COMMUNICATIONS QUARTERLY

THE JOURNAL OF COMMUNICATIONS TECHNOLOGY

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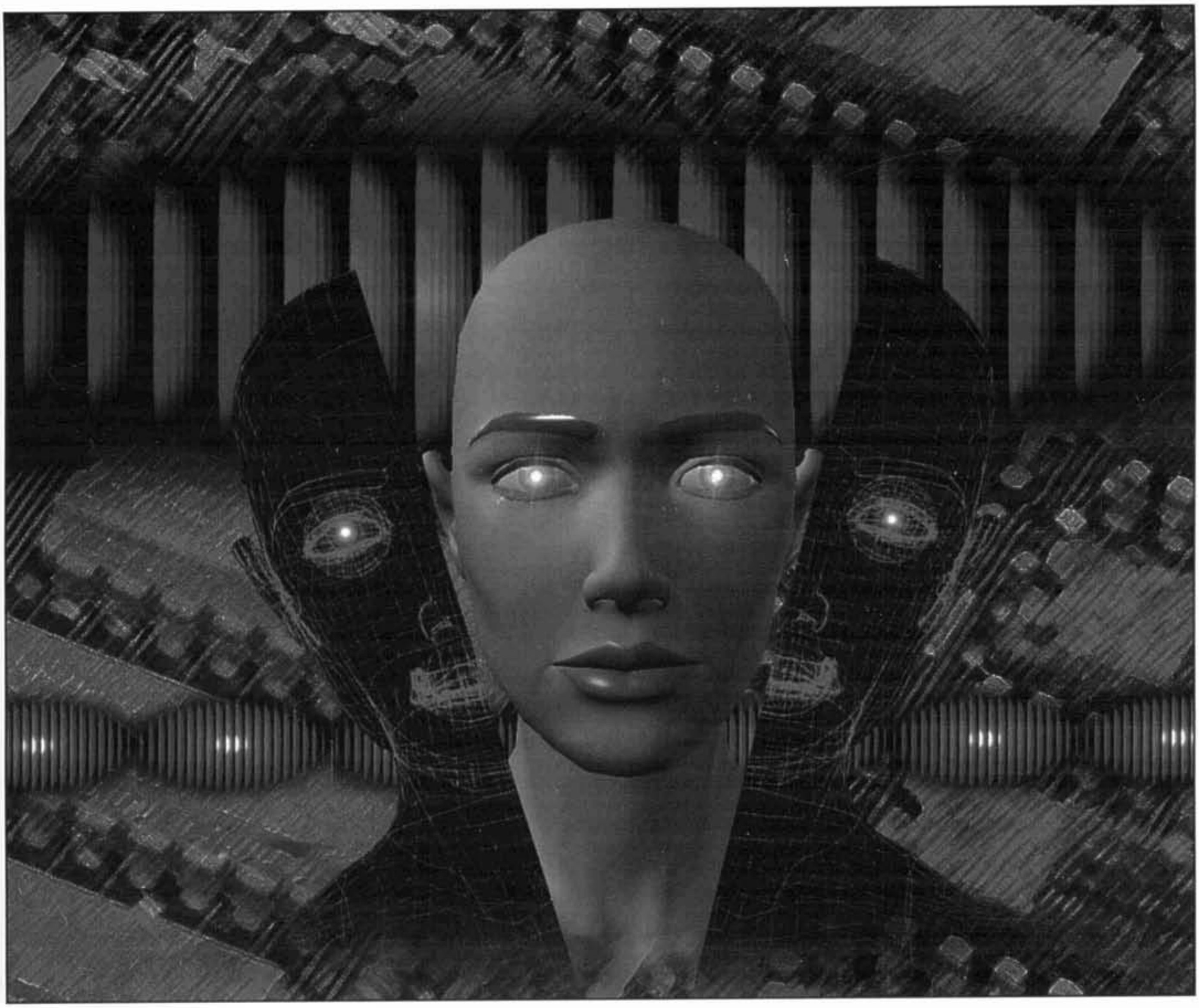

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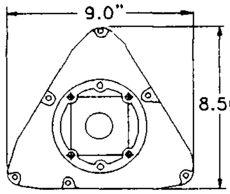

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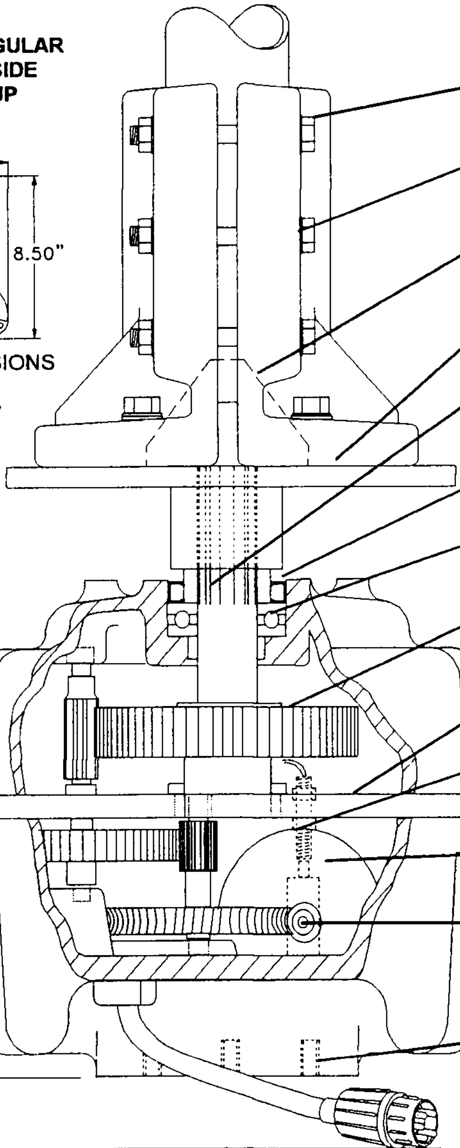

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Communications Quarterly

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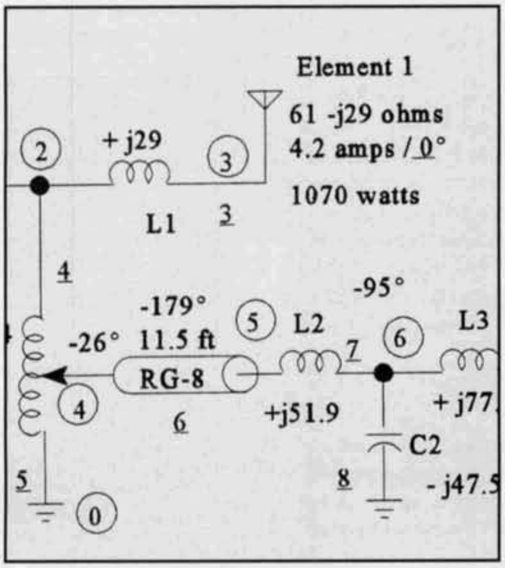
CONTENTS

Volume 9, Number 1 Winter 1999

page 59



page 93

On the Cover: The spiritual component of technology as envisioned by Bryan Bergeron, NU1N. Beneath all the wires and ICs behind the machines that we strive to make more like ourselves, lies our spiritual self. Humankind marches on, for the most part because of technology, which is an overall positive influence. We control the technology we grow with and emerge from—not unlike a butterfly from its cocoon.

EDITORIAL

Toward a radio-computer interface

Back in the 1980s, personal computers began to appear in our shacks. Names like Sinclair, Commodore C-64, and RadioShack TRS-80 were bandied about on the air by those who were interested in the new technology. To these adventurous hams, the new, desktop computers could be interfaced with their rigs to provide even more communications versatility. Looking back over the literature, one notices the advent of computer articles in most of the radio magazines.

Computers have allowed both the technically oriented and the amateur radio manufacturers to refine and improve radio design and performance. We now have fully programmable rigs, down to the handheld level. There are radio-based programs for everything from computerized virtual radios that interface the two pieces of equipment, computerized logs for contesters, optimization programs for antenna designers, and weather and propagation programs.

Over the years, *Communications Quarterly* has featured a number of articles on interfacing your computer to your radio. Back in 1993, and again in 1995, Howie Cahn, WB2CPU, brought us two pieces on "Connecting Computers to Radios"—one on a PC interface for the Ramsey 2-meter transceiver and another on adding DDS frequency control. In the spring of 1996, Bob Haviland, W4MB, brought us "CADA: Computer Aided Design for Amateurs." And, in the summer of 1998, Mike Hall, WB8ICN, made his writing debut with "A BASIC Stamp Project for Amateur Radio." Other computer-radio-related articles are cataloged each fall in the "Quarterly Index."

Is ham radio doomed?

Is ham radio doomed to be swallowed by the computer industry? I don't think so. Listen around. There are still lively discussions on the band. On contest weekends, the frequencies are still jammed. I've heard many ragchewing sessions that revolve around *computer* problems.

I figure that both technologies can coexist and grow together. Right now, ham radio seems to have reached a plateau where confusion regarding digital modes and other computer controls has brought a slowdown in technological advances. Still, it's only a matter of time until we leap over this hurdle and move to the next level of radio discovery.

comm-quarterly.com

In the interest of encouraging the interface of radio and computer technology, and breaking the technology dam that's holding us back, *Communications Quarterly* is finally online. You'll find our new web site at: <comm-quarterly.com>. Contained within the site is information on the magazine and its purpose, a synopsis of articles from the current issue, an online author's guide, a hand-searchable index of nearly all our articles, and an online subscription form, for starters.

The site was developed by Drew Northup, N1XIM, who has served as a summer intern compiling our yearly index. Drew's a senior and Computer Department Assistant at Dover High School in Dover, New Hampshire. He plans to attend Syracuse University in the fall, majoring in computer engineering. He's excited about combining his computer skills with his radio knowledge. Rob Pope, also a Dover High student, assisted Drew with the graphics portion of our site. A junior, Rob designs web sites professionally, but offered his services to us without cost. A hearty thanks to both of these young men for finally getting us online.

Please include us

Please remember to include *Communications Quarterly* as part of your online routine. Use it as a way to combine your love of radio with your love of computers. Look for articles in our index that bring the two technologies together (there are many) and come up with your own ideas. Then write them up for your fellow subscribers to enjoy. For instance, we're looking for articles that identify possible digital formats for speech and graphic communications via ham radio.

On to the millennium!

The amateur radio community has always stressed the importance of technical study and advancement. If we continue to stress excellence in the development of new equipment and technologies, we *will* be a force far into the next millennium!

Terry Littlefield, KA1STC
Editor

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Australia a state?

Dear Editor:

In reference to your Fall 1998 issue, page 83, please note that, in your introduction to "Tech Notes," the initials WIA stand for WIRELESS INSTITUTE OF A-U-S-T-R-A-L-I-A, not AMERICA, as stated!

On page 2 of this fax, a script of my weekly broadcast for the WIA, Queensland Divisional News Service, you will see a reference to your magazine, including a mention of your interpretation of WIA!

This does not matter much; it is all in good fun and I am not game to start writing down what typical OZZIES make of some U.S. initials, if they actually knew what they meant!

With best wishes to all at *Communications Quarterly* for Christmas (white no doubt, as against our heat or storms or flood rains) and a very happy and successful 1999.

John Aarsse, VK4QA
WIA Q-News International Reporter

Here's the portion of John's broadcast that mentions the Quarterly*:*

Communications Quarterly, Volume 8, Number 4, the Northern Fall issue, has, as usual, quite a few interesting articles. I have mentioned before, that most of the articles in this magazine are of a very high caliber and, for the keen experimenters, include deep reaching mathematics. What about a description of a class-E power amplifier and digital driver for 160 meters at 700 watts of CW and 250 watts of AM, using only $20 worth of transistors in your final amp! Or a tunable all-band HF camp mobile antenna. And from ZS-land, ZS6BPZ describes a multiband direct conversion receiver!

Even VK-land contributed with a reprint of an article about a VHF/UHF signal generator by Will McGhie, VK6UU, originally published in *Amateur Radio*, the Journal of the Wireless Institute of, believe it or not, America! I checked the magazine's publication date and it definitely was not April 1! So unbeknownst to us, we may have become the umpteenth state of the USA. Can someone please verify this for me!?!

For LF operators, a most illuminating article by Robert R. Brown, NM7M, of Washington, on unusual low-frequency signal propagation at sunrise around May 1998. The various graphs certainly looked spectacular! Finally, an article about surface-mount technology brought back quite a few, not so happy memories from my days with the ABC. Seven Three for now.

—Assembled and read by VK4QA.

Thanks very much, John, for the review and the gentle ribbing. I hadn't heard that Australia had become a state; but if it were to become one, it would certainly be our largest!—Ed.

In defense of the study of elevated radial systems

Tom Rauch's overpowering comments on the "Elevated Radial Systems for Vertically Polarized Ground Plane Antennas" appearing in *Communications Quarterly* in the Spring 1998 issue has the effect of discouraging rather than furthering the study of elevated radial systems. Consider that, with 5,000 or so AM stations in the U.S. alone (using 120 radials 1/4 wavelength long), there are 26,000 miles of wire buried. If the efforts of amateur radio operators in just the U.S. are considered, doubling that number does not seem out of line. Attempts to reduce the large ground areas required by conventional monopole radiators should be encouraged and welcomed.

The curiosity aroused by elevated ground systems seems to indicate that the researchers cannot seem to give a complete explanation of the effect of ground conductivity on the distribution of the return currents in an elevated (insulated) radial system. Literature in both the professional and amateur press spanning 60 years doesn't give the complete picture either. Hopefully, Mr. Rauch's dogmatic comments will not discourage others from trying to search for the answers that will expand our knowledge of artificial ground systems.

Joe Gagliardi, AA1BU
Holliston, Massachusetts

Feedback

Dear Editor:

Since you have made your e-mail address available (<ka1stc@aol.com>) and, in the same issue (Fall 1998, page 33, "Complex Impedance Measurements"), Michael Gruchalla invited comments regarding complex impedance measurements to be sent to the editor, I will do so. Also attached are comments regarding Ian Poole's synthesizer article (Summer 1998, page 87, "Designing Frequency Synthesizers).

Thanks for the interesting article on measuring impedance. The technique seems fairly obvious once shown, but I would not have thought of it. I would encourage further effort into reducing it to practice. Maybe FAR Circuits or someone else might make a pc board available. I could see real value in a design which included a DSS generator and A/D converters, all connected to a low-cost micro. It might have an LCD display, or an RS-232 port. A couple of relays switching various R and X values might be useful.

One item concerned me regarding **Figure 4A**. I am not sure that the Vrxd output would read correctly if the load did not present a DC path to ground. It would seem that a simple fix would consist of reversing that section of the circuit, so that the 0.001 capacitor had a DC path back to the 75-ohm input termination resistor. One might argue for a similar reversal of the VrRd detector as well, just for symmetry.

Tracking of the various outputs could be optimized in software. Three of the outputs should be identical with no load attached. The VrRd should track Vsd if the output side of Rr were shorted, and shorting Rr and the output should make Vrxd track Vsd.

It seems that an NPO chip capacitor would be ideal for Xr, although, if the piezzo electric effect mentioned is significant, there are chip film capacitors available (see the DigiKey catalog for example).

I enjoyed the visual approach given for analyzing phase noise in synthesizers in Mr. Poole's article. I have had several occasions to design PLL systems, and this approach made the process more intuitive. There was one statement that I questioned, though, and one omission I would have liked to see included.

The statement, "The other major contributor...the phase detector...having the effect of the loop multiplication added," I do not believe is true. The phase detector received the divided down copy of the reference, and thus is after the divider. Its output (filtered) goes directly to the VCO, and any noise passing through the filter will affect the VCO directly, independent of the divide ratio.

What I didn't see was any mention of the effects of the reference frequency itself (not its noise). In many synthesizers, including most used in amateur radio, the reference is divided down so many times to get the fine level of granularity required that the output frequency of the divider is a major noise source, creating sidebands spaced that distance from the carrier on each side. The loop filter must be narrower than this, but the frequency is low enough that this noise source is often near the cut-off frequency of the filter, and thus not attenuated below the level of other noises. This is especially true where other factors, such as agility of the signal or stability of the VCO or even the component values require maximum filter bandwidth. I can't remember designing a PLL where this was not the dominant noise source. I missed seeing the characteristic ears in the spectrum shown in **Figure 6**, etc.

By the way, if you haven't run across it, the most exhaustive and authoritative source I have found on this topic is the book by Rhode. It covers all aspects of the design in great detail, and the accompanying computer disk has useful programs to calculate things like filter design.

Wilton Helm, WT6C
<whelm@compuserve.com>

Key Sentence Missing

Dear Editor:

Larry, VE3EDY, and I were pleased to see the highlight given our article entitled "A Tunable All-bands HF Camp Mobile Antenna," in the Fall 1998 issue of *Communications Quarterly* (pages 47–57).

The presentation, by and large, is well done. There are, however, two minor reproduction mistakes, which could be ignored (1 and 3 below); but a key sentence was left out (2 below), which requires pointing out.

1. On page 50, right column, middle paragraph:

"...this resistance can be changed from fast speed (band change) to slow speed (**fine** tune)."

2. In the section "Comparison with a popular mobile whip," a key sentence was left out on the top of page 54, left column:

"...for the same total antenna height; the measured radiation efficiency was 1.2 percent; i.e., -7.3 dB compared with the author's TABA mobile antenna."

3. In the section "Concluding remarks," the last sentence should read:

"...In fact, **VE3EDY** is continually surprised..."

Jack Belrose, VE2CV
<jack.belrose@dgbt.crc.ca>

A Few Small Errors

Dear Editor:

I am writing to report finding a few small errors in the class-E amp article ("Class-E Power Amplifier and Digital Driver for 160 Meters," Fall 1998, page 9) and have enclosed the corrections for future reference.

In the **Table 2** parts list for the VFO unit: **R103** should be 5-kohms not 5 ohms as listed. Also, in the **Figure 4** schematic of the VFO unit feedthrough capacitor, C104 was not labeled; it is the capacitor next to J102. In the second paragraph on page 17, the sentence should read: "Expand or compress L203 so

maximum output power occurs with the settings in **Table 1**."

The only serious error would be the value of R103. The circuit would not operate if a 5-ohm resistor was used here instead of a 5-kohm.

Thanks again for your kind help and care in publishing the article for me and my co-author Frederick H. Raab.

Todd Roberts, WD4NGG
Hilton Head Island, South Carolina

Network Impedance, Energy Transfer, and the Conjugate Match

Dear Editor:

In our article (*Communications Quarterly*, Fall 1997, pages 25–40) concerning the source impedance of tuned power amplifiers and the conjugate match, we tried to provide an understanding of the basic concepts and a clarification of misunderstandings. Apparently, judging by the correspondence received and a rebuttal written, we have not succeeded. In retrospect we used too many words, and the average reader of *CommQuart* got lost amongst the trees, loosing sight of the forest.

Bruene, reference his article (Spring 1998, pages 23–31), presents his view and states that we are wrong-wrong-wrong. There is little use in our debating his critiques point-by-point claiming that we are right-right-right. We each have had our say. ("We cannot solve a problem by using the same kind of thinking we used when we created them,"—Albert Einstein.)

Instead let us return again to the basics. In particular, let us look at the following definitions from the IEEE Standard Dictionary of Electrical and Electronic Terms, Fifth Edition, 1993 (Bruene, in his article, quoted some of these). The words in parenthesis are mine.

1. Source impedance. The impedance presented by a source of energy (e.g., an RF power amplifier) to the input terminals of a network.

2. The effective output impedance of an electron valve or tube (which from the above must be the source impedance) is the quotient of the sinusoidal component of output voltage by the corresponding component of current when it (the device) is operating normally, usually in steady state mode. Note: Output impedance is sometimes incorrectly used to designate load impedance.

RF power amplifiers are conventionally designed to match load impedances at or near to the characteristic impedance of common transmission lines. No other design would make sense.

3. The effective dynamic output impedance of a power supply (a device) is derived from the ratio of the measured peak-to-peak change in output voltage to a measured peak-to-peak change in load alternating current.

This is the Sabin experiment, which we have used to measure the output impedance Zout of an amplifier delivering design power to the design load impedance (our Fall 1997 article, fourth experiment, pages 35–36 and **Figure 3**); and to show how this output impedance varies with power output (Spring 1998 issue, pages 100–102). As pointed out by Sabin (Summer 1998 issue, pages 107–108), the change in load impedance must be quite small so that the P-tank circuit will not be significantly detuned. Our follow-on measurements to determine Zout used a 13-percent low change in load resistance, which will cause a small detuning effect, but not enough to invalidate an "estimate" of the "magnitude" of the output impedance.

The much larger changes in load resistance corresponding to our initial experiments, which have been criticized, impedance mismatches as large as 2:1 (**Figure 3** of our Fall, 1997, page 34 article), interestingly shows that Zout, attributable to a very small change in load impedance from the reference load, can be quite accurately determined. The reference load (which for our experiment was Rload = 53 ohms) is the load into which the amplifier was first adjusted to deliver maximum available design power. This is because the best fit curve to Zout (so determined) versus Rload is a straight line. While there is a scatter in this initial set of data, the referenced graph shows that Zout = Rload = 51.3 ohms for a reference load equal to 53 ohms.

4. Impedance mismatch factor—antennas. The ratio of power accepted by an antenna to the power incident at the antenna terminals. Note: The impedance mismatch factor is equal to one minus the magnitude squared of the input reflection coefficient of the antenna.

5. Load matching—circuits and systems—is the technique of either adjusting the load-current impedance or inserting a network (our antenna system tuning unit (ASTU)) between two parts of a system to produce the desired (maximum) POWER TRANSFER or signal transmission (Maximum Power Transfer Theorem). The Maximum Power Transfer Theorem is essentially an energy equation, which is an extremely powerful tool of physics used to avoid the pitfalls of nonlinear phenomena. Conservation of energy transcends all bounds. It is true for linear, nonlinear, periodic, and nonperiodic systems.

Simply stated, when the energy being transferred across any linearly behaving connection cannot be increased simply by changing the impedance of either source or load, a conjugate match exists.

The conjugate match theorem tells us that if and only if (IFF) the load impedance equals the conjugate of the source impedance the power transferred will be the maximum amount available from the source in its present configuration. If we make a change to the source then it is no longer the same source and a different value of the load impedance may be required to again reach a new maximum. The effective dynamic output impedance of an RF power amplifier changes as the power output changes (decreases) due to input drive decreasing, but not by the amount suggested by others. The change is least for Class A, and for Class AB and B amplifiers; and is larger for Class-C amplifiers.

6. Impedance, conjugate. An impedance the value of which is the complex conjugate of a given impedance. Note: For an impedance associated with an electrical network, the complex conjugate is the impedance with the same resistance component and a reactance component the negative of the original.

Our experiment 2 (reference the Fall 1997 issue, page 32), shows that if we tune an ASTU for maximum power transfer to a mismatched antenna system, we can then replace the amplifier, connecting to the input of the ASTU, by a real impedance, whose value is the equal to the source or output impedance of the amplifier. Complex conjugate impedances can then be measured by measuring the impedance looking back into the output terminals of the ASTU, and the input looking forward into the input terminals of the antenna system. Keep in mind that one of the tenets of the conjugate match theorem is that once there is a conjugate match anywhere in the system there must be a conjugate match everywhere in the system.

7. Resistance—network analysis. (A) That physical property of an element, device, branch, network or system that is a factor by which the mean-square conduction current must be multiplied to give the corresponding power loss by dissipation or heat, or as other permanent radiation loss, or loss of electromagnetic energy from the circuit. (B) The real part of impedance. Note definition (A) and (B) are NOT equivalent, but supplementary. In any case where confusion may arise, specify definition being used. Our view: Since the impedance of a network deals with energy transfer, it has nothing to do with where the energy came from, or what became of it. The real part Re of this impedance, therefore, does not dissipate energy of itself. It is for this reason that we, the authors, chose to call this

impedance a dissipation-less impedance, implying definition (B) above. That is, power is not dissipated in the amplifier's output impedance Zout (which is measurable), power is available there for transfer to the antenna, where it is radiated. Also, the antenna's radiation resistance Rr (which is also measurable), in accord with my way of thinking, in spite of the fact EM radiation appears to be a "loss" of power (Definition (A)), can be considered to be a lossless impedance, since the Ia2 Rr power delivered to that resistance is not dissipated but radiated (power transfer antenna to propagation medium).

Finally, a comment on matching: Bruene (in one of his initial critiques) states that "a transmission line is matched when its load impedance is equal to the characteristic impedance of the transmission line (say Zo equal to 50 ohms). The source resistance (please source impedance) at the line input has nothing to do with the matched conditions." This statement is true if we are determining "match" by measuring the line input impedance using an RF bridge, or a SWR bridge designed for 50-ohm systems. But if we want to transfer all the available power from the amplifier to the (mismatched) antenna (with zero mismatch loss achieved by the reflection gain obtained with an ASTU matching network) we need to tune the antenna system, with its connected transmission line, so that the input impedance of the transmission line connecting the amplifier to the ASTU is equal to the output, or source impedance of the amplifier. When this condition is met, we indeed have a conjugate match in the entire system, including the RF amplifier.

The experiments we performed, the terms we have used, and the conclusions we reached are all, in our view, in accord with the IEEE way of thinking, when the various contributors to the Dictionary defined the above referenced terms.

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VSWR, REFLECTIONS, AND THE "CONJUGATE" IMPEDANCE MATCH

Reflections and their impact

Any discussion of VSWR, reflections, and mismatch loss is sure to stir up disagreement and confusion within the amateur radio community. The concepts of VSWR, reflections, mismatch loss, reflection gain, and total power delivered to an antenna are discussed repeatedly, yet they are often misunderstood.

This is an examination of reflections and their impact on the power delivered to an antenna located at the end of a transmission line. As background to this discussion, I present a detailed review of fundamental transmission line concepts, including VSWR, impedance, mismatch loss, and power relations.

I also provide a detailed discussion of the operation of a "conjugate" impedance matching network or tuner, including an analysis of all of the reflection interactions that occur between the matching network output and the antenna.

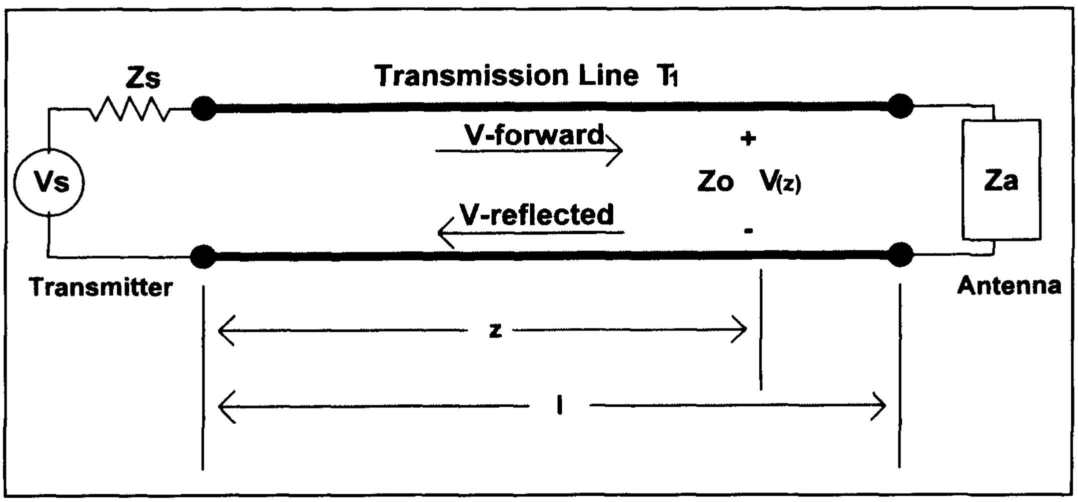

Figure 1. Block diagram for a typical transmission circuit.

The method by which reflections are canceled at the matching network is derived mathematically and validated with an example. It's important to note that I make no initial assumptions in this analysis regarding the method by which the reflections are canceled.

Finally, I present the results of an experiment that verify the concepts presented here.

Transmission line fundamentals

In any communications circuit, the system antenna is defined as being "matched" to the system transmission line when no signal (voltage) reflections exist at the connection point between the transmission line and the antenna.

Any antenna, that has a feedpoint impedance ($Z_A = R_A + jX_A$) different from the characteristic impedance ($Z_O = R_O + jX_O$) of the transmission line to which it is connected, creates an impedance mismatch at the connection point to the transmission line. This impedance mismatch results in a reflected voltage created at the antenna. This reflected voltage travels back toward the transmitter. At any point along the length of the transmission line, the total voltage is the sum of the voltage contained in the forward traveling wave(s) and the voltage contained in the reflected traveling wave(s). The parameter used to characterize the level of impedance mismatch between an antenna and the transmission line is the voltage standing wave ratio, or VSWR. The VSWR is the ratio of the maximum voltage along the length of the transmission line to the minimum voltage along the length of the transmission line.

Consider the block diagram presented in **Figure 1**. In this setup, a transmitter is connected directly to an antenna via transmission line T1. The voltage at point z along the length of the transmission line is found using **Equation 1**:[1]

$$V(z) = V_1 e^{-\gamma z} + V_2 e^{+\gamma z} \qquad \textbf{(1)}$$

Where V_1 and V_2 are phasor voltage coefficients whose values are determined by the voltage and internal source impedance Z_S of the transmitter at $z = 0$, the attenuation (α) and phase (β) properties of the transmission line, the transmission line length in meters (l), and antenna impedance Z_A, which is connected to the transmission line at $z = l$.

Voltage V_1 is the initial voltage delivered to the transmission line by the transmitter and voltage V_2 is created by the reflection at the antenna. Note that $\gamma = \alpha + j\beta$, where α is the transmission line attenuation in nepers per meter and β is simply $2\pi/\lambda$. The wavelength in the transmission line, λ, is expressed in meters. Calculate nepers per meter from dB/100 feet by multiplying the dB/100 feet by 3.7772×10^{-3}.

If the antenna impedance Z_A is not equal to the characteristic impedance of the transmission line Z_O, a reflected voltage is created at the antenna. The value of the reflected voltage at the antenna is determined from the antenna reflection coefficient, ρ_A, found from **Equation 2**:

$$\rho_A = \frac{\text{(value of the reflected voltage wave at the antenna)}}{\text{(value of the incident voltage wave at the antenna)}}$$

$$\rho_A = (Z_A / Z_O - 1) / (Z_A / Z_O + 1) \qquad \textbf{(2)}$$

The value of the reflected voltage at the antenna is equal to the incident voltage at the antenna multiplied by the antenna reflection coefficient. The total voltage applied at the antenna is the sum of the incident voltage plus the reflected voltage. Once ρ_A is known, we can determine the relationship between V_1 and V_2 as follows:

$$V_2 = V_1 \rho_A e^{-2\gamma l} \qquad \textbf{(3)}$$

The antenna VSWR may also be calculated from the antenna reflection coefficient, ρ_A, using the following equation:

$$\text{VSWR} = (1 + |\rho_A|) / (1 - |\rho_A|) \qquad \textbf{(4)}$$

On any given transmission line, the magnitude of the voltage maximum can be found by multiplying the magnitude of the voltage minimum by the VSWR. Remember that the voltage maximum and/or the voltage minimum may not occur at the antenna or at the transmitter.

It is important to note that if the magnitude of the reflection coefficient is 0 (Z_A is equal to Z_O), there is no reflected voltage. Rather, the entire power incident at the antenna is delivered to the antenna. Also, if the magnitude of the reflection coefficient is 1 (Z_A is an open circuit, short circuit or purely reactive), the magnitude of the reflected voltage is equal to the magnitude of the incident voltage, and no power is delivered to the antenna. If the magnitude of the reflection coefficient is between 0 and 1, some power is delivered to the antenna and some power is reflected. This is true for both the forward voltage incident at the antenna and the reflected voltage incident at the transmitter. The amount of reflected voltage or power at either end of the transmission line depends solely upon the terminating impedance and the characteristic impedance of the transmission line.

For any incident voltage arriving at the antenna, the power reflected is equal to the incident power times $|\rho_A|^2$ and the power delivered to the antenna is equal to the incident power times $(1 - |\rho_A|^2)$. The reduction in delivered power due to the reflection at the antenna is defined as reflection mismatch loss. The mismatch loss in dB is found using **Equation 5**:

$$\text{Mismatch Loss (dB)} = 10 \log(1 - |\rho_A|^2) \qquad \textbf{(5)}$$

I will illustrate the significance of mismatch loss in determining the total power delivered to an antenna in subsequent sections of this article.

Consideration for multiple reflections in a transmission line

Once a reflected voltage is created at the antenna, it travels back toward the transmitter. If the transmitter impedance, as seen by the voltage reflected from the antenna, is identical to the characteristic impedance of the transmission line (transmitter VSWR is equal to 1.0:1), which may not be generally true in practice, then all of the power contained in the reflected voltage is delivered to the transmitter. In this case, the effective net power delivered to the transmission line by the transmitter is equal to the forward power delivered to the transmission line less the reflected power delivered to the transmitter. This ensures that the law of conservation of energy isn't violated.

If the transmitter impedance is not identical to the characteristic impedance of the transmission line, which is true in most amateur communication circuits, some of the reflected voltage originating at the antenna is reflected from the transmitter back towards the antenna. This results in multiple reflections in the transmission line. Any power not reflected back towards the antenna is delivered to the transmitter. The amount of power reflected back toward the antenna is a function of the transmitter reflection coefficient.

Once the re-reflected voltage arrives at the antenna, another reflected voltage is created, which again travels back toward the transmitter. Any power not reflected by the antenna is delivered to the antenna for radiation. This process of multiple reflections continues until the reflections are delivered to the transmitter, delivered to the antenna, or dissipated in the transmission line due to transmission line attenuation. In order for all of the incident voltage or power to be reflected from either the antenna or the transmitter, the antenna impedance or the transmitter impedance must be an open circuit, short circuit, or purely reactive—where the reflection coefficient magnitude is equal to 1 and the VSWR is equal to ∞.

To fully analyze the voltages in a system where multiple reflections exist between the antenna and the transmitter, **Equation 1** is modified to include all of the voltage contributions from the initial voltage delivered by the transmitter and the voltages in each of the individual reflections.

Remember that, with multiple voltage reflections on a transmission line, the voltage re-reflected at the transmitter arrives back at the antenna with a time delay, is attenuated, and has a phase change relative to the initial voltage incident at the antenna. These factors are a function of the transmitter reflection coefficient and the transmission line length, attenuation, and phase properties. The interaction of these multiple reflections on a transmission line may have several results; i.e., some level of signal cancellation, echoing, and/or interference may occur.

Due to the nature of HF amateur communications and the extremely short time involved for all the reflection interactions to occur, most communications links do not experience much, if any, signal quality degradation. However, as the antenna VSWR relative to the transmission line increases, these factors may become an important issue when attempting to optimize the quality of the communication link.

High VSWR does impact other aspects of system performance. A high VSWR periodically creates higher levels of voltage and current along the length of the transmission line. This results in an increase in the amount of power lost in the transmission line due to the increased power loss at points of maximum voltage and current along the line. At points along the line where the forward traveling and reflected traveling voltage waves add in phase, there is a voltage maximum. There is a voltage minimum where the two waves add "out-of-phase." At points of maximum voltage, power loss occurs in the conduction loss in the dielectric. At points of maximum current, the power loss occurs in the resistance of the conductors. If the power loss at the points of maximum voltage and current are too high, the transmission line may overheat and actually "burn" through.

The issue with higher voltages is that, at both the antenna and/or the transmitter, higher voltages may exist than would where the antenna impedance is matched to that of the transmission line. The increased voltage that may occur at either the antenna or transmitter may be sufficient enough to cause damage to the transmitter or the antenna (primarily antenna feedpoint matching components). Note that most manufacturers specify device power-handling capability under a matched condition.

