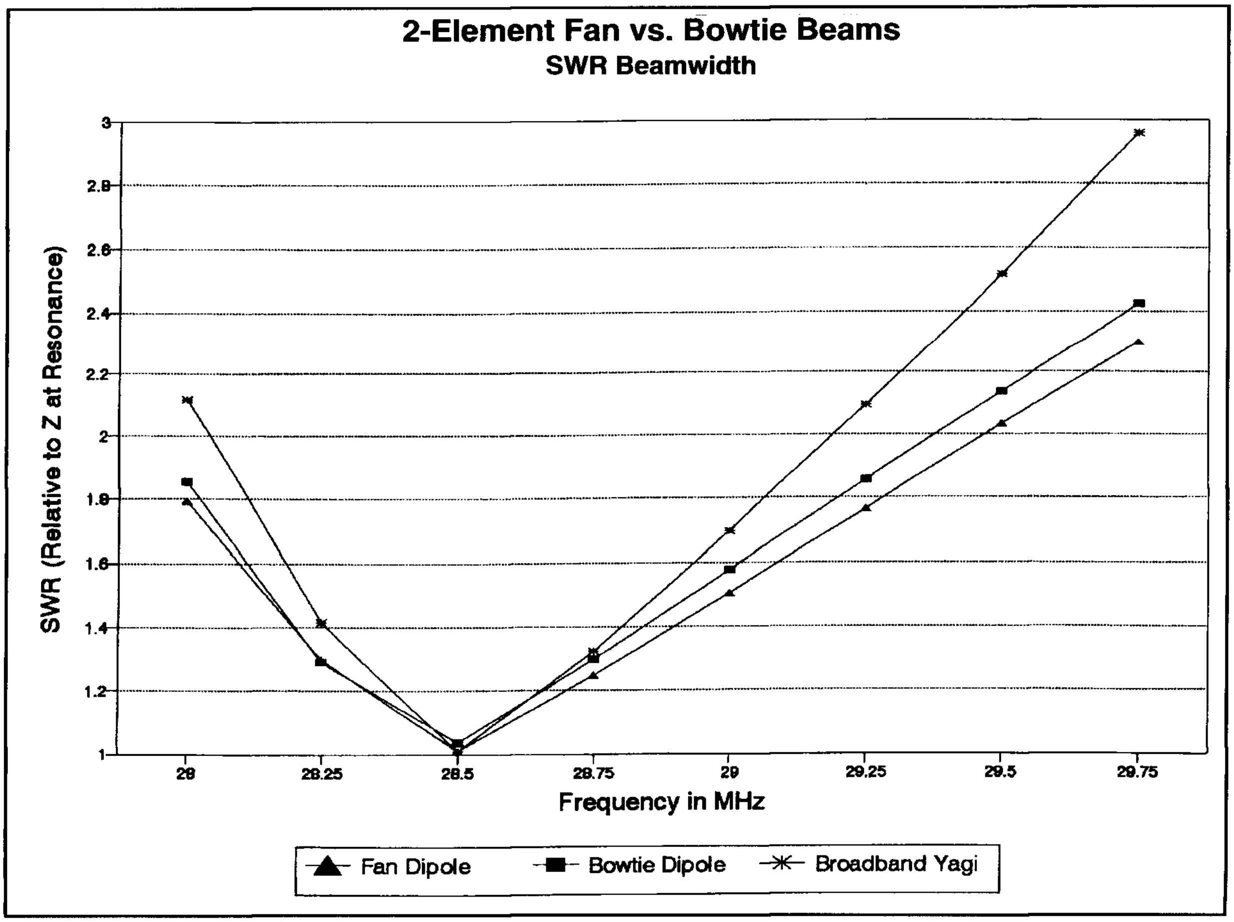

Figure 13. The SWR bandwidth across 10 meters of a fan, a bowtie, and a linear element Yagi.

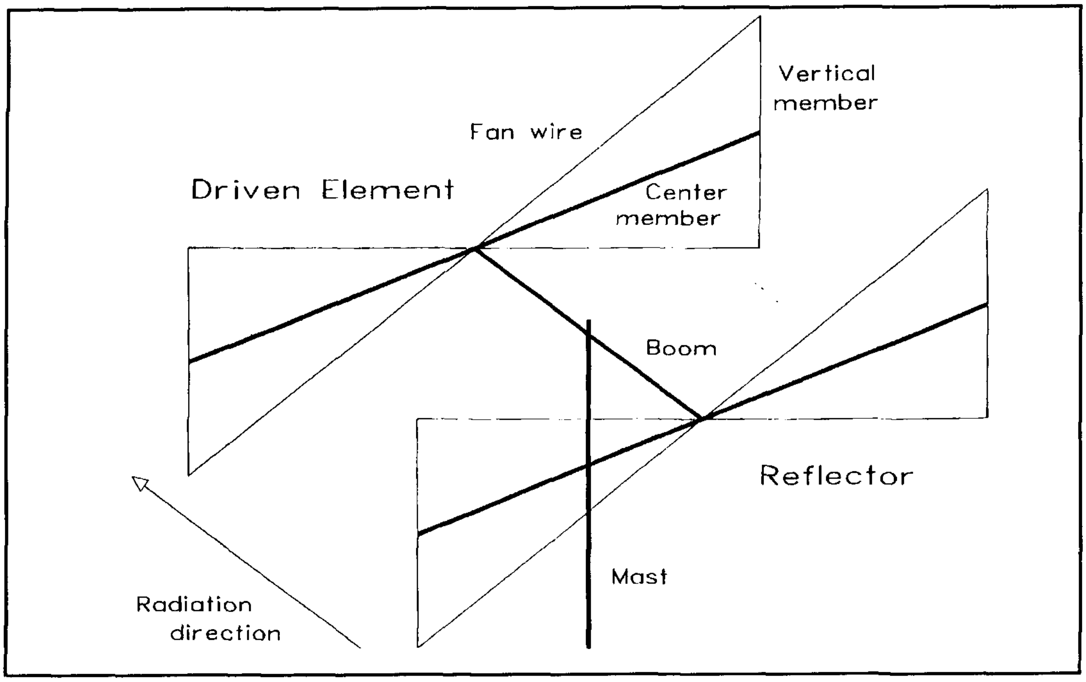

Figure 14. General outline of a two-fan-element Yagi, showing the names used in the text for the various mechanical components.

and element sizes are variable, with #14 wire, 0.375-inch diameter vertical members, and 1-inch diameter center members. Therefore, the antenna was remodeled using these figures to ascertain how close to reality the model would be. Additionally, the model was run at a height of 20 feet above real medium earth, as MININEC defines it (remembering that MININEC calculates feedpoint impedances as if over perfect ground).

A detailed model of a single fan element dipole showed resonance at about 28.7 MHz, with a feedpoint impedance near 50 ohms. Raising the antenna assembly with only the dri-

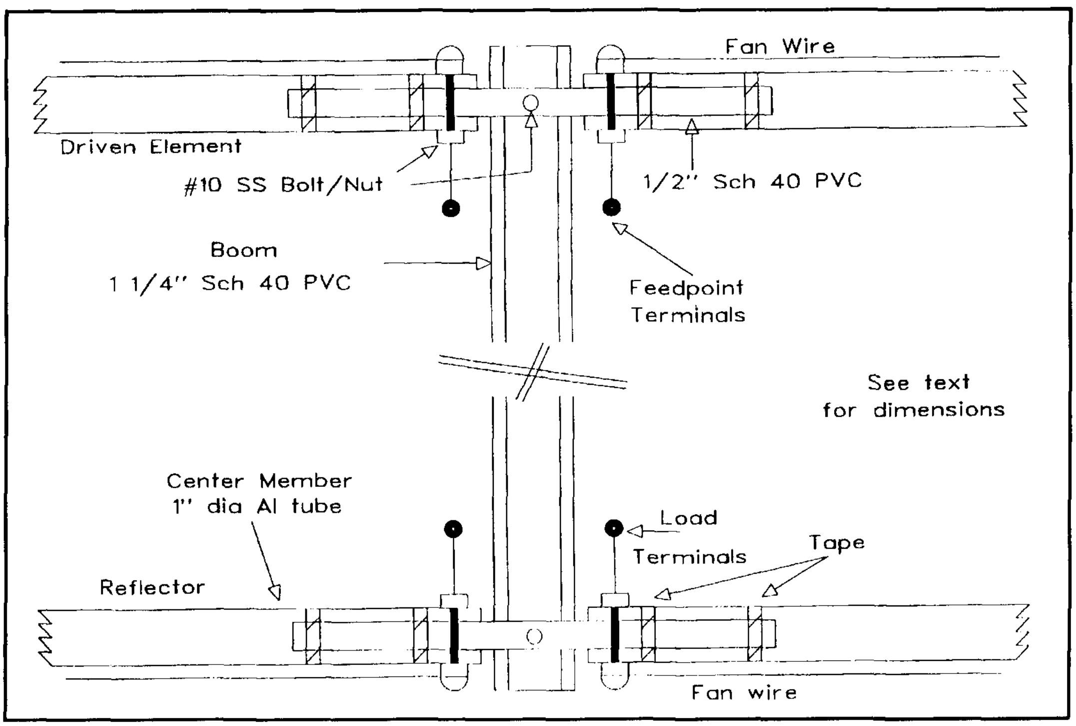

Figure 15. Construction details of the element-to-boom mounting for the fan elements of the test antenna.

