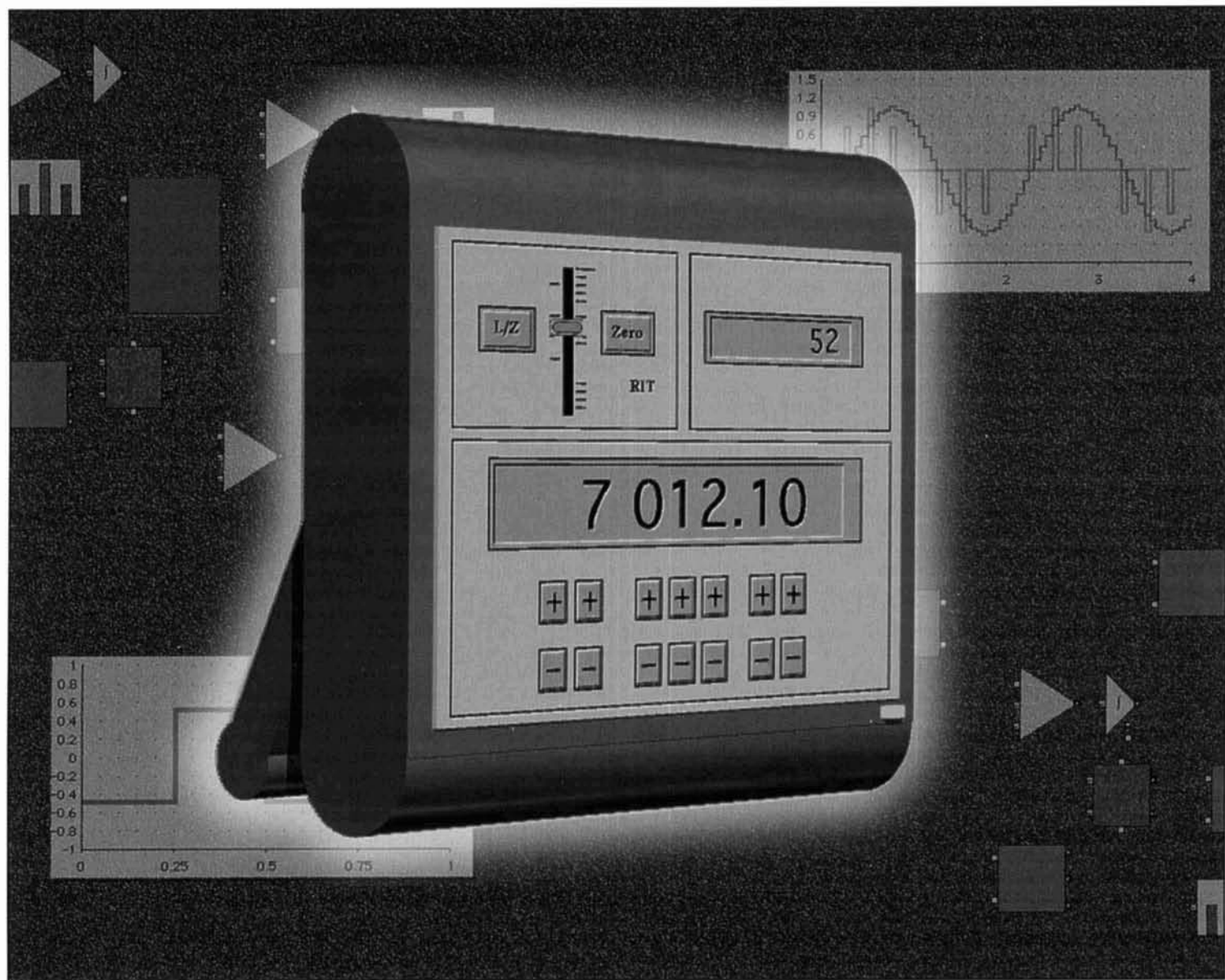


COMMUNICATIONS QUARTERLY

THE JOURNAL OF
COMMUNICATIONS
TECHNOLOGY

Winter 1995

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- Technical Conversations
- A Low-Power 20 Meter Transceiver
- Modeling and Understanding Small Beams
- Beyond the Z-Match
- Power on a Budget

- Quarterly Devices
- Connecting Computers to Radios
- Quarterly Computing
- Aerodynamic Balancing: Part 2
- The Solar Spectrum

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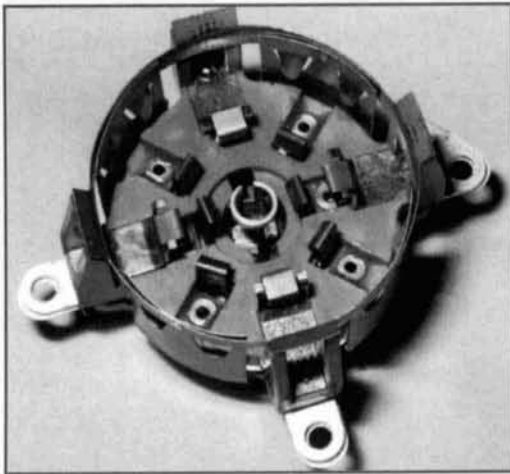
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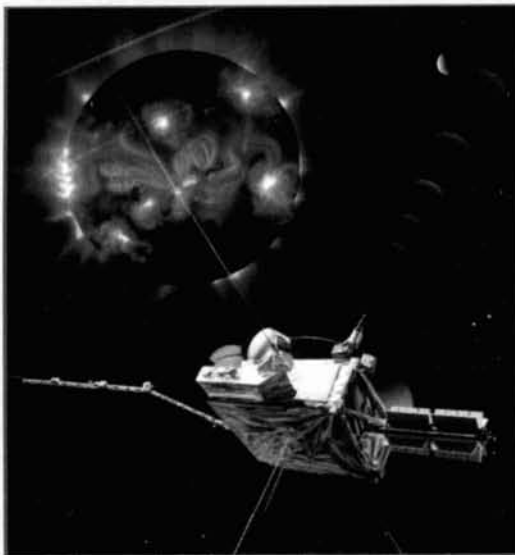
CONTENTS

Volume 5, Number 1

Winter 1995



Gonsior, page 55



Taylor, page 102

- 9 Connecting Computers to Radios**
Howie Cahn, WB2CPU
- 27 Beyond the Z-Match**
Charles A. Lofgren, W6JJZ
- 33 Modeling and Understanding Small Beams**
L.B. Cebik, W4RNL
- 51 Technical Conversations**
*David Barton, AF6S; F.P. Hughes, VE3DBQ;
L.B. Cebik, W4RNL; and Michael Hatzakis, Jr., WB2TBQ*
- 55 Power on a Budget**
Marv Gonsior, W6FR
- 65 Quarterly Devices**
Rick Littlefield, K1BQT
- 69 A Low-Power 20-Meter Transceiver**
Clint Bowman, W9GLW
- 83 Tech Notes**
*Peter Bertini, K1ZJH; Gil Sones, VK3AUI;
Paul Duff, VK2GUT; Ian Keenan, VK3AYK; and
D.R.W. Hutchinson, M(SA)IERE*
- 89 Aerodynamic Balancing: Part 2**
Dick Weber, PE, K5IU
- 99 Quarterly Computing**
Brad Thompson, AA1IP
- 102 The Solar Spectrum**
Peter O. Taylor

Cover Photo: Amateur radio marches on. This month's cover by Bryan Bergeron, NU1N, depicts a laptop computer with touch screen interface displaying the virtual tuner frequency control panel à la Howie Cahn, WB2CPU. For more on "Connecting Computers to Radios," see Howie's article beginning on page 9.



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Connect with us

The Armchair Homebrewer

Forty years ago, the homebrewer's art involved a blend of skills and tools: chassis and metal working, some machinist work, a dose of electronics theory, and the latest *QST* or *ARRL Radio Amateurs' Handbook* to guide one through the latest innovations in the amateur arts. Borrow a circuit here, add one from there, and then cut and tweak away until it worked. And, don't forget a strong back; you needed one to lift those bulky rack-mounted monsters into place!

Add a pinch of black magic. Radio was still an "art." Even the biggest manufacturers had scores of technicians debugging circuits and fine-tuning component values to reach the engineered specifications. When electrical engineers confronted their mechanical counterparts, panel ergonomics usually conflicted with the practical layout. This often resulted in complicated (and costly!) arrays of gears, switches, and levels behind the front panel to mechanically interface the human element with the electronics.

Ah, nostalgia! Those of us who are old enough, fondly remember our first coveted Greenlee chassis punch. Now, we could knock out those octal tube socket mounting holes with ease. No more tedious drilling and fine filing to make things fit. The ultimate goal was to have a complete set of Greenlees, which would let us mount the smallest 7-pin socket or the largest panel meter. Only those with the deepest pockets could afford the costly square punches—incidentally handy tools that would let you knock out clean rectangular holes for IF transformers and cable sockets with ease!

I am a product of the "baby-boom" generation. I'm old enough to remember with fondness the good old days of homebrewing (how time clouds the mind!) and young enough to appreciate the current level of technology. Given a choice—would you prefer the finest Collins receiver made in 1955, or would you opt for the new Watkins Johnson HF-1000? Honestly?*

What I'm leading up to—and the message I'm trying to convey is simple—as radio amateurs, we have a choice. Some of us will bury our heads in the sand and live for the past. Some will hold on to the old methods of building and homebrewing, without giving a thought

to the technological advances waiting to be discovered using today's newest homebrew toys. But, open-minded amateurs—those who read this magazine and are comfortable in the world of computers—will probably be the standard-bearers promoting the techniques of current and future RF technology.

We now have access to powerful CAD programs that help us do everything from designing panel layouts, to drawing schematics, to creating pc board layouts. RF design programs have taken the pain out of circuit and filter design. With a few simple keystrokes, we can alter what we see on the screen, and view how the actual circuit will perform. If we are happy with the results, all we need to do is enter a few more keystrokes and our bubble jet printer will produce the schematic and pc board transparency. Drafting tape and donuts will become a thing of the past. (As a matter of fact, such things are becoming harder and harder to find as electronics designers move to CAD.) Other software lets you own, on paper anyway, the finest antenna farm conceived by man. No antenna range or tedious trimming needed, these antennas do work as designed!

Have I piqued your curiosity? Well, then, tune in as Brad Thompson, AA1IP, author of our new "Quarterly Computing" column takes you on a tour of the American Radio Relay League's new *ARRL Radio Designer* program. This reasonably priced software will provide those amateurs interested in learning computer-aided design a stepping-off place. But I'll let Brad give you the details.

So, stoke the fireplace, settle down in your favorite armchair, and cozy up to your computer with a cup of hot cocoa and *Communications Quarterly* in hand—and enter the world of armchair homebrewing!

Peter Bertini, K1ZJH
Senior Technical Editor

*If the truth be known, our Senior Technical Editor not only lusts after the newest, most technologically up-to-date equipment, he's also an avid collector of antique radios. I think he just might be faced with a bit of a dilemma here!

KA1STC

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RM-35A	25	35	5 1/4 x 19 x 12 1/2	38
RM-50A	37	50	5 1/4 x 19 x 12 1/2	50
RM-60A	50	55	7 x 19 x 12 1/2	60
• Separate Volt and Amp Meters				
RM-12M	9	12	5 1/4 x 19 x 8 1/4	16
RM-35M	25	35	5 1/4 x 19 x 12 1/2	38
RM-50M	37	50	5 1/4 x 19 x 12 1/2	50
RM-60M	50	55	7 x 19 x 12 1/2	60

RS-A SERIES



MODEL RS-7A

MODEL	Colors		Continuous Duty (Amps)	ICS* (Amps)	Size (IN) H x W x D	Shipping Wt. (lbs.)
	Gray	Black				
RS-3A	•	•	2.5	3	3 x 4 1/4 x 5 1/4	4
RS-4A	•	•	3	4	3 1/4 x 6 1/2 x 9	5
RS-5A	•	•	4	5	3 1/2 x 6 1/8 x 7 1/4	7
RS-7A	•	•	5	7	3 3/4 x 6 1/2 x 9	9
RS-7B	•	•	5	7	4 x 7 1/2 x 10 3/4	10
RS-10A	•	•	7.5	10	4 x 7 1/2 x 10 3/4	11
RS-12A	•	•	9	12	4 1/2 x 8 x 9	13
RS-12B	•	•	9	12	4 x 7 1/2 x 10 3/4	13
RS-20A	•	•	16	20	5 x 9 x 10 1/2	18
RS-35A	•	•	25	35	5 x 11 x 11	27
RS-50A	•	•	37	50	6 x 13 3/4 x 11	46
RS-70A	•	•	57	70	6 x 13 3/4 x 12 1/4	48

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MODEL RS-35M

MODEL	Continuous Duty (Amps)	ICS* (Amps)	Size (IN) H x W x D	Shipping Wt. (lbs.)
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RS-12M	9	12	4 1/2 x 8 x 9	13
• Separate volt and Amp meters				
RS-20M	16	20	5 x 9 x 10 1/2	18
RS-35M	25	35	5 x 11 x 11	27
RS-50M	37	50	6 x 13 3/4 x 11	46
RS-70M	57	70	6 x 13 3/4 x 12 1/4	48

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	@13.8VDC	@10VDC	@5VDC			
VS-12M	9	5	2	12	4 1/2 x 8 x 9	13
VS-20M	16	9	4	20	5 x 9 x 10 1/2	20
VS-35M	25	15	7	35	5 x 11 x 11	29
VS-50M	37	22	10	50	6 x 13 3/4 x 11	46
• Variable rack mount power supplies						
VRM-35M	25	15	7	35	5 1/4 x 19 x 12 1/2	38
VRM-50M	37	22	10	50	5 1/4 x 19 x 12 1/2	50

RS-S SERIES



MODEL RS-12S

• Built in speaker

MODEL	Colors		Continuous Duty (Amps)	ICS* Amps	Size (IN) H x W x D	Shipping Wt. (lbs.)
	Gray	Black				
RS-7S	•	•	5	7	4 x 7 1/2 x 10 3/4	10
RS-10S	•	•	7.5	10	4 x 7 1/2 x 10 3/4	12
RS-12S	•	•	9	12	4 1/2 x 8 x 9	13
RS-20S	•	•	16	20	5 x 9 x 10 1/2	18
SL-11S	•	•	7	11	2 1/4 x 7 x 9 3/4	12

Automatic packet revisited

As I was reading your editorial in the Summer 1994 issue of *Communications Quarterly* I kept waiting for you to mention the alternative to having the first-forwarding station check the contents of each message. It never came. Just moaning about how hard it is and how the rules should be changed. The alternative, of course, is "verifying the originator's identity—authentication. This is a technical magazine, so let's get technical.

You did get warm when you mentioned keeping lists of callsigns to delay. The problem with having a list of callsigns that you don't automatically forward from (or even a list of people that you know well enough to always forward automatically) is that someone can always sign on using somebody else's callsign and do whatever they please.

An authentication mechanism could ensure that messages actually originated from a particular station, rather than from an imposter. This would relieve the first-forwarding station from all responsibility for the content of messages sent through his or her BBS. It would also ensure that mail messages waiting to be read could not be read and/or deleted by a third party.

I won't go into all the details of how an authentication mechanism could be put in place, but there are already some in commercial use to protect dial-in modem lines for unauthorized use. Two schemes immediately come to mind; I'm sure there are others.

1. Challenge-Response—When a packet user connects to a BBS the BBS offers a challenge, usually some rather large random number. The user takes the challenge and computes a response using a unique cryptographic algorithm and sends it back to the BBS. If the BBS verifies that it is the correct response for that particular challenge, the user's identity is authenticated and he or she is logged into the BBS. The algorithm (or more precisely the unique parameters of the algorithm) are decided upon between the SYSOP and the user when a new user joins the BBS. This could be done in person to verify the person's identity (eg. show his or her license).

2. One-Time Pad—This scheme is similar to the challenge-response method, except that when a new packet user joins a BBS he or she is given a printout of code words. These would generally be large randomly generated numbers. Whenever the user logs onto the BBS he or she would enter the first unused code word from the sheet and then scratch it off the list. The BBS

would ensure that the code words for each user were used in sequence. If an invalid code word were given the user would not be authenticated and would not be allowed to log in.

Either of these two schemes would be relatively easy to implement in BBS software. It could even be totally automated by using special software on the packet user's side. It would require some up-front time by the SYSOP to set up a new user of a BBS, but this is much easier than filtering every message by hand. And, it doesn't require a judgement call by the SYSOP as to whether or not a message complies with the rules!

These schemes are not 100 percent perfect, however. Someone could break into a ham's home and run his or her software, although simple password protection on the computer itself could take care of that.

It is possible that a user could log into a BBS and then someone with a higher-powered transmitter could take over the link and forge a message while the user is still logged in. To counter this attack, protocols could be developed that would authenticate every packet—but this would require that packet radio change significantly.

I believe that implementing an authentication scheme when establishing a connection to a BBS would virtually eliminate the problems you were concerned with in your editorial. SYSOPs would be off the hook for the content of messages and would not need to review every single one. And, packet users would sleep better knowing that someone wasn't reading and sending messages while logged in using their callsign.

You mentioned that BBS operators' interests are primarily technical. Well, let's solve this problem with a technical solution, rather than a political one. It can be done, and it's not that hard to do.

**Ross Wille, N6SJD
Orem, Utah**

"Online" applause

I picked up *Communications Quarterly* at Barnes and Noble a few weeks ago. Now that local bookstores are carrying a big variety of magazines, I get a chance to see what some of them are like. (I prefer the non-chains. It's too bad the local Swartz Bookstore doesn't carry it.)

All the articles were good. I even ordered two back issues that an article had referenced. I think that I can use some material for a university course that I am teaching.

The library at my full-time work could not

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find any library nearby that subscribed. Too bad. The closest one was in New Jersey. I bet, though, that some technical college departments subscribe, and they aren't listed in the online catalogs.

I will definitely buy the Quarterly again. Based on seeing just one issue, I like the level of detail, and the tutorial character of the articles. There seems to be some depth in basic theory so that I can really understand the physical principles. (That's the reason some of the articles may help my students.)

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Communications Quarterly Goes "Online"

Communications Quarterly readers now have a direct electronic link to our magazine's authors and editors. A new "reader feedback" area has been set up on "America Online," (AOL) in which readers may ask questions and send letters to the editor. A *Communications Quarterly* representative will regularly read the mail, respond promptly to questions, and refer comments, suggestions, etc., to the appropriate editors, columnists, or authors. Reports of subscription problems will be forwarded to our Circulation Department.

AOL members may find the *Communications*

Quarterly feedback area in the Ham Radio Club (keyword: Ham Radio) message center, under the heading "Manufacturers, Dealers, and Vendors." Look for the *Communications Quarterly* folder (along with those for our sister publications, *CQ* and *Popular Communications*, and for *CQ* Books and Videos). Readers who are not AOL members may send e-mail to us via the Internet to N2WL@aol.com.

This reader feedback area is the first step in ongoing efforts by *CQ* Communications, Inc., to better serve its customers through an easily accessible online presence for *Communications Quarterly* and all of its other publications and products. We will keep you informed as we develop more online services for you, either on AOL, on other commercial services, or directly on the Internet.

(A free trial membership in America Online, including software and 10 free hours, is available by calling 1-800-827-6364.)

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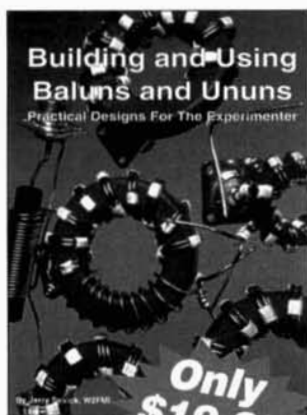
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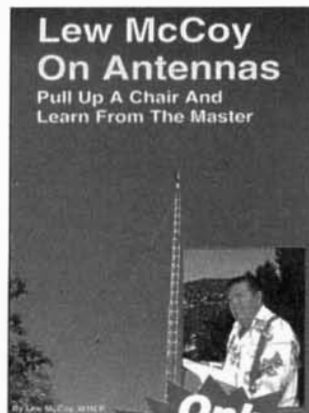
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Lew McCoy on Antennas

by Lew McCoy, W1ICP



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CONNECTING COMPUTERS TO RADIOS

Adding DDS frequency control

New components are available that let you build a frequency generator based on direct digital synthesis (DDS) techniques with ease. The DDS circuit that I'll describe here, together with a personal computer to control it, can add features such as high resolution digital displays, precise tuning, frequency storage memories, incremental offset tuning, and split frequency operation, to many radios that lack them.

Direct digital synthesis was explained in an excellent article by Bryan Bergeron in *Communications Quarterly*.¹ As Bergeron stated, the advantages of DDS frequency generation include: high stability, low phase noise, fast frequency shifting, and a low component count implementation. Because the frequency and other properties of the output of a DDS circuit are set using control words in digital form, it can be readily interfaced to a computer. The computer can be equipped with software that provides a "virtual radio" user interface that replicates the familiar appearance of a radio on the computer screen, but has additional functionality obtainable only with computer control.

This article will provide a brief summary of how DDS works, present a simple DDS circuit, and discuss a Microsoft Windows™ graphical user interface (GUI) for the circuit. A previous article² described the hardware and software for a computer interface to a PLL-based VHF transceiver, the Ramsey FX-146. Here, I'll build upon the techniques described in my earlier article, and in the Bergeron piece, to cover these topics.

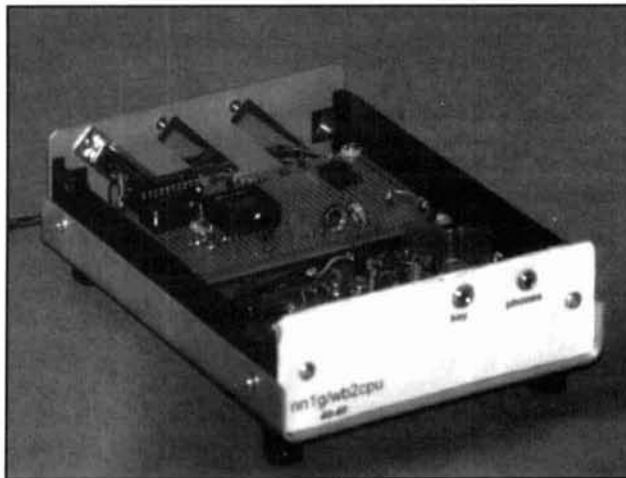


Photo A. 40-40 transceiver board and DDS board in enclosure.

The circuit is a sine-wave frequency generator that uses the Analog Devices AD7008 DDS modulator chip. It can produce any frequency up to approximately 15 MHz and is capable of sourcing several milliwatts into a 50-ohm load. Frequency resolution is about 0.01 Hz although, in practice, the output signal's accuracy, as well as its stability and phase noise, are limited by the characteristics of an on-board, crystal-based oscillator. You can use the DDS board as a general-purpose signal source, or add it to a new or existing receiver and/or transmitter. Frequency control of a target radio is achieved by driving its transmit and receive mixers with the DDS output instead of the

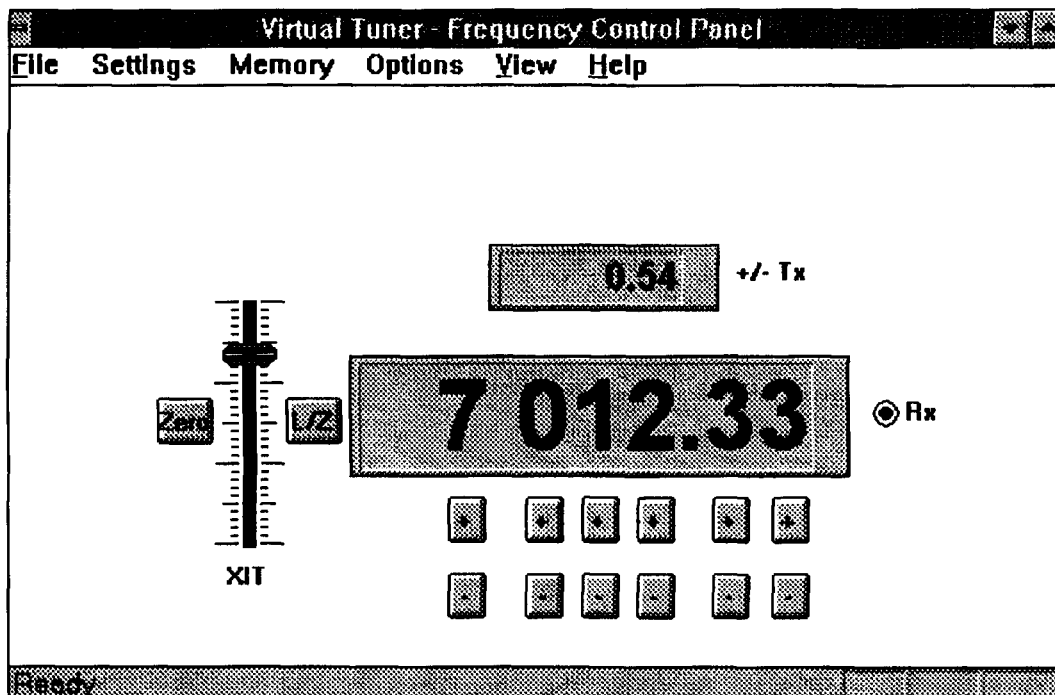


Figure 1. Sample DDS control panel.

radio's existing local oscillator. This will be illustrated on a simple single-conversion, low-power HF CW transceiver, the "40-40," ably designed by Dave Benson, NN1G.³ The 40-40 is representative of similar radios currently popular in the ham QRP community with which the DDS circuit can also be used.* The 40-40 transceiver board and the DDS board fit into a 5 by 7 by 2-inch enclosure with room to spare, as seen in **Photo A**.

The DDS board will generate some spurious signals. The strongest will be about 50 to 60 dB below the desired output signal. This meets spectral purity requirements for amateur radio transmissions. When used with a receiver, a few undesired signals—"birdies"—may be heard. While this may not be acceptable in high-performance receivers, the problem shouldn't be objectionable when used with the kind of equipment mentioned above. "Spurs" are largely a function of the resolution of the digital-to-analog converter (DAC) that constitutes the output stage of a DDS system. As sufficiently fast DACs with increased bit widths become available, this limitation should largely disappear for HF applications.

The computer link is made through the machine's parallel port. Using a combination of hardware and software methods, the features previously mentioned: receive incremental tuning (RIT), transmit incremental tuning (XIT), split transmit/receive frequency operation, and labeled memories, are implemented in a way I

*For example, the technique should also be applicable to the radio described in "A Deluxe QRP Transceiver," *Communications Quarterly*, Winter 1994, page 25.

believe is more powerful and useful than what is provided on commercial radio transceivers.

The graphical user interface combines the look of a standard Windows application with a representation of an actual radio front panel. I tried to make most control functions operable both with a pointing device, such as a mouse, to make it easy to learn, and from the keyboard, for speed. Physical radio panels often appear cluttered and confusing because of the need to cram so many functions onto them. In contrast, a virtual panel is dynamically configurable; that is, only the controls you currently need are displayed. **Figure 1** shows one of the many ways the DDS GUI can be made to appear.

Direct digital synthesis—an intuitive introduction

In a recent *QST* article,⁴ I tried to illustrate how DDS works using a non-theoretical, intuitive approach. Enough people have told me they found it useful that I will summarize the explanation here.

Unlike most other frequency generation methods, DDS is a digital process. The desired waveform is computed on-the-fly by digital hardware instead of being the result of the physical resonance of components like quartz crystals, inductors and capacitors, or mechanical resonators. The mechanism used by a DDS circuit to compute the output waveform is analogous to the way the human mind would perform the same task. In fact, it corresponds to

something you've probably already experienced. In your high school trigonometry class you may have been given an assignment like, "Draw one cycle of a sine wave." Let's review how it's done. First, you must decide how many points to plot to establish the curve. Let's plot, say, 20 equally separated points, corresponding to one every $360^\circ/20 = 18^\circ$. You look up and plot the sine of the angle of the first point; then you add 18° and lookup and plot the next one until the cycle is completed; i.e., $\sin 0^\circ (= 0.00)$, $\sin 18^\circ (=0.309)$, $\sin 36^\circ (=0.588)$, etc. The resulting familiar sinusoid is shown in **Figure 2**.

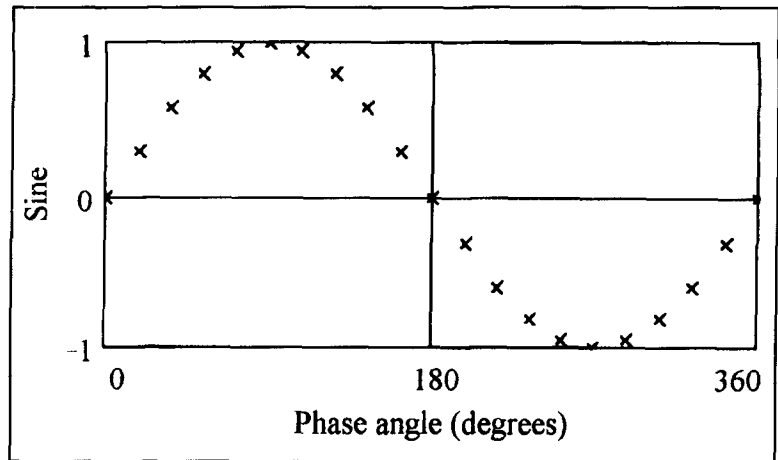


Figure 2. Sine wave from 20 plotted points.

In a real system, each cycle takes a specific time: the *cycle period*. Thus, a point on the x-axis represents both the phase angle and the time. The waveform's frequency is determined by the time interval between the plotted points. Continuing the example, if you could plot the points at a rate of one every 10 seconds, the cycle period would be $20 \times 10 \text{ sec} = 200 \text{ sec}$, corresponding to a frequency of 0.005 Hz. The interval between two adjacent points can be considered as either an absolute time value, or as the specific portion of a full cycle. Whether it is expressed as a fraction of a cycle ($1/20$), or as an angle (18°), this value is the *phase difference* or *phase increment* between the two points. The rate at which we were able to plot the points can be considered the input frequency to our system. It should be clear that the ratio of the input frequency to the output frequency is equal to the number of points plotted per cycle ($0.1/0.005 = 20$), and, the number of points per cycle is the reciprocal of the phase difference, $\Delta\Phi$. Therefore, for a desired output frequency, f_{out} , and a fixed clock rate, f_{in} , the required phase difference between points is:

$$\Delta\Phi = (f_{\text{out}} / f_{\text{in}}) \quad (1)$$

A DDS system performs a similar process using digital logic. **Figure 3** is a block diagram of a DDS circuit. The time between points is determined by an oscillator used as a system clock. At each clock tick, the phase difference, provided on the phase difference input lines, is added to the current phase angle and the sine of this new angle is computed. The circuitry that does this is:

- A **digital adder**—digital logic that adds two values: in this case the current phase angle in the phase accumulator plus the phase difference. The adder's output is fed into the phase accumulator.
- The **phase accumulator**: an n -bit binary register, that stores the current value of the phase angle. The value in the phase accumulator after each clock tick represents the phase angle of the point in the wave being synthesized.

- A **read-only memory (ROM)** that contains a lookup table of sine values. The output of the ROM is a series of values corresponding to the sine of the phase angles at each of the points.
- A **digital-to-analog converter (DAC)** to convert the binary representation of the waveform into a current or voltage.
- A **low-pass filter** to remove high frequency components from the output.

Traditionally, angles are specified in degrees, gradients, or radians. For example, one quarter of a cycle is 90° or 100 grads or $\pi/2$ radians. In DDS or other digital systems there is another way to describe angles such as the phase or phase difference. An n -bit phase accumulator can store one of 2^n possible values. If we represent a full cycle as 2^n , then a fractional part of a cycle can be written as a fraction with 2^n as the denominator. For instance, with an 8-bit phase accumulator, 2^n equals 256; therefore a 90° angle, $1/4$ of the cycle, is expressed as $64/256$. If we multiply both sides of **Equation 1** by 2^n we get: $\Delta\Phi_n = (f_{\text{out}} / f_{\text{in}}) * 2^n$, where $\Delta\Phi_n$ is the numerator of the phase difference fraction. To program the DDS to generate an output one quarter of the input frequency with an 8-bit phase accumulator, you use 64_{10} as the phase increment value. Our example was limited to plotting one cycle of a wave, 360° . Because the waveform is periodic, if we had continued further, the only information we would need to calculate subsequent points is the phase angle relative to the start of the current cycle. We can subtract out any multiple of 360° from the phase angle before computing or looking up the sine. Using a register whose maximum value represents a full cycle does the same thing for us automatically; upon overflow, it just starts again at 0, like a clock does. Also, don't worry that it may not be possible to specify the desired $f_{\text{out}}/f_{\text{in}}$ ratio exactly with a fraction whose denominator is 2^n . In available DDS

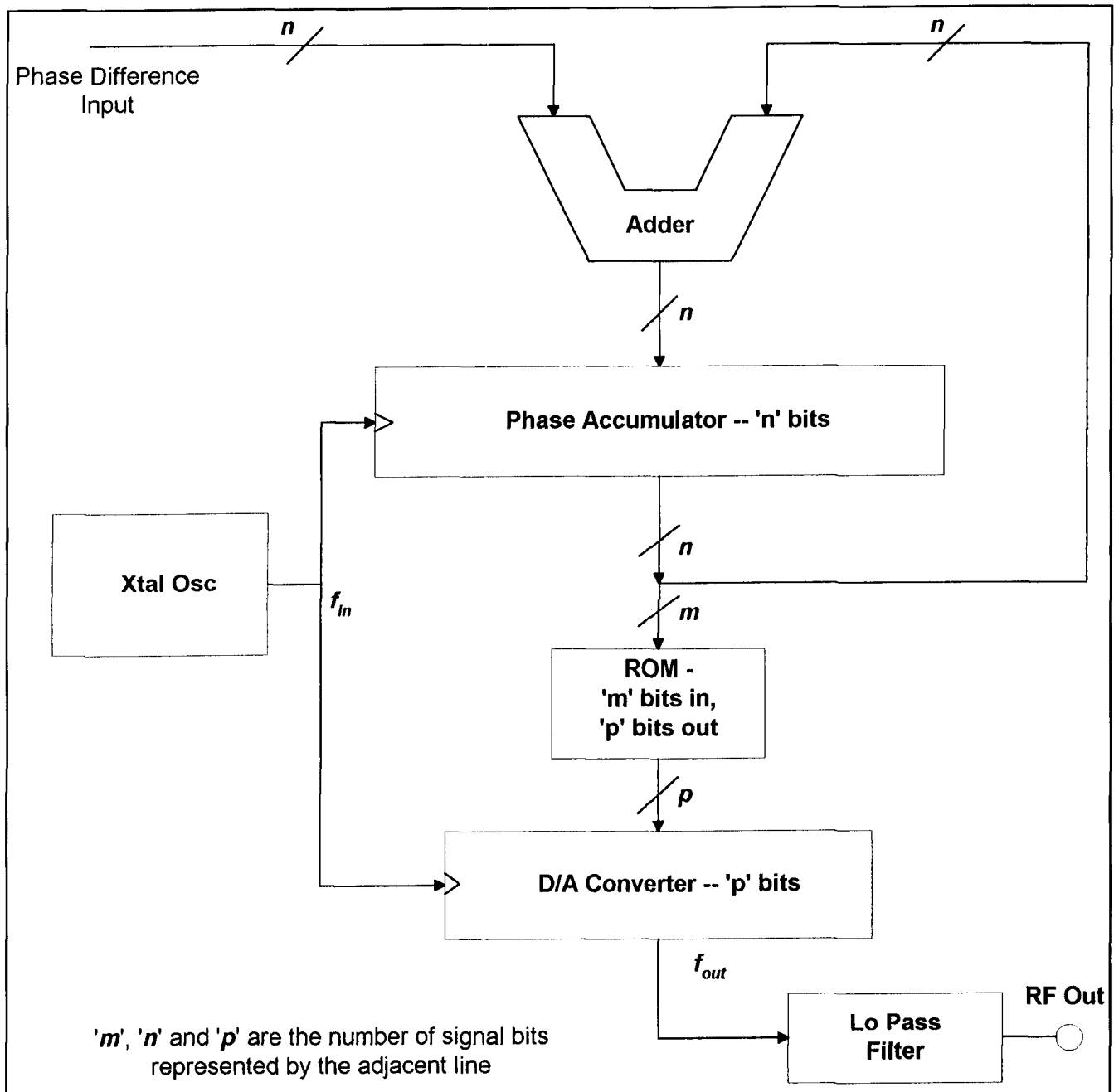


Figure 3. DDS system block diagram.

devices the number of bits in the phase accumulator register is large enough that the resulting error is generally ignorably small.

How small? For any desired frequency within the range of a particular DDS system there are two consecutive values for the phase difference, n and $n+1$, such that the frequency resulting from a phase difference of n is too low and the frequency resulting from $n+1$ is too high (unless the correct frequency is generated exactly by n or $n+1$.) Therefore, the phase difference value can always be set to within one-half bit of the frequency. In other words, the maximum frequency error is less than one half the effect of a

one-bit phase difference change. The frequency change that results if the phase difference is changed by one bit, the resolution of the system, can be derived from **Equation 1**. Since the smallest possible change is $1/2^n$, the value of the least significant bit in the register is:

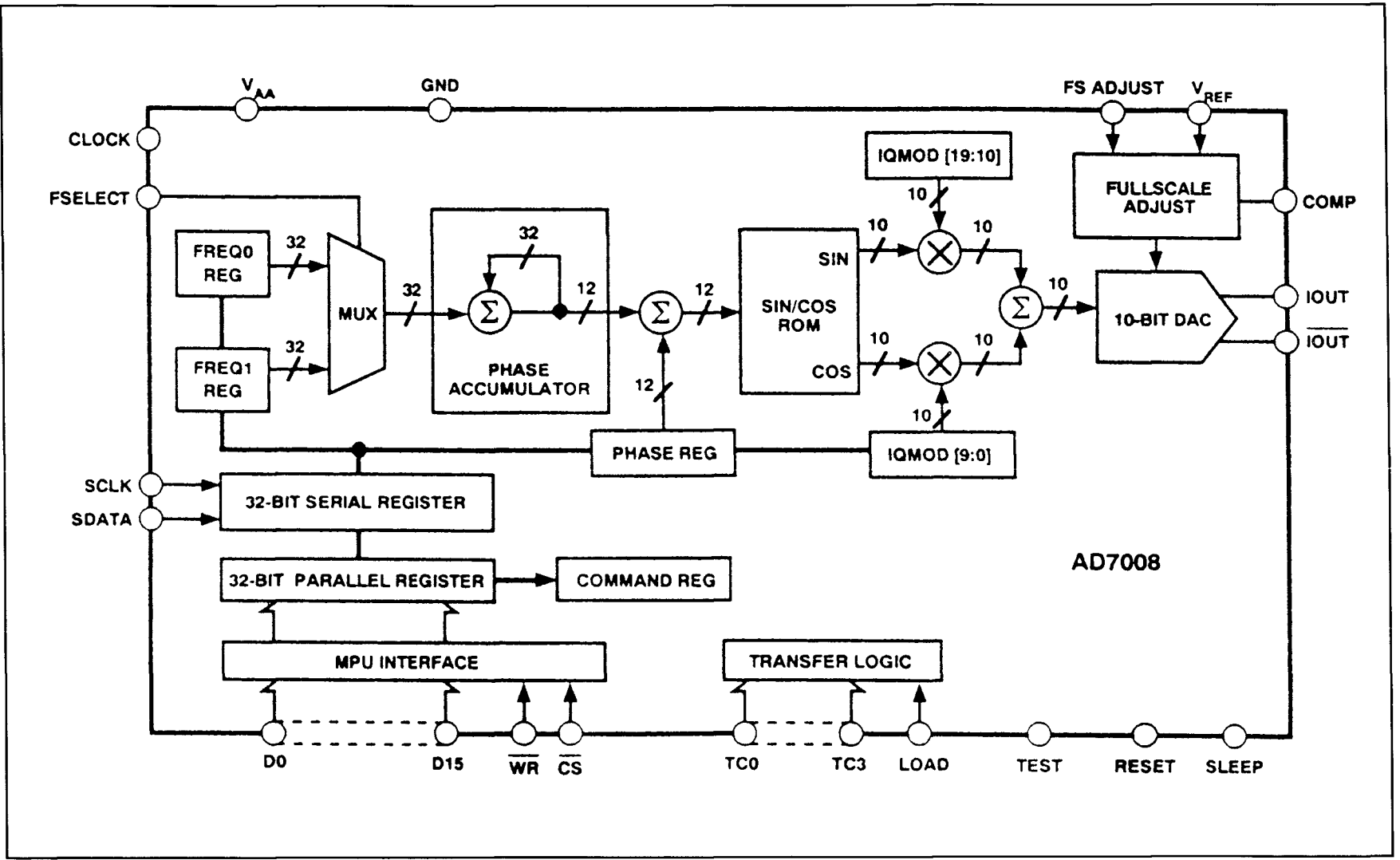
$$\Delta\Phi_{\min} = 1/2^n = \Delta f_{\min} / f_{in}$$

or,

$$\Delta f_{\min} = f_{in} / 2^n \tag{2}$$

Thus, the frequency resolution is dependent

Figure 4. AD7008 block diagram (courtesy of Analog Devices, Inc.).



on the size, n , of the phase accumulator register. Available DDS devices typically have phase accumulators between 24 and 32 bits wide. As an example, with a 30-MHz system clock used as the input frequency, this corresponds to output frequency resolutions of 1.8 Hz down to 0.007 Hz for these accumulator sizes. The maximum output frequency is limited (by the Sampling Theorem) to one half the system clock frequency, f_{in} . In practice, a value of approximately 0.35 to 0.4 f_{in} is obtainable.

Inside the AD7008*

What has made DDS systems practical in recent years has been the development of integrated circuits that incorporate much of the functionality represented in **Figure 3** in a single chip. For this project, I'm using the Analog Devices AD7008. The AD7008 contains a 32-bit phase accumulator, a 10-bit DAC, a flexible interface for programming, and several additional useful features. Some of these can be seen in **Figure 4**. In **Figure 3**, a basic DDS circuit, the phase difference was input into the system on n binary signal lines and added directly to the phase accumulator value. In the AD7008, these lines are routed to and stored in one of two buffer registers, **FREQ0 REG** or **FREQ1 REG**. The register value used for the current phase difference is determined by a multiplexer controlled by the device pin *FSELECT*. Thus, you can store two frequencies in the chip and can switch between them almost instantaneously, without doing any additional programming. This is how offset and split frequency tuning are done in the board and control program described here. You could also use this feature to perform frequency-shift keying (FSK).

The AD7008 makes it easy to add many forms of modulation to the output signal. Much of the circuitry to accomplish this is contained on the chip. To phase modulate, you load a stream of data representing a digitized form of the desired modulating signal into the register labeled **PHASE REG**. This data is summed with the output of the phase accumulator to form a phase modulated output signal. To perform simple amplitude modulation, you load half of the **IQMOD** register (bits 9:0) with data representing the modulating signal, and an internal multiplier generates AM. If you simultaneously feed the other half of the **IQMOD** register (bits 19:10) with a digitized form of the same audio modulating signal, phase shifted by 90°, you get SSB. This quadrature component can be generated by a DSP chip by using a

Hilbert transform. Frequency modulation (FM) is achieved by externally adding the modulating signal to the phase difference value. Thus, with the AD7008, a DSP chip, and only a few other ICs, it should be possible to construct an all-mode radio with a minimum number of components. It's not surprising that the AD7008 contains features that are useful for building amateur radio equipment. The co-designer of the chip, Dr. Neil Weste, is a ham, AA1KR. Neil took an early prototype of the part, added an op amp, an A/D converter and some standard CMOS logic and built a low-power AM transmitter.

Currently, the AD7008 is available in two versions; one supporting a 20-MHz clock frequency, the other 50 MHz. It comes in a 44-pin PLCC package.

Programming the AD7008

The AD7008 contains a group of registers that you program to set the frequency and modulation of the output, and control other functions on the chip. These registers are loaded through either a serial, or a parallel, interface mechanism. Using the serial interface involves clocking in data one bit at a time using the device pins *SDATA* and *SCLK*. The parallel interface transfers data in either 8 or 16 bit units using the chip's data bus and the Chip Select (*CS_*) and Write (*WR_*) pins. In either case data is stored temporarily, in a serial assembly register or a parallel assembly register, until the necessary number of bits to program a register have been received. The following are the programmable registers and the functions they control:

- **FREQ0 REG and FREQ1 REG (32 bits each)**. Allows the user to specify the phase difference values that determine two output frequencies. The maximum frequency is limited by the Sampling Theorem (see discussion).
- **PHASE REG (12 bits)**. Specifies an offset to the phase of the output waveform; used for phase modulation.
- **IQMOD (HIGH and LOW; 10 bits each)**. Specifies values used to amplitude modulate the output with the in-phase (I) and quadrature (Q) components, respectively, of a modulating signal.
- **COMMAND REG (4 bits)**. Used to set four internal functions. The four bits are:
 - CR0: 8-bit (0) or 16-bit (1) parallel data bus transfers
 - CR1: Normal operation (0) or Low power "Sleep" mode (1)
 - CR2: AM enable (ON = 1)—required to use the IQMOD register
 - CR3: Synchronizing Logic Enabled (ON = 1).

*For more complete functional and programming information about the AD7008 see the data sheet available from Analog Devices, Inc., Norwood, Massachusetts.

TC3	TC2	Action
0	0	Transfer Parallel Assembly Register to Command Register
1	0	Transfer Parallel Assembly Register to FREQ , PHASE , or IQMOD register as determined by pins TC1 and TC0 below
1	1	Transfer Serial Assembly Register to FREQ , PHASE , or IQMOD register as determined by pins IC1 and TC0 below
TC1	TC0	Destination register
0	0	FREQ0 REG
0	1	FREQ1 REG
1	0	PHASE REG
1	1	IQMOD (both halves)

Table 1. Register operations controlled by the transfer control (TC) pins.

These registers are loaded using one of the following:

- **SERIAL ASSEMBLY REG (32 bits).**
- **PARALLEL ASSEMBLY REG (32 bits).**

Programming the AD7008 is a two-step process. First, either the serial or parallel assembly register is loaded, and then the data is transferred from there into one of the control registers. As an example, let's program the frequency using the eight-bit parallel interface to load the register **FREQ0 REG**. This is a 32-bit register, so four bytes of data are required. The most significant byte of the Phase Difference data is presented at pins *D7* to *D0* of the data bus and loaded into the parallel assembly register by asserting pin *CS_* and then *WR_*. The data bus data is latched in on the rising edge of *WR_*. This process is repeated with the next three bytes of phase difference data. The second step is to transfer the contents of the assembly register into a control register—in this case **FREQ0 REG**. This destination register is selected by setting the Transfer Control pins, *TC3:TC0*, to 1000, and effecting the transfer with the *LOAD* pin. **Table 1** describes the necessary pin states to accomplish all the possible register transfers.

Computing the phase increment value

As discussed, the value used as the phase difference is:

$$\Delta\Phi_n = (f_{out} / f_{in}) * 2^n$$

For this circuit *n* is 32, the size of the AD7008 phase accumulator, and the system clock is 50.0 MHz. To program the chip to generate, say, 11 MHz:

$$\begin{aligned} \Delta\Phi_n &= (11.0/50.0) * 4\,294\,967\,296 \\ &= 944\,892\,805.1_{10} \\ &= 3851EB85\text{ H} \end{aligned}$$

Therefore, the bytes "38" "51" "EB" "85", in that order, are loaded into the parallel assembly register using the lower eight bits of the data bus. Alternatively, the bit sequence "0011 1000 0101 0001 1110 1011 1000 0101" can be loaded into the serial assembly register.

Connecting to the chip

We've discussed the information needed to configure the AD7008. However, we still have to get this data from a computer to the chip. In the Ramsey article, I described three possible mechanisms to do this: though a computer's serial port, a parallel port, or by taking the data directly from the computer bus. For this project the computer's *parallel* port is used, but data is loaded with the chip's *serial* interface. Here's why. Ideally, we'd like to access the AD7008 data bus plus another nine device pins: *TC3*, *TC2*, *TC1*, *TC0*, *LOAD*, *CS_*, *WR_*, *SCLK*, and *SDATA*. This would give us full flexibility to choose either of the interface formats and allow us to use all the chip's functions.

Unfortunately, the parallel port only provides four control output pins. This forces us to compromise and give up some of the functionality of the AD7008 because there aren't enough port pins to control all the device pins needed for all of the functions. Using the chip's serial interface requires fewer control lines than the parallel interface. I use the chip with pins *TC3*, *TC2*, *WR_*, and *CS_* hard-wired to +5 volts. Connected this way, transfers can only be made from the Serial Assembly Register to either **FREQ0 REG**, **FREQ1 REG**, or **PHASE REG**. The ability to write to the Command Register or to use the parallel interface is lost. The Command Register is initialized to all zeros on power-on reset. The inability to change these bit settings prevents us from using the chip's "sleep" mode, or the **IQMOD** register. If you wanted to use these features, your best option would probably be to redesign the interface to connect to the computer through a bus interface card instead of the PC parallel port.

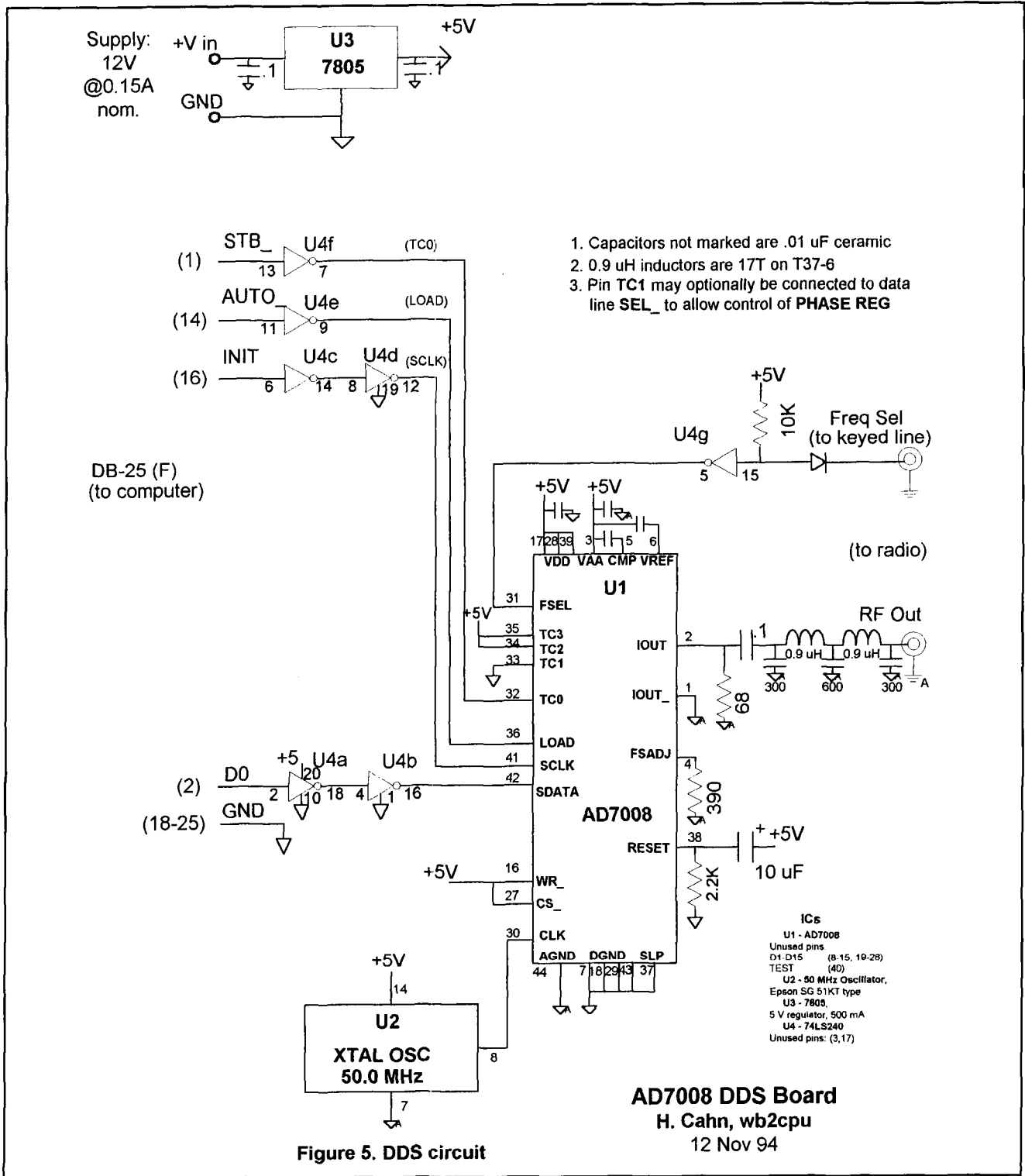


Figure 5. DDS circuit

Figure 5. DDS circuit.

The DDS board—design notes, construction and testing

Figure 5 is the schematic of the DDS board, along with a parts list. Most of the circuit has already been explained. The voltage developed by the RC network connected to U1 pin 38

asserts the *RESET* pin for several milliseconds when the board is powered. The reset function sets all the **COMMAND REG** bits (and all modulation registers) to zero. Since we don't have direct control of the **COMMAND REG** this is necessary to initialize the chip correctly. The *FSELECT* pin is normally at ground so the frequency stored in the **FREQO REG** will be

Register		Cable pin number	AD7008	
Bit name	bit no.		pin name	pin no.
STB_	0(control)	1	TC0	32
AUTO_	1(control)	14	LOAD	36
INIT	2(control)	16	SCLK	41
SEL_	3(control)	17	TC1	33
D0	0(data)	2	SDATA	42

Table 2. Relationship between the signals in the control and data registers of the parallel port, the DB-25 connector pins, and the AD7008 input pin numbers.

used by default. The line may be driven HIGH to switch to the frequency in **FREQ1 REG**. When used with the 40-40 board, this line is tied to the transceiver keying signal to switch between the frequencies needed for transmit and receive (LOW = receive). Buffering of the input lines to the board from the computer isn't strictly necessary, but was included because it was felt that TTL inputs would provide a more rugged interface than connecting directly to the 7008.