Impedance and VSWR at the input to the transmission line

When a transmission line is connected to an antenna, the input impedance to the transmission line depends upon the transmission line length, and the transmission line attenuation and phase properties. The transmission line input impedance is found using **Equation 6**:

$$Z_{IN} = Z_O[(Z_A/Z_O + \tanh(\gamma l))/(1 + (Z_A/Z_O)\tanh(\gamma l))] \qquad (6)$$

Substituting Z_{IN} for Z_A in **Equation 2**, the input VSWR of the transmission line can be determined using **Equation 4**. With any level of transmission line attenuation greater than 0 dB, the input VSWR to the transmission line is less than the antenna VSWR.

In regard to transmission line input impedance and VSWR, input impedance and VSWR cannot be measured or determined until the voltage and current reflections from the antenna arrive back at the input to the transmission line. The total voltage (and current) at the input to the transmission line is equal to the initial voltage (and current) delivered by the transmitter plus the voltage (and current) arriving back at the transmission line input due to the reflection(s) from the antenna. Input impedance is then determined by dividing the total voltage at the transmission line input by the total current at the transmission line input.

Reflections and forward power delivered to the transmission line

When the transmitter is first energized, the only impedance it "sees" at the input to the transmission line is the characteristic impedance of the transmission line, Z_O. The initial voltage, V_I, delivered to the transmission line input, is equivalent to the output voltage the transmitter would develop if a load with an impedance of Z_O were placed across the transmitter terminals. The initial forward power delivered to the transmission line is therefore equivalent to the power the transmitter would deliver to a terminal load with an impedance of Z_O.

After the reflection from the antenna arrives back at the transmitter, a reflected voltage is created at the transmitter, which travels back toward the antenna increasing the total forward traveling voltage (and power) in the transmission line. The level of the reflected voltage (and power) at the transmitter is a function of the transmitter reflection coefficient relative to the characteristic impedance, Z_O, of the transmission line. This process of multiple reflections between the antenna and the transmitter continues as long as "initial" forward voltage is delivered to the transmission line by the transmitter. The condition where the transmitter does not first "see" any reflections from the antenna is called the initial state within the system. The condition where all reflections within the system have reached an equilibrium is called the steady state.

In the steady-state condition, the level of total forward voltage (and power) at the transmission line input increases beyond that of the initial forward voltage (and power) as a result of the voltage from the reflections at the transmitter output being added to the initial forward voltage. Also, the level of reflected power arriving at the transmitter output increases due to the additional voltage from the reflections at the antenna. In this steady-state condition, the effective net power delivered to the transmission line is equal to the total forward power delivered into the transmission line minus the reflected power returned to the transmitter.

The transition from the initial state to steady state is an ongoing process, as the initial voltage, V_I, is continuously delivered to the transmission line by the transmitter. In the steady-state condition, multiple reflections continue to exist as forward voltage is delivered by the transmitter. However, each individual reflection within the system decays because a portion of it is delivered to the antenna or the transmitter, or is dissipated in the transmission line due to transmission line attenuation.

Power delivered to an antenna

Many misconceptions in analyzing reflections within a transmission line are related to the simple subject of power. When multiple reflections exist between the antenna and the transmitter, a number of these reflections arrive simultaneously at the antenna. The total power delivered to the antenna cannot be determined by simply summing the power delivered to the antenna by each individual voltage or power arriving at the antenna.

From **Equation 1**, the total voltage at the end of the transmission line is found as follows:

$$V(l) = V_I e^{-\gamma l}(1 + \rho_A) \qquad (7)$$

The current delivered to the antenna is essentially the difference between the current incident at the antenna and the current reflected at the antenna. This assures that the principle of conservation of energy is not violated.

$$I(l)=(V_I e^{-\gamma l} - \rho_A V_I e^{-\gamma l})/Z_O \qquad (8)$$

The current into the antenna may also be found by dividing the total voltage at the antenna, V(l), by the antenna impedance, Z_A.

$$I(l) = V(l)/Z_A \qquad (9)$$

Once V(l) and I(l) are known, the actual power delivered to the antenna is determined from the following equation:

$$P_{antenna} = |V(l)|\ |I(l)|\ Cos(\theta) \quad \textbf{(10)}$$

Where θ is the phase angle of the antenna impedance, Z_A. Theta (θ) can also be found from the phase angle difference between the total voltage and current.

$$\theta = arg[V(l)] - arg[I(l)] \quad \textbf{(11)}$$

If the antenna impedance is purely resistive, **Equation 10** may be simplified:

$$P = |V|^2/R \quad \textbf{(12)}$$

For example, consider a voltage of 100 + j0 volts arriving at an antenna with an impedance of 50 + j0 ohms. Assume that the characteristic impedance of the transmission line is also 50 + j0 ohms. Because there is no reflected voltage, the total voltage at the antenna is 100 + j0 volts. Using **Equation 12**, the total power delivered to the antenna is calculated to be 200 watts. Now, consider a voltage of –100 + j0 volts arriving at the same antenna. Again, using **Equation 12**, the total power delivered to the antenna is found to be 200 watts. Third, consider a voltage of 0 + j100 Volts arriving at the same antenna. Using **Equation 12**, the total power delivered to the antenna is again 200 watts.

Now consider what happens if two or more of these voltages arrive simultaneously at the same antenna. If all three voltages arrive at the antenna at the same time, will 600 watts of power be delivered to the antenna? In this case, the total voltage at the antenna is 0 + j100 volts. The total power delivered to the antenna is only 200 watts, not 600. The 100 + j0 and the –100 + j0 voltages canceled each other, and no net power is delivered to the antenna from these voltages. If the 100 + j0 and the 0 + j100 voltages arrive at the antenna, will 400 watts of power be delivered to the antenna? The total voltage at the antenna is 100 + j100 volts and, using **Equation 12**, the total power delivered to the antenna is indeed found to be 400 watts. Finally, consider that two voltages, each 100 + j0 volts, arrive at the antenna. Will the total power delivered to the antenna be 400 watts? Using **Equation 12**, the total power delivered to the antenna is actually found to be 800 watts.

This illustration is critical to understanding the power interactions occurring within a system where an impedance matching network or tuner is used between the transmitter and the antenna. When an impedance matching network is used, multiple reflections exist between the antenna and the matching network. Each reflection simultaneously arriving at or leaving the antenna and the matching network has its own individual power level. However, the total power arriving at or leaving the antenna and the matching network may be different from the sum of the individual powers because total power is a function of $|V_{TOTAL}|^2$ and cannot be determined by simply adding individual powers together.

VSWR increases transmission line loss

As VSWR increases, the total power loss in the transmission line increases. In a transmission line system terminated by an antenna with an impedance equal to the characteristic impedance of the transmission line, no reflections are created and all power incident at the antenna is delivered to the antenna. The only power lost between the transmitter and the antenna is that lost due to the attenuation in the transmission line. This attenuation is defined as the matched attenuation of the transmission line. For example, RG-213 coaxial cable has a matched attenuation of approximately 1.05 dB/100 feet at 21.2 MHz.

In a system where the antenna VSWR is greater than 1.0:1, it is common knowledge that the transmission line losses increase due to the increase in VSWR. Let's consider this concept further.

On page 19.4 of *The 1998 ARRL Handbook*, an equation is presented for total mismatched line loss due to increased VSWR.* The equation for total line loss is:

$$\text{Total Line Loss (dB)} = 10 \log[(\alpha^2 - |\rho|^2)/\alpha(1 - |\rho|^2)] \quad \textbf{(13)}$$

where α is the matched line attenuation expressed as $10^{\alpha \text{ in dB}/10}$ and ρ is the antenna reflection coefficient.

As I mentioned earlier, when a transmit circuit is first energized, it "sees" only the characteristic impedance of the transmission line, Z_O. The antenna VSWR is not evident to the transmitter until the reflection from the antenna arrives back at input to the transmission line. When the initial voltage delivered by transmitter V_I reaches the antenna, it has been attenuated by only the matched transmission line attenuation, regardless of antenna VSWR. Any voltage, hence power, reflected by the antenna travels back toward the transmitter. Traveling toward the transmitter, the voltage reflected by the antenna will also be attenuated by only the matched transmission line attenuation, regardless of VSWR.

Consider the following example. Let's make an impractical assumption that the transmitter has a source impedance of 50 ohms, so any reflections from the antenna will not be re-reflected by the transmitter. This assumption does not invalidate the concepts presented below but makes them easier to illustrate. Assume the transmission line connecting the antenna to the transmitter is 100 feet of RG-213 coaxial cable. Also assume the operating frequency is 21.2 MHz. This results in a total one-way attenuation of 1.05 dB between the transmitter and the antenna. Now suppose the antenna connected to the transmission line has a feedpoint impedance of 150 + j0 ohms. This antenna has a VSWR of 3:1 and a reflection coefficient of 0.5.

First, let's use **Equation 13** to determine the total line loss in dB. With a reflection coefficient of 0.5 and a matched attenuation of 1.05 dB, the total line loss is calculated as 1.572 dB.

Now assume the transmitter initially delivers a forward traveling voltage, V_1, of 100 volts into the transmission line. This is equivalent to 200 watts of forward power delivered to the transmission line ($|V|^2/Z_O$). With 1.05 dB of transmission line attenuation, 42.953 watts of forward power is lost traveling from the transmitter to the antenna. Therefore, only 157.047 watts of forward power arrives at the antenna. Based upon the antenna reflection coefficient of 0.5, 39.262 watts of power is reflected at the antenna and 117.785 watts of power is delivered to the antenna. This 117.785 watts of power delivered to the antenna out of the 157.047 watts incident power at the antenna is consistent with the 1.25 dB mismatch loss associated with a 3:1 VSWR (calculated using **Equation 5**). Because the one-way attenuation is 1.05 dB, the reflected power arriving back at the transmitter is only 30.83 watts. The amount of reflected power lost traveling back towards the transmitter is 8.432 watts.

In the analysis above, the one-way cable attenuation did not increase from the 1.05-dB matched attenuation. Of the 200 watts of forward power delivered at the transmitter output, 51.385 watts total power is lost in the transmission line. The 51.385 watts of power lost in the transmission line is a result of the power lost from the 200 watts traveling in the forward direction (42.953 watts lost) and the power lost from the 39.262 watts traveling in the reverse direction (8.432 watts lost). Of the 200 watts of forward power delivered into the transmission line, only 117.785 watts of power is delivered to the antenna. This represents an effective total loss of 2.3 dB (10 log[117.785/200]), which is the sum of the 1.05 dB of one-way attenuation and the 1.25-dB reflection mismatch loss. This example illustrates how mismatch loss can be used to determine the total power delivered to the antenna. In any system, even with multiple reflections present, the power delivered to the antenna can be determined from the total forward power at the input to the transmission line, the one-way matched attenuation, and the reflection mismatch loss.

If the conclusions regarding mismatch loss and line attenuation are true, how does the 1.572-dB of total line loss calculated from **Equation 13** relate to the power discussion? Consider an alternative approach to describing the power relations in the above example. Based upon the 200 watts of forward power delivered by the transmitter and the 30.83 watts of reflected power returned to the transmitter, the effective net power delivered to the transmission line is only 169.17 watts (forward power minus reflected power).

Of the 169.17 watts effective net power delivered to the transmission line, 117.785 watts of power is delivered to the antenna. The loss between the net power delivered to the transmission line and the power delivered to the antenna is therefore 10 log (117.785/169.17), which is 1.572 dB. The 51.385 watts of power lost in the transmission line, as determined above, is consistent with this 1.572 dB loss. When the power delivery in the transmission line is analyzed using effective net power, the 1.572-dB loss of power in the transmission line is the defined total transmission line loss as determined by **Equation 13**. In a practical system, the effective net power delivered to the transmission line can be determined from the total forward power delivered to the transmission line and the total reflected power at the input to the transmission line.

What does the above discussion illustrate? First, the discussion illustrates the relationship between forward power, reflected power, and the effective net power delivered to the transmission line. The relationship between these three powers is a function of the one-way matched transmission line attenuation and the VSWR reflection mismatch loss. The discussion also illustrates the relationship between transmission line attenuation, mismatch loss, and the total attenuation loss concept of **Equation 13**.

Equation 13, illustrating the concept of increased transmission line loss due to VSWR, is really a determination of the total loss in the transmission line when considering the effective net power delivered to the transmission line, not the forward power. The actual one-way attenuation in the transmission line does not really increase for forward or reflected power traveling in the transmission line. Additionally, the discussion shows that mismatch loss does directly impact power delivered to an antenna. Finally, it was demonstrated that there are two somewhat unique but valid methods for determining the total power delivered to the antenna.

In the first method, knowing the forward power delivered to the transmission line, the power delivered to the antenna is found by subtracting the losses due to the one-way matched line attenuation and the VSWR reflection mismatch loss. In the second method, knowing the effective net power delivered to the transmission line (forward power minus reflected power), the power delivered to the antenna is found by subtracting the loss as calculated from the total line loss formula presented in **Equation 13**. Note that when using **Equation 13** to determine power delivered to the antenna, no additional loss due to reflections at the antenna should be considered because they are accounted for in the determination of effective net power delivered to the transmission line.

In practice, we can use both of the above methods for determining the net power delivered to an antenna. Additionally, if the transmitter impedance is not equal to the characteristic impedance of the transmission line, a number of multiple reflections exist between the transmitter and the antenna. In this case, the total forward and reflected powers at the transmitter output increase due the increased magnitudes of the forward and reflected voltages. Because a wattmeter determines power based upon the total forward or reflected voltage in the transmission line, it provides an accurate measurement of total forward and reflected power. Having these wattmeter readings available, the total power delivered to the antenna can be determined if the matched line loss and the antenna VSWR are known.

The "conjugate" impedance match

To describe the operation of an impedance matching network or tuner, it is necessary to perform a detailed analysis of all voltage reflections and interactions between the matching network and the antenna. The following discussion makes no initial assumptions or conclusions regarding the method by which reflections are prevented from traveling back towards the transmitter. The discussion and analysis presented here simply determines all steady-state voltages within the system. Once these steady-state voltages are determined, the method by which reflections are prevented from traveling back towards the transmitter is clearly evident.

Let's consider a practical example. In this case, an antenna with a feedpoint impedance of 6.2 + j14.4 ohms will be matched to the characteristic impedance of the transmission line using an impedance matching network. This feedpoint impedance value was selected because it is the feedpoint impedance of the antenna used in the experimental work. To be consistent with the experimental work, an operating frequency of 21.2 MHz and a matched total transmission line attenuation of 1.64 dB is used in the example. For calculation purposes, the transmission line between the matching network and the antenna is assumed to be 156.19 feet of RG-213 coaxial cable. This length of coaxial cable results in the same 1.64-dB total attenuation consistent with the experimental work.

Now let's consider what happens when we attempt to match the antenna's impedance using a "conjugate" matching network located at the input to the transmission line. Conjugate match theory is based on the Thevenin equivalent circuit with a voltage source, having a complex source impedance equal to the complex characteristic impedance of the transmission line, being connected to a complex load Z_O^*. The conjugate match theory states that more power is delivered to a terminal load impedance Z_O^* that produces a reflected wave on the line than to a terminal load impedance Z_O that produces no reflected wave.

Antenna impedance matching is different in that a lumped circuit element impedance network can be used to "match" an antenna's impedance to the characteristic impedance of the transmission line. If this matching network is located at the antenna, the objective is to match the antenna's feedpoint impedance to the characteristic impedance of transmission line creating a steady state VSWR equal to 1.0:1. If the impedance matching network is located at the input to the transmission line, the matching network matches the transmission line input impedance to the characteristic impedance of the transmission line.

If the matching network is located at the input to the transmission line, the steady-state load impedance seen by the network will be the steady-state input impedance to the transmission line. In general, the input impedance to the transmission line is of the form $R_{IN} + jX_{IN}$, however, it is unlikely that R_{IN} will be equal to the characteristic impedance of the transmission line. In this case, it is necessary for the matching network to cancel out the reactive component of the input impedance while at the same time bringing the resistive component of the input impedance to that of the transmission line characteristic impedance.

As part of this discussion, consider the circuit configuration in **Figure 2**.

Where:

T1 is the transmission line connecting the transmitter to the matching network.

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*Note: There are several typos in the formula presented in the handbook.

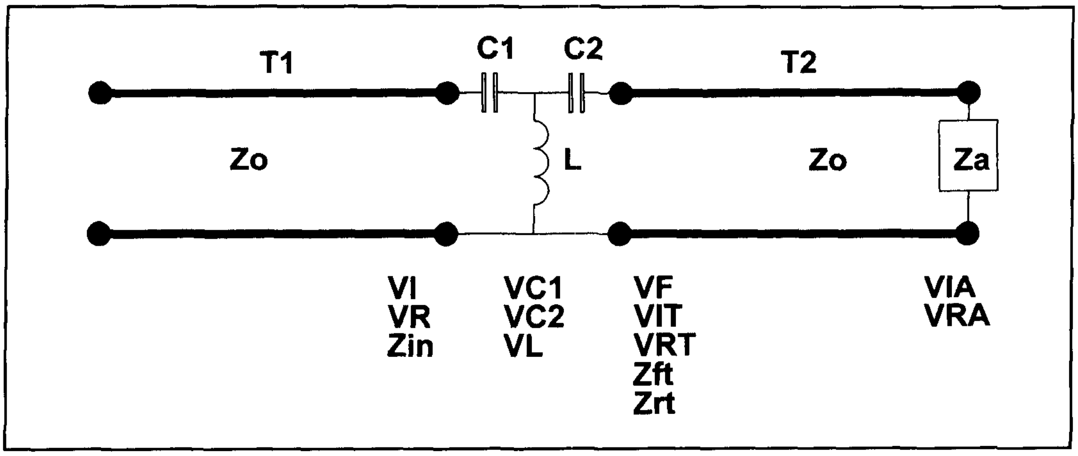

Figure 2. Typical lumped circuit element impedance matching network.

T2 is the transmission line connecting the matching network to the antenna.
VI is the incident voltage arriving at the matching network from the transmitter.
VR is the reflected voltage at the matching network traveling back towards the transmitter.
Zin is the input impedance to the matching network seen from transmission line T1.
VC1 is the voltage across capacitor C1.
VC2 is the voltage across capacitor C2.
VL is the voltage across inductor L.
VF is the initial forward voltage delivered from the matching network into transmission line T2. This will be the voltage phasor V_1 in **Equation 1**.
VIT is the incident voltage at the matching network arriving from transmission line T2. This voltage would have been reflected from the antenna.
VRT is the reflected voltage from the matching network delivered back into transmission line T2. This voltage is the reflected portion of voltage VIT.
Zft is the input impedance to transmission line T2 as seen by the output of the matching network.
Zrt is the input impedance to the matching network as seen from transmission line T2.
VIA is the incident voltage arriving at the antenna.
VRA is the reflected voltage from the antenna.
Za is the antenna impedance.

Antenna load impedance Za is 6.2 + j14.4 ohms. To determine the component values of the impedance matching network, we must first determine the steady-state load impedance seen at the output of the matching network. With the 156.19 feet of transmission line and 1.64-dB total attenuation, the steady-state input impedance seen by the matching network is 36.984 +j54.335 ohms. Again, the operating frequency is 21.2 MHz and the velocity factor for the transmission line is 0.6594.

One set of matching network component values capable of matching this impedance to the characteristic impedance of the transmission line is C1 = 69.229 pF, C2 = 50 pF and L = 0.45 µH. It is assumed that a 100-watt transmitter is used to deliver power into the transmission line system. Therefore, the initial forward voltage delivered by the transmitter into transmission line T1 is 70.711 volts. In this example, the transmitter is assumed to have an impedance of 50 ohms. This does not affect the concepts or conclusions presented here, but it does affect the concept of the "conjugate" impedance match. Additionally, for the purposes of this discussion, assume that the initial voltage arriving at the matching network input is 70.771volts.

The discussion of how the matching network eliminates reflections requires a detailed analysis of the voltages and currents that exist throughout the transmission system. This, in turn, requires determination of the voltages and currents at the input to the matching network, inside the matching network at each component, the voltages on the transmission line between the matching network and the antenna, and the voltages and currents that exist at the antenna. I've detailed this information in 10 steps. Each follows the MathCad® example presented.

Step 1: Determine the input impedance, Zft, to transmission line, T2, so the matching network components can be designed. Using **Equation 7**, the input impedance, Zft is found to be 36.984 + j54.335 ohms.

Step 2: Once we know the input impedance, Zft, to the transmission line, we design a suitable fixed component matching network. One design solution has the following component values: C1 = 69.229 pF, C2 = 50 pF, and L = 0.45 µH.

Step 3: Once the impedance matching network is installed in the system, determine the impedance of the matching network, as seen by transmission line T2. This impedance, Zrt, is 36.984 – j54.335 ohms. This impedance is the conjugate of Zft, hence, the term "conjugate impedance match." Note that the existence of the conjugate condition is entirely dependent upon the 50-ohm impedance seen from the matching network input looking into transmission line T1.

Step 4: To perform a complete voltage and current analysis in the system, it is necessary to determine the input impedance and reflection properties of the matching network in the "initial" state. This is Zin to the matching network before any reflections arrive back at the matching network from the antenna. The initial Zin is calculated to be 16.872 – j18.074 ohms. This impedance represents a 3.393:1 VSWR and the initial input reflection coefficient is 0.545.

Step 5: Now that we know the initial Zin, we can determine the incident, reflected, and total voltage initially delivered to the matching network. With 70.711 volts incident at the matching network, the reflected voltage, VR, is –27.831 – j26.634 volts; therefore, the total initial voltage at the matching network input is 42.88 – j26.634 volts. This is the vector sum of the incident and reflected voltage.

Step 6: Once we know the total voltage at the matching network input, we can find the component voltages and currents in the matching network. We can also determine the initial forward voltage, VF, delivered to the input of the transmission line, T2, and the initial power delivered to both the matching network and the transmission line. Based upon the voltage at the input to the matching network, the initial power delivered to both the matching network and the transmission line is 70.322 watts. The same power is delivered to both the matching network and the transmission line because, in this example, the matching network is assumed to be lossless. VF is found to be –57.568 + j14.214 volts.

Step 7: Now that we have calculated VF, we can determine the voltage, current, and power initially delivered to the antenna. Based upon the reflection properties of the antenna (ρa = 0.795), 17.759 watts of power is initially delivered to the antenna for radiation. At this single instant in time, the 100 watts of initial forward power arriving at the matching network results in 17.759 watts of power radiated by the antenna.

Step 8: Now we must determine the steady-state conditions in the system. The voltage reflected at the antenna will arrive back at the output of the matching network. This incident voltage creates a reflected voltage at the matching network that travels back toward the antenna. Arriving at the antenna, this voltage again creates a reflection based upon the antenna reflection properties. This process continues until the voltage decays as a result of delivery into the matching network, delivery into the antenna, and dissipation in the transmission line. If we sum all of the incident and reflected voltages at the matching network output and at the antenna, we can find the steady-state condition(s). This is illustrated in Step 8 of the MathCad example.

Based upon the steady-state analysis, the input impedance to transmission line T2 is verified by dividing the steady-state voltage at the matching network output by the steady-state current at the matching network output. The steady-state power delivered to the antenna is 35.912 watts. In this example, the forward and reflected power measured by a typical wattmeter is also found in Step 9. The steady-state forward power at the input to the transmission line is 142.203 watts and the steady-state reflected power at the input to the transmission line is 42.203 watts. The wattmeter is located at the output of the matching network. In this case, the effective net power delivered to the transmission line is 100 watts.

This behavior of increased steady-state forward, reflected, and delivered power to the antenna led to the concept known as "reflection gain," a term originated by Philip H. Smith, the developer of the Smith Chart. This concept is also discussed in the text *Reflections: Transmission Lines and Antennas*[2] written by Walt Maxwell, W2DU. In fact, these increased levels of forward and reflected power can be measured by a wattmeter. However, it must be clearly understood that, in the steady-state condition, the net total power delivered to the antenna is the sum of the initial incident power at the antenna (which is not the transmitter output power) plus the power from all of the reflections that resulted from initial incident powers that arrived at the antenna earlier.

Consider an infinitely narrow time pulse of 100 watts average power and the setup described in the example above. The pulse initially arrives at the input to the matching network, where 70.322 watts of power is delivered to the matching network and the input to the transmission line. With 1.64-dB loss in the transmission line, 48.205 watts of power arrives at the antenna. However, due to the reflection mismatch loss, only 17.759 watts of power is radiated.

The 30.266 watts of power reflected from the antenna is continuously reflected back and forth between the antenna and matching network until it decays to 0 power. In this process, some power is delivered back into the matching network (it does not have an infinite VSWR), some power is lost in the transmission line, and some power is radiated by the antenna. Since the pulse has an infinitely narrow time duration, the original 17.759 watts of power delivered to the antenna will have been radiated while the reflected power was traveling between the antenna and matching network. The total radiation from the antenna would therefore be a series of pulses, or a pulse stream, with each successive pulse having a decreased amplitude and relative phase shift. The important concept learned from this illustration is that for every time instant that 100 watts of power is delivered to the input of the matching network, only 17.759 watts is radiated by the antenna.

In the steady-state condition developed in this example, the steady-state voltage delivered to the antenna is comprised of an incident signal plus the summation of the reflections or echoes from previous signals arriving at the antenna. **Considering the initial power delivered to the antenna, the total steady state power delivered to the antenna consists of 17.759 watts of signal power and 18.513 watts of echo power**. The term "reflection gain" is somewhat misleading under these conditions because the initial signal delivered to the transmission line is not undergoing a "gain" process but, rather, is simply having the reflections from other previous signals added to it. These echo signals differ in their content, amplitude, and phase, resulting in the radiation of some form of a combined signal.

The practical significance of this in amateur communications is another issue. In some communication systems, such as those containing video information, echo signals can corrupt the quality of the communication link by creating ghosts in the picture with each reflection. In practice, this is avoided through good antenna and system design. In amateur communications, the signal quality is affected more by atmospheric conditions and "message" corruption; the interference from echoes in the transmission line, is most likely not a significant problem.

Step 10: The last step in the characterization of voltages within the system is to determine the steady-state voltage arriving back at the input to the matching network due to all of the multiple reflections in transmission line T2. Reflections from the antenna arrive at the matching network output and develop a steady-state voltage across the network output. This voltage is applied across the matching network components and therefore, through simple circuit theory, some level of voltage is developed at the matching network input. If we know the total voltage at the matching network output from the multiple reflections, we can determine the voltage at the matching network input. The steady-state voltage at the matching network input, due to the reflections in the system, is the negative of the initial reflected voltage at the matching network input. This voltage is 27.83 + j26.634 Volts. This shows that, in the steady-state condition, a power level equal to the initial reflected power, but 180 degrees out of phase, is developed at the tuner input. **This causes a cancellation of the initial reflected voltage (power) at the matching network input, eliminating the reflections that travel back towards the transmitter. Hence, the steady state 1.0:1 impedance match.**

Mismatch loss considerations

Now let's look at the relationship between the antenna reflection mismatch loss calculated using **Equation 5** and the total transmission line loss calculated using **Equation 13**. Using **Equation 5**, the reflection mismatch loss for this antenna is 4.337 dB. Using **Equation 13**, the total transmission line loss is 4.448 dB.

Let's review how these relate to the above example. First, the forward power measured by a wattmeter at the matching network output is 142.203 watts. Using the one-way matched line attenuation of 1.64 dB and the reflection mismatch loss of 4.337 dB, the total loss is 5.977 dB. With 142.203 watts of forward power, a loss of 5.977 dB indicates that 35.91 watts of power is delivered to the antenna. This is consistent with the total power determined in the example.

Second, determining that the effective net power delivered to the transmission line is 100 watts (142.203 watts forward power minus 42.203 watts reflected power), the total line loss of 4.448 dB indicates that 35.91 watts of power should be delivered to the antenna. Again, this is consistent with the total power determined in our example.

Experimental validation

To provide experimental validation of these results, Danny Horvat, T93M, and I conducted a relatively straightforward experiment at 21.2 MHz. The setup consisted of an antenna connected to a transceiver via a coaxial cable and an antenna tuner located near the transceiver output. The antenna had a measured feedpoint impedance of 6.2 + j14.4 ohms, which calculates to a VSWR of 8.743:1. The coaxial cable used in the experiment had a total matched attenuation of 1.64 dB (approx.). We used an HP8753E network analyzer to make the attenuation and impedance measurements. We performed a full two-port calibration (open, short and load) to calibrate the setup.

After applying transmit power to the system, we tuned out the reflections from the antenna using the antenna tuner. We placed a wattmeter between the transceiver and the tuner input. The forward power was set to 100 watts using the RF output dial on the transceiver (using a 100-watt plug-in meter). The measured reflected power was less than 1/4 watt. We measured forward and reflected power at the output of the tuner. The forward power was near 130 watts. This was a bit inaccurate and difficult to determine because we used a 2500-watt plug-in. The reflected power was 27 watts (100-watt plug-in). Given these measurements, the net steady-state power delivered to the transmission line was near 100 watts. We did not attempt to determine the forward loss through the tuner.

Next, we measured the forward and reflected power at the antenna feedpoint terminals. The forward power was 98 watts and the reflected power was 62 watts. The net steady-state power delivered to the antenna was 36 watts. This represents a relative steady state loss of about 4.44 dB, 100 watts input, 36 watts delivered to the antenna.

These results are identically consistent with the analysis and discussions presented in the previous section.

Summary

I have presented a complete and detailed analysis of the voltage reflections and the power delivered within an antenna system that uses an impedance matching network or tuner. This analysis illustrates that an impedance matching network has an initial state, where an initial voltage reflection exists at the matching network input. The level of initial reflected voltage, and therefore power, is a function of the initial reflection coefficient of the network. The analysis then proves, through detailed determination of voltages within the system, that in the steady state, the initial voltage reflection are canceled as a result of the voltage developed at the matching network input from the multiple reflections within the system.

The analysis also demonstrates that the total steady-state power delivered to the antenna consists of an initial power incident at the antenna plus the reflection echoes of power previously incident at the antenna. Because there is an initial voltage reflection at the matching network input, the analysis shows that the initial power incident at the antenna is not equal to the output power of the transmitter. The transition from the initial state to the steady state is a continuous process as power is being delivered to the matching network from the transmitter.

I have also shown that the analysis accurately predicted the net total power delivered to the antenna. I also demonstrated that the total power delivered to the antenna could be determined using the reflection mismatch loss (**Equation 5**) or the total transmission line loss (**Equation 13**), both determined from the antenna reflection coefficient. Regardless of which method is used, the total power delivered to the antenna is simply a function of the matched transmission line attenuation and the antenna reflection coefficient.

Note that more steady-state power is delivered to an antenna whose feedpoint impedance is matched to the characteristic impedance of the transmission line than to an antenna whose impedance is mismatched to the transmission line. This holds true even when an impedance matching network is used to match the antenna impedance. However, as the transmission line attenuation is reduced, the difference between the two lessen. Additionally, as the antenna VSWR increases, the ratio of steady-state reflection echo power to incident power delivered to the antenna will increase.

It is important to note that the analysis presented in this article makes no assumptions regarding the steady-state condition of the impedance matching network. For instance, the analysis is not based on the assumption that the matching presents a 1.0:1 VSWR at its input. No initial assumption is made that all 100 watts of transmitter output power is delivered into the matching network. In fact, it was shown that this is not how power is delivered throughout the system.

The analysis presented here begins with the voltage incident at the matching network from the transmitter. The analysis then determines the entire voltage distribution throughout the system by considering the reflections occurring at the antenna and at the matching network output. Power delivery and distribution within the system is then determined from the voltages and currents within the system.

Acknowledgments

I would like to thank Danny Horvat, T93M, for all of his valuable discussions related to these topics, as well as his assistance in conducting the experiment.

Although the following individuals may not necessarily support or endorse the opinions expressed in this article, I would like to thank them for their valuable discussions related to this topic and for challenging me to look into this in greater detail. Thanks to Walt Maxwell, W2DU, Ed Sleight, K4SB, Tom Rauch, W8JI, and Jim Reid, KH7M. I would also like to thank Ed for the loan of Walt's book. ■

REFERENCES

1. Robert A. Chipman, *Theory and Problems of Transmission Lines*, Schaum's Outline Series, McGraw-Hill Book Company, New York, 1968.
2. M. Walter Maxwell, W2DU, *Reflections: Transmission Lines and Antennas*, The American Radio Relay League, Newington, Connecticut, 1990.
3. *The ARRL Handbook*, The American Radio Relay League, Newington, Connecticut, 1998.

Practical Example

Consider a 6.2 + j14.4 ohm antenna matched with a "conjugate" impedance matching network. The transmission line has a total of 1.64-dB attenuation. This MathCad program performs a complete and detailed analysis of the voltage delivery and reflections within a system that uses an impedance matching network or tuner to match an antenna's impedance to the characteristic impedance of a transmission line. Once the steady-state voltages are calculated, the total power delivered to the antenna is determined.

First define the known system parameters:

$Zo := 50$ Ohms $\quad Zs := 50$ Ohms $\quad Za := 6.2 + j \cdot 14.4$ Ohms $\quad j := i$

$F := 21.2 \cdot 10^6$ Hz $\quad \lambda := \dfrac{2.997925 \cdot 10^8}{F} \quad \omega := 2 \cdot \pi \cdot F \quad pf := 1 \cdot 10^{12} \quad \mu H := 1 \cdot 10^6$

Define the transmission line characteristics (α, β, and γ) for the line connecting the antenna to the matching network—T2. $\alpha = 0$ assumes a lossless line. Note that α is first entered in dB/ 100 ft and is then converted to nepers/meter.

Enter the transmission line length in meters: $\quad l := 47.607$

Enter the cable attenuation in dB/100 ft.: $\quad \alpha dB := 1.05$

Enter the velocity propagation factor for the cable: $\quad Vp := 0.6594$

$$\alpha := \frac{\alpha dB}{868.59} \cdot 3.28084 \qquad \beta := \frac{2 \cdot \pi}{Vp \cdot \lambda} \qquad \gamma := \alpha + j \cdot \beta$$

Assuming the transmitter is connected directly to the matching network, define the initial voltage incident at the matching network, VI. The power initially incident at the matching network input is equivalent to the power that the voltage VI would deliver to a load of Zo.

$VI := 70.710678$ Volts $\quad Pinitial := \dfrac{(|VI|)^2}{Zo} \quad Pinitial := 100$ Watts

Step 1. Calculate the steady-state input impedance, Zft, to the transmission line, T2.

$$Zft := Zo \cdot \frac{\frac{Za}{Zo} + \tanh(\gamma \cdot l)}{1 + \frac{Za}{Zo} \cdot \tanh(\gamma \cdot l)} \qquad Zft = 36.984 + 54.335i \quad \text{Ohms}$$

Step 2. Define the matching network components that will match the input impedance, Zft, of transmission line T2 to Zo.

$C1 := 69.22892pF \quad C2 := 50.00pF \quad L := 0.449849\,\mu H$

$$ZC1 := \frac{1}{j \cdot \omega \cdot C1} \qquad ZC2 := \frac{1}{j \cdot \omega \cdot C2} \qquad ZL := j \cdot \omega \cdot L$$

Step 3. Calculate the impedance of the matching network, Zrt, as seen by transmission line T2. This is the steady-state impedance looking into the matching network output. Note that for this calculation, it is assumed that Zs = 50 ohms, which results in a 50-ohm impedance seen from the matching network looking back toward the transmitter. Comare impedance Zrt to impedance Zft.

$$Z1 := Zs + ZC1 \qquad Z2 := \frac{1}{\left(\frac{1}{Z1} + \frac{1}{ZL}\right)} \qquad Zrt := ZC2 + Z2$$

$Zrt = 36.984 - 54.335i$ Ohms $\qquad Zft = 36.984 + 54.335i$ Ohms

It is significant to note that Zrt is the complex conjugate of Zft, hence the definition of the "conjugate impedance match." Zrt will ONLY be the conjugate of Zft if the matching network is lossless.