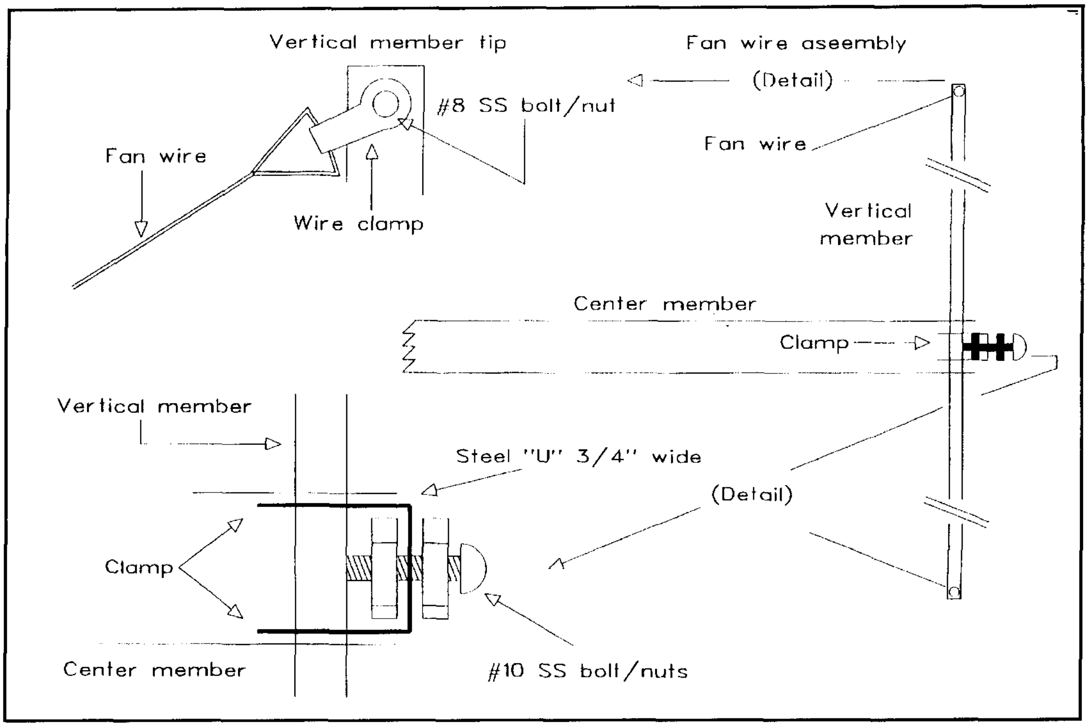

Figure 16. Construction details relating to the vertical member of the fan elements.

ven element mounted on the boom, I found a resonance closer to 28.5 MHz. The differences stem partly from limitations of the modeling program, as very small corners of the fans are cut off, despite extensive element tapering. The greatest part of the variance stems from the inevitable differences between the model and the constructed element. The 1.75-inch diameter boom adds that much to the overall element length, minus part of the 0.5-inch inset at the ends for the vertical members. Additionally, the vertical members are stressed, creating slight bows rather than truly straight vertical members.

Remodeling the Yagi version of the antenna suggested reflector loading of about 18 ohms for maximum front-to-back ratio, which translates into a 0.5-inch inside diameter coil with three turns space to just under a half inch in length. Alternatively, a 2.5-inch wide stub between 3 and 4 inches long, with a shorting bar, matched the terminal spacing of the reflector element of the test antenna. (See the **Appendix** for a utility program for calculating stubs.) The revised model also reported a feedpoint impedance of about 26+j24 ohms, values close to optimum for a beta match with a capacitor of about 110 pF as the parallel reactance. Both the reflector coil and beta-match capacitor would be subject to experimental variation to see how close to reality the model came.

As with most Yagis, the fan-element beam can be initially tuned near the ground by pointing it as straight up as possible, with the reflector toward the ground.[7] Adjusting the reflector load for minimum SWR should place the antenna at its maximum front-to-back ratio operating point. The predicted beta-match capacitor—110 pF—can be varied to further reduce the SWR. The value may be less than predicted if the series inductance in the driven element is less than 25 ohms.

Tune-up of the test antenna began with a 2.5-inch wide stub a bit over 4 inches long, with a shorting bar and a 100 pF beta-match capacitor. Reducing the stub to almost exactly 4 inches and reducing the capacitor to 75 pF produced a broad 2:1 SWR bandwidth between 28 and 30 MHz, with lowest readings (1.1:1) for 100 kHz around 28.75 MHz. Part of the below-predicted capacitance for the beta match stems from stray capacitance across the feedpoint assembly, part from a deficit of inductive reactance in the driven element as the frequency was lowered. At a 20-foot height, the best stub length was 3.5 inches with the same beta-match capacitor, and the region of lowest SWR was narrower. In addition, the center frequency shifted upward by 250 kHz. Although a coil across the reflector produced similar feedpoint readings, the stub proved more rugged and simple and is recommended for similar antennas.

Because the driven element's inductive reactance is marginal for a beta match at the lower end of 10 meters, the SWR curve is much steeper at those frequencies than the curves graphed in **Figure 13**. There are at least two

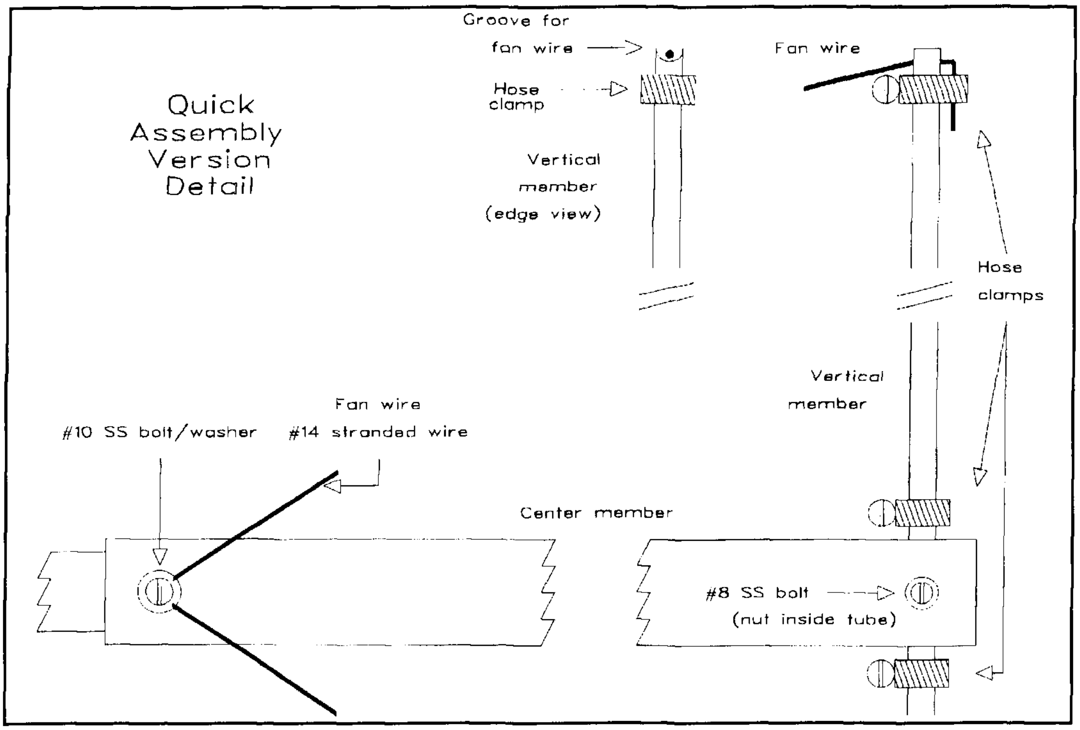

Figure 17. Construction details for a quick assembly version of the fan elements.

solutions to this problem. One is to use a different matching system. Another is to enlarge the elements by a couple of inches, either horizontally or vertically, in order to lower slightly their natural resonant points in the Yagi configuration. This latter measure would also reduce the amount of reflector loading required.

One interesting characteristic of the fan-element Yagi is its apparent greater susceptibility to coupling effects from surrounding objects than other linear antennas tested at precisely the same location. SWR variations as the antenna was rotated were greater than those experienced with any other test antenna. Because the antenna isn't a closed loop (as is the quad), but instead a pair of dipoles with broad vertical end surfaces, its tendency toward unwanted end coupling is no surprise. A fan-element beam should be used in as clear an area as possible.