The on-chip DAC has a current source output that can be used for either single-ended or differential output. The full-scale output current can be adjusted up to a maximum of 20 mA with the resistor connected to the *FS ADJUST* pin. The series-coupling capacitor between the DAC and the output filter network will attenuate low frequencies. If you need to generate output in the low kHz range you will want to increase the capacitance.

The output low-pass filter is a 5-pole Chebychev realization with components selected for 0.1 dB passband ripple. The filter's corner frequency of approximately 11 MHz resulted somewhat arbitrarily from using standard capacitor values. The frequency can be scaled up or down as desired.

The initial implementations of this circuit were breadboarded using a combination of wirewrap and soldered construction. I used a high-quality prototyping board with power and ground planes. Leads should be kept short and supplies well bypassed. This is particularly true for pins in the analog output section: *COMP*, *VREF*, *VAA*, and *AGND*. I couldn't find a 44-pin LCC socket with wirewrap pins, so I used one with standard pc solder leads and soldered wirewrap wire to them. At this writing, plans are underway to have a printed circuit board made for the circuit. Please see the notes at the end of the article for information about the availability of a board and a parts kit. I chose the AD7008 because of its performance and ease of use (other devices require an external DAC). Unfortunately, the 7008 is somewhat pricey, about \$50 each in single quantity. The chip is available from several national distributors.

The board is connected to the computer's parallel port with a cable with male DB-25 connectors on each end. There is a potential "gotcha" involving the cable (at least it got me). Many of the preassembled cables you're likely to find with these connectors are for use with an RS-232 PC *serial* port. Unfortunately, these are unsuitable for this interface because they don't contain all the conductors needed. So, either hunt for a cable with all 25 conductors, or, wire one yourself with just the pins required. Noise is a potential problem, so shielded cable is recommended.

A PC parallel port consists of eight data lines, four output control lines, and five input status lines. The five status lines aren't used in this application. A more complete description of PC ports was given in the Ramsey interface article already mentioned. Briefly, each of these pin groups—data, status, and control—corresponds to a (one byte) location in the PC's I/O address space. When you read or write these locations you read or write the parallel-port lines corresponding to that group. These three parallel-port registers are always located at consecutive addresses, most commonly starting at 378H (=888₁₀). **Table 2** gives the relationship between the signals in the control and data registers of the parallel port, the DB-25 connector pins, and the AD7008 input pin numbers. One other complication is that three of the control signals, *AUTO_*, *SEL_*, and *STB_*, are actually the inverse of what is written to the port. For example, if you write a "1" to bit 0, the signal *STB_*, cable pin 1, will be "0". The inverters on these input lines compensate for this situation.

The disadvantage of using highly integrated components like the AD7008 is that, with so much contained within the chip, diagnosing problems can be difficult. If the circuit doesn't work there are few visible clues to indicate what's wrong. Two suggestions: first, check that the computer interface is working correctly. Make sure you can set each signal line to both HIGH and LOW states by measuring voltages at the AD7008 socket. Try to set each line independently to make sure it's not shorted to

```

' AD7008 DDS test program - H. Cahn, wb2cpu, 6 Oct 1994

' Generates signals on a PC parallel port to program the frequency of an
' AD7008-based DDS board (using the '7008 serial interface).

' Program for use with Microsoft QBASIC - may require modification for other
' environments
'-----
' Program parameters -- CHANGE TO CORRECT VALUES FOR YOUR SYSTEM

' System clock frequency (Hz) -- change to actual measured value
clk& = 49999984

' Locations (decimal values) in PC I/O address space for parallel port
DataAdr = 888
ControlAdr = DataAdr + 2

' Assumes 7008 pin TC1 is connected LOW
'-----
' Get frequency (string) and convert to number (long integer)
INPUT "Enter frequency (Hz): ", freq$
freq& = VAL(freq$)

' Bit assignments for parallel port signals
'
'   Bit   7008 pin   Value Register
'-----
'   0     TC0        1     Control
'   1     LOAD        2     Control
'   2     SCLK        4     Control
'   0     SDATA       1     Data

' Figure out which bits get set for this frequency:
'   Each bit has the value of 'clock/2^(32-n)', where 'n' is bit #.
'   If bit value is less than the amount remaining in temp ; set the bit
'   and subtract the value.
'
remain& = freq&

' Shift out the bits to SDATA...
temp& = (clk& / (2 ^ (32 - i)))
IF (remain& - (temp&) > 0) THEN
  PRINT remain&, temp&, "1' - bit "; i
  OUT DataAdr, 1

  remain& = remain& - (temp&)
ELSE
  PRINT remain&, temp&, "0' - bit "; i
  OUT DataAdr, 0
END IF

' clock it out (SCLK)
OUT ControlAdr, 4
OUT ControlAdr, 0
NEXT

' Output zeros for last eight bits
OUT DataAdr, 0
FOR i = 7 TO 0 STEP -1
  OUT ControlAdr, 4
  OUT ControlAdr, 0
NEXT

' Load it into FREQ REG with pin LOAD
OUT ControlAdr, 2
OUT ControlAdr, 0

PRINT STR$(freq& / 1000!)

```

Figure 6. DDS board test program.

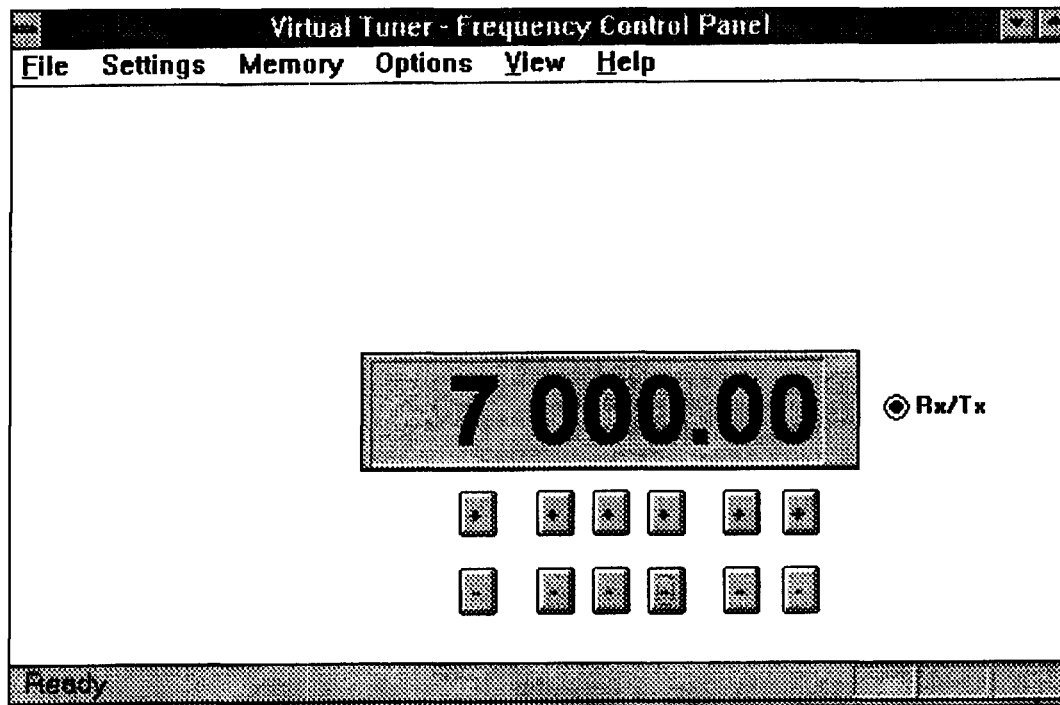


Figure 8. Display at program startup.

window at the bottom of the main QBASIC screen. Commands typed there will be executed immediately after you hit ENTER. Use the BASIC "OUT" statement to send data to an address in the machine's I/O address space. Assuming the port is at location 888_{10} , the statements "OUT(888, data)" and "OUT(890, data)" output one byte to the data and control registers, respectively. (You should verify the address of the parallel port (called "LPT:") in your machine. Usually this is given in the startup messages when the computer is booted.) Once you've confirmed the link between the machine and the AD7008, you can use the BASIC program in **Figure 6** to load frequency information into the DDS board. To keep it simple, the program is neither elegant nor bulletproof. The PRINT statements are just there to help you figure out what's going on. You can remove them when they're no longer needed. You could expand this program to use as your interface for operating the DDS board. I hope, however, you'll consider the more powerful GUI-based approach that I'll describe later.

Connecting the DDS board to the "40-40" transceiver*

Figure 7 shows a portion of the 40-40 circuit that includes the VFO, the receiver mixer, and

the transmitter mixer. The receiver mixer uses the 3-MHz VFO output to generate a 4.0-MHz IF from 40-meter input signals. The VFO is fed into the local oscillator (LO) port of the NE602 mixer (U1) at pin 6 through a voltage divider consisting of C3 and C4. The exact level of this signal isn't very critical. Inside the 602 pin 6 connects to the base of a transistor that's biased so any input greater than 200 mV_{p-p} will provide adequate LO injection. The existing 40-40 VFO supplies about 4 V_{p-p} to the voltage divider. The DDS output is about 1 volt. I changed the value of C3 to maintain the original LO drive level to the mixer, but this probably isn't necessary.

Driving the transmitter mixer is a little more critical. Here the VFO signal is fed to the RF input (pin 2) of U5, another NE602. If the signal level applied to this pin is too high, excessive distortion products will result. If the signal is too low there won't be enough gain in the transmitter's amplifier stages to produce the desired output power. The input to the 602 RF port should be below -25 dBm , about 75 mV into the 2 k impedance at this input. Capacitor C11 should be changed so the DDS board produces this voltage.

In the 40-40, the receiver 4-MHz beat frequency oscillator (BFO) and the transmitter mixer offset oscillator have been tuned to be about 700 to 800 Hz apart to produce the desired receive-to-transmit frequency offset. The frequency of the VFO, therefore, is the same for both receive and transmit functions.

*The 40-40 transceiver kit is available from Small Wonder Labs, 80 East Robbins Avenue, Newington, Connecticut 06111.

When the DDS board is substituted for the VFO, this frequency should be programmed into both **FREQ0 REG** (for receive) and **FREQ1 REG** (for transmit) when no incremental or split tuning is being used. RIT, XIT, or split frequency tuning can be accommodated by changing the values in these registers.

I've assumed that you are modifying an existing 40-40 or similar transceiver. If you're starting from scratch you can take advantage of the fast frequency shifting property of the DDS board to simplify and improve the transceiver circuit. The 40-40 transmit mixer stage generates the transmit frequency as the sum of the VFO frequency plus an oscillator at the IF frequency. Since it's just as easy to shift the DDS several MHz as it is to shift several hundred Hertz, the transmit frequency can be produced by simply shifting the DDS between transmit and receive frequencies—eliminating the need for this mixer stage. You just program the two AD7008 frequency registers with frequencies that differ by the receiver IF. For operation at 7100.0 kHz, for example, **FREQ0 REG** would be programmed to 3100.0 kHz and **FREQ1 REG** to 7100.0 kHz. Not only would this scheme remove a whole stage and its possible distortion products, but it would also remove the need for the input voltage divider—reducing the gain requirements in the transmitter amplifier chain.

User-interface philosophy

Communications engineers and hams have spent much of this century perfecting radio equipment to allow us to reliably exchange information with distant places around the globe. Today there is another challenge, one that has been mostly ignored in the amateur radio literature: how to communicate information between the operator and the radio; that is, how to build equipment that is easier and more efficient to use. The appearance of a typical home-built radio hasn't changed much in decades, despite the significant changes in the technologies available. The front panel of commercial equipment looks more modern, but is often cluttered, confusing, and inflexible. The problem isn't limited to radios. Most of us have consumer products—TVs, digital clocks, or VCRs—that are frustratingly difficult or unintuitive to use. (Maybe someone can explain why my car has a lever you push *forward* to spray the rear windshield, and *backward* to spray the front.)

The arrival of personal computers has led to an examination of these issues. There's certainly no unanimity of opinion about which computer programs and operating systems are the easiest to use, but a body of knowledge about

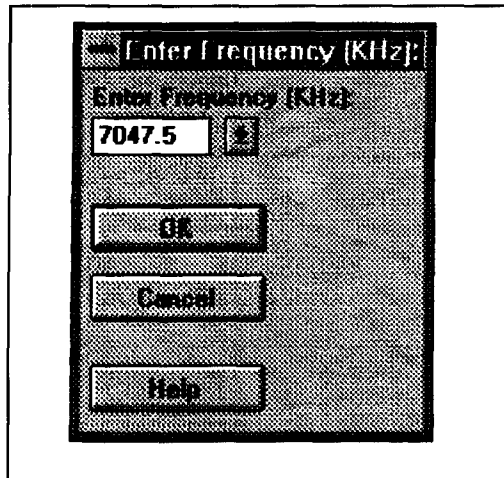


Figure 9. Frequency entry dialog box.

what constitutes good user-interface design is developing in the computer science community. While a discussion of this material is beyond the scope of this article, I would like to address the topic in the future if there is sufficient interest. For now, I'll just describe the user interface program I've come up with to control the DDS board and provide some reference material for anyone interested in exploring the larger question further.^{5,6} This design is just the start of a hoped-for dialog about user interfaces for radio equipment.

The DDS GUI

The primary components of a graphical user interface are menus, keyboard accelerators (shortcut keys), controls, and dialog boxes. Pull-down menus are pretty familiar by now and operate similarly in all the popular GUI-based operating system environments: MS-Windows, Macintosh, X-Windows, and OS/2. Menus provide an easy-to-learn method for accessing commands and are a way to hide functions not currently needed. They're easy to use because the functionality of a menu choice is usually evident from its name. In MS-Windows, additional descriptive text about a menu command can appear in a "status bar" at the bottom of the window. Navigating through levels of menus can be time consuming so, for widely-used commands, alternative, keyboard-based, selection methods are often provided. An accelerator key is a key or key combination that provides a faster way to invoke a menu function.* "Controls" are additional visual elements in the GUI provided to exchange information with the user. Commonly used controls are various kinds of buttons, labels, edit boxes, and list boxes. Dialog boxes are small (usually)

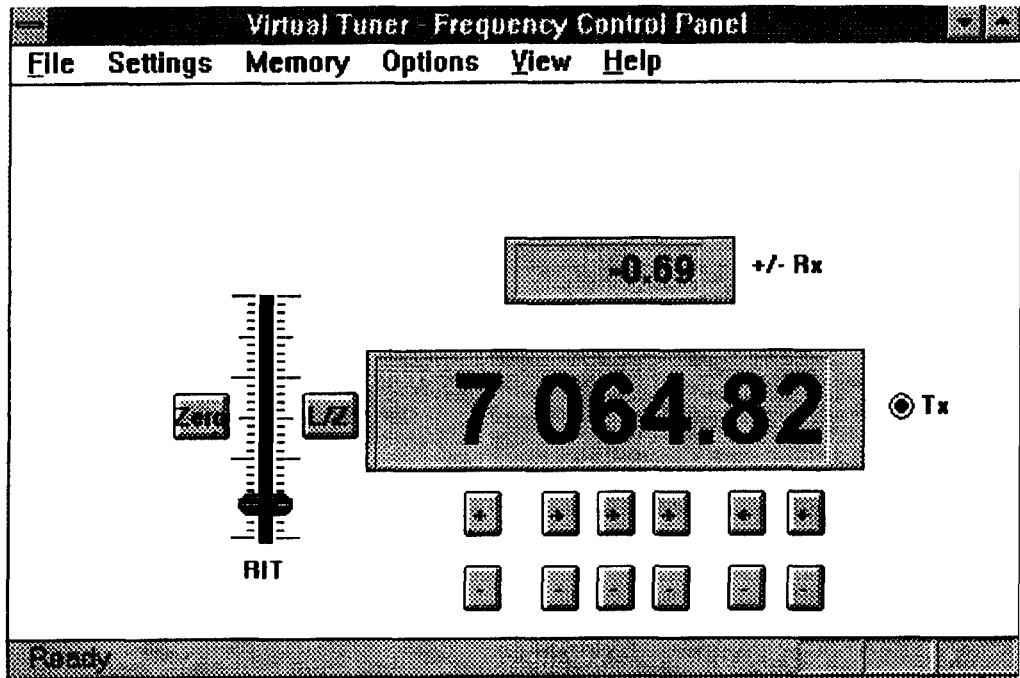


Figure 10. RIT tuning mode.

windows that pop up for a specific task. Dialog boxes are themselves controls that contain other controls.

Software is easiest to learn and use when it is based on familiar models. I tried to combine two paradigms: the Windows desktop environment and the operation of a real radio front panel. It's always desirable to follow existing conventions and standards when designing something new. Unfortunately, sometimes there are conflicts between applicable standards. This program was created loosely following the Microsoft guidelines for Windows programs,⁷ together with some elements from existing amateur radio software—including the widely used contest logging program *CT*, by K1EA (a DOS program), and the Windows logging program *WriteLog* by W5XD.

Basic tuning

Tuning is the most important function the GUI must provide; replacing the knob that is, after all, the *only* way of changing the frequency of a non-digital radio. A physical tuning knob has several definite advantages. It provides visual and tactile feedback, is well adapted to the hand for precise control, and is a good

mechanism for translating a large, linear frequency space into a localized, rotational space. I've incorporated into the GUI panel several mechanisms to try to replace the functionality of the knob. I think they produce a way to tune the radio that's quite usable. At this point it may not match the knob in applications where the highest performance and operating efficiency are required, such as in competitive amateur contesting. I'm hoping that as we explore computerized user interfaces further we can eventually come up with tools that are demonstrably as good as or better than a tuning knob.

Two primary tuning mechanisms are included in the GUI, with a third to be added later. The simple display, shown in **Figure 8**, is seen on program startup. It consists of the standard Windows menu bar, a "text" control to display the current frequency, and a series of "push buttons" to increment or decrement the frequency. The buttons work like auto-repeat keys on an electric typewriter. If you hold one of the push buttons down, the frequency continues to increment or decrement (at a user-selectable rate) until the button is released. As with most Windows controls, it's not necessary to use a pointing device to operate the buttons. You can get to the desired button with the TAB key and activate it with ENTER—although this can be slow. A better way of setting the frequency completely from the keyboard is provided with the Ctrl+F key combination. This brings up the dialog box in **Figure 9**. It takes the same number of keystrokes to use as a pure "command-

*Accelerator keys are often confused with "access keys." If you look at a Windows menu or dialog box, often one letter is underlined in each function name. This is the access key. Typing "Alt" plus that key will invoke the function. Access keys are slow and are usually only used when no pointing device exists, whereas accelerator keys are used for faster access.

line” approach would. To enter a particular frequency you type: “Ctrl+F” (invokes the dialog box), “7047.5” (the frequency appears in the dialog’s “edit box”), and ENTER (the dialog box disappears and the new frequency is shown in the main display control). An additional benefit of this approach is that the edit box is part of a “list box” that stores the last several frequencies entered into it for later recall. The tuning method I plan to add uses only the cursor arrow keys. You can assign these key pairs to increment/decrement the frequency. For instance, you might have the right and left arrows control the frequency in 0.1 kHz steps and the up and down arrows producing 1.0 kHz changes. Alternatively, you could use an arrow pair to switch between frequency ranges, bands, or to select from the memorized frequency lists described below.

Advanced tuning—RIT, XIT, split frequencies

The AD7008 can switch almost instantaneously between the two frequencies that have been programmed into the **FREQ REQ** registers. This is the basis for several forms of offset tuning: receiver incremental tuning (RIT), transmitter incremental tuning (XIT), and pure split frequency operation. The screens that allow you to control these modes are shown in **Figure 10** for RIT and **Figure 11** for split frequency operation (the similar screen for XIT

can be seen in **Figure 1**). As with most other DDS GUI functions, you can select one of these modes either through a menu selection or with a Ctrl key + letter key accelerator. In the *incremental tuning modes*, you’ll notice a slider control to adjust the offset. For split operation, the main frequency readout displays the receiver frequency and a smaller text box appears above it for the transmit frequency. You can bypass the mouse completely with some additional accelerator keys. There are key assignments to bring up dialog boxes that change the RIT or XIT value, or, change one of the components of the split frequency pair. Two additional buttons were added to aid incremental tuning. The “Zero” button provides a quick means of zeroing out the offset slider control. Similarly, the “L/Z” button first loads the current frequency (the main display plus the offset) into the main display, and then resets the offset slider and offset readout to zero.

Memory

I’ve never been able to figure out the value of radios that have, say, 100 memory locations that are accessed by number. It seems to me that it just complicates things to have to either remember that memory location “67” contains “14.260 MHz,” or have to scroll through the whole memory to find the frequency. A computer-controlled memory system, on the other hand, can add memory features that are truly

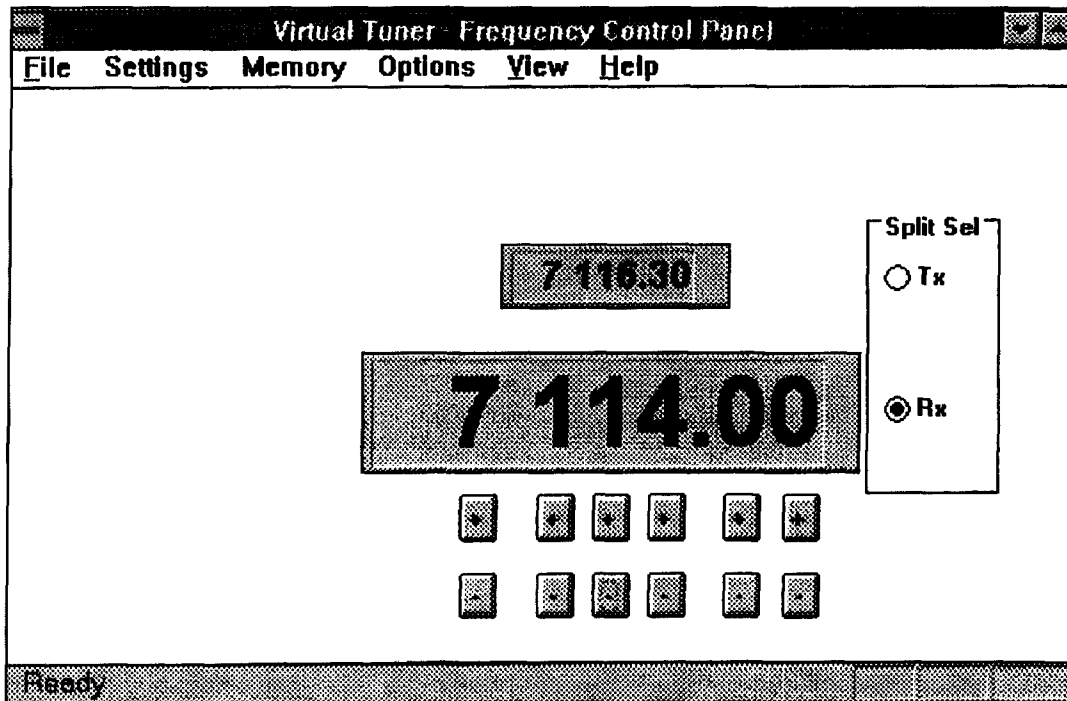


Figure 11. Split frequency mode.

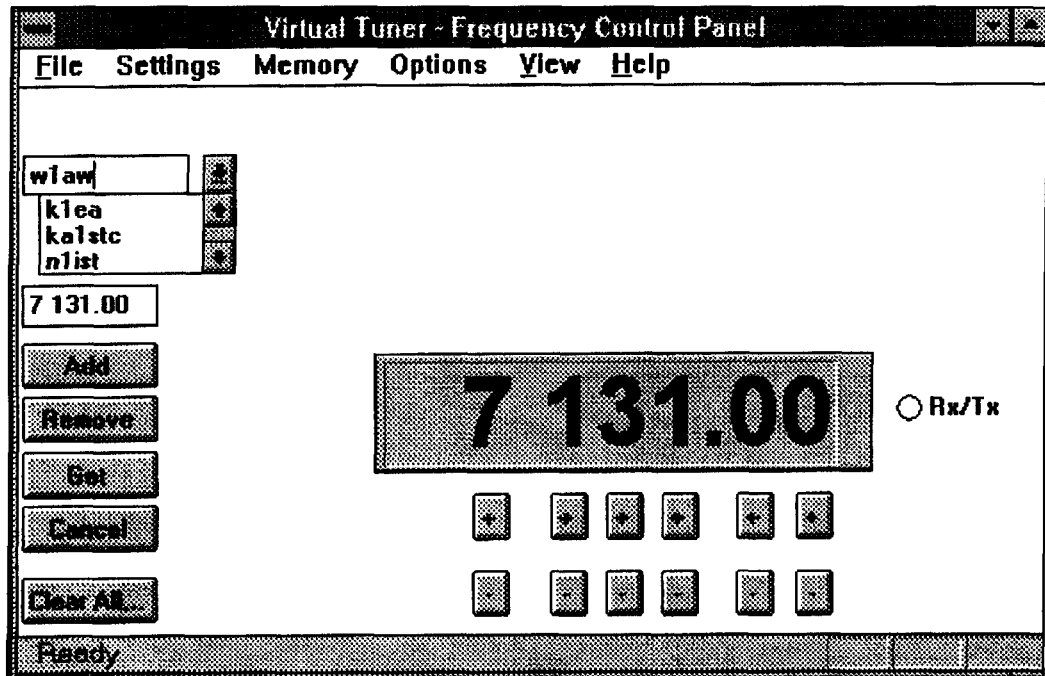


Figure 12. “Tag” memory controls.

useful. I provide two mechanisms for recalling previously stored frequency information; one, I call “tag” memory, and the other “scratch pad” memory. Both are primarily accessed using a pointing device, but keyboard access is provided for some operations where speed is desirable or when no pointing device is available.

Tag memory is typically used for long-term storage of things like net frequencies (XYZ Traffic Net, 3.892 MHz), the station you talk to every Tuesday morning (K1EA, 14.002), or standard calling frequencies (QRP-40M, 7.040). To add to your list, invoke the screen shown in **Figure 12** from the menu or use Ctrl+T. You enter the text (the “tag”) you want associated with the frequency in the edit box at the top of the dialog box. When you click on the “Add” button the name is added to the tag list and complete configuration information (frequency, offset, program mode, etc.) is stored with it. Recall information either by selecting an item in the tag list box and clicking on the “Get” button, or, with the key sequence Ctrl+G, “tag name,” ENTER. Only as much of the tag name to make it distinguishable from the other list items need be entered. Tag memory information is saved between program sessions.

The other form of memory provided is meant for short term storage—say, you’re tuning around a band and want to make note of stations or frequencies to come back to later. Scratch pad memory is meant to be used instead of writing the information on a slip of

paper. The scratch pad dialog box is shown in **Figure 13**. There are four slots for stored frequencies. If you just want to record the frequency, click on the “Store” button of the desired slot. If you want to save additional information there is a line at the top of the box resembling a line from a log book where you can type in this data. When you click on one of the four “Recall” boxes, the corresponding stored frequency is loaded into the display and sent to the radio. Scratch pad information is not saved between program sessions.

Program configuration

The program lets you supply information needed to work with your specific radio: the receiver IF frequency, the exact DDS clock frequency, band definitions, etc. Also, you can specify other operating parameters like the repeat rate of the increment and decrement push buttons and the arrow cursor key assignments mentioned above. These are all set with menu choices in the “Options” sub-menu. It would be tedious to have to reenter all this information every time you started the program. To avoid this the program stores the data in a file using a method called *serialization*, the mechanism used by most Windows programs to store configuration data. In the Windows directory structure you may have noticed many “.ini” files. These contain the serialized data

from Windows programs you have installed on your system. Using this program will automatically create and maintain the file "vir_tun.ini".

The programming environment

I've described the functions provided by the user interface without saying anything so far about how to create it. There are several approaches to programming Windows GUIs. The techniques can be grouped in a three-level hierarchy ranging from most powerful and difficult to the easiest but most limited. In the first category is a method of programming that requires low-level function calls directly to the operating system to control all aspects of creating and maintaining the displays. This is typically done using the languages C or C++ with the Microsoft System Developer's Kit, so it's often referred to as SDK-style programming. At the next level up, the functionality needed to create the displays is encapsulated into *classes*, the basic components of *object-oriented* programming. Classes are supplied by compiler providers in class libraries such as Borland's Object Workbench Library (OWL) or Microsoft Foundation Classes (MFC). Classes are used within an object-oriented language such as C++. Finally, there are tools that automate the process of building GUI-based applications, simplifying programming even further. Microsoft's Visual Basic is probably the most popular example in this group. Someone who has done some programming might find it takes months to learn SDK programming, weeks to learn to use class libraries, and days, or even hours to use the higher-level tools. The level of program complexity that can be supported, as well as the degree of control you have over the environment, generally goes down as you move up the methodological hierarchy. In the Ramsey article, I described programming in Visual Basic. This interface was done in Visual C++ using MFC. Because the number of readers who are likely to want to recreate the program using these exact tools is probably small, I won't provide listings here. My source code is available for anyone interested.

Summary

DDS is already a powerful technology that will become even more so as DAC clock speeds and resolutions increase and chip costs come down. Devices with sample rates greater than 50 MSPS and with 14-plus bits of output resolution probably already exist in some manufacturer's lab. When they become commercially available the remaining arguments against DDS as the method of choice for HF frequency generation should be eliminated.

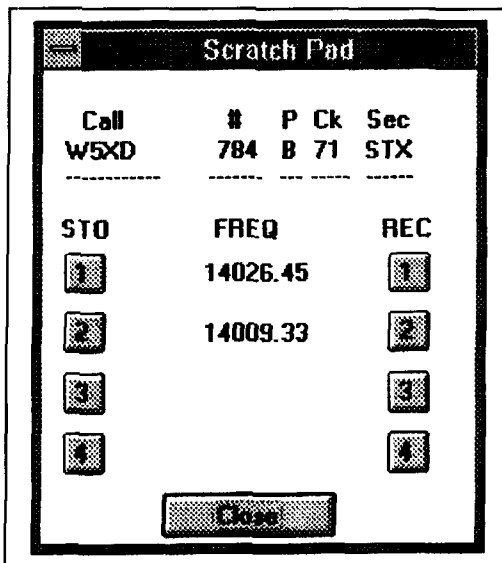


Figure 13. "Scratch Pad" memory dialog box.

This article tried to expand on ideas presented by previous writers in *Communications Quarterly*: Rick Littlefield on digitally adjusted components and Bryan Bergeron on computer-controlled "virtual instruments".^{8,9} I have talked about what I think are the advantages of these approaches.¹⁰ The project was an attempt to lay the foundation for building radios based on these techniques. The design of the DDS graphical user interface continues to evolve. As I use it, I realize how something could be done in a better way, or I think of a new feature that would be useful. The ultimate goal is a radio that offers cost, ease-of-use, performance, and operating efficiency advantages over existing equipment. I have several projects in the works along these lines. One uses a commercial "SoundBlaster"TM type board to provide DSP audio filtering and sideband generation; another replaces the "40-40" board with a radio core that achieves greater performance and simplicity by using more integrated components. This project has reinforced my belief that computer-controlled radios using these new devices make it feasible for hams to build equipment comparable to or better than what is available commercially.

Acknowledgments

Thanks to Brad Brannon, N4RGI, of Analog Devices, Inc., and Neil Weste, AA1KR, of TLW, Inc. for assistance with the AD7008. Thanks to Mike Ardai, N1IST for proofreading the article.

Conventions used

Three different number systems, decimal, hexadecimal, and binary are used in this article.

A hexadecimal number ends with an uppercase 'H' and a binary number consists of only 1s and 0s. All other numbers, or numbers with the subscript "10" are decimal. An "*" within formulae indicates multiplication.

Digital signal or pin names are in *italics*; a trailing underscore ("_") indicates that the named signal function is active when in the LOW state. For example, the TTL signal "RESET_" will perform the reset function when driven to zero volts. Names of variables are in **bold** type. Register names are **BOLD CAPITALIZED**.

"PC" denotes an IBM-compatible computer and "pc" is a printed circuit board; although these abbreviations may not be "politically correct."

Addendum

At press time, the author was working on a source for pc boards. Check with Howie about

the possibility of a board and/or parts kit. A disk containing an executable version of the program described (requires MS-Windows) together with Visual C++ source code is available for \$5.—Ed.

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PRODUCT INFORMATION

S-COM Repeater Controllers

S-COM Industries offers three repeater controllers (5K, 6K, 7K).

The 5K repeater controller offers a CMOS HD63B09 (MC6809 compatible) microprocessor; 8 K bytes non-volatile CMOS RAM; 16K bytes CMOS EPROM; an MT8870, full 16-digit, crystal controlled DTMF decoder; 3 logic inputs, 3 logic outputs; Tranzorb™ suppressor and ferrite bead on power input; and Tranzorbs™ on all power MOSFETs, and a 1 year warranty on parts and workmanship.

The 6K is based on the model 5K. The 6K retains the 5K's powerful feature set and has nearly the same connector pinout—but many new features were added. You get the 6K main board and an FCC-registered telephone interface module, mounted in a low-profile 1-3/4" by 19" rackmount cabinet. The cabinet has additional room for up to two optional audio delay modules.

New features include: an autopatch that dials in pulse (rotary) format at 10 or 20 PPS, or regenerated DTMF (TouchTone™); accepted and rejected number tables for complete dialing security; autodial features, using S-COM's exclusive Macro command capability; reverse patch features; phone line control and phone line programming; larger memory capacity to accommodate future software upgrades; and pro-

grammable speed and pitch of CW messages.

The 7K consists of a main board and rack-mount cabinet as standard equipment. It accommodates two optional boards mounted in the chassis—the telephone interface module (for autopatch, reverse patch, and control by phone line), and the speech synthesizer module, and two optional audio delay modules (one each for repeater audio and link audio) may be mounted in the cover.

For more information, contact S-COM Industries, P.O. Box 1718, Loveland, CO 80539 (303-663-6000).

New AES 1995 Catalog

Antique Electronic Supply is now distributing its 1995 catalog covering electron tubes and parts and supplies for collectors and hobbyists. The catalog has been expanded to 36 pages listing over 3000 types of receiving, audio, transmitting, and industrial tubes plus an extensive line of capacitors and other items needed for repair or construction of tube type electronic equipment. Also in the catalog are over 75 book titles covering collecting, restoration and related subjects.

For further information, contact Noreen Cravener, Antique Electronic Supply, 6221 South Maple Avenue, Tempe, AZ 85283 (phone: 602-820-5411; Fax: 800-706-6789).

BEYOND THE Z-MATCH

The IBZ Coupler

The hunt continues for an ideal “antenna tuner” or, more accurately, an antenna system tuning unit (a.s.t.u.). Both Bill Orr, W6SAI, and I have recently reported on new versions of the Z-match coupler, which dates back to the 1950s and uses a multiband tank circuit to cover 80 through 10 meters.^{1,2} Here I offer another improvement—one that draws on two classic designs to produce the improved balanced-Z match or “IBZ Coupler.”

The need to press on arises from a weakness that the Z-match shares with the more common approach of using an input or output balun in conjunction with a single-ended network (such as the popular T-match). These designs all provide balanced output—but only within limits. When the feedpoint impedance seen at the tuner’s “balanced” output terminals exceeds several hundred ohms, current imbalance on the feedline begins to appear. When the impedance at the tuner reaches several thousand ohms, the imbalance may become serious, with significant results. Also, with highly efficient antenna systems using open-wire or ladder line (like the multiband “center-fed Zepp”), high impedances do occur at the tuner under some conditions.

Although the extent of the current imbalance depends on the tuner’s design and construction, along with the frequency of operation, imbalance proves unavoidable with the Z-match circuit and with single-ended networks using baluns. The current on one side of the feeder may be 50 percent higher than on the other side—approximately a 3.5 dB difference, assuming the load on each side is equal. Occasionally, the difference is greater. (For a fuller assessment, see the appendix on “A.S.T.U. Output Imbalance.”)

While the practical significance of such feed-

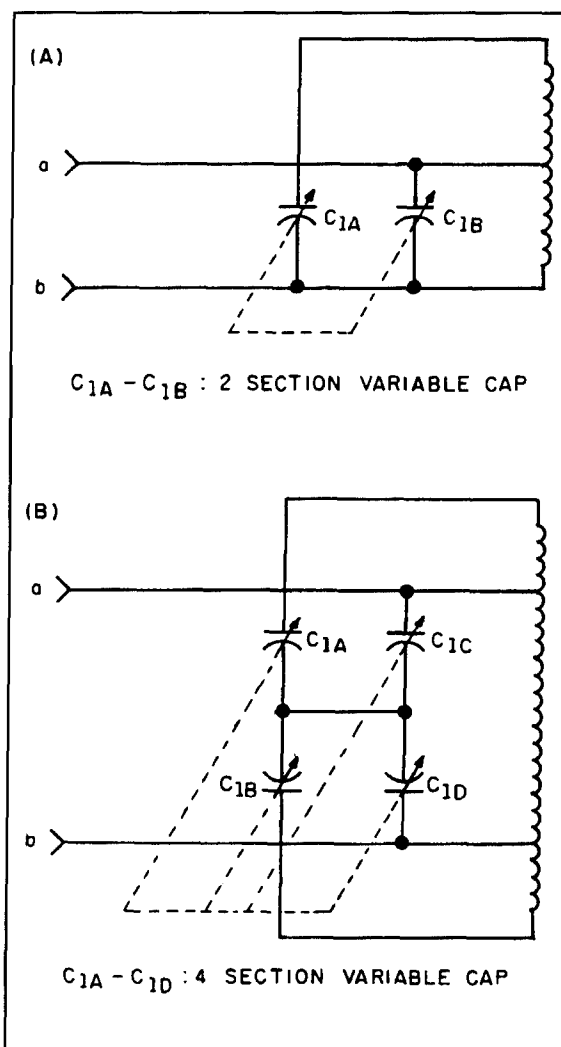


Figure 1. Single-ended and balanced multiband tank circuits. (A) Single-ended tank circuit, simultaneously tuned through two ranges (see Reference 5). (B) Balanced tank circuit, simultaneously tuned through two ranges.

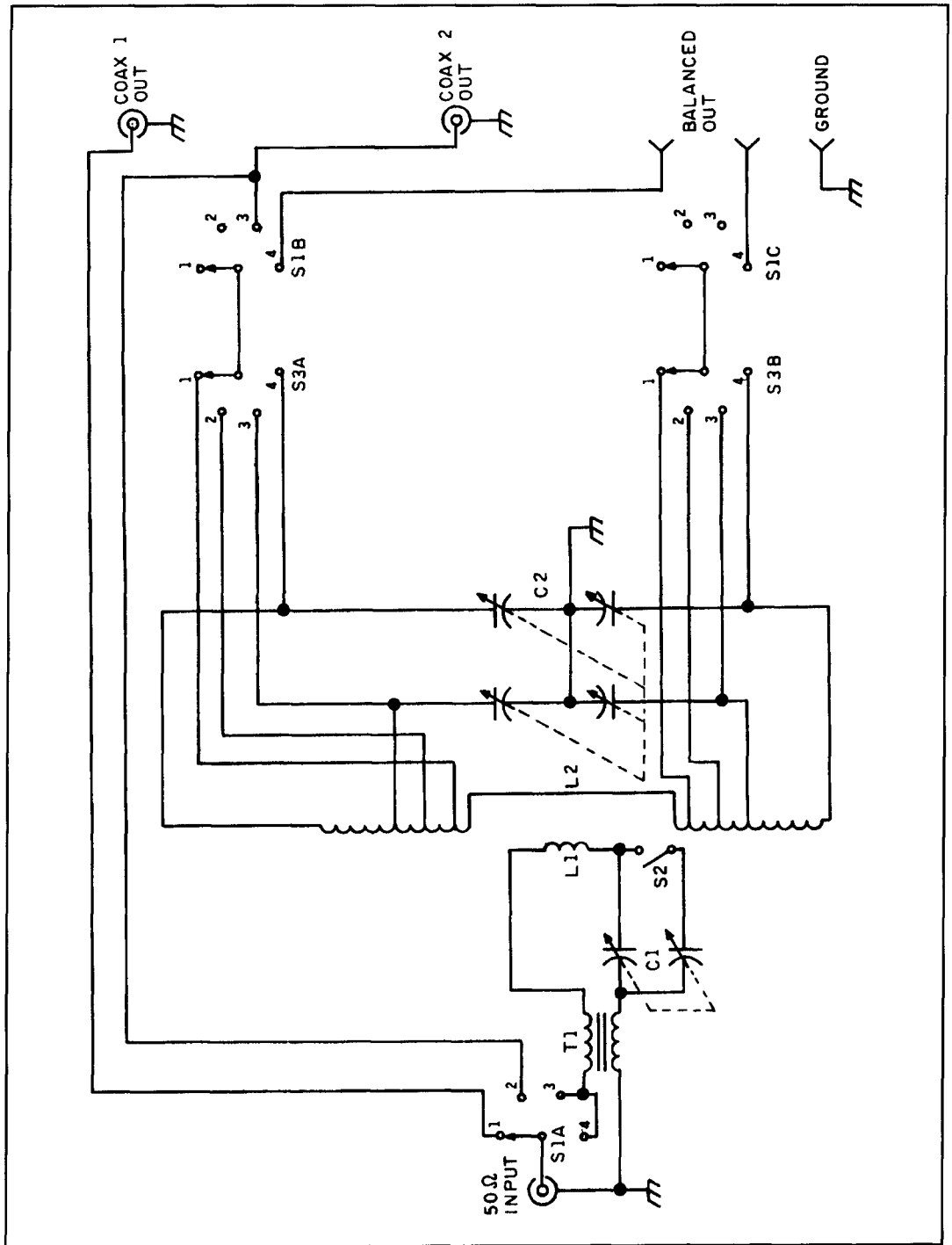


Figure 2. The IBZ Coupler.

line imbalances can be exaggerated, it is hardly trivial. With a substantial imbalance, the feedline radiates nicely, RFI problems sometimes increase, and RF more readily shows up in the shack. In addition, antenna directivity may suffer, owing to the combination of feedline radiation, current imbalance at the antenna feedpoint, and often-overlooked current phase shift that accompanies the current imbalance and

results from unbalanced reactances to ground in the tuner.

The "IBZ" solution

Happily, there's a solution. It combines the virtues of the Z-match with those of the standard link-coupled, balanced output a.s.t.u. that's been around since pre-World War II

days,^{3,4} and eliminates the weaknesses of each. As every ham knows—or should know—the “old-fashioned” link-coupled tuner offers the best approach to feeding balanced line. Its drawback is inconvenience due to a need for coil changing, bandswitching, and/or tap changing. The remedy is to replace its tank circuit with a balanced version of the multiband circuit that constitutes the heart of Z-match. The result is the IBZ Coupler, which provides convenience of operation *and* good feedline balance—even at high impedances. With minimal switching, the IBZ Coupler also allows coaxial output and is as flexible as the T-match and its cousins.

For background on the IBZ design, refer to **Figure 1A**. It shows how a single inductor may be tuned simultaneously through two ranges, with a parallel resonant circuit appearing between terminals “a” and “b.”⁵ In **Figure 1B**, the single-ended circuit has been modified into a balanced version. Besides providing the basis for balanced output, this version has an important advantage from the standpoint of a.s.t.u. design. It has a lower C/L ratio than the single-ended version, and hence is less susceptible to losses from high circulating currents in the resulting tank circuit (another problem with the Z-match). With the appropriate component values, the circuit will simultaneously cover approximately 3.5 to 12 MHz and 10 to 40 MHz.

In **Figure 2**, the multiband tank circuit of **Figure 1B** replaces the link-coupled a.s.t.u.’s usual tapped-coil bandswitching. As a result, band switching is accomplished by tuning C_2 through its full range, which simultaneously tunes the multiband circuit from 80 through 30 meters and 30 through 10 meters. (In practice, the 30-meter band appears at either the high or low capacitance end of C_2 , or at both ends, depending on the reactance reflected from the feedline into the tuner.) The extended high-end frequency range of the circuit allows easy matching on 10 meters in the presence of unpredictable reactances. The coil taps in **Figure 2**, along with switch S_3 are *not* for bandswitching, but instead provide a choice of high and low impedance outputs. Residual reactance is tuned out with C_1 .

The balun in the input circuit (T_1) is a 1:1 current balun consisting of 10 turns of RG-174 on an FT-140-61 ferrite core. Unlike a balun on the input side of a *single-ended* a.s.t.u. (such as a conventional L-network or a T-network), a balun at this point does *not* also appear as a reactance between one side of the *output* and ground. Here the balun really is in a place for which it is designed. True, the standard link-coupled a.t.s.u. does not have an input balun. Tests indicate, however, that including the balun improves output balance under some

Parts List

C_1	350 pF or higher per section, “broadcast” variable (see text and footnote on switching).
C_2	20–500 pF per section, 4 section (Fair Radio Sales no. 4G-535).
L_1	4 turns #18 or #20 enameled on T-225-6 core, close-wound between two halves of L_2 .
L_2	28 turns #14 copper or tinned, 14 turns on each side of L_1 . Tapped at 1, 4, and 7 turns each side of center.
S_1	3 pole, 4 position ceramic rotary.
S_2	See text and footnote on switching.
S_3	2 pole, 4 position ceramic rotary.
T_1	1:1 balun (see text).

conditions, even in a balanced design like this one. (In fact, anyone with a standard link-coupled tuner might consider including an input balun to improve overall balance.)

Regarding the switching arrangements, S_1 allows both coax and balanced output, as well as tuner-bypass operation. Single-wire feed is accommodated by using either side of the balanced output. A more elaborate switching arrangement at S_1 would allow additional output options (such as two tuner-or-bypass coax ports, two balanced-feed ports, and/or a separate output for single-wire feed).

Alternatively, if you don’t want the option for coax output, you may omit S_1 and the associated circuitry. The result would be a “dedicated” tuner for balanced output. In this instance, the frame of C_2 may be left “floating” above ground, because there is no need to provide a ground-return circuit for coax and single-wire output. (However, tests show that grounding the frame of C_2 provides somewhat better output balance at high impedances *when the load itself is well-balanced*. If the load is somewhat unbalanced to ground, better output-current balance results when the frame “floats.”)

Switch S_2 provides the additional capacitance at C_1 often needed on the lower frequency bands. The switching arrangement at S_2 may be varied as required by the capacitance and number of sections available in C_1 . It should allow one section of C_1 to be used alone when necessary for low minimum capacitance, along with the option of switching in additional sections and/or padding capacitors, for a maximum capacitance of 1200 pF.*

*For example, if you are fortunate enough (as I was) to find a two-section capacitor with 600 pF per section for C_1 , the most suitable arrangement is to use a d.p.d.t. switch for S_2 , wiring it to switch the two sections of C_1 from series to parallel. With the series setting, the maximum capacitance becomes 300 pF, and the minimum capacitance is significantly less than allowed by one section alone. This is sometimes useful on the higher bands. If C_1 has three or more sections, then a three-pole, double-throw switch can be used to change between two of its sections in series and all the sections in parallel. (In either case, working out the appropriate wiring is good mental exercise.)



Photo A. The IBZ Coupler mounted in bottom-opening cabinet.

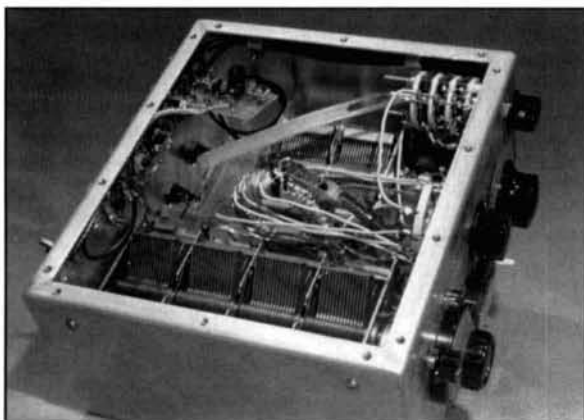


Photo B. The variable capacitors and T_1 are mounted on a subchassis of Plexiglass™.

Switch S_3 selects either high or low impedance outputs. Position 4 is the highest impedance output on the multiband circuit's low-frequency range, and position 3 is the high impedance output on the high-frequency range. As explained above, S_3 is *not* a bandswitch. "Bandswitching" is accomplished using C_2 .

As for component selection, C_1 is a "broadcast"-type variable with two or more sections. Because C_1 is in a low impedance circuit, its plate spacing is not critical, but adequate rotor wiper contacts are important. Capacitor C_2 is a four-section unit that is available on the surplus market and often shows up at swap meets.* (The same type of capacitor may also be used for C_1 by employing three of its sections.***) Because of the high RF voltages in the multiband tank circuit, C_2 's plate spacing sets the tuner's power-handling limit.

Switches S_1 and S_3 should be ceramic rotary units. For S_2 , which is at a low-voltage point, it's okay to use phenolic or plastic insulation or

casing. (I find that a small toggle switch is adequate at the 50-watt level.)

Construction and operation

Construction details aren't critical, but leads should be kept short. Where this is difficult owing to layout, copper strap or shield taken from coaxial cable should be used for low-inductance leads *within* the multiband circuit as well as between the balun/ S_1 and link coil L_1 . In the interest of balance on the higher frequencies, parts placement should allow approximately equal-length leads on each side of the output of the multiband circuit.

A vernier drive is useful for C_2 , the four-section capacitor, because tuning can be sharp. For the inductors on the T-225-6 powdered-iron toroid, L_1 is wound first, and then one half of L_2 is wound on each side of L_1 . Leave about 1/8 inch as the minimum spacing between the outside turns of L_1 and the inside turns of L_2 . (Care in this regard helps output balance on the higher frequencies.) Space the turns of L_2 so they are about a wire diameter apart on the inside of the toroid, in order to leave a spacing of about 3/4 inch between the opposing ends of L_2 . Because the core is nonconducting, insulated wire is not required for L_2 , and either bare or tinned copper wire is suitable. In the interest of low loss in the tank circuit, **Figure 2** specifies no. 14 wire for L_2 , but the difference in practice will be slight if no. 16 is used, and no. 16 is much easier to wind on the core. (If no. 16 is used, the turns spacing will be greater than a wire diameter.) If the switch used for S_3 has five or six positions per pole, rather than the four that are specified, use the extra positions for more taps on L_2 . This will provide additional mid-range impedance options. In any event, retain the taps at seven turns each side of center that each go to a section of C_2 , as well as to S_3 . For T_1 , the input balun, strip the outer insulating cover from the RG-174, and space the 10 turns so they do not short. (The FT-140-61 ferrite core itself is nonconductive.)

In the unit shown in **Photos A** and **B**, the variable capacitors and T_1 are mounted on a subchassis of 1/4-inch Plexiglas™. The subchassis, in turn, is mounted on short spacers inside the top of the bottom-opening cabinet. The cabinet, which I found at a swap meet, measures 4-1/2" H x 9" W x 8" D. (A slightly larger cabinet would allow for less crowded construction.) The toroid for L_1/L_2 is not mounted on the subchassis as its short, heavy leads to C_2 make it self-supporting. The unit, as pictured, has additional output options at S_1 (as suggested above) and includes a built-in SWR bridge for operating convenience. The meter came from an old Heathkit SWR bridge and,

*Fair Radio Sales lists a suitable capacitor in their catalog (part number 4G-535).

**Refer to footnote on switching.

like the cabinet itself, went through several earlier iterations.

To adjust the IBZ Coupler, set the input capacitor at midpoint and peak the multiband tank circuit on receiver noise. Then apply low power and adjust the input and tuning capacitors to obtain an SWR of 1:1 between the IBZ Coupler and the rig. Interaction may occur between the adjustment of the input capacitor and the tuning capacitor. If a low SWR (1.5:1 or less) proves impossible to obtain with the impedance switch set on the high impedance position (position 4 for the low-range bands, position 3 for the high-range bands), use a lower impedance position. For greatest efficiency, select the highest impedance tap that produces a match; however, do not use position 4 on the high bands (from 30 meters at the high-capacitance end of C_2 on through 10 meters). If it's possible to tune 30 meters at both the low and high-capacitance ends of C_2 , use the low-capacitance setting for maximum efficiency. After the appropriate settings are logged, band switching is quick and easy.

I've used the IBZ Coupler with good results on several antennas, and the Zuni Loop QRP group has put it through its paces on three successive Field Days—with great success. The moral of the story is that you need not sacrifice versatility and performance to properly feed coaxial cable *and* balanced lines.

Acknowledgements

For their suggestions, help, and encouragement in moving this project from design and construction to the written word, I am indebted to Fred Bonavita, W5QJM, Richard Fisher, KI6SN, and Cam Hartford, N6GA. Bob Spidell, W6SKQ, now a Silent Key, also provided much-appreciated encouragement.

Appendix: A.S.T.U. Output Imbalance

Experiments with both Z-matches and more common tuner designs using baluns show that substantial output imbalance results when the a.s.t.u. sees a high output impedance. The impedance level at which the imbalance becomes significant depends on the tuner and may be anywhere from several hundred ohms up to a thousand or several thousand ohms. Such impedance levels at the tuner's output can easily occur in commonly used balanced-feed antenna systems. This appendix briefly explains the sources of the problem and how to check tuner output performance.