Step 4. Calculate the initial input impedance and reflection properties of the matching network. This initial input impedance differs from the 50-ohm steady-state input impedance because, in the initial state, no reflections from the antenna have yet arrived back at the matching network input.

$$Z1 := Z0 + ZC1 \qquad Z2 := \frac{1}{\frac{1}{Z1} + \frac{1}{ZL}} \qquad Zin := Z2 + ZC1$$

$$Zin = 16.872 - 18.074i \quad \text{Ohms}$$

$$\rho network := \left| \frac{\frac{Zin}{Zs} - 1}{\frac{Zin}{Zs} + 1} \right| \qquad \rho network = -0.394 - 0.377i \qquad \left| \rho network \right| = 0.545$$

$$VSWRnetwork := \frac{1 + \left| \rho network \right|}{1 - \left| \rho network \right|} \qquad VSWRnetwork = 3.393$$

Step 5. Calculate the initial voltage, VR, reflected at the matching network input. In the initial state, the input impedance to the matching network will not be 50 ohms. For this reason, there will be some level of reflected voltage and power at the matching network input. This reflected voltage will be canceled in the steady-state resulting in the 50-ohm steady-state impedance.

$$VR := \rho network \cdot VI \qquad VR = -27.831 - 26.634i \quad \text{Volts}$$

Step 6. Calculate the initial power delivered to the matching network and transmission line T2. Calculate the power reflected at the matching network input. Determine the initial forward voltage, VF, delivered to transmission line T2.

Calculate the matching network component voltages and currents. The initial total input voltage at the matching network is equal to the incident oltage, VI, plus the reflected voltage, VR.

$$Vtotal := VI + VR \qquad Vtotal = 42.88 - 26.634i \quad \text{Volts}$$

$$Itotal := \frac{Vtotal}{Zin} \qquad Itotal = 1.971 + 0.533i \quad \text{Amps}$$

Calculate the net power, PI, initially delivered to the matching network and the input voltage, VF, to the transmission line connecting to the matching network to the antenna. Also, calculate the initial power reflected, PR, at the matching network.

$$\theta := \arg(Vtotal) - \arg(Itotal)$$

$$PI := \left| Vtotal \right| \cdot \left| Itotal \right| \cdot \cos(\theta) \qquad PI = 70.322 \quad \text{Watts}$$

$$PR := Pinitial - PI \qquad PR = 29.678 \quad \text{Watts}$$

Calculate the matching network component voltages and currents as necessary to determine VF.

$$VFC1 := Itotal \cdot ZC1 \qquad V2 := Vtotal - VFC1 \qquad IL := \frac{V2}{ZL}$$

$$VFL := IL \cdot ZL \qquad I2 := Itotal - IL \qquad VFC2 := I2 \cdot ZC2$$

$$VF := I2 \cdot Zo \qquad VF = -57.568 + 14.214i \quad \text{Volts}$$

Calculate the power initially delivered to the transmission line.

$\theta := \arg(VF) - \arg(I2)$

$Power := |VF| \cdot |I2| \cdot \cos(\theta)$ $\qquad$ $Power = 70.322$ Watts

Step 7. Calculate the initial incident voltage, VIA, and the initial reflected voltage, VIR, at the antenna. Also calculate the initial power delivered to the antenna.

Calculate the initial voltages at the antenna based upon the antenna reflection properties.

$$\rho a := \frac{\frac{Za}{Zo}+1}{\frac{Za}{Zo}-1} \qquad \rho a = -0.67+0.428i \qquad |\rho a| = 0.795$$

$$VIA := VF \cdot e^{-\gamma \cdot l} \qquad V2 := VF \cdot \rho a \cdot e^{-2 \cdot \gamma \cdot l} \qquad VRA := V2 \cdot e^{\gamma \cdot l}$$

Calculate the initial total voltage and power delivered to the antenna.

$Vantenna := VIA + VRA$ $\qquad$ $Vantenna = -26.533 - 0.227i$ Volts

$Iantenna := \frac{VIA}{Zo} - \frac{VRA}{Zo}$ $\qquad$ $Iantenna = -0.683+1.549i$ Amps

$\theta := \arg(Vantenna) - \arg(Iantenna)$

$Power := |Vantenna| \cdot |Iantenna| \cdot \cos(\theta)$ $\qquad$ $Power = 17.759$ Watts

Step 8. Determine the steady-state conditions in the system. There will be a great number of reflections in the system that result in steady-state volgates at the matching network output, matching network input, and the antenna. Steady-state power at the antenna in the transmission line can be determined.

When the reflected voltage from the antenna arrives at the matching network output, it will "see" impedance Zrt. Therefore, in order to perform a complete and accurate reflection analysis, the reflection properties of Zrt must be determined.

$$\rho N := \frac{\frac{Zrt}{Zo}-1}{\frac{Zrt}{Zo}+1} \qquad \rho N = 0.173-0.517i \qquad |\rho N| = 0.545$$

Because ρN is not 0, there will be a reflected voltage created at the matching network that travels back toward the antenna. This voltage will then experience a reflection at the antenna. This process will continue until the system reaches the steady-state condition. Note that the magnitude of ρN cannot be 1 unless Zrt is an open circuit, short circuit or purely reactive.

The total voltage at the matching network will be the sum of all forward and all reflected voltages at the matching network.

$V1 := VF$

$$VA := V1 + V1 \cdot \rho a \cdot e^{-2 \cdot \gamma \cdot l} + V1 \cdot \sum_{n=1}^{1000} \rho N^{n} \cdot \rho a^{n} \cdot e^{-2 \cdot n \cdot \gamma \cdot l}$$

$$Vnetwork := VA + V1 \cdot \sum_{n=1}^{1000} \rho N^{n} \cdot \rho a^{(n+1)} \cdot e^{-(2n+2) \cdot \gamma \cdot l}$$

$$IA := \left(\frac{V1}{Zo} - \frac{V1}{Zo} \cdot \rho a \cdot e^{-2 \cdot \gamma \cdot l}\right) + \frac{V1}{Zo} \cdot \sum_{n=1}^{1000} \rho N^{n} \cdot \rho a^{n} \cdot e^{-2 \cdot n \cdot \gamma \cdot l}$$

$$Inetwork := IA - \frac{V1}{Zo} \cdot \sum_{n=1}^{1000} \rho N^{n} \cdot \rho a^{(n+1)} \cdot e^{-(2n+2) \cdot \gamma \cdot l}$$

Verify the steady-state input impedance to transmission line T2 using the steady-state voltage and current developed at the matching network output. This should be equal to Zft determined earlier.

$$Z := \frac{Vnetwork}{Inetwork} \qquad Z = 36.984+54.335i \quad \text{Ohms}$$

$$Zft=36.984+54.335i \quad \text{Ohms}$$

The total voltage at the antenna will be the sum of all forward and all reflected voltages at the antenna.

$$VA := V1 \cdot e^{-\gamma \cdot l} + V1 \cdot e^{\gamma \cdot l} \cdot \rho a \cdot e^{-2 \cdot \gamma \cdot l} + V1 \cdot e^{-\gamma \cdot l} \cdot \sum_{n=1}^{1000} \rho N^{n} \cdot \rho a^{n} \cdot e^{-2 \cdot n \cdot \gamma \cdot l}$$

$$Vantenna := VA + V1 \cdot e^{\gamma \cdot l} \cdot \sum_{n=1}^{1000} \rho N^{n} \cdot \rho a^{(n+1)} \cdot e^{-(2n+2) \cdot \gamma \cdot l}$$

$$IA := \left(\frac{V1 \cdot e^{-\gamma \cdot l}}{Zo} - \frac{V1 \cdot e^{\gamma \cdot l}}{Zo} \cdot \rho a \cdot e^{-2 \cdot \gamma \cdot l}\right) + \frac{V1 \cdot e^{-\gamma \cdot l}}{Zo} \cdot \sum_{n=1}^{1000} \rho N^{n} \cdot \rho a^{n} \cdot e^{-2 \cdot n \cdot \gamma \cdot l}$$

$$Iantenna := IA - \frac{V1 \cdot e^{-\gamma \cdot l}}{Zo} \cdot \sum_{n=1}^{1000} \rho N^{n} \cdot \rho a^{(n+1)} \cdot e^{-(2n+2) \cdot \gamma \cdot l}$$

Calculate the total steady-state power delivered to the antenna to include all reflections.

$$\theta := \arg(Vantenna) - \arg(Iantenna)$$

$$Power := |Vantenna| \cdot |Iantenna| \cdot \cos(\theta) \qquad Power =35.912 \quad \text{Watts}$$

Step 9. Determine what a typical wattmeter would measure on the line as forward and reflected power under a steady-state condition. The wattmeter is located at the matching network output.

$$Vforward := V1 + V1 \cdot \sum_{n=1}^{1000} \rho N^{n} \cdot \rho a^{n} \cdot e^{-2 \cdot n \cdot \gamma \cdot l}$$

$$Iforward := \frac{Vforward}{Zo}$$

$$\theta := (Vforward) - \arg(Iforward)$$

$$Pforward := |Vfotward| \cdot |Iforward| \cdot \cos(\theta) \qquad Pforward = 142.203 \quad \text{Watt}$$

$$Vreflected := (V1 \cdot \rho a \cdot e^{-2 \cdot \gamma \cdot l}) + V1 \cdot \sum_{n=1}^{1000} \rho N^{n} \cdot \rho a^{(n+1)} \cdot e^{-(2n+2) \cdot \gamma \cdot l}$$

$$\text{Ireflected} := \frac{\text{Vreflected}}{\text{Zo}}$$

$$\theta := \arg(\text{Vreflected}) - \arg(\text{Ireflected})$$

$\text{Preflected} := |\text{Vreflected}| \cdot |\text{Ireflected}| \cdot \cos(\theta)$ Preflected=42.203 Watts

Calculate the effective net power delivered to transmission line T2.

$\text{Peff} := \text{Rforward} - \text{Preflected}$ Peff=100 Watts

Step 10. Calculate the voltage delivered back to the input of the matching network, Vback. VT is defined as the total voltage developed at the matching network output as a result of the forward and reflected voltages developed at the matching network due to the reflections from the antenna. Once VT is determined, the steady-state voltage, Vback, can be determined using simple circuit theory. In the steady state, Vback will add to the initial incident and reflected voltages at the matching network. In order to prevent the initial reflection, as described in Step 6, from traveling back toward the transmitter, Vback must be the negative of the initial reflected voltage, VR.

$\text{VT} := \text{Vnetwork} - \text{VF}$ $\text{IT} := \dfrac{\text{VT}}{\text{Zrt}}$

$$\theta := \arg(\text{VT}) - \arg(\text{IT})$$

$\text{PT} := |\text{VT}| \cdot |\text{IT}| \cdot \cos(\theta)$ PT = 29.678 Watts

Determine the voltages delivered to the matching network components as a result of the multiple reflections within the system. Deter-mine the voltage, Vback, and the power delivered back to the matching network input, Pback.

$\text{VRC2} := \text{IT} \cdot \text{ZC2}$ $\text{V2} := \text{VT} - \text{VRC2}$

$\text{IL2} := \dfrac{\text{V2}}{\text{ZL}}$ $\text{VRL} := \text{IL2} \cdot \text{ZL}$ $\text{I2} := \text{IT} - \text{IL2}$

$\text{VRC1} := \text{I2} \cdot \text{ZC1}$

$\text{Vback} := \text{I2} \cdot \text{Zo}$ Vback = 27.83+26.634i Volts

VR = −27.831− 26.634i Volts

Note that Vback is the negative of the initial reflected voltage, VR, causing a cancelation of voltage VR and Vback. In the steady state, this prevents the reflected voltage, VR, from traveling back toward the transmitter. This also prevents the reflected voltage from the antenna from traveling back toward the transmitter.

Calculate the power delivered back to the matching network input and compare it to the power reflected at the matching network input.

$$\theta := \arg(\text{Vback}) - \arg(\text{I2})$$

$\text{Pback} := |\text{Vback}| \cdot |\text{I2}| \cdot \cos(\theta)$ Pback = 29.678 Watts

PR = 29.678 Watts

Note that these two powers are exactly 180 degrees out of phase, and therefore cancel each other.

$\text{Vback} + \text{VR} = -3.494 \cdot 10^{-4} - 1.049 \cdot 10^{-7} i$ Volts

$\text{Vback} + \text{Vtotal} = 70.71 - 1.049 \cdot 10^{-7} i$ Volts

Hence, in the steady-state, no reflections travel toward the transmitter. This results in the steady-state, 50-ohm impedance (1.0:1 VSWR) at the matching network input.

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A 222-MHZ TRANSVERTER

"Roll your own" and get on the band

Why aren't you on 222? I can think of a number of reasons: There's nobody on the band. It's impossible to buy new rigs, or even old junkers. There's no time, no propagation, etc. Still, there is activity on 222 and you'll be pleasantly surprised at the amount of propagation. I started on 223.5 FM simplex some time around 1977, and in detailed comparisons between 135 centimeters and 2 meters, 223.5 was almost always better!

The lack of available equipment has been a problem for years, even on FM. But it's much easier to build high-quality, high-performance gear for VHF and UHF than for HF. The engineering challenge of building something comparable to even the low-end cost range of commercially available HF equipment is formidable. On the other hand, homebrew VHF/UHF transverters can perform as well or better than commercial transceivers at considerably less cost. Plus you have the satisfaction of having "rolled your own."

Checking the literature

There are several transverters in the literature[1–3] designed to extend the use of rigs that operate in other amateur bands to 222 MHz. Most use an intermediate frequency (IF) of 28 MHz. Power inputs and outputs vary considerably, as do the construction techniques. Printed circuit boards may make construction easier, but interference can occur when you put many circuits on one board. I wanted a transverter for 222 MHz that was relatively easy to build, and as clean as I could reasonably make it on both receive and transmit.

Several years ago Joe Reisert, W1JR, published a series of articles in *Ham Radio*[4–6] that addressed the problem of building your own gear in the VHF and UHF ranges. He recommended, and I heartily agree, the use of a small RF-tight box for essentially every circuit in a transverter—especially for the local oscillator (LO). However, it's not necessary to use a bunch of small pc boards, even at 432 MHz; "dead-bug" construction works fine. Of course, while the use of separate boxes for every stage greatly simplifies testing and tuning, you will face considerable added cost for the several boxes, connectors, and cables.

Reisert's designs have always worked right for me the first time, so I used many of them for some of the individual parts of my trans-

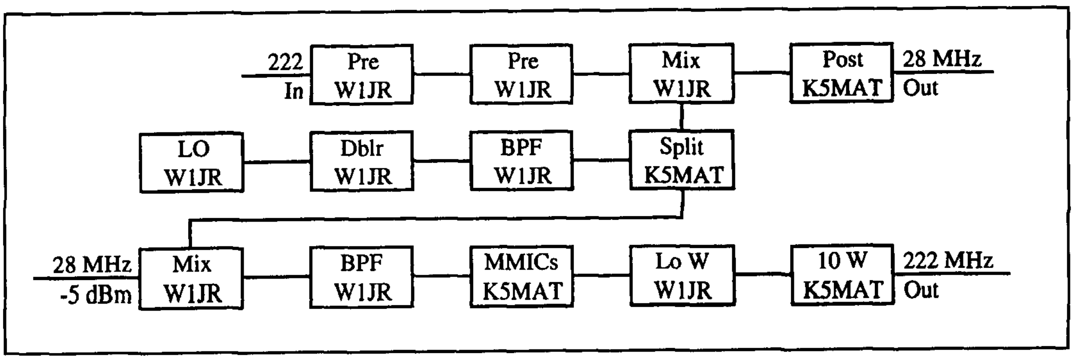

Figure 1. 222-MHz transverter.

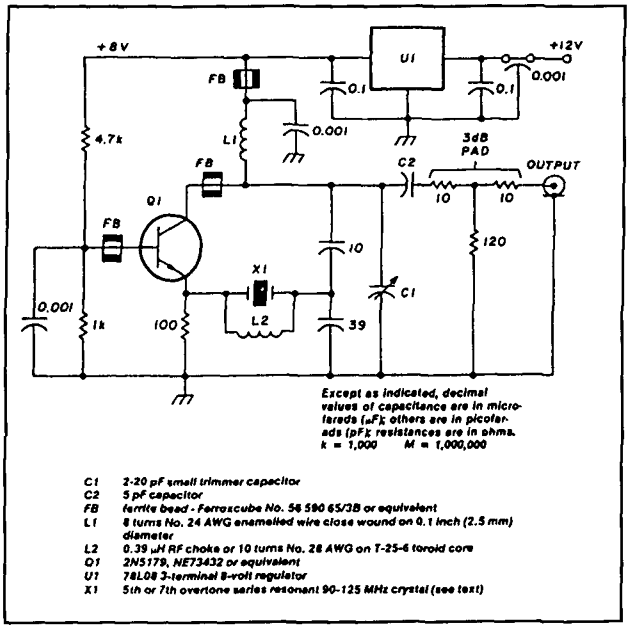

Figure 2. W1JR's LO.[4]

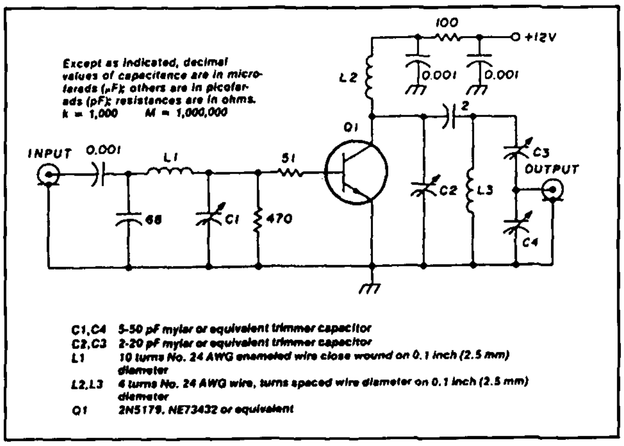

Figure 3. W1JR's doubler.[4]

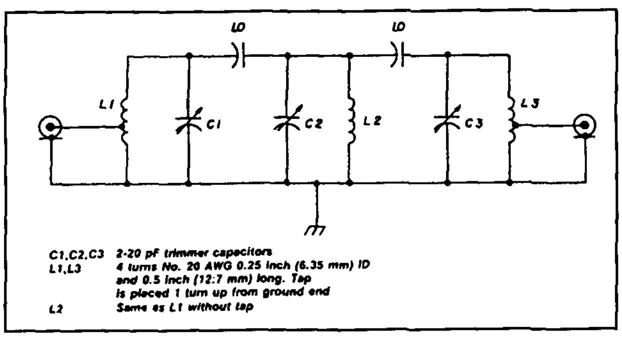

Figure 4. W1JR's band-pass filter.[5]

verter. However, I also took advantage of some more modern technology—especially in the transmitter. The block diagram of my 222-MHz transverter is shown in **Figure 1**. The blocks that are simply copied from Reisert's articles are labeled W1JR. Read his work if you're curious about any of the specifics. Anything that is a substantial modification of his design, or is of my own design, is labeled with my call.

As **Figure 1** shows, you'll need to find some way to perform the necessary transmit-receive switching because inputs and outputs are totally separate. I tap off a low power level ahead of the final amplifier in my transmitter and disconnect screen voltage from the final tubes so I don't need to worry about the extra power consumed by that stage. That may be difficult if you're using a transceiver and you need to prevent putting too much drive into the transmit mixer. A system of relays and dummy load attenuators of sufficient power handling capabilities may be necessary to accomplish the switching with a transceiver of any frequency range as an IF.

The use of separate boxes for every stage greatly simplifies testing and tuning. Even if you don't choose to put each stage in a separate enclosure, you should still build, test, and tune one stage at a time. The importance of this type of testing cannot be overemphasized.

Test equipment

The single most useful piece of test gear I had at my disposal, other than the usual small scope, volt-ohmmeter, etc., was a microwattmeter that provided full scale readings of +10 dBm (10 milliwatt) down to -30 dBm (1 microwatt) with reasonable accuracy. While you probably don't have such a thing in your junk pile, it's not inconceivable that you can borrow one. Alternatively, you might consider building one,[7] they're really handy gadgets to have around if you're interested in doing developmental work at VHF/UHF. I also have a useful set of low power attenuators (1, 2, 3, 6, 10, 14, and 20 dB). In fact, the clever use of those attenuators and a simple detector lets a builder do almost as well as he would with the microwattmeter.

I borrowed a reasonably good spectrum analyzer to prove out the final results. Such equipment is nice to have, but not really necessary. It's also a good idea to be able to measure the final output power as a function of input power to ensure high linearity if a suitable scheme can be found to do so. It's quite interesting to do that with so-called linear commercial brick amps!

The transverter design

The heart of a transverter is the local oscillator, and Reisert's design[4] has become a de facto standard (see **Figure 2**). I used a 28-MHz IF with a crystal frequency of 97.0000 MHz. If you want to use 50 (LO = 86.0000) or 144 (LO = 78.0000) MHz for the IF, you'll have to design your own LO and multiplier chain. It's important to purchase a good crystal as it will be more likely to produce adequate output power and be stable. I used Reisert's design for the LO multiplier[4] without change. The first doubler (the only one needed for 222) is shown in **Figure 3**. I used a supply voltage of 13.6 volts throughout.

A bandpass filter[5] (**Figure 4**) is tuned to the desired output frequency of 194 MHz to minimize problems with spurious signals in both the receive and transmit sections of the transverter. For this frequency, I used 10-pF Johanson variable capacitors and 1-pF fixed chips. All the inductors are five turns of #18 tinned wire wound on a 0.25-inch drill bit with an overall length of 1/2 inch. The taps are one turn up from the ground end. I tuned the filter for maximum output at 194 MHz. This filter gave maximum spurious output frequency responses of more than 60 dB down.

As there are many kinds of power splitters that can be used, I won't discuss them all here. **Figure 5** shows one of the simplest splitters you can use, but in my case it was necessary to add monolithic microwave integrated circuit (MMIC) amplifiers to overcome the extra loss in the resistors.

The receive mixer[4] is shown in **Figure 6**; it is for 28 MHz only. A different diplexer tuned to another frequency would be necessary for a different IF. The diplexer helps reduce spurious responses in the receiver. The 3-dB pads at the input and output of the mixer assure good termination of the mixer at all frequencies. My station uses Drake B-line equipment for the IF, and the receiver needs a bit of gain ahead of it for optimum performance. I used an MMIC amplifier followed by a bandpass filter[8] as shown in **Figure 7**. Many operators with more modern equipment than mine may have no use for the postamp. **Reference 8** contains some excellent construction tips.

The U310 preamp[5] (**Figure 8**) is strictly W1JR. For 222 MHz, the variable capacitors are all 10-pF Johansons: C4 is 2 pF, and L1 and L2 are three turns of #18 tinned wire. Preamps for the 135-centimeter band are easy to build, but tuning them for lowest noise figure is nontrivial. If you use a GaAsFET preamp ahead of the U310, it's okay to tune the U310 for maximum gain. If not, try to tune it for best readability on an extremely weak signal. If you can take your preamps to one of the various VHF conferences around the country, you may be able to get it tuned there. Your only expense will be a plane ticket, hotel, and meals.(Hi!) I haven't included a schematic of my GaAsFET preamp[9] because many of you will probably

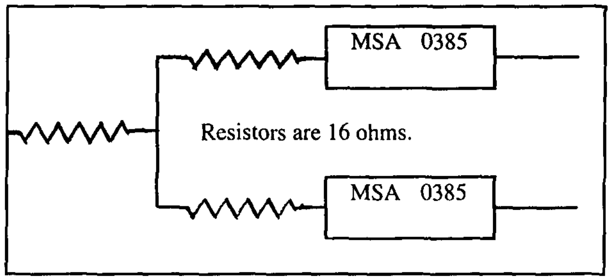

Figure 5. Simple power splitter.

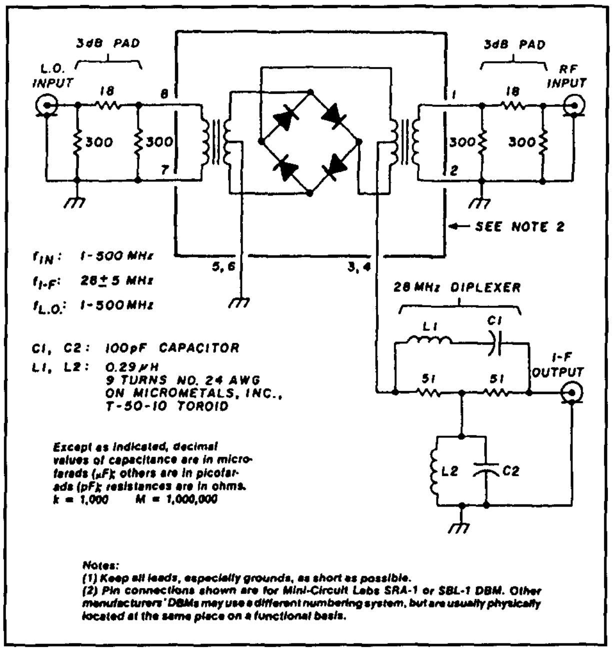

Figure 6. W1JR's receive mixer.[4]

have your own favorite, or will have one in a brick power amplifier.

I made some tests with a similar 2-meter receiver as compared to the receiver in a commercial transverter using dual-gate MOSFETs for the RF amp and mixer. A friend who lives about a kilometer from me made some transmissions while I tuned around. With the commercial rig, I stopped tuning for garbage when I was 100 kHz above and below his frequency—even though the noise was still quite annoying. With my homebrew receiver, I couldn't tell he was on the air until his sidebands were inside the first IF filter in the R-4B, which was about 5 kHz out. Yet on-the-air listening tests showed my receiver (even without a GaAsFET in front of it) could copy weak distant signals just as well, if not better.

The transmitter

The transmitter is only slightly more complex than the receiver! Again, I started with the transmit mixer[6] shown in **Figure 9**. Af this point, I departed from some of Reisert's

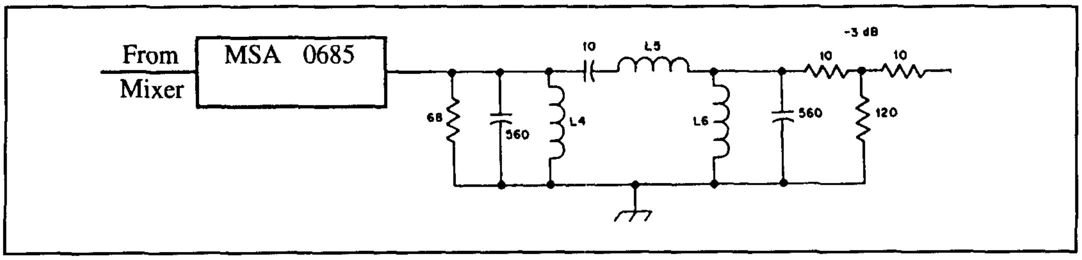

Figure 7. 28-MHz post amp. Note: L4 and L6 are 4 turns of #18 enameled wire 5/32-inch I.D., close wound. L5 is 27 turns of #24 enameled wire on a T-50-6 core.

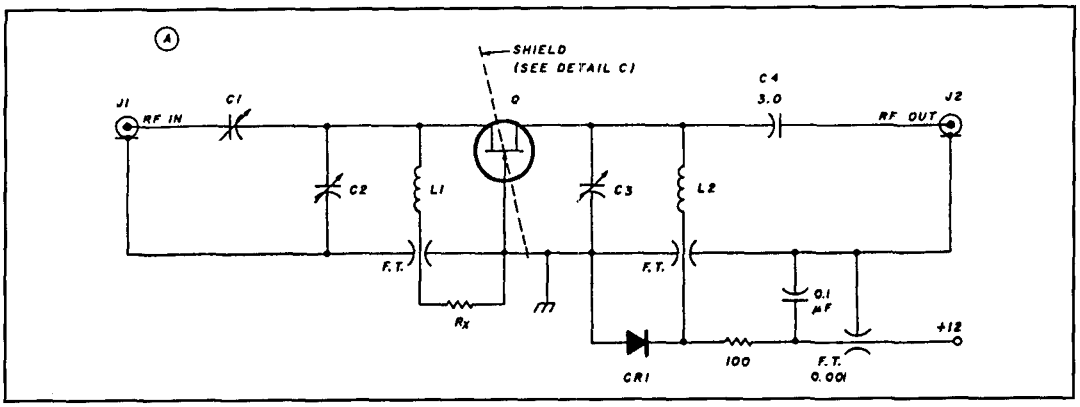

Figure 8. W1JR's U310 pre-amp.[5]

designs. The transmit mixer is followed by another bandpass filter[5] (see **Figure 4**). This mixer must be good enough to reject the LO and image frequencies as well as all the other mixing products. The coils are four turns of #18 tinned wire wound on a 0.25-inch drill bit, 1/2 inch long, with the taps at 3/4 turn from ground.

Ideally, tuning should be done by sweeping the filter, but it's not too far off to tune for maximum signal at the frequency of interest, checking occasionally at the LO frequency to make sure it remains low in intensity. This is what I did. The RF level at 222 MHz coming out of the filter should be about -16 dBm with a 28-MHz drive level of 0 dBm. This level of 25 microwatts isn't enough to allow you to work

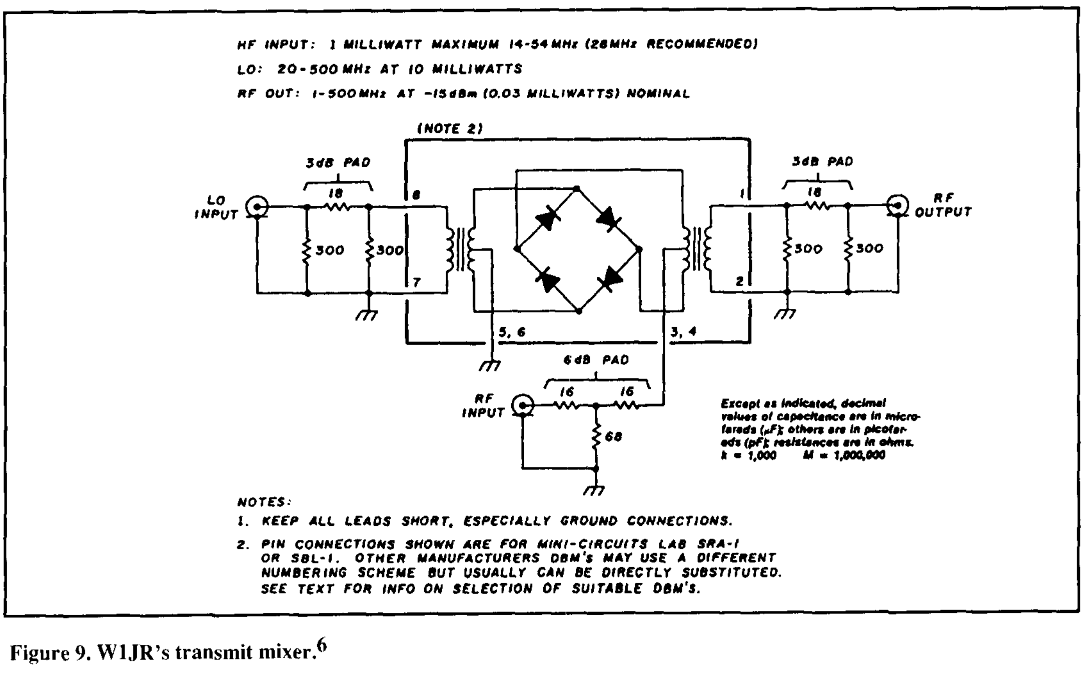

Figure 9. W1JR's transmit mixer.[6]

much DX, but modern MMICs let you get to the 25 milliwatt level with ease. This is still quite low, but should allow real communications with locals.

The block diagram of my low-level MMIC amplifier is shown as **Figure 10**. If you build this transverter, you may need to use different MMICs because of the variability of the gain in production units of the M67712 modular amplifiers. This amplifier is a high-gain amp in a small box. Good bypassing and shielding techniques are necessary, or you may just find that you have a super-GHz-plus oscillator instead of an amplifier!

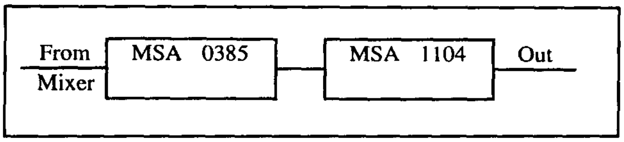

Figure 10. Low-level transmit MMIC amp.

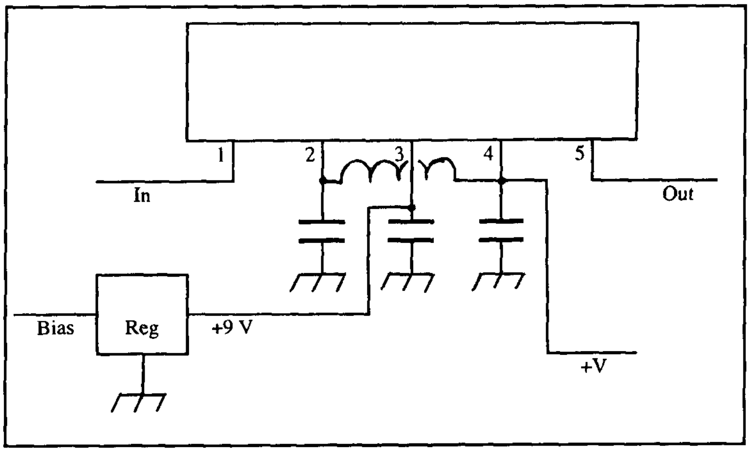

Figure 11. Mitsubishi 67712 modular amplifier.

Using the Mitsubishi M67712

Figure 11 shows the use of a Mitsubishi M67712 modular amplifier to get to the 10-watt level. It's specified to have >20-dB gain and 25 watts output at the 1-dB compression point. I couldn't measure a departure from linearity at the 12-watt level (my way of trying to assure a clean transmitter), and the gain of the one I bought was measured at 22 dB!

This circuit is straight out of the Mitsubishi manual and the device was chosen to maximize linearity and gain at the 10-watt level rather than to minimize the cost. This amplifier was the first piece of hardware I built (dead-bug style) for the transmitter portion, so I would know exactly how much drive power would be needed to get the desired output. If I could get to that drive level without "stretching" MMICs, it would simplify construction. The capacitors are 33-μF tantalums in parallel with 0.001 chips mounted as close as feasible to the module, and the RF chokes are about 10 turns of #26 enameled wire close wound on a 0.25-inch drill bit.

I built similar transverters for 50, 144, and 432 MHz (all dead-bug construction) using appropriate modular amplifiers. None of them required anything but MMICs to drive the modular amp, but they all had higher gain than the M67712. I used another of Reisert's circuits[5] (**Figure 12**) using an old CATV transistor. Unfortunately, these may not be available today, but there are many possible substitutions. It's necessary to use a low-pass filter of 230 to 240 MHz rather than the 150 shown on the schematic. This amp is highly linear up to 250 mW out with a value of 20 ohms for the resistor R1, and has a gain of about 12 dB.

The MMIC amplifiers[10] are hooked up as shown in **Figure 13**. The resistor is calculated by subtracting the device operating voltage from the supply voltage, then dividing by the recommended device current in amps. If the resistor value is less than 500 ohms, you may need to add the choke shown. Note that a single coupling capacitor is all that's needed between the output of one stage and the input to a following stage. The capacitors should have a reactance of less than 1 ohm at the operating frequency. See **Reference 11** for a complete description of how to use MMICs at VHF and higher frequencies.

The overall gain of my transverter is about 46.5 dB, which made it difficult for me to measure linearity at low levels. However, an increase in drive of 3 dB resulted in around 25 watts of output, which is still within the 1-dB compression range. I operate at the 10- to 12-watt output level on SSB, and, if I need CW, I can crank up the output to 25 watts without generating trash.

Parts procurement can always be a problem. RF Parts Co. in California is a good source for the M67712 amplifier. Down East Microwave has the capacitors and MMICs. I have a limited number of capacitors and MMICs, as well, and will be happy to send a list, with prices, if the request includes a self addressed stamped envelope (SASE).