Operationally, the antenna appears to be a typical two-element beam, similar in gain and front-to-back ratio as any other. It remains reasonably matched as one moves up 10 meters well past the point where gain and front-to-back ratio have diminished noticeably. In short, the fan-element beam behaves close to the predictions of the model, despite limitations on exact correspondence imposed by both MININEC constraints and construction deviance. Indeed, without a computer model with which to begin, I wouldn't have known just where to begin the process of shortening the elements as I spread their ends. The ham antenna builder doesn't require lab-grade predictions, but reasonable expectations for construction and testing.

Summary

A fan-element Yagi may be the most practical way to obtain full two-element Yagi gain from a 12-foot wide antenna without sacrificing SWR bandwidth. In fact, the bandwidth is improved, which is a plus for a band as wide as 10 meters. The 2-S-unit front-to-back ratio doesn't equal that of the narrower-bandwidth linear-loaded Yagi or the superb figure of the Moxon rectangle, both comparably sized antennas. However, as has been noted along the way, the final selection of antenna features and specifications requires attention to the major operating goals for a given station. For some operators, a high front-to-back ratio is the key element to success (if the gain among eligible antennas varies by only a dB or so), while for others, excessive front-to-back ratio can actually hinder operations.

The fan-element Yagi has its best home on 10-meters (and perhaps on 6 and 2, for some styles of operating). The smaller frequency range of the lower bands from 20 through 12 meters permits successful use of linear elements, even linear-loaded elements, with their simplified construction. However, for 10-meter operators whose interests require spanning the

Appendix: Calculation of Inductive and Capacitive Stubs

Inductive (shorted) and capacitive (open) transmission line stubs less than 1/4 wavelength long are very useful for introducing frequency-sensitive amounts of inductance (or inductive reactance) or capacitance (or capacitive reactance) into circuits. Their main use is in antenna work, where they serve to load antenna elements (electrically lengthen or shorten them). They also serve as inductors or capacitors in matching networks.

The simple GW BASIC utility program that follows allows you to calculate the length of either an inductive (shorted) or capacitive (open) stub by entering the frequency of interest and either a reactance value or a capacitance or inductance. Then you may enter either the known line impedance (Zo) or the physical dimensions (wire size and spacing in inches) of a feedline (along with its velocity factor). The program will then provide the stub length. The program likely replicates similar utilities written by large numbers of folks, but because I found none among my freeware stock, I wrote this one. I list it on the assumption that some few others might also need such a program. A visually enhanced version of the program appears in the latest version of HAMCALC by George Murphy, VE3ERP.[1]

Even number wire sizes that are greater than 3 and less than 19 (for example, 10, 14, 18) will be automatically recognized as AWG sizes. Decimals (0.25, for example) and whole numbers less than 3 are wire diameters in inches.

For somewhat greater accuracy, especially with large wires and close spacing, substitute the following for line 250:

```
250 SD=SP/DIA:ZO=120*LOG(SD+SQR((SD*SD)-1))
```

The chief error trapping is for division by zero whenever a potential denominator is accidentally entered as a zero. An arbitrary minuscule number substitution avoids the program break, but the results will be meaningless and a retry is necessary.

Program

```
10 'file XMSNSTUB.BAS
20 CLS:COLOR 11,1,3
30 CLS:PRINT"Calculation of Inductive and Capacitive Stubs":PRINT"L. B. Cebik, W4RNL"
40 PRINT:PRINT"For any value of Lx, Cx, L, or C, calculates the length of a transmission line
stub for a given frequency."
50 PRINT:INPUT " Enter frequency of interest in MHz",F:FB=F*(1000000!)
60 PRINT:PRINT" Select desired starting point:":PRINT"A. Inductive Reactance   B. Capacitive
Reactance":PRINT"C. Inductance   D. Capacitance"
70 PI=3.141592654#:T=0
80 A$=INKEY$:IF A$="A" OR A$="a" THEN 110 ELSE IF A$="C" OR A$="c" THEN 90
ELSE   IF A$="B" OR A$="b" THEN 140 ELSE IF A$="D" OR A$="d" THEN 120 ELSE 80
90 'Inductance to inductive reactance
100 INPUT" Enter Inductance in microHenries",L:LB=L*.000001: LX=((2*PI)*(FB*LB)):
T=1:GOTO 150
110 INPUT" Enter Inductive reactance in ohms",LX:T=1:GOTO 150
120 'Capacitance to capacitive reactance
130 INPUT" Enter Capacitance in picoFarads",C:CB=C*(1E-12):CX=1/((2*PI)*(FB*CB)):
T=0:GOTO 150
140 INPUT" Enter Capacitive reactance in ohms",CX:T=0:GOTO 150
150 'Stub construction
160 PRINT:PRINT" Select desired starting point for stub construction":PRINT"1. Feedline Zo
2. Line structure"
170 B$=INKEY$:IF B$="1" THEN 180 ELSE IF B$="2" THEN 200 ELSE 170
180 INPUT" Enter Zo of transmission line",ZO
190 INPUT" Enter Velocity Factor of transmission line",VF:IF T>0 THEN 280 ELSE 370
200 INPUT" Enter Line spacing in inches",SP
210 INPUT" Enter Wire diameter in inches or in even AWG",DIA
220 IF DIA>3 THEN IF DIA=4 THEN DIA=.257 ELSE IF DIA=6 THEN DIA=.186 ELSE IF
DIA=8 THEN DIA=.144 ELSE IF DIA=10 THEN DIA=.1019 ELSE IF DIA=12 THEN
DIA=.0808 ELSE IF DIA=14 THEN DIA=.0641 ELSE IF DIA=16 THEN DIA=.0508   ELSE
```

```
IF DIA=18 THEN DIA=.0403
230 IF DIA>3 THEN 210
240 IF DIA=0 THEN DIA=.0000001
250 ZO=(276*.43429)*(LOG((2*SP)/DIA))
260 INPUT" Enter Velocity Factor of transmission line",VF
270 IF T>0 THEN 280 ELSE 370
280 'Inductive stub
290 PRINT:PRINT" Inductive stub for a reactance of";LX;"ohms and a Zo of";ZO;"ohms"
300 IF ZO=0 THEN ZO=.0000001
310 LR=ATN(LX/ZO):LDG=(180*LR)/PI
320 IF F=0 THEN F=.0000001
330 LFT=(VF*LDG)/(.366*F)
340 PRINT"Length of Inductive stub =";LFT;"feet"
350 PRINT:PRINT" <A>nother run or <Q>uit?
360 C$=INKEY$:IF C$="A" OR C$="a" THEN 30 ELSE IF C$="Q" OR C$="q" THEN END
ELSE 360
370 'Capacitive stub
380 PRINT:PRINT" Capacitive stub for a reactance of";CX;"ohms and a Zo of";ZO;"ohms"
390 IF CX=0 THEN CX=.0000001
400 LR=ATN(ZO/CX):LDG=(180*LR)/PI
410 LFT=(VF*LDG)/(.366*F)
420 PRINT"Length of Capacitive stub =";LFT;"feet"
430 PRINT:PRINT" <A>nother run or <Q>uit?
440 C$=INKEY$:IF C$="A" OR C$="a" THEN 30 ELSE IF C$="Q" OR C$="q" THEN END
ELSE 440
```

NOTES

1. HAMCALC (Version 9.8 as of this writing, but undoubtedly higher by now) is available from VE3ERP, 77 McKenzie Street, Orillia, Ontario L3V 6A6. Murph requests a $5 donation to cover disks, mailers, and postage, and he donates the excess of donations over costs to a Canadian institute for the blind which offers amateur radio services. The program includes a structured menu system, its own copy of GW BASIC, and nearly 100 handy calculation programs of interest to hams. They range from basic Ohm's law computations to more complex antenna, transmission line, inductor, and basic matching system calculations, with tables of wire, cable, and other construction materials, as well as many more features.

band from end-to-end, fan elements are worth serious consideration.

The vertical dimension of the fan-element Yagi, especially the commercial Butternut HF5B Butterfly, approaches that of a small quad. A two-element quad has the further space-saving benefit of reducing the horizontal dimension to about half that of a linear Yagi. Some experimenters have tried to reduce the quad dimensions even further. This series would be incomplete if I did not take a good look next time at shrunken quads. ■