With the Z-match, the main culprit is capacitive coupling between the single-ended tank

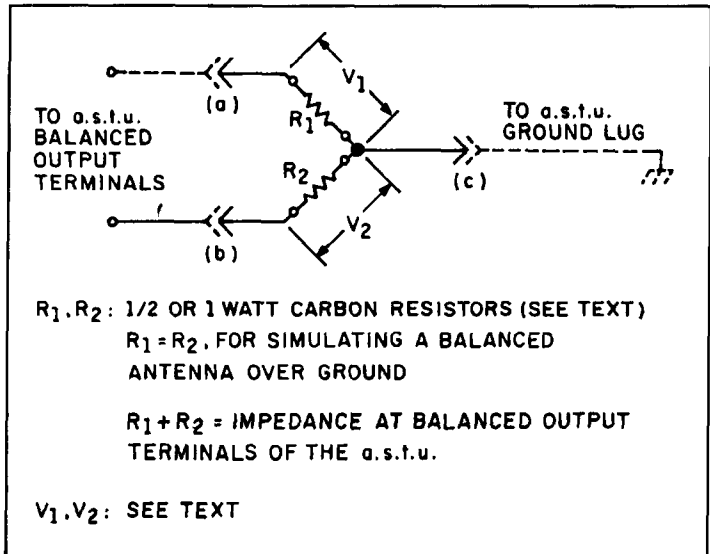


Figure 3. Antenna system simulator (see appendix).

circuit and output link. This is especially a problem when the link is wound directly over the tank coil rather than spaced concentrically away from the coil either inside or outside of it by 1/4 inch, or so.

Regarding the T-network, which is today's most common matching network in commercial and homebrew designs, the problem comes from use of a balun in circumstances for which it is not suited. The same is true for the L-network and the SPC match. In each of these single-ended networks, as many critics have noted, the use of a balun to provide balanced output is poor practice. Depending on power level, the balun's design, and the output impedance seen by the a.s.t.u., a balun may suffer from heating or core saturation or produce feedline imbalance. To be sure, core saturation is not a problem at low power levels; but whatever the power, serious feedline imbalance is likely under some conditions when using a balun with tuned feeders. A current balun on the *input* side of an L-network or a T-network has sometimes been suggested as a solution, but it actually is not a cure.

In the case of either input or output baluns, the imbalance results from insufficient isolation between one side of the supposedly balanced output and ground. This is fairly easy to conceptualize regarding an output balun, but even with a balun placed at the input of a single-ended (unbalanced) network, the balun also appears as a reactance or leakage path between the "lower" side of the output and ground. (As conventionally drawn, the schematic of a single-ended network with an input balun does not immediately disclose this leakage path, but the schematic can easily be redrawn to show the balun appearing between the lower side of the

output and ground.) In other words, it is only partly true to say that when a balun is placed at the input of a single-ended network, it is used in the kind of low or medium-impedance, resistive circuit for which it is designed.*

It isn't difficult to check a.s.t.u. output balance. I've described how to measure feedline imbalance in actual antenna systems when feeding a line at a high impedance point in **Reference 6**. However, the simple "antenna system simulator" shown in **Figure 3** provides an easier approach to checking a.s.t.u. output performance. For balance-checking purposes, the circuit adequately replicates an antenna system comprised of a feedline and a horizontal antenna over ground (as opposed to one in free space). R_1 and R_2 should be 1/2 or 1-watt carbon resistors. For easy replacement, connect them to binding posts on a scrap of perfboard. In a real system, of course, reactances are often present in the impedance seen by the tuner, but the simulator still provides revealing "first-cut" results.

To simulate a balanced antenna system, R_1 in **Figure 3** should equal R_2 , with the sum of these two resistances equaling the impedance that the a.s.t.u. sees at its output terminals. To use the simulator, attach leads "a" and "b" to the balanced output terminals on the tuner, and attach lead "c," from the junction of the two resistors, to the tuner's ground lug. Then, applying just

enough RF power to obtain a reading, use an RF probe and a digital voltmeter to measure the voltage drops across R_1 and R_2 . (Less than a watt suffices.)

If the voltage drops across R_1 and R_2 (indicated as V_1 and V_2 in **Figure 3**) are equal, this indicates equal current through the two resistors. Unequal voltage drops indicate proportionately unequal currents and thus the presence of a leakage path between one side of the "balanced" output and ground. (Strictly speaking, the leakage path appears in parallel with one of the resistors, lowering the net impedance to ground on that particular side of the supposedly balanced output.)

For interesting results, make R_1 and R_2 2000 ohms each, to simulate a balanced resistive load of 4000 ohms. Then test various antenna tuners, both commercial and homebrew. The balance between the two sides of the output should be within about 12 percent (or 1 dB) across the full frequency range of the tuner. You'll see why the search for the ideal a.s.t.u. continues, and why the IBZ design presented here is so advantageous.

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5. R.W. Johnson, W6MOR "Multiband Tuning Circuits," *QST*, July 1954, pages 25-28 and 122; provides a full explanation. For applications of the design in single-ended Z-match circuits, see **References 1 and 2**.
6. Charles Lofgren, W6JJZ, "A note on feedline balance," in "The Z-Match Coupler—Revisited and Revised," *The ARRL Antenna Compendium*, Volume 3, page 194.

*In this regard, the usual rule-of-thumb is that the series (or winding) reactance of a balun should be at least five times the impedance of its load. We can take this figure as a guide for the minimum reactance that can be safely placed between the lower side of a single-ended network and ground without causing significant feedline imbalance. If we assume that the impedance at high voltage points along the feedline may easily reach 2000 ohms (a low estimate) in a tuned-feeder antenna system, then the balun's series reactance should be a minimum of 10,000 ohms, which is unlikely to be realized across the full frequency range of the a.s.t.u.

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MODELING AND UNDERSTANDING SMALL BEAMS

Part 1: The X-beam

Between miniature loop antennas and full-size arrays exists a special class of antennas: the small beams. To some, small beams represent the search to achieve something for nothing. Like lotteries and sweepstakes, they receive as much moral disapproval as technical discrediting. To others, they are a necessity, because some improvement is better than none. Space and money restrictions are the chief motivations to build a mini-beam of one design or another.

For me, small beams are fun. Many types of small beams have received only casual empirical development. Their performance limits remain a mystery. Computer modeling can estimate with buildable accuracy antenna gain, front-to-back ratio, overall azimuth and elevation patterns, and feedpoint conditions. For under \$30 per antenna, with hardware store parts, I can test what modeling tells me ought to be the case.

In this series, I'll look at some contemporary small beams. My purpose is not to recommend any of the designs. Rather, the goal is to understand what each design can and cannot do. Only then can you assess what each beam's place may be in the scheme of amateur antennas.

Standards of comparison

Small beams appeal to the builder because they are manageable. They are "hands-on" devices, not set-and-forget store items. On 10 meters, where all the beams in this series were

modeled and tested, dimensions are reasonable as one-person projects. Moreover, the results are generally scalable down to at least 20 meters with little adjustment. Since 10 meters is closed part of the day, even during sun spot maxima, it's possible to test the antennas without causing QRM.

This series will contain extensive notes on the modeling aspects of these investigations to show the evolution of various designs. These notes may also prompt kindred spirits to try some directions not taken here, and to uncover some possibilities that have eluded me. Modeling data in this series will often be given for free space (to permit ready general comparisons), at 20 feet (a common height for portable antennas), and for 35 feet (a common permanent installation height).

There are several general propositions about small beams that underlie both our knowledge and our mythology:

1. Small beams generally have narrower bandwidths than full-size beams. This proposition is true, but it's not always the limiting factor we may think it is.
2. Small beams present complex construction problems. The truth of this proposition depends upon the sort of antenna construction with which you are comfortable. If you have only slipped tubing in and out of other tubing in normal Yagi construction fashion, then some closed geometry beams will seem ungainly at first sight. However, if you look at plumbing and electrical supplies and see antenna possibilities, then small beams offer a pleasant chal-

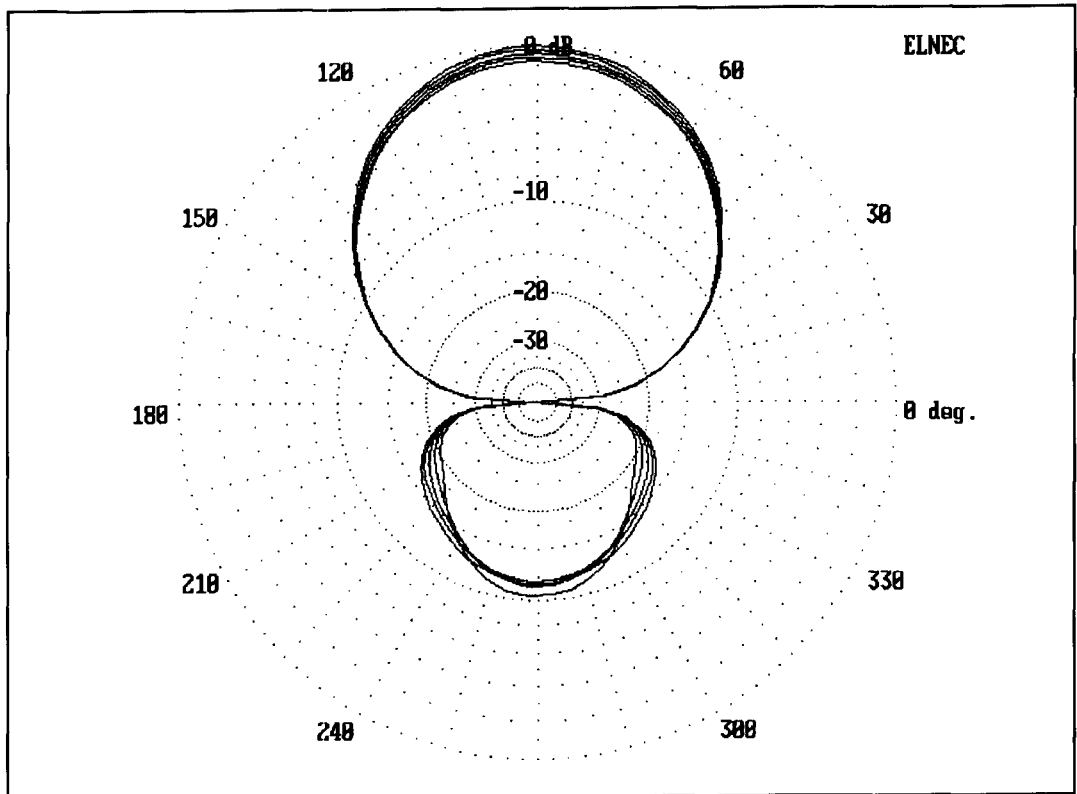


Figure 1. Composite free space azimuth patterns of a wide-band 2-element Yagi from 28 to 29 MHz.

lence. But, admittedly, a few require careful construction.

3. Small beams give reduced performance. With respect to forward gain, this proposition is generally true. However, some of the beams in this series—all two-element parasitic types—

will outperform the standard two-element Yagi in front-to-back ratio with little loss of forward gain over a small bandwidth. And, for certain operators, this property may be just the one desired. Every beam must be measured against a particular set of applications.

Wide-Band 2-Element Yagi Performance					
Frequency (MHz)	28.00	28.25	28.50	28.75	29.00
Free Space					
Gain (dBi)	6.7	6.5	6.3	6.1	5.9
F-B ratio (dB)	10.3	11.0	11.2	11.0	10.6
Impedance (R±X)	24 - 17	28 - 8	32 + 1	36 + 9	40 + 18
20' high					
Gain (dBi)	11.7	11.5	11.3	11.2	11.1
F-B ratio (dB)	13.3	12.2	10.9	9.7	8.8
Impedance (R±X)	27 - 21	30 - 13	34 - 5	37 + 2	39 + 10
35' high					
Gain (dBi)	12.0	11.9	11.7	11.5	11.4
F-B ratio (dB)	12.7	13.7	13.5	12.6	11.5
Impedance (R±X)	27 - 17	31 - 8	35 ± 0	40 + 8	43 + 15

Note: Performance figures for 20 and 35 foot heights were taken over average ground (also called medium earth: dielectric constant = 13, conductivity = 5 milliSiemens/meter). See text for antenna dimensions and material.

Table 1. Performance characteristics in free space, at 20 feet, and at 35 feet of a wideband 2-element Yagi using a driven element and a reflector.

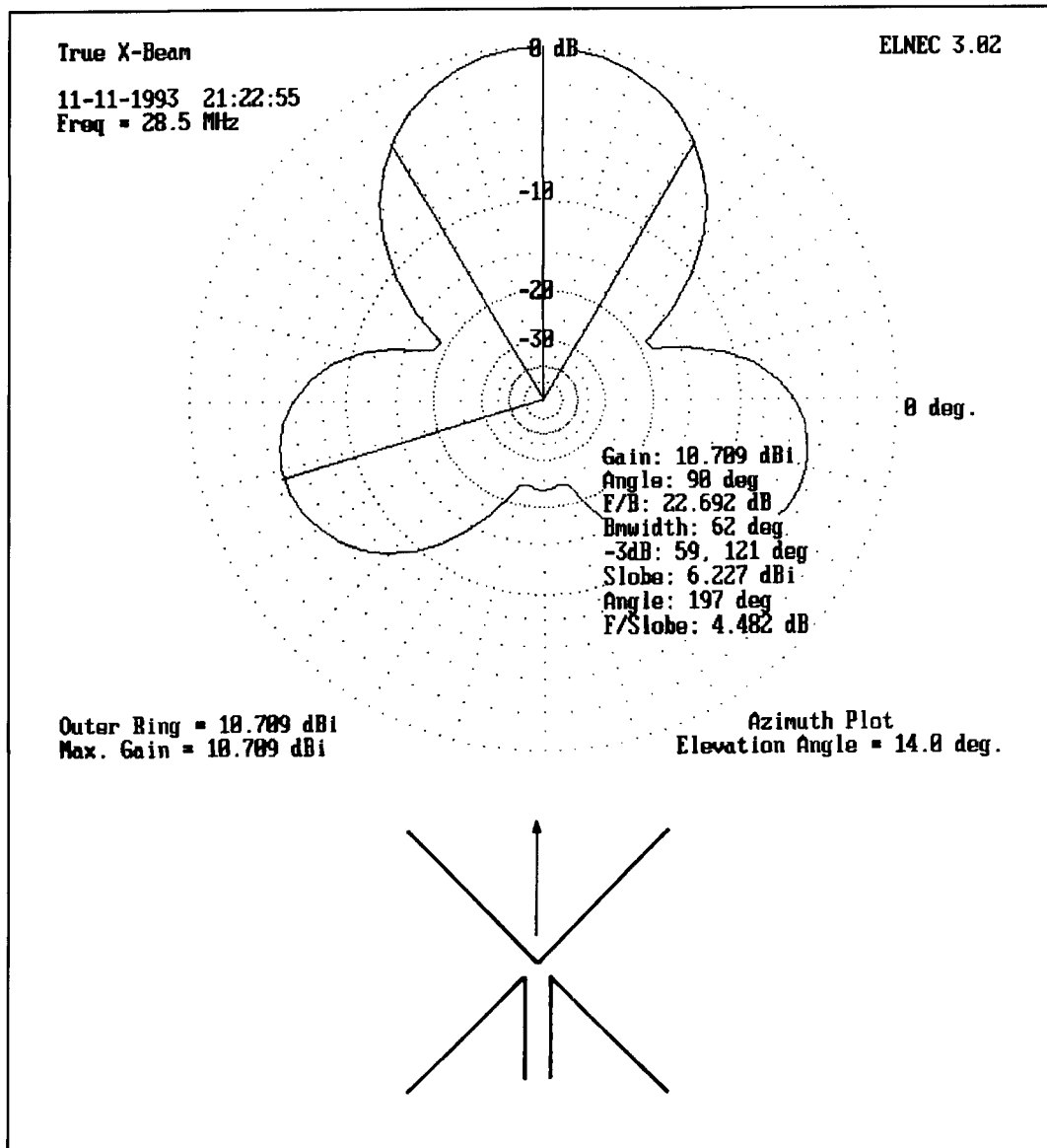


Figure 2. TRU-X beam configuration and azimuth pattern at 35 feet above average earth.

As a standard of comparison, let's present a well-known 2-element Yagi with a driven element and a reflector. Using 1-inch aluminum tubing, the driven element is 16-feet long, the reflector is 17.5 feet long, and the spacing is 4.25 feet.* The resonant feedpoint impedance is about 32 ohms at 28.5 MHz. What makes this antenna a good standard of comparison for understanding the mini-beams in this series is that it's well-behaved. As **Figure 1** and **Table 1** demonstrate, the beam maintains its characteristics across a 1 MHz span of 10 meters with excellent consistency. Free space forward gain varies by about 0.76 dBi, while the front-to-back ratio varies by less than 1 dB. Variations over real ground at 20 and 35 feet are somewhat greater, but still quite minor.

The antenna's performance in terms of raw gain and front-to-back ratio at design center is

modest. However, 6.3 dBi forward gain (free space) and 11.2 dB front-to-back ratio are quite serviceable figures compared to the free space gain of a wire dipole (about 2.15 dBi). You can design Yagis for another dB of forward gain or, alternatively, for a somewhat greater front-to-back ratio—but only at the expense of bandwidth. Thus, the given design seems well suited to its role as a standard of comparison. With it in mind, we can turn to the first of our small beams.

The folded X-beam

In Volume 1 of *The Antenna Compendium*, Brice Anderson presented dimensions for a

*This beam is quite obviously a minor variation of the Yagi presented by Bill Orr, W6SA1, in *CQ*, December 1990, pages 83-84. For modeling purposes, the driven element has been lengthened to resonance.

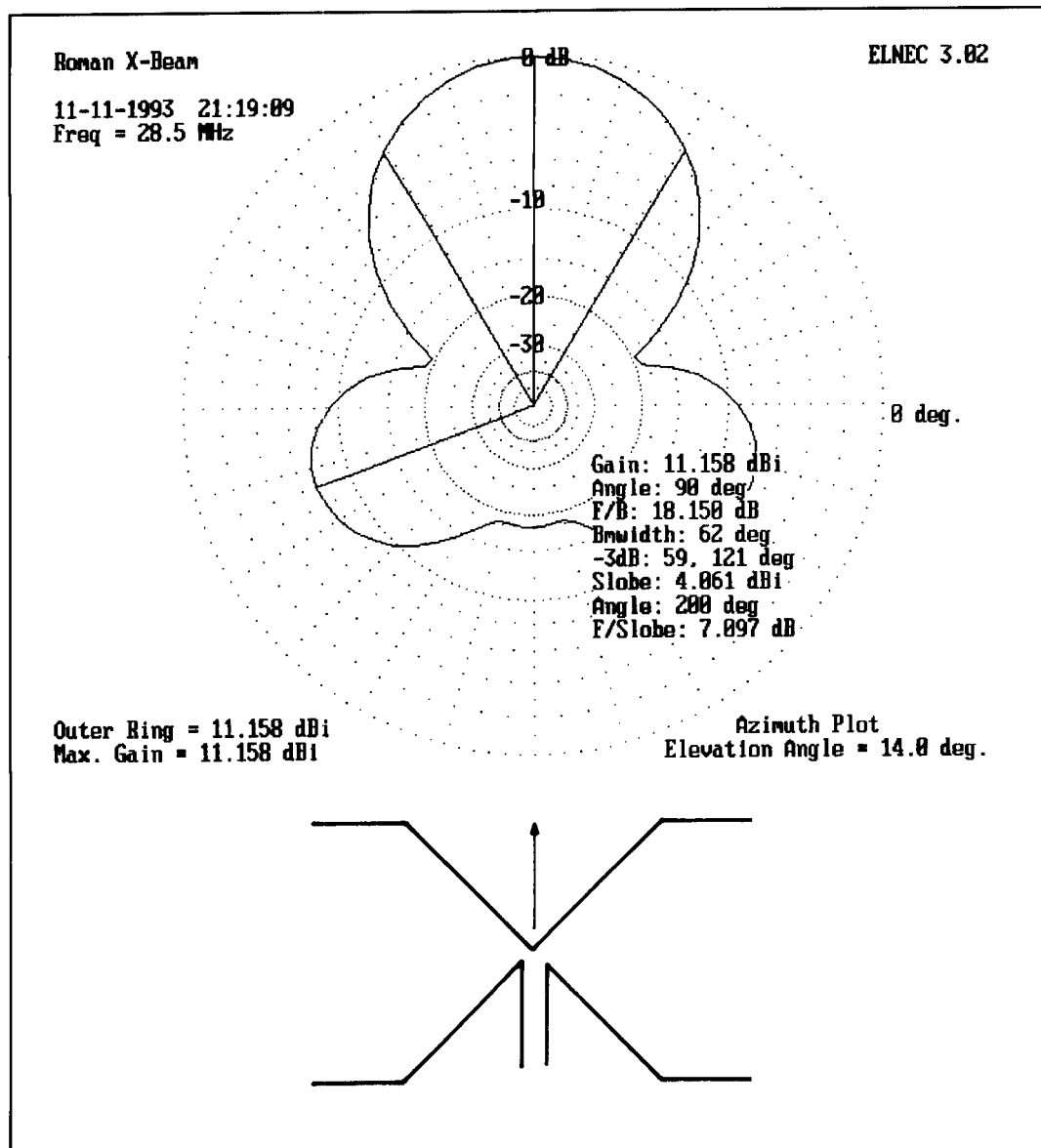


Figure 3. Roman-X beam configuration and azimuth pattern at 35 feet above average earth.

folded X-beam for the upper HF bands.¹ Anderson built his X-beams using a driven element and a director element to match as closely as possible a 50-ohm line without additional components. His aim was a design that home builders could replicate and use reliably across a chosen band. In achieving this goal, he produced a 2-element parasitical beam with less than its full potential for gain and front-to-back ratio. Nevertheless, the X-beam as Anderson presented it may well serve hams operating in very limited space. With dimensions approximately a quarter wavelength on a side, the X-beam is a compact way to achieve some gain

**The modeling data and pattern diagrams used in this study were developed via ELNEC 3.02, an enhanced version of MININEC available from Roy Lewallen, W7EL, P.O. Box 6658, Beaverton, Oregon 97007. Other versions of MININEC are available.

and some front-to-back ratio in a compact arrangement.

Modeling the X-beam

Computer modeling via MININEC offers us a chance to better understand some of the characteristics of the X-beam. However, due to its complex geometry, full of acute angles, it's not the simplest beam to model.**

Where does the X-beam come from? Any two wires, one of which is fed, will show some degree of parasitic element properties if they are not too dissimilar in length, not too far apart, and not at right angles to each other. Paralleling elements is efficient, but not necessary. One can construct a true X-beam (or Tru-

Performance and Dimensions of the Tru-X and Roman-X Antennas

Performance (in free space):				
Antenna	Gain (dBi)	Front-to-Back (dB)	Band width (degrees)	Source Impedance R ± X
Tru-X				
Max. Gain	5.5	16.4	58	42.8 + 3.1
Max. F-B	5.2	29.5	62	51.8 + 2.5
Roman-X				
Max. Gain	6.3	12.1	58	41.2 + 27.3
Max. F-B	5.7	23.1	62	59.8 + 25.3
Dimensions:				
Antenna	DE Length	Dir Length	Total Side-side Length	Total Front-back Length
Tru-X	20.6' DE = 2 x Arm	20.4' Dir = 2 x Arm	14.5'	14.5'
Roman-X	21.4' DE = 2 x (Arm + Tail)	20.4' Dir = 2 x (Arm + Tail)	17.4'	9.8'
Typical 2-element Yagis				
DE + Dir	17.0'	15.6'	17.0'	4.0'
DE + Ref	16.0'	17.5'	17.5'	4.3'
X-Beam	6.9' + 3.8' DE = 2 x (Arm + Tail)	6.9' + 3.3' Dir = 2 x (Arm + Tail)	9.8'	9.8'

Table 2. Dimensions of Tru-X, Roman-X, DE + Dir Yagi, DE + Refl Yagi, and X-beam antennas.

X), as shown in **Figure 2**, and alter the dipole pattern to show a bit of gain in a favored direction. With over 3 dB gain above a dipole in the same setting (35 feet in **Figure 2**), the Tru-X might be a useful beam were it not so large and were the pattern something less than a 3-leaf clover. The side lobes are only 4 dB down from the main lobe, inviting QRM.[†]

Now, let's bend back the ends of the X-elements to form a Roman X, as shown in **Figure 3**. This simple act pushes the side lobes toward the back, reduces them, and increases forward gain. A comparison of the patterns in **Figure 2** and **Figure 3** reveals the differences. Forward gain can improve by up to 1 dB over the Tru-X, while the side lobes are down 7 dB from the main lobe. However, as **Table 2** shows, the Roman-X is a big antenna relative to these marginal improvements. Both the Tru-X and the Roman-X beams are, in general, about the same breadth as, but unequal in performance to, a standard 2-element Yagi. **Table 2** compares the dimensions for the X configurations and for the Yagis.

If we bend the ends of the Roman-X wires still further back until the director ends and driven element ends point at each other, we can retain the performance of the Roman-X in a square only slightly greater than a quarter wavelength on a side, as shown in **Figure 4**. This is the X-beam Anderson presented in Volume 1 of the *Antenna Compendium*, and **Figure 4** shows the dimensions he recommended for 10 meters.*

Basic Anderson performance: To evaluate

[†]Incidentally, it is this configuration—and not the Anderson configuration—against which Moxon recommends in his study of small antennas. Indeed, Moxon might appreciate the folded model of the X, given the many folds in some of his recommended antennas, such as the VK2ABQ square, an interesting antenna in its own right. However, he might object to the quartering rear lobes of the X-beam. See L.A. Moxon, *HF Antennas for All Locations*, R.S.G.B., 1982, pages 81–83 and 172–175.

*For those wishing to scale the X-beam to other bands, Anderson provides the following formulas. Arm length (ft) = 195/f (MHz); Total driven element length (ft) = 603/f (MHz); Driven element tail (ft) = 106.5/f (MHz); Total director element (ft) 575/f (MHz); and Director tail (ft) = 92.5/f (MHz). (Anderson, page 64) Although the choice of materials—tubing diameter for the arms and wire size for the tails—will alter the required dimensions, no. 18 wire and tubing with a length-to-diameter of about 200:1 will make these dimensions a good starting point for modeling.

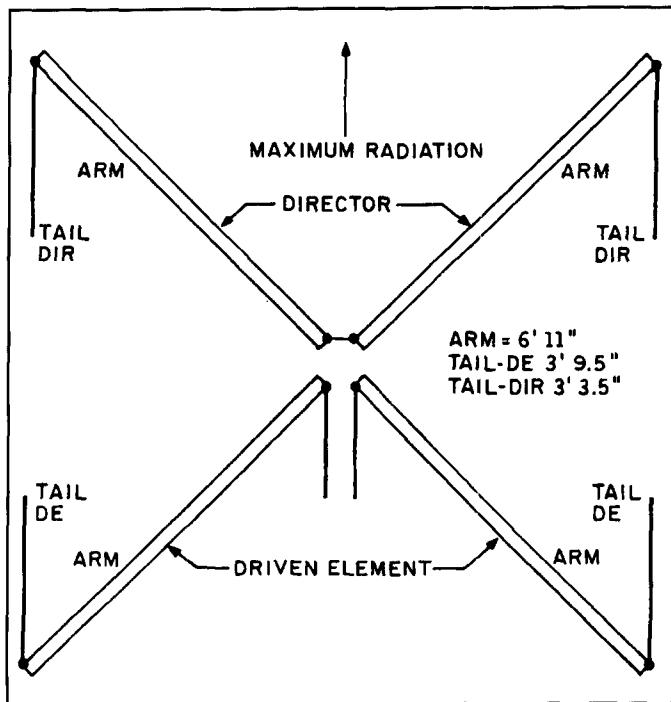


Figure 4. Configuration and terminology of the X-beam.

folded the X-beam, I constructed numerous computer models of Anderson's design and variations upon it. Throughout, I retained Anderson's terminology, calling the angular tubing the "arms" and the bent-back wire ends the "tails."

The initial stage in the work was to see what MININEC might tell about Anderson's beam. I modeled his dimensions for 10 meters (28.5 MHz), using 0.5" diameter aluminum tubing and no. 18 wire. The result was a beam with modest gain over a real dipole in free space, at 20 feet (a common height for portable beams on 10), and at 35 feet (a common 10-meter fixed installation height). The results appear in **Table 3**, with **Figure 5** showing a representative azimuth pattern for the "best match" configuration at 35 feet above average earth (dielectric constant 13, conductivity 5 mS/m). Three aspects of the figure are especially notable. First, the gain over a dipole in the same position is less than 3.5 dB and not up to average 2-element Yagi figures. The front-to-back ratio is also modest, but odd. The side lobes quartering to the rear are less than 10 dB down from the main lobe, which limits signal rejection in these directions. These rearward lobes will remain, no matter how much front-to-back ratio is achieved in the 180-degree direction. Third, the input impedance is in the mid-30-ohm range, which permits a direct coax connection with an SWR ratio of under 2:1. One can either live with or cancel the remnant reactance at the feedpoint without too much trouble.

Attempts to improve performance led to some alteration of the element dimensions, while adhering to Anderson's advice to keep the tail lengths a certain amount different—on 10 meters, 6 inches. This procedure lent consistency to all my models. My first attempts lengthened all the elements by 0.2 feet per end. As **Table 3** shows, the resulting antenna increased in gain by a small but useful amount. However, the front-to-back ratio increased by much more. Nonetheless, the rearward side lobes did not shrink materially. **Figure 6**, a representative azimuth pattern, displays the side lobes quite well. Finally, the input impedance decreased into the mid-20-ohm range. This antenna promised improved performance at the cost of requiring a 2:1 matching system.

Increasing the tail lengths by 0.3 feet over the original Anderson dimensions produced further changes in X-beam characteristics. Gain increased to equal or surpass average 2-element Yagis. However, as **Table 3** shows, the front-to-back ratio decreased dramatically. Yagis with 2 elements show a similar phenomenon when optimized for maximum gain. **Figure 7** shows a representative "max gain" X-beam azimuth pattern. Finally, the source impedance dropped into the mid-teens range.

None of these models represents the absolute peak of coax match, front-to-back ratio, or gain that might be obtained with an X-beam. Rather, they are models designed to find out about where various parameters end up with changes of element length. Initial observations of the data in **Table 3** suggested that with the X-beam, one cannot wholly rely on a free-space model. Modeling over the anticipated real terrain of the antenna is a must. Because MININEC calculates the source impedance over perfect ground, not the modeled ground, slight differences between the models and reality are the order of the day.

Constructing an X-beam without loading requires a tuning process that opposes what we normally do. If we wish to move from best 50-ohm match to maximum front-to-back ratio, we must lengthen the element tails, a process that requires more forethought than merely snipping dipole wires.

Loading the X-beam: If we could achieve some desirable improvements in X-beam performance by changing element lengths, the next question is whether similar results might be obtained by loading the director. This idea came from W2EEY's X-beam that I encountered in *Antenna Roundup*, Volume 2.² **Table 4** shows some initial results. Using the original Anderson dimensions, we can achieve the performance of the maximum front-to-back ratio model and the maximum gain model by placing a series inductor of the proper reactance at the

Properties of a 10-Meter X-beam					
Antenna A: Basic Anderson design: Arms 0.5" dia., 6.92' long, Tails #18, DE = 3.79', Dir = 3.29'.					
Environment	Gain (dBi)	Take-Off Angle	F-B Ratio (dB)	Beam width (degrees)	Source Impedance R ± X
Free Space	5.5	--	20.9	66	36.3 - 19.2
20', ave. gnd.	10.7	24	11.8	74	38.6 - 12.1
35', ave. gnd.	10.9	14	17.1	68	34.3 - 16.1
Antenna B: +0.2' to Tails: Arms 0.5" dia., 6.92' long, Tails #18, DE = 3.99', Dir = 3.49'.					
Free Space	6.6	--	13.0	62	20.3 + 0.4
20', ave. gnd.	11.6	23	24.2	66	22.9 + 1.9*
35', ave. gnd.	11.9	14	15.4	62	20.3 + 1.9
Antenna C: +0.3' to Tails: Arms 0.5" dia., 6.92' long, Tails #18, DE = 4.09', Dir = 3.59'.					
Free Space	6.9	--	7.0	58	14.1 + 15.1
20', ave. gnd.	11.9	23	10.6	62	15.3 + 15.1
35', ave. gnd.	12.1	14	7.9	60	14.3 + 15.6
* Dimensions of antenna B were optimized for a 20' height above average earth. Free space models require adjustment for low heights and require real earth models.					
Performance figures for 20 and 35 foot heights were taken over average ground (also called medium earth: dielectric constant = 13, conductivity = 5 milliSiemens/meter).					

Table 3. Properties of a 10-meter X-beam in free space and at 20 and 35 feet over medium earth.

center point of the director. The models use a lossless inductor as the loading element.

Further azimuth patterns proved unnecessary, as progressions began to emerge. Best match, maximum front-to-back, and maximum gain patterns each look alike, whatever means are used to arrive at them. Moreover, the resistive component of the source impedance remains in the same ball park for each category: the mid-30s for best match, the mid-20s for maximum front-to-back ratio, and the mid-teens for maximum gain.

Inductive loading is usually less preferable than capacitive loading, wherever a choice is available. Therefore, following W2EEY's lead, the next set of models lengthened the director tails to require a capacitive load for each X-beam category. By increasing the capacitive loading, one can move from best match through maximum front-to-back ratio to the maximum gain level. At the same time, one passes smoothly through the three source impedance ranges. The result is an X-beam that permits

adjustment with the turn of a knob.

Anderson recommended a constant length difference between driven element and director tails. However, since a variable capacitor is capable of tuning the system to the desired characteristics and since the driven element reactance remains stable across the span of adjustments, we may depart from Anderson's empirically derived constraint and resonate the capacitively loaded X-beam. **Table 5** shows in finer detail the result of returning the driven element to an approximately resonant length while leaving the director long for capacitive tuning. The results are precisely comparable to those for all other models.

In fact, the representative azimuth patterns shown in earlier figures are taken from the 35-foot model of this X-beam. Moreover, rearward side-lobe angles shown in **Table 5** apply to all the models within a few degrees and with close matching of their level below the main lobe. Additionally, the X-beam does not lend itself to close stacking or interlacing. W9PNE confirmed

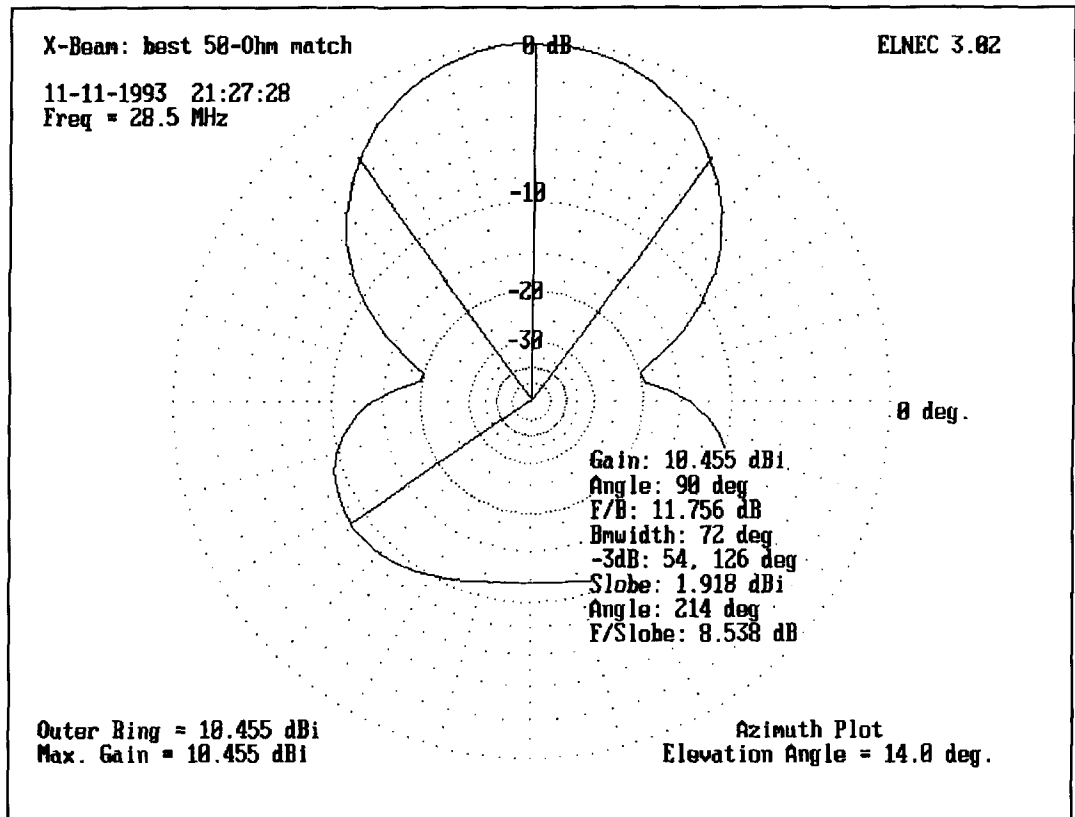


Figure 5. Representative azimuth pattern of the X-beam, configured for best match to 50-ohm coaxial cable.

my own modeling failures in correspondence: no one he knew had had any luck trying to create a compact multi-band X-beam assembly.

Changing materials and frequencies:

Because not everyone has access to the same materials, I checked the sensitivity of the X-beam in **Table 6** to variations in arm diameter and tail thickness. Increasing the arm diameter to 1 inch and the tail wires to no. 12, while preserving the dimensions of the antenna in **Table 5**, produced the results shown in **Table 6**. Those who are only used to working with linear elements, where fat equals electrically long, may be surprised to find that both the driven element and the director play short. Both require lengthening to match the antenna in **Table 5**. (Those who have modeled quads with wires and tubing of varying diameters won't find this phenomenon strange at all.) The requisite lengthening—on the order of 0.2 to 0.3 feet per tail—approaches the point of requiring slightly longer arms to keep the tails from almost touching. In the end, varying the materials of an X-beam by any significant amount requires either a new model for a blueprint or some empirical experimentation.

SWR bandwidth: Developing 2:1 SWR bandwidth curves does help us determine which of the design variations will best serve a specific application. The bandwidth of the X-

beam appears to be little better or worse than most antennas. **Table 7** charts two models of the X-beam at 35 feet, the first of which is the original Anderson specifications. Referenced either to the center frequency resistive component of the source impedance or to a 50-ohm cable, the Anderson design provides a broad curve within the usual 2:1 SWR limits. The curves are graphed in **Figure 8**. Note that as the frequency increases, the design transitions from a "best match" condition to a "max gain" condition, with the "max front-to-back" condition occurring about three-fourths of the way through the 1 MHz spread. Although the gain and front-to-back ratio drop off seriously at the lowest end of the band, performance is reasonably adequate throughout most of the range.

Using the capacitively loaded antenna at 35 feet of **Table 5**, I converted the load reactance at 28.5 MHz into a capacitance for insertion into the model as a LaPlace figure. The results of a selective frequency sweep appear in **Table 7** and are graphed in **Figure 9**. Again, the lower line references the SWR to the center frequency, while the upper line references it to 50 ohms. The figures suggest that the 2:1 SWR bandwidth exceeds 750 kHz on 10 meters. However, most of the bandsread lies below the design center. Therefore, when planning an antenna for a specific bandsread, the designer

Properties of a 10-Meter X-beam with an Inductively Loaded Director

Antenna: Arms 0.5" dia., 6.92' long, Tails #18, DE = 3.79', Dir = 3.29'.

Environment	Dir XI (Ohms)	Gain (dBi)	Take-Off Angle	F-B Ratio (dB)	Beam width (degrees)	Source Impedance R ± X
Antenna A: Anderson best 50-Ohm match design						
Free Space	0	5.5	--	20.9	66	36.3 - 19.2
20', ave. gnd.	0	10.7	24	11.8	74	38.6 - 12.1
35', ave. gnd.	0	10.9	14	17.1	68	34.3 - 16.1
Antenna B: Maximum front-to-back ratio design						
Free Space	+ 6	5.9	--	27.6	64	29.7 - 19.6
20', ave. gnd.	+12	11.4	24	24.0	68	26.1 - 15.7
35', ave. gnd.	+10	11.5	14	27.2	64	24.9 - 16.6
Antenna C: Maximum gain design						
Free Space	+23	6.8	--	6.9	58	13.1 - 9.9
20', ave. gnd.	+23	11.7	24	10.3	62	14.3 - 10.2
35', ave. gnd.	+20	12.1	14	10.3	60	15.5 - 11.8

Note: Loading permits each antenna to be optimized for each environment.

Performance figures for 20 and 35 foot heights were taken over average ground (also called medium earth: dielectric constant = 13, conductivity = 5 milliSiemens/meter).

Table 4. Properties of a 10-meter X-beam with an inductively loaded director in free space and at 20 and 35 feet over medium earth.

should center the model about two-thirds to three-fourths of the way up the spectrum. Like the Anderson design, this X-beam changes its characteristics from a "best match design" through a "max F-B" design to a "max gain" design. If a particular characteristic is especially needed at a specific frequency, one must design for that frequency and live with the results at other frequencies.

Building an X-beam

Because a folded X-beam for 10 meters is under 10 feet on a side, garage construction is easy. A 10-pound package is easy to achieve, which means that a simple mast will hold the antenna, and antenna raising is a simple matter. At the risk of repeating some of what Brice Anderson has said about building an X-beam, let's look at some possibilities.

Basic structure: For a number of reasons relating to the bandwidth of the antenna, "permanent" X-beam installations should consider

using Anderson's original recommendations for dimensions. Although the rear lobe of the antenna will be wide and gain will not be maximum, matching the antenna to a 50-ohm cable and transmitter is vastly simplified. In fact, a simple sleeve balun (such as the W2DU model or those available through Radio Works) may be the only device between the coaxial feedline and the antenna terminals or connector.

Refer to **Figure 10** for some construction details. Although this figure shows elements relevant only to a hilltopper version of the X-beam, it also reveals the basic construction technique for all X-beams. For 10 meters, a plywood plate 18 inches on a side supports the X members. Scrap plywood 3/8 to 1/2 inch thick will easily bear the antenna load. My technique for attaching the antenna to the mast is to fasten an 18-inch length of 1-inch nominal diameter Schedule 40 PVC with two 2-inch L-bracket supports. Stainless steel bolts (no. 10 x 2 inches) run through the pipe to clamp the L-brackets, while 1-inch no. 10 bolts hold the plate.

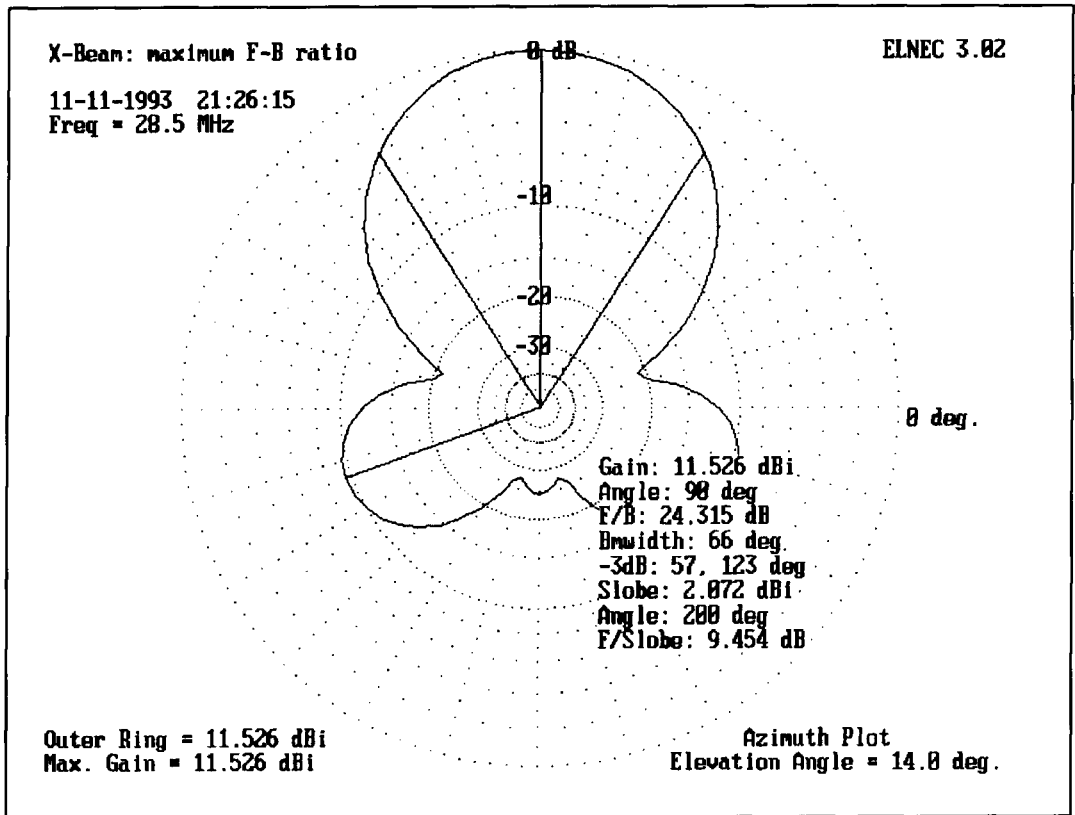


Figure 6. Representative azimuth pattern of the X-beam, configured for maximum front-to-back ratio.

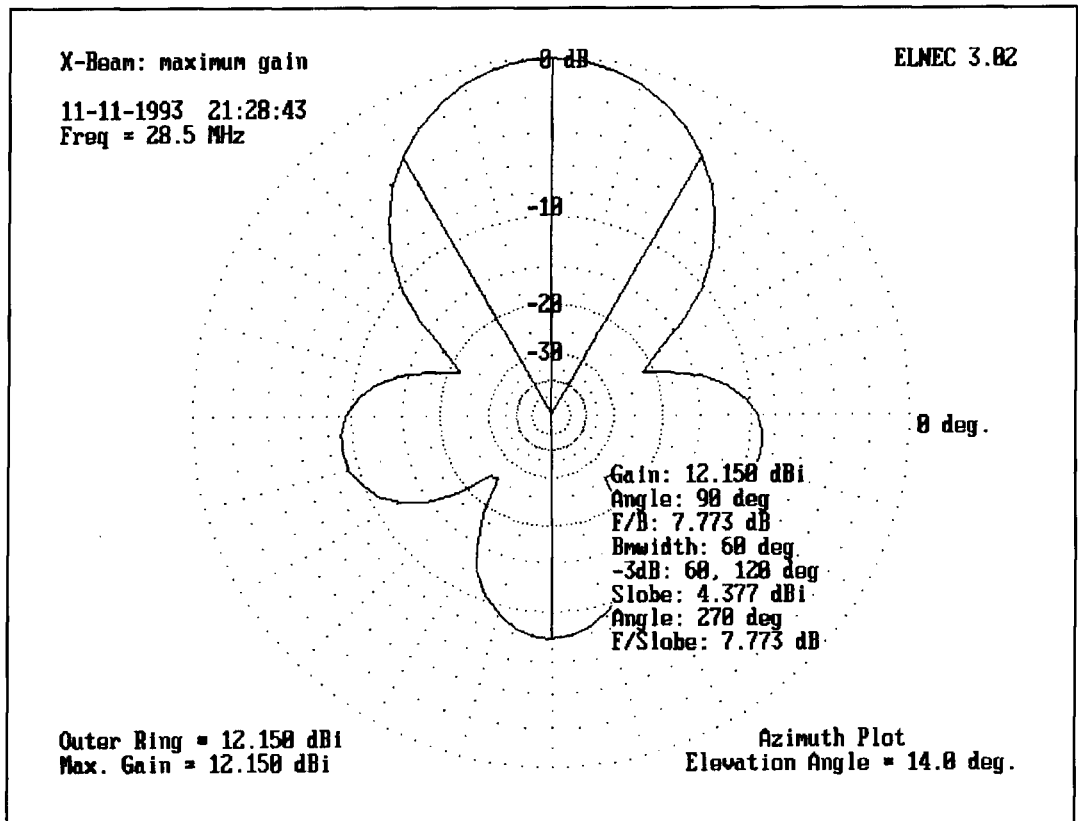


Figure 7. Representative azimuth pattern of the X-beam, configured for maximum gain.

Properties of a Resonant 10-Meter X-beam with a Capacitively Loaded Director

Antenna: Arms 0.5" dia., 6.92' long, Tails #18, DE = 3.99', Dir = 3.79'.

Dir Load -Xc	Gain (dBi)	F-B Ratio (dB)	Band Width (degrees)	Source Impedance R ± X	Side Lobe Angle (deg)	Side Lobe Ratio (dB)	
Environment 1: Free Space							
-15	6.9	6.9	58	13.7 + 8.1	100 (est)	-9.0 (est)	n. 1
-20	6.8	10.1	60	17.0 + 4.4	101	-9.1	
-25	6.5	14.4	62	21.1 + 1.5	105	-9.4	
-30	6.2	21.1	64	26.2 - 0.3	108	-9.4	
-35	5.9	51.5	66	31.4 - 0.9	112	-9.4	n. 2
-40	5.6	22.2	68	35.6 - 0.4	115	-9.2	
-45	5.3	16.8	68	39.7 + 0.8	118	-9.0	
-50	5.0	13.9	70	43.1 + 2.3	121	-8.7	n. 3
Environment 2: 20' above average earth, takeoff angle 23 degrees							
-15	11.8	10.4	62	14.9 + 8.2	101	-9.2	n. 1
-20	11.8	16.3	66	19.1 + 5.1	105	-9.4	
-25	11.5	33.0	68	23.9 + 3.5	109	-9.4	n. 2
-30	11.3	21.2	70	28.9 + 3.4	114	-9.2	
-35	11.0	15.2	72	33.3 + 4.5	118	-8.8	
-40	10.8	12.2	74	37.0 + 6.5	123	-8.4	
-45	10.5	10.2	74	39.7 + 8.9	127	-8.0	n.3
Environment 3: 35' above average earth, takeoff angle 14 degrees							
-15	12.2	7.8	60	13.9 + 8.7	100	-9.0	n. 1
-20	12.1	11.6	62	17.2 + 5.4	102	-9.2	
-25	11.8	16.8	64	21.3 + 3.1	106	-9.5	
-30	11.5	24.3	66	25.5 + 1.8	110	-9.5	n. 2
-35	11.2	22.1	68	29.7 + 1.6	113	-9.3	
-40	10.9	16.9	68	33.4 + 2.2	117	-9.1	
-45	10.6	13.8	70	36.6 + 3.5	120	-8.8	
-50	10.5	11.8	72	39.2 + 5.0	124	-8.5	n. 3

Note 1: maximum gain

Note 2: maximum front-to-back ratio

Note 3: best coax match

Performance figures for 20 and 35 foot heights were taken over average ground (also called medium earth: dielectric constant = 13, conductivity = 5 milliSiemens/meter).

Table 5. Properties of a resonant 10-meter X-beam with a capacitively loaded director in free space and at 20 and 35 feet over medium earth.

Drill the plywood for the X members, using 2-inch no. 10 bolts for the fasteners closest to the mast stub and 1-1/2 inch no. 10 bolts about a half inch from the plate corners. (All hardware is stainless steel, now more readily available in hardware stores.) I used a drill press to drill the tubing, keeping the two holes per piece aligned by dropping a long nail through the first hole and matching it to a reference vertical. The only other holes needed are at the X-member far ends, about 1/4 inch in for attaching the perimeter line and the tails.

Disassemble the basic X and weather-treat the plywood. Spar varnish has been a standard, but scrap plywood often admits moisture through gaps in the edge. A more weather-proof alternative is to fiberglass epoxy the board. Before coating the board, enlarge all

holes slightly. Then coat their edges heavily. When everything is set, use the correct drill bit to reestablish bolt holes and some sandpaper to refit the mast stub. An orbital sander with coarse paper will make quick work of smoothing and leveling the surface. The result is weather-proof nearly forever.

When the hub and X members are reassembled, run a perimeter line of 1/8-inch nylon rope or cord, using the end bolts as attachment points. Solder ring connectors on the tails and place them on the end bolts with an additional nut. Add very short leads (with ring connectors) to a coax connector and attach the ring connectors to the underside bolt extensions for the driven element, using additional no. 10 nuts. Add a shorting bar across the bolt extensions of the director.