Spurs and harmonics

I now have a very linear, highly frequency stable 10-watt transverter driven with less than 1 mW at 28 MHz. So far so good; but how about spurs? Unfortunately I didn't have access to a good spectrum analyzer that could be used to take photographs, but a friend lent me a Tektronix unit with which to make measurements. The highest non-harmonic spur was

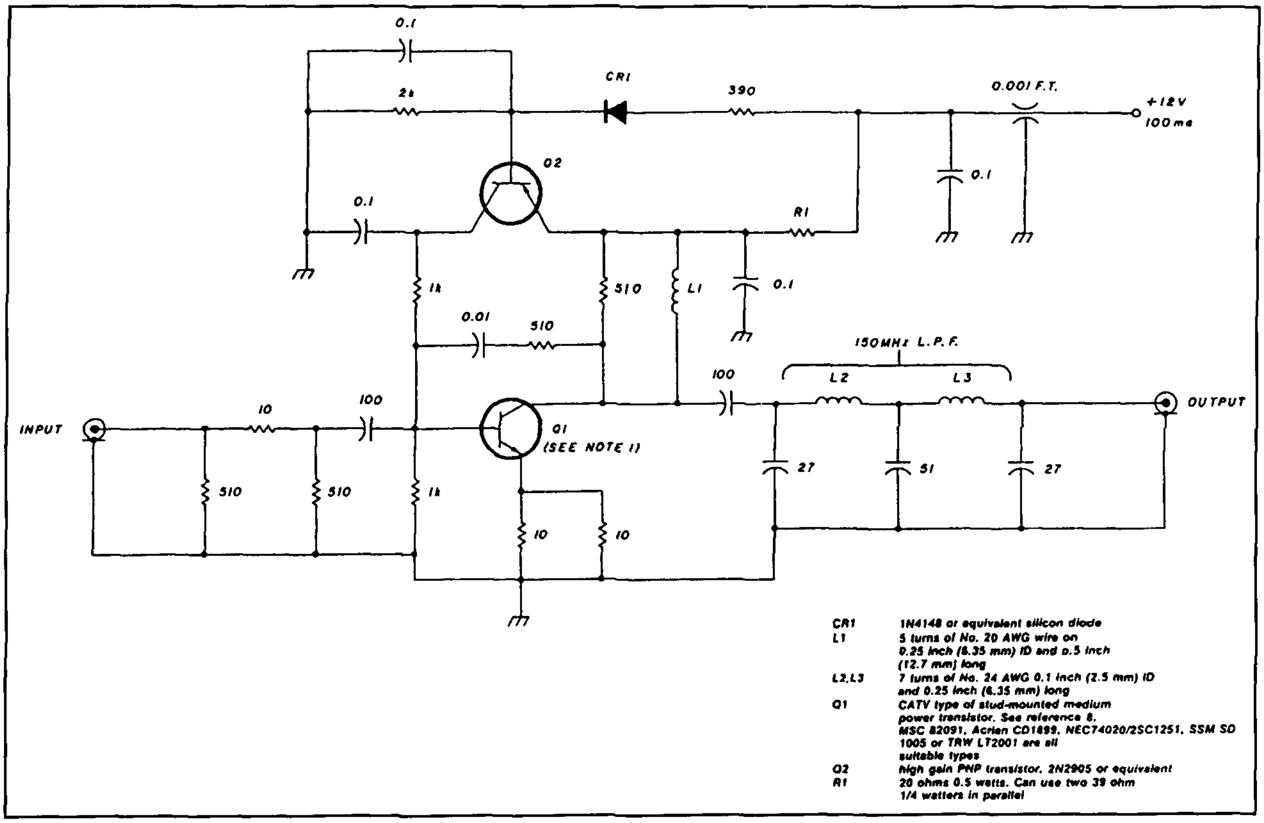

Figure 12. W1JR's 250-mW amplifier.[5]

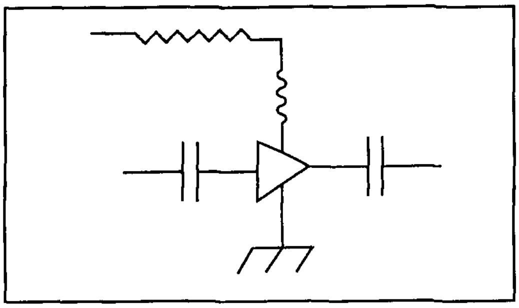

Figure 13. MMIC components. This bias resistor must be determined for each circuit. The capacitors are 1 nF, and the choke should have sufficient inductance so that the impedance of the biasing is about 500 ohms or more.

over 70 dB down. Unfortunately I lost the harmonic measurements, and I no longer have access to the analyzer, but I do remember that harmonics were considerably further down than specified for the modular amplifier itself. If harmonics may be a problem, follow this unit with a good low-pass filter.

Acknowledgments

I'm grateful to Joe Reisert for his wonderful series of articles in *Ham Radio* magazine. They inspired me to build a largely homebrew station for 144, 222, and 432 MHz. I'd also like to thank Dick Lewis, K5RHR, for his helpful review of my manuscript (although he is *not* to blame for any errors). I'll be happy to try and answer any questions upon receipt of a SASE.

REFERENCES

1. Zak Lau, KH6CP, 1995 *ARRL Handbook*, pp. 17.88 to 17.92
2. Paul Drexler, WB3JYO, 1991 *ARRL Handbook*, pp. 31-17 to 31-28
3. Richard Stroud, W9SR, 1985 *ARRL Handbook*, pp. 31-18 to 31-25
4. Joe Reisert, W1JR, *Ham Radio*, March 1984, pp. 42 to 46
5. Joe Reisert, W1JR, *Ham Radio*, November 1985, pp. 54 to 62
6. Joe Reisert, W1JR, *Ham Radio*, April 1984, pp. 84 to 88
7. Denton Bramwell, K7OWJ, *QST*, June 1997, pp. 33 to 35
8. Dave Eckhardt, W6LEV, 1991 *ARRL Handbook*, Figure 35, page 32–20
9. Joe Reisert, W1JR, *Ham Radio*, December 1987, pp. 72 to 81
10. 1995 *ARRL Handbook*, page 24.22
11. Al Ward, WB5LUA, *The ARRL UHF/Microwave Experimenters Manual*, page 7-32

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10-MHz CLASS-E POWER AMPLIFIERS

200 watts from a $6 transistor

A new HF, high-efficiency class-E radio power amplifier has been developed at Caltech that is smaller and less expensive than existing amplifiers with similar characteristics. This amplifier operates on the 30-meter (10-MHz) band and is designed with an output power of 200 watts. It is based around the International Rectifier IRFP440 MOSFET, which is a device commonly used in switching power supplies and costs under $6.

We were able to use such a relatively inexpensive, low-power device because of the high efficiency characteristic of the class-E amplifier—typically around 90 percent. This makes the power dissipated in the switching device small enough to allow operation without a cooling fan. The 30-meter amplifier requires only 3 to 8 watts of drive power for reliable, efficient operation. We used a modified NorCal-40A low-power radio transceiver to supply this drive.

This new amplifier is a continuation of work done at Caltech by undergraduate and graduate students over the past three years. Last year, a pair of amplifiers which had output powers of 300 and 500 watts was developed specifically for hams operating on the 40-meter band. A paper about these amplifiers was published in *QST* magazine.[1] Collections of parts were assembled and sold to people who wanted to try building the amplifiers. It was found that there was much interest among people in amplifiers for other bands, and this led to the work done during the summer of 1997 and the development of the new 30-meter, 200-watt amplifier. Like its predecessors, this new amplifier is easy to build, costs about the same, and has most of the same characteristics and limitations.

Background

It's often necessary to amplify weak signals to make them more useful, and there are different methods for carrying out this amplification. Each class of amplifier has advantages and disadvantages that make it more suitable in certain situations. Most conventional amplifiers fall into one of the three classes: A, B, or C. The following is a short description of each of these classes:

Class A: These amplifiers use a single transistor, which is always active. An active transistor has considerable amounts of voltage across and current through it. The product of the voltage across the transistor and current through it translates into heat dissipated in the transistor. For this reason, the efficiency of class-A amplifiers is limited to about 30 percent. However, class-A amplifiers introduce very little distortion in the amplification process. This means that the output of a class-A amplifier looks almost exactly like its input, only greater in amplitude.

Class B: Class B builds on class A by having two transistors, which take turns amplifying the input signal. One transistor usually handles the

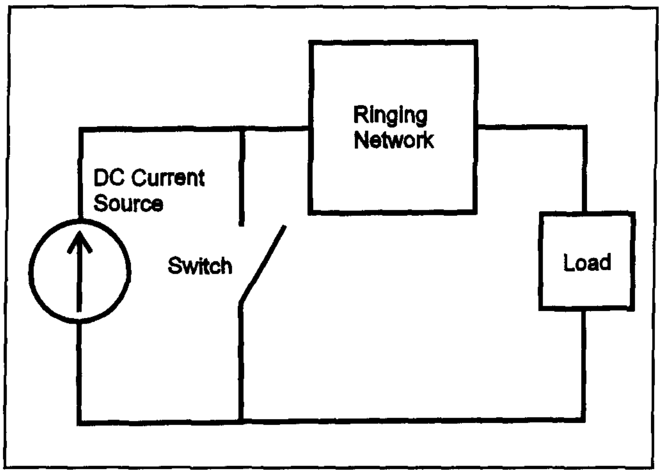

Figure 1. Block diagram of class-E amplifier.

positive half of the output, and the other the negative. This way, only one transistor is active at any given time, while the other is off. This doubles the efficiency to around 60 percent, at the cost of small amounts of distortion at the signal crossover, where one transistor turns off and the other takes over the amplification.

Class C: The class-C amplifier is known as a switching amplifier. In traditional class-C amplifiers, the switching device is active less than half of the time, making it even more efficient. In modern, transistor-based class-C amplifiers, the transistor often acts like a switch, either completely on or completely off. When on, a large current flows through the transistor, but the voltage across it is small. When off, voltage builds across the transistor, but the current through it falls nearly to zero. This way, there is very little overlap between voltage and current in the transistor, which translates into much less power being wasted in the device and a higher efficiency.

Class-C amplifiers normally have efficiencies reaching 75 to 80 percent. Most of the loss in a class-C amplifier occurs during the actual switching. The transistor does not switch from on to off (or vice versa) instantaneously. As the transistor switches, it goes through an active region, where it has both a current through it and voltage across it. This translates into power wasted as heat in the transistor every time it is switched, and this wasted power increases as the frequency of operation is increased. A disadvantage of the class-C amplifier is that because of its switching mechanism for amplifying signals, it is not at all linear. The output may look nothing like the input. It will be at the same frequency, though with a great deal of harmonic distortion.

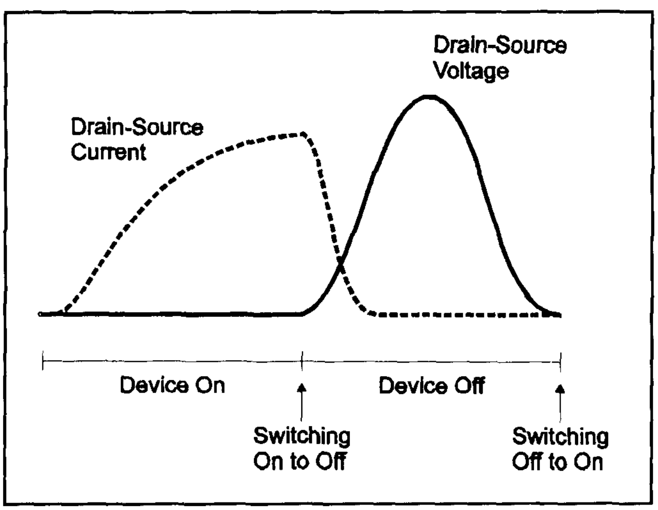

Figure 2. Ideal class-E waveforms.

Class E

The Caltech power amplifiers rely on a less well-known class of operation: class E. The class-E amplifier design was first explored in 1964 by Gerald Ewing[2] and later by Nathan and Alan Sokal in 1975,[3] who patented it. The use of class-E amplifiers in high-frequency amateur radio transmitters prior to the Caltech designs is unreported.

An idealized diagram of a class-E amplifier is shown in **Figure 1**. It consists simply of a DC current source feeding a switch, followed by a ringing network and the load. The class-E amplifier is very similar to the class-C amplifier. However, a class-E amplifier achieves even higher efficiency (typically around 90 percent) by minimizing the current-voltage overlap that occurs during switching in a class-C amplifier. This is done by the addition of a ringing network following the switching device. With the ringing network in place, the voltage across the transistor rises smoothly from 0 to a maximum, and falls smoothly back to zero just before the transistor switches on again. A complete cycle of the class-E switching amplifier can be broken down into the following four phases (see **Figure 2**):

Device On: Transistor conducts current through itself to the ground. The voltage across the transistor is very small, keeping power dissipation in the device low.

Switching On to Off: Voltage across transistor rises smoothly as current is redirected from the device into the resonant load network. While the switching is occurring, power dissipation is kept low because the voltage across the device rises slowly and remains small during the short switching period.

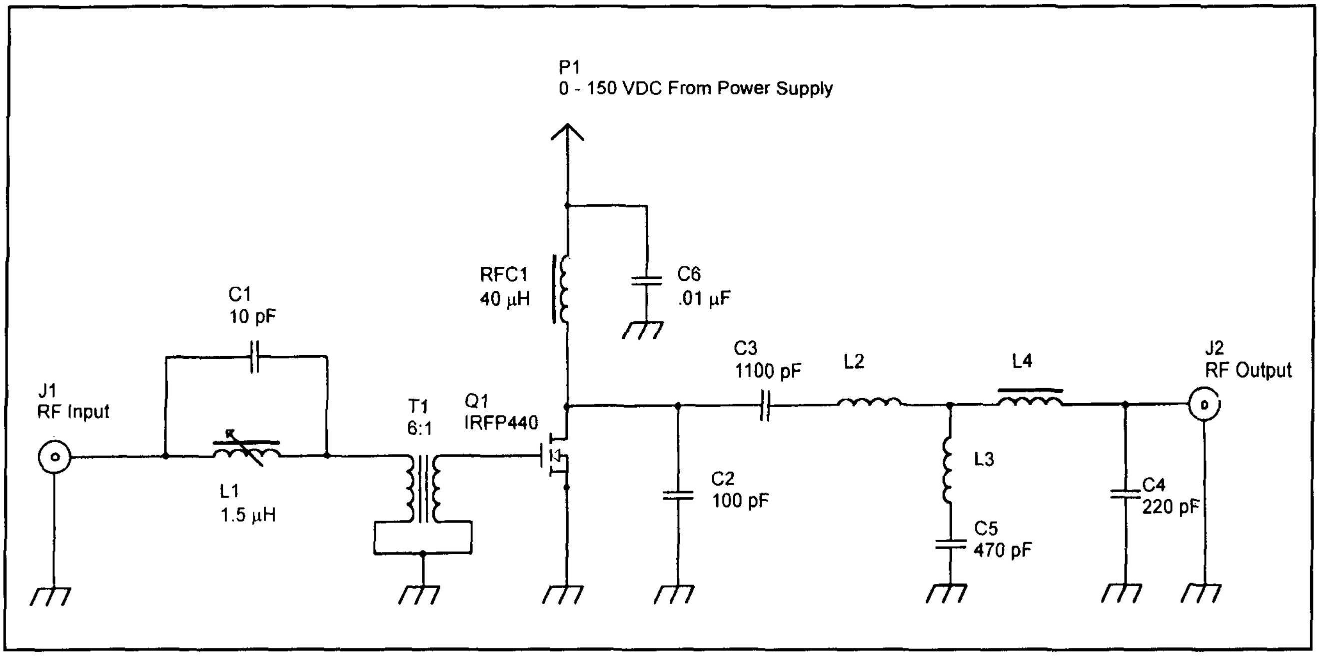

Figure 3. Circuit diagram of 30-meter, 200-watt amplifier.

Device Off: Voltage rises smoothly to a maximum, then falls smoothly back to zero.

Switching Off to On: Because there's near-zero voltage across the transistor, little power is wasted as current is again directed into the device instead of the load network.

In this way, the class-E amplifier achieves efficiencies of around 90 percent, which allows operation with a relatively small, inexpensive transistor. Power dissipated in the transistor is small enough that heat can be removed with a simple heatsink. No fan is required for low-duty cycle use. Because of the relationship between efficiency, dissipated power, and output power, an amplifier running at 90-percent efficiency can output 21 times more power (at a given dissipation level) than a 30-percent efficient amplifier.[1]

There are some disadvantages to the class-E design. The amplifier must be driven with enough power to reliably switch the transistor on and off. The gates of the MOSFET's are capacitive and present a low impedance at high frequencies. The drive must be powerful enough to switch the transistor on and off quickly and completely. If the transistor is not switched on and off reliably, or if it is switched

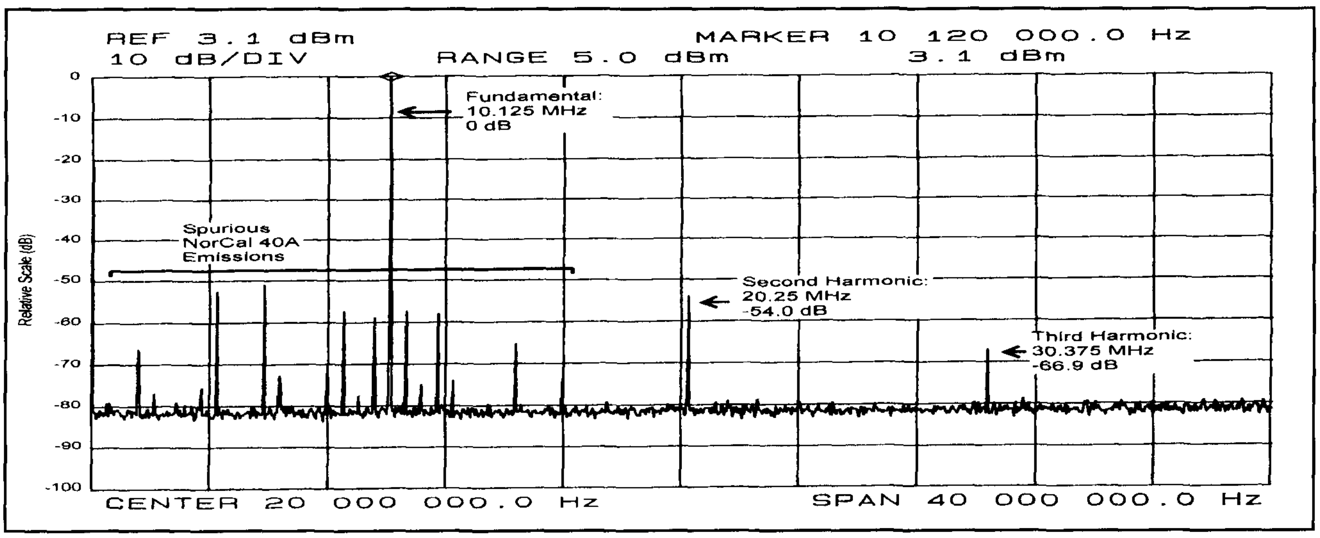

Figure 4. HF spectrum of 30-meter amplifier driven with NorCal 40A transceiver.

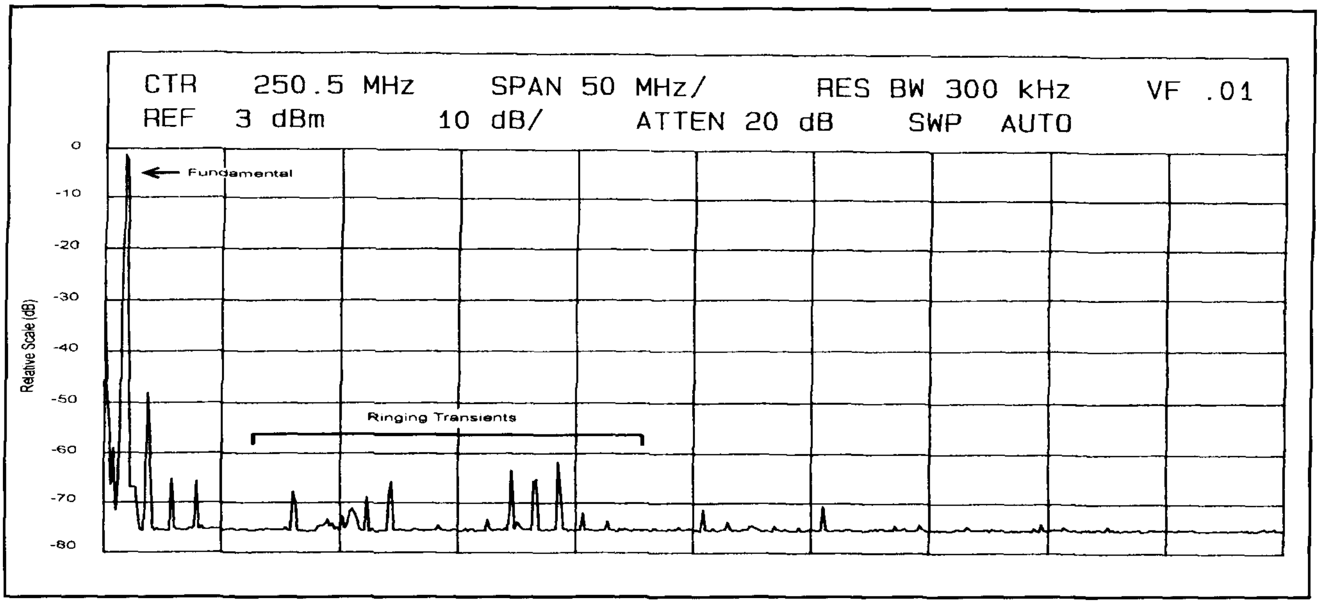

Figure 5. VHF spectrum of 30-meter amplifier driven with modified NorCal 40A transceiver.

too slowly, inefficient operation results which can cause a buildup of heat in the transistor, leading to its failure.

The class-E amplifier is completely non-linear, and its operation is limited to a single band by the ringing circuit on its output. The power output of the amplifier (assuming it is driven properly) depends almost exclusively on the DC supply voltage. This makes it unsuitable for use in SSB transmitters, without external circuitry.[4,5] However, it works very well in CW, FM, and FSK applications, where information is transmitted either by turning a carrier on and off completely (CW) or by changing the frequency, not amplitude, of the carrier by small amounts (FM, FSK).

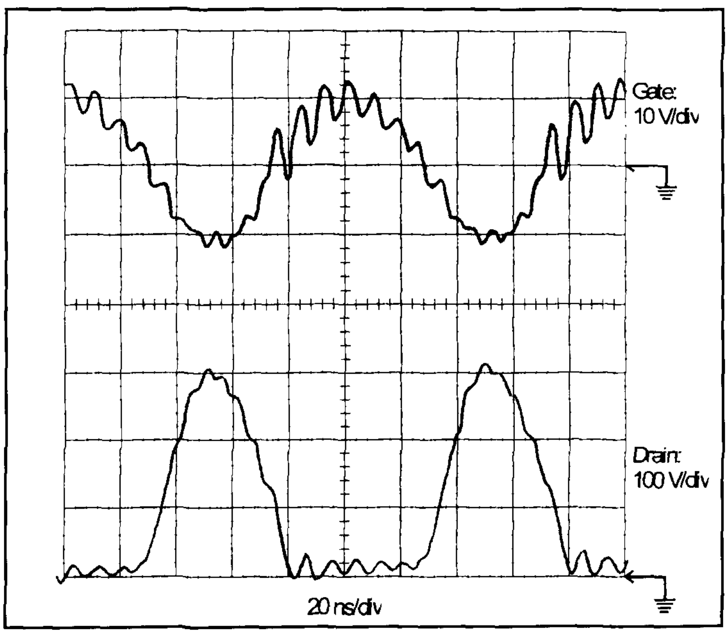

Figure 6. Gate (top) and drain (bottom) waveforms of 30-meter amplifier, being driven at 4.78 watts with a modified NorCal 40A transceiver. The output power is 204 watts. The gate is at 10 V/div, and the drain at 100 V/div. The time scale is 20 ns/div. Note the ringing on both waveforms.

The 30-meter amplifier

The newly developed 30-meter, 200-watt amplifier is very similar in design to the existing 40-meter amplifiers. For the most part, only the component values were changed to operate at the higher frequency. (See **Figure 3** for a circuit diagram of the amplifier and the **Appendix** for a more detailed description of the components used.) Components L1 and T1 match the input of the transistor to the output of the driving transceiver. The impedance of the MOSFET gate is low and mostly capacitive (although there is a small parasitic resistive component). The transformer matches this low impedance to the nominal 50-ohm impedance of the driver. The variable inductor L1 is tuned to cancel the gate capacitance at the operating frequency. The input SWR is typically below 1.4. Capacitor C1 is a high-frequency bypass of L1 to reduce VHF ringing.

RFC1 converts the 0 to 150 volts DC from the power supply into a current source. In conjunction with C6, RFC1 also helps keep RF voltages away from the power supply. C3 and L2 form the main resonant circuit and cause the voltage at the drain to rise smoothly to a maximum and return to zero before the transistor turns on again. C5 and L3 serve two purposes. The first is to form a notch filter for the second harmonic. Together, C5 and L3 resonate at twice the operating frequency (around 20 MHz) and therefore provide a direct path to ground for the second harmonic. Their second function is to transform the 50-ohm impedance of the antenna down to around 10 ohms, a better match for the transistor in this type of circuit.

C4 and L4 form a low-pass filter. They ensure that all of the harmonics fall well below –40 dB of the carrier, the limit set by the FCC. **Figures 4** and **5** are the HF and VHF spectra of the amplifier's output, respectively. These spectra were taken with the amplifier output set to 200 watts. The driver was a modified NorCal 40A transceiver, which will be discussed in the next section. The drive level was 4.8 watts. **Figure 6** shows a picture of the MOSFET gate and drain waveforms.

The 30-meter amplifier requires from 3 to 8 watts of drive power for optimal operation. **Figure 7** is a plot of efficiency versus drive power. We have noticed problems with keying the amplifier at levels below approximately 4 watts, so drive powers exceeding this amount are recommended. The drain efficiency is determined by dividing the output RF power by the input DC power. A better indicator of how much power is being dissipated in the transistor, however, is the total efficiency. The total efficiency is defined as the RF output power divided by the sum of the input DC power and RF drive power. As can be seen from **Figure 7**, although the drain efficiency continues to increase, the total efficiency drops for drive powers above approximately 8 watts. This happens because the drive power doesn't couple to the output, but is dissipated in the transistor as heat. Therefore, once the drive power is high enough that the transistor is switching optimally, further increasing the drive power only increases the amount of power dissipated in the transistor and does not benefit the total efficiency of the amplifier.

An interesting feature of this amplifier, which it shares with the existing 40-meter amplifiers, is the way in which signals couple backwards from the output to the input. When the supply voltage is at 0 volts (this occurs during reception), there is a large capacitance between the gate and drain of the transistor. Therefore, in-band signals couple backwards through the amplifier, including the transistor, with a measured loss of only 8.5 dB. Because of the high sensitivity of the modified NorCal 40A driver (we measured its MDS to be -133 dBm), even with this 8.5 dB of attenuation placed before the antenna, the transceiver is still limited by atmospheric noise. This way, no additional transmit/receive switch circuitry is required for the amplifier. The amplifier operates nearly transparently with the driving transceiver.

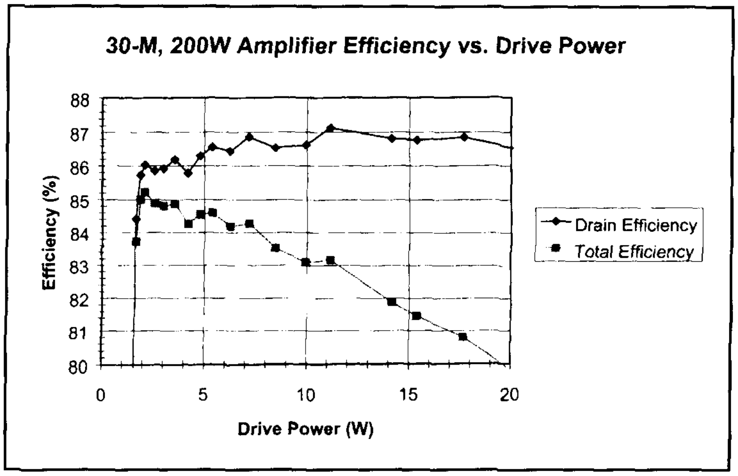

Figure 7. Efficiency versus drive power for 30-meter amplifier.

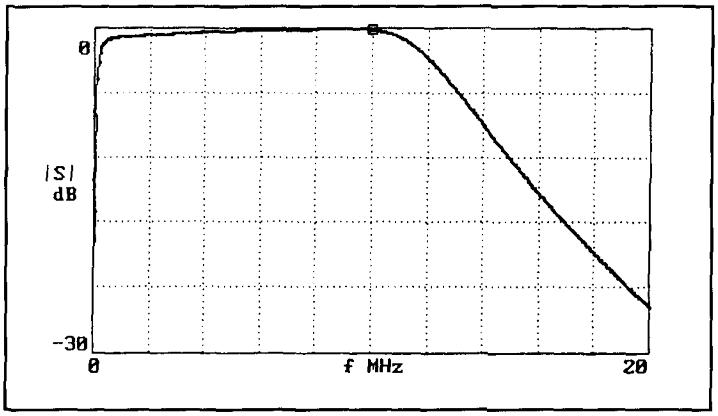

Figure 8. PUFF simulation of modified NorCal 40A low-pass filter.

The NorCal 40A modifications

Because the 30-meter amplifier was aimed at an audience of amateur radio operators, one of our goals was to find a suitable transceiver for driving the amplifier that would be commonly available to hams. We tried using a modified NorCal 40A (NC40A) as the driver. The NC40A is a 40-meter (7-MHz) transceiver; however, Ed Burke outlines a way of modifying the NC40A to change its operating frequency from 7 to 10 MHz.[6] At the heart of this modification is the changing of the IF from 5 to 8 MHz. Both the original and modified NC40A's have a local oscillator at 2 MHz. In the original, 2 MHz is mixed with 5 MHz to give the operating frequency of 7 MHz. In the modified NC40A, 2 MHz is mixed with 8 MHz, and the resulting radio works at 10 MHz.

The modified NC40A, although functional on the 30-meter band, suffered some losses in per-

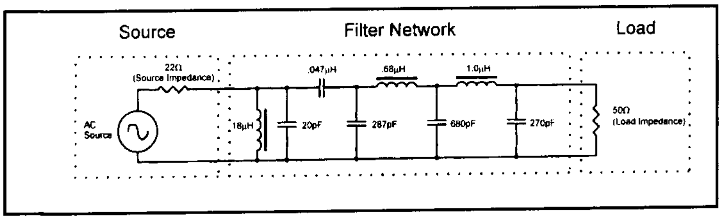

Figure 9. The modified NorCal 40A low-pass filter.

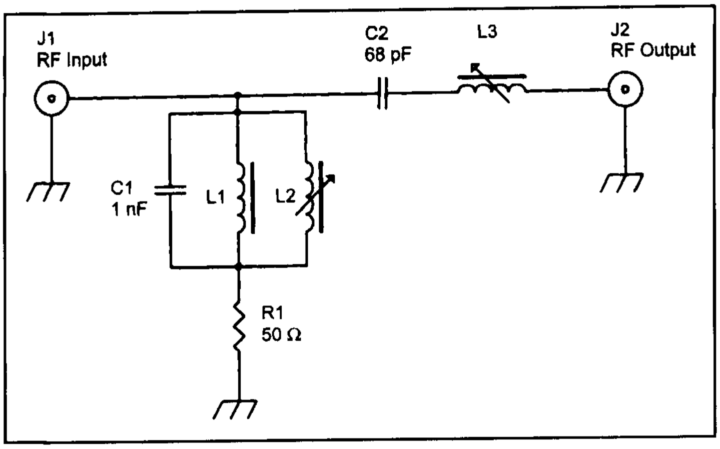

Figure 10. Circuit diagram of the small, low-power diplexer. C1 and C2 are silver micas. L1 is 12 turns of magnet wire wound on a T37-2 core. L2 and L3 are 0.68 µH Toko variable inductors, respectively. R1 is a 1/4-watt precision resistor.

formance. The MDS went from –132 to –126 dBm, the gain (RF input to audio output) from 104 to 96 dB, and the output power from 2.5 to 1.4 watts. To solve these problems, the following changes need to be made to Ed Burke's list of modifications (all component numbers refer to the Wilderness Radio NorCal 40A).

1. A 15-pF capacitor is added in series with C2. This decreases C2's capacitance enough so it can resonate with the secondary of T2 at 10 MHz.
2. C4 is left at 5 pF.

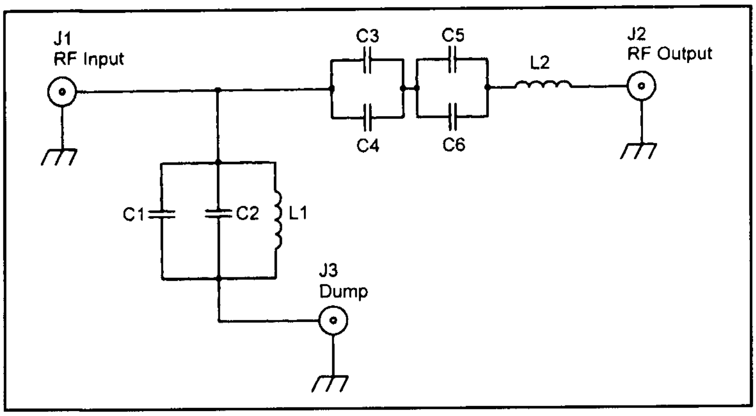

Figure 11. Circuit diagram of the large, high-power diplexer. C1 and C2 are both 560 pF. C3 through C6 are 150 pF. All six capacitors are Cornell Dubilier mica, 1000 volts. L1 is three turns of copper tape with inner diameter of 1 inch. L2 is 12 turns of copper tape with inner diameter of 1-3/8 inch. For the dump, a 5-watt BNC 5-ohm load is used.

The above two steps bring the gain and sensitivity of the modified NC40A back close to their original values. In our NC40As, the sensitivity after the modifications was –133 dBm and the gain was 102 dB.

3. T1 is wound with 21 turns for the primary, four turns for the secondary.
4. Output transistor Q7 is changed to a Motorola MRF237.
5. L7 is wound with 13 turns.
6. L8 is wound with 16 turns.
7. C45 and C47 are replaced with 270-pF capacitors.
8. C46 is replaced with a 680-pF capacitor.

These steps greatly improve the power output of the NC40A. Step 3 matches the output of the driver stage of the transmitter to the input of the power amplifier. Steps 5 through 8 lower the impedance of the low-pass filter, increasing the current and therefore the power from the power amplifier. The values for these components were determined through computer simulations of the low-pass filter using PUFF.[7] Refer to **Figure 8** for the low-pass filter's (**Figure 9**) simulated profile. The output power of the two NC40As we modified was increased to 5.5 and 6.5 watts, making the modified NC40A perfect for driving the amplifier.

9. C18 is replaced with a 680-pF capacitor.

This last change, though not essential, gives an audio frequency of around 600 Hz. Use of the 200-pF capacitor suggested in Ed Burke's modifications provides a higher pitched audio tone and may be desirable to some.

Diplexers

To keep spurious emissions from the NorCal 40A driver and the 30-meter amplifier low, two diplexers were built. A diplexer is a band-pass filter that has an impedance of 50 ohms at all frequencies, terminating out-of-band spurious emissions in a 50-ohm dummy load.[8] **Figures 10** and **11** are schematic diagrams of the diplexers we used. The smaller of our two diplexers, designed to handle low power, is placed between the NorCal 40A driver and the power amplifier. This diplexer is constructed with the same type of variable inductors found in the power amplifier's input network. The capacitors are small silver micas. The case is a small metal box, manufactured by Pomona Electronics, measuring 2.25 x 1.4 x 1.1 inches, with female BNC connectors at either end. This diplexer ensures that the power amplifier is driven with a clean signal at the correct frequency. It attenuates the spurious emissions from the driver which would otherwise be amplified by the power amplifier.