NOTES

1. NEC-2, while having advantages over MININEC in many respects, has difficulty in dealing with complex geometries involving wires of significantly different diameters. The models of fans, bowties, and their two-element Yagi counterparts do not tax the limits of MININEC. All patterns and figures graphed were derived from ELNEC 3.02. Initial models used all #14 wire, while subsequent models—especially of antennas based on the fan—replaced the center member with a 1-inch diameter tube in order to model the practical antenna actually built. Because the antenna elements have many right and acute angles, elements were extensively segment-tapered. The resulting Yagi models required 198 segments for fan versions and 190 segments for bowtie versions.
2. I have discovered that some hams believe that a half wavelength dipole shows maximum gain at its resonant frequency. This belief likely represents a confusion between the gain of the main lobe and our ability to get the antenna to perform due to the increasing SWR off its resonant frequency. The bidirectional lobes of a dipole show reduced gain (in dBi) as they are shortened below resonance at a frequency of interest and more gain as they are lengthened. The gain increases with wire length until the antenna is about 1.25 wavelengths long (the EDZ), after which the main pattern lobes begin to decrease as the formerly minor lobes increase.
3. For this reason, fans have not found widespread use on the lower bands. On 80 meters, a better choice may be dual dipoles—one cut for the CW end of the band, the other for the phone end. The dipoles may be connected at the feedpoint or you can use close-coupled spacing for similar broadband results.
4. Although not an insignificant task to assemble and tune, the HF5B performs to specifications supplied by the maker. My 6-year-old model has survived a local area house move (thanks to a very careful mover) and continues to perform as specified. Although its dimensions are a bit larger than the fan Yagi given here in order to accommodate multiband operation, the principles remain the same. Butternut records in their instruction manual that "BUTTERFLY" is a trademark, although the status of the lower-case version of the word is uncertain. (In any event, we shall not have to revert to the original term "flutterby" when referring to certain showy insects.)
5. See the **Appendix** for a short utility program to calculate transmission line stubs from required capacitive or inductive reactances. The program also permits use of capacitor or inductor value inputs and allows for commercial or home-constructed transmission lines.
6. See Thomas Cefalo, Jr., WA1SPI, "The Hairpin Match: A Review," *Communications Quarterly*, Summer 1994, pages 49–54; L. B. Cebik, W4RNL, "Technical Conversations: Further Notes on the Beta Match," *Communications Quarterly*, Winter 1995, pages 51–54; Gooch, Gardiner, and Roberts, "The Hairpin Match," *QST* (April 1962), 11–14, 146, 156; and Jerry Hall, K1TD, editor, *The ARRL Antenna Book*, 16th Edition, pages 26-21–26-23.
7. See Brian Beezley, K6STI, "Adjusting HF Yagi Matches," "Technical Correspondence," *QST*, April 1995, page 74.

Compiled by Peter Bertini, K1ZJH
Senior Technical Editor

Antennas for Dxing

Here are two antennas for the Dxers among us. Those who can't put up a up a beam or who want a compact spotter antenna might like to try K1BQT's Dx-pole. Readers interested in building a DX receiving antenna for the low bands should check out W8JI's reversible beverage antenna.

—K1ZJH

Build a 20-Meter DX-pole Antenna

This no-radial vertical for HF is patterened after the VHF discpole antenna.

Rick Littlefield, K1BQT

The DX-pole is a no-radial vertical for HF patterned after the VHF discpole antenna.[1,2,3] With a vertical element measuring just under 17 feet and weighing in at 4-1/2 pounds, this compact monobander mounts easily on a sturdy chimney support or RadioShack telescoping mast. Like its VHF and UHF predecessors, the DX-pole is a capacitively loaded, off-center-fed (OCF) antenna designed to deliver full-size half-wave efficiency and performance from a quarter-wave radiator. If you can't put up a beam, or if you want a compact spotter antenna, the DX-pole might be for you.

Theory of Operation

According to theory, electromagnetic radiation is induced when RF current travels over linear distance along a conductive element. For a full-size dipole where L = l/2, a moderate amount of current is distributed over a relatively long conductive path (**Figure 1**). For a loaded dipole where L = <l/2, proportionately more current is distributed over a shorter conductive path. Theoretically, a loaded dipole may induce as much radiation as a full-sized dipole, as long as RF current is allowed to flow in a straight line and losses are minimal.* To satisfy those conditions, the DX-pole's radiator is straight, round, and very low in resistive losses along its entire length.

Whether loaded or unloaded, current distribution is seldom uniform along the length of a half-wave dipole (an exception being when the dipole is *extremely* short, as in a small magnetic loop antenna). If the radiator is balanced, RF current is highest at mid-element, and RF voltage is highest at the ends. End segments primarily support charge storage and radiate very little electromagnetic energy. The DX-pole takes advantage of this characteristic by replacing end tubing with symmetrical capacitive loading structures. These structures serve three important functions. First, they reduce element length by approximately 50 percent through capacitive loading. Second, they lower energy

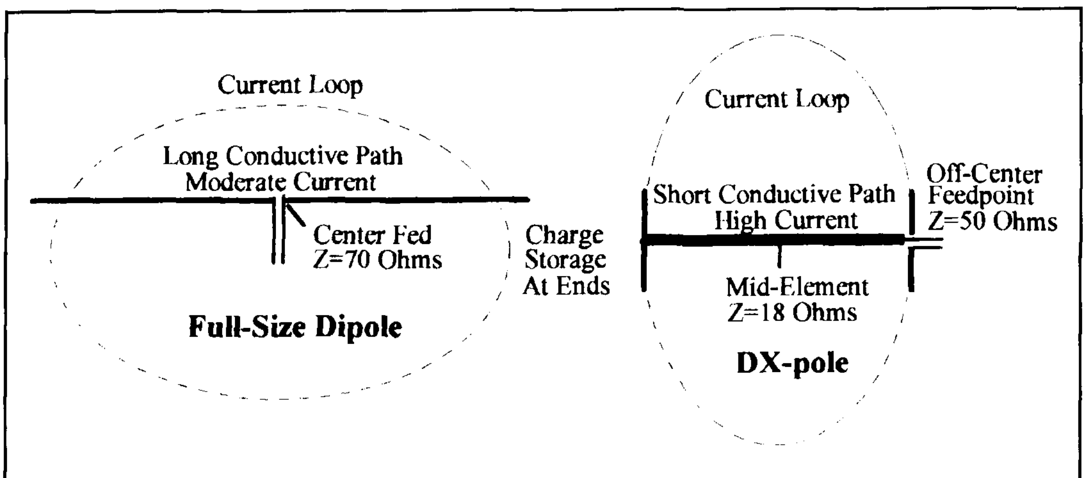

Figure 1. Current distributed over a conductive path for a full-size dipole and the DX-pole.

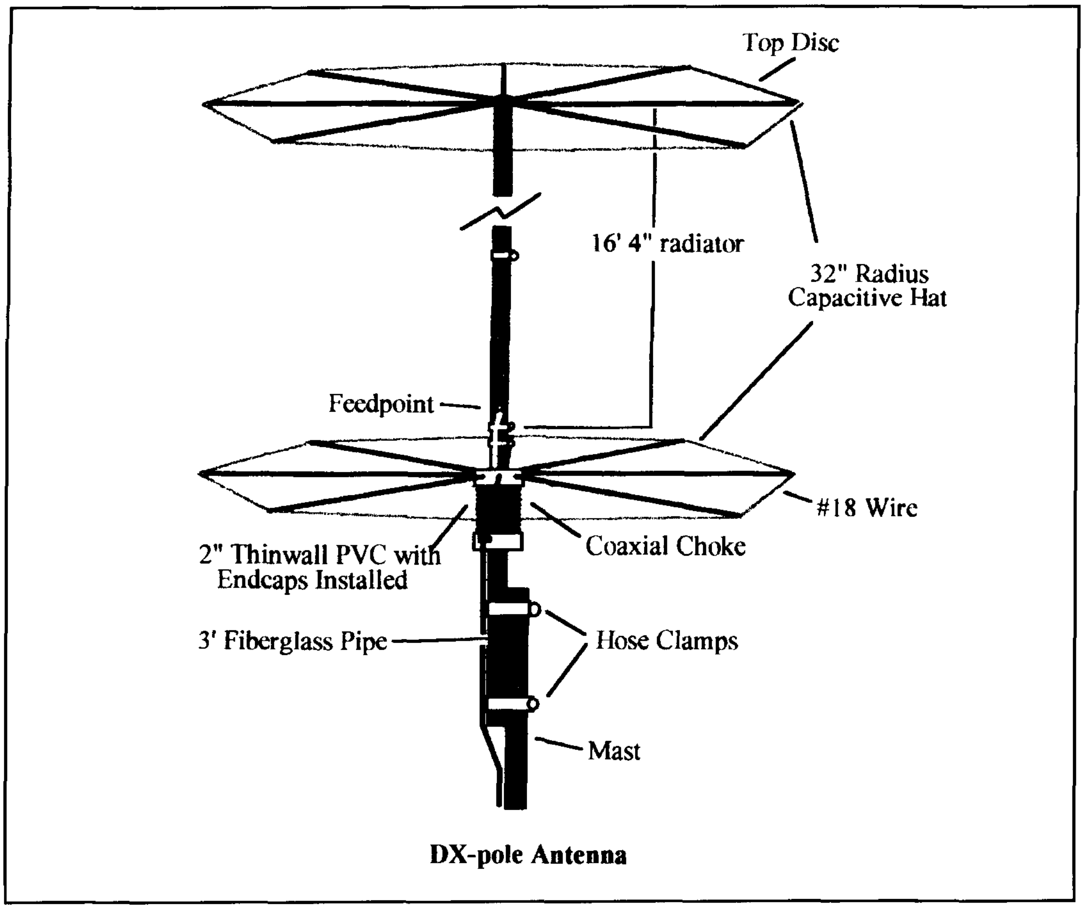

Figure 2. The DX-pole antenna.

storage around the element—a modification that recovers much of the bandwidth normally lost to shortening effects. Finally, because of their size relationship to the vertical element, they provide a convenient 50-ohm off-center feedpoint (OCF) at the element junction. Of all loading methods, capacitive-end loading is preferred because it contributes the least amount of loss to the system. Also, as symmetrical opposing end structures, capacitive hats do not radiate. Virtually all electromagnetic radiation is confined to the vertical radiator.