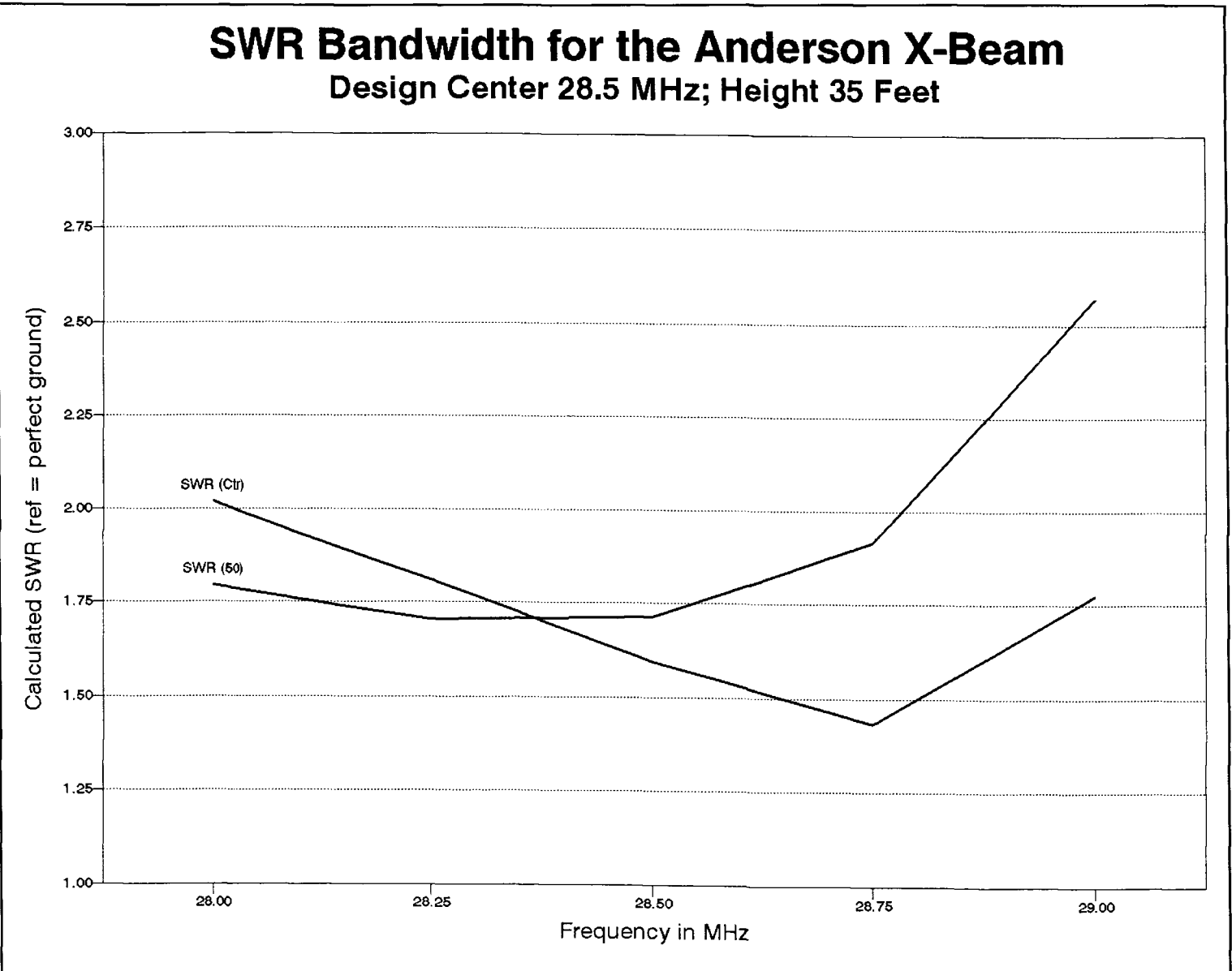


Figure 8. 2:1 SWR bandwidth graph for the Anderson 10-meter X-beam in Table 2.

My model bolts the mast stub inside a 3-foot length of 1-1/4 inch PVC, which then slips over the main mast. A pair of heavy sheet metal screws locks the larger diameter stub to the mast. Alternatively, the larger diameter stub

can be cut to match the length of the inner stub and the pair locked in a rotator. Although the result is light, it is mechanically sound. My test beams have survived winds in excess of 45 miles per hour without damage.

SWR Bandwidth: Director-Loaded X-Beam

Design Center 28.5 MHz; Height 35 Feet

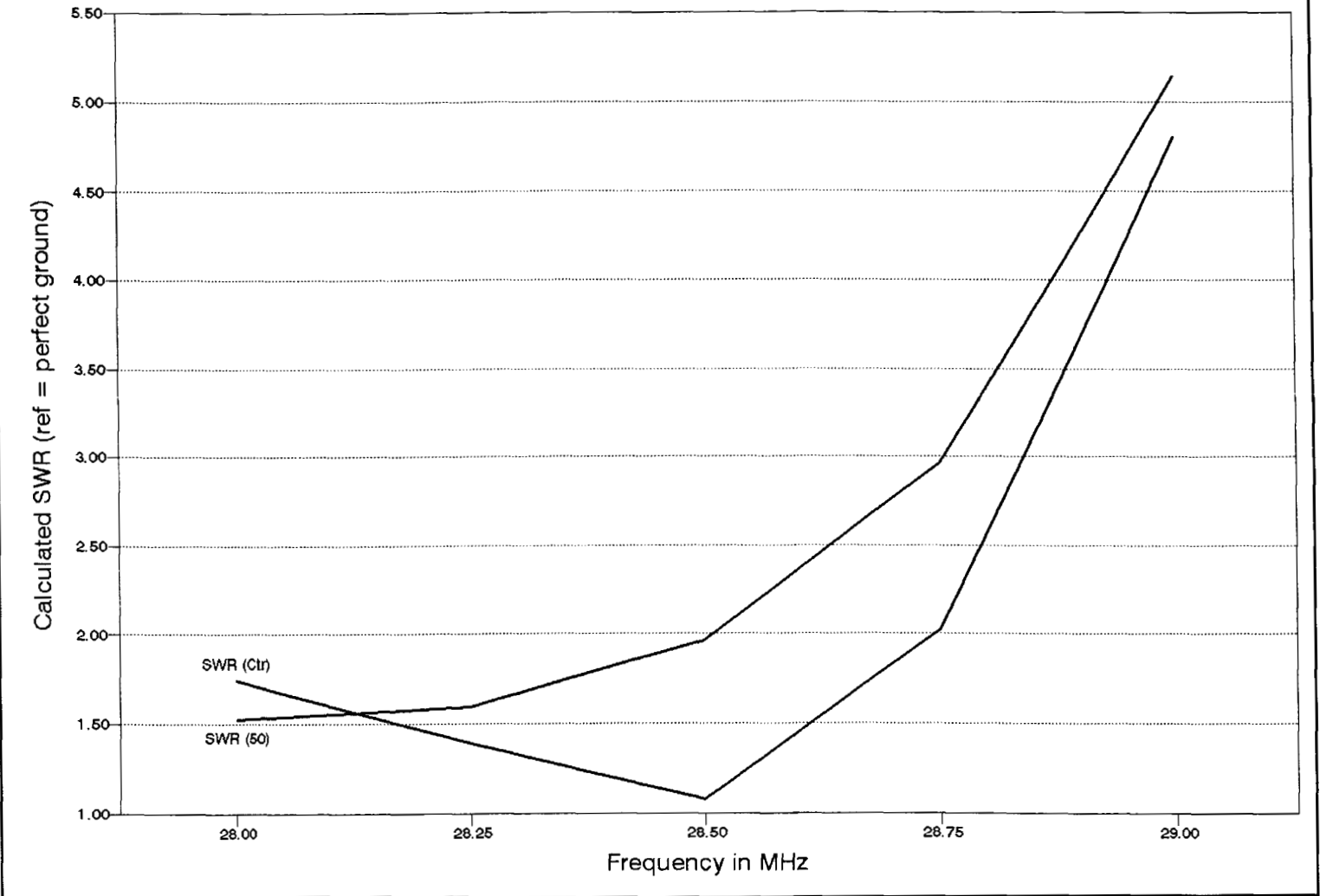


Figure 9. 2:1 SWR bandwidth graph for the director-loaded 10-meter X-beam in Table 5.

The perimeter line dampens vibrations and reduces metal fatigue. Rain cannot reach the coax connections on the underside of the plate. The end result is simple and long-lived.

Adjustment: Because the lowest obtainable

SWR doesn't coincide with the antenna's best performance, careful adjustment is required. Adjusting the tail lengths of the director will affect the resistive component of the driven element impedance, while changing the tail length

Properties of a 10-Meter X-beam with Thicker Arms and Tails and with a Capacitively Loaded Director

Antenna: Arms 1.0" dia., 6.92' long, Tails #12, DE = 4.29', Dir = 3.79'.

Environment	Dir XI (Ohms)	Gain (dBi)	Take-Off Angle	F-B Ratio (dB)	Beam width (degrees)	Source Impedance R ± X
Antenna A: Best 50-Ohm match design						
Free Space	-25	5.3	--	15.7	70	39.9 - 2.9
20', ave. gnd.	-25	10.5	24	9.6	76	39.5 + 5.3
35', ave. gnd.	-25	10.7	14	13.0	70	36.7 - 0.4
Antenna B: Maximum front-to-back ratio design						
Free Space	-15	5.9	--	33.0	66	31.4 - 4.8
20', ave. gnd.	- 5	11.6	24	53.0	68	24.0 - 0.4
35', ave. gnd.	-10	11.5	14	24.2	66	26.5 - 2.3
Antenna C: Maximum gain design						
Free Space	+ 5	7.0	--	6.9	58	13.4 - 4.4
20', ave. gnd.	+ 5	11.9	24	10.5	64	14.5 + 4.3
35', ave. gnd.	+ 5	12.3	14	7.7	60	13.6 + 4.8

Performance figures for 20 and 35 foot heights were taken over average ground (also called medium earth: dielectric constant = 13, conductivity = 5 milliSiemens/meter).

Table 6. Properties of a 10-meter X-beam with thicker arms and tails and with a capacitively loaded director in free space and at 20 and 35 feet over medium earth.

of the driven element will alter its reactive component. If the center frequency is to be in the lower portion of the 10-meter phone band, alter the tail lengths so the lowest SWR (whatever its value) is just about at the 28.0 MHz point. Then alter the driven element tail lengths to reduce the SWR at the center frequency further by bringing the driven element to resonance. Since initial adjustments may require lengthening elements, do not install ring connectors until the tail length is set.

A hilltopper version of the X-beam: The ability to capacitively load the director for optimum performance suggested that the folded X-beam would make a compact collapsible antenna for portable hilltop operations. **Figure 11** shows the theoretic coincidence of five patterns across the first MHz of 10 meters with the antenna director adjusted in each case for maximum front-to-back ratio at a 20-foot height level over medium ground, the average conditions for my portable operations.

Because local hardware stores carried only 3/4-inch, 0.050-inch wall thickness aluminum

tubing, I remodeled the antenna for this material. With 6 foot 11 inch arms, the director and driven element tails became functions of the tuning and matching system. Since a beta match requires driven element shortening, its tails worked out to be 3 feet 11 inches long to provide a requisite 25 ohms of capacitive reactance. The director element had to be lengthened so the tuning capacitor could electrically shorten it; 4 foot 1 inch tails provided for this factor. (Note: despite the fact that the physical length of the driven element is shorter than the parasitic element, the electrical lengths are reversed and the beam remains a driven element-director arrangement.)

Beta matches are easily calculated from information in the *ARRL Antenna Book*.³ I set 25 ohms as the desired feedpoint impedance, because this is close to optimum for maximum front-to-back performance from the antenna. A beta match was then calculated for a 2:1 reduction in impedance. Coil construction utilizes data from recent editions of the *ARRL Handbook*.⁴ The coil required a 50-ohm inductive

reactance, which works out to about 0.28 μH , or 6 turns spaced over 3/4 inch with a 1/2 inch diameter, using no. 12 wire.

Antenna construction is identical to that described for the "permanent" model. Only a few changes are needed to accommodate collapsing the antenna and adding the tuning and matching elements. First, the beta-match coil has ring connectors on its very short leads and connects across the driven element inner bolt extensions. Second, a half wavelength (adjusted for velocity factor, about 16 feet) of 450-ohm ladder line attaches to the director bolt extensions in place of the shorting bar. At the other end of line, which is within easy operator reach, are two pin jacks. The tuning capacitor has pin terminals to mate with the jacks. The line is spaced from the mast by plastic spray can tops in which four holes are punched. Cable ties hold the tops to the mast every few feet, and smaller cable ties hold the line to the tops.

Every use of the antenna sacrifices a few cable ties during disassembly. However, the remote tuning system mounts and dismounts

quickly. The capacitor goes into a protective box when not in use, and the line (with caps attached) rolls up semi-neatly.

The bolts marked "P" on **Figure 10** are pivot bolts; they are loosened for antenna transport and storage. The others are removed to free the arms to pivot and then replaced in their plate holes for storage. The perimeter line is looped over the arm point marked "L" and removed when the antenna is not in use. The entire assembly is then pivoted so all arms point the same way. Fold the tail lines neatly along the lengths of the nested arms, and the resulting compact mass fits in my hatchback.

Adjustment: Using the antenna is simply a matter of putting it up and tuning the remote director capacitor for minimum SWR. More accurately, SWR minimum is optimum at the design center frequency only. Below center frequency, use slightly less capacitance than for absolute minimum SWR (within the usual 2:1 transmitter limits) and above center frequency, use slightly more. These guidelines are a function of the complex resistance-reactance combi-

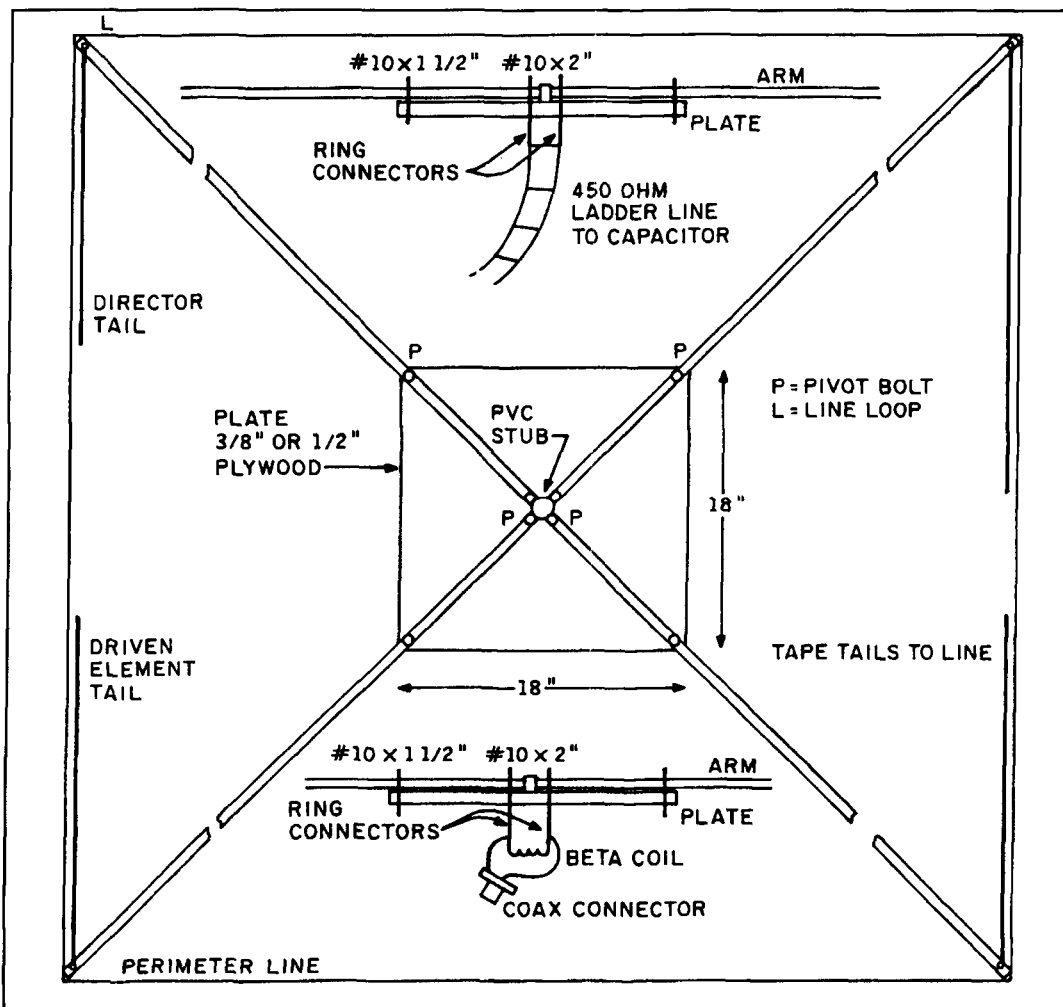


Figure 10. Construction details of a collapsible hilltopper X-beam with beta match and remote director tuning.

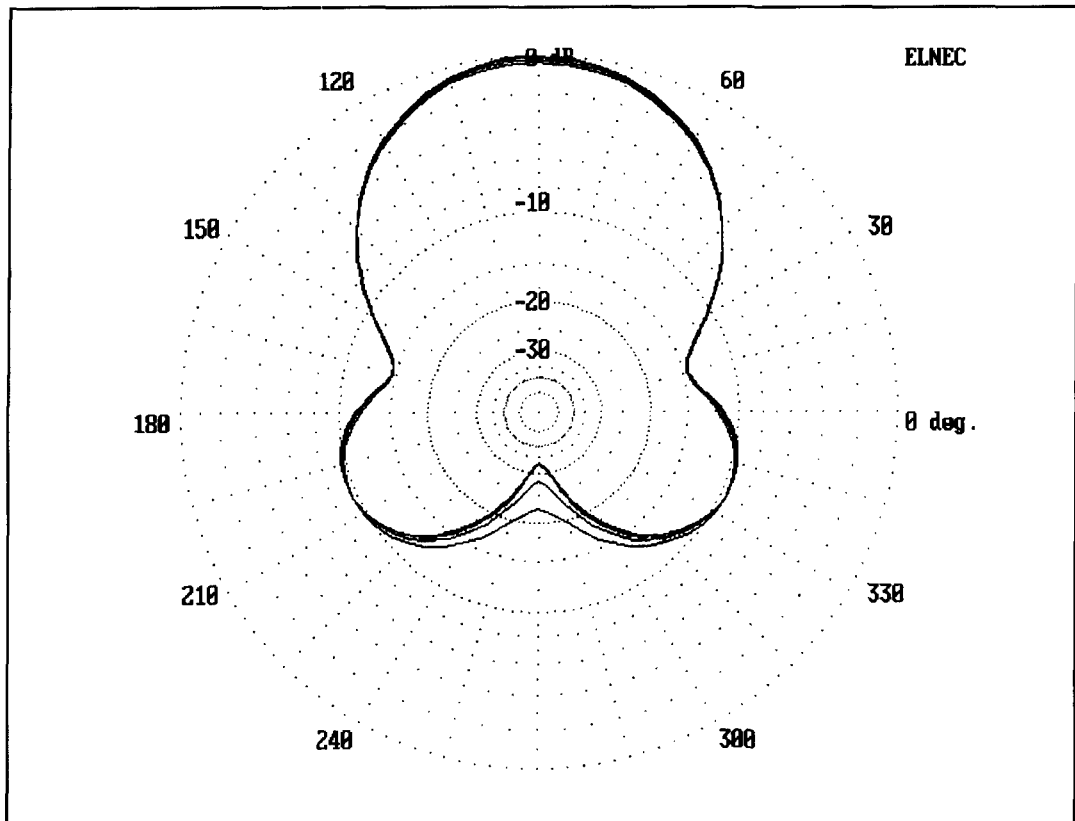


Figure 11. Azimuth patterns of the hilltopper X-beam at 20 feet from 28.0 to 29.0 MHz.

nations that occur off center frequency and the desire to hold the antenna at the maximum front-to-back ratio point.

Initial adjustment is similar to the procedure used for the permanent model. Tune the capacitor for the lowest SWR at the center frequency and then trim the driven element tails for a further drop.

Variations in materials can create significant differences from modeling figures. For example, the hilltopper model used plastic covered speaker wire as a convenience in cutting identical tails. However, this material yielded tails that were electrically long for both elements. I anticipated a need for a capacitor covering a range of 10 to 45 ohms capacitive reactance (about 125 to 570 pF) to tune the range from 28 to 29 MHz. With the wire used, a 100 pF variable capacitor covered the resultant 80 to 220 ohm range (25 to 70 pF).

The tails required 6 inches of pruning each to place the SWR curve well within the desired band. However, the result was a curve that went below 1.15:1 at 28.300 MHz and rose to 1.5:1 at 28 MHz and to 1.9:1 at 29 MHz. Local point-to-point contacts verify the pattern as generally in accordance with the shape shown in Figure 11. Performance is detectably better than with a reference dipole placed on the same

mast, and the reasonably sharp null to the rear is more pronounced than the forward gain.

Overall evaluation

The X-beam is a parasitic 2-element beam, but not a 2-element Yagi in either appearance or performance. However, for those with restricted budgets and space, it may provide significant improvements in performance over a dipole or other simple antenna. The X-beam is flat and compact. As an inexpensive (\$25 to \$35 dollars, depending on local hardware store prices), easily built antenna that fits on a mast (instead of a tower) with a TV rotator (or simply an "Armstrong" rotator), the antenna lends itself to some home applications. Its turn radius is a little less than that of a rotatable dipole. But, for me, it comes into its own as a collapsible, tunable hilltopper.

One factor to consider in making the decision to build an X-beam should be the pattern of usual QRM that besets your station. If that QRM comes directly off the rear of a dipole, then a "max F-B" X-beam set-up may be quite helpful. However, if the QRM angles in off the rear (for example, QRM in St. Louis from New England and Florida when the beam points due west), then it may be wise to hold out for a small Yagi.

X-Beam Bandwidth Characteristics

Antenna: Arms 0.5" dia., 6.92' long, Tails #18, DE = 3.79', Dir = 3.29' @ 35'
Design center: 28.5 MHz (original Anderson design)

Frequency (MHz)	28.00	28.25	28.50	28.75	29.00
Gain (dBi)	9.9	10.3	10.9	11.5	12.0
Front-to-back ratio (dB)	8.9	11.6	17.1	33.8	13.1
Beam width degrees	72	70	68	66	62
Source Impedance R ± X	42.1 - 26.1	39.4 - 21.5	34.3 - 16.1	27.2 - 8.4	19.5 + 3.1
SWR (50) ¹	1.797	1.706	1.710	1.913	2.569
SWR (Ctr) ²	2.020	1.808	1.594	1.433	1.766

Antenna: Arms 0.5" dia., 6.92' long, Tails #18, DE = 3.99', Dir = 3.79' @ 35'
Director load (Xc) = -30; Load = 1.86•10⁻¹⁰F (186 pF) @ design center: 28.5 MHz

Frequency (MHz)	28.00	28.25	28.50	28.75	29.00
Gain (dBi)	10.3	10.9	11.5	12.1	12.1
Front-to-back ratio (dB)	11.6	16.9	24.3	12.0	5.5
Beam width degrees	72	68	66	62	58
Source Impedance R ± X	37.0 - 12.9	32.3 - 6.7	25.5 + 1.8	18.4 + 14.0	13.3 + 29.5
SWR (50) ¹	1.524	1.596	1.961	2.959	5.141
SWR (Ctr) ²	1.744	1.390	1.074	2.030	4.800

¹ SWR (50) gives the calculated SWR relative to a 50-Ohm feedline.

² SWR (Ctr) gives the calculated SWR relative to the approximate center design frequency impedance (25.5 Ohms).

Table 7. Frequency-related properties of the original Anderson design and of an X-beam optimized at 28.5 MHz for maximum front-to-back ratio.

In a comparison with the standard 2-element Yagi, the X-beam has some deficiencies. First, X-beam gain only approaches the 2-element Yagi as the pattern deteriorates. Moreover, the X-beam has rearward lobes that can't be eliminated and that are only about 1.5 S-units down from the main lobe.

However, the Yagi standard rests upon some operating assumptions that may not hold true for all hams. The standard presumes a desire to work one station in a favored direction, with all other signals reduced, if not eliminated. Hence, maximum forward gain and front-to-back ratio are assumed to be desirable in all cases. Some

amateur radio operations may not benefit from this standard. For example, local and even wide-area nets may suffer from an inability to hear stations in "off" directions. Casual band sweeping in search of a rag chew may benefit from a less directive antenna, while still profiting from a little gain and front-to-back ratio. In short, an antenna must be judged by reference to an application as well as to certain performance specifications.

The folded X-beam is only one of several small beams that one might choose to build. Before putting an X-beam in the attic, you may want to compare it with another interesting beam that uses a closed geometry: the modified

Moxon rectangle. I'll look at this unjustly neglected mini-beam next time.

REFERENCES

1. Brice Anderson, W0PNE, "Designing X-Beams," *Antenna Compendium*, Vol. 1, American Radio Relay League, Newington, Connecticut, pages 64-66.
2. W2EY, "The X Beam on 20," *Antenna Roundup*, Tom Kneitel, K2AES, Editor, Cowan Publishing Corp. Port Washington, New York, 1966, pages 60-62; originally published in *CQ*, June 1965. Incidentally, W3EY's X-beam has different proportions than Anderson's design. Modeled without the all-metal support spider, from which the arms are insulated by plastic hose, the beam shows a free space gain of 6.0 dBi and a front-to-back ratio of about 26 dB in a "max F-B" model. The input impedance is low and capacitively reactive: not a good direct match to 50-ohm coax. Moreover, the director required considerable inductive loading, not the capacitive loading specified by the author. The differences between the model and actual antenna are likely due to W2EY's construction method.
3. *The ARRL Antenna Book*, American Radio Relay League, Newington, Connecticut, 1992, pages 26-21 to 26-23.
4. *Ibid.*, page 2-18.

Book Review

Price Wars: How To Win The Battle For Your Customer!

By: **Thomas J. Winninger**

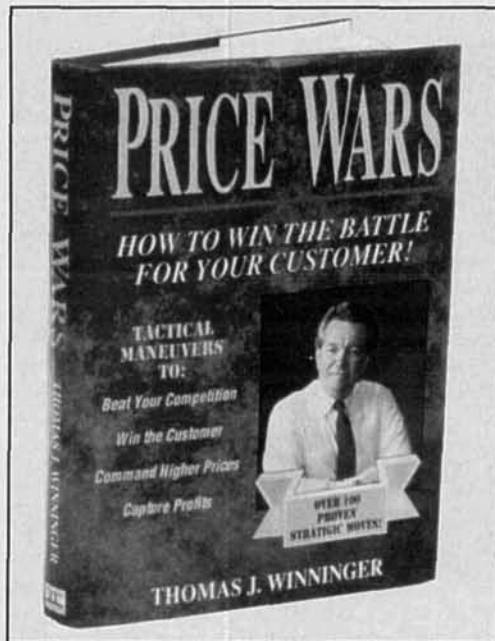
If you're a small business owner, now there's a book just for you! *Price Wars* by Thomas J. Winninger is an informational book that gives the small business owner the tools and resources to compete in a market that is rapidly bowing to price oriented services rather than value oriented services.

This upbeat and reassuring book covers such topics as defining niches, packaging, partnerships, pricing strategies, micro-marketing, and how to "restructure your organization to become consumer driven instead of product-driven." One key point is made in the foreword to the book: a quote from John Ruskin, English Author and Critic; "There is hardly anything in the world that some man cannot make a little worse and sell a little cheaper and the people who consider price only are this man's lawful prey." This point is used as the basis of creating tactics to promote the small business' value and expertise in an economy that coddles large impersonal, warehouses that trade value and expertise for lower prices.

Price Wars helps small business owners review their services and products and create an environment that will produce more consumers and subsequently, more sales. It also includes key points highlighted within the chapters and boxes titled "Battle Tactics." The "Battle Tactics" boxes list specifics covered in the chapter in an order that retailers can use to implement them in steps. Another key to *Price Wars* is its unending advice on how to keep customers coming back. It emphasizes promoting value and services to your customers not so much to rationalize your prices, but to help customers realize that smaller businesses mean more personal service, and a greater familiarity with the products or services sold.

Price Wars won't make your small business grow by itself, but I believe if a small business owner/retailer reads the book and implements some or all of the suggestions, they will improve their ability to compete with the ever-growing popularity of price oriented warehouses.

Price Wars is well written and easy to understand. After reading the book cover to cover, it becomes an



excellent reference book. My advice to small business owners, is to read *Price Wars* and mark the pages that have "Battle Tactics" for future reference. It may not save your business, but then again . . .

Price Wars is published by St. Thomas Press, measures 7 x 9 inches, is 288 pages in hardcover, and costs \$24.95 plus \$3.50 postage & handling. For more information, contact St. Thomas Press, 4510 West 77th St., Suite 210, Minneapolis, MN 55435; or call Kimberly Williams at 1-800-899-8971. To order (24 hours a day; seven days a week) contact Publisher's Distribution Service, 6893 Sullivan Road, Grawn, MI 49637. Consumer orders: 1-800-507-2665; Consumer Fax orders: 1-616-276-5797; Trade orders: 1-800-345-0096; Trade Fax orders: 1-800-950-9793.

Nancy Barry
Assistant Editor

TECHNICAL CONVERSATIONS

AF6S responds to Coke Darden's letter in the Summer issue of Communications Quarterly.

Dear Editor:

Obviously Coke Darden, W4DHA, has devoted much time and energy to the problem of measuring transmission-line length. His careful work has produced beautiful data that make sense. And his speculation, that harmonics of the MFJ-249's internal test signal produce the odd-even "staggered" results, seems reasonable and likely to be correct.

One could test that idea by using a directional coupler with an ultra-clean signal generator—or with a spectrum analyzer-tracking filter combination. Perhaps some reader will do just that, and report the results right here. I hope so.

David Barton, AF6S
San Jose, California

VE3DBQ has something to add to W0IYH's Summer 1994 article, "Designing the Long-Wire Antenna System."

Dear Editor:

W0IYH describes an intriguing project in the Summer 1994 issue, and I'd like to make a point about it.

He measures the ground resistance between his building ground and two ground rods. With another couple of measurements, he could find the actual resistance of each rod to true ground, or T.

Call his ground rods and building ground A, B, and C. Call the resistance of each ground to true ground, in order, x, y, z. Measure (AC) resistances C to A, A to B, and B to C: call them in order, k, l, and m. Then $x = 1/2(k + l - m)$. The remaining resistances y and z are easily found. (For a visual representation, see **Figure 1**.)

F.P. Hughes, VE3DBQ
Ontario, Canada

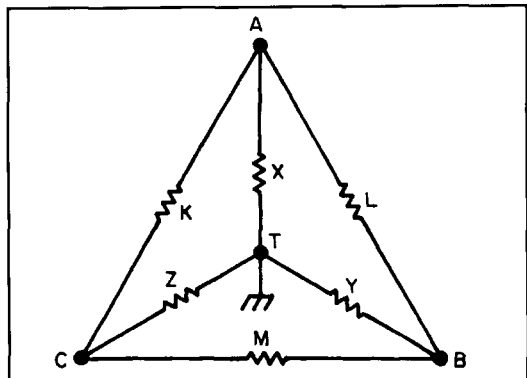


Figure 1. Making some additional measurements to find the actual resistance of each rod to true ground, T.

L.B. Cebik offers some further notes on the beta-match.

Dear Editor:

WA1SPI's recent review of the beta-match ("The Hairpin Match: A Review," by Thomas Cefalo, Jr., *Communications Quarterly*, Summer 1994, pages 49 to 54) provided both an excellent summary of the foundations of this L-section match and a productive computer program for calculating the hairpin match for Yagis. However, there are a few other facets of the beta-match that may be worth adding to the review.

The beta- or hairpin match is, as Cefalo notes, simply a special case of an L-section used to reduce the impedance at the source (the transmitter or feedline, R_o) to a lower antenna impedance. When shortened, the driven element shows a capacitive reactance (X_a) in series with the load resistance (R_a). By carefully selecting the degree of shortening, one can obtain the correct value to go with an inductive reactance (X_L) placed across the terminals (parallel to R_o) to create a matching section where $R_o > R_a$, as shown in **Figure 1**. Cefalo derives two classic formulas for calculating both of the reactances:

$$X_a = \text{SQR} (R_o R_a - R_a^2) \quad (1)$$

$$X_L = R_o R_a / \text{SQR} (R_o R_a - R_a^2) \quad (2)$$

In other literature, you may find R_o also called R_i (input impedance), R_1 , or R_p (impedance at the parallel reactance). Likewise, R_a has been called R_L (load impedance), R_2 , or R_s (impedance at the series element).

It is useful to derive some other forms of these two equations, because they will help us answer some questions about beta-matches and L-sections in general.

X_a in **Equation 1** can undergo some simple manipulations, as follows:

1. Collect terms and multiply by R_a/R_a to get:

$$X_a = \text{SQR} ((R_a^2(R_o - R_a))/R_a)$$

2. Bring out R_a to get:

$$X_a = R_2 \text{SQR} ((R_o - R_a)/R_a) \quad (3)$$

X_L in **Equation 2** can also undergo some manipulation:

1. Collect terms and bring the numerator under to radical to get:

$$X_L = \text{SQR} ((R_o^2 R_a^2)/R_a(R_o - R_a))$$

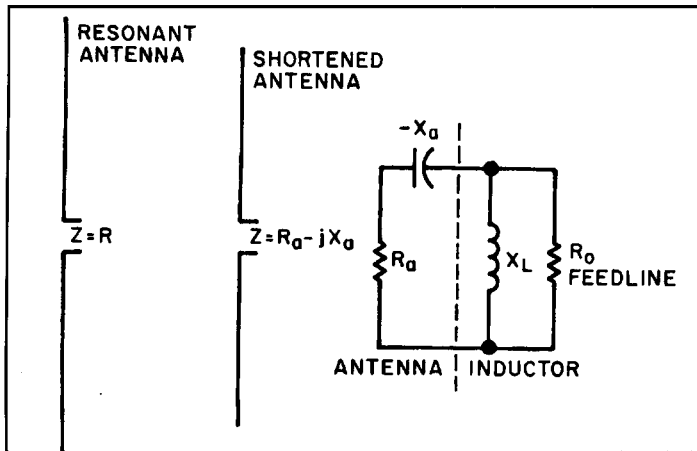


Figure 1. Beta-match network.

2. Cancel and bring out R_o to get:

$$X_L = R_o \text{ SQR} (R_a / (R_o - R_a))$$

3. Invert the radical to get:

$$X_L = R_o / \text{SQR} ((R_o - R_a) / R_a) \quad (4)$$

This exercise reveals the factor common to the calculation of both the series and parallel reactances: $\text{SQR} ((R_o - R_a) / R_a)$. The common factor has in more recent time been called the "network Q," but was earlier defined simply as δ (delta).¹ By collecting terms, δ becomes the more familiar:

$$\delta = \text{SQR} ((R_o / R_a) - 1) \quad (5)$$

so that the ratio:

$$R_o / R_a = \delta^2 + 1$$

Moreover, calculating the series and parallel reactances becomes simplified to:

$$X_a = \delta R_a \quad (6)$$

$$X_L = R_o / \delta \quad (7)^2$$

Because $R = 50$ ohms for most amateur cases, network Q or δ can be tabulated for most of the likely range of values to be encountered in beta-match work. **Table 1** shows the values of R_o / R_a in terms of resistance and as a ratio, with consistent values of δ , X_a , and X_L . Although values are shown for down to 10:1 matches (50 to 5 ohms), the more likelier cases involve ratios of less than 5:1 (50 to 10 ohms), with consequent values of δ between 1 and 2.

Besides slightly easing a set of calculations, δ also provides a means of estimating network losses. Cefalo notes in passing (page 51) that "Lumped-constant (coil-type) inductors can be impractical to use and may have a lower Q due to greater losses." He also uses equations for his hairpin "that are based on a lossless line because the hairpin will consist of open wire line that has negligible loss." While we might concede the general point that coils are lossier than open-wire transmission lines, we may well ask by how much they might be lossier.

If we let Q_u (the component Q) be the Q of the inductor (assuming, as usual—and as is sometimes dangerous at VHF—that the Q of capacitors of whatever sort is too high to be of concern), then:

$$\text{Power lost in the network} / \text{Power delivered to the network} = \delta / Q_u \quad (8)^3$$

It should be reasonably straightforward to construct a coil for X_L with a Q of 100. On this

Table of Beta-Match Values									
Ro = 50 Ω									
	<- More likely values ->				<- Less likely values ->				
Ra (Ω)	25	17	12.5	10	8.3	7.1	6.25	5.55	5.0
Ratio	2:1	3:1	4:1	5:1	6:1	7:1	8:1	9:1	10:1
δ	1.0	1.4	1.7	2.0	2.2	2.4	2.6	2.8	3.0
Xa (Ω)	25.0	23.6	21.7	20.0	18.6	17.5	16.5	15.7	15.0
X _L	50	35.4	28.9	25.0	22.4	20.4	18.8	17.7	16.7

(Note: All values are very rounded, since the table is designed to give only an indication of values.)

Table 1. Values for beta matches with 50-ohm inputs.

basis, losses in the network for the use of a solenoid inductor rather than a transmission line will be about 1 to 2 percent. This value compares well with the ballpark estimates of losses usually thrown without calculation into articles on antenna tuners and amplifier output matching sections: namely, around 3 to 5 percent loss. Whether or not this is an acceptable loss, compared to smaller losses for a section of transmission line, depends upon one's application and the conditions surrounding it.

In his summarization of the original Gooch-Gardiner-Roberts paper, Hall notes that calculations suggest a broader SWR curve for solenoid inductors than for hairpin inductors.⁴ The higher Q obtainable by transmission line sections supports such calculations, because real lines may have values of Q_u in the several hundreds. As Cefalo shows with his 20-meter example, the hairpin is often broad enough to cover all of a band with an SWR under 2:1. Note that his data reveals a prematch impedance of 22 ohms, which yields a figure for δ of just over 1; that is, a network Q at its broadest. More radical ratios of feedline-to-load resistance may show narrower bandwidths for the SWR curve. The lower Q_u of the solenoid inductor thus may prove useful for situations where one needs to transform lower antenna impedances and remain safely operational over a wider bandwidth. The cost is a couple of percent of the power, compared to a well-constructed hairpin.

Realistically, viewing open-wire transmission line sections as virtually lossless applies—if at all—only to freshly installed, tightly connected, unweathered assemblies. Component Q is a function of the sum of all ohmic losses associated with an inductor. We might loosely catalog the effects on the Q_u of inductors this way:

1. Inductor shape: Solenoid geometry, although capable with the proper materials of Q_u values well above 100, is no match for a transmission line section.

2. Wire losses related to current: Currents that exceed the capacity of any given diameter of wire to handle without significant heat, add to losses that lower the Q_u of any type of inductor. The solenoid and the hairpin require equal attention to wire size. Copper losses are not wholly negligible in real assemblies.

3. Skin-effect: Losses due to concentration of RF currents at the surface of the wire add to the previous category of losses and indeed may be considered an important part of it. Both types of inductors are susceptible to such losses.

4. Corrosion and weathering: New antennas do not stay new for very long. Corrosive salts and acids in the atmosphere eat at metals, while oxidation tends in many metals to increase the resistance at the surface, just where the skin effect would concentrate the current. In this

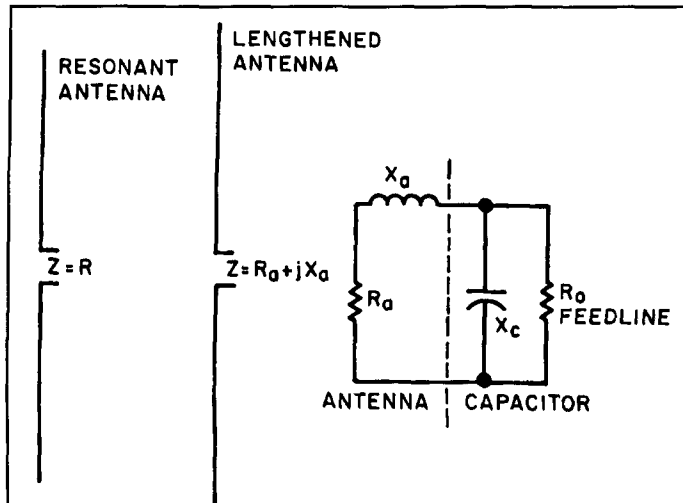


Figure 2. An alternative beta-match network.

regard, it may often be more practical to weather-protect a solenoid inductor than a hairpin loop. However, periodic maintenance is the guardian against excessive weathering losses.

5. Connections: Often neglected are the mechanical connections between a matching component and the antenna and feedline; yet, ohmic losses here are often the culprit in amateur antenna tales of woe. The fewer connections, the better, assuming each connection is of equal merit. A true hairpin suffers no more than a solenoid, because each has only end-point connectors. The sliding bar version of the hairpin presents two additional loss sources that become more significant the longer the antenna stays in the air (or the chemical soup, as some call it).

The gradual reduction in Q_u of an inductor—hairpin or solenoid—usually does not reveal itself as a radical change in antenna impedance. Instead, it often shows up as a gradual broadening of the bandwidth of an antenna. That is why SWR curves need to be run periodically to keep track of the condition of exposed parts. Lower Q is only sometimes a blessing: at others, it is a sign of impending trouble.

Two closing thoughts: first, the hairpin inductor gave rise to the beta-match, but need not be the only L-section matching scheme. In principle, one may reverse the reactance types. By lengthening the antenna driven element, the feedpoint impedance becomes complex with a resistance in series with an inductive reactance. Adding a parallel capacitive reactance across the line would complete the match, as shown in **Figure 2**. Reactance values would be the same as those in **Table 1**, with the series inductance represented by X_a and the parallel X_c equal to the tabulated value of X_L .

In practice, this scheme is unlikely to be especially useful, even if there are no mechanical objections to lengthening the driven element. Although well-chosen capacitors at HF are high-

Q devices, finding a suitable capacitor may be difficult. Most variable capacitors do not weather well, and most of the easily obtainable fixed capacitors have leads (both inside and outside the hard casing) that are inadequate for the current levels of high-power amateur RF.

Finally, it is not necessary to begin the driven element shortening process from scratch; that is, from a resonant length. Any of the antenna modeling programs will permit some close estimates of the length that will provide the correct measure of capacitive reactance. Then, the wire constructor makes the element only a bit too long.

The upshot of this discussion is twofold. First by rearranging the L-section formulas, we can obtain further information on the potential performance of a beta-match. Second, with that data in hand, we are better positioned to decide whether a hairpin transmission line section or a solenoid inductor is the better selection for a particular application.

**L.B. Cebik, W4RNL
Knoxville, Tennessee**

REFERENCES

1. See, for example, Jerry Hall, editor, *The ARRL Antenna Book*, 16th edition, The American Radio Relay League, Newington, Connecticut, 1991, page 25-11, or William Orr, *Radio Handbook*, 23rd edition, Sams, Indianapolis, Indiana, 1987, page 14-18, for references to "Q." See Terman, *Radio Engineers Handbook*, McGraw-Hill, New York, 1943, page 213, for information on the use of δ .
2. Some readers may recognize in these formulas Grammer's starting point for introducing impedance-matching networks in his little classic, *A Course in Radio Fundamentals*, The American Radio Relay League, Newington, Connecticut, 1972, pages 69-70.
3. Terman, page 213.
4. Gooch, Gardiner, and Roberts, "The Hairpin Match," *QST*, April 1962, pages 11-14, 146, and 156; compare Hall, editor, *The ARRL Antenna Book*, 16th edition, pages 26-21 to 26-23.

WB2TBQ is looking for assistance with a project he's designing.

Dear Editor:

I have subscribed to *Communications Quarterly* for almost two years now and have enjoyed both the practical project-oriented articles and the theoretical communications articles. I am currently stuck on a hobby design and thought I might ask your assistance either helping me with the design or recommending an article or publication, etc.

I have been trying to build a microcontroller-based receiver and am currently designing the VFO. I have build a great double-conversion superhet receiver and want to replace the analog oscillators with synthesized ones for stability, ease of use on all ham bands, and also just for fun to understand how synthesizers work. I have started my design using an MC145151 and an MC4024 or 4324 VCO according to **Figure 1**.

The op-amp would be a standard low noise op-amp moderate bandwidth device, because the speed at this stage would be slow...perhaps an LF 356 or so. I plan to hook up an MC68HC11 to the digital end of this synthesizer for the LO and another of the same setup for the mixing oscillator. I will use the 68HC11 to generate the divide ratios for the synthesizers and for computing frequency for a display and to perform functions such as scanning, memories, etc. The IRQ input would be hooked to a very smooth optically encoded dial for a frequency knob. The increment steps and frequency would be controlled in the microprocessor. The design would be quite inexpensive for a hobbyist design.

Unfortunately, my communications theory is over 10 years old and I am having some difficulty understanding the theory behind the component selection of the low-pass filter that is used between the ϕR and ϕV outputs of the synthesizer and the VCO. I understand the concept of the system, but need help understanding the formulas needed for component selection. I would like one oscillator system for 5 to 5.5 MHz LO in small steps and one for all the mixers; that is, 10.5, 7.9, and 13 MHz. I have an article on the design, but it is too theoretical for me to plow through all the mathematics to arrive at an understanding of component selections.

This project is an ambitious one for me, but I am using it as an educational tool for me to learn more about receiver and synthesizer design. Thanks, in advance, for any help you might be able to offer me.

**Michael Hatzakis, Jr., WB2TBQ
44 Newell Road, Apt. 11
Palo Alto, California 94303**

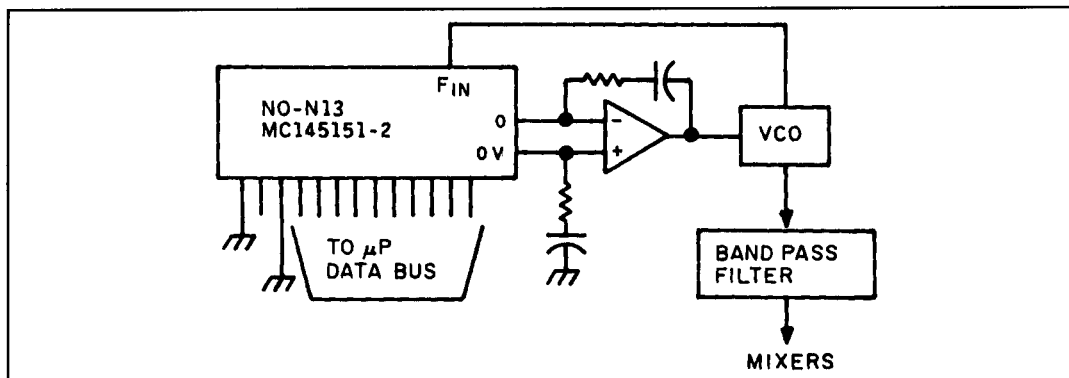


Figure 1. Block diagram of author's design.

POWER ON A BUDGET

Using the Russian Svetlana 4CX1600B power tetrode in modern amplifier designs

Something new has been added for high-power linear amplifier designs. It's from Russia with love—a conservative legal limit, cost-effective power tetrode tube.

Background

There was a film some time ago titled, "The Russians are Coming." The introduction of a rather complete line of high quality RF amplifier tubes manufactured in St. Petersburg, Russia, which employ the modern external anode technology, makes this a reality. A very large company—Svetlana Electron Devices, Inc., privatized in 1992—now sells its products worldwide. Recent descriptions in *Communications Quarterly*¹ of two of their tubes, gave me the incentive to try one to revitalize my needy homebrewed Class AB1 amplifier. The application data and results are presented here.

Svetlana 4CX1600B characteristics

The tube, and its custom SK3A socket, are shown in **Photos A and B**. It's a ceramic-to-metal external anode tetrode whose original application was in a military transmitter, which attests to its ruggedness and quality construction. (This tube was called the 4CX1600A, and had a much smaller cooler.) Thanks to several unique design features, the 4CX1600B exhibits high-performance when operated in class AB1 with relatively low anode voltage.

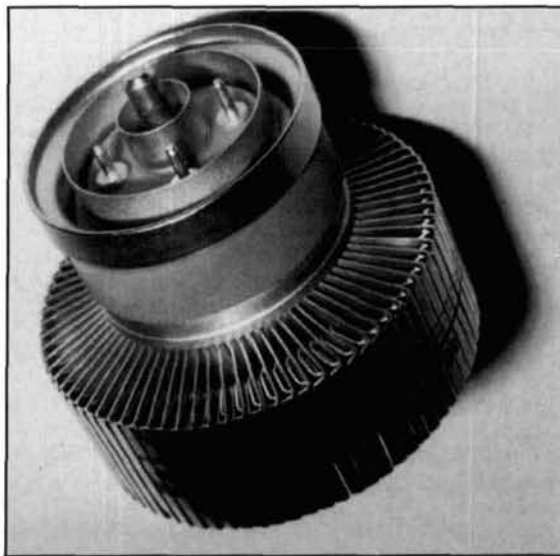


Photo A. Svetlana 4CX1600B. Photo by W6FR.

The anode was recently enlarged and is now essentially identical to the 8877 in size and configuration. Unfortunately, its matching chimney hasn't yet been modified to fit. To overcome this problem, I designed one of my own. I've been told that a compatible chimney will be available in the near future. For the general tube mounting outline, dimensions, and construction details of my homebrewed chimney, please refer to **Figure 1**.

Figure 2 shows the tube's specifications, along with my actual operating parameters, while running the tube as a grid driven ampli-

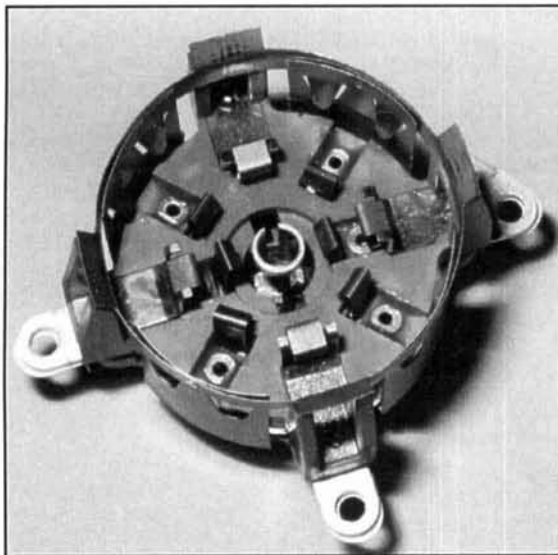


Photo B. The SK3A socket. Photo by W6FR.

er under single tone conditions. As you can see, it's conservatively rated for 1500 watts of continuous output at 75 MHz—employing a recommended 50-ohm passive grid configuration that greatly simplifies design and construction while minimizing the risk of parasitics and the need for neutralization. It requires only a net of 25 watts of PEP drive in class AB1, with grounded cathode for full output in passive grid. Because of its internal geometry, the 4CX1600B performs efficiently at a recom-

mended relatively low anode voltage of 2.4 kV for 1.5 kW output.

The tube requires the use of Svetlana's SK3A socket. This is a very modestly priced custom unit with a built-in, removable 0.01 mF annular screen bypass capacitor (mine measures 0.0114 mF). The upper frequency limit of this tube, at full power, is specified at 250 MHz. The bottom views of the socket and tube base are shown in **Figure 3A** and **B**.

The intermodulation distortion (IMD) specification for the 4CX1600B is a very respectable -36 dB for the 3rd order products, as measured by the conservative method to one of two equal tones. In amateur applications, which are referenced to the PEP, this would translate to -42 dB which, to my knowledge, exceeds all current transceiver specifications. If operated properly, without "all knobs to the right," you shouldn't experience any significant degradation of the tube's performance. This assumption is based upon the manufacturer's specified use in a grounded cathode configuration. Using my HP141T spectrum analyzer at 1500 watts PEP output at 14 MHz into a dummy load, I was essentially able to verify this, as the two-tone test results showed no change—with or without the final amplifier. I used my Kenwood TS950SDX as the driver, with 30 watts PEP (see **Photo C**). The scale was 10 dB/div and the sweep was 1/2 kHz/div.

Invariably, most amplifiers—that is, those in transceivers when operated at significantly

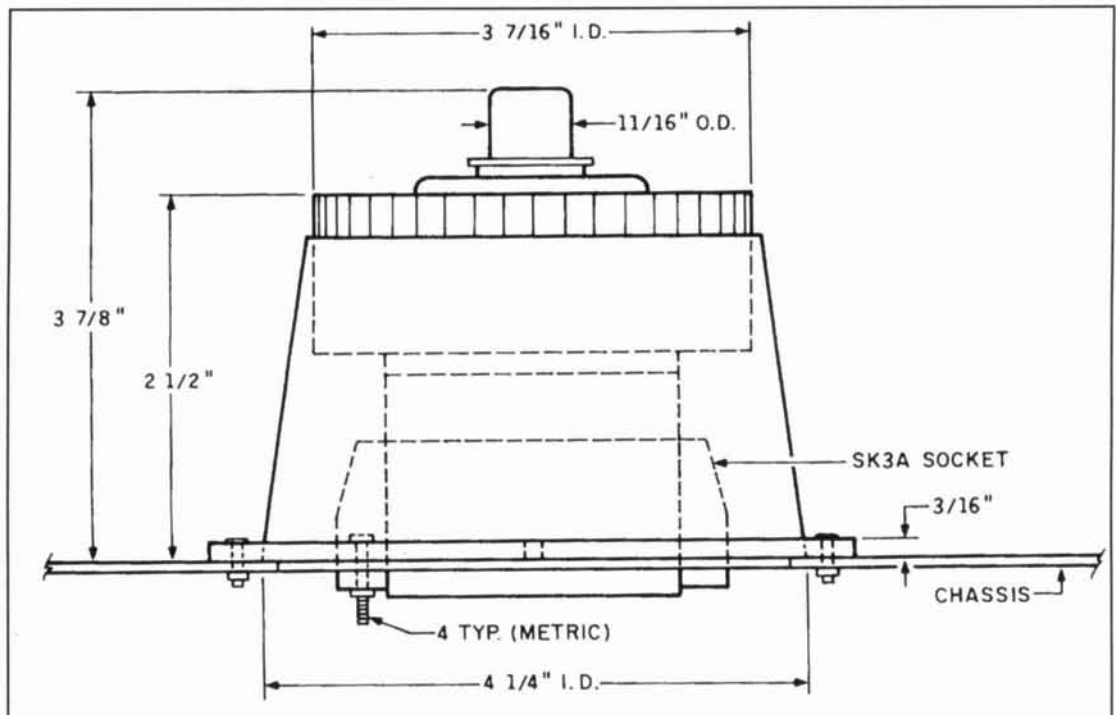


Figure 1. Tube mounting with chimney.

reduced power—will exhibit a corresponding improvement in IMD performance as evidenced in **Photo C**. As specified by the manufacturer, but not tested (see **Figure 2**), with increased grid bias and degenerative cathode bias of 24 ohms, the tube's specified efficiency rises from 61 to 72 percent. Unfortunately the IMD would suffer, increasing to -30 dB on the third order products from -36 dB. On the plus side, the zero signal anode current decreases from 500 to 200 mA; however, the drive requirement increases from 25 to 77 watts.