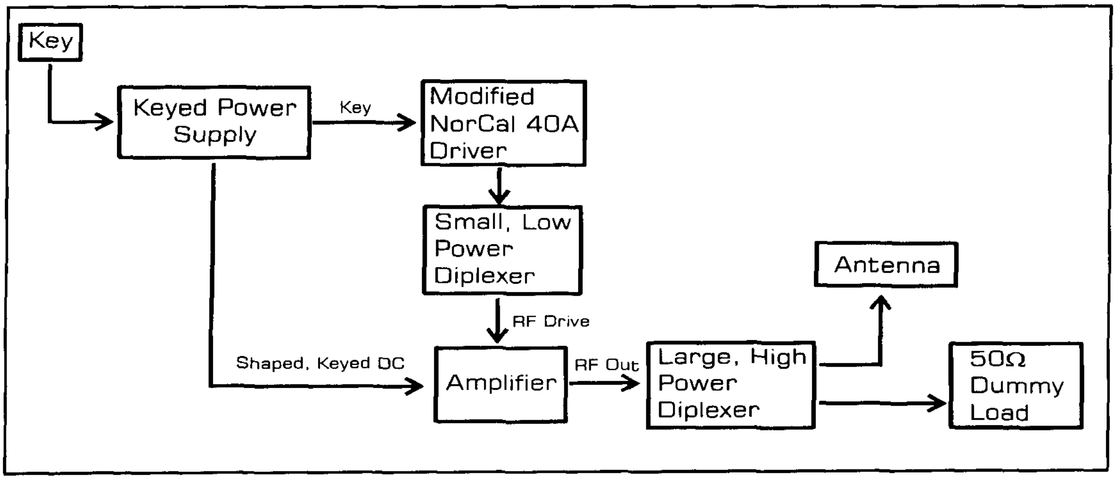

Figure 12. Block diagram showing complete 30-meter CW radio station.

The larger, high-power diplexer is placed after the output of the power amplifier, and helps in attenuating spurious emissions generated by the power amplifier. It is constructed with the same types of air-core inductors and 1000-volt mica capacitors found in the power amplifier and is housed in a 3.5 x 6 x 10-inch box. **Figure 12** shows a block diagram of the complete 30-meter CW radio station.

Future work

There is still much room for improvement on these class-E power amplifiers. One possibility is a built-in pre-amp, which would allow the amplifiers to be driven with smaller amounts of power, allowing their use with low-power QRP transceivers without modification. Another useful addition would be a TR switch of some sort, which would allow for better reception by allowing incoming signals to bypass the amplifier. A small, switching power supply could be developed to replace the current one, which is very large and expensive. It may even be possible to use the RF drive, perhaps divided in frequency, as an oscillator for such a power supply. Finally, it would be interesting to develop class-E amplifiers for other bands and higher powers. Work

Appendix. Description of components used.

Name	Component	Notes
Q1	IRFP440	Use International Rectifier only
C1	10 pF	Cornell Dubilier mica, 500 volt, 5 percent
C6	0.01 μF	Ceramic disc, 1000 volts
C2	100 pF	Cornell Dubilier mica, type CDV19, 1000 volts, 5 percent
C3	1100 pF	Cornell Dubilier mica, type CDV19, 1000 volts, 5 percent
C4	220 pF	Cornell Dubilier mica, type CDV19, 1000 volts, 5 percent
C5	470 pF	Cornell Dubilier mica, type CDV19, 1000 volts, 5 percent
L1	1.5 μH	Toko variable inductor
RFC1	40 μH	Radio frequency choke
L2	9 turns	0.2-inch copper tape wound on 1-1/4 inch O.D. pipe, spread 2.6 inches to fit holes on the pc board
L3	5 turns	18-gauge wire, closely spaced on a 7/32-inch drill bit
L4	10 turns	20-gauge wire, wound on T80-2 type toroidal core
T1	6 turns	26-gauge, stranded hookup wire

on developing a 15-meter class-E amplifier is already underway.

Acknowledgments

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REFERENCES

1. E. Lau, K.W. Chiu, J. Qin, J.F. Davis, K. Potter, and D.B. Rutledge, "High-Efficiency, Class-E Power Amplifiers: Part 1," *QST*, May 1997, pages 39–42; Part 2, *QST*, June 1997, pages 39–42. Packages of amplifier parts only, including pc board, components, connectors, heatsink, and chassis for both 40-meter (300- and 500-watt) and 30-meter (200-watt) amplifiers are available at cost from Puff Distribution, Department of Electrical Engineering, MS 136-93, Caltech, Pasadena, California 91125. Contact Dale Yee by e-mail at <yee@systems.caltech.edu> or by fax at (626)395-2137 for more information. Visit <http://www.systems.caltech.edu/EE/Faculty/rutledge/poweramp.html> for an order form.

2. G.D. Ewing, "High-Efficiency Radio-Frequency Power Amplifiers," Ph.D. Dissertation, Oregon State University, Corvallis, Oregon, 1964.

3. N.O. Sokal and A.D. Sokal, "Class E—A new class of high-efficiency tuned single-ended switching power amplifiers," *IEEE Journal of Solid-State Circuits*, Vol. SC-10, June 1995, pages 168–176.

4. D. Freedman, "Signal Envelope Elimination and Restoration," *QST*, October 1997 page 82.

5. N.O. Sokal, "Signal Envelope Elimination and Restoration in Class-E, High-Efficiency Linear RF Power Amplifiers," *QST*, January 1998 page 80.

6. E. Burke, "NC40 to NC30 Conversion," *QRPp*, Vol. III, Number 4, December 1995, page 63.

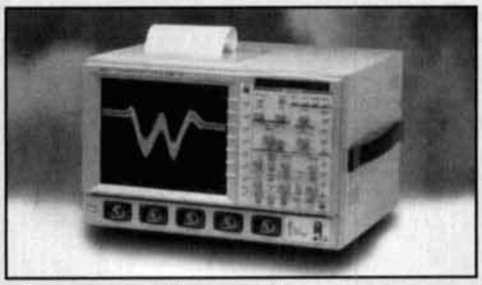
7. PUFF is a program developed at Caltech for design and simulation of microwave integrated circuits, but is very useful for RF circuits, as well. It is available from Puff Distribution, Department of Electrical Engineering, MS 136-93, Caltech, Pasadena, California 91125. The cost is $10. See <http://www.systems.caltech.edu/EE/Faculty/rutledge/puff.html> for more information.

8. D. Newkirk and R. Kariquist, "Mixers, Modulators, and Demodulators," *The ARRL Handbook*, 1996 edition, Chapter 15, page 15.22.

PRODUCT INFORMATION

LeCroy Waverunner™ Color Scopes

LeCroy has introduced the Waverunner series of two- and four-channel digital oscilloscopes. These instruments offer large color displays, 500-MHz bandwidth, 500-MS/s sampling rate, a full complement of I/O capabilities, long data acquisition memory, and an analysis package. The series also include the LeCroy proprietary technology for capturing multiple waveforms in an Analog Persistence™ Oscilloscope display mode.



As an optional feature, Waverunner scopes can include the JTA (Jitter and Timing Analysis) package. This provides the ability to make precision calculations using all the data in a persistence display.

LeCroy Waverunner scopes are priced from $5,990 to $12,990. Additional information is available from the LeCroy Customer Center at (800) 453-2769. You can also check out LeCroy's web site at: <www.lecroy.com/waverunner>.

Active Differential Probe Available

LeCroy has also introduced a high-performance active differential probe for use with oscilloscopes. The new probe, model AP034, offers differential measurement capability with a bandwidth of 900 MHz. It has a choice of x1 gain or 10x attenuation factor, low noise (less than 16 nanovolts per root Hertz from 1 to 100 MHz) and 80 dB, common mode rejection (>10,000:1 at 60 Hz). The AP034's sensitivity range is continuously variable from 2 mV/div to 1 V/div when used with a LeCroy scope. The probe can apply up to ±1.6 volts of calibrated offset (±16 volts when 10x attenuator is used).

The AP034 is small with 0.1-inch input spacing. The probe also offers 1-Mohm impedance and 1.5-pF capacitance loading of each input to ground (2 Mohm and <1 pF between inputs) to minimize degradation of high-frequency signals. It may be operated using power and control signals supplied through LeCroy's ProBus interface. An inexpensive ADPPS adapter is available for non-ProBus oscilloscopes, network analyzers, and spectrum analyzers. The probe inputs are protected from over-voltage and ESD damage.

The AP034 active differential probe can be used in the testing of data storage devices (hard drives/heads), telecom, and high-speed digital and analog design.

Engineers who would like more information on the AP034 can contact LeCroy at (800) 453-2769 or visit the LeCroy Web site at: <www.lecroy.com>.

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THE INVENTION OF THE SUPERHET RADIO

From its early development until its acceptance in the early 1930s

The invention of the superhet radio is without doubt one of the most important developments that has taken place in the field of radio. Although several improvements have been made in terms of adding extra conversions, the basic principle remains the same. In fact, it has remained as one of the core techniques in radio technology for more than 70 years, during which time it has seen many other technologies come and go. Despite its age, there is little to indicate that the superhet radio will not be with us for years to come, making it one of the oldest and most widely used techniques in the radio and electronics industry.

The need for the superhet

The two main functions of a radio set are to provide sensitivity and selectivity. In the early days of radio, the need for these was found time and again. Marconi proved this in his early experiments. However, it was probably most crucial for his first transatlantic tests in December 1901. The Marconi Company was under extreme pressure to succeed in this venture. His company had invested huge sums of money in establishing new records and showing the use of radio, and these tests were no exception. After many setbacks, the tests began. The transmitter in England was to send the letter "S" for three hours each day. Eventually, at 12:30 p.m. on December 12th, faint signals were heard. While this was a major success, it demonstrated that the major weakness in a wireless link was the receiving equipment.

Selectivity was also an issue and was found to be a major problem. It was understood that by reducing the bandwidth, the level of static interference could be reduced. This was one of the major driving factors for some of the developments that later led to the realization of the superhet. Other instances, like the sinking of the Titanic, proved many of the shortcomings of radio and its use at the time. Not only was the radio operator on a ship just within visual distance of the Titanic asleep, and therefore he did not hear the distress call, but it was reported later that many well-meaning amateurs crowded the airwaves—hampering essential communications. This would not have occurred if the receivers had sufficient selectivity and some coordination of communications had been possible.

Beginnings

A number of developments had to be made before the superhet radio could become a reality. One of the major milestones was the development of the thermionic valve (or tube). This came about through Marconi's need to be able to receive weak signals. Professor Sir Ambrose Fleming of University College in London was Scientific Adviser to Marconi's Wireless Telegraph Company, and it was he who designed the transmitter that sent the first transatlantic message on December 13, 1901. He knew that one of the main problems impeding further radio development was the lack of sensitivity in the receiving apparatus. This was chiefly due to the detector.

Some years earlier, Fleming had come across a phenomenon known as the "Edison Effect." Edison had discovered this while

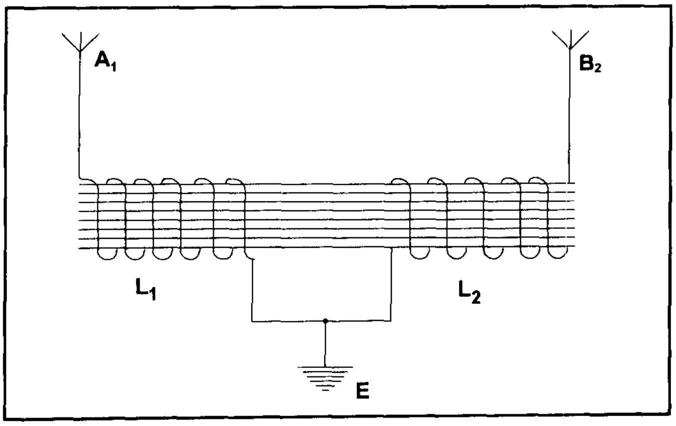

Figure 1. The circuit arrangement for the first beat-frequency receiver.

investigating methods of increasing the life of light bulbs. He placed a second electrode into the bulb in the hopes that he could place a charge onto this electrode and repel what he thought was carbon atoms from the filament that caused blackening on the inside of the glass. While he was performing these experiments, he noticed that a current would flow when a battery was connected between the two electrodes, but only in one direction. Somewhat surprisingly for this great inventor, he never found a use for this phenomenon.

Fleming met Edison on a visit to the U.S. in 1884, and, in the years that followed, Fleming performed a number of experiments on the Edison Effect. He even rectified alternating currents up to frequencies around 100 Hz. These experiments were aimed primarily at the operation of electrical power machinery. However in the fall of 1904, while he was walking down Gower Street outside University College London, he had what he described as a "sudden very happy thought." He asked his assistant, G.B. Dyke, to put his idea to the test and it worked. Then, in November of the same year, he patented the idea, and Fleming's oscillation or diode valve was born. Using his oscillation valve, he was able to rectify the incoming signals so they could be heard on a transducer.

Shortly afterwards, Lee de Forest, a Marconi competitor, started to investigate thermionic technology with a view to designing a rectifier that did not infringe Fleming's patent. He devised a three-electrode device, which he called an audion. Several patents for this idea exist with dates between 1905 and 1907, although the idea is generally dated at 1906.

The addition of this third electrode was to have far-reaching effects. Initially, the idea was only used as a rectifier. It took some time before it was fully used for amplification and in oscillators.

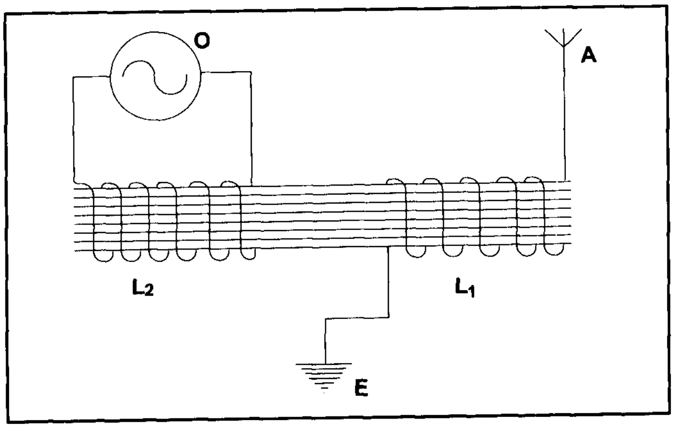

Figure 2. The circuit arrangement for the first heterodyne receiver with the beat signal generated internally.

Idea of heterodynes

One of the first major developments along the road to discovering the superhet was made by R.A. Fessenden. As early as 1901, he reasoned that improvements in wireless signaling (using Morse) might be made if the transmitting station sent out signals in such a way that their combined effect produced the required audio signal at the receiver. One way of achieving this was to transmit two signals differing by a small amount so they were combined and detected in the receiver to give an audio beat note. Fessenden patented this idea in September 1901. His was the first recorded proposal for using the beat-frequency phenomenon.

In the descriptions of the time, the idea for the receiver was like that shown in **Figure 1**. In this diagram, the two signals are shown as being received on different antennas. These are connected to coils L1 and L2, which are wound on a common former. The combined currents flowing in the coils blend their effects to provide a magnetic variation that actuates the diaphragm with an audio note.

Fessenden did not let his idea remain in this state. He realized that transmitting two signals was wasteful of power, and his next idea was to generate the beat signal within the receiver, as shown in **Figure 2**.

The idea was revolutionary, but it was not developed for several years. Although it offered many improvements over existing systems, its efficiency was still very low. Also, there were no valves or other active devices at that time, and it could only be used with a non-polarizing type of transducer. Despite these limitations, its effectiveness was still proved when a range of 3,000 miles was recorded using this type of receiver.

Later, in 1910, the next foundation stone was laid. Transmission tests were carried out between two American cruisers, and the radio operator noticed that the received signal strength greatly increased when the ship's transmitter was in use. This occurred even though the difference in frequency between the transmitter and the received signal was above the audible range.

Investigations into the reasons why this happened were made, and a far more sensitive heterodyne receiver was developed. The diagram for this is shown in **Figure 3**. The operation of the circuit is fairly obvious today. L1 acts as loading coil for the antenna, and signals are coupled into the rectifier or mixer circuit via L2 and L5. The local oscillator is first coupled into the antenna circuit and then into the rectifier of the mixer circuit.

The heterodyne receiver showed itself to be far more efficient than the other methods used at the time. In 1913, tests were carried out between Arlington, Virginia, and the Naval ship, *Salem*; a range of 6400 miles was achieved. During these tests, the heterodyne method of reception not only proved to be more sensitive, but it was also far superior under bad atmospheric conditions. It is interesting to note that, for these tests, a high-frequency alternator and not a valve generated the internal beat-frequency signal, although thermionic technology was beginning to be used more widely.

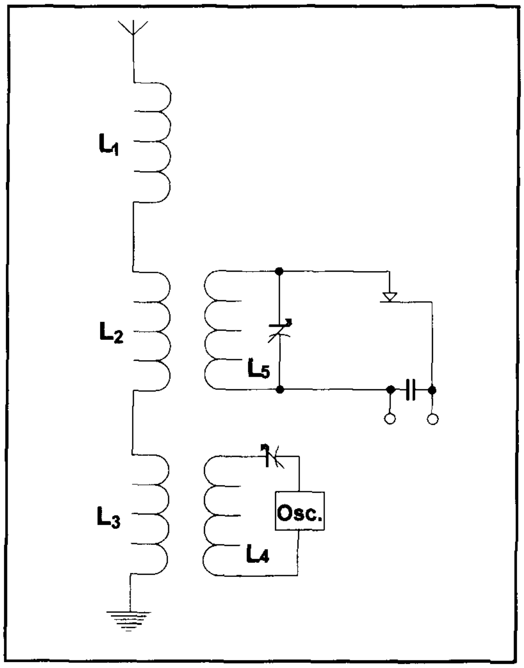

Figure 3. The diagram of an early heterodyne receiver using a rectifier.

The valve as an amplifier and oscillator

Initially, the triode was only used as a detector. Its principles of operation were not fully understood. Although de Forest tried to use it to provide low frequency amplification, his initial attempts failed. It was discovered later that the grid-blocking capacitor used for rectification was not required in an amplifier. There was also a suggestion that he used the wrong capacitor value.

Eventually, de Forest succeeded and was able to use the triode as an amplifier. In 1912, he built a cascade amplifier using first two then three audions. With the aid of his friend, John Stone, he demonstrated the idea to a telephone company. Despite the fact that the performance was poor, the company saw the potential of the idea. They took it on and considerably improved the performance of the triode. Development was fast now that companies including AT&T and GE were involved. Less than a year after the initial demonstration by de Forest and Stone, these companies had transformed the performance of the amplifier valve—enabling them to be used as telephone repeaters and to provide significant improvements in telephone performance.

First regenerative receiver

Once the triode had been shown to amplify, the number of uses were apparent. It was shown that by adding positive feedback or "regeneration," the degree of amplification could be considerably increased. The idea came to light in 1913, when a number of people demonstrated the regenerative receiver. In the U.S., a very bright young engineer named Edwin Armstrong described his new circuit to the Institute of Radio Engineers (founded in 1912, the forerunner of today's IEEE). Shortly after this, de Forest applied for a patent for the idea, and Irvin Langmuir, as well as Meissner in Germany, also demonstrated the concept. It is difficult to determine who was actually first. The courts ruled in favor of de Forest, while the engineering community honored Armstrong.

Round's autodyne

Valves were expensive, and many people developed circuits using as few as possible to achieve the required performance. An innovative type of receiver designed by H.J. Round in 1913, called an autodyne, used the heterodyne principle with just one valve. The diagram in **Figure 4** shows the basic circuit. In this receiver a single valve is used to amplify the incoming signals, generate the heterodyne signal, and then detect the signals. Effectively, it was a single-valve direct-conversion receiver using the valve as a self-oscillating mixer. Signals from the antenna were coupled into the input of the autodyne via L1 and L2. L3 acted as the feedback path for the oscillator. For operation, C3 was tuned so the circuit oscillated slightly above or below the incoming signal. C1 and C2 were adjusted to give the strongest signal.

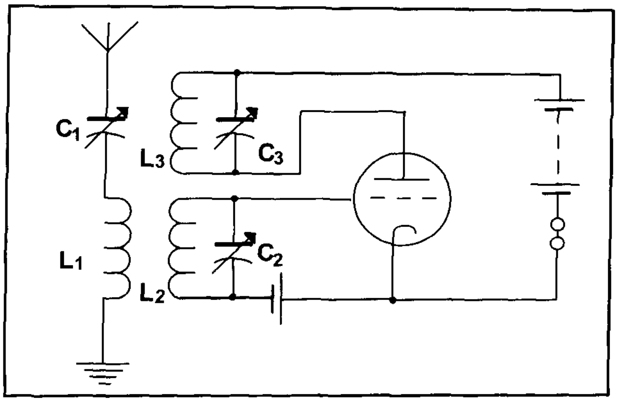

Figure 4. Round's autodyne.

Although crude by today's standards the circuit was successful, enabling a heterodyne receiver to be built with the minimum of valves.

Wartime developments

The start of hostilities in 1914 gave new impetus to the development of radio receivers. The army quickly saw the possibilities for the receiver's use in everything from communications to detecting the presence of aircraft. For all of these requirements, the need for higher amplification and selectivity was of paramount interest. A further requirement was for operation on what were then considered to be very high frequencies between 500 and 3000 kc/s (kHz). Operation on these frequencies was particularly difficult because of the instability arising from the inter-electrode capacitances within the valves. At this time, only triodes were available. While operation below about 500 kHz was possible, keeping any circuits stable at frequencies above this became increasingly difficult. Apart from this, low-frequency noise was a major problem. Additionally, the detectors were such that for low-level signals there was very little response.

A number of people started to address the problem. On the Allied front, there were H.J. Round, one of Marconi's engineers, M. Latour, and later Edwin Armstrong. For the Germans, W. Schottky spent a considerable amount of time working at it and made some progress.

H.J. Round spent a considerable amount of time improving the performance of valves. One of the major problems was the capacitance between the anode and grid, which caused instability and limited their gain at high frequencies. Trying to reduce this, Round devised a number of new types of valve where the anode and grid were taken out through opposite ends of the glass envelope. In this way, he made improvements in performance, but he did not completely remove the problem of spurious capacitance.

The performance was improved in other areas as well. For many years the conduction mechanism within valves was not understood properly. It was widely believed that some gas molecules were required inside the envelope for correct operation. A number of scientists in the U.S. and Germany began to realize that this was not the case. In 1915, Irvin Langmuir proved conclusively that gas molecules were not required. As a result, a new generation of what were termed "hard" valves with high vacuums were introduced. These provided a number of advantages over their "soft" predecessors. They were able to operate with higher anode voltages, and it was now possible to coat the cathodes to improve their emission. Previously, the gases soon contaminated any coatings.

Levy's receiver

A French electrical engineer named Lucien Levy produced a set that used the principles of the superhet. He did not succeed in devising the full solution in the form of the superhet as we know it today, but his design did use the principle in a form that he claimed would completely eliminate "parasites and ordinary interference." His method of reception involved converting the signals down to a frequency where they could be easily separated from one another. For

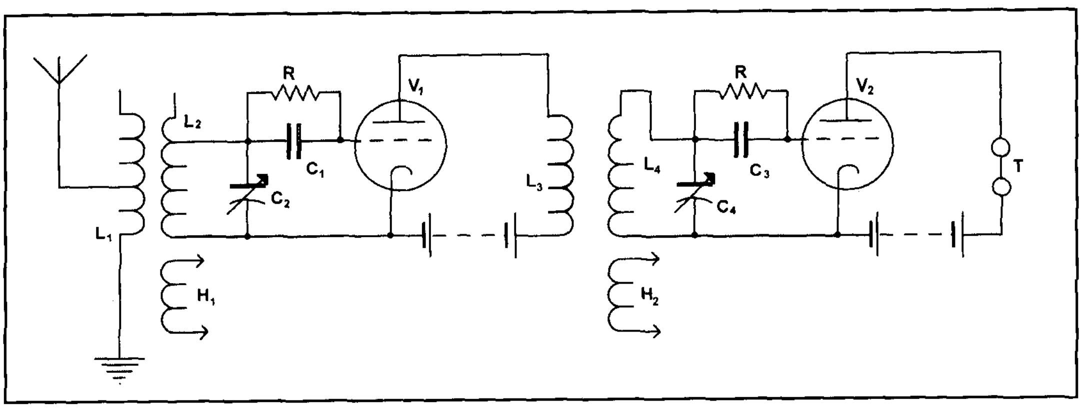

Figure 5. Levy's idea for a receiver using the superheterodyne principle.

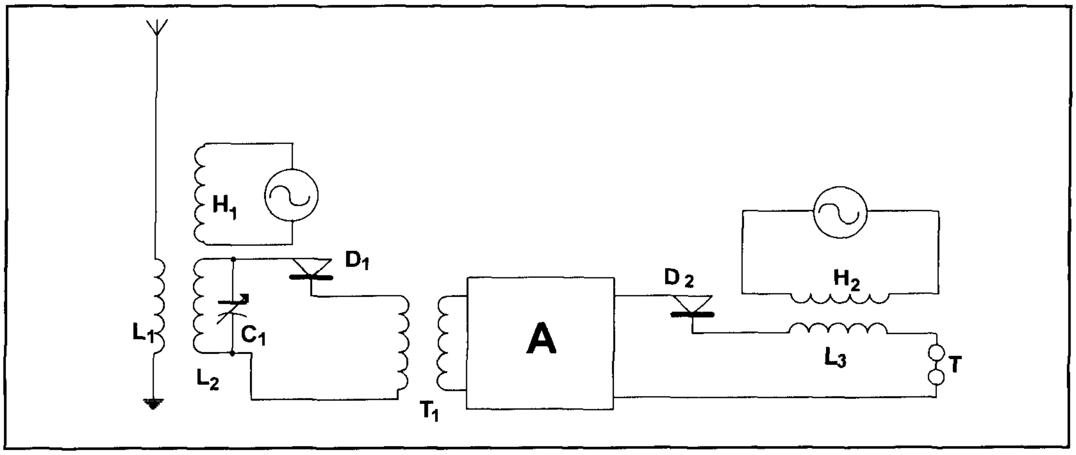

Figure 6. The circuit arrangement for Armstrong's idea of a superhet receiver.

this he used a tuned or variable intermediate frequency amplifier, as shown in **Figure 5**.

In the circuit, L1 and L2 provide the antenna coupling and tuning. The heterodyne signal is also applied here. Valve V1 is set up for grid detection, as indicated by the R1/C1 resistor capacitor combination. The anode of the valve is tuned to the intermediate frequency, and this is applied to the grid of the second valve. This, too, has a further oscillator to bring the signals down to audio frequencies for conversion to audible signals via a transducer. In his documentation, Levy indicated that a stage of amplification could be inserted between the two rectifiers of mixer valves to provide additional levels of amplification.

There were several advantages to Levy's receiver. The lower frequency meant that selectivity could be improved. Additionally, valves could be used to provide far greater levels of amplification. Operating at the lower intermediate frequency, they were able to run at higher levels of gain without bursting into oscillation. A patent covering Levy's idea was filed on August 4, 1917.

Armstrong's superhet

Although Levy's receiver used the basic principles of the superhet, it did not use them in exactly the same form we use today because his receiver had a variable intermediate frequency. Armstrong, at that time a Major in the U.S. Army, was approaching a slightly different problem. His needs were for much greater levels of amplification as well as selectivity. As a result, he needed further stages in the intermediate frequency area of the set. Armstrong reasoned that better performance could be achieved by converting the signals down to a lower frequency where they could be amplified and tuned more easily. However, he also reasoned that a fixed frequency intermediate frequency stage would enable several stages of tuning to be placed in series and, as a result, greater levels of selectivity as well as the achievement of high levels of amplification.

The basic idea for the receiver is shown in **Figure 6**. Here the incoming signals are applied to the transformer consisting of L1 and L2, and the heterodyne signal is applied via H1. RF selectivity is provided by the fact that the resonance of the input circuit can be tuned by C1. The combined input and heterodyne signals are applied to D1, which acts as a mixer. The action of the mixer generates signals at the intermediate frequency that are tuned and applied to the intermediate frequency amplifier by T1.

Once through the intermediate frequency amplifier, the signals are detected using a diode. It was stated that if continuous wave signals were used, a second heterodyne, as shown in the diagram, was required to make the signals audible.

Armstrong was the first to realize the superheterodyne principle in a practical receiver. Experiments with the Armstrong's proposals were undertaken by the Division of Research and Inspection of Signals Corps of the American Expeditionary Force. An eight-valve set was comprised of first detector (mixer), heterodyne oscillator, three stages of intermediate frequency amplification, second detector, and two stages of low frequency amplification.

On the German side, W. Schottky proposed a similar scheme. Schottky was manager of the Siemens Laboratory (Schwachstromkabel Laboratorium), and was investigating receiver selectivity and amplification. Outlines of his idea were published in the Journal of the Schwachstromkabel Laboratorium between February 25 and March 16, 1918. He applied for a patent to cover his ideas on June 18, 1918. While this predates Armstrong's patent

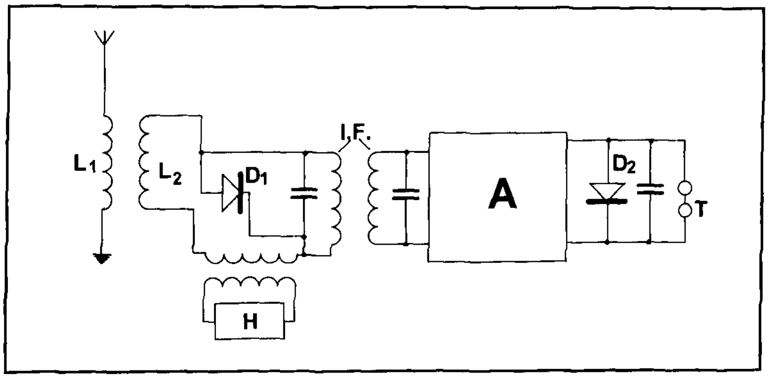

Figure 7. Schottky's proposed circuit for a superheterodyne receiver.

application, Schottky could not develop his ideas in a practical form, and, as a result, Armstrong is rightly credited with the invention of the superhet.

First amateur transatlantic tests

After the end of the First World War, the need for the performance of the superhet receiver was not as great. Valves were expensive, so the idea lay almost dormant for some time. In fact, comparatively little was heard of the idea until it surfaced again for the first shortwave transatlantic tests between the U.S. and Britain. Here it was again used because of its superior performance in terms of selectivity and sensitivity.

With the resumption of amateur activity in Europe, many new ideas started to surface. Amateurs were increasingly using the shortwave bands as the congestion on the longer wavelengths was increasing.

It was quickly noticed that radio amateurs in North America were achieving very long-distance contacts. A 500-watt telephony station had been heard at a distance of about 2,000 miles. Although commercial operators started to notice the distances that were being achieved, it was the radio amateurs who took much of the initiative. It was decided that experiments would be undertaken to see if the Atlantic could be spanned on the shortwave bands.

There were a number of difficulties. World War I had hit Britain very hard, and, in addition, the authorities were very slow in allowing amateur operation to recommence after the hostilities. Accordingly, the first licenses were not granted until May 1920. When amateur operation did resume, there was little equipment and receiver technology was not as advanced as it was in the U.S.

The first transatlantic tests took place in February, 1921. They aroused considerable interest in the radio press and within the amateur radio fraternity. However, despite the support they received, no signals were heard. One of the reasons given for the failure was the poor receiving equipment used in Britain.

To overcome any fears of failure due to poor equipment, an American amateur named Paul Godley (2ZE) was sent to Britain with an Armstrong superhet. The next tests were arranged to take place between December 8 and 17, 1921. In preparation, Godley first set up his equipment in Wembley, North London, but he found the levels of electrical noise too high. Accordingly, he moved to a far better location just outside Ardrossan, not far from Glasgow in the southwest of Scotland. Here he was able to erect a 1,300-foot-long wire on poles about 12 feet above the ground.

Using this station Godley successfully managed to copy signals from across the Atlantic. He picked up his first station just after midnight on December 9th. Although he only managed to decipher the callsign on the first night, he did manage to copy the whole of the message a couple of days later.

Although other British stations managed to copy some transatlantic signals, the tests did prove the need for sensitive and selective receiving equipment.

Other receivers

Despite the success of the superhet in these transatlantic tests, the receiver still did not catch on. Cost was one of the major limiting factors. Much of the reason behind this was that the superhet used more valves than other types of receiver, and it appeared that valves were being wasted on functions apart from amplification. The oscillator provided no amplification and, in the listener's view, was a passenger providing little purpose for the set.

The number of valves was of major importance because they were very expensive. Additional items, like batteries for the valves, were also costly. A typical valve filament would consume about 3/4 of an amp, and this had to be supplied from an accumulator because indirectly heated valves were not generally available at this time. This meant that the cost of a filament supply for an eight-valve receiver was virtually prohibitive for most applications. It was not possible to use an AC derived supply because hum was transferred to the signal.

Another reason for the lack of popularity of the superhet was that its superior performance was not required. In the early 1920s, there were comparatively few stations broadcasting, so the higher levels of selectivity provided by the superhet were unnecessary.

Several attempts were made to overcome the problems of the superhet by designing the circuits in which one valve performed more than one function. Round's autodyne was an early example of this type of receiver; however, its shortcomings meant that it was not widely used after the war. Not only were all the tuned circuits interdependent, making it difficult to operate, but it also radiated the oscillator signal that caused interference to other nearby sets.

Other techniques of using one valve for more than one function were devised. Harry W. Houck came up with a system called the harmonic heterodyne in 1923. This was loosely based around the idea of Round's autodyne, but it overcame a number of the problems previously encountered.

The basic circuit of this design is shown in **Figure 8**. The input circuit consisting of L1, L2, and C1 is tuned to the incoming signal. However the feedback circuit (L3, L4, and C2) is tuned to a sub-multiple of this frequency. The required harmonic (normally the second harmonic) is used to provide the required local oscillator signal, which mixes with the incoming signal to convert the signals down to the intermediate frequency and pass them on to the IF amplifier via the first IF transformer consisting of L5, L6, C3, and C4.

The advantage of this circuit was that the interdependence of the oscillator and input tuning circuits would be greatly reduced as their resonant frequencies were separated by a very wide margin. It was also said that radiation of unwanted spurious signals was reduced. This would certainly be true for the fundamental. In addition to this, signal loss compared to Round's autodyne was reduced by using separate tuned circuits for the oscillator feedback and input tuning.

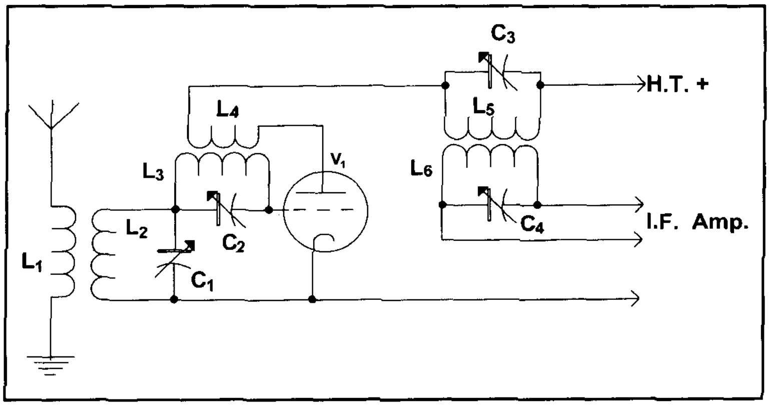

Figure 8. Houck's harmonic heterodyne.

Valve improvements

There was a significant amount of work taking place with thermionic technology. Earlier work by Langmuir and Round had sparked significant improvements, but there were still some major limitations—particularly that of the anode grid capacitance that gave a feedback path and the resulting oscillation.