Because the DX-pole is an OCF design, its feedline must be carefully decoupled from the element to prevent detuning and unwanted common-mode radiation. A self-resonant solenoid-type coaxial choke made from 50-ohm feedline provides near-complete isolation.

Construction

The DX-pole base is constructed around an insulated 3-foot length of 1-1/4 inch fiber glass pipe (see **Figure 2**). This material is sold in 8-foot sections at most home-supply outlets. The choke balun and lower capacitive hat are built on a PVC form that slips over the pipe.

To build the form, cut a 6-inch section of 2-inch thin wall PVC pipe and install PVC end-caps at each end. Glue the bottom end-cap in place with Genova Cement™. The top end is removable, and will be secured later with three #6 sheet metal screws. Cut a 1-1/4 inch clearance hole in the center of each end cap. My lower element section was the same ID as the fiber glass tube, so I cut a 12-inch length of 1-inch OD wooden dowel to serve as a transition. If needed, wrap the top part with a layer or two of electrical tape to ensure a snug fit. If your lower element section slips inside the fiber glass tube, you don't need the dowel. Either way, use two hose clamps to reinforce the top of the fiber glass tube; this will prevent it from splitting under lateral stress from the element.

My vertical element is made from three scrap lengths of telescoping aluminum tubing (1-1/4, 1-1/8, and 1-inch OD). Use whatever you find in your junk pile (free), at the local scrap metal yard (cheap), or from the hardware store (expensive). Slot one end of each tube and fit it with a hose clamp, then join each section to make up a 16-foot radiator.

To construct the top capacitive hat, cut a 6-inch length of 1-1/8 inch tubing to serve as a

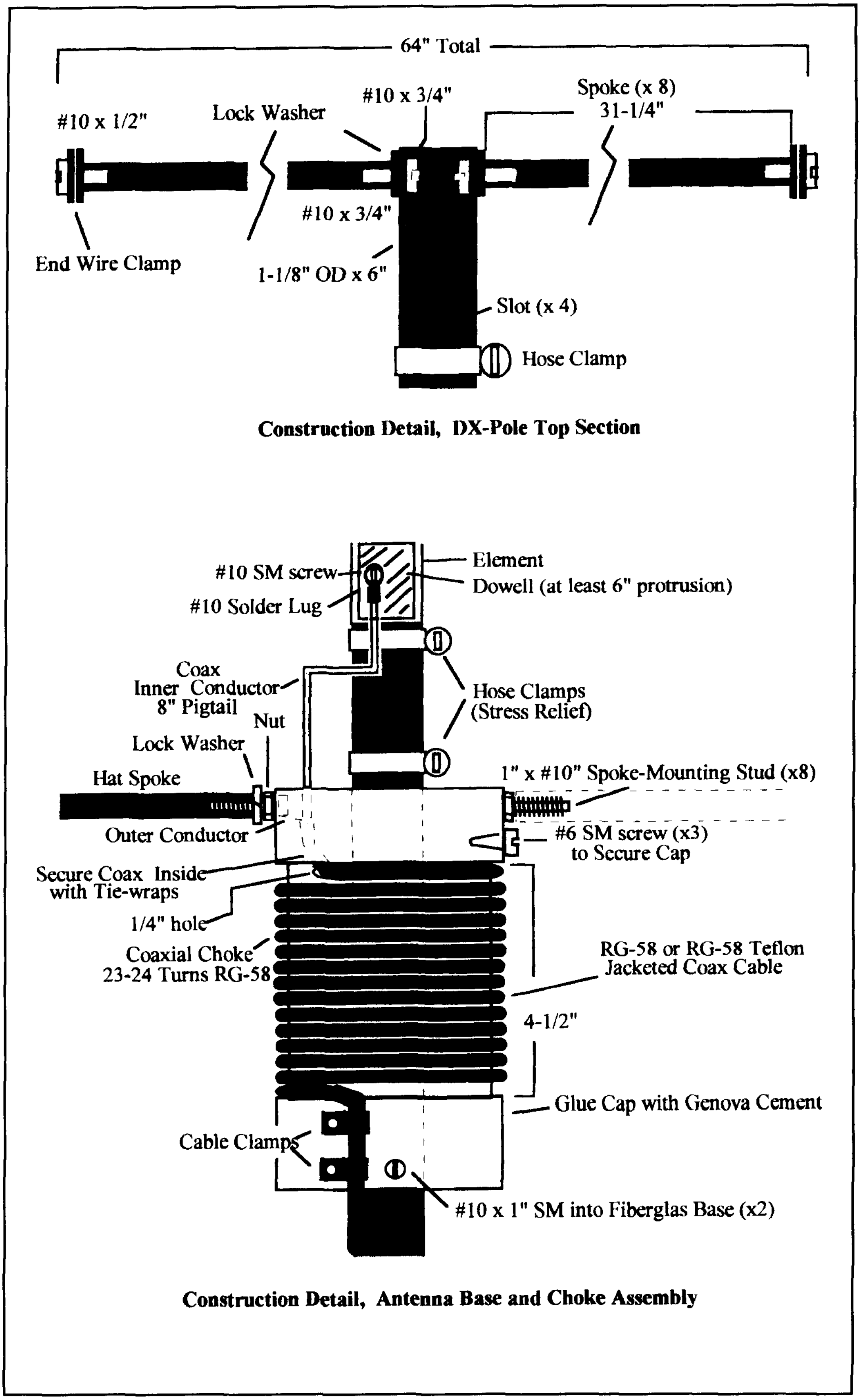

Figure 3. (A) Construction detail, DX-pole top section. (B) Construction detail, antenna base and choke assembly.

hub (**Figure 3A**). Calculate the circumference of the tube (D) and divide by 8 to obtain the spacing between each spoke. Mark off each location 1/2 inch in from the end, then drill eight equally spaced holes for #10 hardware. To complete the hub, slot the opposite end and install a hose clamp. The spokes consist of eight lengths of 1/4-inch OD tubing cut to 31-1/4 inches in length. Each is tapped at both ends for #10-24 hardware. Secure one end of each to the hub by inserting a 3/4-inch screw from the inside out. Use lock-washers to prevent loosening. Install a 1/2-inch screw and two flat washers at the opposite end of each spoke.

The wire portion of the hat is made from #18 soft-drawn bare copper available at most hardware stores. When installing, string the wire straight between each spoke tip, but not taut enough to distort the spoke. Wrap one turn around each screw and move onto the next spoke, tightening each clamping screw as you go. Pull the wire tight when making the last connection. When done correctly, the wire will pull up tightly all around—without obviously bending or distorting any of the spokes. The radius of the completed hat should measure 32 inches.

The bottom capacitive hat is built in similar fashion (**Figure 3B**). To start, remove the top end cap, measure the outer circumference, and divide by 8. Now mark off eight equally spaced radial positions 1/2 inch down from the top of the cap, and drill for #10 hardware. Since PVC is an insulator, the spokes must be electrically bonded. Use a 1/2-inch wide strip of light-gauge aluminum strap for this, wrapping it around the inside of the end-cap and marking each hole location. When marked, remove it from the cap, cut off excess length, and drill eight holes for #10 clearance. Now, reinstall with 1-inch #10-24 screws and lock-washers in each hole—around the circumference of the cap. Secure each screw on the outside with a nut to form eight mounting studs. Cut the bottom spokes to 30-1/4 inches and tap each end. Set these aside for later installation.

For transmitters up to 100 watts, RG-58 coax is adequate for the balun. However, for power ratings greater than 100 watts, use a Teflon™-jacketed coax, such as RG-58 plenum cable. I used a 21-foot length to wind a 24-turn choke, but 12 turns should be sufficient for series resonance at 14 MHz. Feed the prepared end into the form hole and secure in place with a couple of tie wraps. Thread the center conductor up through the top and install a #10 solder lug. Cut the shield to about 1 inch and install a #10 solder lug. Connect this to the bonding strap at one of the spoke studs. Wind the coax tightly down the form, securing at the base with wire clamps. Finally, install a connector on the end of the coax.

To complete the base section, install and tighten the eight spokes. String the outer wire using the same technique described for the top. Now, slip the form onto the fiber glass pipe, as shown, and lock in place with two #10 x 1-inch sheet-metal screws (drill small pilot holes in the fiber glass before installing).

Finally, install the element and top hat section onto the base—and connect the coax center conductor to the main element using a #10 x 3/4-inch sheet metal screw. When completed, the aluminum element and top-hat should measure 16-feet, 4-inches (plus the length of your coax center lead). Each hat should have a total diameter of approximately 64 inches. When installing the antenna, leave at least 4 inches of spacing between the bottom of the choke and the top of the metallic support mast. Secure the antenna to the mast with three hose clamps rather than with TV hardware: TV clamps will crush the fiber glass.