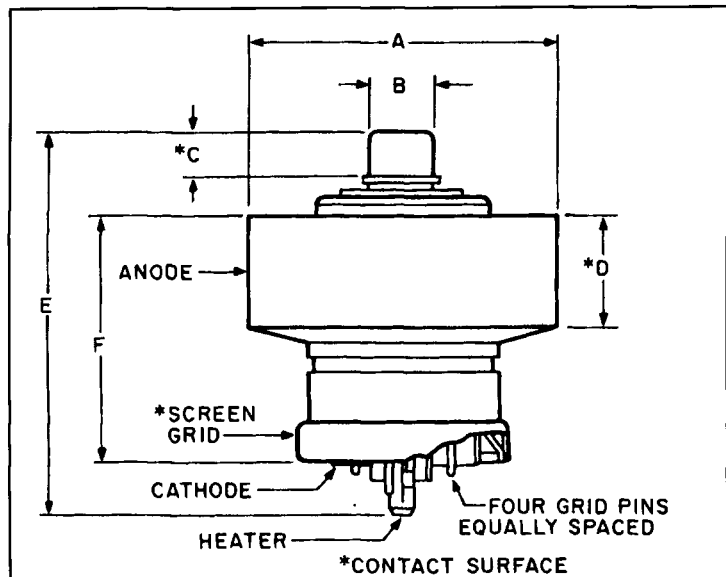
The chimney fabrication

As **Figure 4 A** and **B** illustrates, the chimney may be constructed inexpensively using commonly available materials. The parts are attached using clear silicone rubber cement (RTV). Scuff mating parts with a coarse, 100-grit sandpaper, apply the adhesive sparingly, and an excellent bond will result. My base mounting plate was fabricated from a sheet of 3/16-inch Lexan®, a polycarbonate. Almost any similar material will do, as there is no mechanical nor thermal stress on the mounting plate. I used a fly-cutter to fabricate the base mounting ring. Of course, this concept is applicable to almost any tube requiring a chimney for forced air cooling.

Some provisos

Svetlana recommends that the regulated screen grid power supply be capable of handling any negative current that may occur. This is relatively common to the tetrode class of tubes because they lack a suppressor grid. As a result, the supply—no matter what type—should be capable of maintaining good regulation under all conditions, e.g. positive or negative current. A very simple shunt regulator is therefore recommended, and could consist of a pair of OA2s or VR150s (OD3A) in series. This setup should just do the job, but only for the grounded cathode configuration, as they are specified at 30 and 40 mA, respectively. I'm cautious about zeners because of their inherent positive drift characteristic. Drift occurs with zeners unless they are extremely well heat sunk, oversized, and air cooled. I obtained good stability with eleven 30-volt, 5-watt zeners in series exposed to forced air cooling inside the plenum. A shunt regulator is a preferred means of control for the source of the screen voltage because it offers the fail-safe advantage of fold back. It may also be available from the high-voltage power supply, providing additional safety for purposes of interlocking the two.

It's possible to use an adjustable, electroni-



Dim	Inches		Millimeters	
	Min	Max	Min	Max.
A	2.83	3.76	84.8	85.6
B	0.67	0.75	17	19
C	0.394	—	10	—
D	0.787	0.866	29	31
E	—	3.740	—	109
F	1.968	2.155	66.6	68.6

Typical Operation, Grid Driven, Single Tone

Frequency	14 ⁴	75	75 MHz
Power output	1500	1600	1500 W
DC plate voltage	2450	2400	2400 V
DC screen voltage	330	350	350 V
Bias voltage ¹	-52	-53	-70 V
Zero-signal plate current	500	500	200 mA
DC plate current	1080	1100	870 mA
DC screen current	25	20	48 mA
DC grid current	0	0	0 mA
Driving voltage	51 ⁵	50	88 V
Driving power ²	30	25	77 W
CW plate input power	2646	2640	2068 V

Intermodulation distortion:³

3rd order	-36 ⁵	-36	-30 dB
5th order	-58 ⁵	-58	-34 dB
Efficiency	57	61	72 %
Cathode resistance	0	0	24 ohms

Notes:

1. Approximate value, adjust to specified zero-signal plate current.
2. The driving power is determined by the 50-ohm resistive input circuit.
3. Intermodulation distortion measured by two-tone method referenced to one of two equal tones at 500 kHz.
4. Author's parameters.
5. Presumed, not measured (see text).

Figure 2. 4CX1600B specifications.

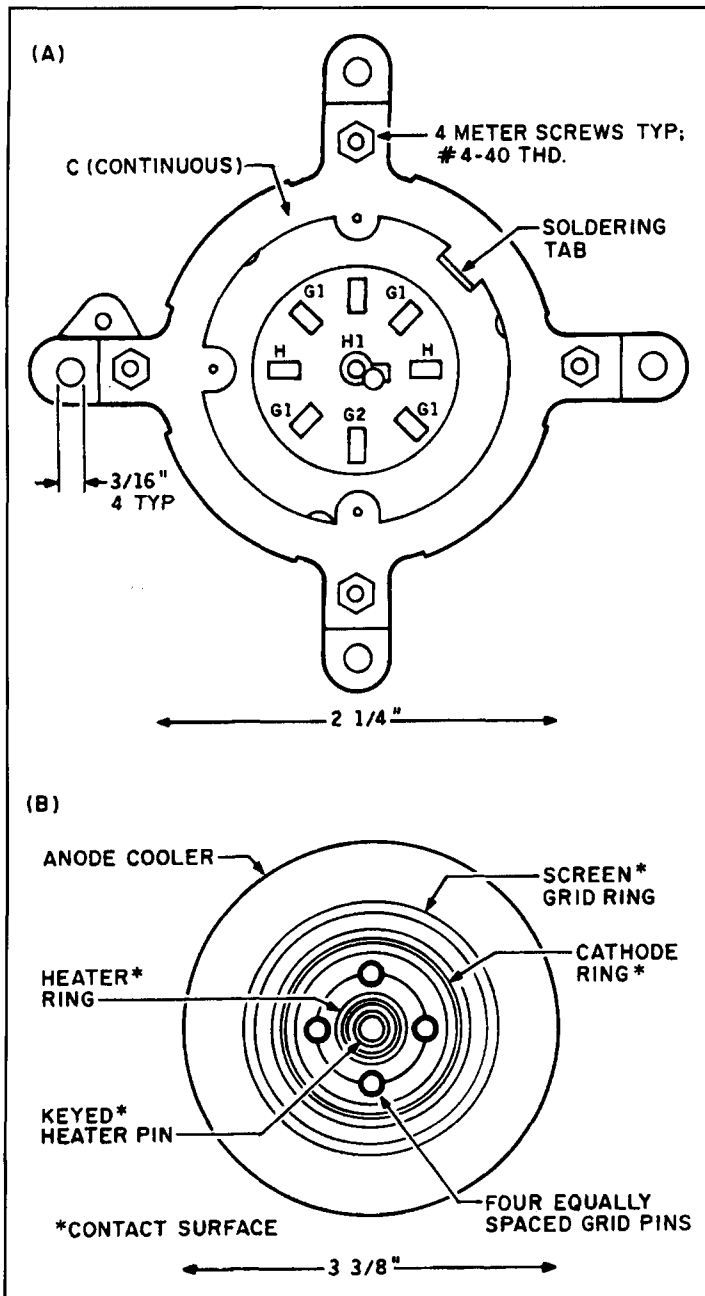


Figure 3. Bottom view of the (A) SK3A socket and (B) 4CX1600B.

cally, series-regulated power supply (I've used one in the past²), but it must be capable of handling possible negative screen current. To that end, I'd bleed a fixed amount of current consistent with an assumed worst-case negative current of 20 mA. Make sure the output voltage is clamped to prevent over-voltage. I'd use a pair of 1N2845Bs in series (180-volt 50-watt zeners) for this purpose. Also, to protect the screen from excessive dissipation, set the "roof" voltage about 90 volts above the regulated output voltage. This is just at the edge of the required differential for regulation while using a 40-mA

transformer. In the event of excessive screen current, that combination causes the voltage to fold back, or "starve," the regulator—thereby falling out of regulation. The manufacturer recommends that a source impedance of not more than 9,000 ohms be established as a current path from the screen to the cathode. Also, use a minimum of 50 ohms source impedance to the screen grid to further protect the tube in case of an arc. Both of these requirements are easily met with a shunt regulator. To read the screen current properly in either direction, I reset the zero forward manually—fully beyond the normal range of the adjustment screw—to 20 mA on a separate zero-to-one meter movement and calibrate for 100 mA full scale. I did this as a precaution in case a negative screen current occurred in addition to my normal, internal multimeter setup.

Svetlana also recommends that a 50-watt fixed resistor of at least 25 ohms be placed in series with the anode supply as a safety measure. This resistor is not used as a fuse, but to provide protection from infrequent internal arcs that are known to strike tubes at this power level. As with all tetrodes, the tube should never be operated without the anode voltage if the screen voltage is applied. It's an essential practice to only activate the two supplies together, or to interlock them accordingly. The heater voltage should never be applied without the proper cooling. The manufacturer has suggested the inclusion of a small inductance, as an L section in series with the 50-ohm input load resistance to ground, to cancel the 86 pF plus strays of the input capacitance of the tube for operation at 10 meters. I would recommend that this be modified into a fixed Pi section with a 91-pF input capacitor with a 0.33-μH inductor (5 turns #16 spaced to occupy 1 inch on a Micrometals T130-6; yellow core), which is effectively followed by my estimated total of 90 pF of grid input capacitance—allowing for strays. The calculated value for 28.5 MHz alone would be 0.27 μH. This Pi section will have negligible effect on frequencies below 10 meters, and will enhance the match on that band while offering a small compromise for a slightly better match at 14 MHz. I provide this information for the purist, as most modern tuners will accommodate the basic mismatch resulting from the total input capacitance at the grid.

The heater is specified at 12.6 volts plus or minus 5 percent at 4.4 A. It's best to be slightly on the low side, as this enhances tube life along with providing some minor IMD reduction. To test for this, reduce the heater voltage *just* until the power decreases slightly, then increase it a few tenths of a volt above that point until full output is restored. You can repeat this test periodically during the life of the tube.

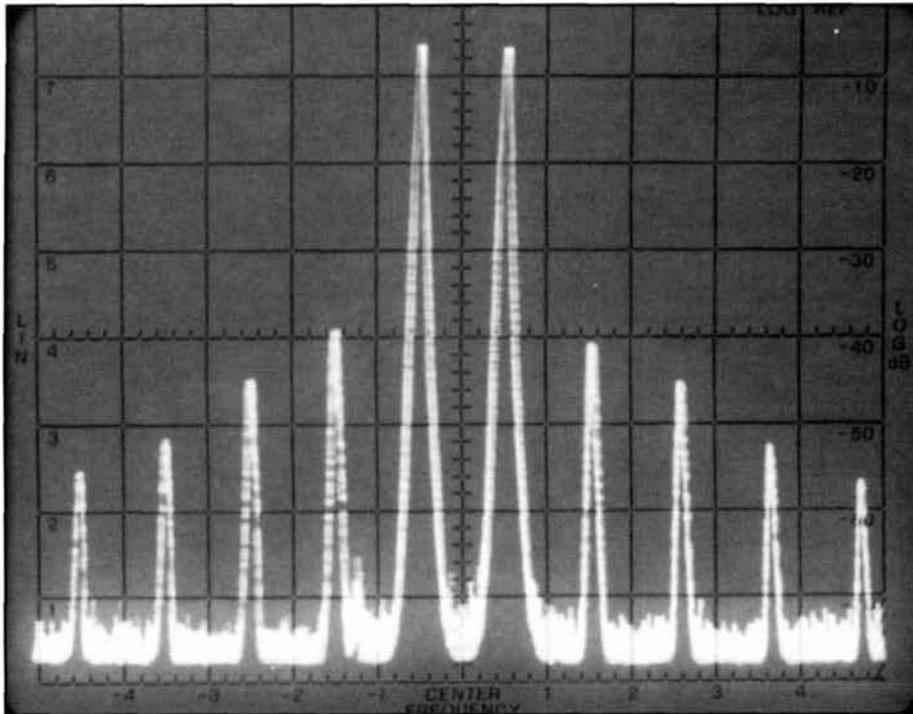


Photo C. IMD performance. Photo by W6FR.

The heater voltage can be obtained conveniently through series connection of two properly phased 6.3-volt windings, or by commonly available transformers for 12 volts DC power supplies followed by a precise adjustment with an autotransformer or variable resistance. I used an old UTC S-70—a dual 6.3-volt, 5-A transformer with no problem; it's air cooled in the plenum, too. Two and one half minutes of warm up time is required for the heater to reach 90 percent of its safe current demand, although I conservatively elected to use three minutes. The tube may be operated in any position.

To enhance tube life, I recommend a Keystone Carbon Company Inrush Current Protector type CL-60, a negative coefficient thermistor, in series with the heater circuit. This device is specified to offer 10 ohms at 25 degrees C, dropping to 0.18 ohms at 5 A. These specifications closely fit the heater current drain and offer a smooth ramp of 10 seconds to bring the heater up to its rated voltage and temperature—reducing thermal stress to a reasonable minimum. Bear in mind that the thermistor must be placed *outside* the cooling airstream to function, and away from heat-sensitive objects for safety. In my amplifier, it was necessary to use Keystone's CL-190 in series with the primary because of my air cooling system. This device is rated at 25 ohms at 25 degrees C and ultimately falls to 0.41 ohms. In my case, it drops to 1.79 ohms at less than full load; that is, about 0.7 A. This unit also provides a nice time constant of approximately 10 seconds. The

Electrical

Cathode:	Oxide-coated
Voltage	12.6±0.6 volts
Current, at 12.6 volts	4.4±0.3 A
Voltage, cathode-heater, max.	±100 volts
Warm-up time	2.5 minutes
Transconductance	50 mA/V

Direct interelectrode capacitances (grounded cathode):

Input	86 pF
Output	12 pF
Feedback	0.15 pF
Maximum frequency	250 MHz

Mechanical

Maximum overall dimensions:

Length	95 mm (3.51 in.)
Diameter	72 mm (2.81 in.)
Net weight	600 grams (1.21 lb.)
Operating position	Any
Cooling	Forced air

Maximum operating envelope

temperature	225 degrees C
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Radio Frequency Linear Amplifier, Class AB1

DC plate voltage	3.0 kV
DC screen voltage	350 volts
DC grid voltage	-150 volts
DC plate current	1.4 A
Plate dissipation	1.6 kW
Screen dissipation	20 watts
Grid dissipation	0.1 watts

Table 1. 4CX1600B general characteristics.

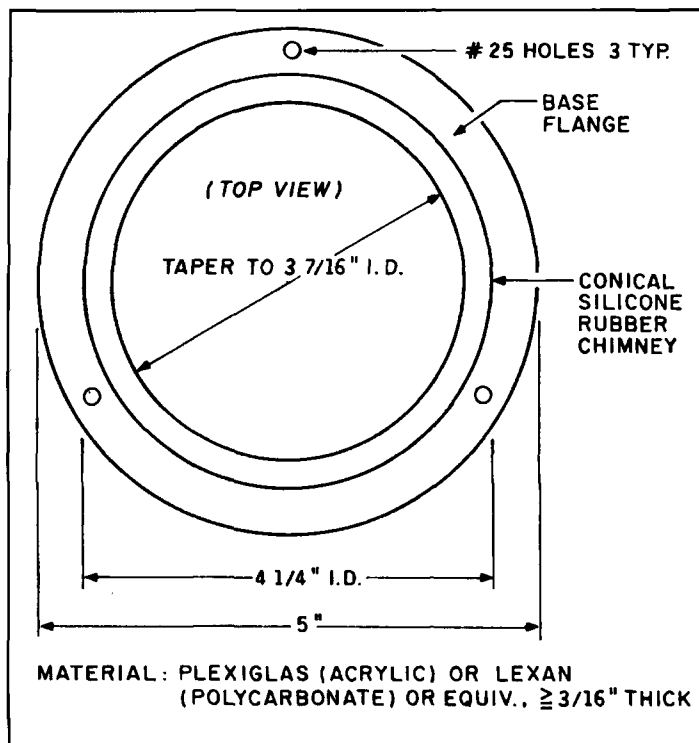


Figure 4A. Chimney mounting flange and design pattern, top view.

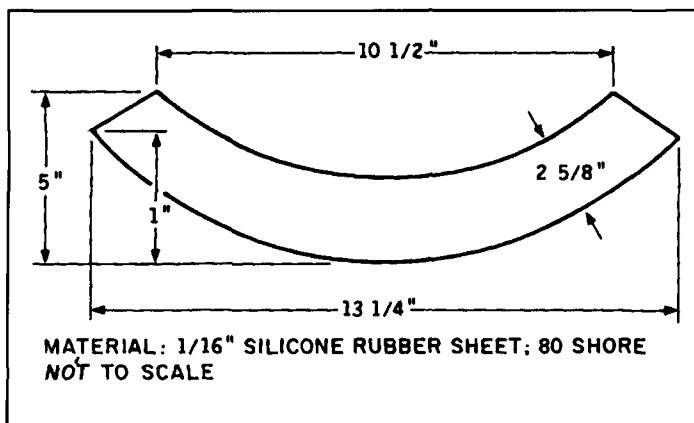


Figure 4B. Chimney design pattern.

cathode, when operated above ground, is specified at a maximum of 100 volts—regardless of polarity, between it and the heater.

Last, but definitely not least, the cooling requirement for 1600 watts dissipation at sea level, at 25 degrees ambient intake air temperature, is 44 c.f.m. with a back pressure of 0.50 inches. The anode temperature must be limited to 225 degrees C (433 degrees F). It's a well-known fact that tube life is enhanced with low envelope temperatures. I found that the recommended air flow specification is most conservative for ICAS service. In my amplifier, the exit air temperature only increased 46 degrees F at 1500 watts output during heavy duty SSB operation.

The SK3A socket

I like the way the tube's element configuration and related socket contacts (all of which are spring loaded, see **Photo B**) are designed for low inductances. The heater contacts are brought out coaxially. The cathode is a ring surrounding the control grid pins, brought out through four equally spaced pins connected in parallel inside the base. The screen grid is also a ring at the base of the tube to which contacts are made via eight points around its circumference. Four contacts are made in the tube socket to the cathode ring. These are brought out to the four mounting ears so, with operation such as mine using an electronic bias switch (EBS),³ the cathode contacts must be isolated from ground. Otherwise, these terminals provide a nice, low-inductance means of directly grounding the cathode.

To isolate the four cathode connections that also serve as the socket mounting holes, I chose nylon insulating bushings used with the TO-3 type transistor. These fit the existing socket mounting holes precisely. By cutting off their stems, I could also use them to make insulating washers for the nut end of the mounting screws. These bushings are readily available at parts houses and are very inexpensive. The mounting job to the deck was completed using flat head 4-40 machine screws. The general characteristics for the 4CX1600B are given in **Table 1**.

Construction

In my homebrew amplifier, the layout and format are quite conventional with a Pi-L output. The amplifier is essentially designed for 20 meters. All capacitors and RF relays are vacuum type for reliability and low inductance. One of the two 4-inch panel meters serves as a center zero 500- μ A tuning phase detector⁴ and the other, a zero-to-one milliammeter, serves as a multimeter that reads all the currents and low voltages. Their layouts are shown in **Photos D** and **E**. For details of the Pi-L output network, refer to **Table 2**. By today's conventions, my unit is pretty large and probably overdesigned. However, I intended it to run continuously at 1500 watts without a whimper! I use a low velocity axial exhaust fan located at the top of the cabinet to enhance cooling. This type of construction is a near necessity for making IMD measurements because of the heat generation, and results in higher reliability. The schematic of my amplifier is shown in **Figure 5**.

Amplifier performance

By generally following Svetlana's typical operating voltage and current guidelines (2.45

Zin ^{1,2} (ohms)	Freq. (MHz)	C1 ³ (pF)	L1 ([MU]H) (pF)	C2 ([MU]H)	L2
5000	1.80	239	40.011	1696	8.917
5000	2.00	190	40.011	1316	8.917
5000	3.50	123	20.272	872	4.518
5000	4.00	95	20.272	658	4.518
5000	7.00	57	11.108	387	2.476
5000	7.30	52	11.108	360	2.476
5000	14.00	29	5.651	186	1.259
5000	14.35	27	5.651	183	1.259
5000	21.00	19	3.780	125	0.843
5000	21.45	18	3.780	123	0.843
5000	28.00	15	2.730	95	0.609
5000	29.70	13	2.730	89	0.609

Notes:

Zin from the formula:

$$R_L = \frac{2500}{1.5 \times I} \approx 1666 \Omega$$

1. ARRL Handbook, 1989 edition, page 15-3, 15-8.
2. Collins, *Amateur Single Sideband*, 1962, page 68.
3. Includes output capacitance plus strays (12 + 5 pF).

Table 2. Pi-L values¹ Q=12

kV on the anode with 330 volts on the screen), as given in **Figure 2**, I was able to generate 1500 watts of continuous carrier at 14 MHz. The tube ran quite cool into the dummy load. In SSB or CW service, using my EBS (see **Figure 5**), it ran much cooler than before. It saves power consumption, thereby reducing heat, increases tube life, and eliminates all background noise like fans, etc.—so I wouldn't be without it. Note that this tube's quiescent anode current is specified at 500 mA, under the condition of grounded cathode for Class AB1, which at 2.45 kV results in just over 1200 watts of pure power dissipation. That, alone, justifies the use of the EBS. From the manufacturer's specifications, along with its low inductance mechanical configuration, it appears that this tube would make an ideal 1.5 kW amplifier for VHF.

One caveat: the control grid dissipation is rated at 0.1 watts, so you must ensure that essentially zero grid current flows—and for good reason. First, it precludes one form of nonlinearity that creates distortion and resultant splatter and second, it protects the grid dissipation and enhances the life of the tube. I recommend high-level ALC, as shown in **Figure 5**. It senses a few microamperes of grid current to generate bias that automatically reduces the drive from the exciter, ensuring the integrity of the grid dissipation and linearity while providing the correct drive power level. Further, due to the scales and/or ballistics of most meter movements, especially under voice conditions, it is very difficult, if not impossible, to detect

grid current. This current, if present, will seriously and almost exponentially degrade the linearity of the amplifier. At the same time, it will cause an upswinging of the wattmeter due to the distorted, ("flat-topped") waveform. It's almost impossible to see small increments of this phenomena on an oscilloscope, but the result is very apparent over the air and quite visible on a spectrum analyzer. As a simple rule, if you observe *any movement* of any grid current meter, you're either underloaded and/or overdriving the amplifier—creating distortion.

The typical Christmas tree pattern isn't totally reliable for detecting nonlinearity, even if precautions are taken. However, it can be more informative if you follow the procedure below, which is illustrated in **Figure 6**. First, lower the baseline of the oscilloscope to the bottom of the CRT, doubling the effective size of the resulting half pattern. Second, increase the sweep speed to display no more than two or three tree patterns. Now, load the amplifier under voice conditions, for the highest and sharpest single peak. If you want an alternative sensor for improper or inadequate loading, watch the screen current. Look for times when the current appears high or takes an abrupt, severe upward excursion at the onset of grid current. I know of one ALC system that employs this characteristic as a means of generating its control voltage.

For best combined SSB power out and IMD performance, it's necessary that the amplifier be heavily loaded; not by a single tone, as is often practiced, but by voice, a high speed

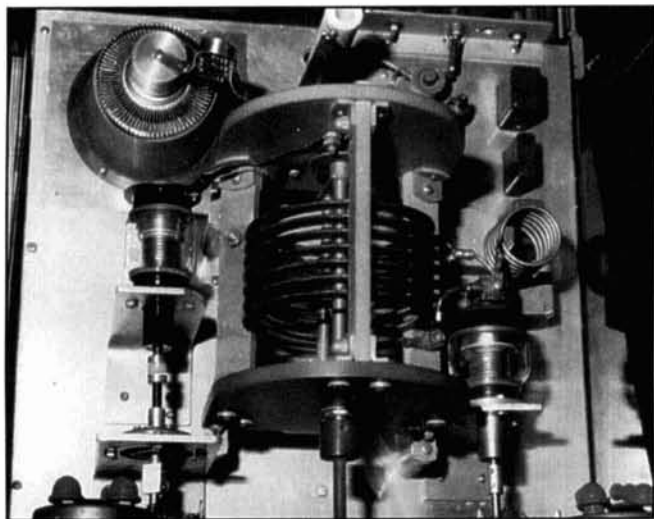


Photo D. PA layout, interior. Photo by W6FR.

pulsar that simulates the voice, or two tones—in that order of preference. As an alternative approximation *and* a last resort, load up fully using a single tone and continue loading approximately 10 percent beyond that point until a slight power decrease is noted. This will only be in effect under single-tone (CW) operation; under SSB conditions it will represent the maximum linearity and power out because of the differing anode impedances between a single tone versus the complex, voice waveform. The power output phenomenon, under voice operation, is easily verified on an oscilloscope, but often overlooked—causing needless and annoying distortion products.

Initial operation

There are a few things you can do to make

C	0.1/600 volt discs
C1	0.01 integral to SK3A
C2	0.001/5 kV Centralab 850
C3	0.33/100 volt paper
C4	0.001/600 volt discs
R1	Globar Type A, 25 watt, 50 ohm, noninductive
R2	25 ohms, 50 watts, wire wound (see text)
R3	25 k, 10 watt, wire wound
R4	100 k, 2 watts (fail safe for R3)
T1	Audio transformer; 500 ohms, 1:1
D1	1N914 or HP2800
S1	Test/Disable switch
Q1	2N3439 or 2N3440
Q2	2N3902
D2	1N56644 "Transorb" for optional spike protection

Keystone Inrush Limiter type CL-60 negative coefficient thermistor (see text).

All resistors are 1/4 watt, except as noted.

initial operation a bit easier and safer. First, grid dip the anode tank circuit for the frequency of operation. Then, check all circuits for continuity, shorts or opens, etc. Next, place a 100-ohm, 100-watt resistor in series with the high voltage, which I would initially reduce to around 1800 volts.

Insert a 5-k, 20-watt resistor in series with the screen voltage, which would also be preliminarily reduced to 250 volts. Allow proper time for warmup, and measure the heater voltage. Now, with the anode and screen voltages applied and cathode grounded, adjust the grid bias and/or screen voltage for approximately 325 mA of quiescent anode current. Then, slowly bring up the grid drive power to a few watts and watch for any peculiar meter readings or other signs of instability while peaking the power out. If all goes well, cautiously bring up the anode and screen voltages, and the drive, and reduce the values of the series current-limiting resistors until all the normal values are obtained. Of course, make sure to use a good dummy load and remember *high voltage kills!* Always short the high voltage lead to ground after any high voltage has been deactivated, and before commencing any other work regardless if the high voltage meter reads zero. It's also a good safety precaution to work with just one hand when any high voltage is on, or even if it has been deactivated.

Conclusion

The tube performed precisely to my expectations, and proved itself a very cost-effective means for providing a stable legal-limit linear amplifier using a passive grid input. The trade-offs, as Littlefield pointed out,¹ of using

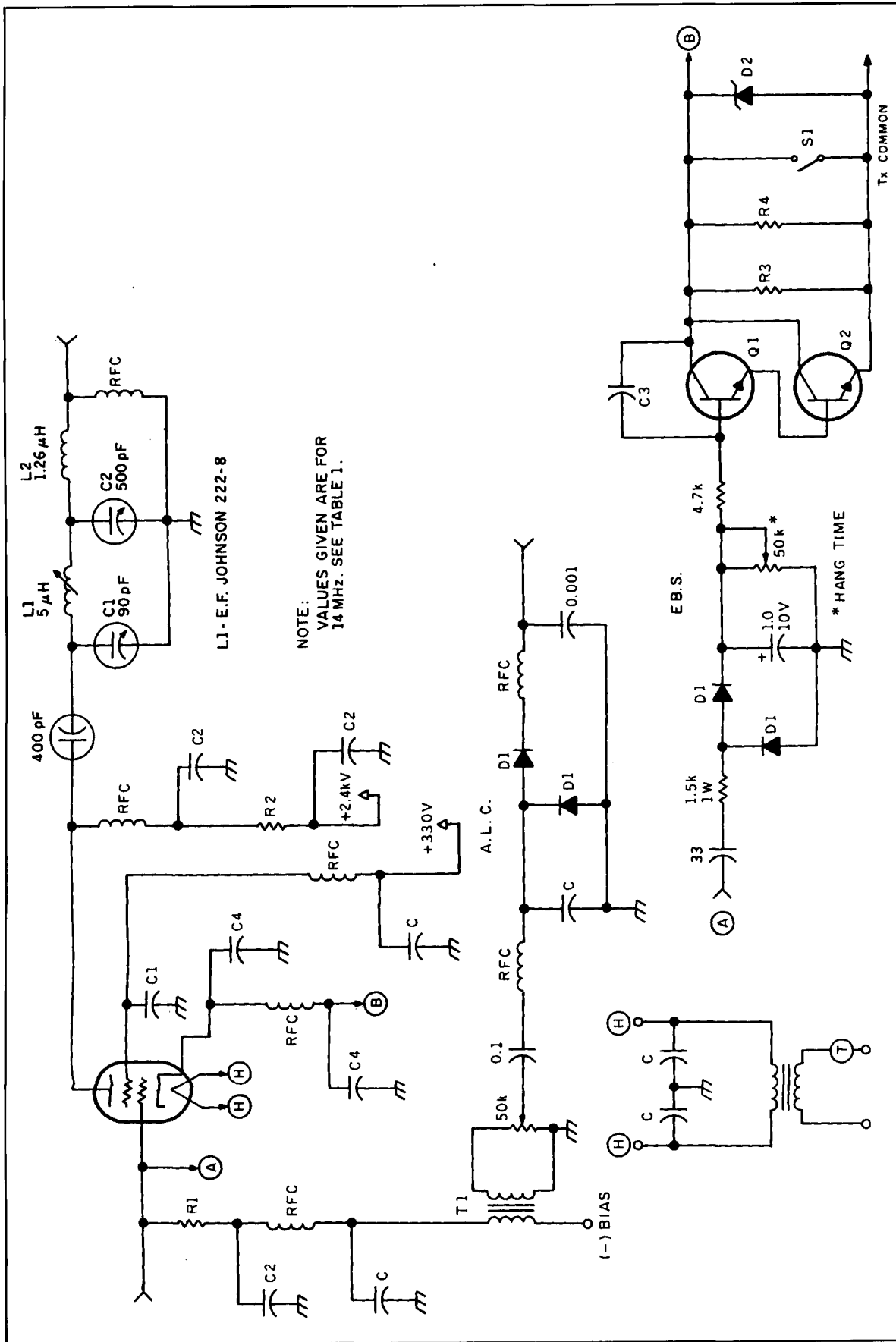


Figure 5. PA schematic.

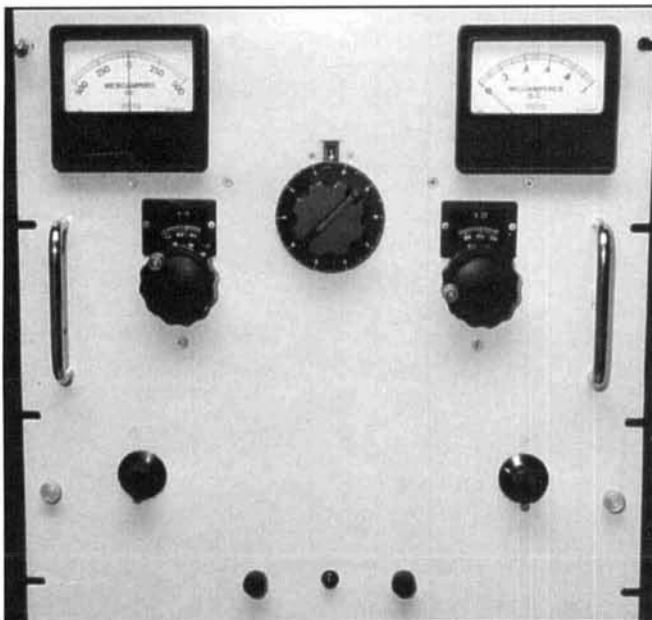


Photo E. PA layout, front panel. Photo by W6FR.

tetrodes instead of triodes have become more significant in recent years. As a result, tetrodes may now be given much more serious consideration for projects such as mine. Using one in the manner described here is a significant way to provide a high quality signal, while avoiding a number of the common electrical and mechanical problems associated with amplifier construction and performance. I highly recommend its consideration.

Acknowledgements

I wish to gratefully acknowledge and express my sincere appreciation to W6JAZ and W6TC. Their contributions significantly assisted me in my work.

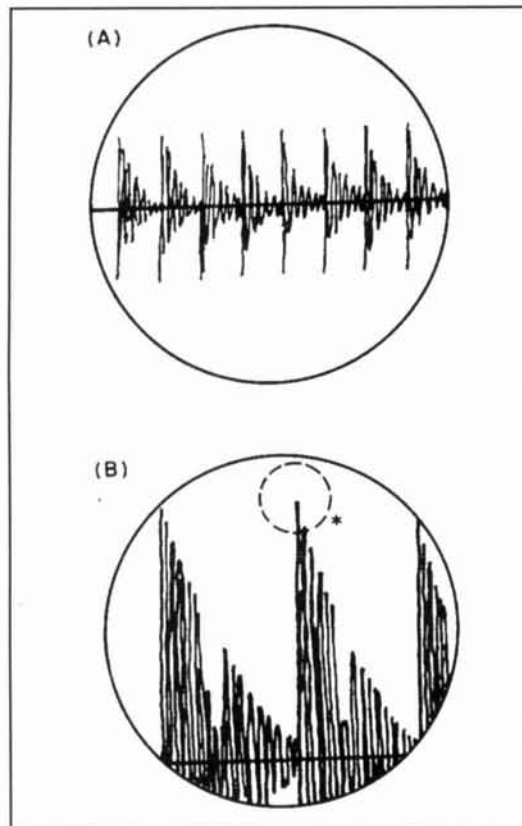


Figure 6. Improved loading procedure. (A) Common presentation; (B) enhanced presentation (see text). *Focus only on one peak for greatest, sharpest amplitude while progressively increasing loading and drive.

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1. Rick Littlefield, K1BQT, "Quarterly Devices," *Communications Quarterly*, Summer 1993, page 89.
2. Richard Measures, AG6K, "A Regulated Screen Grid Power Supply," *Ham Radio*, June 1986, page 51.
3. Marv Gonsior, W6FR, "Electronic Bias Switching," *Ham Radio*, March 1975, page 58.
4. *Amateur Single Sideband*, 1st edition, Collins Radio Company 1962, page 77.

The manufacturer

Svetlana Electron Devices, Inc. manufacturer of the 4CX1600B. The word svetlana means light in Russian, which was appropriate enough when the company initially produced light bulbs. Svetlana is a relatively large, old line company, organized in 1898. The company began producing electron tubes in 1928. Svetlana's vacuum tube engineering and production staff, alone, consists of more than 2,000 persons. Apparently little was known about them until recently, as their production was solely for domestic and military consumption. The company manufactures a wide variety of quality tubes from triodes to pentodes. They range in size from the smallest tubes to extremely large devices, manufactured under

rather impressively strict quality control procedures. For instance, Svetlana tubes are processed at higher temperatures than those in the West. An extensive two year warranty is offered and their products, including compatible sockets, are far more inexpensive than those prevailing for similar devices. A number of their tubes are directly interchangeable with existing Western types, and more will be coming out in the future.

For more information on Svetlana Electron, Inc. and their line of tubes, write: Svetlana Electron, Inc., 8200 Memorial Parkway, Huntsville, Alabama 35802. You can also phone the company at (800) 239-6900.

QUARTERLY DEVICES

Pin diodes

PIN diodes are nothing new: commercial equipment designers have been using them for several years in a variety of RF-switching and attenuator applications. However, these devices have been conspicuously absent from RF construction projects published in the popular literature. There's really no technological reason for this, because PIN diodes are easy to obtain and use. In this edition of "Quarterly Devices," we'll look at two popular low-power switching diodes from Motorola that are especially well suited for receiver and low-power transceiver projects.

How PIN diodes work

The PIN diode is, essentially, a diode switch that exhibits progressively less resistance to RF energy as forward bias is increased. When fully biased, the PIN diode acts like a piece of wire, exhibiting a very low-resistance RF path. At zero (or reverse) bias, it acts like a small-value capacitor—exhibiting 40 dB or more attenuation to signals in the HF or VHF range. Power handling ability depends upon the size of the diode's internal geometry. A large diode structure will exhibit *very* low forward resistance (perhaps 0.1 ohm). This allows the diode to handle a lot of RF energy without breaking down. However, the larger geometry also yields more capacitance at zero-bias—and this progressively diminishes the diode's ability to attenuate RF at progressively higher frequencies. A small diode structure has the opposite effect. Forward-bias resistance is higher (perhaps 1.0 ohm), and power-handling capability is less. However, the small structure yields lower capacitance at zero bias, and offers far better high-frequency attenuation. When selecting a PIN diode, it's important to match diode characteristics to your application.

Advantages of PIN diodes

When it comes to routing RF, PIN diodes provide several significant advantages over relays and switches. Two of the more obvious

benefits are physical; PIN diodes can't become oxidized by the environment, and they won't fail from mechanical wear after long-term use. In terms of board layout, you can place PIN diodes anywhere without regard for control panel location. Leads are short, so their impact on high-frequency RF circuitry is minimal. PIN diodes fit easily into small RF-tight subenclosures, and multiple switch points may be activated simultaneously for complex routing or for bandswitching. Perhaps the biggest advantage, in this age of digital communication, is speed. PIN diodes provide much faster switching times than electro-mechanical T/R relays—making them especially valuable as switches in high-speed data or digital-mode transceivers.

PIN diodes also offer superior switching performance over conventional small-signal switching diodes such as the 1N4148 at RF frequencies. Recent articles document the advantages and degree of improvement realized from making substitutions in commercial equipment.^{1,2} For a more thorough discussion of PIN-diode physics and operating characteristics, read the footnoted articles and refer to the PIN-diode section in a recent edition of *The ARRL Handbook* under "Solid State Basics."

The Motorola MPN3700 and MPN3404

Although some of the more exotic instrumentation-grade PIN diodes are a bit pricey for casual experimentation, there are a number of economical small-signal PIN diodes available from foreign and domestic sources. Devices like the BA482 and BA389 are quite popular, and have found their way into several low-power amateur and commercial products. Two offerings from Motorola, the MPN3700 and MPN3404, have similar characteristics and fall into this category. Both were designed primarily for VHF bandswitching, but are equally applicable to general purpose RF-switching and attenuator circuits. See **Figures 1** and **2** for complete technical specifications.



CASE 182-02
TO-226AC
TO-92
MPN3700



CASE 318-07
TO-236AB
SOT-23
MMBV3700L

HIGH VOLTAGE SILICON PIN DIODES

... designed primarily for VHF band switching applications but also suitable for use in general-purpose switching and attenuator circuits. Supplied in a cost effective plastic package for economical, high-volume consumer and industrial requirements.

- Long Reverse Recovery Time
 $t_{rr} = 300$ ns (Typ)
- Rugged PIN Structure Coupled with Wirebond Construction for Optimum Reliability
- Low Series Resistance @ 100 MHz —
 $R_S = 0.7$ Ohms (Typ) @ $I_F = 10$ mA
- Reverse Breakdown Voltage = 200 V (Min)

MAXIMUM RATINGS

Rating	Symbol	MPN3700	MMBV3700L	Unit
		Value		
Reverse Voltage	V_R	200		Volts
Total Device Dissipation @ $T_A = 25^\circ\text{C}$ Derate above 25°C	P_D	280	200	mW
		2.8	2.0	mW/ $^\circ\text{C}$
Junction Temperature	T_J	+ 125		$^\circ\text{C}$
Storage Temperature Range	T_{stg}	- 55 to + 150		$^\circ\text{C}$

ELECTRICAL CHARACTERISTICS ($T_A = 25^\circ\text{C}$ unless otherwise noted)

Characteristic	Symbol	Min	Typ	Max	Unit
Reverse Breakdown Voltage ($I_R = 10 \mu\text{A}$)	$V_{(BR)R}$	200	—	—	Volts
Diode Capacitance ($V_R = 20$ Vdc, $f = 1.0$ MHz)	C_T	—	—	1.0	pF
Series Resistance (Figure 5) ($I_F = 10$ mA)	R_S	—	0.7	1.0	Ohms
Reverse Leakage Current ($V_R = 150$ Vdc)	I_R	—	—	0.1	μA
Reverse Recovery Time ($I_F = I_R = 10$ mA)	t_{rr}	—	300	—	ns

Figure 1. Motorola's MPN3700 high-voltage silicon diode.

The MPN3700 and MPN3404 exhibit nearly identical path loss (about 0.7 ohm) when forward biased, and both fit the profile of a small-signal device—offering excellent high-frequency attenuation, but limited RF power-handling capability. However, this doesn't mean the

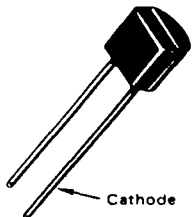
devices are identical. The MPN3700 carries a reverse-voltage rating of 200 volts, which is ten times greater than that of the MPN3404. From a practical standpoint, this means the MPN3700 can fend off a bigger voltage hit without breaking down—a particularly desirable characteris-

tic for protecting sensitive receiver front-ends and MOS-amplifier gates. On the other hand, the MPN3404 is rated at 400-mW total device dissipation, while the MPN3700 is only rated at 280 mW. As a result, the MPN3404 can be forward biased at a higher DC current level and can pass more RF power without breaking down. For these reasons, the MPN3700 is probably a better choice for the receive side of a T/R switch, and the MPN3404 is probably a

better choice for the transmit side.

Biasing PIN diodes

The MPN3404 and MPN3700 pass low-level RF signals with relatively little forward bias. For signals below +20 dBm, 10 mA forward bias is probably sufficient. However, at progressively higher RF power levels—on the



SILICON PIN DIODE

... designed primarily for VHF band switching applications but also suitable for use in general-purpose switching and attenuator circuits. Supplied in a cost effective TO-92 type plastic package for economical, high-volume consumer and industrial requirements.

- Rugged PIN Structure Coupled with Wirebond Construction for Optimum Reliability
- Low Series Resistance @ 100 MHz – $R_S = 0.7 \text{ Ohms (Typ) @ } I_F = 10 \text{ mA Dc}$
- Sturdy TO-92 Style Package for Handling Ease

MAXIMUM RATINGS

Rating	Symbol	Value	Unit
Reverse Voltage	V_R	20	Volts
Forward Power Dissipation @ $T_A = 25^\circ\text{C}$ Derate above 25°C	P_F	400 4.0	mW mW/ $^\circ\text{C}$
Junction Temperature	T_J	+125	$^\circ\text{C}$
Storage Temperature Range	T_{stg}	-55 to +150	$^\circ\text{C}$

ELECTRICAL CHARACTERISTICS ($T_A = 25^\circ\text{C}$ unless otherwise noted)

Characteristic	Symbol	Min	Typ	Max	Unit
Reverse Breakdown Voltage ($I_R = 10 \mu\text{A}$)	$V_{(BR)R}$	20	—	—	Volts
Diode Capacitance ($V_R = 15 \text{ Vdc}$, $f = 1.0 \text{ MHz}$)	C_T	—	1.3	2.0	pF
Series Resistance (Figure 5) ($I_F = 10 \text{ mA}$)	R_S	—	0.7	0.85	Ohms
Reverse Leakage Current ($V_R = 15 \text{ Vdc}$)	I_R	—	—	0.1	μA

Figure 2. Motorola's MPN3404 silicon PIN diode.

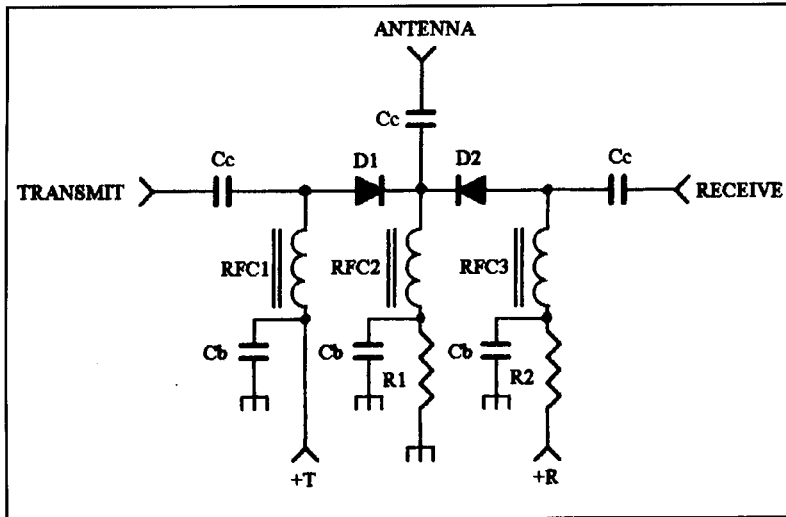


Figure 3. Typical PIN-diode T/R switch using the MPN3404 (D1) and MPN3700 (D2).

transmit side of a PIN-diode T/R switch for example—it takes progressively more forward bias to keep the PIN junction in conduction on RF-waveform peaks. If the diode drops out of conduction due to insufficient bias, some *very nasty* things begin to happen. First, harmonic and spurious emissions rise rapidly—due to rectification and mixing in the diode's PIN junction. Second, saturation products that are *not* transmitted as spurs dissipate as thermal energy, causing the device to overheat. Third, VSWR increases, possibly to the point of destabilizing the PA. Many a frustrated RF designer (myself included) has torn apart perfectly healthy RF strips to ferret out spurs and harmonics that were actually emanating from the radio's PIN diode T/R switch!

The ultimate power-handling capability of a PIN diode is determined by its Total Dissipation Rating. The diode dissipates energy from two sources; RF conduction loss, and IR drop in the DC bias circuit. To measure RF loss, replace the PIN diode with a good-quality coupling capacitor and note any increase in power output on a sensitive wattmeter. To calculate IR dissipation in the bias circuit, multiply forward bias current times the voltage drop across the device ($P = IE$). Add the two losses together, and you'll have a good estimate of how much power your diode must shed as heat. In practice, I've found the MPN3404 will comfortably switch up to 5 watts RF into a well-matched load at VHF frequencies. This requires a forward DC bias level of 60 mA—about maximum for the device in intermittent service.

A practical PIN-diode T/R switch

The diagram in **Figure 3** illustrates a simple PIN-diode SPDT T/R switch using the MPN3700 and MPN3404. Molded RF-chokes

RFC1-RFC3 and bypass capacitors, C_b , isolate the RF line from DC bias circuitry. DC-blocking capacitors, C_c , prevent bias voltage from exiting the various RF ports. Actual choke and capacitor values are selected for optimum effectiveness at the intended frequency of operation. When +R voltage is applied, D2 (the MPN3700) is biased on, opening a low-resistance RF path from the antenna to the receiver port. Bias current for D2 is set by the combined resistances of R1 and R2. During receive, D1 (the MPN3404) appears as a very small capacitance in parallel with the antenna line. When +T voltage is applied, D1 becomes biased on, opening a low-resistance path from the transmitter to the antenna port. Bias current for D1 is set by R1 alone. D2 appears as a very small-value coupling capacitor to the receiver port—too small to provide significant energy transfer. For 13 volts at +T and +R, $R_1 = 220$ ohms and $R_2 = 1$ k. Maximum RF power level is 5 watts.

Many other T/R switching configurations are possible with these devices. Two diodes may be operated in parallel at higher power although there is some loss in isolation at higher frequencies. If more receiver isolation is needed, a second PIN switch can be added on the receiver side to provide a shunt path to RF ground. In addition to T/R switching, the MPN3700 and MPN3404 are useful for a host of other applications. As mentioned, recent articles advocate using PIN diodes in bandpass filter boards to replace low-cost PN switching diodes. PIN diodes are also useful in a wide variety of other passive RF circuits, including RF attenuators and phase shifters. Once you start using them, you may become hooked!

The bottom line

The Motorola MPN3404 and MPN3700 have been on the market for some time, so you should have little difficulty finding them. Experimenters may order in small quantities from Newark Electronics at \$.98 each for the MPN3700 and \$.62 each for the MPN3404. Large-quantity pricing from Motorola-authorized distributors runs around \$.78 per thousand for the MPN3700 and \$.42 per thousand for the MPN3404. Give them a try. If you come up with any interesting applications, *Communications Quarterly* would be happy to share it with our readers.

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2. Ulrich Rohde, KA2WEU. "Key Components of Modern Receiver Design, Part 3. QST, July 1994, pages 42-45.

A LOW-POWER 20-METER TRANSCEIVER

Attention homebrewers! Here's one man's experimental transceiver.

I began this project in January 1989 as an experiment to construct a 20-meter SSB/CW receiver. From time to time, I made alterations and additions. The final result is the transceiver presented here. The completed unit is shown in **Photo A**.

The original 5-1/4 inch wide enclosure remains the same—except for the addition of the “penthouse” to accommodate the transmit final and driver amplifiers. This alteration isolated these heat-producing components from the VFO, aiding long-term frequency stability.

Making the circuit boards

The original “receive only” circuit board now includes the transmit modulator, mixer, and low-level power circuits (see **Photo B**) needed for transceiver operation. I use the term “circuit board” because its electrical traces were routed to design specifications using a high-speed rotary tool and dental burr, rather than an etching process. This fabrication method left most of the original copper on the board, making the later addition of the transmit circuitry easy.

I have used this method of fabricating circuit boards for many years. It is fast, clean, and permits alterations and additions as desired. With a little practice, electrical connections for components such as dual-in-line integrated circuits with as little as 0.1-inch mounting centers can be readily performed. (Caution: The burrs available



Photo A. Completed low-power 20-meter transceiver.

from the rotary tool manufacturers do not retain their sharpness and their shanks are too long to enable precision manipulation of the tool. Dental burrs are far superior for this application.)

I don't have any formal lab performance numbers for my transceiver. However, based on observations made over three years of “on the air” use, I offer the following subjective measurements. So far, I've made two-way contacts with 35 states, 4 Canadian provinces, and several European and Central American countries. My operation followed a seasonal pattern as I alternated residences between Chicago and near the Inland Waterway south of Clearwater, Florida. Competition from nearby amateur

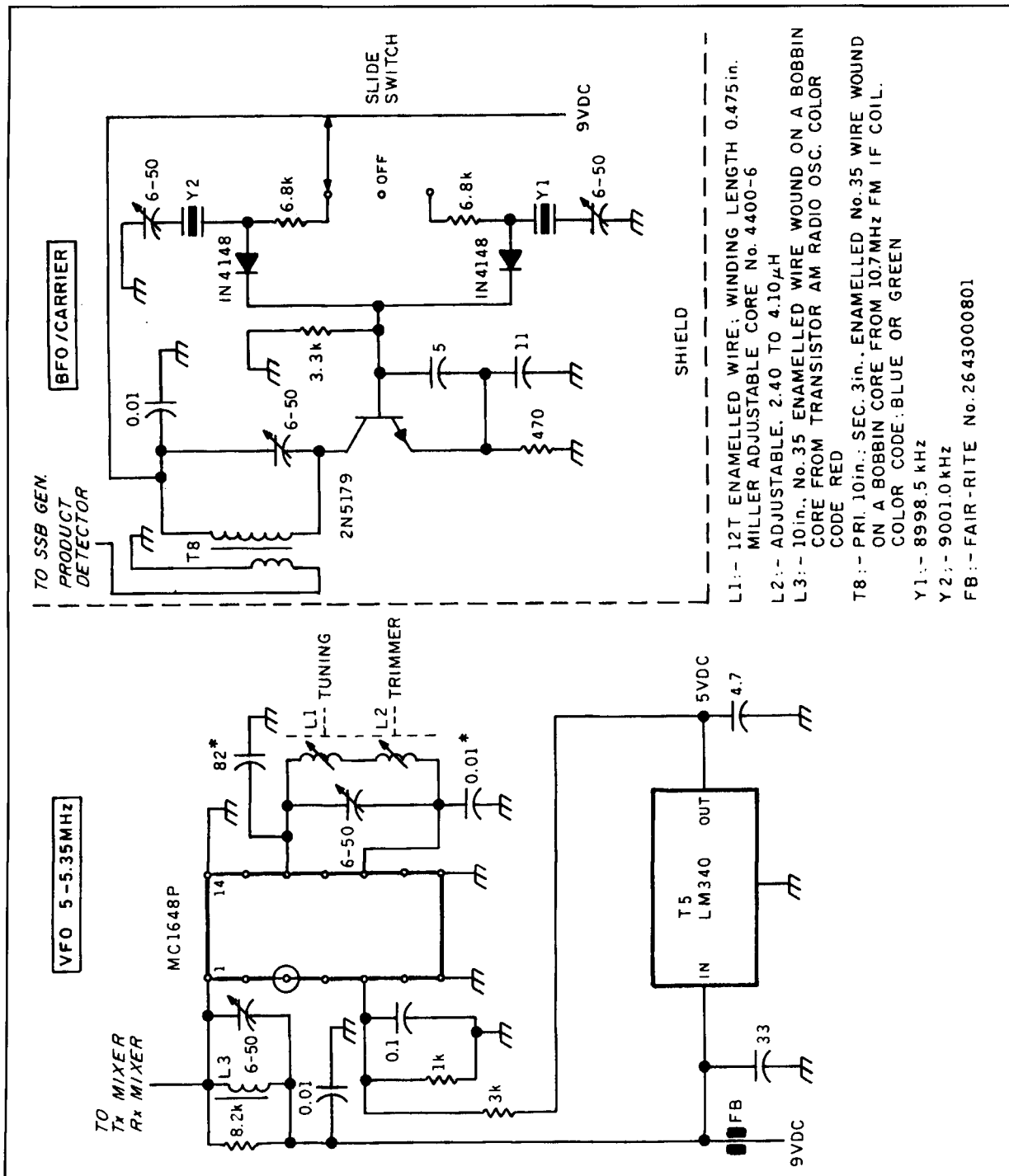


Figure 2. Circuit diagrams of the (A) VFO and (B) BFO/carrier. (See text for temperature considerations.)

radio stations, as well as from a myriad of commercial RF polluters, provided an ideal test bed to put the homemade transceiver through the "wringer."