The solution came in 1925 when a fourth electrode was added to the triode. The additional grid was named the screen grid and was placed between the original or control grid and the anode. This screen grid had a positive potential applied to it, but less than that on the anode. It effectively reduced the parasitic capacitance between the control grid and the anode to zero.

These new tetrodes quickly became popular, but they had an annoying discontinuity in their response curve. This was caused by electrons passing through the screen grid, hitting the anode, and bouncing off. This problem was solved by the introduction of the pentode, which incorporated a suppressor grid that was connected to zero volts. This grid was placed between the screen grid and the anode as shown in **Figure 9**.

Another major improvement in valves took place with the introduction of indirectly heated cathodes. Originally, the heated element also formed the cathode. This restricted the way in which the cathode could be biased. Individual filament batteries were usually required, and AC could not be used or very significant levels of hum were introduced into the circuits. By placing a separate heating element inside the cathode to heat it, the two elements were electrically separated, enabling AC to be used to heat the cathode, and also enabling the cathode to be biased as required.

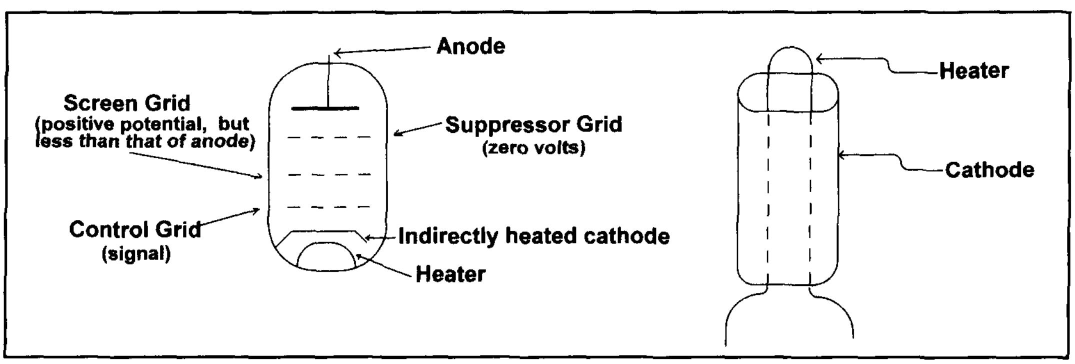

Figure 9. Make up of a valve.

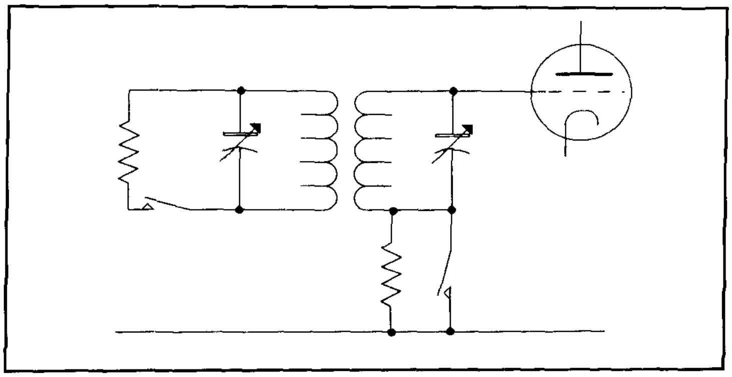

Figure 10. A circuit arrangement used in a 1931 set for varying its selectivity.

Once this idea was introduced, mains-driven equipment became possible and the cost of owning and operating valve equipment dropped rapidly.

Unfortunately, the improvement in the RF performance of valves prolonged the use of TRF receivers, particularly in Europe where the need for the superhet was not as great at this time. However, the performance of the superhet was soon required.

Acceptance

With the enormous rise in the number of broadcast stations in the U.S., the selectivity of the superhet receiver was needed and there was a large increase in its use at this time. Not only was there a great increase in the number of stations, but the power levels they were using also increased. Both of these factors heightened the need for increased levels of selectivity. In addition, the developments that had occurred in valve technology made receivers somewhat less expensive—although they still represented a significant cost and were not affordable by all the households of the time.

A similar need did not arise in Europe for some years. Here, little interest was shown in superhets. In fact, a design that appeared in *The Wireless World* (a UK radio magazine that survives today as *Electronics World*) in 1923 produced little interest. There were significantly lower numbers of stations, which was partly due to the more restrictive nature of legislation associated with radio. This may have been due in part to the recent war. However, at the beginning of the 1930s, broadcasting levels increased significantly and the need for superhet receivers was felt.

Receiver problems

Naturally, the receivers used in these early days suffered from many problems. Initially, two tuning controls were required: one for the RF or antenna tuning and the other for the oscillator. This was overcome by the development of the ganged-tuning capacitor.

Another problem was that superhets were prone to the reception of spurious whistles. Often stations could be heard at several places on the dial—especially if the signal was strong. These problems arose from the harmonics generated within the oscillator and from the poor strong signal handling characteristics of the amplifiers and mixers.

Image rejection was also poor. Prior to about 1928, many sets used an intermediate frequency (IF) of between 40 and 50 kHz. This enabled a very stable and selective amplifier to be constructed, but it did nothing for the image response rejection. As techniques improved, the favored intermediate frequency rose, and a frequency of about 180 kHz became popular. Frequencies rose again to around 450 to 470 kHz, where they have remained for AM sets to this day.

The audio quality was usually poor. As a result of the quest for selectivity, relatively narrow-band filters were used, although with not particularly good shape factors. Accordingly, many sets used an arrangement like that shown in **Figure 10** to vary the selectivity according to the reception conditions. By reducing the Q of the tuned circuits using the shunt resistors, the bandwidth of the circuits was widened allowing improved fidelity.

Summary

By the early 1930s the superhet was firmly established in all corners of the world as the optimum type of receiver. Although TRFs and direct conversion sets would continue to be used in niche areas, it was the superhet that was to dominate the market until the present day. Since the 1930s, many improvements have been made to receiver technology, but the principle of the superhet still forms the basis of virtually all sets in use today, both professional and amateur. ■

BIBILOGRAPHY

1. A.T. Witts, *The Superheterodyne Receiver*, Chapter 1, Pitman, London, 1942.
2. Albert Hall, *The True Road to Radio*, Ferranti, UK, 1931.
3. John Clarricoats, *World at their Fingertips*, R.S.G.B., 1967.
4. J.T. MacGregor-Morris, "Sir Ambrose Fleming," *Notes and Records of the Royal Society*, Volume 2, 1955, pages 134-144.
5. W.A. Atherton, *From Compass to Computer*, Macmillan, 1984.
6. K.R. Thrower, "Evolution of circuit design for a.m. broadcast receivers: 1900–1935," *Journal of the Institution of Electronic and Radio Engineers*, Volume 56, (UK), Oct/Dec 1986, pages 325–341.
7. Ladner and Stoner, *Short Wave Wireless Communication*, Chapman and Hall Ltd., London, 1938.
8. Keith Geddes, *Guglielmo Marconi 1874–1937*, Science Museum, London, 1979
9. *The Wireless World*, Volume XI, October 1922 through March 1923.
10. Anthony Constable, *Early Wireless*, Midas Books, UK, 1980.

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THE EASY MLA-2500 CONVERSION

A replacement tube solution

The MLA-2500 power amplifiers available for purchase have become scarce since the articles for conversion to more popular tubes appeared in *QST*[1] and *Communications Quarterly*.[2] Also, the 8875 tubes are less in demand because, except for amateur use, there's little industrial/commercial use for the tube. As production decreases, we can expect the price of a new 8875 to continue to climb from its present level, which is near $450 each.

The MLA-2500 series is one of the most compact linear amplifiers in its power class. Its intermod performance at 1250 watts RF output, with 100 watts of drive to 8875 tubes, as measured by the American Radio Relay League (third and fifth order distortion products of –35 and –37 dB, respectively) is relatively good by the standards of modern amplifier designs.[3] Therefore, it seems reasonable for MLA-2500 owners to try to find an alternative, more economical and readily available replacement power tube.

Replacement tube solution

The 3CX400A7 and the 3CX800A7 are both candidates for consideration. Neither is nearly as economical as the tetrode tubes of previously published conversions. However, these tube types are both very similar to the original 8875. The 3CX400A7 is electrically identical to the 8875. Both the 3CX400A7 and 3CX800A7 use the same socket, so this conversion is very easy. There's a good chance both tubes will be available for some time, as they are both used in numerous types of commercial, industrial, and medical equipment. The continuous and ongoing use of these tubes will maintain the demand, resulting in a reasonable supply and price stability. At today's prices, the 3CX400A7/8874 is about $100 less than the 8875, and the 3CX800A7 is about $80 less. Therefore, to retube the amplifier at current prices, there's a net savings of $150 to $200, and this may increase due to greater competition. In the future, as demand for the 8875 is further reduced and manufactured quantity decreases, the savings may be even greater.

If you've recently purchased your MLA-2500 or MLA-2500B, you'll want to check out the power supply portion to determine the starting point. Prior to turn-on, make a visual inspection for dirt, loose, wires, or damaged components. Ideally, you would have done this prior to purchase. Because many of these amplifiers are available without tubes, or with weak or bad tubes, you'll need to take some additional precautions. Take off the top and remove the existing tubes. Make certain the plate connections aren't near or touching anything. Initial turn-on should be at less than full-line voltage. I prefer to use a variac or a 150-watt light bulb in series with the primary power line. Half-voltage may easily be achieved by strapping the PA primary power for 220 volts AC operation and using 110 volts AC primary power. A low-voltage/soft start is a good idea when the MLA-2500 hasn't been operated for some time. This allows the electrolytic capacitors to "form" and not destroy themselves. Depress the plate voltage meter switch and depress the tune/operate switch to the tune position.

Observe **EXTREME CAUTION**, as there is access to **LETHAL VOLTAGES**. With the amplifier connected to a primary power source, turn the power switch to "on"; nothing should happen. With one hand behind your back, and

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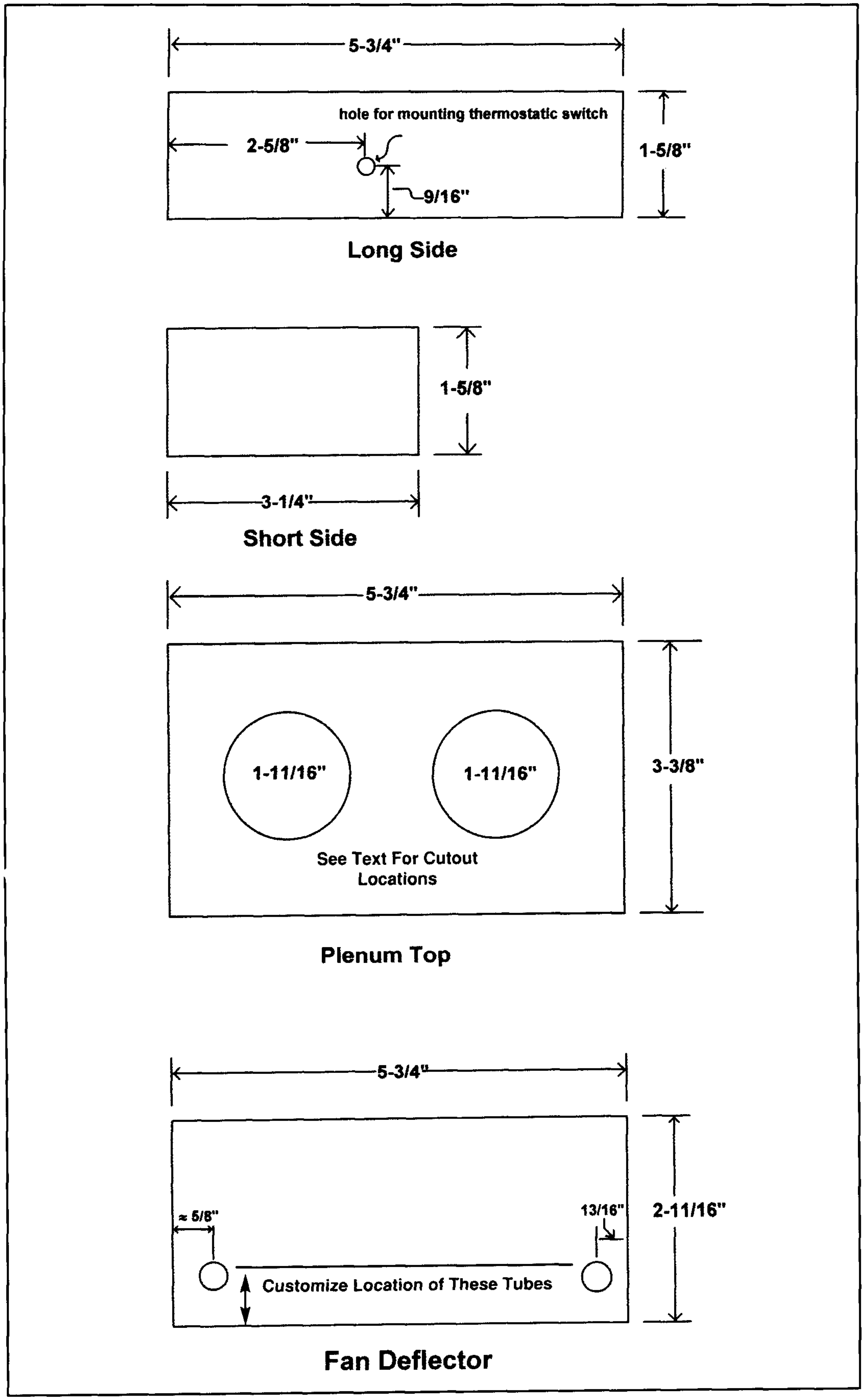

Figure 1. Plenum fabrication.

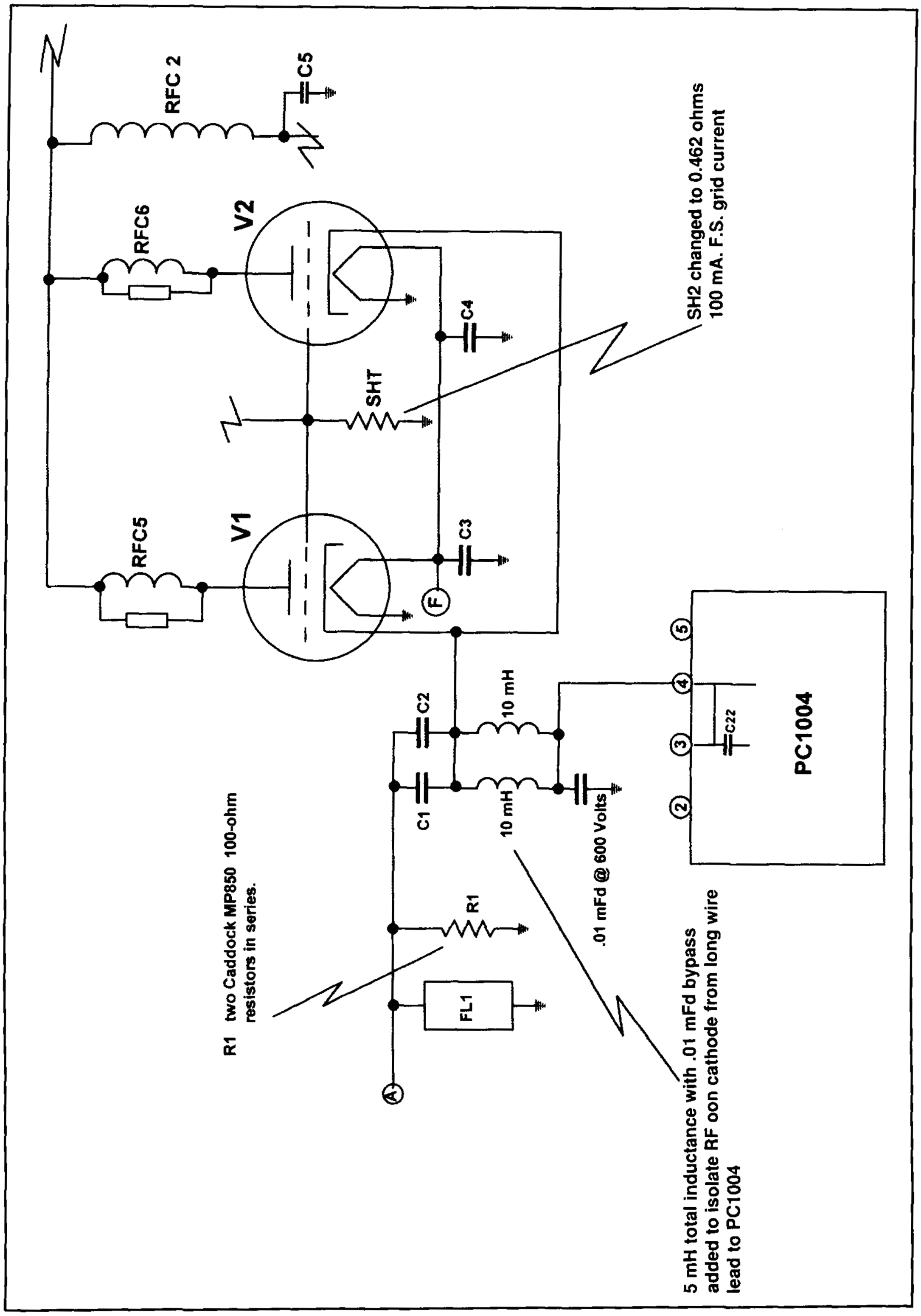

Figure 2. Modified MLA-2500B schematic.

using only one finger of the other hand, momentarily close the micro switch (interlock) at the top right rear of the PA and observe the plate voltage meter. If there are no problems, it should rise to 1000 to 1500 volts. The fan will also operate. If there's no arcing or other problem, close the micro switch and hold it in place for a minute or so to assure operation.

After the initial turn-on, operate the amplifier at full voltage for a longer period to verify power supply performance. Make sure the micro switch is closed temporarily. I used a

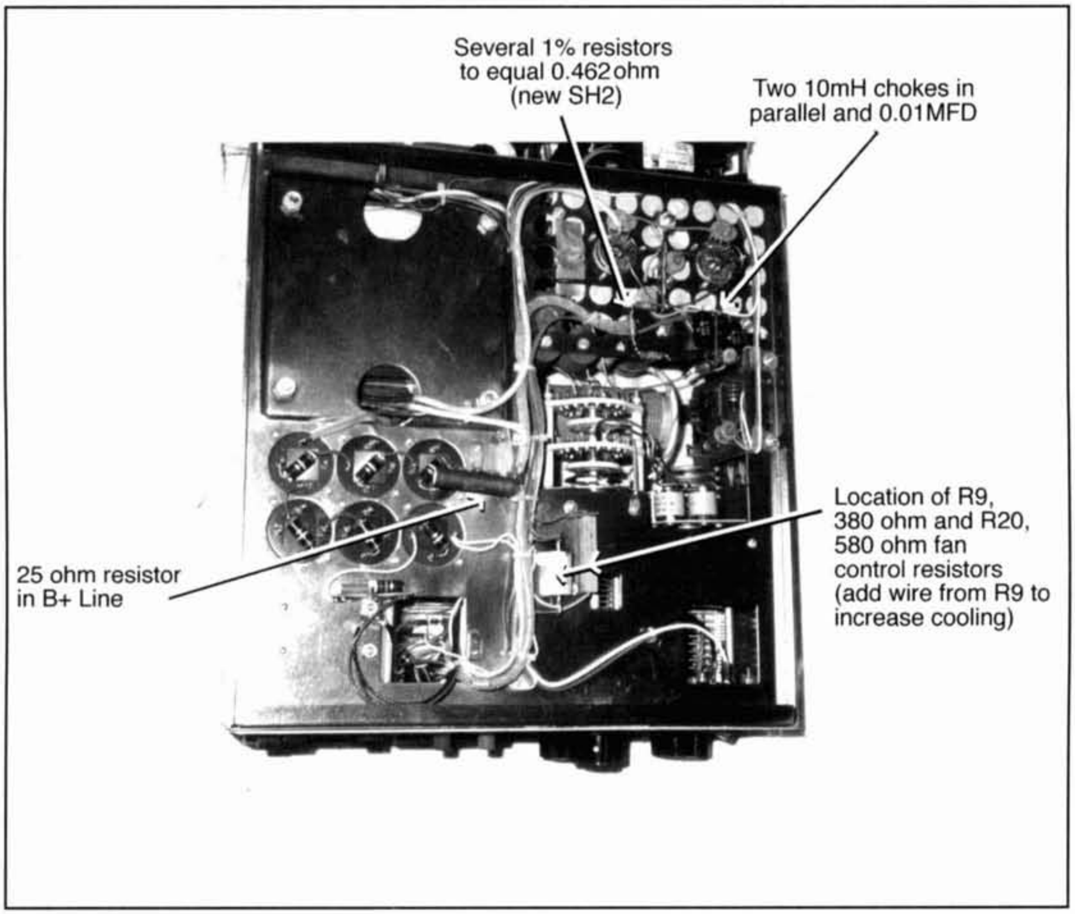

Photo A. Bottom view without cover.

plastic bag tie wrap. Carefully make needed measurements to verify that all the appropriate voltages are available.

If you wish to try operation with tubes received with your amplifier, this is the time to do so.

The 3CX400A7 conversion

The 3CX400A7/8874 is electrically identical to the 8875, but there are two mechanical differences. First, and most important, the 3CX400A7 cooling air must be applied axially through the radial anode cooling fins, like the famous 4CX250 series. The 8875 cooling air is applied transversely across the horizontal fins. Second, the plate power dissipation rating of the 3CX400A7 is 400 watts, while that of the 8875 is 300 watts. The additional 100 watts per tube provides additional margin for cooling; however, it does not provide additional RF power output.

We need to construct a plenum or other device to change the direction of the air flow from parallel to the chassis to vertical, and through the cooling fins of the 3CX400A7/8874 anode. A similar change in direction of the air flow was made in the 4CX400A conversion.[2] The solution also involved making a plenum.

Seal the holes in the bottom of the chassis. I used duct tape, as I did for the 4CX400A conversion. It should be possible to use a very thin fiberglass material; however, I found the duct tape satisfactory.

Next, fabricate the side panels of the plenum. While G-10 is an ideal material, I used a good grade of Masonite for the 4CX400A conversion, and it has been operational for two years. **Figure 1** shows the approximate dimensions of the side panels. When cutting the two pieces for the side panels, leave an extra 1/8 inch in each dimension, so you can file them to fit. Two "L" brackets are fabricated from soft aluminum strips about 1/2 inch wide. Install these brackets to hold the side panels and to assure there will be metal to drill holes for mounting to the bottom of the MLA-2500 chassis. Moving the thermostatic switch to approximately the center of the long panel, allows it to respond equally to both tubes. The hole vacated by the stud holding the switch now provides a convenient mounting hole for one of the "L" brackets to hold the long panel in place.

A second long panel "L" bracket and the new horizontal mounting position of the thermostatic switch may be combined using the same screw that holds the solder tie point. Because the mounting position on the plenum long side is now very close to the plate RF choke, be certain to use a nylon screw to mount the thermostatic switch. (I didn't and I had to repair the RF choke; see further discussion on the plate RF choke below.)

The plenum top is made from thin G-10 material. The approximate size is shown in **Figure 1**. Custom fit the dimensions to the plenum. Your dimensions may be slightly different from mine. Locate the centers of the two tubes and cut two holes in the G-10 material top to fit the anode coolers of both tubes.

With the two tubes in their sockets, adjust the height of the side panels to ensure the tubes' cooling jackets project above the G-10 plenum top by at least 1/8 of an inch. This will force the cooling air through the anode cooler of each tube. Check to verify that the tubes can be removed from their sockets with the plenum in place. There should be enough room in the plenum top to allow a limited rocking motion to remove each tube. The pin contacts of the 11-pin socket should be loose enough to allow tube removal. If necessary, a small pointed object may be used to release the tension of the socket pin contacts.

Mount the fan thermostatic switch in the hole on the long side panel, so it's approximately centered with respect to both tubes. The thermostatic switch wiring, when rerouted, will reach the new mounting position.

Bolt the plenum to the floor of the amplifier chassis, using existing holes, and to the screws holding capacitors or other components to the underside of the amplifier. You may also drill new mounting holes. Seal all plenum joints with duct tape. A larger seal is necessary to close the opening between the fan deflector and the G-10 top plate. I used another piece of G-10 material mounted to the screws holding the fan. I bent it out slightly to meet the plenum top plate. The dimensions of this piece are also shown in **Figure 1**. Close the openings with duct tape to force all of the air from the fan through the tube cooling fins.

Plate RF choke

My experience with the MLA-2500, when performing the 4CX400A conversion, prompted me to check the resonance of the plate RF

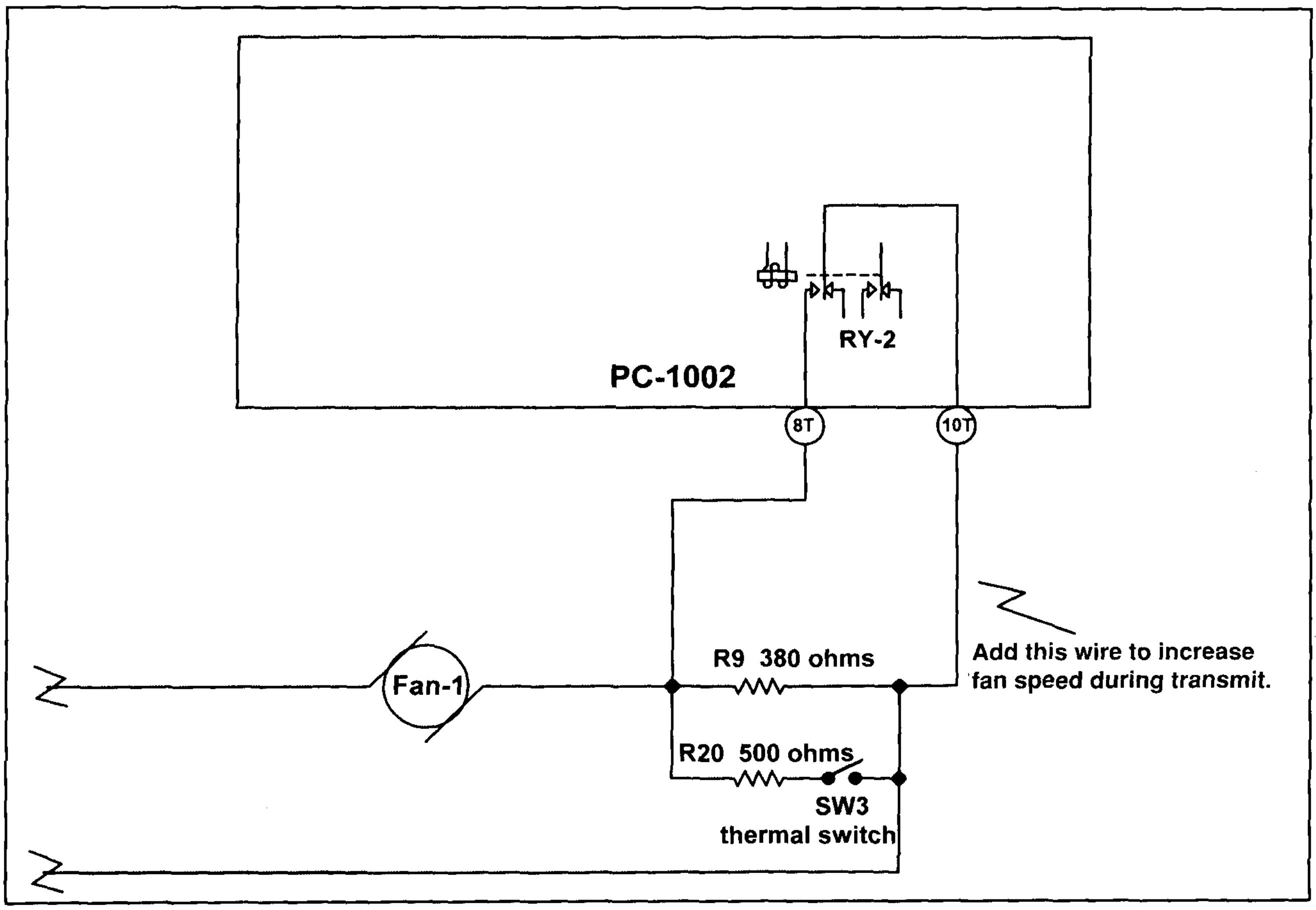

Figure 3. Modified MLA-2500B schematic, continued.

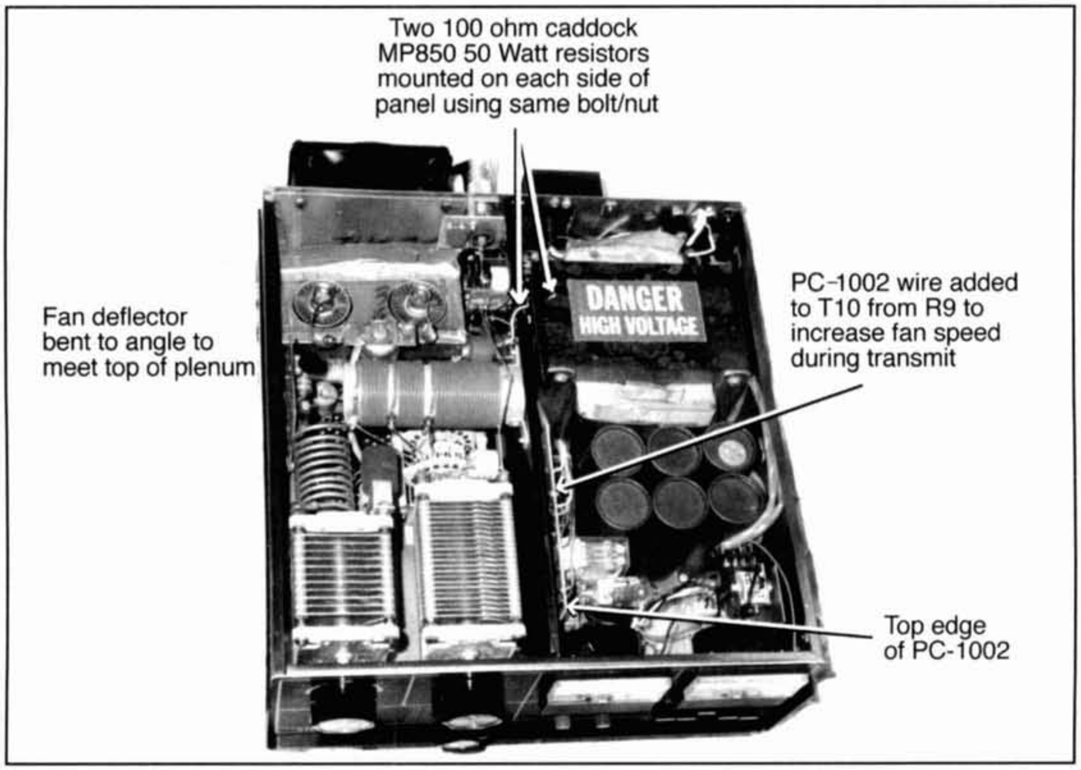

Photo B. Top view without cover.

choke in the 17-meter band. I've found the grid dip oscillator (GDO) to be a reliable (but imperfect) instrument for this check. Some interpretation of the data is necessary. A high "Q" resonance (deep dip of grid current) is almost always a sign of trouble.

The resonance of the plate RF choke in my MLA-2500B was about 17 MHz. I prefer about a 2-MHz separation between the resonance and the band edge. To move the resonance down in frequency to the 16-MHz point, I made a capacitor by adding three turns of #20 Teflon™-insulated wire on top of the RF choke winding at/near the top of the RF choke. These three turns are spread out over about an inch, and the two ends are ***not*** shorted. The ends of these turns are twisted together to hold them in place. Several attempts at positioning the over-winding were required to make certain that no high Q (deep) resonances occurred in ***any*** HF amateur band. Fasten the turns in place with low-loss adhesive. While working on the plate choke, make certain the primary power is completely disconnected.

The minimum amount of air required for the 3CX400A7 at 400 watts anode dissipation is 8.6 CFM. The MLA-2500 (not the "B" version) has two fan speeds, slow and normal. The slow speed is used during standby. The keying line increases the speed to normal during transmit. The temperature sensor, when activated, also increases the speed to normal. The MLA-2500B normally operates at the slower speed and only the temperature sensor activating will increase the fan speed.

Applying AB6YL's idea of using a plastic bag of known volume,[1] I roughly measured/estimated the MLA-2500B's fan's capacity. Running at its slow speed, it produces about 4 CFM with the back-pressure caused by the plenum. This is adequate air for 200 watts anode dissipation. Again, using the plastic-bag method, I roughly estimated 8 to 9 CFM cooling at normal speed. I therefore decided to activate the fan to the higher speed when transmitting. This is accomplished by paralleling the unused contacts of RLY-2 with the series fan resistors R9 and R20. A single wire from pin 10T (top) on PC 1002 to the 380-ohm resistor is all that's necessary.

The MLA-2500 manual didn't specify a typical grid current rating; the only comment made is that 200 mA should not be exceeded. The Svetlana 3CX400A7 data sheet shows 5 watts maximum grid dissipation. It doesn't have a specified maximum grid current limit. The 3CX400A7 has about 80 percent of cathode area of the 3CX800A7, which is rated at 60 mA grid current (CCS). Thus, 50 mA is a conservative maximum grid current for the 3CX400A7. A higher grid current during a brief tune-up and for voice and CW is reasonable; however, 100 mA average for both tubes during operation should not be exceeded. Therefore, for my conversion, I changed the MLA-2500 grid shunt from 0.0462 ohms (F.S.=1A) to 0.462

PERFORMANCE CHART
TWO 3CX400A7/8874 IN MLA-2500 (B)

Operating Freq MHz	28.5	24.9	21.3	18.15	14.2	10.1*	7.2	3.8	1.9
Bandswitch	28	21	21	14	14	7	7	3.5	1.9
Power Output Watts	820	1020	920	920	1000	940	1000	1050	800
Drive Power Watts	115	115	120	120	110	60	80	85	60
Plate Voltage	2000	1950	2000	2000	1950	2000	1950	2000	2000
Plate Current Amps (single tone)	.800	.930	.880	.920	.930	.740	.950	.900	.860
Grid Current mA	70	95	75	80	80	95	90	90	100
Input VSWR (Auto Antenno-ON)	1.2	1.2	1.5	1.3	1.3	1.1	1.2	1.1	1.1
Tune Control	21	27	18	25	12	23	6	3.8	6
Load Control	7.	7.5	6.2	7	3	10	3	6.4	10

* Maximum RF Power Output in U. S. is 200 watts at present. Operation from DX locations should be governed by local regulations.

Conditions: Primary Power = 119V AC
Exciter = ICOM 765; Auto Antenna Tuner On
RF Load = 50 ohm Cantenna
RF Power Meter = Bird 43 1 KW element

Table 1. Performance chart.

(F.S.=100mA). This modification serves several purposes; the two-tube grid current limit is now full scale, causing the operator to exercise more care when tuning, and it eliminates the 10 percent of full-scale maximum value of the original meter shunt—which was difficult to monitor and, therefore, put the tubes at risk.

A series resistor in the HV B+ lead to limit current due to an internal arc in the tube is another worthwhile addition. A 20- to 50-ohm, 50-watt resistor is adequate for this protection. In the event of an arc, the power supply energy is dissipated in the resistor instead of the tube. This is an important addition to the MLA-2500. I mounted one end of this resistor with one lead on the top electrolytic capacitor, after unsoldering the heavily insulated B+ lead to the RF choke (see **Photo A**, bottom view). I made a well-insulated solder tie point from a scrap of G-10 material and an aluminum angle bracket. I mounted the other end of the 25-ohm resistor between the tie point and the connection to the RF choke.