Setup and Performance

I tuned my prototype with its base approximately 35-feet above ground, supported by a RadioShack telescoping mast that was wall-mounted against the back of a single story building. When tuned, the antenna's resonant frequency (Fr) fell at 14.1 MHz with minimum VSWR too low to measure on my MFJ-259 Analyzer. In terms of bandwidth, the 1.5:1 VSWR points occurred at 13.8 MHz and 14.35 MHz for a commercial-standard bandwidth of just under 4 percent. Its 2:1 VSWR bandwidth measured 7.3 percent, extending well beyond the margins of the band. For all intents and purposes, VSWR was flat across the CW and Extra class phone subbands. The DX-pole is a balanced antenna and should never be mounted with one end too close to ground. Indeed, lowering the antenna's base to 10 feet above the roof increased VSWR at resonance to 1.2:1 and moved Fr by about 30 kHz—indicating that the building and its surroundings were beginning to disrupt the antenna's electrical balance and tuning (closer proximity to ground will pull the current loop off-center toward the antenna's base and severely detune the element). Because of this, it's important to mount the antenna as high as safety considerations permit, keeping it away from wires, foil-backed insulation, metallic guys, or metallic masses.

To give the DX-pole a good, though perhaps unfair, test, I installed a horizontal dipole at the opposite end of my yard. The dipole was positioned at 65 feet (10 feet higher than the top of the DX-pole) and broadside to Europe and South America. I routed both feedlines to my radio through a two-position switch, allowing for instantaneous signal comparisons. The results were interesting but inconclusive. For example, a JA at S3 on the dipole became unreadable on the vertical, while just 5 kHz away a VK at S5 on the vertical literally disappeared when I switched to the dipole. And so it went, up and down the band—a 5M7 was stronger on the vertical while a YV5 was stronger on the dipole. Overall, the dipole seemed to do a little better on more signals, but it was hard to tell. In reality, anecdotal comparisons such as these prove little and make for poor science. Faraday rotation and other uncontrolled variables work constantly to skew the outcome in unpredictable ways. However, the vertical performed nicely and I made a lot of QRP-SSB contacts with it.

Conclusion

The DX-pole provides a simple demonstration of what can be accomplished by a small antenna when you apply capacitive loading techniques and pay attention to resistive losses. The hat structure makes construction a bit more complex than it might have been for a simple dipole or inductively loaded vertical. However, I think this is effort well spent because capacitively loaded antennas tend to be more efficient and exhibit greater bandwidth than their inductively loaded cousins. If you build the DX-pole carefully and install it in the clear, you'll be rewarded with a scrappy little 16-foot vertical that can take on antennas twice its size—and win!

REFERENCES

1. Rick Littlefield, "The 2-Meter Discpole Antenna," *Communications Quarterly*, Summer 1996, pages 77–81.
2. Rick Littlefield, "Build the Six-Meter Discpole Antenna," *CQ VHF*, March 1997, pages 44–50.
3. Rick Littlefield, "The Discpole Antenna," *RF Design*, publication pending.

FOOTNOTE

* NEC-based gain predictions for the discpole and DX-pole vary from 1.8 to 2.5 dBi in free space, depending upon the investigator and specific version of the program used. Range tests conducted by the author at 146 MHz using a MFJ-224 field-strength analyzer indicate no measurable advantage or disadvantage between the discpole and a half-wave reference antenna of known performance.

A Practical Reversible Beverage

Improve your DX capabilities on low frequencies with this popular antenna

Tom Rauch, W8JI

A Beverage is a longwire antenna installed very close to the Earth. To perform as a true Beverage, the antenna must be longer than 1 wavelength and installed at a height of less than 0.05 wavelength above Earth. The Beverage is usually terminated in a resistance at the end opposite the feedpoint. This termination ensures that traveling waves (as opposed to standing waves) appear on the antenna. The reduction of reflected waves from the far end of the antenna produces a unidirectional pattern.

Beverage antennas provide one of the least expensive and most reliable ways to improve DX capabilities on low frequencies. Even though their low height greatly increases losses (making them very inefficient antennas), their directivity can be used to advantage for receiving. Because of their simple construction and predictable performance, Beverage antennas have become one of the most common DX receiving antennas used on the lower amateur bands. With careful attention to design and construction, a single Beverage antenna can be made to cover two directions with excellent performance over a very wide frequency range.

Unterminated Beverages

A Beverage's front-to-back (F/B) ratio is equal to the one-way attenuation or "signal loss" (in decibels) over the length of the antenna. If the Beverage antenna had no loss or attenuation along its length, and wasn't terminated, it would provide a true bidirectional pattern.

Unterminated Beverages always exhibit some front-to-back ratio. The inherent F/B ratio in an unterminated Beverage is caused by attenuation of signals traveling the length of the antenna's conductor. The unavoidable reduction of signal along the length of the antenna is mostly due to ground-induced losses and radiation effects.

The actual attenuation (and unterminated F/B ratio) is dependent on antenna conductor size and resistance, height, length, and soil conditions below and around the antenna. Measurements of several Beverages, installed six to eight feet high over various soil types, have indicated one-way losses of approximately 6 dB per wavelength on 160 meters. Because the unterminated F/B ratio is equal to the one-way power "loss" in decibels, these antennas typically provided unterminated F/B ratios of 6 dB.

The Beverage's inherent distributed attenuation (signal loss) also limits performance improvements as the antenna is made longer. In properly terminated systems, directivity (and signal-to-noise ratio, or S/N ratio) generally stops increasing as antenna length is extended beyond one or two wavelengths.

Terminated Beverages

The use of a termination resistance greatly improves front-to-back (F/B) ratio by ensuring that all waves traveling toward the termination are absorbed. Proper termination removes or attenuates signals and noise arriving from the unnecessary "back" direction without affecting

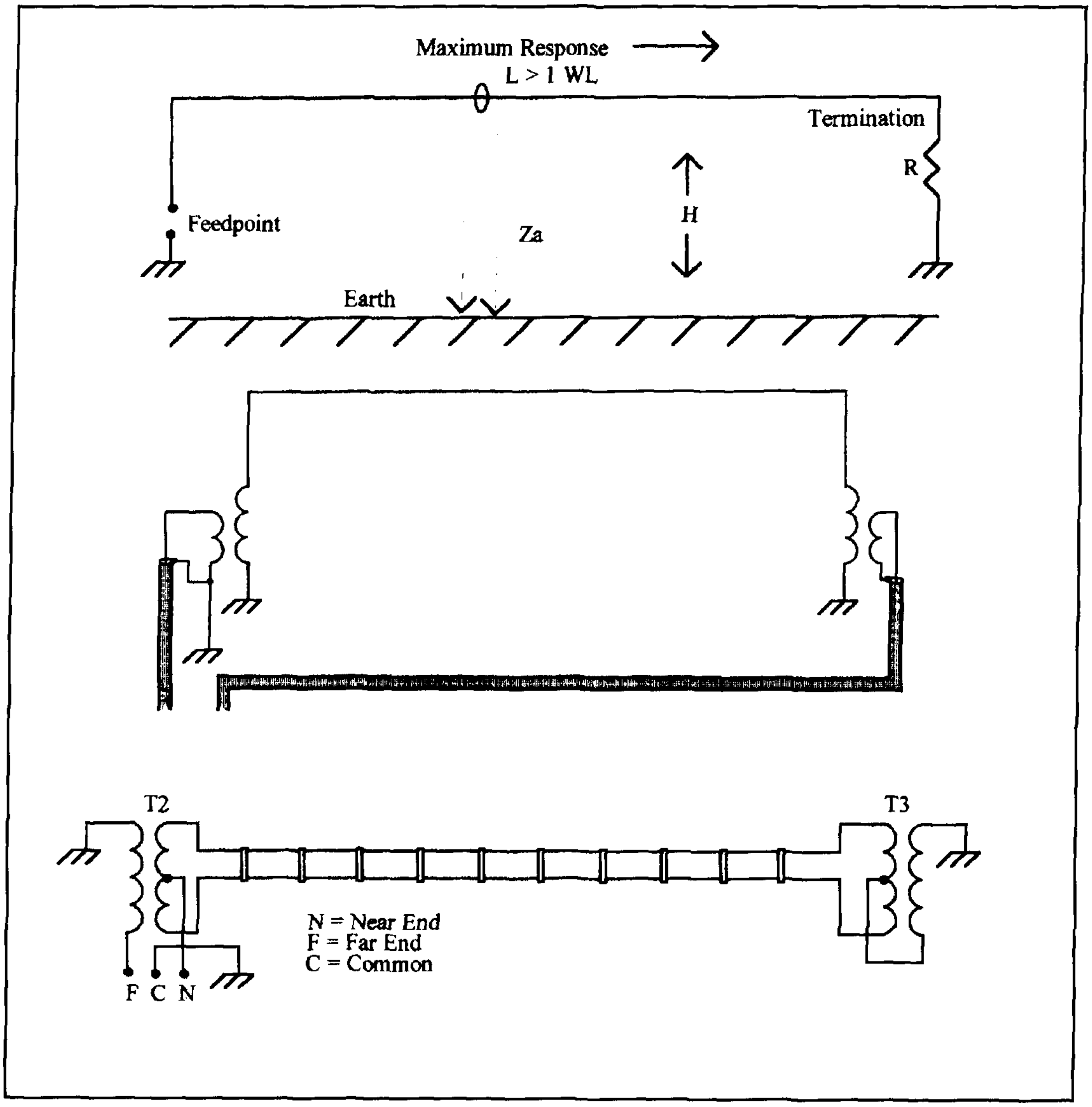

Figure 1. (A) Beverage antenna terminated at the far end. (B) Addition of a second feedline allows the Beverage to be terminated at either end. (C) The two-wire reversible Beverage.

sensitivity in the desired direction. Proper termination almost always improves signal-to-noise or signal-to-interference ratios.