The receiver noise floor is low enough to permit detection of atmospheric noise with the antenna connected. Desensing or reciprocal mixing problems have rarely been observed—and certainly not from signals 20 kHz removed from the passband. Stability is good. Only

slight retuning is required at 20-minute intervals after turn-on, with complete stabilization in about 60 minutes. No spurious receive signals, birdies, or image responses were detected.

Transmitter output power is 6 watts on CW, and 6 watts peak on SSB. Carrier leakage is about 250 μW, or at least -43 dB referenced to the peak output. Lacking a spectrum analyzer, I was unable to measure the harmonic or spurious content of the transmitter; however, no har-

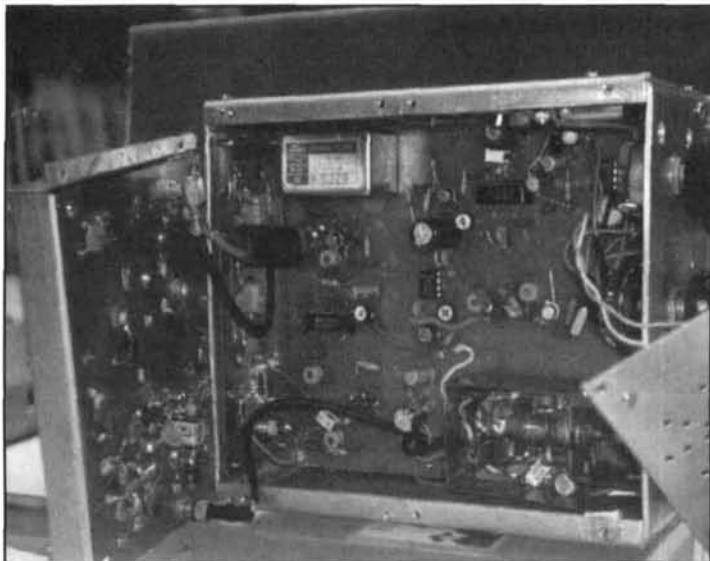


Photo B. Top view of the main circuit board, the bottom of the RF driver, and the final amplifier.

monics or spurious signals are detectable while viewing my shack TV set.

Receiver signal path

A MRF901 serves as the RF amplifier. Although the use of such a high-gain UHF device may at first seem unorthodox at 14 MHz, note that the gain is reduced to a point where the system noise figure is just below that detected on a 20-meter antenna. Transformers T9 and T10 (Figure 1) provide impedance matching and selectivity. The MRF901 is under AGC control.

First mixer uses MECL device

The first mixer is a MECL device (more on

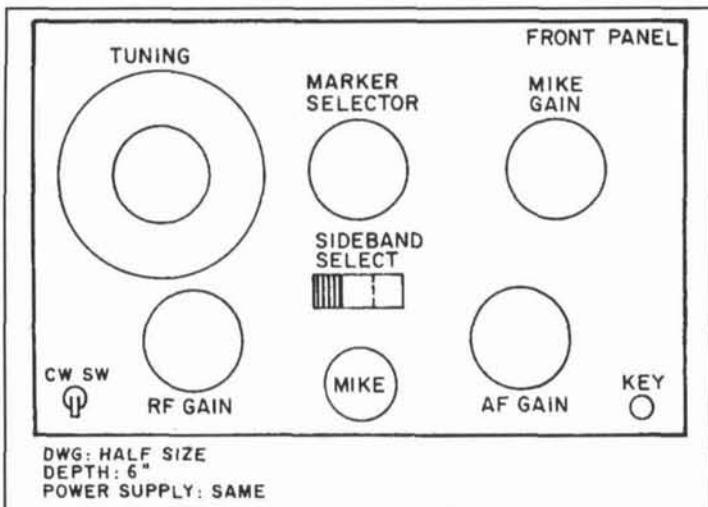


Figure 3. Diagram of the transceiver's front panel. RF gain control was deleted in final design.

this later), and is followed by a 2N5179 IF preamp stage. Transformer T11 is tuned to the IF frequency of 9 MHz. The 2N5179 is operated in a common base configuration. The 330-ohm collector resistor sets the terminating impedance for the crystal filter, and is also under AGC control. A 2N2369 IF buffer/post IF amplifier stage follows. The IF amplifier and product detector is accomplished via a CA2111A IC. More on this stage later.

AGC is audio derived

An automatic gain leveling amplifier protects the receive mixer from widely varying antenna signal levels. Similar protection is provided after the mixer to the crystal filter, IF system, and product detector by a second leveling amplifier working in tandem with the first. AGC is audio derived from the output of the CA2111A product detector. A 2N2222 audio amp drives a voltage doubler rectifier, the output of which is filtered and, in turn, drives a 2N2222 DC amplifier. The DC amplifier provides AGC control voltage to the MRF901 RF stage and 2N5179 IF preamp. At one point during development, I included an RF gain adjustment, but I eliminated it because it offered no improvement in reception. Calibration accuracy is enhanced with panel selectable 5, 10, and 50-kHz markers derived from a 100-kHz standard. The 9-MHz SSB crystal filter, VFO, and BFO/carrier oscillators are common to the receive and transmit circuits. The SSB crystal filter is diode switched.

MC12002 notes

The MC12002 is an active mixer similar to the LM1496. It is an exceptional device and easy to use. Unlike the LM/MC1496, all bias levels are set internally, resulting in a low external parts count. It may be used as a double-balanced mixer to at least 100 MHz; it may also serve as a suppressed carrier modulator, product detector, or frequency doubler. Two of these devices are used in this rig—one as the receiver mixer, the other as the exciter DSB generator. A common emitter amplifier, included on one of the ports of the MC12002, may be used to amplify either the local oscillator or the signal source as desired. The MC12002 common emitter amplifier in the receiver mixer is used at the 14-MHz signal input. The MC12002 DSB mixer uses the common emitter stage for mike AF amplification.

LO drive levels from the VFO and carrier oscillators are adequate and require no further amplification. A 50-ohm potentiometer, added

Color code	Frequency range
Black;yellow	400 kHz to 600 kHz
Red	600 kHz to 6 MHz
Blue;green	6 MHz to 100 MHz

Table 1. The useful frequency range of the salvaged ferrite bobbin cores is indicated by the color codes found on the adjustment cups of the IF transformers listed here.

across pins 5 and 6 of the MC12002 DSB generator, will permit further carrier reduction if desired. This was not needed in my unit.

IF stage and product detector

The RCA CA2111AE (Motorola MC1357) is a broadband 3-stage, 60-dB limiting amplifier with a quadrature detector and emitter coupled audio output. While intended for FM receiver applications, I have found that it performs well for SSB/CW reception if one applies a low-level carrier (BFO) signal to pin 11 for product detection. Pin 11 provides direct access to the quadrature detector circuitry of the device. The exact BFO drive level is set via the variable 2.7 to 10 pF coupling capacitor shown in **Figure 1**.

Two precautions must be taken when using this circuit: The input circuit (pins 4, 5, and 6) must be treated with respect because of the overall high gain of the device. Also, the quadrature detector will not tolerate excessive BFO voltage at pin 11; in this situation, distortion will occur.

Additional gain can be obtained by moving the 4.7-pF capacitor from pin 9 to pin 10, but overloading on strong signals may result. The LM386 low-power audio amplifier works well with the CA2111AE. The layout I used is almost identical to that recommended by the manufacturer. One modification I found desirable was the addition of a 0.05- μ F capacitor across the LM386 speaker output to the ground plane.

Transmitter summary

Mike audio is amplified by a fixed gain LH101 IC audio amp. Amplified mike audio is applied to one input of a MECL MC12002 active mixer, while the other input is provided from the 9-MHz BFO. A 9-MHz DSB signal results. The DSB signal is passed through the 9-MHz SSB filter (diode switched and shared with the receiver) where the unwanted sideband is removed. A NE602 mixer follows the SSB filter, and combined with the 5-MHz VFO signal provides the desired 20-meter signal. A two-stage tuned amplifier follows, reducing the frequency, and bringing the desired 14-MHz

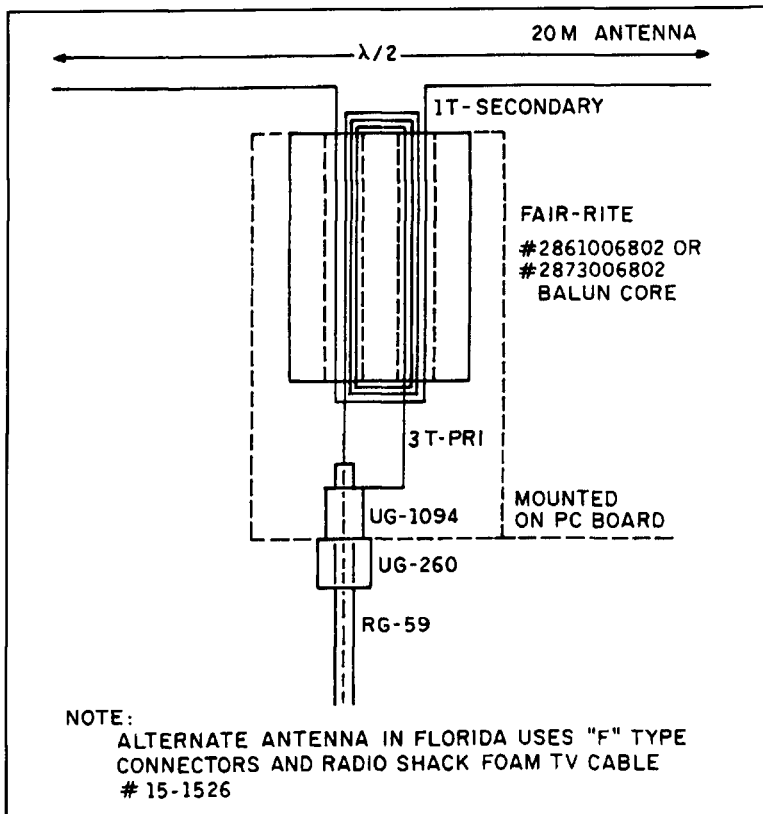


Figure 4. The author's 20-meter dipole antenna.

SSB signal to a level capable of driving the two-stage driver and PA section to an output of 6 watts peak. A half-wave harmonic filter follows the PA stage.

The heart of the VFO, the MC1648

The MC1648 integrated circuit is a 2-terminal voltage-controlled oscillator. Suitable LC circuits will allow a frequency range from 50 kHz to more than 200 MHz. As a result of the saturated nature of the waveform, output amplitude is almost constant regardless of frequency. When used as a VFO, the internal AGC bias of the device is altered by applying an external voltage to pin 5. This changes the square wave to a simulated sine wave. An LC parallel-tuned circuit at the output provides additional filtering of the waveform. Varactor tuning for narrow-band service, such as that required by SSB/CW, isn't recommended. Internal noise generated in a varactor diode translates into an uncontrolled frequency shift that becomes progressively worse in direct relation to frequency. At 5 MHz, the shift translates to a delta swing of about 50 cps. Accordingly, I use variable L or C tuning with the MC1648. Several modules have been constructed for VFO service, as well as

calibrated test generators for frequencies from below 50 kHz to somewhat over 260 MHz.

VFO is inductively tuned

The VFO is inductively tuned (see **Figure 2A**). I avoided air dielectric trimmers and tuning capacitors in order to reduce the possibility of microphonics. Of course, frequency stability with temperature change is most important. There are several ways in which this matter can be addressed. First, the VFO should be fabricated into a separate housing assembly and mounted directly to the front panel. (Layout of the front panel is illustrated in **Figure 3**.) It should also be shielded from incidental air movements and heat producing components such as power transistors, bias resistors, and the like. The MC1648P operates from 5 volts. A LM340 voltage regulator provides regulated 5 volts for the device. Note that the LM340 is a heat-producing device, and should be mounted away from frequency-determining components in the VFO housing.

"Fine tuning" VFO drift

Remaining drift can probably be reduced by focusing on capacitors connected to pins 10 and 12 of the MC1648. Mica or polystyrene dielectric capacitors are recommended. Various combinations of NPO and negative temperature disc ceramics should be tried for the 82-pF capacitor until reasonable long-term thermal stability is obtained for the VFO environment.

Simple tuning mechanism

The tuning mechanism is quite simple. A 1/4-inch diameter brass shaft is coupled to the threaded lead screw of coil L1 in **Figure 2A**, and brought out through the front panel. A knob, numerically inscribed with figures from 0 to 100 over 360 degrees, completes the assembly. Covering from 14 to 14.35 MHz requires 8.76 turns of the knob—or somewhat less than an average of 400 cps per dial division. The 1/4-inch tuning shaft should be allowed to "play" in and out of the panel bushing as the tuning slug is adjusted.

Notes on the NE602

The NE602, manufactured by Phillips (formerly Signetics), is a Gilbert Cell derived mixer designed for cellular telephone service. However, it is also useful for a variety of mixer type applications. An internal bias network eliminates the need for external parts. This device has excellent high frequency character-

istics. I have used it as a down converter to 3 GHz. I make good use of this device in the transceiver to convert the 9-MHz SSB signal to 14-MHz.

Bobbin coil forms

One of the most satisfying aspects of this project was the opportunity to put ferrite bobbin cores—salvaged from IF transformers found in discarded AM/FM radio sets—to good use. The average AM/FM radio will yield 5 or 6 of these cores, possibly more. The useful frequency range of these bobbins is indicated by the color codes found on the adjustment cups of the IF transformers listed in **Table 1**.

Winding jig aids assembly

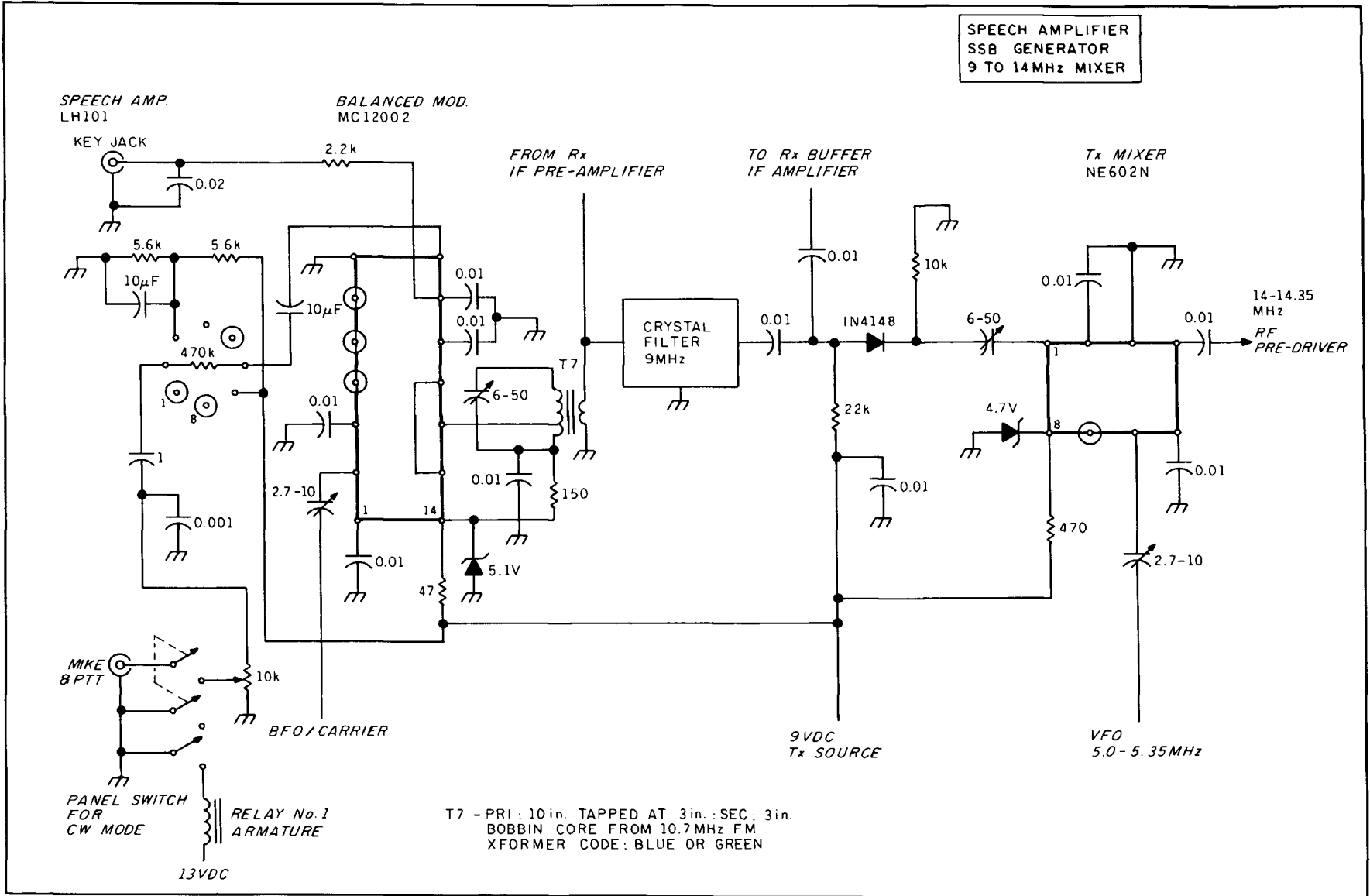
Winding wire on these diminutive forms is somewhat awkward, but it can be done with a little patience and preparation. I use a simple tool consisting of a spring-loaded wooden clothespin with "V" grooves cut into the opposing jaws to grasp one of the flanges while the winding progresses. Don't be concerned with neatness; scrambled windings work fine. However, all windings should be made in the same direction on each form and the indicated start/end sequence should be maintained per the schematic from which you are working. I find these bobbins much easier to wind than toroidal cores. They exhibit the same self-shielding benefits, and the coefficient of coupling of windings appears to be better. These cores are not suitable for power applications because they will saturate at comparatively low values of direct current flow or RF.

Coils and transformers made with these cores are so small and light that they can be secured by their leads. A drop of candle wax will also seal these assemblies and hold them firmly in place. Just image an RF coil for 160 meters that's so small you hardly notice it on the circuit board!

Alignment procedure, start with the VFO

Alignment can be performed without a signal generator; but, you will need a frequency counter or calibrated, general coverage receiver. For the VFO, begin by turning the knob fully counterclockwise, until the shoulder of the iron core just contacts the coil form bushing. Now, position the coil on the form so the leading edge of the core just enters the coil. Then spread the turns of the coil evenly so that with

Figure 5. Speech amplifier, SSB generator, and TX mixer.



Initial BFO settings

For the BFO/carrier alignment, set the crystal slide switch to activate the 8998.5-kHz crystal circuit. Slowly adjust the trimmer capacitor associated with transformer T8 until oscillation

occurs. No additional adjustments are made at this time.

Look for regenerative conditions

With both the VFO and BFO activated, the

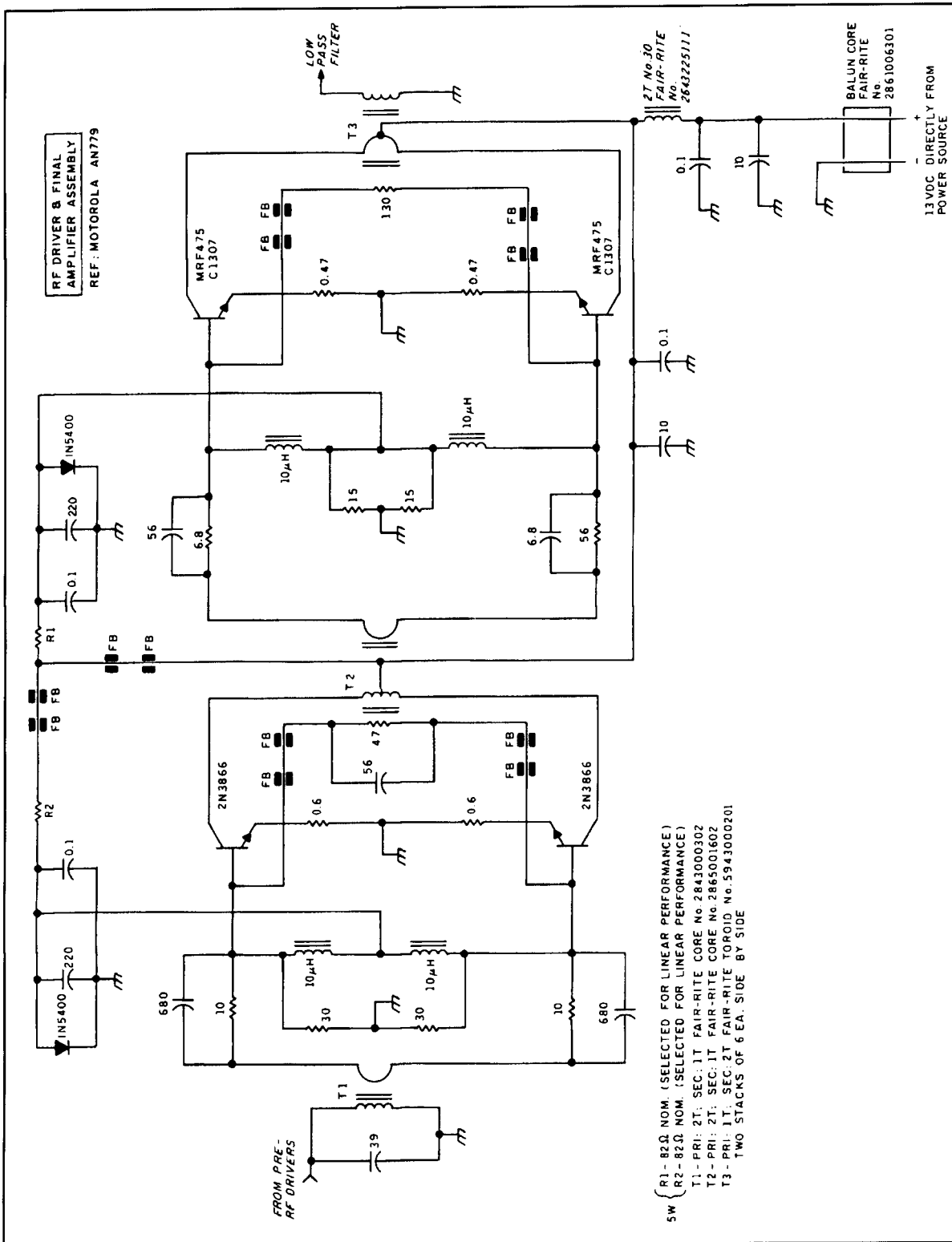


Figure 7. RF driver and final amplifier assembly.

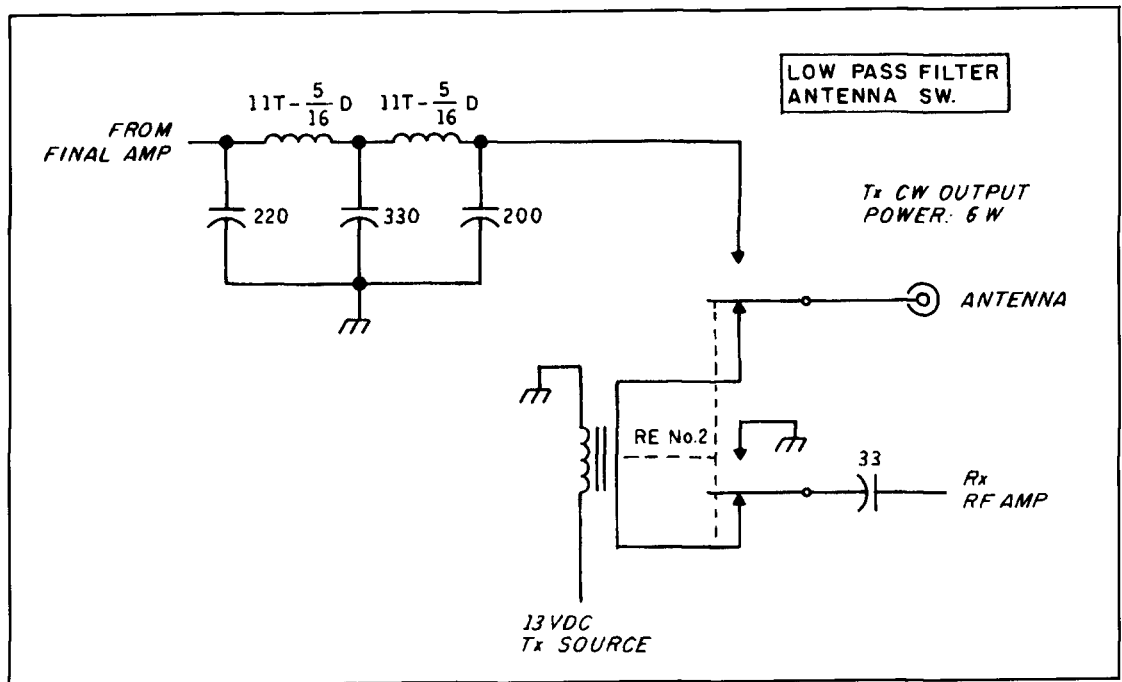


Figure 8. Low-pass filter and antenna switch.

antenna (shown in **Figure 4**) connected, and the audio gain control advanced, noise or signals should be audible. Peak transformers T9, T10, T11, and coil L4 for maximum output. If instability is noted, increase the value of the 750-ohm resistor in the emitter common ground return path (2N2222 collector to ground) for the MRF901 and 2N5179 stages. Tune the VFO to midrange and repeak all circuits for maximum noise. With the antenna disconnected, tune the entire 20-meter band and note if any sharp noise peaks are found. Such peaks would indicate a regenerative condition that must be eliminated, otherwise system intermodulation distortion will result—especially when strong signals are present. No additional operating benefits will be obtained by increasing the gain of the MRF901 beyond the point at which it establishes the noise figure for the system. Note that additional DC decoupling is provided by supplying the base, emitter, and collector of the MRF901 from three different regulated voltage sources. Likewise, the voltage source for the mixer is separate from that of the base of the RF transistor.

Final BFO alignment

The final alignment of the BFO consists of selecting quartz crystal Y2 (9001.0 kHz). The trimmer associated with coil T8 should be readjusted, if necessary, to initiate oscillation. The trimmer must be adjusted so oscillation occurs as each crystal is selected, without additional trimmer adjustment. The BFO oscillator should

start reliably, with either crystal selected, when power is applied and removed. Carrier frequency adjustment may be performed while listening to a moderate strength, unkeyed signal. Each crystal series trimmer capacitor is adjusted until the appropriate upper or lower sideband is eliminated or reduced to an acceptable level. The shape factor of the crystal filter will determine the exact settings for Y1 and Y2. (Editor's note: The normal setting of the BFO crystal is at a point -20 dB down on the filter skirts.)

Care must be taken to prevent stray BFO/carrier energy from entering the receiver IF input, transmit modulator output, and subsequent converter/power amplifier RF circuits. Use additional shielding as necessary.

Heart of the transmitter

The speech amplifier, SSB generator, and TX mixer are shown in **Figure 5**. In this assembly, the microphone audio is amplified by a LH101 operational amplifier and then mixed with the 9-MHz BFO carrier in the MC12002 to produce a DSB signal. The subsequent 9-MHz SSB filter removes the unwanted sideband, and a NE602 mixer combines this SSB signal with the VFO to produce 14-MHz.

CW operation

CW operation is accomplished by unbalancing the MC12002 transmitter mixer by grounding pin 9 through a 2.2-k resistor, permitting

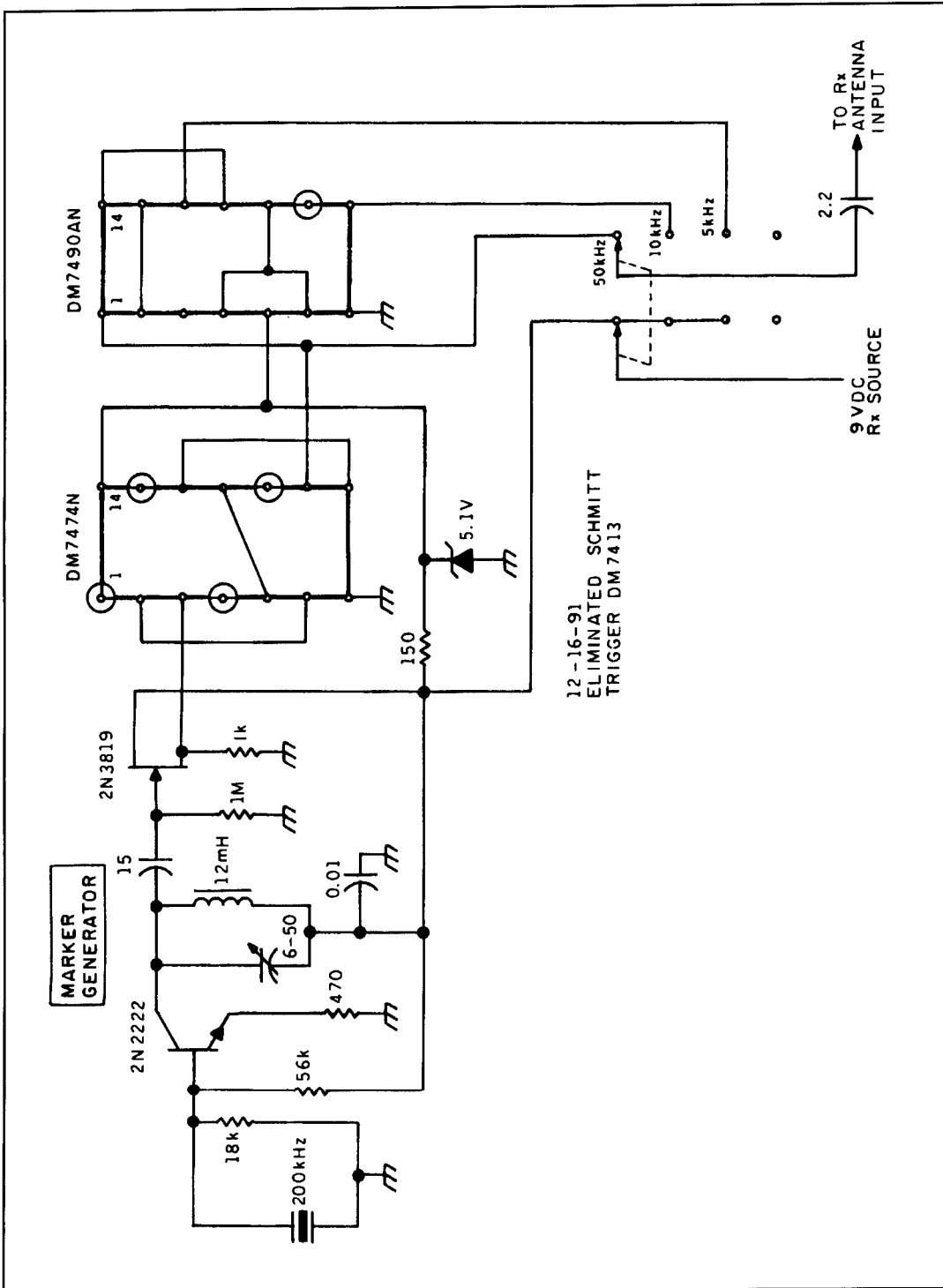


Figure 9. Marker generator schematic.

the 9-MHz signal to pass through to the mixer output. The CW key is connected in series with this 2.2-k resistor, and grounds it on contact closure. Transformer T7 is peaked approximately midway between 8998.5 and 9001.0 kHz. The VFO and BFO/carrier frequencies should not shift more than a few cycles between receive and transmit. These oscillators

should be checked with a frequency counter for unwanted "pulling."

Do not overdrive the BFO carrier injection applied to pin 2 of the MC12002 DSB modulator. This is set by a 2.7 to 10-pF trimmer capacitor. (With the key jack open, and no voice modulation, final power output must be a bare minimum.) Also, be sure that the NE602 mixer stage

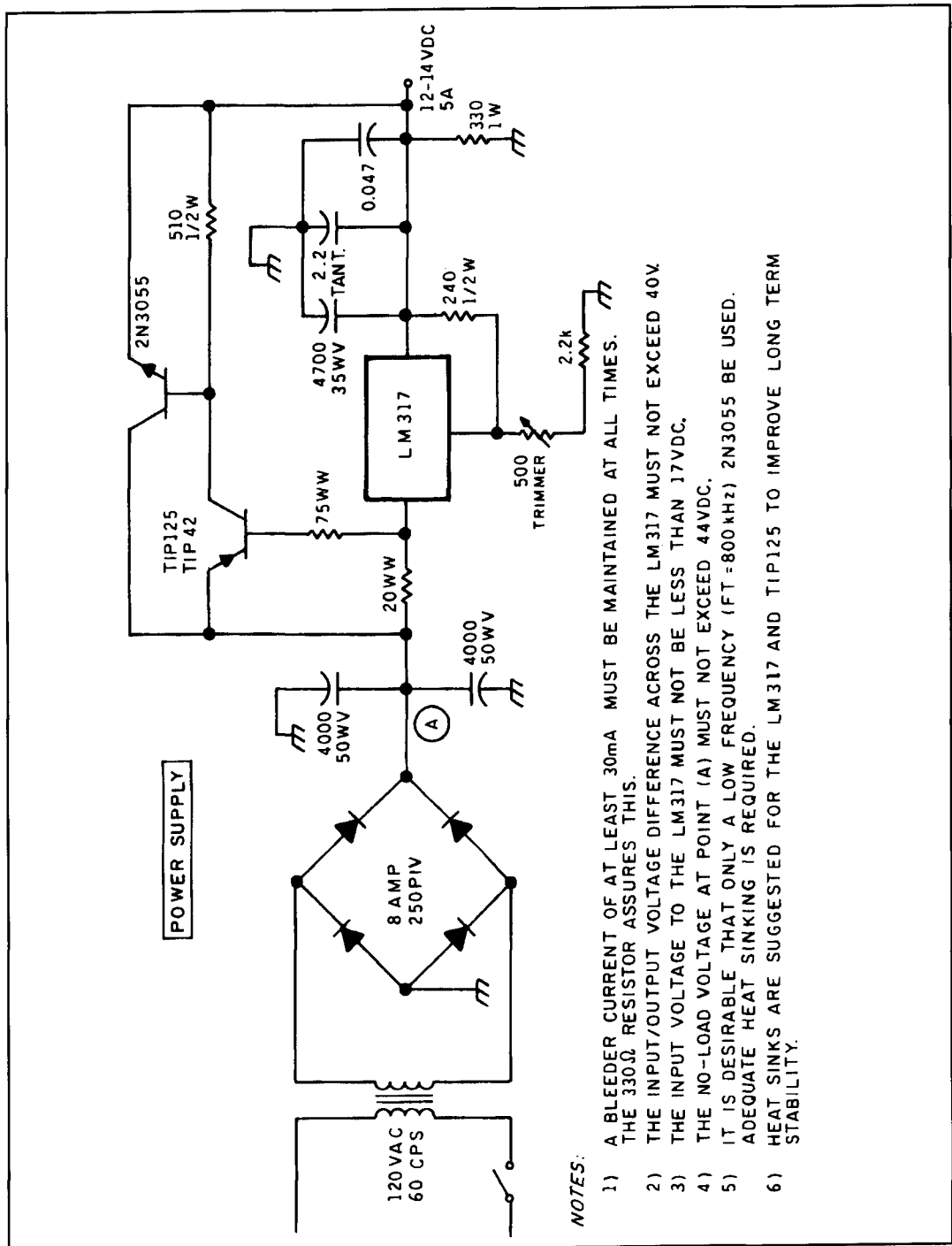


Figure 10. Power supply diagram.

isn't overdriven. Look for signs of flat-topping on the outputs of the MC12002 and NE602 mixer stages using a good oscilloscope. The drivers and PA stages may have to be shut off to observe the low level signals at these points and to eliminate stray pickup by the scope probe. The VFO injection level for the NE602 is set by a series 2.7 to 10 pF capacitor on pin 6.

Pin 4 of the NE602 transmitter mixer is extremely vulnerable to unwanted feedback from the large amount of in-line gain and high

signal levels presented by the transmitter power chain. I suggest DC decoupling of common source lines as shown in **Figures 6** and **7**. Good layout and shielding are a must.

Tuned stages are used in predriver

The transmitter predriver amplifiers are shown in **Figure 6**. Transformers T4, T5, and

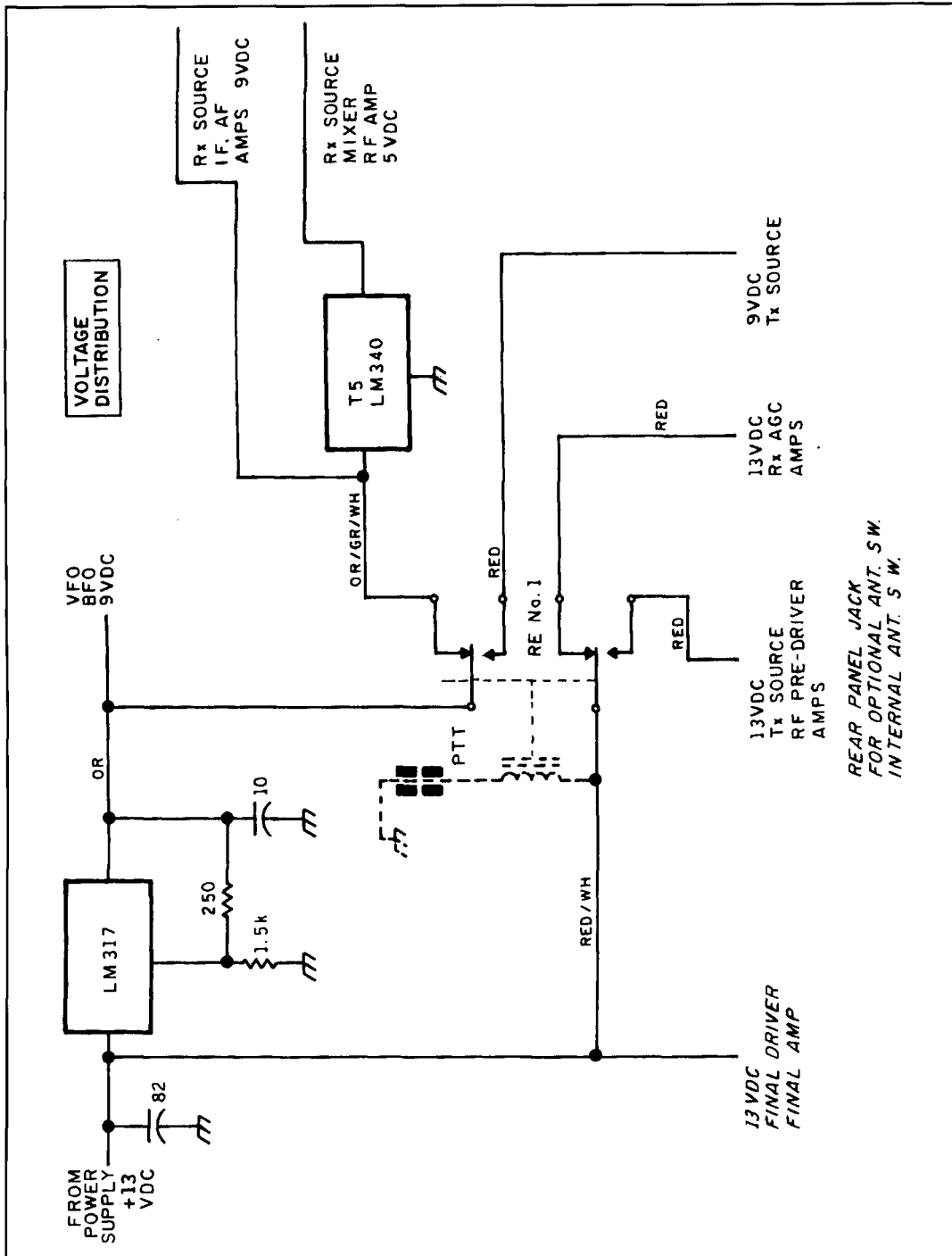


Figure 11. Voltage distribution diagram.

T6 should be stagger-tuned to produce constant output across the 20-meter band. Transformers T4 through T6 provide rejection of the transmitter image frequency at 7 MHz. Transistor 2N5109 operates as a Class "A" amplifier. The 2N3866 also operates in a linear mode. Accordingly, the bias resistor (R3) may need to be changed from the nominal value of 82 ohms to achieve best linearity. Heat sinks are required for both transistors. Linearity can be tested by using a two-tone test oscillator and

oscilloscope. Refer to *The ARRL Handbook* for details on this procedure.

Push-pull stages are used in driver and PA

The RF driver and final amplifier stages are shown in **Figure 7**. This is a broadband linear amplifier assembly; no tuning is required. The 82-ohm bias resistors may need to be altered

for best linearity due to gain tolerances allowed during manufacture of the transistors used. Heat sinks are necessary for both sets of transistors. Note the amplifiers are run in a push-pull configuration, providing improved even-order harmonic suppression and easing the second harmonic rejection burden placed on the output harmonic filter.

I've been using an identical assembly for the past 10 years as the driver for a pair of Motorola MRF454 power transistors running a peak input of about 250 watts. For more information, see Motorola application notes AN762 and AN779.

LP filter and antenna relay

The low-pass filter and antenna switch are shown in **Figure 8**. This half-wave low-pass filter has a cutoff frequency of about 16 MHz. A Radio Shack no. 275-249 DPDT relay serves well as the antenna relay (RE2). The marker generator is shown in **Figure 9**.

Power supply

My base station power supply is shown in **Figure 10**. Note that there is no short circuit production or current limiting provided by this simple design! I have built several supplies using this basic circuit as a model, and all have performed completely to my satisfaction. However, care must be taken when using this circuit. Accidental shorts will surely result in device failure. A fast-acting 6-amp fuse should be used in line with the 13.8 volts DC output.

Voltage distribution for the various assemblies of the transceiver are covered in **Figure 11**. Note the LM317 regulator IC provides 9 volts for the VFO and BFO circuits, as well as various RX and TX circuits. The regulated 9 volts is also fed to T5—an LM340 regulator, which provides 5 volts for the MC12002 RX mixer. (The 5 volts for the MC12002 DSB generator is produced via a 5-volt Zener.) This figure is included as a convenience—to facilitate repairs or alterations, and for clarification of an important point. In this design, the push-to-talk (PTT) switch on the handheld microphone controls transfer of voltages between the receiver and transmitter sections via the contacts of DPDT relay RE1 (another Radio Shack no. 275-249 relay). The 13.8 volts for the driver and final amplifier stages is fed direct and is not switched.

Postscript

No circuit board drawings are available for this transceiver. The boards have undergone constant revisions over the years, and continued experimentation will undoubtedly yield more changes as time passes. My transceiver was basically conceived of as a one-of-a-kind rig, and it has never been my intent to present it for others to build. Rather, the intent of this article is to stimulate amateur construction on a level above that of simply making a carbon copy clone of the material presented. I assume however, that anyone attempting this circuit has experience in troubleshooting and aligning QRP SSB transceivers. ■

PRODUCT INFORMATION

Hex-Wik Desoldering Braid

The Tool Resource has made available the Hex-Wik desoldering braid. The Hex-Wik desoldering braid works quickly while safely desoldering components from PC boards and terminals. The braid removes the solder so that the desoldering joint is left clean and ready for resoldering.



For more information contact The Tool Resource, P.O. Box 1106, W. Dundee, IL 60118 (Phone/Fax: 708-468-0849).

New Hameg HM8042 Curve Tracer

Hameg Instruments, Inc. has introduced the HM8042 Curve Tracer. This is a module that plugs into Hameg's HM8001-2 Mainframe Power Supply and is used in conjunction with any oscilloscope that has X-Y display capabilities. It can test diodes, transistors, FETs and thyristors. The instrument is microprocessor based, and measures and displays characteristic curves of semiconductor devices. Digitized data is used to generate a 5-curve, flicker-free X-Y oscilloscope display. A 4-digit LED numeric readout indicates parametric data.

For additional information, call Hameg toll-free at 800-247-1241.

Antenna dreams for snowbound hams

I know it's February, and there's still too much ice on the roof to climb up and check on how your antennas have fared during the winter thus far. However, it's never too early to begin planning this year's antenna farm.

In this installment of "Tech Notes," you'll read about a triband antenna, coaxial cable traps, a VHF/UHF combiner for mobile use, and a direction-finding method. Enjoy!

*Peter Bertini, K1ZJH
Senior Editor*

Triband Dipole

*Reprinted from Technical Abstracts
Gil Sones, VK3AUI, Editor
Amateur Radio
May 1994*

A simple triband dipole was described in the Swiss magazine *Old Man* for February 1994 by Robert Kägi, HB9KL. The dipole consists of parallel 80 and 40 meter dipoles with 15-meter operation using the 40-meter dipole as three half waves.

Of interest is the shortening of the 80-meter dipole by folding back the ends. Some adjustment of the length would be required in VK as our band is different. The 80-meter section being resonated at 3.7 MHz would require some extra length to lower the frequency for local use.

The spacers used were plastic and some care would be needed to select a plastic which is not too badly affected by ultra-violet light. An alternative would be to use dowel or canes, possibly with plastic insulator tips. The pipes used in garden watering systems may provide suitable materials.

The antenna was built with 400 mm spacers. The center was supported 10 meters off the ground and it was hung as an inverted Vee with an included angle of 115 degrees. This means the ends were approximately 2 meters off the ground.

The total length of the 80-meter section was 42.48 meters for a frequency of 3.7 MHz. The length of the 40-meter section was 20.18 meters for a frequency of 7.05 MHz.

To adjust lower in the band, lengthen the 80-meter section by around 2 meters and prune to get a suitable resonance point in our band. This will only be around a meter overall as the ends are folded.

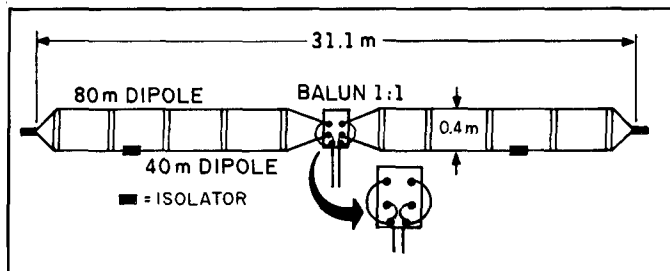


Figure 1. Triband dipole.

A bandwidth of 142 kHz was obtained on 80 meters and a bandwidth of 255 kHz on 40 meters. The bandwidths on both bands were for an SWR of 2:1. At resonance an SWR of 1.05 was obtained at 3.7 MHz and an SWR of 1.1 was obtained at 7.05 MHz. A 1:1 balun was used at the feedpoint and one of the current balun designs that uses ferrite beads would be suitable.

Figures were not given for 15 meters, but the bandwidth should be reasonable. Some compromise may be needed between the 40-meter center frequency and the 15-meter center frequency, but this should be fairly easy.

The antenna is shown in **Figure 1**.

Coaxial Cable Traps—In Search of the Perfect Antenna

*Paul Duff, VK2GUT
Reprinted from Amateur Radio
October 1993*

Ever hear the one about the amateur who pulled down all his antennas and never got around to putting them up again? About two years ago, I got fed up with trying to get out on 80 meters with a dipole 20 feet off the ground and no decent earth. A change of QTH seemed the perfect chance to look at my options, especially as the new site, a small semi in the suburbs, had very little ground space. What I really wanted was an antenna which was easily built, easily mounted and demounted, easily matched, had broad multiband function, and an omnidirectional pattern of radiation. It also had to be portable.

See why I never got back on the air?

Last year, however, I got my full call and, fired with enthusiasm, I tackled the problem afresh. A multi-trapped vertical, chimney mounted and fed with open-wire line via a transmatch, such as the Z Match, seemed to fill most of my criteria except for the broadband response

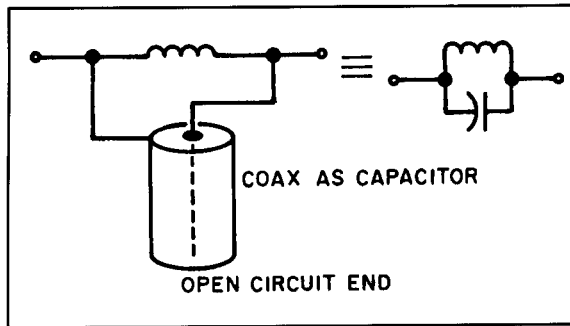


Figure 1. Basic coax trap.

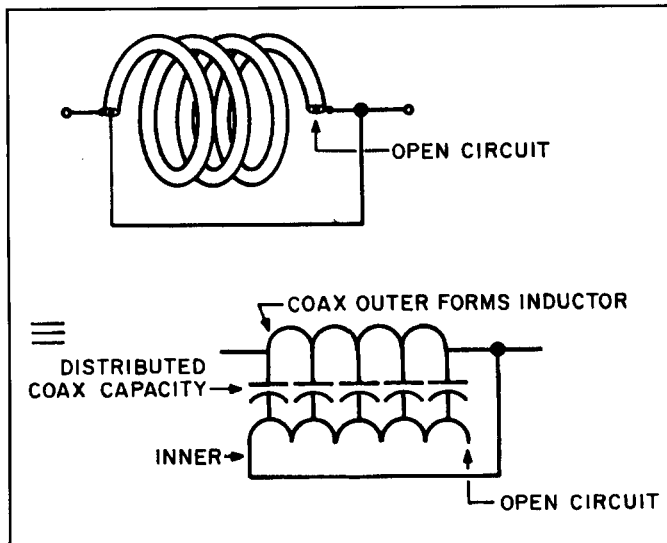


Figure 2. Coax also forms coil.

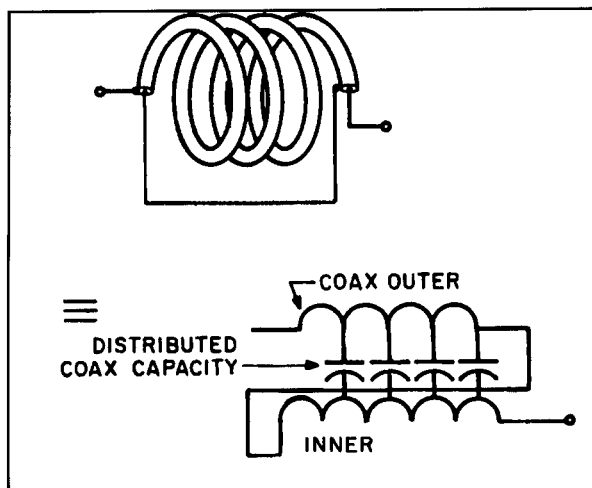


Figure 3. Inner plus outer in series.

and ease of construction. Every design that I came across seemed to be a mass of high-voltage capacitors, difficult to tune coils, and hard to shape phenolic spacers. That was when I came across the idea of using coaxial cable in traps.

Theory

The most basic coaxial cable trap uses the coaxial cable as a capacitor, see **Figure 1**. This is simple, but cumbersome, and offers no great advantage over a standard trap.

The idea of using the coaxial cable as both coil and capacitor first appeared in *QST* in May 1981, see **Figure 2**. The shield acts as the coil. The inner conductor, because each turn is shielded from the next, has no more inductance than a straight wire of the same length, but the coaxial cable still maintains its inherent capacitance. The result is a much more elegant trap, but electrically it has little more to offer than the simple trap. The electrical equivalent is shown in **Figure 2**.

In *Ham Radio* for October 1981, Gary O'Neill, N3GO, described a coax trap, **Figure 3**, which looked the same but, in fact, was very different. This I have tried to represent in the circuit diagram. In this trap, the shield and the inner conductor are magnetically linked and the current thus induced means that this circuit has much more inductance than that in **Figure 2**. This current, however, changes the whole phase arrangement of the capacitive element which also consists of the shield and the inner conductor. The result is that the capacitance decreases in direct proportion to the increase in inductance. If the coupling is perfect, which it nearly is, the inductance is increased by four and the capacitance is decreased by four, but the resonant frequency stays the same (refer to *QST*, August 1985).

To follow why these traps are superior, you may refer to the equations in **Table 1**.

The perfect trap would give 100 percent isolation across the band of interest and 100 percent conduction outside that band. An adequate trap, being one which will give an SWR less than 2:1 across the band, must provide at least 7 k of isolation at the band edges. This band-edge impedance is directly proportional to the reactance at resonance of either the coil or the capacitor (see **Equation 1** in **Table 1**). This is, in turn, directly related to the inductance (see **Equation 2** in **Table 1**). But below 30 MHz, the losses from a trap are all from the coil. If these losses are represented as a resistor in parallel with the trap, then a lossless trap will have infinite resistance and a lossy trap will have a low resistance.

Design

Again refer to **Table 1**. If you can synthesize **Equations 3, 4, and 5** to give a value for n for a given value of L , C , D , F_0 , and t , then you are better at mathematics than I am. For those of you like me, a simpler answer is to program

your computer to substitute for values of n using a for/next loop until a given value of f is reached. In writing such a program, two points should be noted. The first is that $D+t$ is used as the coil diameter rather than $D+2t$ so as to give the mean coil diameter—and this works well. The second is that a value of D should be chosen to provide the shortest length of coax possible in the coil as this will reduce losses to their minimum. This occurs when the denominators in **Equation 3** are equal. That is, when $(D+t)/nt = 2.22$. In reality, any value between 2.5 and 1.0 provides an acceptable coax length (see *QST*, December 1984). A BASIC program for the C64 is given in **Table 2**. This should be adaptable to other computers. A set of optimized trap dimensions is given in **Table 3**.