Most modern power amplifiers use a tuned circuit of some sort between the exciter and the cathodes in a grounded-grid configuration. The tuned circuit provides several benefits, in addition to lower RF-drive requirement and improved IMD. First, it provides a good termination/load for the exciter and provides isolation from the variation of input impedance to the cathodes due to the plate tuning and loading. The lack of isolation becomes obvious with exciters that have an automatic antenna tuner. With the auto tuner out of the circuit, the VSWR may be high, causing the power foldback protection circuit of the exciter to protect and reduce RF output. With the auto tuner in the circuit, it's difficult to optimize the RF output of the amplifier. Adding a tuned circuit for each of the nine amateur bands, while nice, isn't practical with the MLA-2500. I made an improvement in input VSWR by adding an RF choke and a bypass capacitor near the tube socket cathodes (see **Photo A**). This reduces the adverse effect of the long cathode connection running to PC 1004, the ALC board.

A second improvement is to replace the existing 75-ohm load resistor, which is supposed to be non-inductive (but isn't) with a true non-inductive resistor. I used two Caddock 100-ohm, 50-watt resistors, type MP 850, in series to realize 200 ohms, 100 watts. These resistors are mounted on each side of the solid wall near the antenna changeover relay (**Photo B**, top view). Use plenty of heat-conducting compound to assure heat removal from these compact devices. The input VSWR performance would be improved by placing the resistors closer to the cathodes, except there's no solid surface at this point to mount them, for conducting the heat away. These two additions do improve the input VSWR and provide sufficient isolation so the antenna tuner in my ICOM 765 matches the input of the MLA-2500 on all bands.

We're now ready for *initial testing*.

Testing

Before applying power, carefully inspect around the plenum. Close the safety micro switch at the right rear of the chassis. Make your initial turn-on without tubes. If all is satisfactory—after full safety precautions are taken, including removing the primary power plug, and after the plate voltage meter reads zero, and shorting the B+ supply to ground—place the 3CX400A7 tubes in their sockets. Use care, because the top and bottom are open and the safety switch is disabled. Set the amplifier multi-meter switch to plate voltage and the tune/CW switch to tune. Reapply primary power. B+ should be about 1700 on the MLA-2500B and should read about 2200 volts DC on the MLA-2500.

After an initial warm-up period of five to 10 minutes (I prefer a longer than normal initial warm-up period for a new out-of-the-box tube), close the keying line without RF drive. The grid current should read zero, and the zero signal plate current should be about 10 mA. The MLA-2500 fan should increase in speed. This is now true for both models, provided the MLA-2500B was modified as above.

Connect a wattmeter and a dummy load to the RF output connector. Set the band switch to 10 meters. With the keying line closed, rotate the plate-tuning control completely (360 degrees), first with the load control at minimum, and then at maximum capacity. Make sure there are no settings of the tune-and-load controls where there is a change in plate or grid current. Perform the same test for each of the band-switch positions.

With the amplifier band switch again set to 10 meters, connect the RF exciter set for 10-meter CW and 10-watts RF output. Now key up the combination and adjust the amplifier plate-tuning and load controls for maximum RF output. Follow the MLA-2500 instruction manual.

With the amplifier and exciter unkeyed, increase the B+ on the "B" version to 2400 volts DC by placing the tune/CW-SSB switch in the CW-SSB position. Depress the multi-meter switch for the grid-current position. Again, key the amplifier-exciter combination. Recheck the plate tuning. There should be little change.

Increase the RF drive, alternately adjusting the tuning and load control to increase power output and keep the grid current below 100 mA. Watch the plate current occasionally as a check; however, the grid current is the most important parameter to observe. On the higher HF bands, you may operate the exciter at full 100 watts RF output. On the lower bands, exciter RF output must be reduced to maintain the grid current below 100 mA and the plate current below 1 amp. I operate the tune-and-load controls—where the RF power output is at peak, the grid current is no more than 100 mA, and plate current is less than 1 amp.

You may use the MLA-2500 manual as a guide for tune-up, but follow this article for adjustment of grid current. This parameter must **NEVER** exceed 100 mA. You'll find a tuning chart is very handy.

Table 1 shows the performance of my MLA-2500. Operation from 220 volts AC primary power will undoubtedly yield higher RF output, providing better regulation and stable B+ voltage. A schematic of the modified MLA-2500B appears as **Figures 2** and **3**.

Acknowledgments

My appreciation to several fellow amateurs; in particular, George Badger, W6TC, who encouraged this article and contributed to some of the project ideas. A future article, using a pair of 3CX800A7s (improved cooling, but no increase in RF output) is under consideration.

REFERENCES

1. George T. Daughters, AB6YL, "New Life for Dentron MLA-2500s," *QST*, May 1996, page 45.
2. B-N. Alper, W4OIW/6, "4CX400A Russian Tubes for the MLA-2500 Amplifier," *Communications Quarterly*, Summer 1996, page 29.
3. ARRL Laboratory Test Report, HF Comparative Table, Test Date: March 1978, Copyright 1977.

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SCIENCE IN THE NEWS

Molecular magnets and the photonic lattice

Innovation in Molecular Magnets Attracts New Class of Microelectronics

A new class of magnetic materials made of clusters of inorganic molecules has been found to contain such unusual properties they could open up an entirely new wave of applications in the field of microelectronics. In research conducted at the Weizmann Institute in Rehovot, Israel, and described in the September 24, 1998 issue of *Nature*, scientists used molecules of nickel-dichloride to make the new magnets, which are much smaller than the metal-organic compounds currently used to create most molecular magnets.

Nickel-dichloride magnetic materials, unlike most other molecular magnets, are semiconductors. This means they can be used to create switches operated by an electrical current. They can also be operated optically because they selectively absorb light of a certain wavelength. This combination of properties makes the new molecular magnets extremely versatile.

Molecular magnets represent the ultimate in miniaturization for the microelectronics industry, which is perennially seeking new ways to create ever smaller devices. In particular, these magnets permit more computer memory to be packed into limited spaces.

Hard-disk computer memory is usually constructed of millions of magnetic switches, in which a change between the "on" and "off" (one and zero) positions is performed by altering the switch's magnetic polarity. The ideal magnetic switch must be operated by a relatively weak magnetic force, so its polarity can be altered with relative ease, yet at the same time it must be sufficiently stable to preserve its polarity for extended periods.

Numerous types of molecular switches are currently under development, but researchers continually run into problems when they are placed next to one another. Magnetic forces reach over a relatively long range, so when the miniature magnets are packaged tightly, interference results. Thus, when the polarity of one of these magnets is switched, the orientation of neighboring magnets changes as well. Such interference makes it impossible to store information reliably over a long period of time.

There is a solution

The new nickel-dichloride structures created at Weizmann promise a solution to this problem. They are expected to be much less influenced by the magnetic fields of their neighbors and to be relatively indifferent to other environmental influences, such as temperature. Their seamless structure also suggests they should not be sensitive to hostile chemical effects of the environment, such as oxidation.

In addition, because these structures contain no impurities and because their spatial structure is well defined, their magnetic properties can be deployed in a precise manner according to predetermined needs.

Significantly, the new molecular magnets are also distinguished by their shapes. Some have a cage-like structure and resemble fullerenes (the soccer-ball like molecules named after the architect R. Buckminster Fuller), while others are shaped like tubes, called nanotubes, which are desirable as the Age of Nanotechnology approaches.

A nanotube is a honeycomb lattice rolled into a cylinder. A nanotube diameter of about 1 nm is much smaller than the most advanced semiconductor devices, thus nanotubes open the way for new solid-state physics. One of the most significant physical properties of nanotubes is their electronic structure which depends only on their geometry. Specifically, the electronic structure of a single-wall nanotube is either metallic or semiconducting, depending on its diameter and chirality (asymmetry), and does not requiring any doping (addition of impurities to achieve desired properties). Further, the energy gap of semiconducting nanotubes can be varied continuously from 1~eV to 0~eV, by varying the nanotube diameter. Thus, the smallest possible semiconductor devices imaginable are likely to be based on carbon nanotubes.

Bottom-up construction

The method by which the nickel-dichloride clusters at Weizmann are constructed is also unusual. Rather than producing chunks of magnetic material, the scientists built the magnetic molecules from individual atoms. The molecules then self-assemble into a spherical layer one molecule thick. This bottom-up method of creating a magnetic material gives scientists precise control over the size and structure of the magnet's molecules and the number of layers. This, in turn, allows the material to be tailored to specific needs.

"The bottom-up approach gives us enormous flexibility," said research leader Reshef Tenne of the Weizmann Institute of Science's Materials and Interfaces Department. "It's like constructing a building from individual bricks as opposed to moving around the walls within a prefabricated house."

The study of fullerenes and nanotubes in inorganic materials was pioneered by Tenne and his Weizmann Institute colleagues in the early 1990s, creating a new avenue of research in materials science. The production of the nickel-dichloride molecular magnets further expands this field by introducing to it a totally new family of compounds.

Apart from switches for computer memory, Tenne's molecular nanotube magnets may have a variety of other industrial uses. Thanks to their small size, they may be used for extremely fine "etching" of information onto magnetic disks—a process known as lithography—and for "reading" this information. Such reading is performed, for example, during the quality control of computer chips, and a nickel-dichloride nanostructure could do this at far greater resolution than any existing device.

Because the magnets are new and are expected to display intriguing magnetic behaviors that have not yet been fully investigated, they may find other, unexpected applications in the future. Tenne's team is now developing methods for synthesizing large quantities of nickel-dichloride magnetic materials in order to study their magnetic properties in greater detail and pave the way for industrial testing.

The Photonic Lattice: Bending the Light Fantastic

Government scientists believe they have scaled a major technical barrier to optical communications by learning how to bend light easily and cheaply with no leakage—no matter how many twists or turns are needed for optical communications or, potentially, for optical computing.

The achievement was reached at Sandia National Laboratory in Albuquerque by interlocking tiny slivers of silicon into a lattice that, under a microscope, appears to have been constructed with miniature Lincoln Logs™.

The lattice, dubbed a photonic crystal (crystals have regularly repeating internal structures) and shown in **Photo A**, now works in the infrared range at approximately 10-micron wavelengths. Researchers Shawn Lin and Jim Fleming, members of the Sandia technical staff, are now preparing a 1.5-micron crystal for the region in which almost all the world's optically transmitted information is passed.

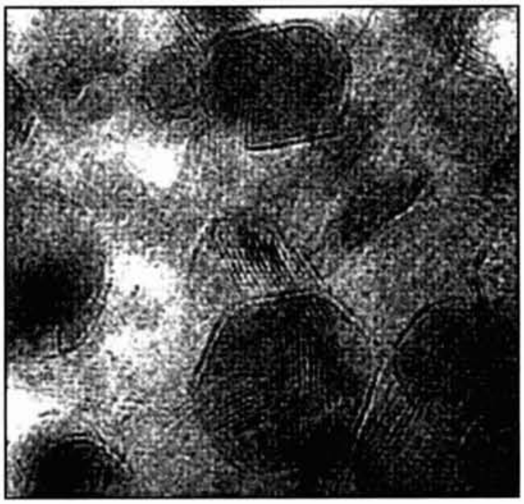

Photo A. The photonic lattice created at Sandia National Laboratories acts like a crystal in guiding light because of its tiny, regularly placed silicon "logs." The logs are 1.2 microns wide. Control of different wavelengths is achieved by changing the lattice dimensions.

The Sandia researchers believe the improvement—which bends far more light in far less space at considerably less cost than current commercial methods—will make possible tinier, cheaper, more effective waveguides to combine or separate optical frequencies at the beginning or end of information transmissions. It will find wide application in data transmission and in more compact and efficient sensors.

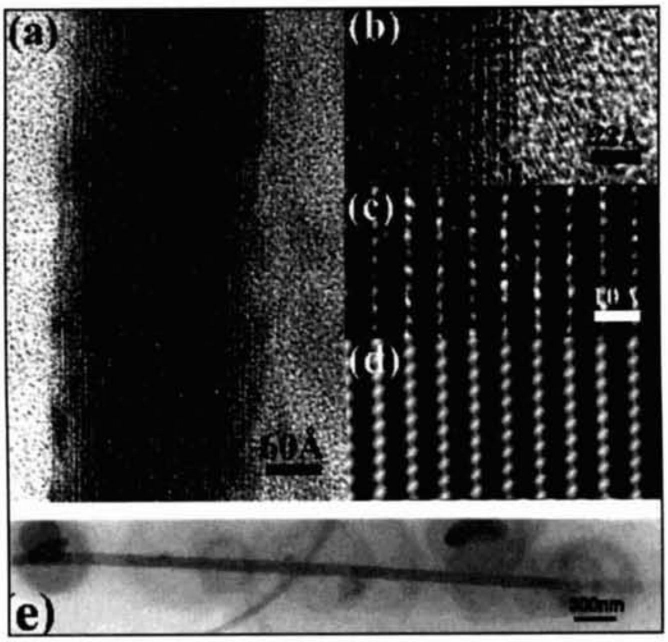

Photo B. The NiC12 nanotube in different magnifications.

The creation of a photonic "band gap"

What is significant about the work is that the researchers have created the equivalent of a photonic "band gap," which forbids certain frequencies of light from exiting the lattice.

"This work represents a breakthrough in that researchers have been attempting to realize a 3-dimensional photonic band-gap structure in the infrared and visible region of the spectrum for over a decade without substantial success," said Del Owyoung, Unit Director, Advanced Semiconductor Technologies at Sandia. ("Band gap" is a term usually applied to electrons, not photons, and signifies a range of energies in which electrons are absent because their presence would contradict the laws of quantum mechanics.) "Our recent work shows a method for fabricating these structures in the infrared spectral region (12-micron) with promise for extension to the near-infrared (1.5-micron) which is the wavelength of interest for optical communications. Previous embodiments of the concept have only been realizable down to millimeter wavelengths."

The work has amazed other researchers working on the same problem. According to Rama Biswa, a researcher at Ames Lab, "We had built the same structure ourselves, but more than 100 times larger, in the microwave frequency range. I think it is quite remarkable that Shawn Lin's group could do it at these wavelengths in the infrared and at this size." Pierre Villeneuve, a research scientist at the Massachusetts Institute of Technology, who has theorized about uses for photonic crystals for much of the 1990s, praised the Sandia group for "showing that what we're doing is valid. People would say, 'It's wonderful that you're coming out with all these great devices that make use of photonic crystals, but your building block—the photonic crystal—doesn't even exist.' That is no longer the case," he said.

The Sandia researchers say the photonic lattice structure, in the regularity and spacing of its "log" components, traps light of a particular frequency within the cavity of the structure, preventing it from escaping, or leaking. Instead, light is forced to follow along the twists or turns designed into the log structure. By designing the distance between logs carefully, a chosen wavelength is reflected instead of passing out of the space, as longer or shorter wavelengths can. With introduction of an impurity like air or much thicker polysilicon "logs" to provide routes for preselected wavelengths introduced into the crystal, light travels along the impurity as it twists or bends. No matter how sharp the turns, light of frequency roughly in the middle of the band gap cannot escape.

The nearly leak-proof lattices form a cage that traps and guides approximately 95 percent of the light sent within them, over three times more efficiently than conventional waveguides, and the Sandia lattice takes only one-tenth to one-fifth the space to bend the light. (The Sandia photonic lattice's turning radius is currently in the 1-wavelength range, rather than the traditional waveguide bend of more than 10 wavelengths.)

Standard integrated-circuit manufacturing technology used to fabricate micromachines—machines nearly too small to see—at Sandia's Microelectronics Development Laboratory, can create tens of thousands of waveguides from a single, 6-inch silicon wafer. This is 10 to 100 times denser than can be fabricated using more expensive gallium arsenide with current commercial technology.

A revolution in the design of guided-wave optical systems

"In the IR [infrared] there has been a long-standing interest in these types of structures for tailoring the infrared emission properties of objects," Owyoung said. "At the shorter wavelengths, they could be a key building block in realizing much more efficient semiconductor lasers as well as the material for building ultra-compact optical circuits consisting of very low-loss waveguides and other optical elements. This would represent a revolution in the way we design and build guided-wave optical systems."

What's next? "A future goal," said Owyoung, "is to study and establish the emissivity characteristics of these materials and learn how to further tailor them in the infrared. We also want to design and study optical structures such as cavities, waveguides, beamsplitters, etc., to learn how to utilize these materials to innovate new devices and systems."

Because little light is lost in the lattice's three-dimensional mirroring that sends light back at itself, a new type of microlaser requiring little start-up energy is now theoretically possible. (Most conventional lasers require large jolts of energy to begin operating because so much light is lost in the lasing start-up process.)

The photonic lattice achievement also brings closer the advent of optical computing—using computers that transmit information using photons rather than electrons. Present-day computers use electrons to pass information, but, as more circuits are included on new chips, they become more difficult to cool. Photons, the stuff of light, are faster and cooler than electrons. The problem was that previously no one has been able to bend useful frequencies of light around tight corners (as navigated by electrons through a million turns on a computer chip the size of a postage stamp) without experiencing substantial loss of data.

Till now, the more tightly light was turned in present optical communication techniques, the more it has leaked. One principle of optical communications is that differing frequencies of light are bent by different amounts. Waveguides first combine the frequencies of a number of information streams—for example, voice and data—by bending the streams into the combined "white" light passing through an optical cable. At the other end, similar waveguides separate (bend) the light back into its original component frequencies.

The Sandia photonic crystal lattice could be the solution to this light loss, making optical computing a reality! ■

PRODUCT INFORMATION

Svetlana 3CX2500A3 and 3CX2500F3

Svetlana Electron Devices, Inc. has introduced the 3CX2500A3 and 3CX2500F3 high-performance tubes. These ceramic/metal power triode tubes are designed for use in amplifier, oscillator, or modulator service. A modern mesh filament replaces the hairpin construction in both the 3CX2500A3 and 3CX2500F3. The

3CX2500F3 has flying leads for the filament and grid connections that eliminate the need for a socket in low-frequency RF and AF modular applications.

For more information, visit the Svetlana web site at: <www.svetlana.com>.

Pentek Lowpass Anti-Aliasing Filter

Pentek Inc.'s anti-aliasing, lowpass filter board is designed to eliminate out-of-band frequency components and reduce noise prior to A/D conversion. When connected with Pentek's Model 6606 or 6109 A/D boards, it provides an eight-channel A/D converter system for VMEbus data acquisition. The unit is offered with cutoff frequencies at either 800 kHz or 8 MHz. Other custom frequencies are available.

Model 6106 delivers filtered output signals to a 37-pin front-panel "D" connector that is pin-compatible with the A/D input connectors of Pentek Models 6106 and 6109. Model 6106 also offers 14-bit resolution, 2-MHz maximum sampling rate, and optional differential inputs. Model 6109 offers 12-bit resolution, 20-MHz maximum sampling rate, along with single-ended inputs.

For details on Pentek's anti-aliasing filter, contact Pentek, Inc., One Park Way, Upper Saddle River, New Jersey 07458-2311.

HP Capacitance Meter

The HP 4268A 120-Hz/1-kHz capacitance meter from Hewlett Packard is a measurement solution for high-value (10 to 100 μF) multilayer ceramic capacitors (MLCC). The meter's features and capabilities include fast auto-level control function; standards-compliant test signal level; and high-speed measurement of 25 ms/pt.

For more information, contact Hewlett-Packard Company, Test and Measurement Organization, 5301 Stevens Creek Blvd., MS 54LAK, Santa Clara, California 95052.

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QUARTERLY REVIEW

The Tektronix TDS-210 digital real-time scope

The dual-trace, high-frequency scope is the staple of every amateur RF experimenter and designer's toolbox. In the last few years, analog scopes that cost thousands when new have been made available to us for a few hundred dollars and are an invaluable tool for radio circuit evaluation, development, optimization, and signal tracing.

On the professional front, digital scopes are now the leading edge for state-of-the-art measurements. As analog scope writing speeds are limited by CRT voltages, the limit of a visible analog display is about 1 GHz, using special CRTs with microchannel plates. Today, the most advanced new generation scopes are digital and have real-time bandwidths of a gigahertz or more, sampling at four gigasamples per second.

Taking the plunge

In time, these units will hit the flea markets, but meanwhile the first low-cost, real-time digital scopes have made their debut. Several months ago, seeing the Tektronix TDS-210 offered at under $950, I gave in to temptation, reached for my plastic, and ordered one.

Although the display screen is the same size, overall the TDS-210 is smaller than a comparable analog scope, and, at less than five pounds, is a lot lighter. You can get an idea of the size from the BNCs in **Photo A**.

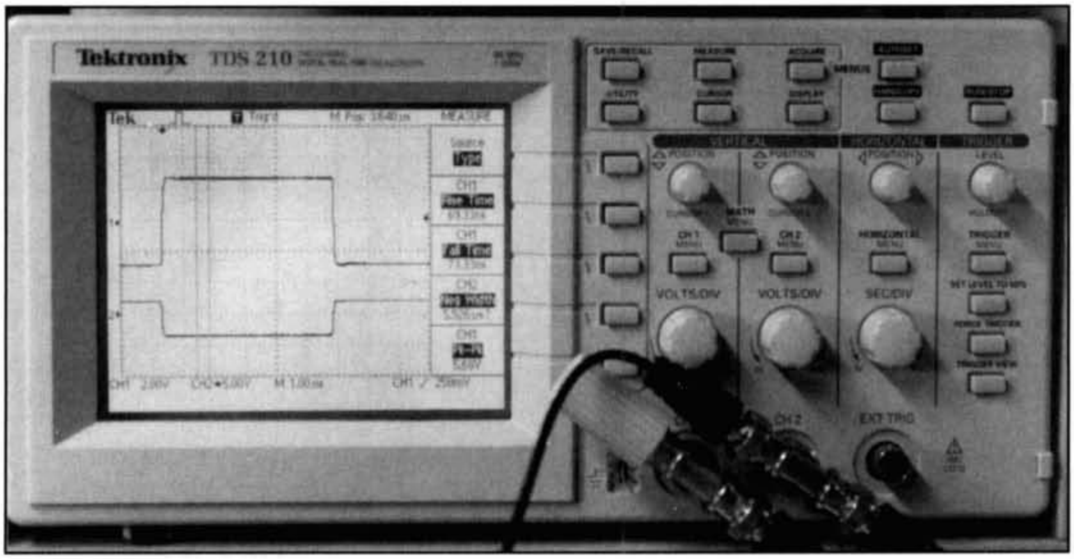

Photo A. Front view of the TDS-210.

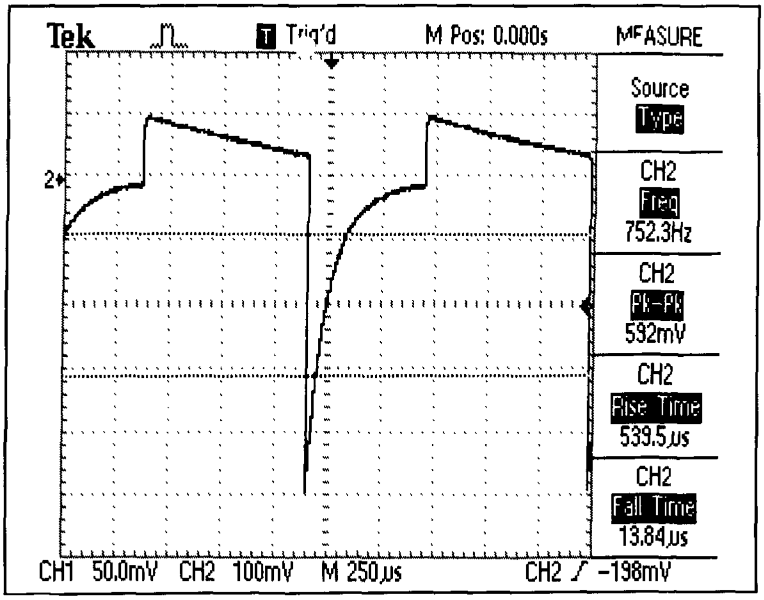

Figure 1. Quick evaluation of complex waveform.

With many features and capabilities not available in analog, the TDS hasn't disappointed me. But, don't pitch your trusty 30 to 100-MHz analog scope yet. It still does several things better than its digital counterpart. Although I haven't had occasion yet to use every capability of this new instrument, it has both advantages and disadvantages.

The fine points

The TDS is a Digital Real Time (DRT) scope, with a technology different from older digital scopes. Earlier digital waveform acquisition technologies depended on successive sampling of repetitive waveforms known as Equivalent Time Sampling. The TDS DRT scope constantly processes the input signal whether a trigger event occurs or not, at a rate up to one gigasample per second. A steady stream of data flows through the scope; the trigger merely allows the data to be presented. This enables display of fast, single-shot events or infrequently changing signals.

The fact that the input signal is analog-to-digital, converted and then processed digitally, makes available many interesting capabilities not feasible with an analog scope. A notable one, Fast Fourier Transform, is described later.

The display on the TDS is a backlit liquid crystal with conventional graticule markings; i.e., 10 centimeters wide by 8 centimeters high, providing a visual interface familiar to users of analog scopes (see **Figure 1**). As opposed to a CRT, where the image is written by a moving electron beam, the LCD's display image consists of dots or vectors (lines connecting successive samples) generated and refreshed by a microprocessor that provides a constant-intensity display. On an analog CRT, where the image brightness is a function of the writing speed, rapid transitions such as fast pulse rise-times, parasitic oscillations, or high frequency ringing, are faint compared to other, slower changing features, such as the top of the pulse.

With the TDS, numerical readouts and menus in the display offer status indicators, scale factors, and parameter readouts. The normal display includes symbols and numbers indicating the trigger status, level and timing, the DC baselines of the signals displayed, and other measurement information in addition to your selection of directly read-out parameters.

The TDS-210 has a bandwidth of 60 MHz. (The TDS-220 bandwidth is 100 MHz.) Both units have a maximum sample rate of 1 gigasample per second and a record length of 2500 samples per channel. Sweep speeds range from five seconds/div to 5 nanoseconds/div. This is a very fast sweep and, because it's generated by a digital clock, it is also very accurate—specified at ±0.01 percent. The time base in an analog scope is often quite non-linear at the leading edge as it is hard to design a sweep circuit that begins instantaneously. Because it's effectively continuous, the TDS sweep is precisely linear and provides good pretrigger viewing, with the trigger event initially displayed in the center of the display.

A small segment of a signal can be examined in detail with the WINDOW feature. The Window Zone is easily set on the main sweep and then a portion of the signal can be examined in much greater detail. This offers a capability similar to delayed sweep and sweep magnification on an analog scope, but is easier to use and provides a unique capability: the display of pre-trigger events.

The panel layout is similar to a conventional dual-trace scope, with separate position and sensitivity controls for channels one and two on the left and horizontal controls on the right. There's a conventional calibration and grounding terminal on the front panel for probe adjustment (see **Photo B**). A significant difference from traditional analog scopes is that there are far fewer controls and buttons; details are revealed to the user via menus displayed on the screen.

The vertical amplifier ranges are similar to a good analog scope with a sensitivity from 2ms/div to 5V/div. Offset ranges are generous, ±2 volts on the 2 through 200-mV ranges and 50 volts beyond. Accuracy is specified at ±3 percent with a vertical scale resolution of 1 part in 250, or about 1/30 of a centimeter—much closer than I can read an analog scope. As with vertical amplifiers in analog scopes, the bandwidth can be restricted to reduce noise. Even

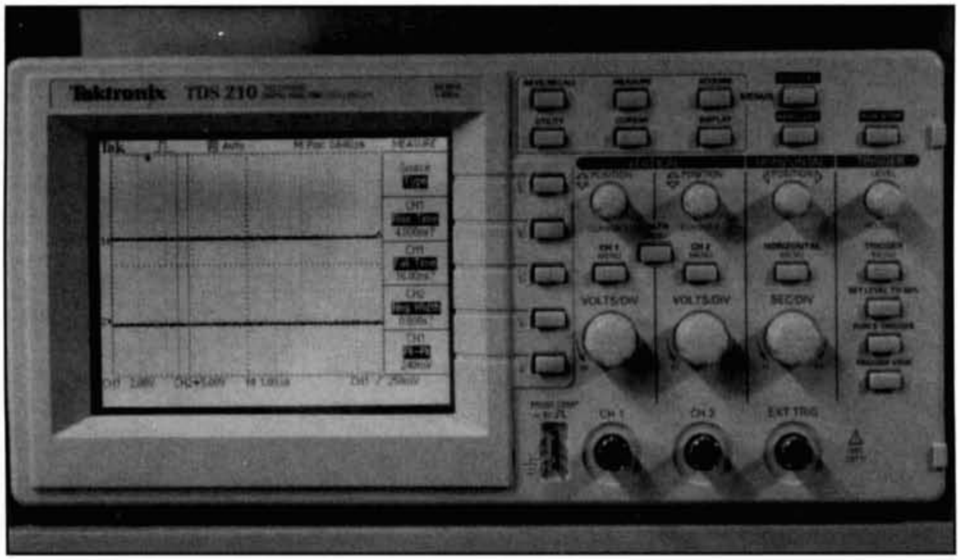

Photo B. TDS-210 controls.

better, you can use the display-averaging feature to clean the noise out of repetitive low-level signals.

The TDS automatically calibrates itself on boot up. After an hour or so of use, you can initiate a SELF CAL. The specified accuracy is over a 0 to 50-degrees C temperature range and my impression is that the accuracy on my bench is far better than spec. On the analog scopes I've used, the DC base line moves as you go through gain settings. On the digital, of course, it doesn't move at all.

Tek furnishes 100-MHz 10X probes. My TDS is mounted on a shelf directly in front of me, and I found their 2-meter-long leads got tangled in my work, so I substituted 1-meter long Probemasters. The long Tek probes are just right for my dolly-mounted analog scope.

Measurement parameters

Direct readout is provided for the following signal measurement parameters:

- Peak-to-Peak Volts
- RMS Volts
- Period
- Frequency

In addition to the above, with the TDS2MM option installed, the following additional parameters can be displayed:

- Rise Time
- Fall Time
- Positive Pulse Width
- Negative Pulse Width

Four parameter measurements can be selected and displayed simultaneously for channels 1 and 2.

For measuring other waveform features like ringing, overshoot, ripple, etc., two calibrated cursors are provided, usable for either level or time measurement. With either direct parameter readout, or the use of the cursors, there's no need to make eyeball estimates of trace positions and multiply them by scale factors. This common source of error is eliminated.

Figure 2 shows the use of the cursors in the time mode to measure the delay of a 7-MHz signal through a special coaxial line.

A very convenient feature, standard in other Tek digital scopes, is AUTOSET. Instead of having to fiddle with beam intensity, "Brightline" or "Beam Finder," trigger level, baseline centering, amplitude, etc., on an unknown sig-

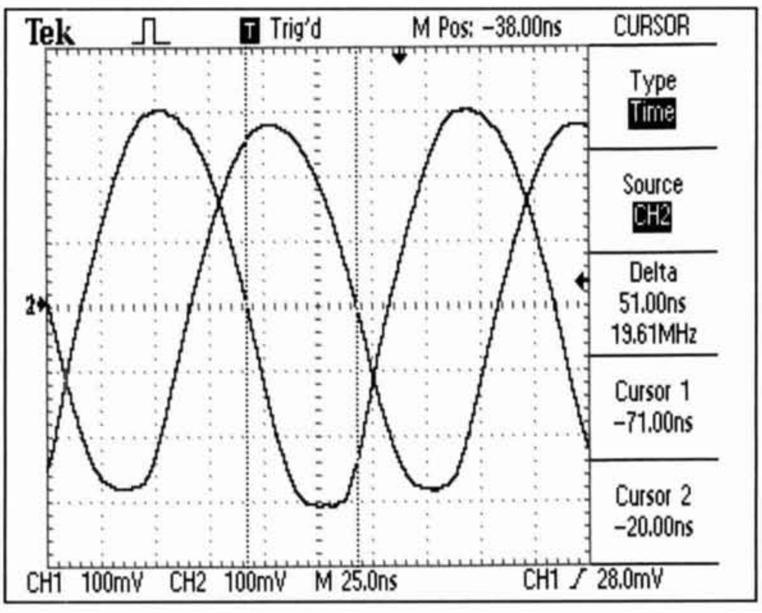

Figure 2. Delay measurement.

Photo C. Note that *all* functions and settings are included on the front panel of a traditional analog scope.

nal, you just press this button, wait for a second, and the scope's microprocessor stabilizes and displays several cycles of the signal.

With careful adjustment of an analog scope, you can just see the bottom of a fast leading edge. Distortion of the first 5 or 10 percent of the leading edge is also common due to sweep start up. In the TDS, the initial display centers at the trigger point. The horizontal display can then be adjusted to view pretrigger events as far in advance as the sweep is set following the trigger point.

With an analog scope, slow waveforms are a problem to synch and to display. Signals under 20ms/div or so appear to the eye as a wavy trace or just a moving dot. A Polaroid scope camera plus a lot of skill is necessary to study slow signals. The TDS has a minimum speed of 5 sec/cm, which means 50 seconds of data can be displayed. This is a virtual chart recorder. The trace brightness at slow speeds is the same as the fastest 5ns/div sweep. Using the RUN/STOP button in the single-shot mode, the waveform image can remain as long as you choose, or be printed out.

Evaluating relay closure

Fast, but infrequent, signals require pushing up the intensity on an analog scope and watching carefully—the "blink and miss" technique. An example of this is evaluation of relay closure. With the TDS, reed relay bounce lasting only microseconds can be observed in detail at a rep rate of 1 second and, once you've adjusted your setup, on a single-shot basis.

I'm building a general-purpose time-sequenced keyer for several QRP rigs, including a classic tube model. For this, I needed to evaluate several relays, a reed for keying a buffer tube in the B+ lead and a heavier relay for antenna T/R. I needed to determine the closing and settling time for my T/R relay, so I wouldn't be switching it hot. It took me only a few minutes to select the fastest one in my junkbox.

The printout for the reed keying relay is shown in **Figure 3**. With a function generator cycling the relay at a rate of one or two cps, I adjusted the display and observed that the bounce pattern was independent of coil voltage and frequency. As you can see, it completes its closing cycle in less than 360 microseconds, which is fast enough to key at 25 wpm (faster than most QRP ops). This information will also minimize cut and try on the RC network for the keyed signal shaping.

Varying coil voltages showed a significant effect on the time. Using the same setup as the slower relay, and the WINDOW feature of the horizontal system, I could examine the bounce in great detail and learn that was very stable.

Printer attachment

You can attach a printer to the TDS with any of the options. I placed a small Canon Inkjet next to the TDS. A captured (or live) waveform can be printed by simply pushing the HARDCOPY button. It takes a few seconds to for the printer to complete the copy, but meanwhile you can continue using or adjusting the scope.

Screen parameters

A major advantage of digital is that you set the screen parameters for your ambient light and forget about CRT intensity control. I haven't changed my screen in three months. The flip side of this is that black and white digital doesn't provide an important feature of analog scopes called "Intensity Grading." This is the characteristic where the phosphors respond to the speed and repetitiveness of the CRT beam. I first discovered this when I tried looking at a modulated carrier, a "Dynamic Signal," as opposed to a steady carrier or pulse train. Conventional CRTs display dynamic signals by varying the trace brightness (or "fuzziness"). Digital doesn't do this very well.

A good example of a dynamic signal is NTSC Video on a moving scene. On the TDS or other conventional digital scopes, it's very difficult to get a satisfactory display without trying various combinations of the acquisition and display controls and display persistence. (High-end scopes deal with intensity factor using color. Red indicates the brightest ranging and blue the least, but it'll be a long time before these will be on many ham benches.)

With care and experimentation, intensity-graded signals may be observed by adjusting two controls not found on analog scopes: ACQUISITION and DISPLAY.

There are three Acquisition modes: Sample, Average, and Peak. Sample is the basic mode, with instantaneous response to changes. Peak is used for catching infrequent and high frequency glitches. Averaging is a great way to clean up noisy signals. The number of averages from 4 to 128 can be selected. A small amount of averaging usually makes a big difference. Use of 128 successive measurements on slow (i.e., audio range) signals makes changes in the signal sluggish, so I normally use Sample unless I want to get a clean screen shot.