The termination resistance is generally made equal to the antenna's common-mode surge impedance (Za) in **Figure 1A**. Although a perfect termination requires the compensation of reactive components in the antenna, the reactive components of Za are generally so small that they are ignored.

Figures 1A through **C** show the progression from a typical single wire unidirectional Beverage to a two-wire reversible Beverage. In **Figure 1A**, the Beverage is terminated at the far end. The termination reduces response in the direction opposite the end of termination; in other words, it reduces response in the direction of the feedpoint. Termination affects forward sensitivity very little, but can greatly improve S/N ratio if the noise is coming from the feedpoint's direction.

Reversible Beverages

Figure 1B shows the addition of a second feedline, allowing the Beverage to be terminated at either end. The transformers are designed to match the Beverage's impedance (Za) to the coaxial feedline's impedance. The direction of maximum response reverses whenever the receiver and termination resistance's feedline connections are exchanged. In a system of this type, termination resistance is made equal to the coaxial feedline's impedance. The transformers step the termination impedance up to the proper value, just as they step the receiving

Table 1. Reflection Transformers

Total Turns		
Primary	**Secondary**	**Impedance Ratio**
6	4	0.44
5	4	0.64
7	6	0.73
4	4	1.00
5	6	1.44
3	4	1.78
4	6	2.25
Zb/Za		

end's impedance down to match the cable.

Figure 1C is the next generation of the reversible antenna: the two-wire reversible Beverage. In a two-wire reversible Beverage antenna, the antenna element performs two very different functions. The two-wire antenna element not only functions as a simple longwire antenna, it also doubles as a balanced transmission line. This balanced transmission line provides the connection to the far end of the system.

A two-wire reversible Beverage requires the antenna element to operate a balanced transmission line, as well as a conventional longwire antenna. This means the parallel wire line used to construct the antenna must be excited under near-perfect balanced transmission line conditions, without disturbing normal antenna mode operation of the system. This is accomplished by exciting the two-wire transmission line making up the Beverage's element with equal 180 degree out-of-phase currents. Multi-frequency operation requires standing waves be minimized, so the impedance presented to the transmission line by the transformers must match the line's differential (transmission line) mode impedance. Any sacrifice of these parameters affects operation, and will generally result in a noticeable reduction of system performance.

The simultaneous requirement of operating the two-wire transmission line (making up the antenna's element) as a simple longwire antenna while it acts as a transmission line, means that the antenna's conductors must be excited by equal in-phase currents without disturbing transmission line (out-of-phase) currents. These separate functions are accomplished by connections made to the antenna's conductors through T2 and T3 (**Figures 1C** and **2A**). These transformers must be carefully designed to precisely isolate the two distinct antenna functions. At the same time, they must match system impedances properly and have low loss. The transformer primaries provide the transmission line (differential) mode out-of-phase connections, while the transformer secondary center taps provide the antenna mode (parallel or common mode current) in-phase connections.

The connections around T3 are of particular interest, especially the one at point "A." This is where the antenna mode connection (taken from the center tap) feeds directly into the differential or transmission line mode connection (the primary) of T3. The primary then excites the center tap of T3 with a differential mode signal, allowing the antenna's remote end to be accessed via its differential mode connection that eventually appears at point "F." Point "N" (the center tap of T2) provides the antenna mode connection at the feedpoint end of the system.

Simply grounding one wire of the balanced two-wire line at the far end won't produce correct termination, or the necessary differential mode currents for proper operation. "Self-reflecting" Beverages, or systems using poorly designed transformers, provide inferior performance when compared to systems using well-constructed reflection transformers.

Transmission line selection

The antenna mode impedance (Za) of a Beverage antenna is set by the effective conductor diameter and height. Za is typically in the range of 400 to 500 ohms. The transmission line used to construct the Beverage can easily be made any standard impedance value, but the logical impedance choice for the Beverage element is 450 ohms. A 450-ohm transmission line element allows use of simple 1:1 transformers in the system.

The advantages of prefabricated 450-ohm line are clear:

1. The line making up the Beverage element is available off the shelf from many vendors.
2. The line is easy to install, repair, or replace.
3. The line operates with a very low standing wave ratio using simple 1:1 "reflection" transformers.
4. Undesirable signal loss and unwanted signal ingress in the transmission line is minimized by the line's construction.

While 450-ohm line is readily available, it's important to use caution during the selection process. At least one line using 1-inch spaced heavy-duty #14 conductors is advertised and sold as a 450-ohm line, even though its impedance is just 370 ohms. Be sure you know the

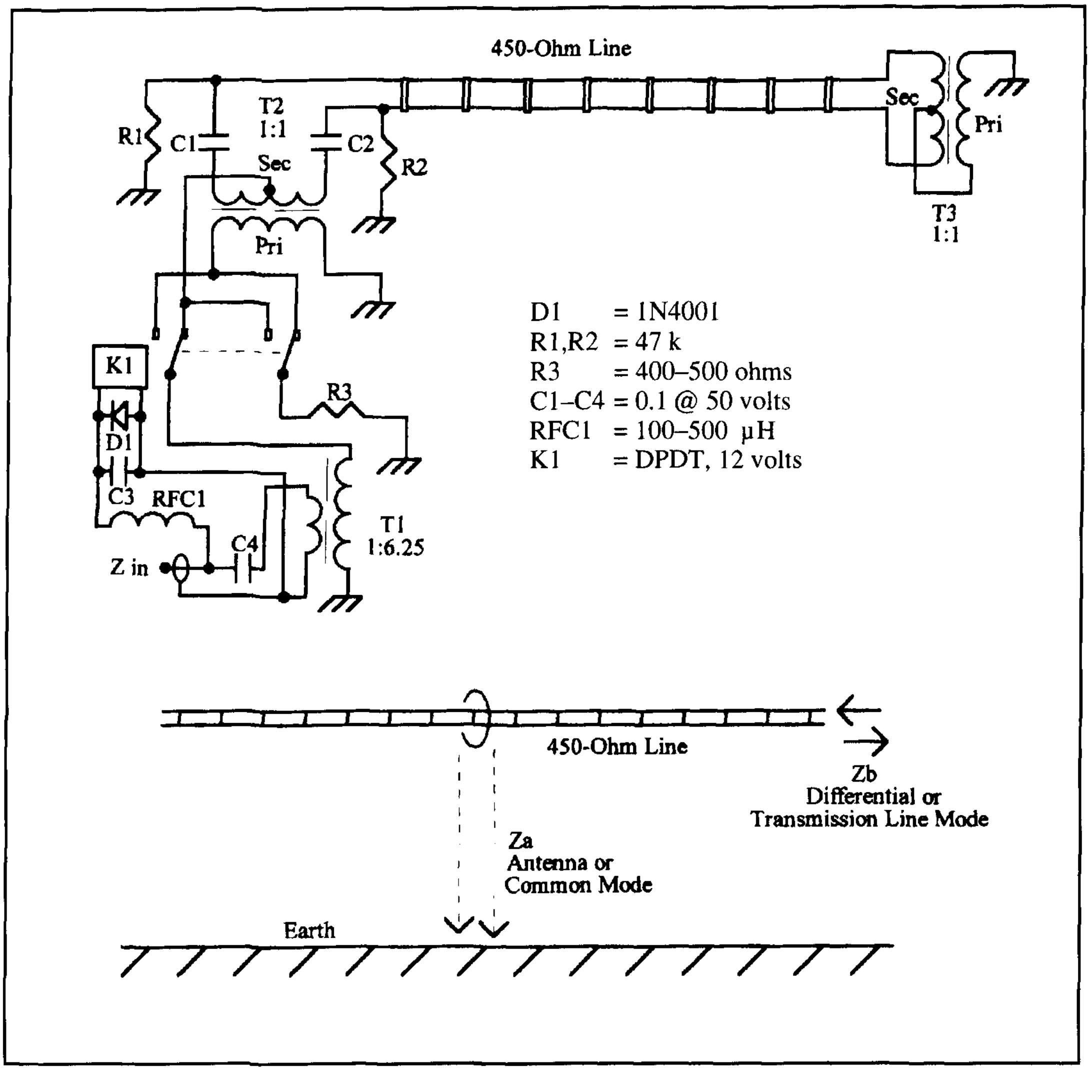

Figure 2. (A) Diagram of a typical reversible Beverage system. (B) Feedline (Zb) and Beverage common mode (Za) impedances.

real impedance of the line you select.