Construction

Construction is easy if you just follow the steps below:

1. Choose a former of suitable diameter white PVC pipe and cut a short length.
2. Drill a hole for the coax near one end of the former.
3. Cut a piece of coax 6 cm longer than the calculated length needed for the coil. Strip 3 cm of cover from each end.
4. Insert one end of the stripped coax into the drilled hole in the former. Tightly coil the insulated length of the coax around the former.
5. Mark the point on the former where the coax insulation finishes. Do not worry too much if this does not exactly correlate with your calculated value of n .
6. Remove the coax and drill a second hole at the marked spot.
7. Separate the exposed braid and conductor back to the insulation. Strip and trim the inner conductor as necessary and then rewind the coax trap. Finally, solder the braid at one end to the conductor at the other.

The result should dip very closely to the required frequency, but if any fine tuning is required this can be done by slightly separating the turns of the coil before applying glue to fix them in place.

Conclusion

The potential uses of such a simple device are endless. My own plans are to build a top-loaded vertical for 80 through 10 meters for under \$60. Multiband dipoles not requiring an ATU, dual or triband beams, or simple dual or triband mobile use would be easy to build. Experiments with miniature coax may extend their use to VHF at least at low power. I can see these traps being the basis of antenna projects for years to come.

$Z = \frac{-j(X_o)}{(F/F_o) - (F_o/F)}$	(Equation 1)
$X_o = 2\pi F_o L$	(Equation 2)
$L = \frac{(D+t)^2 n^2}{18(D+t)+40nt}$	(Equation 3)
$C = \frac{C_o \pi n (D+t)}{12}$	(Equation 4)
$F_o = \frac{1000}{2\pi \sqrt{LC}}$ MHz	(Equation 5)
All lengths in inches (with apologies) as sourced from U.S. articles and books.	
<p>X_o = reactance at resonance of either the inductor or capacitor L = inductance C = capacitance (total) C_o = capacitance of coaxial cable per foot F = operating frequency F_o = resonant frequency Z = tuned circuit impedance n = number of turns (assume close wound) t = thickness of the coax with insulation on D = inner diameter of the coil</p>	

Table 1. Parameter equations.

```

10 PRINT "FORMER DIAMETER" :ENTER D
20 PRINT "COAX THICKNESS IN INCHES" :ENTER T
30 PRINT "COAX CAPACITANCE IN pF PER FOOT" :ENTER Co
40 PRINT "DESIRED FREQUENCY IN MHz" :ENTER X
100 FOR N = 1-500step0.25
200 L = ((D = T)EXP2*nEXP2)/(18*(D*T)) = (40*n*t)
300 C = Co*3.142*n*(D = t)/12
400 f = 1000/2*3.142*(sqr(D = t))
500 IF f<=X THEN GO TO 700
600 NEXT n
700 PRINT "NUMBER OF TURNS EQUALS" :N
800 PRINT "COAX LENGTH EQUALS" :N*(D + t)/2.54:"mm"
900 END

```

Table 2. C64 BASIC program to calculate trap dimensions.

Frequency (MHz)	Former ID (mm)	Former OD (mm)	Turns Tight Wound (n)
3.6	70	75	10.2
7.1	60	64	6.8
14.2	40	44	5.5
21.5	32	36	4.75
28.8	32	36	3.75

Table 3. Dimensions of optimized traps used RG58C/U coax and commonly available PVC formers.

VHF/UHF Combiner for Mobile Use

Ian Keenan, VK3AYK
Reprinted from *Technical Forum*
with John Walker, ZL3IB
Break-In
March 1994

The function of the antenna combiner is to provide rejection or isolation of the opposite band in use.

Looking at the circuit (Figure 1), the operation of the combiner is quite straightforward. There are two discrete filters, one for each band of operation. The outputs of each filter are connected in parallel for connection to the antenna.

Filter 1 is a low-pass filter, allowing a VHF signal to pass almost unobstructed to the antenna, but providing a high-impedance path for any UHF signal.

For UHF, filter 2 is a high-pass type allowing UHF signals to pass, but inhibiting the passage of any VHF energy.

Construction

The device was built into a die-cast box approximately 1.5 by 1.25 by 2 inches in size. A small piece of printed circuit board was cut to fit neatly into the bottom of the box. This was then bolted to the bottom of the box. The screw heads were countersunk into the box. The nuts, after being tightened, were then soldered to the circuit board to provide a good electrical earth.

The capacitors and inductors are supported entirely by their own leads. They are soldered directly to each other and to the circuit board where required—no standoffs are used. The inductors should be mounted at right angles to each other. This will significantly reduce mutual coupling and provide greater isolation. I

found no shielding partitions were necessary provided this was done.

Component details are as indicated in Figure 1. For coils, 18 gauge tinned copper wire was used. All coils are close wound—with sufficient space so adjacent turns do not touch.

It is extremely important that 600 volt NPO disc ceramic capacitors are used. All component lead lengths (inductors and capacitors) are kept as short as practical—in effect, just long enough for the components to touch so they can be soldered. Remember, this is a VHF/UHF circuit!

I preferred to use chassis-mounted coaxial connectors. This facilitates an easier and more compact unit, as the start of each filter element for both bands can be soldered to the rear of the coax connector. Coaxial cable flying leads could be used (as do the commercial combiners) with suitable holes drilled in the box. The cable braid should be neatly trimmed and tinned, then soldered to the circuit board with suitable coax connectors at the other end for connection to the transceiver. Small standoffs should be used inside the box to support the coax center conductor where it is attached to the filter components. Keep the leads short!

Testing

Now comes the most important part—the testing. Check the unit and observe that all the wiring is correct.

Connect a nonreactive 50-watt dummy load to the VHF port at the combiner. Connect a sensitive UHF/VHF power meter terminated in a 50-ohm load to the UHF port. Then connect a 146-MHz power source (transceiver) to the antenna connector of the combiner. Key the transmitter on and read the power on the meter. It should be at least 40 dB down on the VHF power level (i.e., 40 dB is a power ratio of 10,000 to 1). Therefore, for 50 watts at the

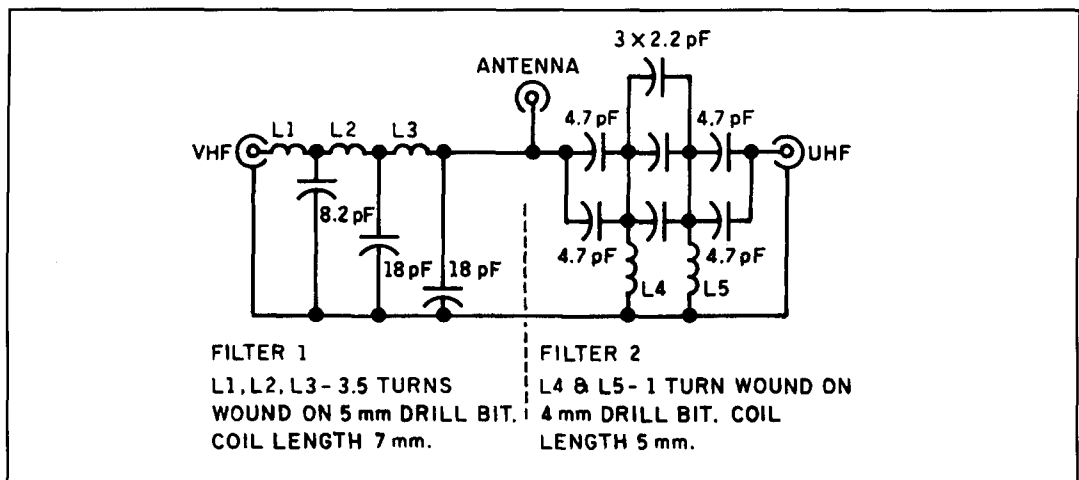


Figure 1. Schematic diagram of the combiner circuit.

antenna port, there should be in the vicinity of 5 mW or less observed on the power meter. If you feel it is necessary, try slightly squeezing or expanding the inductors associated with the high-pass filter section to optimize the above figure. This is not usually necessary.

Next, connect the dummy load to the UHF port of the combiner and the 50-ohm power meter to the VHF port of the combiner. Apply RF power (50 watts) at 440 MHz to the antenna port of the combiner and observe that the power level is at least 40 dB down.

Finally, check the insertion loss. It will probably be around 1 dB. This is done by connecting the power meter to the antenna port and applying a known power level to the appropriate VHF and UHF ports, respectively, and measuring the result. The loss for a power level of 50 watts should be around 10 watts.

When carrying out all of the above, try placing the lid on the die-cast box to make sure there are no violent changes in performance—indicating a possible problem.

Specifications (measured)

Insertion loss—approximately 0.8 to 1 dB (VHF and UHF) to antenna.

Attenuation—45 dB at 400 MHz for the low-pass filter. Around 44 dB at 149 MHz for high-pass filter.

Test instrument used—Wiltron Network Analyzer.

Conclusion

Commercial dual-band antennas for mobile use are available. To save money and retain what I think is a fairly modest antenna, I use a 2-meter quarter-wave whip. This whip is harmonically related to the 70-cm band and works reasonably well. RF power levels of 50 watts should not be exceeded. The unit has been running now for a number of months with no sign of problems.

Determining True North Accurately Without Instruments

*D.R.W. Hutchinson, M(SA)IERE
Condensed and Reprinted from
RADIO ZS
November/December 1993*

I have known the magnetic axis of the needle on a cheap compass to be 5 degrees or more out of alignment with the geometric axis. Additionally, the compass is influenced by the proximity of large ferrous masses such as the galvanized iron roof of your house. Even assuming that you have a good compass and no

large masses of iron nearby, once you have found magnetic north you can only find true north if you know the magnetic deviation for your area.

The method I am going to describe, utilizing the shadow cast by a rod, requires only the rod (a long knitting needle will do), a piece of thin string, a nail or pencil, and a level surface. It can be used anywhere in the world where the sun shines.

An original method

I think my method is original. I have certainly never seen it in print. The theory behind it goes like this. The length of a shadow cast by a thin vertical rod at 't' time before local noon will be exactly the same length at the shadow cast by the same rod at time 't' after local noon and the line bisecting the angle produced by the two shadows from the base of the rod, will lie exactly north-south. From this can be derived a method which will give true north in which no exact measurement is required. You don't have to know how long the rod and its shadow are, nor do you have to have a watch, let alone a chronometer.

Start the hour or so before noon (the earlier the better). Find a level piece of ground—I leveled an area about 3 feet square in an empty flower bed—and push the rod into the center of the side nearest the sun, so that it casts a shadow across the leveled area. For real precision, check that the rod is vertical with a plumb line. Mark where the end of the shadow falls. Make a loop in one end of the string and slip it over the rod and tie the large nail or pencil to the other end of the string so that it can be used to describe an arc across the level piece of ground.

Go away

Now go away and see what's happening on 15 or 20 meters for an hour or so, but keep checking to see whether the afternoon shadow has reached the line of the arc. As soon as it does, mark the point where it touches the arc. The two points marked on the arc and the base of the rod form the angle to be bisected. This is done by moving the rod first to the point at the end of the morning shadow, describing a small arc with the nail or pencil remote from the point where the rod was originally pushed in, then moving the rod over to the point at the end of the afternoon shadow and making a second small arc to cut the first.

Straight line

A straight line drawn between the point where these two arcs intersect and the point

where the base of the rod was originally, will lie north-south. The accuracy of the bearing will depend on the accuracy of your geometrical construction but should be, at worst, ± 2 degrees.

Southern Hemisphere

In the Southern Hemisphere, below Capricorn, north will be towards the original position of the rod and vice versa in the Northern Hemisphere above Cancer. In the tropics it will depend on the time of the year, but you should have some idea of where north is—the final line will lie north-south, but you will have to work out which end is which.

Semi-permanent indicator

I extended the line I scratched on the flower bed on to the adjacent garden path with a line of paint to give me a semi-permanent indicator.

If you would like a more permanent record, cast a small slab of concrete before you start and use that as your level surface, smoothing out all the construction marks and leaving the north-south line, which you can embellish as you think fit.

If you haven't got a garden, the procedure can be carried out on any level flat surface—an odd piece of plasterboard on a table near a window for instance—but if you do use a loose level surface, make sure that it doesn't move at any time during the operation. If it does, you will have to start again.

Now you know!

Well, now you know just where North is, so you can set up your beam and be reasonably sure that you know where it is pointing.

This article originally ran in the November 1969 issue of RADIO ZS. Ed.

PRODUCT INFORMATION

TP-3200 Shared Repeater Tone Panel

Communications Specialists, Inc. has announced the availability of their new Shared Repeater Tone Panel, model TP-3200. The unit is the complete tone and control interface between the repeater receiver and transmitter. It supports 157 repeater subscribers in CTCSS and DCS signaling formats.

The menu driven RS-232 programming is accomplished locally by computer or remotely with a modem over landline. DTMF programming on the repeater channel is also included. The TP-3200 is available in Desk Top or Rack Mount versions and is priced at \$279.95. The

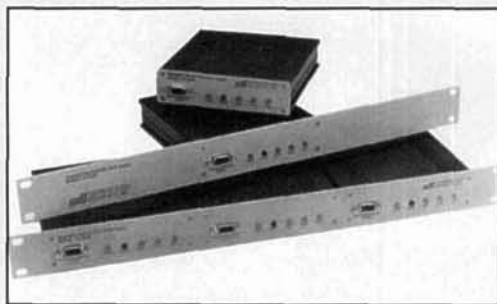
TP-3200 is available for immediate delivery from stock. A catalog is available on request.

For more information, contact Communications Specialists, Inc., 426 West Taft Avenue, Orange, CA 92665-4296; or phone: 800-854-0547; Fax: 800-850-0547 (USA/Canada); 714-998-3021; Fax: 714-974-3420 (Int'l).

QUALCOMM Introduces Low Power 1.6 GHz Phase-Locked Loop Frequency Synthesizer

QUALCOMM Incorporated, VLSI Products, introduced the new Q3216 Phase-Locked Loop (PLL) Frequency Synthesizer, a high-speed, low-power, single chip solution for high performance communications systems. The Q3216 PLL can interface with various system controls. It supports three modes of digital processor interface including 8-bit bus, serial loading and 16 TTL/CMOS-compatible parallel inputs.

For further information, contact James Botdorf, Marketing Supervisor, VLSI Products at 619-597-5005.



AERODYNAMIC BALANCING: PART 2

More on Yagi antennas with slightly drooped horizontal elements

In part 1 of this article, I discussed wind loading, torque, and boom balancing. In this second and final part, I continue my discussion of aerodynamic balancing by addressing imbalance due to elements, element torque balancing, the design and use of an offset mast and side arm mounted Yagis. I'll end with a summary of information presented in both parts 1 and 2.

Imbalances due to elements

Any attempt to correct an imbalance due to elements should only be tackled after all possible measures have been taken to balance the

boom. Assuming that the boom has been balanced, what measures can you use to minimize an imbalance due to elements where there is a boom-to-mast offset? One method is to build the antenna without a boom-to-mast offset. This is easier said than done; but a special boom-to-mast mounting arrangement can, with some difficulty, be built.

Let's consider the two antennas shown in **Figure 1**. Both have balanced booms with Antenna 1 having a net element projected area of 6 square feet and Antenna 2 having 8 square feet. Both have a 3-inch offset. Using a wind speed of 50 mph, **Graph 1** shows the mast

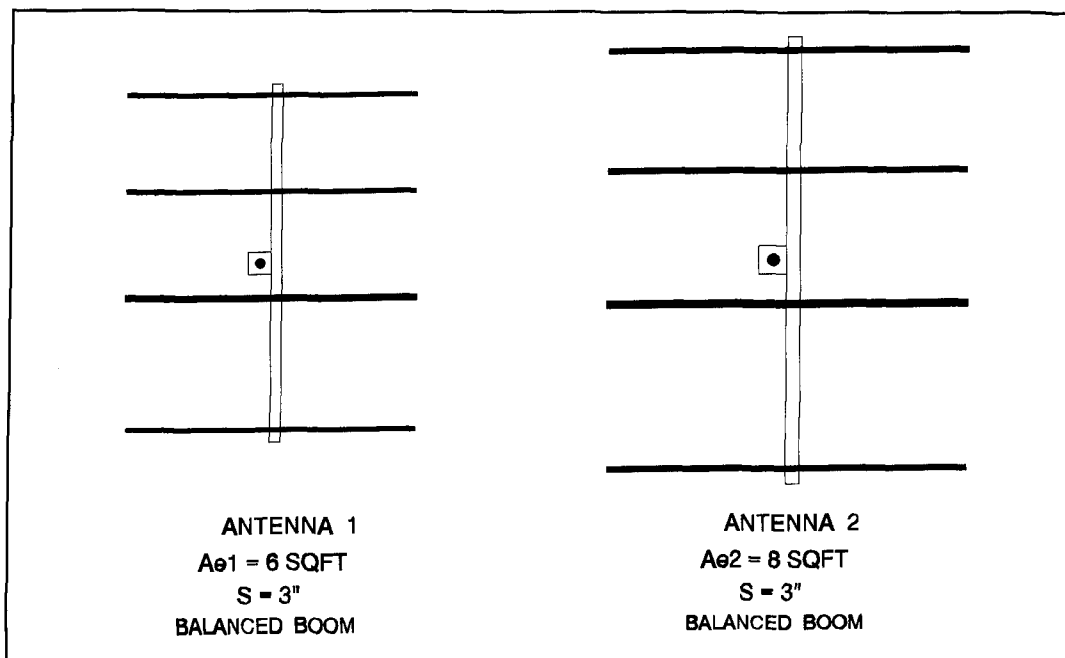
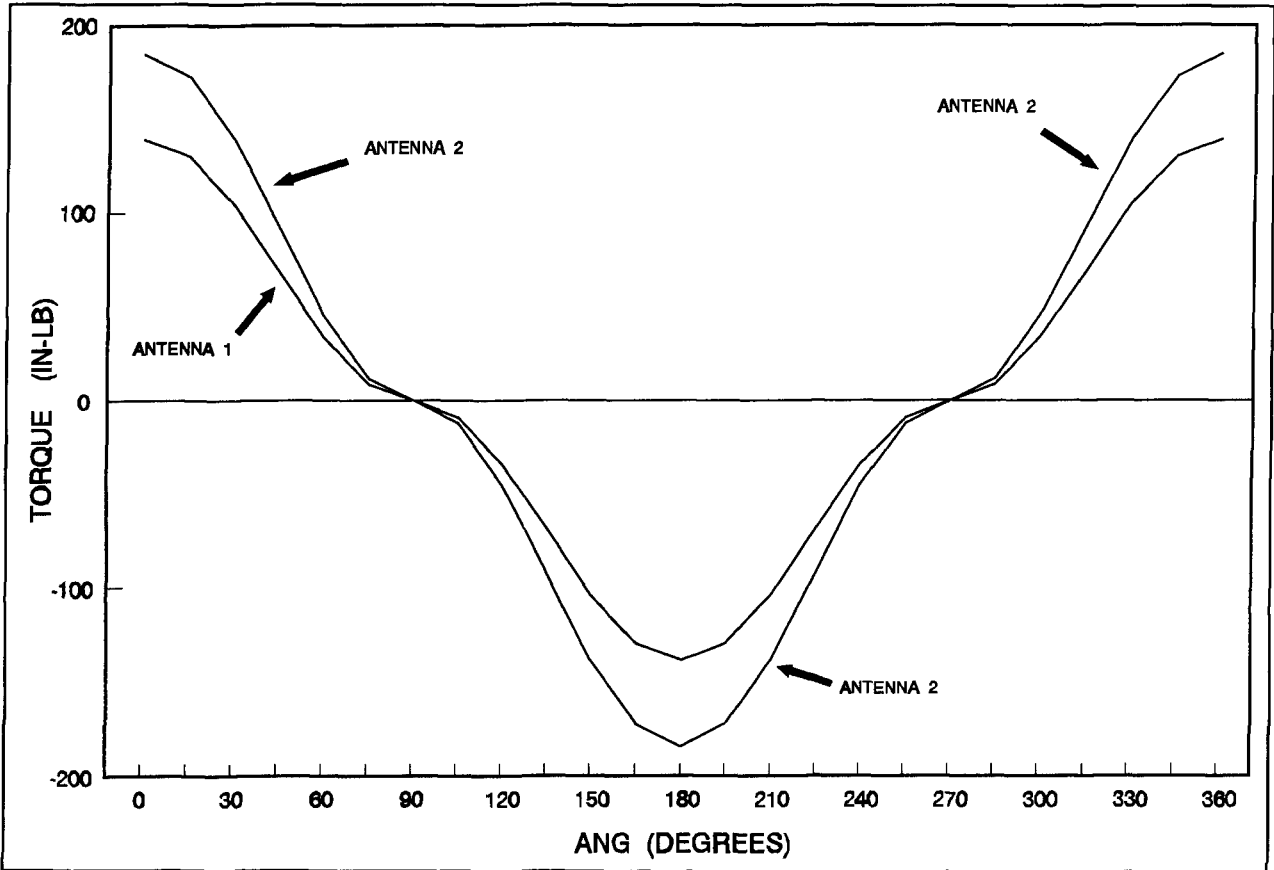


Figure 1. Two balanced boom antennas. Antenna 1 (Ae1) has 6 square feet; Antenna 2 (Ae2) has 8 square feet.

NET TORQUE



Graph 1. Net torque for Antenna 1 and Antenna 2.

torque for each antenna. The antennas are mounted as shown in **Figure 2**. Note the relatively low torque levels for each. Because a typical amateur antenna rotor is capable of producing 800 inch-lbs. of torque, there should be no problems turning either of these antennas in high winds.

Consider a typical tower installation that may have two antennas like these, or larger, mounted on a common mast. With two antennas mounted on a common mast, there's an opportunity to mount them in such a way that element torques from individual antennas counteract. **Graph 2** shows two cases. In the first,

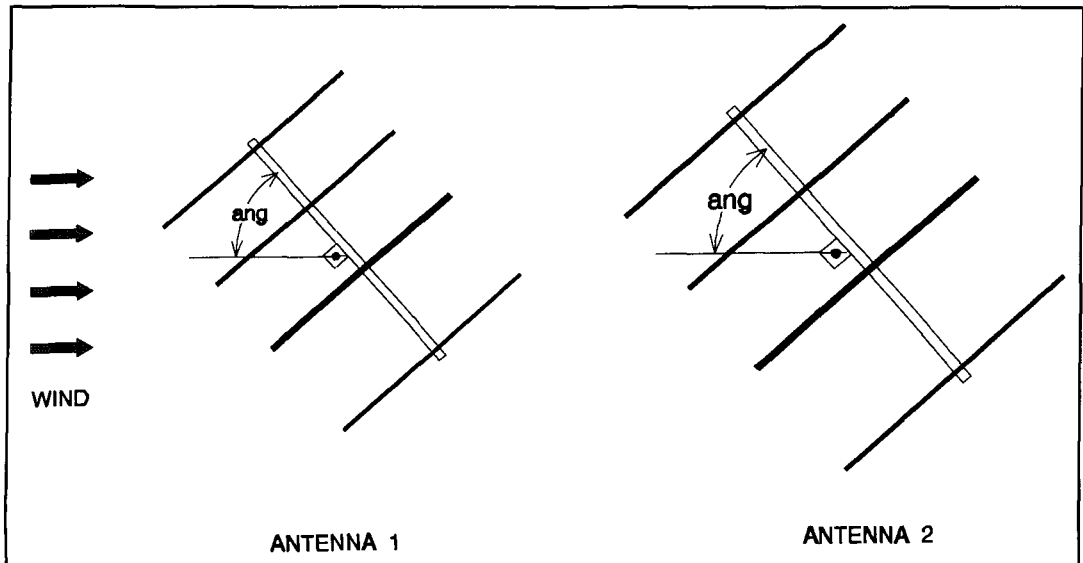
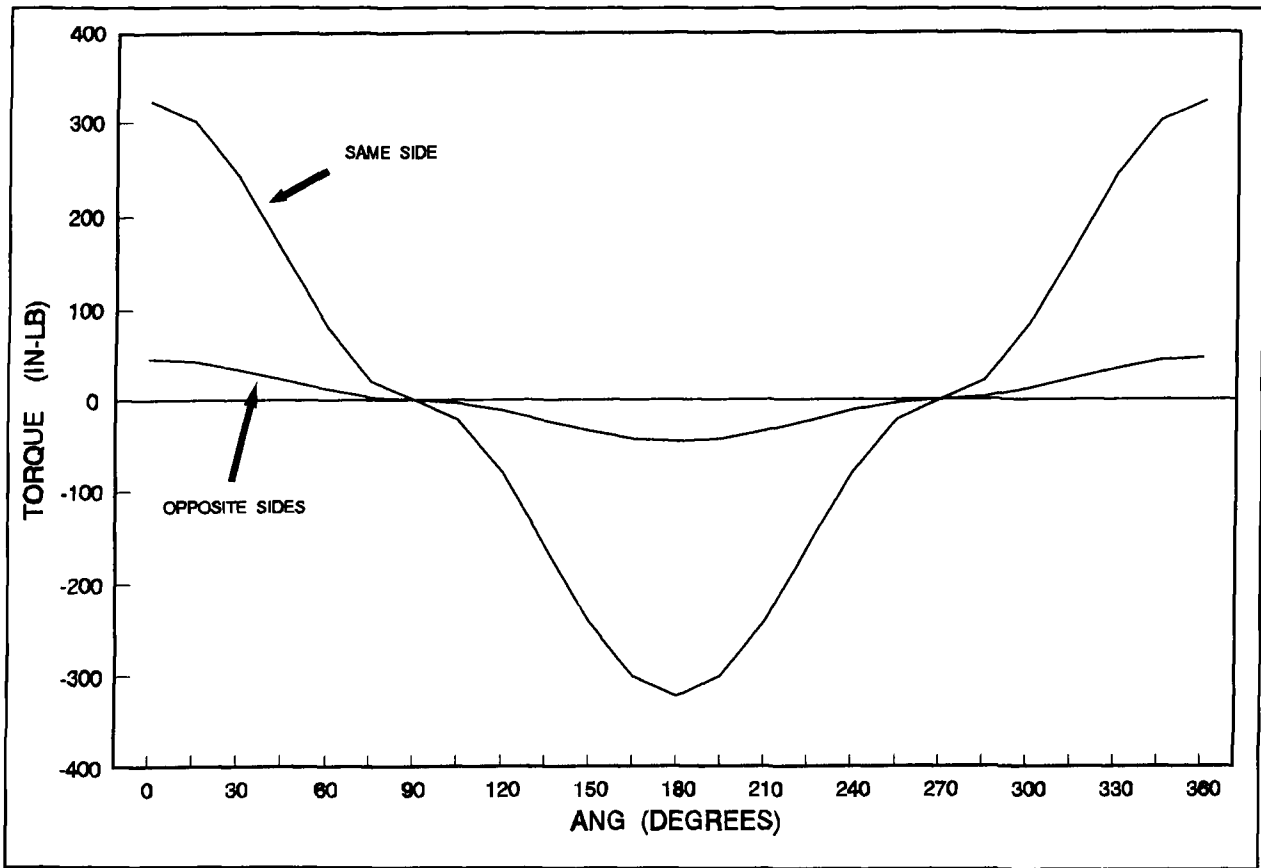


Figure 2. Antenna 1 and Antenna 2 mounted on the same side of the mast.



Graph 2. Net torque for antennas mounted to the same side of the mast compared with antennas mounted to opposite sides of the mast.

both “boom balanced” antennas are mounted on the same side of the mast, as shown in **Figure 2**. In this case, the torques due to boom-to-mast offsets for the individual antennas add. In the second case, shown in **Figure 3**, the antennas are mounted on opposite sides of the mast. Here, the torques counteract and yield a

reduction in the mast torque. When two antennas are mounted on a common mast, placement is a simple matter. They should be mounted on opposite sides of the mast.

From a mathematical viewpoint, one determines T_e for each antenna and sums the results in a manner that depends on the mounting

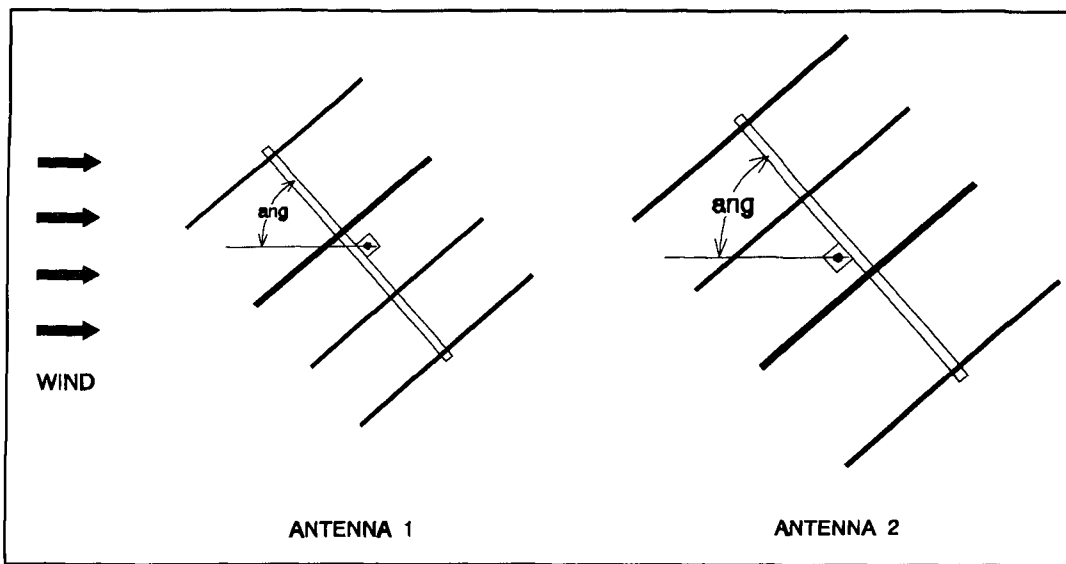


Figure 3. Antenna 1 and Antenna 2 when the antennas are mounted on opposite sides of the mast.

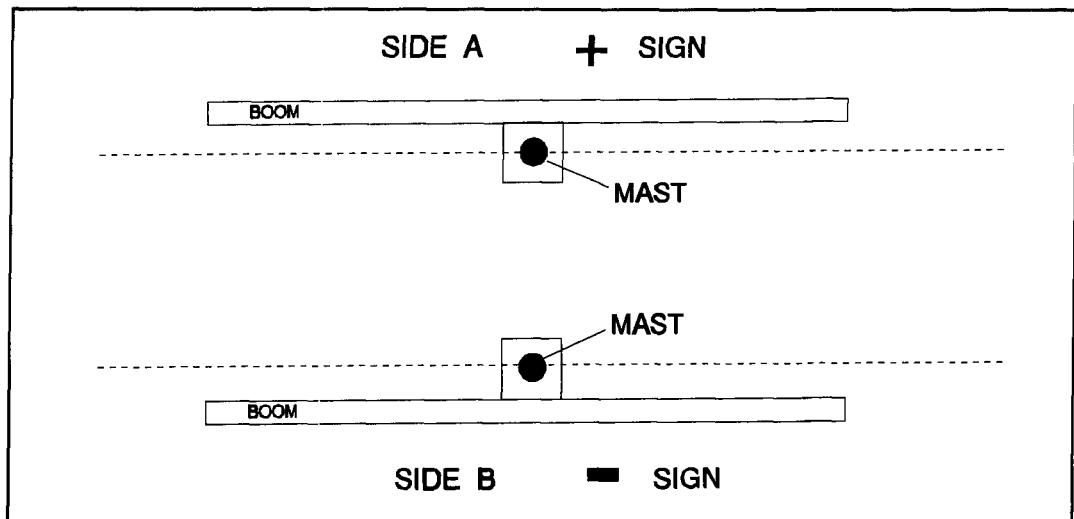


Figure 4. Mast side placement of antennas.

Antenna	S (ft)	Ae (sqft)	S Ae (cuft)
1	0.25	6.0	1.500
2	0.25	8.0	2.000
3	0.25	4.5	1.125

Table 1. Antenna "S Ae" values.

scheme. Equations for T_e for each antenna follow the form of **Equation 22** from part 1.

$$T_{e1} = 1.2 S_1 A_{e1} P \cos^2(\text{ang}) \quad (1)$$

where:

T_{e1} = torque due to elements on antenna 1
 S_1 = antenna 1 offset
 A_{e1} = antenna 1 net element projected area

$$T_{e2} = 1.2 S_2 A_{e2} P \cos^2(\text{ang}) \quad (2)$$

where:

T_{e2} = torque due to elements on antenna 2
 S_2 = antenna 2 offset
 A_{e2} = antenna 2 net element projected area

When antennas are mounted on the same side of the mast, as shown in **Figure 2**, the total torque due to elements is found by summing the contributions from both antennas as shown in **Equation 3**.

$$T_{es} = T_{e1} + T_{e2}$$

$$T_{es} = 1.2 S_1 A_{e1} P \cos^2(\text{ang}) + 1.2 S_2 A_{e2} P \cos^2(\text{ang})$$

Simplification provides:

$$T_{es} = 1.2 P \cos^2(\text{ang}) (S_1 A_{e1} + S_2 A_{e2}) \quad (3)$$

where:

T_{es} = total torque due to elements, antennas mounted same side

When the antennas are mounted on opposite sides of the mast, as in **Figure 3**, the total torque due to the elements is obtained by finding the difference of the two contributors following the sign convention in **Figures 11** and **19** of part 1. This results in **Equation 4**. In **Graph 2**, plots for values of T_{es} and T_{eo} as a function of angle (ang) show that a reduced torque can be achieved by mounting the antennas on opposite sides of the mast.

$$T_{eo} = T_{e2} - T_{e1}$$

$$T_{eo} = 1.2 S_2 A_{e2} P \cos^2(\text{ang}) - 1.2 S_1 A_{e1} P \cos^2(\text{ang})$$

Simplification provides:

$$T_{eo} = 1.2 P \cos^2(\text{ang}) (S_2 A_{e2} - S_1 A_{e1}) \quad (4)$$

T_{eo} = total torque due to elements, antennas mounted opposite sides

What if there are more than two antennas? In cases with more than two antennas, a mounting arrangement can be found that reduces the mast torque due to the elements. To determine the best mounting arrangement, a simple methodology can be derived based on features of **Equations 3** and **4**. From these equations, it's obvious that both antennas are subjected to the

same dynamic wind pressure. Also, both have round elements, so a drag coefficient of 1.2 applies to each. For a general case, the only discriminators are the magnitude of the net element projected area, the boom-to-mast offset for each, and the side of the mast to which they are mounted. In either **Equation 3** or **4**, the way to minimize the resulting torque is to force the sum of the terms within the parenthesis as close to zero as possible. To do this, use the magnitude of the "S Ae" term for each antenna and specify its sign using the sign convention in **Figure 4** for the proposed mounting arrangement. For each mounting scheme, sum these terms and pick the arrangement that has the terms within the parenthesis closest to zero. This is basically a trial and error method to find the mounting scheme with the least residual imbalance.

As an example, let's add a third antenna with a boom-to-mast offset of 3 inches and a net element project area of 4.5 square feet. Here the goal is to mount the three antennas on a common mast in a manner that provides the least resulting imbalance due to the elements.

Equation 5 shows terms whose absolute value when they are summed must be as small as possible within the mounting alternatives. It isn't necessary to try all possible combinations. "All antennas mounted to side A" has the same effect as "all antennas mounted to side B." The same holds true for combinations that are mirror images of each other. Recognizing this, one antenna can be assigned to one side and the other two antennas can be moved around to find the minimum absolute value of SUM as defined by **Equation 5**. In this equation antenna 1 is assigned to side A, which results in a positive sign.

$$\text{SUM} = + S1 Ae1 \pm S2 Ae2 \pm S3 Ae3 \pm \dots (5)$$

where:

$$\text{SUM} = \text{sum of "S Ae" terms}$$

For the three-antenna example, "S Ae" values for the each antenna are shown in **Table 1**. Using these values and the mounting sign convention in **Figure 4**, I evaluated four mounting combinations. **Table 2** shows the mounting configurations and evaluation methodology. Antenna 1 was assigned to side A, and antennas 2 and 3 were then mounted in three unique combinations. Next, the sign of each antenna's "S Ae" term was determined from **Figure 4** and the three "S Ae" terms summed for each mounting scheme. Results of the summation are listed in **Table 2** as SUM. The last step was to list the absolute values of SUM as ABS(SUM), as shown in the far right column. The mounting combination, or its mirror image, with the smallest value of ABS(SUM) results in the least torque during wind loading.

The first combination in **Table 2** with all antennas on side A is the worst case mounting scheme. Although I knew it to be the worst case, I evaluated the combination to ascertain the amount of improvement between it and the optimal mounting scheme. In **Table 2**, the least value of ABS(SUM) is 0.625 for the case where antennas 1 and 3 are mounted opposite antenna 2. If this mounting scheme is used rather than the traditional approach of putting all the antennas on one side of the mast, an 86.4 percent reduction is realized.

Although torques due to elements aren't very large for individual antennas, they may be con-

Antennas	Side	Sign	SUM	ABS (SUM)
Antenna 1	A	+	+1.500	4.625
Antenna 2	A	+	+2.000	Worst Case
Antenna 3	A	+	+1.125 = 4.625	
Antenna 1	A	+	+1.500	2.375
Antenna 2	A	+	+2.000	
Antenna 3	B	-	-1.125 = 2.375	
Antenna 1	A	+	+1.500	0.625
Antenna 2	B	-	-2.000	
Antenna 3	A	+	+1.125 = 0.625	Optimal Case
Antenna 1	A	+	+1.500	1.625
Antenna 2	B	-	-2.000	
Antenna 3	B	-	-1.125 = -1.625	

Table 2. Alternative mounting schemes.

siderable when multiple antennas are mounted to a common mast. Selecting a mounting arrangement using this methodology results in an optimal mounting configuration for very little effort. No computations are required for two antennas because there's only one solution. Two antennas should be mounted on opposite sides of the mast.

In practical applications, use of the net projected element area for each antenna, while at the same time ignoring boom end area effects and contributions due to element mounting plates, introduces very minor errors when evaluating mounting schemes. If a more accurate evaluation of mounting alternatives is desired, the "E" parameter discussed in **Reference 1** can be used with an equation similar to **Equation 5**. To make use of "E" parameters, use **Equation 6** in a manner identical to that in the three-antenna example above. A word of caution, do not use "E" for one antenna and Ae for another. Do the evaluation using only Ae values for each antenna or "E" parameters for each antenna. If you're interested in using the "E" parameter, a detailed explanation is given in **Reference 1**.

$$ESUM = +S1 E1 \pm S2 E2 \pm S3 E3 \pm \dots \quad (6)$$

where:

ESUM = sum of "S E" terms
 E1 = E parameter, antenna 1
 E2 = E parameter, antenna 2
 E3 = E parameter, antenna 3

Determining "E" parameters for antennas can be very time consuming and, as it turns out, when they are used to determine the best mounting scheme, there is no apparent benefit. This is because there are only a very limited number of possible mounting schemes and the very slight improvement in accuracy has no impact. However, in the future, "E" parameters will be included in antenna specification sheets by some manufacturers. In this case, an "E" parameter will be provided that will make finding ESUM an easy task.

Element torque balancing

There are some other ways to reduce the resulting torque due to elements. One approach is useful when there are multiple Yagis, or a single Yagi, already mounted on a mast and you wish to reduce the torque due to elements. In this case, remounting may be out of the question. Assuming the booms are balanced, a torque balancing plate like the one shown in **Figure 5** can be installed. In this case, the plate offsets the torque due to elements. Making use of ABS(SUM), you can find the size of a square element balancing plate with **Equation 7**. By looking at the sign of SUM, you can determine on which side of the mast to install the plate and arm assembly. If the sign of SUM is positive, put the assembly on side B—which has a negative sign using the sign convention in **Figure 4**. If SUM has a negative sign, put the assembly on side A, which has a positive sign.

Antennas	Side	SUM	ABS (SUM)	Y (ft)
Antenna 1	A			1.15
Antenna 2	A	4.625	4.625	(13.8 inch)
Antenna 3	A			on side B
Antenna 1	A			0.83
Antenna 2	A	2.375	2.375	(10 inch)
Antenna 3	B			on side B
Antenna 1	A			0.42
Antenna 2	B	0.625	0.625	(5 inch)
Antenna 3	A			on side B
Antenna 1	A			0.68
Antenna 2	B	-1.625	1.625	(8.2 inch)
Antenna 3	B			on side A
Antenna 2	either side	2.000 or -2.000	2.000	0.75 (9 inch) other side

Table 3. Element balancing plate sizes, with Rpe = 3 feet.

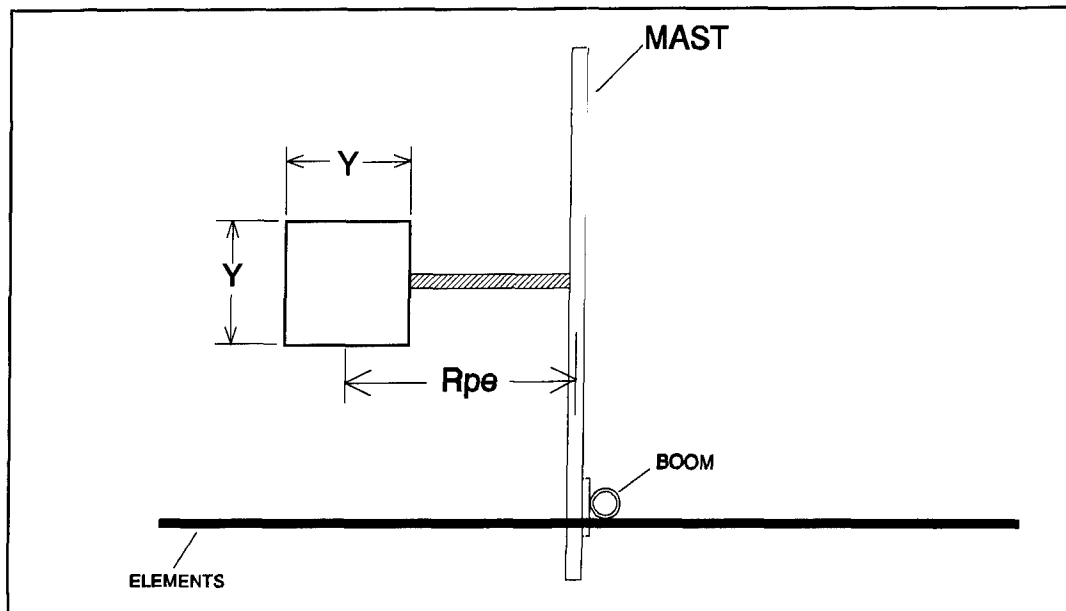


Figure 5. Element balance plate.

(The torque contribution due to the arm holding the plate is assumed negligible and not accounted for in Equation 7.)

$$y = \left[\frac{0.86 \text{ ABS (sum)}}{R_{pe}} \right]^{0.5} \quad (7)$$

$$\text{or, } y = \frac{\sqrt{0.86 \text{ ABS (sum)}}}{R_{pe}}$$

where:

Y = side of square element balancing plate

R_{pe} = distance from centerline of mast to centerline of element balancing plate

ABS(SUM) = absolute value of SUM

For each mounting combination in **Table 2**, element balancing plate sizes are shown in **Table 3** for an arm length of 3 feet. The plate size for antenna 2 mounted alone is also included. With this method, torque due to elements can be reduced for an existing installation, for a single Yagi, or where a mounting scheme doesn't reduce the residual torque to an acceptable level. A word of caution. Don't use "E" parameters with **Equation 6**. If "E" parameters are used for the antennas, **Equation 8** should be used.

$$y = \left[\frac{0.71 \text{ ABS (esum)}}{R_{pe}} \right]^{0.5} \quad (8)$$

$$\text{or, } y = \frac{\sqrt{0.71 \text{ ABS (sum)}}}{R_{pe}}$$

where:

ABS(ESUM) = absolute value of ESUM

Design and use of an offset-mast

There is another way to reduce the torque due to elements using a specially fabricated mast like the one in **Figure 6**. With this type of mast, the portion of the mast to which the Yagis are mounted is offset by the amount S_m. By using this type of mast with a single Yagi, the centerline of the boom can be positioned in line with the lower mast portion axis of rotation. As a result, the boom-to-mast offset S is effectively set to zero. Although there is an imbalance created by the portion of the mast that is offset, it can be ignored for relatively short masts. In the case where there's one Yagi on a short mast, S_m can be set equal to S.

For longer masts with multiple Yagis that have similar or different boom-to-mast offsets, it's possible to find a value of S_m which takes into account the imbalance of the mast and each antenna. The value of S_m that results in zero element torque can be determined using **Equation 9**. If "E" parameters are available for your antennas, find S_m using **Equation 10**. Needless to say, the booms must be balanced first, or the improvement in element torque loading with an offset-mast will not be realized. For example, suppose we had a 2-inch diameter mast that we wanted to make into a 16-foot off-

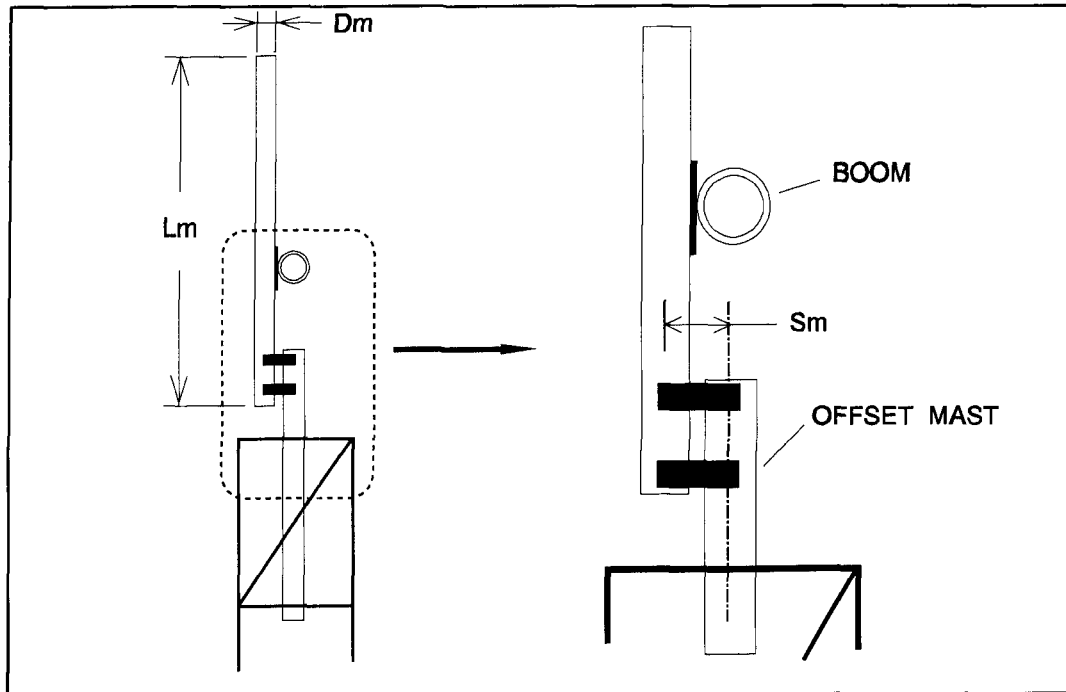


Figure 6. Specially fabricated mast helps reduce the torque due to elements.

set-mast, on which we would mount the three antennas shown in **Table 1**. Using **Equation 9**, the optimal value of mast offset, S_m , is 2.62 inch.

$$S_m = \frac{Ae_1 S_1 + Ae_2 S_2 + Ae_3 S_3}{D_m L_m + Ae_1 + Ae_2 + Ae_3} \quad (9)$$

where:

S_m = mast offset
 D_m = mast diameter
 L_m = mast length

$$S_m = \frac{E_1 S_1 + E_2 S_2 + E_3 S_3}{1.2 D_m L_m + E_1 + E_2 + E_3} \quad (10)$$

Before going to the effort of building an offset-mast, balance your Yagis' booms. If you plan to mount multiple Yagis, first find a mounting scheme based on the methodology demonstrated in **Table 2**. If the boom-to-mast mounts on your antennas can't be changed easily to allow "other side of the mast mounting," consider an offset-mast.

Side arm mounted Yagis

The use of side arms (see **Figure 7**) is quite common in efforts to mount more Yagis on existing towers. Although typically 300 degrees of rotation can be achieved, inherent torque imbalances result. For this mounting situation, the nature of resulting torques are no different than mast-mounted Yagis with boom-to-mast

offsets. However, in this case, offsets are much larger, resulting in far greater element torques. While a mast-mounted Yagi may have $S = 3$ inches, the same Yagi mounted on a side arm may have $S = 30$ inches.

Using the Yagi in **Figure 7**, **Graph 3** shows boom and element torques at 50 mph for a 32-foot boom, 2 inches in diameter, with an 8 square foot net element area while mounted on a side arm 30 inches long. First, let's consider the case where the boom is not balanced, but has $L_1 = 180$ inches and $L_2 = 204$ inches. As shown in **Graph 3**, the element torque is almost four times the boom torque. In previous examples, when the offset was 3 or 4 inches, the element torque was quite low. Now, however, with a 30-inch offset, the element torque is very large and is above torque ratings of many commonly used rotors.

Before proceeding, balance the boom, as we did before. This adjustment leaves the element torque as it is. The overall effect of boom balancing can be seen in **Graph 4**, which shows the net torque for the antenna before the boom is balanced and when the boom torque has been zeroed. Although there is improvement in parts of the array rotation relative to the wind, the peak net torque is the same for this antenna design. Beyond balancing the boom, there's nothing else that can be done with this mounting arrangement.

If you are considering using a side arm, make sure the rotor you're planning to use has sufficient torque to the turn the antenna in the anticipated winds. Also, make sure the rotor

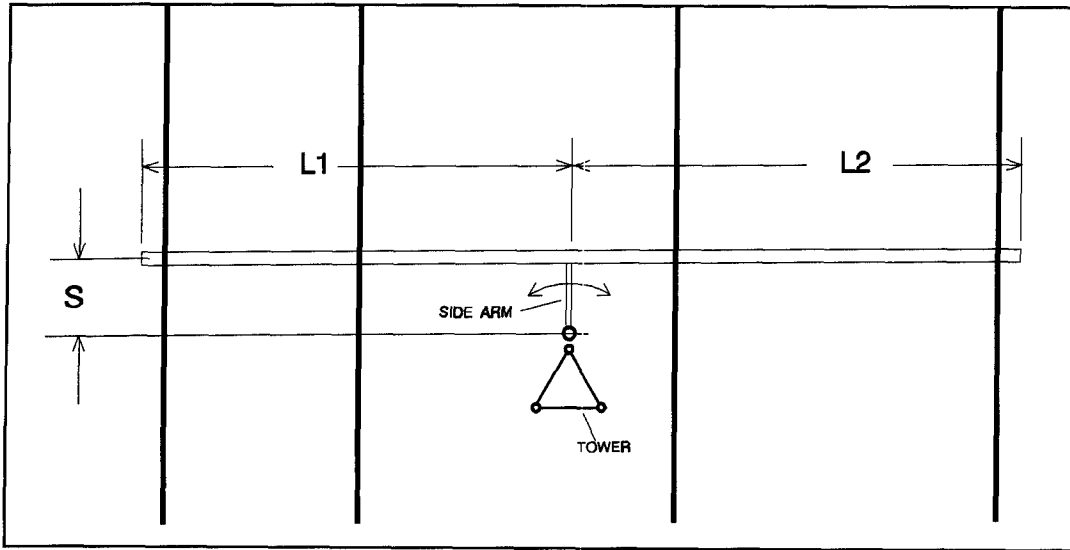
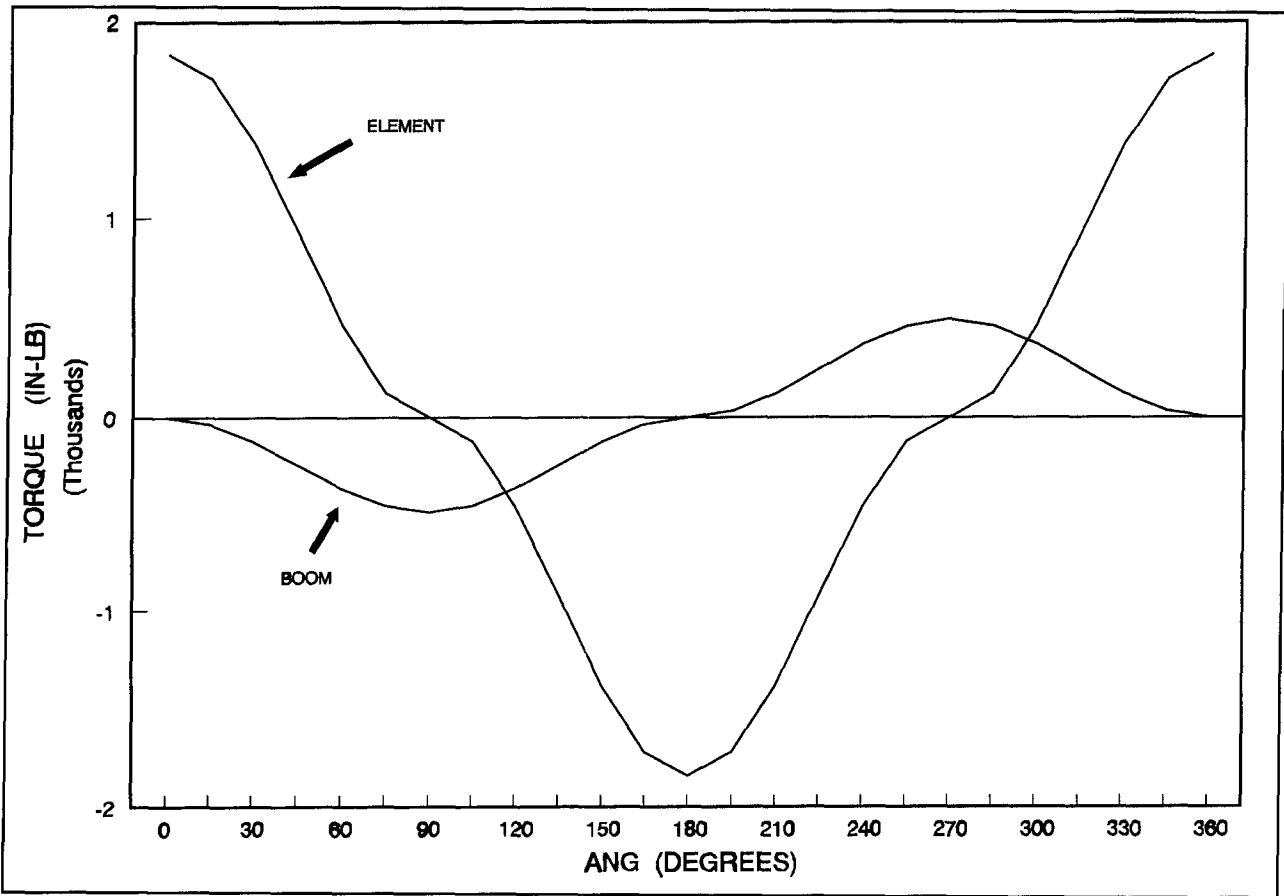


Figure 7. Side arms are used to mount additional Yagis.

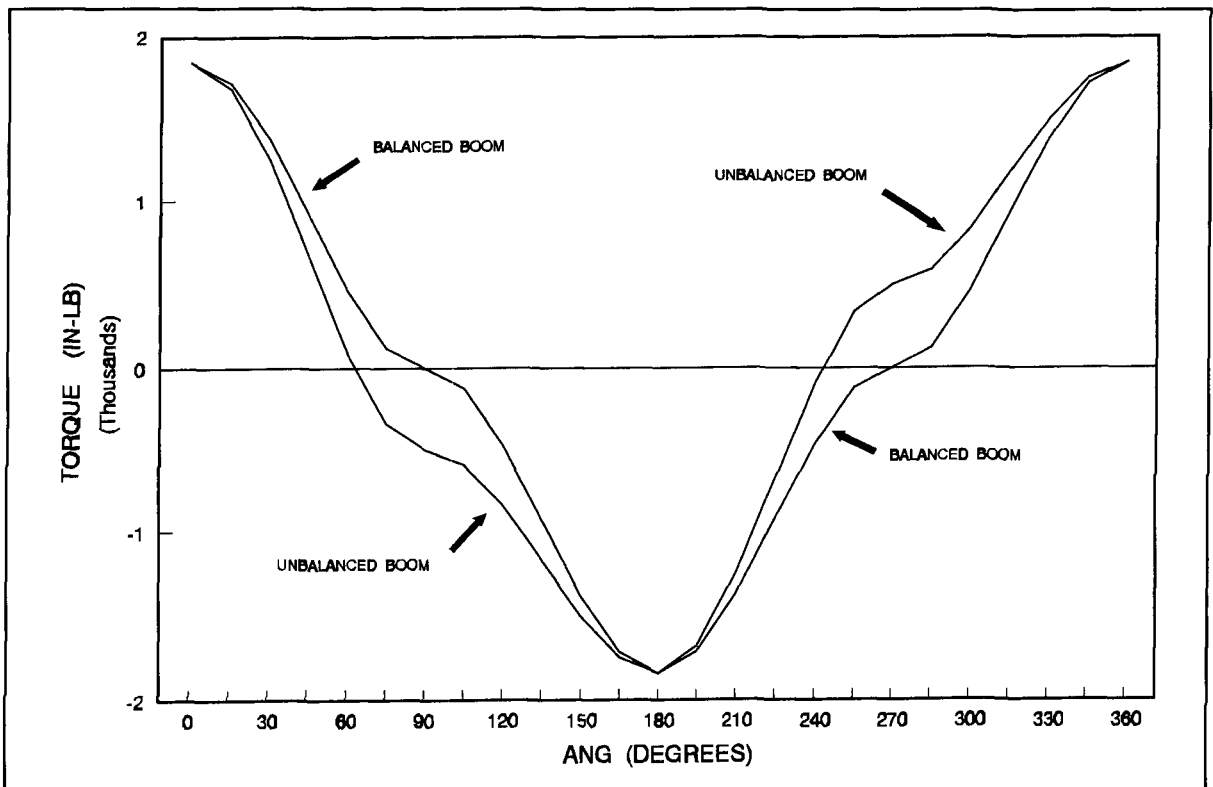


Graph 3. Element and boom side arm torque.

brake has sufficient capacity to hold the antennas during storm wind conditions. Use Equation 24 from part 1 to find these torques and to evaluate the effect of a balanced boom. You may find the peak net torque isn't affected by balancing the boom. In this case, balancing the boom isn't worth the effort.

Summary

In trying to reduce the torque required to rotate Yagi arrays in winds, minor design changes can be made to existing antennas to create a more aerodynamically balanced structure. For Yagis mounted to masts, the biggest



Graph 4. Net side arm torque.

torque imbalances are due to boom asymmetries. If possible, the boom-to-mast mounting bracket for constant diameter booms should be located at boom midpoint. A boom extension can also be used. If the boom has a stepped construction as shown in **Figure 13** of part 1, one of two approaches can be taken. Either the boom can be the same on each side of the mast or a construction can be used that satisfies **Equation 9** (see part 1). Most likely, when the boom-to-mast mounting plate is properly placed, it will be necessary to install a weight to provide a weight balance at the mast. If boom modifications aren't possible, a boom balancing plate that forces the boom into balance should be considered.

Perhaps one rather surprising feature of Yagis is the contribution of the elements to the net torque. The location of the elements on the boom plays no role in determining the amount of torque, but the element's net effective projected area and the magnitude of the boom-to-mast offset do. For multiple Yagis mounted on a common mast, the effect of torque due to elements during winds can be minimized by using a properly selected mounting scheme. For those cases where the optimal mounting scheme can't be implemented, you can use either an element balancing plate assembly, or build an offset-mast. In any case, attempts to reduce torque effects due to element wind loads should only be considered after all booms have been balanced. This applies equally to the case where

only one Yagi is used. If the booms aren't balanced, efforts spent to minimize element-induced torques will be masked and no improvement will be realized.

Although balancing isn't possible when a Yagi is mounted to a side-arm, the rotor turning and braking capacity can be predicted. In this case, you'll need "brute" torque and a strong brake to make this type of mounting arrangement work successfully.

A couple of final notes. First, there are a number of secondary factors that affect the torque to turn a Yagi in a wind. Such things as line chokes made from coiled coaxial cable taped to the boom, boom mounted baluns, driven element matching devices, feedlines, and anything else mounted to the Yagi, can create imbalances. I have not addressed these factors here, but rather have concentrated on defining the primary contributors, their effect on Yagi aerodynamic balancing, and the means to minimize their contributions. Most importantly, I have used the "cross flow principle" to define the magnitude and direction of forces that result when Yagi boom and elements are in a wind stream.¹

For the sake of brevity, derivations for **Equations 32** (part 1), **7, 8, 9,** and **10** were not included. I will be happy to send a copy of these derivations to any interested party who sends me an SASE.

REFERENCE

1. Dick Weber, K5IU. "Determination of Yagi Wind Loads Using the Cross Flow Principle." *Communications Quarterly*, Hicksville, New York, Spring 1993, pages 13-30.

QUARTERLY COMPUTING

Software for homebrewing plus an updated CD-ROM

Thanks in part to inexpensive and readily available semiconductors and an influx of technically savvy code-free

Technician licensees, homebrew construction of amateur radio equipment, an art declared dead at least once a decade, is very much alive.

In this issue, we'll examine two software packages that can help the homebrewer in diametrically different ways. We'll also revisit Buckmaster Publishing's amateur radio CD-ROM, which now features a Windows version of its ICALL callsign-retrieval program.

From Beverages Thru Oscar

For many amateurs, a construction project begins with a literature search to see what's been done before. While a visit to the public library can unearth amateur radio-related magazines and books, most amateurs maintain their own collections of back issues of popular journals. However, problems arise when you actually try to find relevant articles in a stack of magazines—you can skim each issue's table of contents, or locate and plow through several years' annual indexes (if they're available).

Compiled by Rich Rosen, K2RR, *From Beverages Thru Oscar*, abbreviated FBTO, can shorten search times and actually add value to those stacks of amateur radio magazines that you never had the heart to recycle.

A bibliography covering topics ranging from antennas through Yaesu upgrades, FBTO contains over 3 MB (megabytes) of data supplied in compressed form on one or two floppy disks. The software comprises over 60,000 entries taken from 297 sources including perennial favorites like *CQ*, *QST*, and *73*, discontinued

journals (e.g., *Ham Radio*), books, and bulletins. Publication dates range from 1901 through December 1990. Each entry contains basic information—title or description, source, date of publication, page number, author, and any comments entered by the archivist.

To install FBTO, you'll need an IBM-compatible PC that's equipped with 640 KB (kilobytes) of main memory and a hard disk drive with 3.2 MB of unused space to install FBTO, and either a color or monochrome display. A plain-vanilla XT clone works fine, albeit slowly.

You use FBTO's setup routine to install and expand the software. The program runs under PC DOS or as a DOS task under Windows. If your PC includes an extra 4 MB of RAM, you can create a RAMdisk drive for FBTO and enjoy nearly instantaneous reference retrieval.

You query FBTO by selecting a Search menu, from which you choose global (keyword), author name or callsign, date or name of publication as a search topic. For example, suppose that you're interested in building a converter for the 2-meter band. You enter "2M" as a global topic, and FBTO finds 2000 citations—too many to peruse. From the Main Menu, you select "Narrow Down Resulting Data" and enter "CVTR," which yields 127 citations. Finally, you select a further narrowing by requesting only those citations that appeared in, say, *Ham Radio* magazine. This time, you're presented with a list of 12 references to view onscreen or print.

FBTO's built-in search engine is easy to use, and a 54-page printed manual lists keywords and sources, and clearly explains the program's functions. As with other databases, it pays to try several search strategies—looking up "converter" yielded no entries, as FBTO prefers "con-

verter" as the correct spelling. Also, entering "2M" instead of "2M" improves search accuracy because the surrounding spaces eliminate many records that include the characters 2 and M as part of a citation (e.g., "...A 4.32 MHz).

While its contents clearly deserve an update, *From Beverages Thru Oscar* represents an excellent value for dedicated amateurs or researchers who maintain an extensive magazine backfile, or who have access to a library's collection. FBTO will quickly earn its \$85 purchase price in time and effort saved. (DiDah Publishing also offers databases covering individual topics or journals at prices ranging from \$10 to \$25—a bargain for anyone who collects only one publication.)

ARRL Radio Designer

Until recently, radio amateurs interested in owning their own computer-aided circuit design (CAD) programs enjoyed few choices—commercial software typically cost too much, and shareware products either lacked capabilities or suffered from awkward user interfaces.

All that changed in late 1994 when the American Radio Relay League, working with Compact Software of Paterson, New Jersey, released *ARRL Radio Designer*, a \$150 program comprising a significant subset of Compact's *Super Compact* professional RF CAD software.

ARRL Radio Designer analyzes active and passive linear circuits of up to 999 nodes (i.e., component junctions) at up to 513 frequencies. You can use built-in component models for bipolar and field-effect transistors, operational amplifiers, and transmission lines. You can also invoke one- and two-port "black boxes" modeled by S-, Y-, Z-, H-, and ABCD-parameter data.

To prepare a paper schematic for analysis, redraw the diagram with power and bias feed-points connected to ground—a detail accomplished by RF bypass capacitors in the actual hardware. Next, assign numbers to all nodes except ground, which is always node 0. You "build" a circuit for *ARRL Radio Designer* via the program's built-in editor or an ASCII text editor of your choice, creating a netlist that identifies all components' connections and electrical connections. You also specify other program blocks that control the frequency range of analysis, statistical analysis and optimization of circuit performance, data for one- and two-port network modeling, and group-delay modeling.

ARRL Radio Designer includes several databanks (libraries) of models that cover many RF devices offered by Motorola, Avantek, and other manufacturers. However, the databanks

don't include many older 2N-family devices; but given a data sheet, you can use a text editor to create a custom databank for your favorite components.

To do this, you import a device model from a databank via Windows' Notepad editor, cutting and pasting data into a circuit file via Window's Clipboard. Notepad won't open the large Avantek library, however. Use Windows Write or another editor. After completing your circuit's netlist, run *ARRL Radio Designer's* analysis and specify the style and format of the results displays. For example, you can view output amplitude in dB versus frequency.

Now for the special effects: you can invoke a tuning routine in which the software varies values of selected components and plots circuit response for each setting—a function equivalent to watching a filter's response on a spectrum analyzer while you adjust a trimmer capacitor.

Using the program's statistical capabilities, you can "build" 100 filters with components of various tolerances. Then you can examine the yield of good circuits and the distribution of their center frequencies.

Last, but not at all least, a built-in utility called Circles provides an automated Smith chart and interactive design of matching networks—a capability that alone is worth the program's purchase price.

If you've labored in the CAD vineyard—squinting at asterisks and plus signs denoting traces on crude 80-character-wide teleprinter plots—here's your reward. Running under Windows, *ARRL Radio Designer* presents gorgeous, multicolored plots featuring crisp, sharp, and colorful lines and legible titles. You can easily direct output to disk files, a printer or plotter, or the Windows Clipboard, making the program a hands-on pleasure to use.

To run this program, you'll need an IBM PC-compatible computer equipped with an 80386 or higher microprocessor, 8 MB of RAM, a mouse, a hard disk drive with from 6 to 11 MB of unused space, and Windows 3.1 or later. To load the software, your PC must include a 3-1/2 inch floppy disk drive. For fastest results, ARRL also recommends a math coprocessor—if your PC doesn't already include one.

ARRL Radio Designer's inch-thick, 400-page manual suffers from a split personality. Written by David Newkirk, WJ1Z, the first 100 pages ease the reader through program installation and a walkthrough of eight examples that showcase many of the program's features. Newkirk's free-flowing, easy-to-read prose makes these examples readily accessible to amateurs who aren't electrical engineers. At this level, virtually anyone can use the software and get results quickly.

The remainder of *ARRL Radio Designer's*

notes lapses into manual-ese, a writing style that is characterized by convoluted sentences and heavy use of the passive voice which is harder to read. Also, readers would probably benefit from some additional circuit-analysis examples and an expanded explanation of time-domain analysis.

Finally, the manual does a sketchy job of relating a "real world" circuit to its analytical model. While the program easily models lead inductance and stray capacitance effects, it's likely that beginners will oversimplify circuit models and wonder why physical circuits don't perform as well as the models predict.

On balance, *ARRL Radio Designer* deserves a place on the virtual workbench of every amateur who takes pride in homebrewing professional-quality equipment. It's also a worthwhile investment for anyone who will expend the time and effort to learn what this remarkable program can do.

Buckmaster Ham Call CD-ROM

Featuring over 400 new or updated amateur radio-related shareware and freeware programs, the October 1994 edition of the *Buckmaster Ham Call* CD-ROM now includes ICALLW, a Windows version of its ICALL U.S. and international callsign lookup program.

In operation, double-clicking on ICALLW's icon makes a window pop up that shows a field of white space and a blue data field. You type the callsign you're searching for in the blue window, and a few seconds later, licensee data appears in the white window. From there, you can build a queue of labels for later printing—a handy feature for creating a one-time list of QSL card address labels.

One minor problem surfaced during installation of ICALLW; its SETUP program couldn't copy a file called VER.DL_ to the review system's WIN subdirectory until an existing version of the file was renamed. Also, remember that various nations may restrict release of their citizens' callsign data, making international callsign coverage somewhat spotty.

Shareware highlights of this edition of *Buckmaster Ham Call* include version 6 of GeoClock, a superb map display that shows sun illumination of the Earth's surface in real time. This update includes a new gazetteer for city location and sundry database improvements. GeoClock's registered version can access *Buckmaster Ham Call's* callsign database. Updated in April and October, this \$50 CD-ROM is well worth consideration as your station's callsign database.

In upcoming columns, we'll track recent CD-ROM releases and other software focused on amateur radio. If you'd like to suggest software for review, please contact me at the address at the beginning of this column.

For more information

To obtain a copy of *From Beverages Thru Oscar*, contact DiDah Publishing, P.O. Box 7368, Nashua, New Hampshire 03060, or call (603) 878-3628.

You can purchase a copy of *ARRL Radio Designer* from the American Radio Relay League, 225 Main Street, Newington, Connecticut 06111-1494, or call (203) 666-1541.

To purchase a copy of the *Buckmaster Ham Call* CD-ROM, contact Buckmaster Publishing, Route 4, Box 1630, Mineral, Virginia 32117, or call (703) 894-9141. ■

PRODUCT INFORMATION

New Model 232FC RS-232 Frequency Counter

B&B Electronics Manufacturing Company has announced its new RS-232 Frequency Counter, Model 232FC. The frequency counter is powered from the RS-232 port handshake lines, and the 232FC allows you to make frequency measurements of TTL level signals. If for any reason the handshake lines are not available, a 12 volts DC external supply can be used to power the converter.

The 232FC makes frequency measurements from 5 Hz to 2 MHz and duty cycle measurements from 5 Hz to 50 kHz. Simple commands let you make measurements with your own pro-

gram. The 232FC is shipped with software that logs and plots frequency and duty cycle information, as well as display histograms. Sample code in BASIC, Pascal, and C that demonstrates the 232FC commands is provided.

For more information about the RS-232 Frequency Counter and to receive B&B Electronics' free catalog of *Serial Communication Interface and Control Equipment* contact: B&B Electronics Manufacturing Co., 707 Dayton Road, P.O. Box 1040, Ottawa, IL 61350 (phone 8a.m. to 4:30p.m. Central: 815-434-0846; 24-hour Fax: 815-434-7094; 24-hour BBS: 815-434-2927 Internet: techsupt@b&b_elec.com).

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4523 Thurston Lane, #5
Madison, Wisconsin 53711-4738

THE SOLAR SPECTRUM

An update

I thought I'd play a little catch-up this issue, with updates on solar cycle 22, and on the Ulysses mission, now in polar orbit above the Sun. First, let's turn to the current cycle.

Updating Solar Cycle 22

Figure 1 pretty well tells the story of cycle 22, which is now declining towards an expected 1996 minimum. (The end, or "minimum," of each cycle occurs at the intersection of the old declining cycle and the ascending branch of the new cycle, as measured by the smoothed monthly mean relative sunspot number.) At this

stage of development, some astronomers predict that the coming minimum will be a somewhat protracted one; perhaps similar to the lull that occurred in 1954, prior to the highest-amplitude cycle on record.

Just 24 class M, and no class X solar flares were recorded between January and November 1994. In fact, the most recent class X flare erupted over two years ago, towards the end of 1992. Some fairly intense geomagnetic storms have taken place during the past year, but most have been spawned by the long-lived recurrent coronal holes that are more prevalent during a cycle's descent, or in some cases, by fast moving coronal mass ejections.

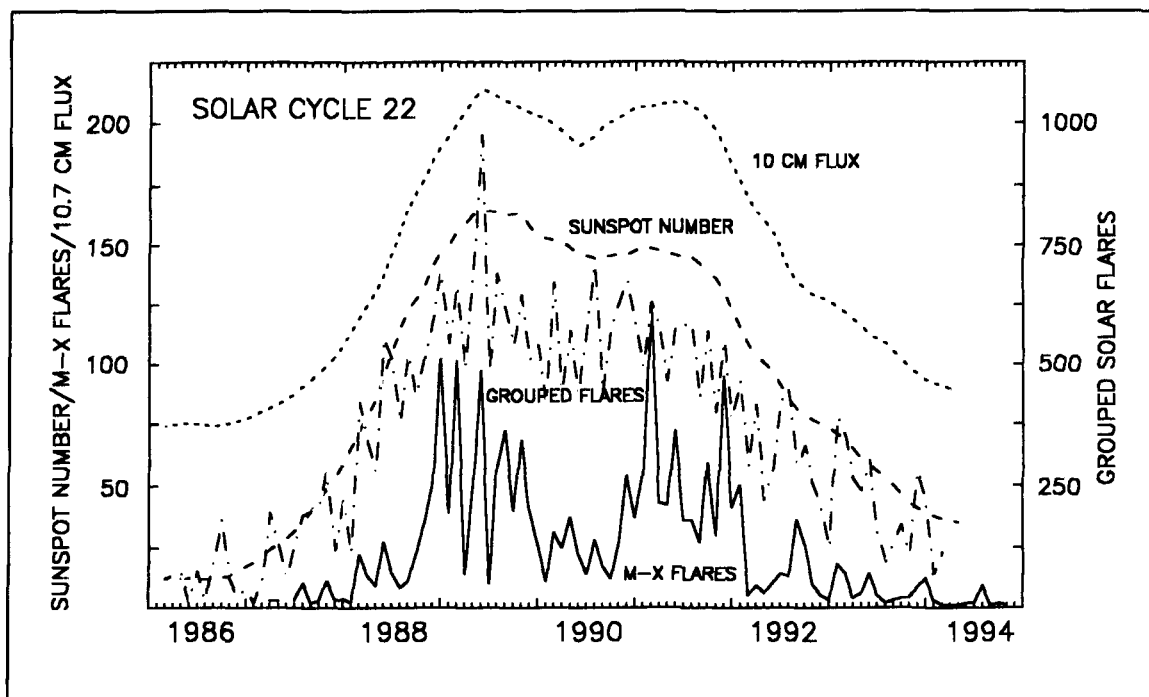


Figure 1. Some monthly indices of solar activity during cycle 22. The numbers of class M and X solar flares are shown according to the hemisphere where they occurred along with the hemispherical monthly relative sunspot number in *Figure 2*.

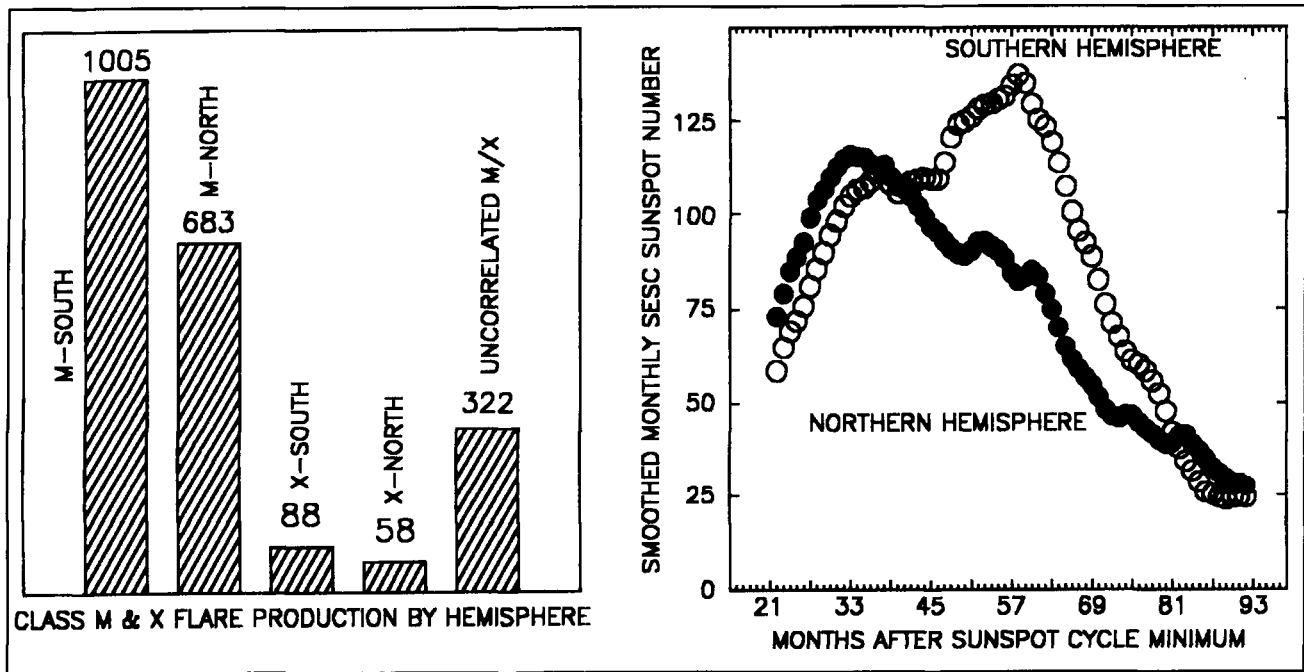


Figure 2. Solar flare eruption (left-hand diagram) and smoothed monthly sunspot numbers (unadjusted SESC estimates) by hemisphere during cycle 22. Note that sunspot activity in the Sun's Southern Hemisphere trailed that in the north until after cycle 22 reached maximum (July 1989), then went on to reach an even higher peak two years later. The Southern Hemisphere has been the site of a majority of class M and X flares throughout cycle 22.

Thus far during cycle 22, the sum-total of newly emerging sunspot groups is quite a bit lower than for cycles 20 and 21. However, when viewed in their entirety, cycle 22 sunspot numbers are roughly comparable to cycle 21, and considerably higher than those from cycle

20. Since even-numbered cycles tend to have lower maximum amplitudes than the odd-numbered cycles that bracket them, we would expect this effect for cycle 20, but cycle 22 numbers are only marginally less than those for cycle 21. Class M and X solar flare eruptions

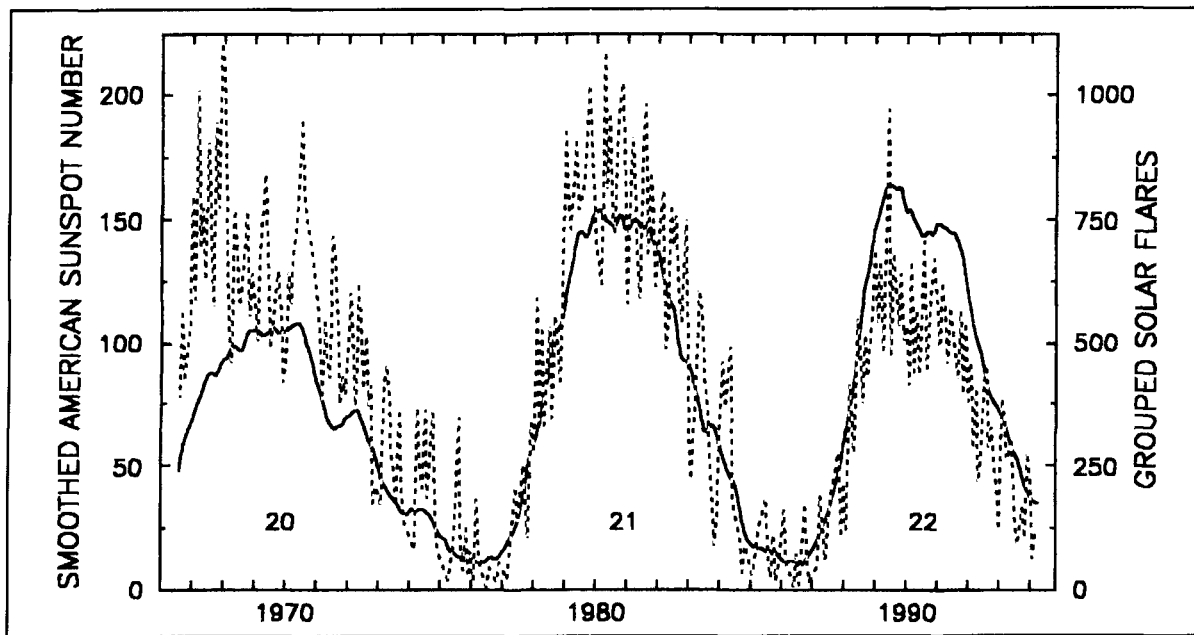


Figure 3. The total number of solar flares recorded each month, after allowance for duplication when events are reported by more than one observatory, is known as the "grouped solar flare index." As shown above, this index (dotted line) is in excellent agreement with the sunspot cycle (solid line).

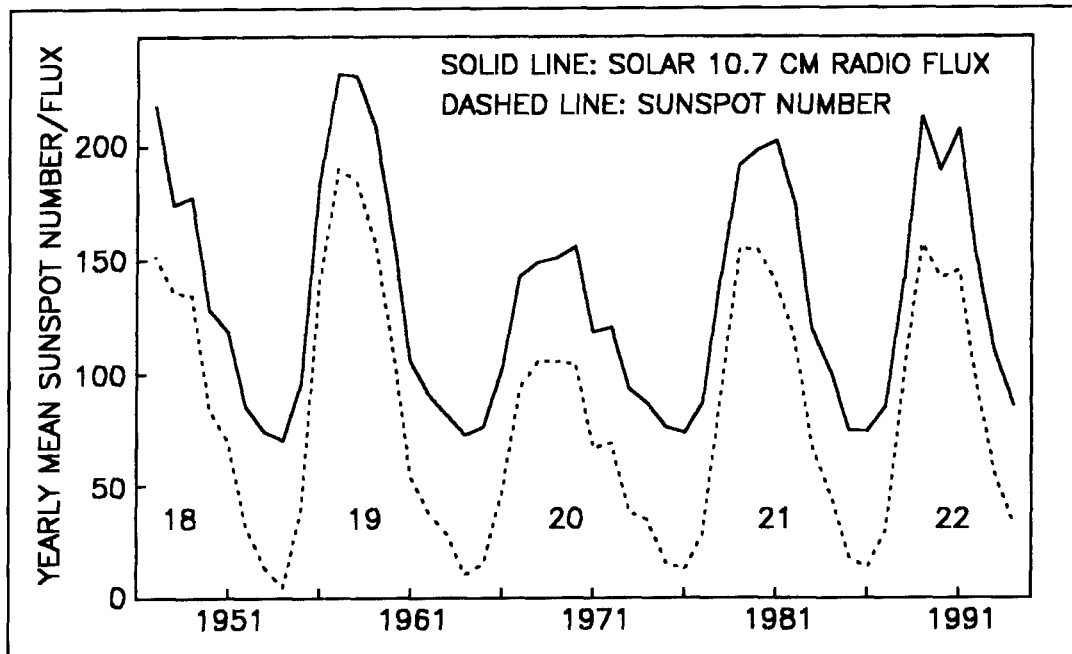


Figure 4. The solar 10.7 centimeter radio flux and yearly mean sunspot numbers, circa 1947 to 1994 (October).

and smoothed monthly sunspot numbers for cycle 22 are shown according to their hemispherical occurrence in **Figure 2**.

The number of individual solar flares recorded during each month of cycles 20-22 (through May 1994) is shown in **Figure 3**. Note that the overall number of flares is considerably lower for the current cycle. On the other hand, many more satellite proton events have occurred during cycle 22 than during either cycle 20 or 21. (The Space Environment Services Center in Boulder, Colorado, defines such events as periods in excess of 15 minutes when the >10 mega-electronvolt proton flux remains above 10 particle flux units per cubic centimeter, per second, at geosynchronous altitude.) However, in spite of the larger number of proton events at satellite altitude, cycle 22 ground level events have thus far been fewer in number, and more isolated, than during cycle 21.

Figure 4 shows that the monthly solar 10.7 centimeter radio flux was similar during cycles 21 and 22, and in both instances much higher than for cycle 20. The close agreement between these indices is very apparent in this diagram.

David Hathaway, NASA/Marshall Space Flight Center, provided the superb "butterfly-diagram" shown in **Figure 5**. The diagram clearly shows how sunspots first appear in two latitude bands at about 30-degrees north and south at the beginning of a cycle, and how these bands grow and move toward the equator as the cycle progresses. The diagram also shows the percent of the Sun's visible hemisphere that was covered by spot-groups during

each solar rotation since 1874.

When will the new cycle—number 23 in the long series of observed solar cycles—begin? The first sunspots of a new cycle typically emerge 12 to 18 months before the old cycle officially ends. New cycle spot-groups are relatively easy to detect, since they have a reversed magnetic polarity when compared with groups from the current cycle, and appear at an uncharacteristically high solar latitude for late-cycle spots; generally above 25-degrees. As of early December, no such guide posts had yet emerged and the exact onset of cycle 23 remained uncertain. However, most forecasts continue to specify mid-1996, as the more likely date for this event.

News from the Ulysses spacecraft*

One of the more striking attempts to learn the secrets of the Sun began in 1990, when a vehicle was launched to study solar regions never before seen—the Sun's polar regions. Carried aloft by the Space Shuttle "Discovery," this unique spacecraft was routed to the Sun the long way around...via the giant gas-planet, Jupiter (**Photo A**). A joint undertaking between NASA and the European Space Agency (ESA), the project was initially called the "International Solar Polar Mission," but in 1984 its

*A portion of this information was taken from bulletins issued by ESA, NASA, and the Jet Propulsion Laboratory.

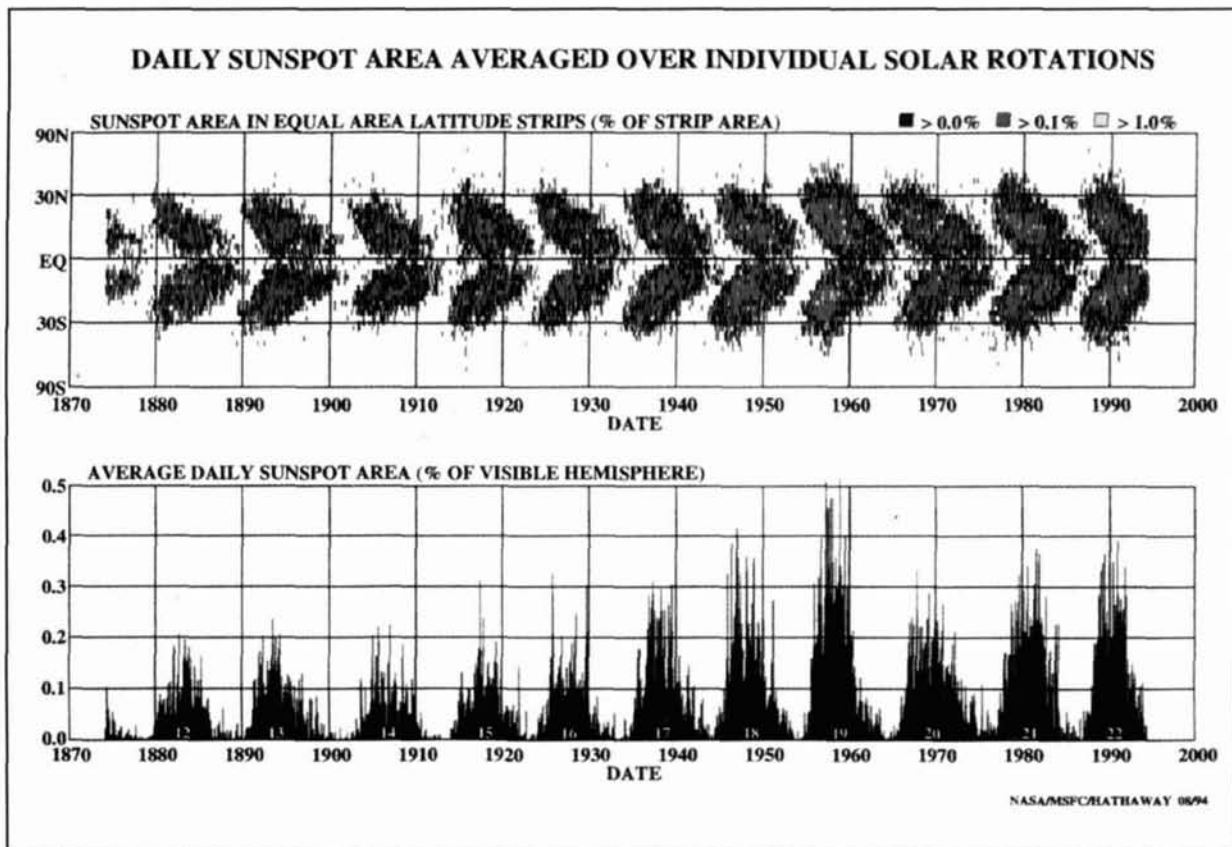


Figure 5. A sunspot cycle “butterfly-diagram” for 1874-1994 from data obtained by Greenwich Observatory and the NOAA/USAF SOON network. The bottom panel shows how much of the visible surface of the Sun was covered by sunspots for each rotation of the Sun since 1874. The upper panel shows how much of each of 50 equal-area latitude strips from the South to the North Pole was covered by sunspots. Black indicates the presence of small spots; red indicates between 0.1 and 1.0 percent of the strip was covered; and yellow indicates that more than 1.0 percent of the strip was occupied by sunspots during that rotation of the Sun. Diagram courtesy of D. Hathaway, NASA/MSFC.

name was appropriately changed to Ulysses, the hero of Greek mythology whose adventures are described in Dante’s *Inferno*.

Although Ulysses was designed to study the Sun itself, the project obtained new information about the heliosphere, the effects of the solar wind on Jupiter’s magnetosphere, and the origins of cosmic dust, before beginning its direct observations of the Sun. In mid-1994, Ulysses ended a journey of nearly 2000 million kilometers, finally soaring over the Sun’s South Pole where it spent four months gathering data. Ulysses has now completed its exploration of the regions surrounding the Pole, providing astronomers with their first opportunity to obtain measurements from a location that has never before been visited by spacecraft or ground-based instruments.

Several months after Ulysses began its polar fly-over, scientists at the ESA space research and technology center in Noordwijk (the Netherlands) reported that the polar region displayed some unexpected magnetic and dynamic

characteristics, and hinted that our thinking on the magnetic structure of the Sun might need to be revised.

The first findings to surprise astronomers were the unexpectedly low cosmic radiation activity above the South Pole (cosmic ray intensity in this high latitude region increased, but not nearly to the extent that scientists had predicted), the unexpected uniformity of the Sun’s magnetic field, and the remarkable apparent absence of a south magnetic pole. Richard Marsden, ESA’s project scientist, explained: “We expected that we were going to find a local increase in the solar field’s intensity. But the probe did not detect any such thing, and all the evidence so far suggests that the Sun has no south magnetic pole.”

It appears that the way Ulysses sees the Sun’s magnetic field is very different from the way it is seen by observations from Earth. Measurements from Earth show the Sun having a magnetic field with magnetic poles that are not the same as the Sun’s north and south geo-

graphic poles (axis); as the Sun rotates, the magnetic poles go around. Ulysses, however, found a uniform magnetic field at the Sun and did not detect any magnetic pole whatsoever.

Could it be that the Sun truly has no such poles? This and other thought-provoking questions are being studied by the team of scientists working on the project, who are keenly looking forward to comparing these results with those that will become available when Ulysses passes over the Sun's North Pole in mid-1995.

However, it is already manifestly clear that the structure of the solar magnetic field in the southern polar region is not as predicted by the traditional models. In particular, the instruments on Ulysses detected a new type of very slowly varying electromagnetic waves, with oscillation periods of 10 to 20 hours. The experts' theory is that this is due to an unexpected phenomenon

which conveys the solar magnetic field into space via the solar wind. Furthermore, in the polar regions this hot, ionized flow of gases and energetic particles emanating from the Sun was found to be moving at the very high velocity of 750 kilometers per second (about two million miles per hour), approximately double the speed at which the solar wind is known to flow at lower latitudes.

It is still far too early to determine what effect these new findings are going to have on our overall understanding of the Sun and the interplanetary solar wind it generates. A vast volume of data running into thousands of millions of bits—all of it of enormous scientific value—has been acquired during Ulysses' initial polar pass. A detailed analysis will require many months, but one thing is certain; there is no doubt about the complexity of the phenome-

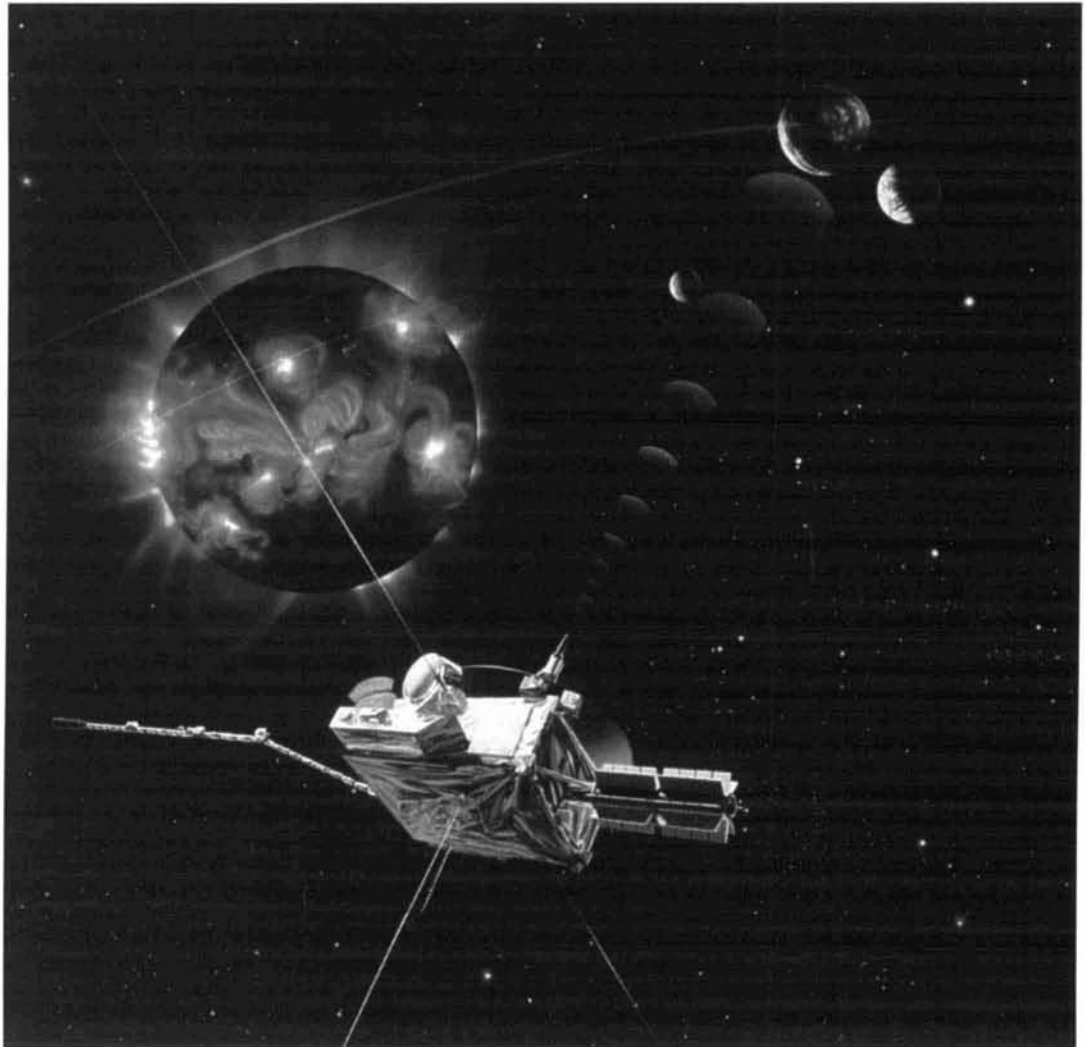


Photo A. This artist's rendering depicts the Ulysses spacecraft as it began its pass below the Sun's South Pole. The spacecraft reached 70-degrees south of the Sun's equator on June 26, 1994, marking the beginning of its primary mission to study these previously unexplored regions of the Sun. Ulysses completed its southern pass at the end of 1994 and looped back around the Sun, continuing on to the northern polar region, where it will arrive in mid-1995. Illustration courtesy of the Jet Propulsion Laboratory—California Institute of Technology/NASA.

na that have been observed, which involve the combined effects of the solar wind, magnetic field, electromagnetic waves, and fast-moving bursts of particles. Further insights will almost certainly be gained when Ulysses passes over the Sun's North Pole, and, if the mission is extended, when it makes two additional polar passes in 2000-2001, during a period of increased solar activity.

At this is written, Ulysses is heading toward

the Sun's equator, where it will make its closest approach of approximately 192 million kilometers (120 million miles) on March 12, 1995. At that time, scientists plan to conduct a "coronal sounding" experiment in which they will use the spacecraft's radio beam to measure the electron content of the Sun's corona. Ulysses will then continue its northern ascent to traverse the Sun's northern polar region beginning on June 19, 1995. ■

PRODUCT INFORMATION

New AEA DM-1 Deviation Meter

Advanced Electronic Applications, Inc. has introduced the DM-1 Deviation Meter. The DM-1 is designed for measuring the deviation of FM transmitters operating in the 144, 220, or 440 MHz amateur bands. A unique feature of the DM-1 is that the tuning is crystal controlled, which provides stable measurement and eliminates the need for manual tuning. Two deviation ranges allow sufficient resolution for accurate measurement of voice, data, DTMF, and subaudible deviation. Handhelds, mobiles, and base stations can be checked for correct audio deviation.



Power is supplied by one standard 9 volt alkaline battery. When battery voltage drops below 5.5 volts, the DM-1 is automatically disabled to prevent erroneous readings.

For more information, contact Advanced Electronic Applications, Inc., P.O. Box C2160, Lynnwood, WA 98036 (Phone: 206-774-5554; Fax: 206-775-2340).

AEA's New Log Windows Version 2.0

AEA's new version of Log Windows is now compatible with their PC PakRatt for Windows version 2.0. Users now have TNC control of PC PakRatt for Windows 2.0, coupled with the logging and tracking features of Log Windows. Antennas can be turned to the short path, long path, or in an arbitrary direction, with the click of a mouse. A new Database Browser enables users to sort and print logs by any criteria. Users can also query on-line callbook databas-

es such as SAM, QRZ, or HAMCall for a call-sign at any time.

Log Windows 2.0 was created to automatically display DX spots and allow users to move to the designated frequency quickly, log the contact, and save the information in a log. Log Windows 2.0 can announce DX spots with a voice-synthesized DX announcement. A filter can be turned on so Log Windows 2.0 will only display and sound an alarm for DX that are needed. The stand-alone LW Import program allows CT, DXLog, Log Master, Easy DX, Hyperlog, DX Base, N6RJ 2nd Op, Log View, DX Desktop, PC PakRatt, and any ASCII log to be imported into Log Windows 2.0. Logging and award tracking are supported for ARRL DXCC, WAS, VUCC, and *CQ Magazine's* CQ Zone and USA-CA awards.

For more information, contact: Advanced Electronic Applications, Inc., P.O. Box C2160, Lynnwood, WA 98036 (Phone: 206-774-5554; Fax: 206-775-2340).



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- FCC General Class Question Pool
- Updated Records for all Major Contests
- Proposed Vanity Callsign Rules
- Revised U.S. Ham Census
- All-new Color Photography
- U.S. Radio Clubs by State & City
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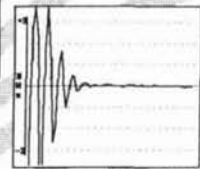
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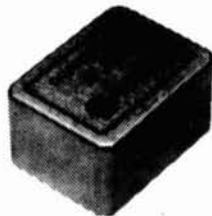
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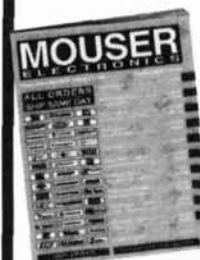
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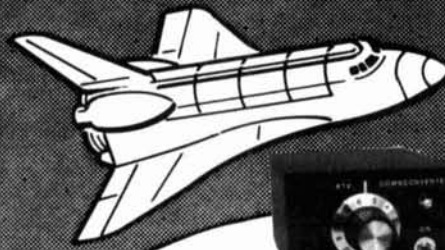
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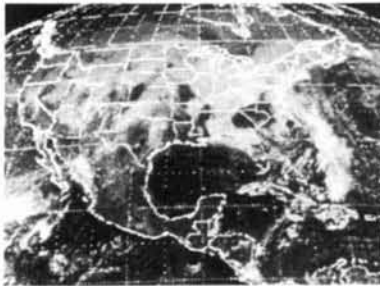
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ADVERTISER'S INDEX

AEA/Adv. Elec. Applications.....	3
Alinco Electronics	Cov.II
Amidon Associates	111
AMSAT.....	112
Astron Corporation	5
Beezley, Brian, K6STI.....	109
C & S Sales	109
Communications Concepts	109
CQ Almanac & Calendars.....	108
CQ Equipment Buyer's Guide	110
CQ New Books	8
Down East Microwave.....	111
Hewlett Packard.....	7
Kenwood	Cov.IV
Lewallen, Roy, W7EL	111
Motron Electronics.....	109
Mouser Electronics	110
Nemal Electronics	110
Nye Engineering	110
PC Electronics.....	112
Software Systems Consulting	112
Spectrum International, Inc.....	110
TX RX Systems Inc	111
Yaesu	Cov.III

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H F T R A N S C E I V E R S



TS-950SDX •160m to 10m amateur band operation, 100kHz to 30MHz general coverage receiver •Built-in DSP (Digital Signal Processor) •MOS-type FET final section •Dual frequency receive •Menu system



TS-850S •160m to 10m amateur band operation, 100kHz to 30MHz general coverage receiver •DDS (Direct Digital Synthesizer) & digital PLL system •AIP (Advanced Intercept Point) system •IF slope tuning •Electronic message keyer circuit with weighting control



TS-450S/690S •160m to 10m amateur band operation, 500kHz to 30MHz general coverage receiver •6m amateur band and 50-54MHz general coverage receive (TS-690S) •DDS (Direct Digital Synthesizer) & digital PLL system •AIP system •IF shift function tuning •AF notch filter



TS-140S •160m to 10m amateur band operation, 500kHz to 30MHz general coverage receive •IF shift circuit •31 memory channels with multi-scan functions •Dual-mode noise blanker •Speech processor with audio compression amplifier



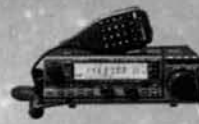
TS-50S •Super-compact, 100-watt, 160m to 10m transceiver, 500kHz to 30MHz general coverage receiver •DDS (Direct Digital Synthesizer) with "fuzzy logic" control •AIP system •100-memory channels •CW reverse •Menu system



TS-790A •144/440MHz dual-band operation •1200MHz unit (option) •All-mode operation •Satellite communications with Doppler effect frequency correction •59 multi-function memory channels with lithium battery back-up



TS-60S •50MHz all-mode operation •Max. 90W RF output (SSB, CW & FM) •100 memory channels •DDS with "fuzzy logic" control •AIP, IF shift (SSB & CW) and optional 500Hz CW filter •Dual-menu system



TM-255A/455A •All-mode operation (TM-255A: 144MHz; TM-455A: 430MHz) •101 memory channels •DDS with "fuzzy logic" control •TF-SET (TX frequency set) function •DTSS selective calling with page •1200/9600bps packet terminal

F M M O B I L E T R A N S C E I V E R S



TM-742A/642A/942A •Multi-band operation (TM-742A: 144/440MHz; TM-642A: 144/220MHz; TM-942A: 144/440/1200MHz) •Optional FM band units for TM-742A/642A (TM-742A: 28/50/220/1200MHz; TM-642A: 28/50/440/1200MHz) •Dual/triple receive capability •101 memory channels per band •Detachable display & control panels (option)



TM-733A •144MHz/440MHz dual-band operation •Dual receive on same band (VHF +VHF or UHF +UHF) •"6-in-1" programmable memory •72 memory channels •DTSS selective calling with page •1200/9600bps packet terminal •AIP system •Cross-band repeater function



TM-241A/331A/441A •Single-band operation (TM-241A: 144MHz; TM-331A: 220MHz; TM-441A: 440MHz) •20 multi-function memory channels plus call channel •Multi-scan capability •Selectable CTCSS tone encoder •Multi-function microphone supplied



TM-251A/451A •Single-band operation (TM-251A: 144MHz; TM-451A: 440MHz) •41 memory channels (optionally expandable to 200) •Dual-band receive •Digital recording system •DTSS selective calling with page •1200/9600bps packet terminal

COMMUNICATION RECEIVER



R-5000 •100kHz to 30MHz in 30 bands, with optional 108-174MHz coverage •SSB, CW, AM, FM & FSK modes •Dual IF crystal filters •IF shift circuit •100 memory channels with multi-scan functions

F M H A N D H E L D T R A N S C E I V E R S



TH-79A(D) Deluxe •144MHz/440MHz dual-band operation •Mos FET power module •Dot-matrix LCD, guide function & menu system •Wide range flexible antenna •Dual receive on same band •82 non-volatile memory channels in E²PROM •ID memory & DTMF memory •Built-in CTCSS encode/decode •Cross-band repeater function



TH-28A/48A •Single-band transmit and dual-band receive (TH-28A: TX 144MHz/RX 144/440MHz; TH-48A: TX 440MHz; RX 144/440MHz) •Alphanumeric memory •Alphanumeric message paging •40 multi-function memory channels (non-volatile) •Tone alert system with indicator



TH-22A/42A •Single-band operation (TH-22A: 144MHz; TH-42A: 440MHz) •MOS FET power module •Built-in DTMF keypad •40 memory channels in E²PROM (plus 1 call channel) •Multiple scan functions •Dual scan stop modes (CO & TO) •Channel Display function

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