System construction

Figure 2A shows a diagram of a typical reversible Beverage system. Notice that the primary (receiver end) of T1 isn't connected to the Beverage antenna ground. T1 functions as an isolation and impedance matching device isolating the feedline from the antenna's ground. The floating primary ensures that equal and opposite currents excite the feedline. This confines all current to the inside of the cable. It also prevents common mode currents flowing on the feedline from being conducted into the Beverage's ground connection, reducing system noise. The feedline is grounded to a separate ground rod several feet from the Beverage's ground connection.

Proper feedline treatment reduces unintentional coupling of noise and unwanted signals into the Beverage system. If the feedline is routed near a transmitting antenna or a noisy environment, or if the receiver is located in a house with noise sources, additional decoupling of the feedline is advisable. A separate ground rod should be installed 50 to 100 feet from the Beverage, and a string of ferrite beads with a few hundred ohms of total impedance can be installed over the feedline on both sides of the ground rod. The combination of beads and a ground rod improves common mode isolation by providing both a high series impedance and low shunt impedance for common mode signals flowing on the shield.

If the ground system is nonsymmetrical, or is constructed using a few long counterpoise wires, it can actually behave like a small separate receiving antenna and diminish front-to-back ratio and noise performance. It's better to use several ground rods spaced several feet apart, rather than one ground rod or a few long radials or counterpoise wires. Because impedance Za of the Beverage is several hundred ohms, even 20 to 30 ohms of RF ground resistance is sufficiently low.

Transformer design

I prefer to use CATV-type cables, such as RG-6 or RG-59, because connectors are inexpensive and easy to install. A variety of hardware, such as ground blocks, is also available. For this reason, and because I also use 450-ohm ladder line about eight feet above ground to build my antennas, the transformers I use are designed for 75-ohm feedline with 450-ohm impedances for Za and Zb.

If you prefer to use other impedances, **Table 1** gives turns ratios for various feedline mode (Zb) and Beverage common mode (Za) impedances (**Figure 2B**). The secondary windings of T2 and T3 must have an even number of turns, allowing the center tap to be placed in the electrical center of the winding. This winding is a bifilar pair wound separately from the primary-ensuring balance and symmetry.

Three transformers are required. They are made from strings of FB73-801 or Fair-Rite 2673000801 beads. These beads are approximately 0.3-inch OD and 0.3-inch long, with a 0.1-inch center hole (window). A stack of five beads is used to make each transformer. The window is large enough to accept up to 12 passes (turns) of #26 wire.

The 75- to 450-ohm impedance matching transformer (T1) in **Figure 3A** uses two passes of #26 AWG formvar or enamel wire through the center hole of the bead stack for the primary. The secondary is five passes of #26 wire.

The "reflection" mode transformers (**Figure 3B**) use four-turn windings on both primary and secondary. The secondary winding is a loosely twisted pair making two (or in some cases three) passes through the core. The start of one wire is connected to the finish of the remaining wire, and this point provides a connection (center tap) between the second and third turns of the secondary.

When constructed in the manner described, 450-ohm transformers offer exceptional performance. Measurements of properly terminated transformers indicate an SWR error of less than 1.15: 1 from 1.5 to 10 MHz, with less than 1 dB loss. Reversible Beverage antennas require near perfect voltage balance in the reflection transformer secondary windings. These transformers have almost no measurable secondary winding voltage imbalance.

The slight impedance transformation error is primarily due to losses in the transformer's core. This inherent defect shows up as a slight amount of internal resistive loading of the transformer windings. In a practical antenna system, transformer SWR errors produce negligible performance changes from the lower AM broadcast band to perhaps 15 MHz, while offering acceptable performance mid-VLF to 30 MHz. A 500-foot long,160-meter Beverage system can be used for receiving, with some sacrifice in directivity on lower frequencies, from a few hundred kilohertz to 30 MHz.

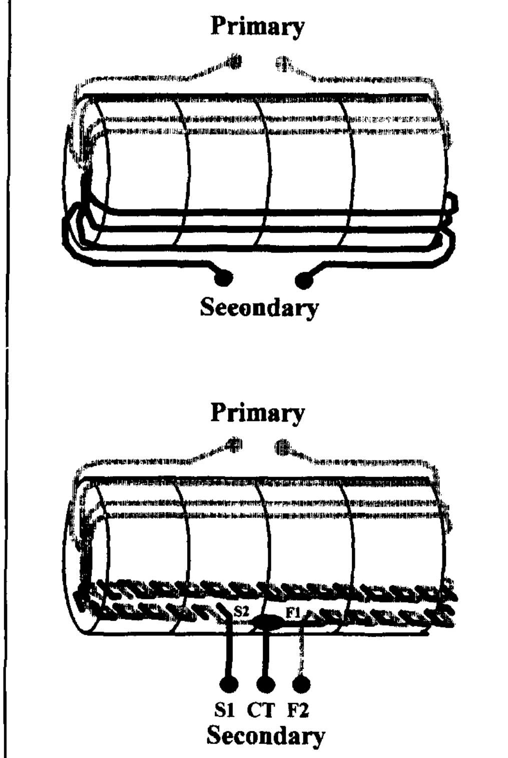

Figure 3. (A) The 75- to 450-ohm matching transformer (T1). (B) "Reflection" mode transformers use four-turn windings on both primary and secondary.

It's unnecessary, and actually undesirable, to isolate the primary and secondary windings with a Faraday shield. Stray capacitance must be minimized to ensure maximum performance and bandwidth of this system. Unnecessary conductors and dielectrics placed between the primary and secondary windings only serve to increase undesired winding-to-ground capacitances. Measurements have confirmed a Faraday shield actually increases common mode coupling and voltage imbalance.

Switching and control

In the design example shown in **Figure 2A**, a 12-volt sealed DIP relay is used to switch the termination resistor and feedpoint connections. The relay is activated by imposing a DC control voltage on the feedline. Capacitors isolate the line from the transformers, allowing measurement of system resistance from the feedpoint or far end of the antenna. A simple DC

ohmmeter is used to verify the integrity of the antenna conductors.

Operation

The system can be adjusted with a SWR analyzer, receiver, or small transmitter and clamp-on RF milliammeter. The termination resistors are adjusted for minimum SWR change with frequency, maximum null depth, or smoothly tapering common mode current in the antenna.

If the clamp-on meter is used, remember that ground and radiation losses will result in almost 50 percent reduction of current over a distance of 1 wavelength. The goal is to eliminate peaks and dips caused by standing waves. Don't attempt to achieve uniform current at both ends of the antenna.

A properly installed and adjusted system will provide a typical front-to-back ratio of 15 to 25 dB over a frequency range of 3:1 or greater. I've used 450-ohm systems at locations ranging from wet, black, sandy loam with high conductivity to poor, rocky clay. In all cases, these systems offer exceptional performance. ■

TECHNICAL CONVERSATIONS

(from page 4)

A late catch

Dear Editor:

One of my corrections did not make it [into my article "Modern Receiver Design," *Communications Quarterly*, Winter 1997, page 22]. The capacitor C28 of 1 nF should go to ground, but should not be in series with the 221-k resistor.

Also, on page 28, there was an error on one of the equations. The equation:

F = 10 (970) 30 dB

should read as follows:

F = 10 log (970) 30 dB

Sorry for the late catch.

Ulrich L. Rohde, DJ2LR, KA2WEU, HB9AWE
Upper Saddle River, New Jersey

Notes from WØIYH

Dear Editor:

Here is an item that I put on the Internet and that you might consider for the next *Communications Quarterly* issue. This note refers to my article on a logarithmic speech processor in the Winter 1997 issue of the *Communications Quarterly*.

In order to use a low-impedance dynamic microphone, with a lower level of audio volts than a high-impedance mic, two circuit changes can be made. I tested these changes using a RadioShack 33-3005 mic and also a 33-984 (dual hi/low Z) mic.

In the microphone amplifier circuit, solder a 1-k resistor across R4. This increases the gain of the mic amplifier. Solder a 0.1-µF capacitor across C3. This preserves the frequency response of this stage. Connect a 680-ohm resistor at the mic jack, across the audio line. This terminates the microphone properly.

It is also possible to install a low-Z to hi-Z miniature transformer instead of the R and C changes. The hi-Z side should have a 12-k resistor load. I tested this also. These transformers are available from various sources. RadioShack has the 274-016 adapter that does the same job.

Other microphones may require somewhat different values of R4 for the correct amount of speech processing.

There is a problem concerning the parts list for my speech processor. The AD633JN is the preferred, as shown in the schematic of **Figure 10**. The correction that I made to the parts list was not incorporated in the article. One reader had a question about that.

The AD633JN multiplier chip is available from Newark Electronics. They have a large quantity in stock for $6.50 in small quantities. They also have a $25 minimum order requirement that can be filled with other parts for the processor. Call 800-463-9275 for the phone number of the nearest Newark office.

Bill Sabin, WØIYH
Cedar Rapids, Iowa

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