There are three Display modes: Vector, Dot, and Dot Persistence. It's easy to toggle back and forth between them to see which mode shows the signal features you want. The fastest risetime is easily observed (and printed) in Vector mode. After setting up in Sampling and Vectors, I normally use Dots with No Persistence and Averaging for making accurate measurements. Persistence can be set from 1 second to infinite providing some interesting displays.

Fast Fourier Transform

The TDS2MM option provides the scope with a very interesting feature: Fast Fourier Transform (FFT). The FFT function displays the harmonic structure of a signal against a horizontal frequency reference. I regard this as the "poor man's spectrum analyzer" and have used it on audio and RF signals. The useful dynamic range is about 60 dB, or about 0.1 percent distortion. I first tried this when evaluating a sound meter calibrator picked up at a hamfest. Not only was the GenRad calibrator in spec, so was my equally inexpensive sound level meter!

Because of the scope's bandwidth, the FFT function is useful through the HF range. I've looked at various transmitter signals and have obtained consistent results with various sources and low-pass filters. The TDS measures harmonics well below the FCC approved ranges of 30 and 40 dBc. If they are there, you can observe harmonics out into the low VHF range. Of course, the accuracy degrades as you go beyond the scope's vertical bandwidth.

Figure 4 shows the harmonic structure of a 7-MHz QRP rig. As you can see, the delta reading of the cursor shows that the magnitude of the fourth harmonic, 10 meters, is down less than 35 dB from the fundamental. This indicates that a low-pass filter, or an antenna coupler, is needed following the low-impedance, broad-band collector circuit.

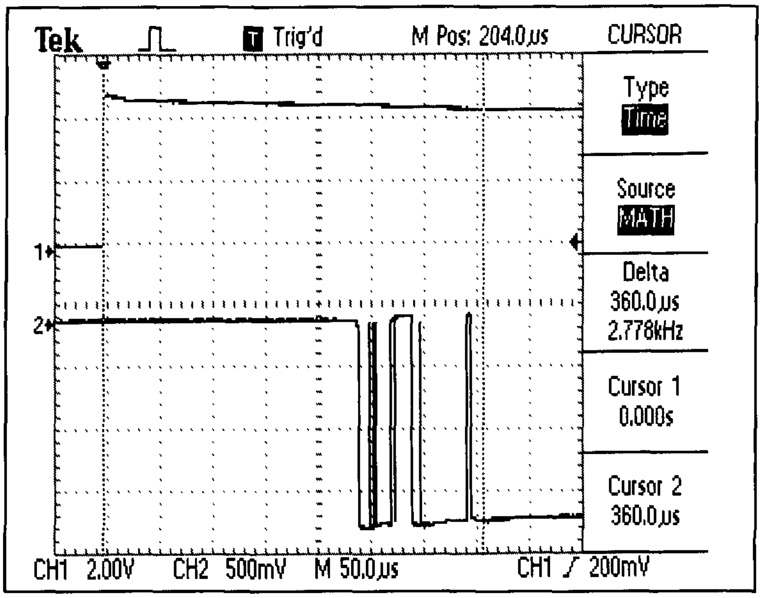

Figure 3. Reed relay bounce.

Figure 5 shows an AM modulated 92.75-kHz carrier from a Surplus USM-25 signal generator. The modulation is a very distorted 1000 Hz. The display shows multiple sidebands going out beyond the display. **Figure 6** shows the use of the cursor in frequency mode to identify the sidebands.

The good news on FFT is that the resolution bandwidth (RBW) can go down into the Hz

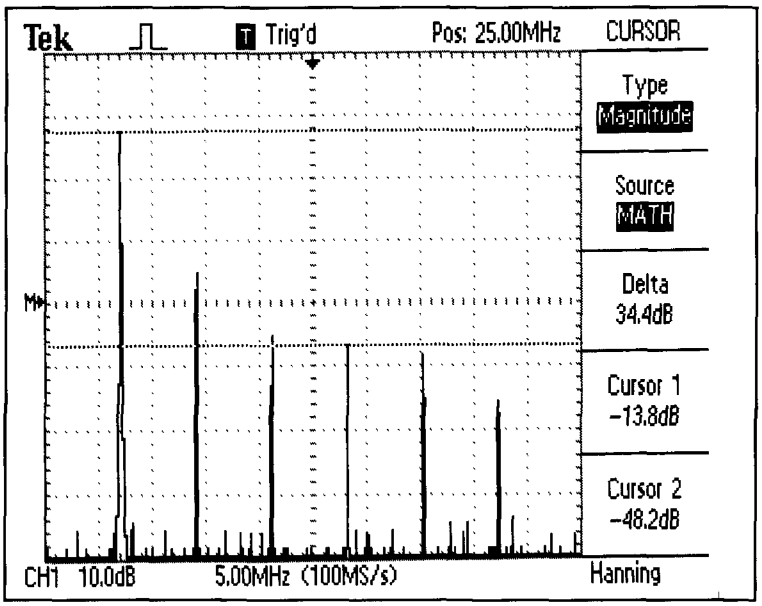

Figure 4. Measurement of 4th harmonic of 40-meter signal at 10 meters.

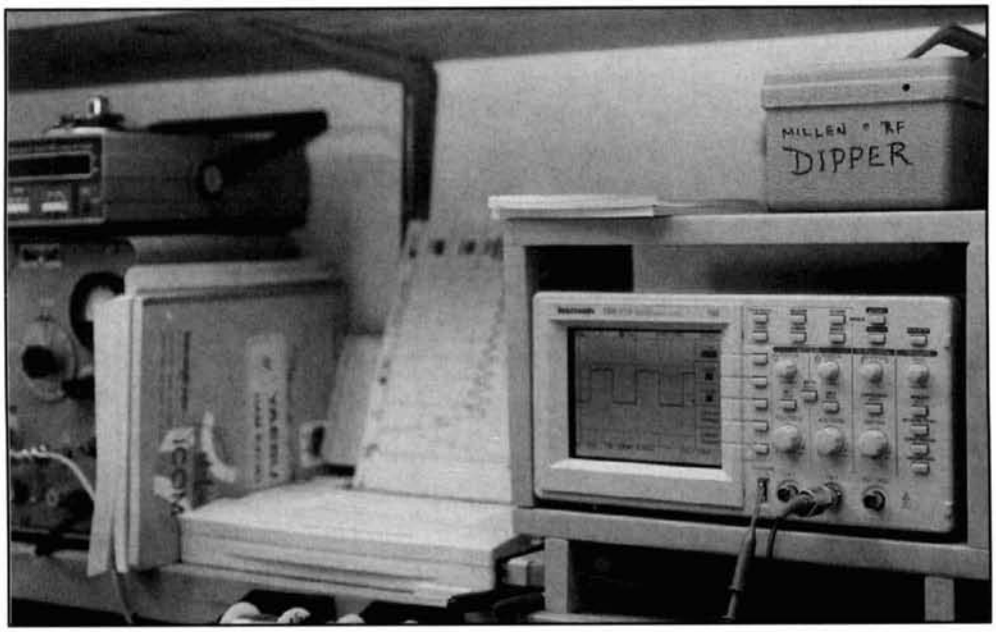

Photo D. The TDS-210 on its wooden stand.

region, suggesting its usefulness for looking at such things as intermodulation distortion (IMD) on an SSB signal. The bad news is that the span is related to the RBW and starts at DC (the "zeroeth" harmonic). As you decrease the RBW, the available span goes down accordingly. Using the window and zoom modes enables you to expand this by a factor of several times; but for useful measurements, say of IMD of an RF signal, the "center frequency," i.e., the first harmonic, needs to be around 100 kHz, or lower.

It's possible that a heterodyne technique could be used to make useful close-in measurements at ham frequencies. Beating the test signal with a well-filtered generator a few tens of kHz offset from the carrier followed by a low-pass filter (LPF) may be a way of making this measurement.

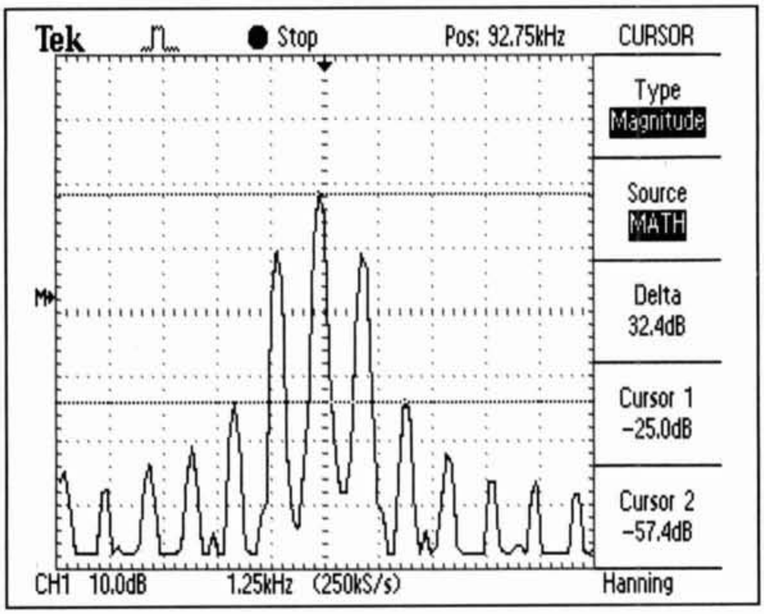

Figure 5. Level measurement of 2000-Hz AM sidebands.

In **Figures 5** and **6**, the harmonics or sidebands are displayed in terms of dB down from the reference carrier. You can't tell from the hardcopy printouts, but the "grass" in the lower portion of the display moves vigorously, while the desired signals, harmonics, and spurs remain steady. To use the FFT, you capture the signal in YT (conventional amplitude versus time) mode. After this, the FFT display is "live" and displays the results of changes in the signal, filter adjustment, etc. The dynamic range of the FFT measurement and display depends on the noise of the signal and other factors but more than 60 dB, allowing measurement of distortion to 0.1 percent or better.

Reasons to keep your analog scope

The TDS does have a few shortcomings, and you'll want to keep your trusty analog scope in service. The most significant advantage of analog is the display of dynamic; i.e., modulated signals. This is just about the earliest ham application of scopes (described in ARRL handbooks in the '30s) and is easy to do with a wideband analog scope. As mentioned above, you can fiddle with its controls, but the TDS won't deliver a clean display of a modulated signal comparable to an analog version. The same thing is even more true of live video.

On the other hand, as a probe for signal tracing or optimizing steady state RF, the TDS is very good. Due to the very high sampling rate and the fast sweep, the relative strength of RF signals can be usefully observed to 200 MHz or so. Of course, as you go above the nominal frequency range of the scope, the vertical sensitivity goes way down, but the LCD display allows "tweaking" as good or better than an RF probe with VOM. My 100-MHz analog scope shows only a white blur, while sine waves are displayed on the TDS.

The display refresh on an analog scope is essentially real time, limited only by sweep retrace. Although the digital scope samples the signals up to one billion times a second, the display refresh is limited by the time required for microprocessing this bit stream. In the TDS scopes, this is done at a maximum of 100 times per second. With normal signals, display refresh takes place well above the flicker response of the eyeball.

The most significant difference between analog and digital scopes lies in the controls. On an analog scope, every function and value is associated with the knobs and buttons and appears on the front panel as in **Photo C**. On a digital scope, the panel is much less cluttered, but under every button or knob there's a menu—sometimes nested menus—which are displayed on the screen.

Some of the settings on an analog scope are more directly controlled. For example, setting a channel to ground involves simply moving a switch, without even looking at it. On the TDS, unless your probe has a ground button, you have to go to the Vertical Menu and select GROUND. The TDS does have a little numbered arrow on the screen showing where the ground level is for each channel.

A ham analogy

A friend, who has taught EE for many years, and has used the TDS as well as other digital scopes, feels that analog is still the best "entry" scope for beginners. Here's a ham analogy: The ICOM-706 is a very powerful and versatile little transceiver but has only a handful of knobs, compared to the ICOM-765, or other comparably featured "analog" radios. I found the 706 very frustrating when I first used it. By the time I mastered it, the instruction book was well thumbed. When I took it on a mini DXpedition, it did a great job. At the moment the 706 is set aside, but it's instruction book, and ICOM's excellent "prompt card," remain nearby for the next time I take it somewhere and get "stuck."

In both scopes and radios, if you don't know what the various functions are supposed to do, presumably learned from their analog equivalents, it will be more difficult to realize their

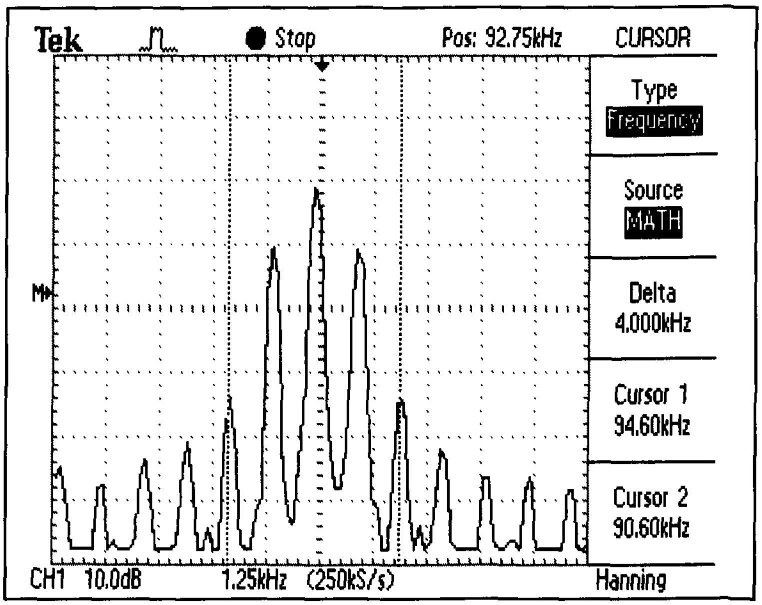

Figure 6. Identification of sidebands using frequency cursor.

full potential. As with the TDS, the 706's software-oriented panel is essential to its small size, but the features are buried in menus.

A significant difference between the radio and the scope, which is about three years newer than the radio, is that Tektronix has done a superior job of logically relating the scope's buttons and menus. After getting started, I've only had to use the book to learn about new features.

Still, the TDS-210 takes center stage

The TDS quickly became my primary instrument, so I decided to locate it in the center of my work area. The light weight, which is certainly a big advantage when you take the scope to the work, is a disadvantage on the bench. You can't push the buttons without putting your hand around the case. Also, it has a rounded case, much like a portable radio or small boom box, which precludes putting anything on top of it.

With little effort, I made a wooden stand with a shelf on top (see **Photo D**). The scope recesses perfectly and the controls can be adjusted with ease. The printer connector is a DB-25 and the AC cord has a standard IEC receptacle (like your PC), so it's very easy to remove the scope from its stand and use it portable, leaving the two cables in place for quick return to bench service.

The display is bright and easy to read, but the viewing angle seems to be somewhat less than CRT scopes. This is why I mounted the TDS directly in front of me, at eye level.

There is one feature standard in analog scopes that's missing on the TDS: continuously variable sweep speeds. After fruitlessly looking for this feature, I finally called Tek, where I learned that it's just not found on DRT scopes. They have an 800 number (1-800-835-9433) which does NOT have the usual…"if all else fails, dial F for frustrated…." I've used this free support line a couple of times and gotten helpful information as well as additional applications literature.

There are features in the TDS that don't seem to have much relevance to RF design. With any of the options, you receive a software disk and a fat manual for interfacing with a PC. I haven't dug into this yet, but it should enable an appropriately skilled user to process information in a variety of ways, record and incorporate data in reports, etc.

Features in the TDS which should be useful for digital troubleshooting include HOLDOFF, for stabilizing a complex signal and viewing it in detail and PEAK triggering, which can capture glitches as narrow as 10 ns through the full range of sweep speeds.

An interesting feature, which I haven't used yet, is the capability to store two reference waveforms. By toggling a button, these waveforms can be recalled and compared with the current display. This feature would be useful for comparing two digital patterns and for production test or troubleshooting.

I won't cover aliasing, or the errors possible with undersampling, here. This subject is well described in the literature. The only applications where I've seen this are in looking at modulated waveforms and some FFT measurements. In both cases aliasing, which causes false information to be displayed, is easily recognized.

All in all, I think the TDS-210 is great and, for work in the HF range, I wouldn't trade it for any analog I've seen. It represents the future and it won't be too many years before instruments of this type will be on every serious builder's bench.

Acknowledgments

Many thanks are due to my friend, Dr. Frank Hankey, KB8VGM, who assisted me in the preparation of this report. I would be interested in receiving comments or questions and can be reached at my *Callbook* address or via E-mail at <71020.711@compuserve.com>. ■

PRODUCT INFORMATION

Analog Devices Publications

Analog Devices, Inc. has published Volume 3, Number 3 of "Communications Direct," a newsletter dedicated to digital communications topics. Articles in this issue discuss ADSL and xDSL.

Copies are available by calling Analog Devices' Literature Center at 1-800-ANALOGD, by clicking on Quick Links/Magazines/DS Patch on Analog Devices' home page or accessing <http://www.analog.com/publications/magazines/comm/index.html>.

Analog Devices has also published issue 32-1 of *Analog Dialogue*. Featured are an article on CCD signal processing from a single integrated circuit; a description of two high-speed, 16-bit, wideband oversampling ADCs; and part 4 of the DSP101 tutorial series.

Copies of *Analog Dialogue* are available by calling Analog Devices' Literature Center at 1-800-ANALOGD. This is and a number of back issues can be viewed on Analog Devices' Web site at <www.analog.com/publications/magazines/Dialogue/dialog.html> or by clicking on Quick Links/Analog Dialogue on the home page.

HP Capacitance Meter

Hewlett-Packard's HP 4268A capacitance meter can measure the speed and test-signal level required for high-value multilayer ceramic capacitors (MLCCs). The HP 4268A 120 Hz/1 kHz has fast auto-level control function, standards-compliant test signal level, and high-speed measurement of 25 ms/pt.

For more information write: Hewlett-Packard Company, Test and Measurement Organization, 5301 Stevens Creek Blvd., MS 54LAK, Santa Clara, California 95052.

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THE DOUBLE RECTANGLE

Three variations on a rectangular theme

The double rectangle (DR) is a fascinating antenna for the low bands. It's a vertically polarized rectangle with a vertical middle element. As **Figures 1** through **3** illustrate, the middle vertical can be placed anywhere from one end of the rectangle to its center. The feedpoint can be located in the center of that middle element or in any of the rectangle's corners.

When the antenna is relatively short, and the center element is near one of its sides, it exhibits the properties of a simple rectangular antenna; but because of its higher gain, I'd call it an enhanced rectangle or ER (see **Figure 2**). When it's long, as in **Figure 3**, and the middle element approaches its geometric center, it becomes a symmetrical double rectangle or SDR—an antenna which has also been called a double magnetic slot or DMS. The antenna in **Figure 1** lies in the mid-range length. It's the asymmetrical double rectangle, or ADR—a highly flexible, high-gain antenna that can be tailored to fit between any two available supports. If a site can support a dipole, it can support one of the DRs with a much higher low-angle gain and high-angle rejection.

In this article, I'll outline the background of the DR and its historic roots. Then I'll show how the antenna's performance varies as it gets longer and how it undergoes transformations from an ER to ADR to SDR. I'll analyze and explain these transformations and provide construction suggestions to aid in choosing lengths for best performance. Finally, I'll examine how well the 80-meter antennas, which are the main focus of this article, can be scaled to 40 and 160 meters.

Background

DX operators have always searched for a "magic" antenna for the low bands which would provide gain, interference rejection, work at practical heights, and use a minimum amount of territory. By gain, I mean "useful" gain at the elevation angles covering DX propagation on 40, 80, and 160. N6BV has published exhaustive data on the incoming propagation angles on all of the ham bands.[1] It's the DXer's goal to try to match his antenna's characteristics to the appropriate range of these angles.

This ideal antenna would provide significant gain over a reference dipole at these takeoff angles, be of comparable length, and be at an attainable height over ground. The only kind of radiator with these properties is a vertically polarized one. Unfortunately, with this type of radiator, one must deal with the ground losses in the near field and the possibility of radials.

There is, however, a class of vertical radiators which appear to be relatively independent of near-field ground losses due to their radiation dependence on their magnetic field. These

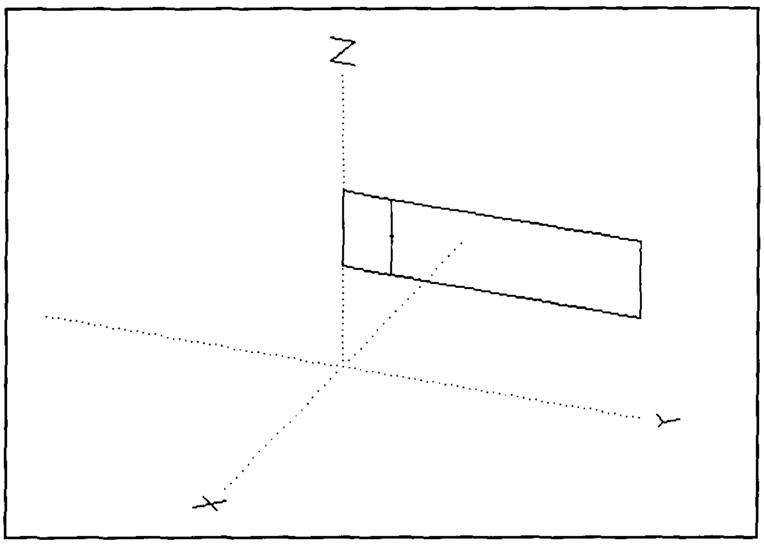

Figure 1. Asymmetric double rectangle.

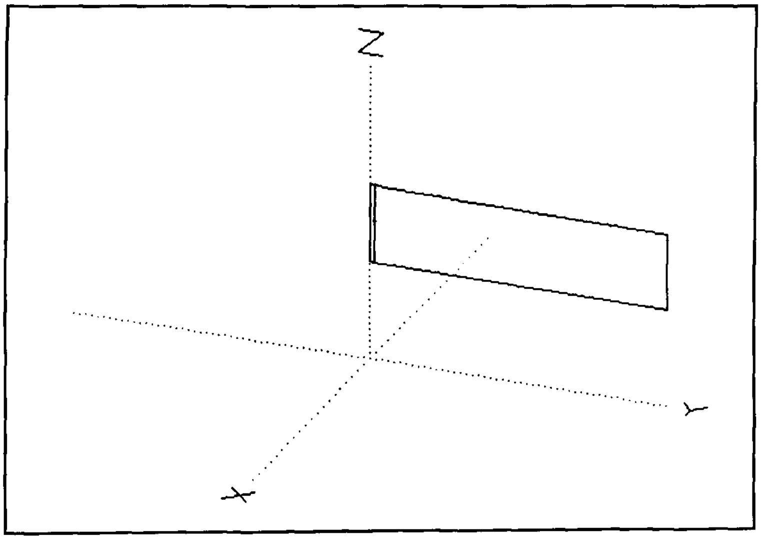

Figure 2. Enhanced rectangle (ER).

are the self-contained verticals or SCVs, as they are called by L.B. Cebik, W4RNL.[2] Antennas with this ground-independent property include the half-wave vertical, all types of loops from quads to deltas, and rectangles and loop-derived antennas like the half square and bobtail curtain. In examining the requirements I have outlined, Cebik concluded that the rectangular loop best meets these criteria on the low bands.

Rectangular loops

The simple rectangular loop for 80 meters (**Figure 4**), which I have modeled after Cebik's data and tested, provides 4.3 dbi gain in free space (fs) and about 3 dbi over ground (og). This is at a length of 115 feet which, at 3.5 MHz, is considerably shorter than that of a dipole (which is about 133 feet). When strung at reasonable heights with the top wire at anywhere from 50 to 90 feet, this antenna has a tremendous advantage over a comparably situated dipole in usable gain for DX. **Figure 5** illustrates the vertical radiation patterns of the dipole and the rectangle. Although the dipole has more gain, the gain is over a range of vertical radiation angles that are useless for DX. The rectangle is clearly the superior antenna for DX and for high-angle rejection.

We also know that other types of rectangles provide even more gain. K5RP published the so-called magnetic radiator[3] antenna and K4VX[4] did the same for the double magnetic slot (DMS). At one point, I used K5RP's antenna for over a year with success but, due to its wiring complexity and the prevailing winds and ice, it didn't last. The longer DMS is simpler to construct and more likely to survive the northern winters, but takes up a large amount of territory.

I then set about to look for a gain antenna that met the above criteria, but which would be simple to construct and could be tailored to any length between 100 and 200 feet. This would enable a builder to make the antenna as long as the distance between any two available supports (trees, towers, etc.).

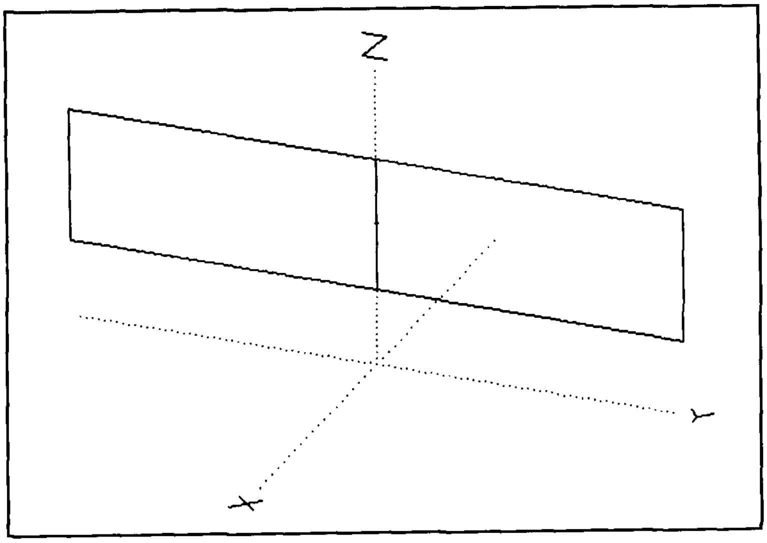

Figure 3. Symmetrical double rectangle (SDR).

The double rectangle or DR

I'd been playing with a model[5] of the Hentenna[6] for 10 meters. As shown in **Figure 6**, this is a horizontally polarized tall rectangular loop with a middle horizontal wire. The sides are 1/2 wavelength long and the spacing between the verticals is 1/6 wavelength. The middle wire is fed at its center and can be moved up or down to cancel out the feedpoint reactance. I found that when the loop is made taller and inductively reactive, moving the center feedpoint toward the geometric center of the antenna canceled out the reactance. The gain kept increasing as the antenna was made taller, and the reactance could be canceled up to the point that the middle horizontal wire finally reached the geometric center of the loop. The gain of the antenna is directly related to its length so the movement of the middle wire, by canceling out the inductive reactance due to lengthening, is the key to maintaining resonance at a higher gain. Once the middle wire is centered, you can see, by looking at the antenna, that no further change in its properties can be achieved because of its symmetry.

I then decided to look at the Hentenna on its side, vertically polarized and scaled for 80 meters. In essence, this is a double rectangle with the middle wire positioned anywhere from

Table 1
Enhanced Rectangle At Spacing of 0.9'
Compared to Simple Rectangle

Length	Height	ER Gmax	ER Gmin	Rect Height	Rect Gain	Rect
100	52.6	4.13	4.08	46.3	4.02	47.1
105	46.8	4.30	4.27	40.8	4.16	36.2
110	40.7	4.45	4.38	35.2	4.27	26.8
115	34.5	4.55	4.47	29.6	4.30	19.0
120	28.1	4.51	4.43	24.0	4.18	12.9
125	21.4	4.23	4.20	18.4	3.78	8.5

Table 1. Enhanced rectangle at spacing of 0.9 feet compared to simple rectangle.

one of its sides to the center. The first thing I noticed was that, when using the same horizontal dimensions as the simple rectangle (lengths ranging from 90 to 120 feet), the gain was greater than that of the rectangle when its length was 100 feet or larger. **Table 1** provides a comparison of the gains of the rectangle and the DR with the smallest usable spacing of 0.9 feet. (Note that all modeling lengths used here are in whole number increments of 5 or 10 feet, except in areas where more resolution is necessary to illustrate important changes. The heights, as well as the location of the middle wire, can be fractional.) What **Table 1** also shows is that, if you don't quite get the dimensions for the narrow-spaced DR just right, there will be some loss of gain; however, the gain (Gmin) in the end will always be greater than that of the simple rectangle. More on this later.

When you compare the performance of the

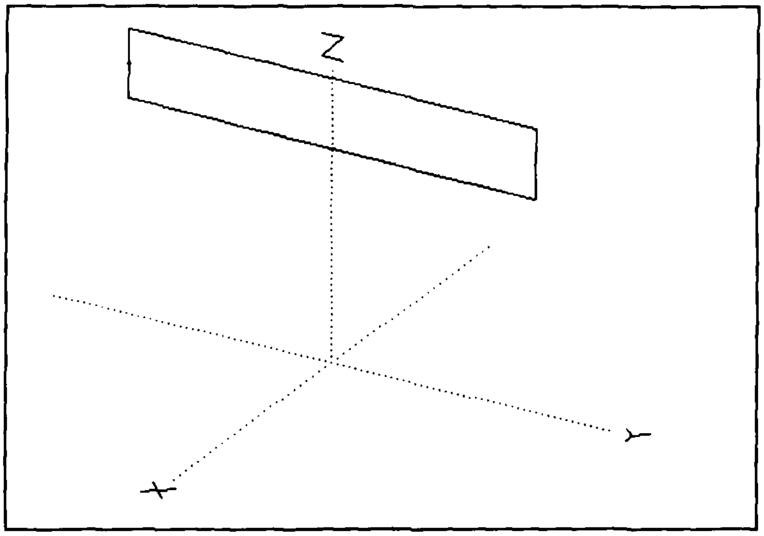

Figure 4. Rectangle.

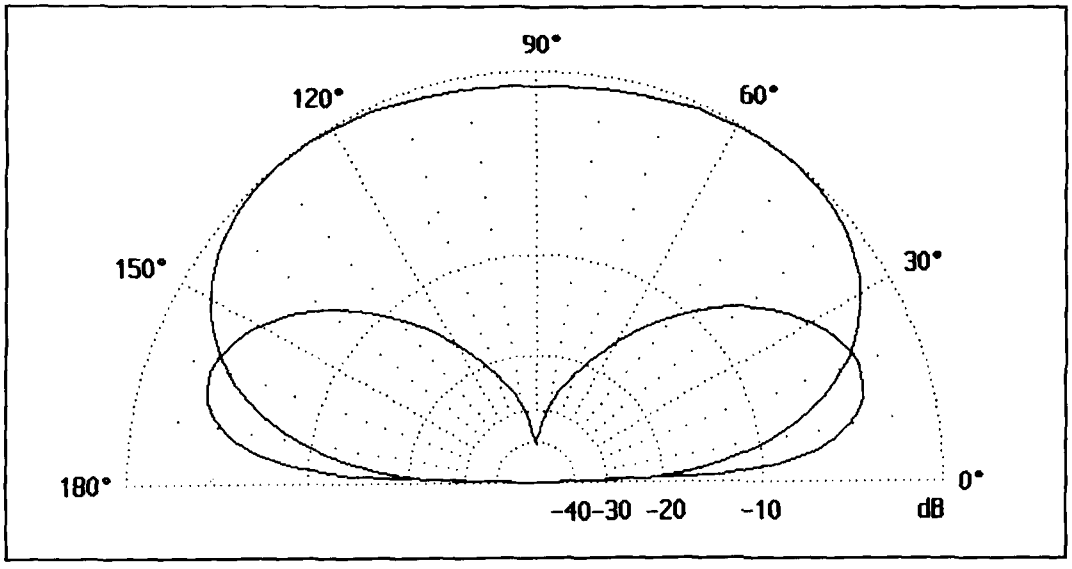

Figure 5. Rectangle and dipole: Top wire height of 80 feet.

Table 2

DOUBLE RECTANGLE ANTENNA DATA

L	Spc	Ht:	Gcent	Zcent	Gnear	Znear	Gfar	Zfa r
100.0	0.9	52.5	4.13	213.0	4.01	339	4.06	88
105.0	0.9	46.8	4.30	165.0		251		63
110.0	0.9	40.8	4.44	123.0		182		44
115.0	0.9	34.5	4.54	86.5		127		29
116.0	0.9	32.2	4.55	80.0				
117.0	0.9	31.5	4.55	73.8		110		25
118.0	0.9	30.7	4.54	67.9				
118.0	4.1	35.5	4.56	74.5				
119.0	6.0	36.7	4.58					
120.0	6.0	35.3	4.59	67.3	4.56	166	4.58	30
121.0	7.7	36.0	4.60	66.3				
122.0	7.7	34.6	4.61					
123.0	7.7	33.8	4.62					
124.0	11.3	36.3	4.63					
125.0	11.3	35.0	4.64	53.8		210		30
130.0	15.4	33.2	4.67	45.0		226		27
135.0	22.0	34.5	4.71	43.4		280		30
140.0	26.6	33.5	4.74	37.9	4.71	296	4.73	28
145.0	32.7	34.3	4.78	36.8	4.75	330	4.77	30
150.0	38.5	34.7	4.82	35.5	4.79	350	4.81	31
155.0	44.2	35.0	4.86	34.3	4.83	360	4.85	33
160.0	50.7	36.0	4.91	34.6	4.87	361	4.90	36
165.0	58.5	38.1	4.97	36.8	4.93	348	4.96	44
166.0	60.0	38.5	4.98	37.1				
167.0	64.0	40.9	5.00	41.2				
168.0	63.0	39.0	5.01	37.5	ADR to SDR convergence of gain			
168.0	84.0	48.9	5.01	56.3	At 168'			
169.0	68.0	42.0	5.03	42.6				
170.0	73.5	44.6	5.05	47.1	5.01	265	5.03	87
170.0	85.0	47.9	5.05	53.2	5.02	166	5.02	166
171.0	78.0	45.9	5.07	47.5	5.03	226	5.05	107
172.0	82.0	46.4	5.09	49.5	5.05	190	5.07	127
173.0	83.0	46.2	5.11	48.5				
175.0	80.0	43.9	5.14	43.5	5.11	207	5.12	95
175.0	87.5	45.4	5.14	46.1	5.12	145	5.12	145
180.0	90.0	42.9	5.23	39.7	5.21	127		
190.0	95.0	38.2	5.39	29.4	5.37	97		
200.0	100.0	33.6	5.51	21.5	5.50	73.5		
205.0	102.5	31.3	5.56	18.3	5.54	63.8		
210.0	105.0	29.1	5.59	15.5	5.57	54.6		
215.0	107.5	26.9	5.59	13.1				
220.0	110.0	24.8	5.56	10.9	5.56	39.9		

Table 2. Rectangle: heights and gains versus length and optimum length/height ratio.

110-foot rectangle with a similar DR over ground and at a top-wire height of 80 feet, the gains are 2.98 dbi and 3.08, respectively, at a takeoff angle of 18 degrees. **Figure 5**, which compares this rectangle with a dipole at the same height, shows the dipole has most of its gain, or what I would call interference acceptance, at much higher elevation angles. At an angle of 10 degrees, which is right in the middle of the range of useful angles for DX, the performance advantage of the rectangle over the dipole is large: -3.06 dBi for the dipole and +1.76 dBi for the rectangle, or a gain difference of almost 5 dB. Below this takeoff angle the advantage of the rectangle increases so, at 5 degrees, it's about 7.2 dB.

Analysis

Note that between the extremes in size of the short, simple rectangle and the long, symmetrical double rectangle, the DR is the antenna best able to use the middle range. The following data details the analysis of the DR, which begins as an enhanced gain rectangle, or ER, then becomes an asymmetrical DR, ADR, and finally transforms into the symmetrical DR, SDR. At the end, I'll focus on what I call the "sweet spots." These are the areas with easily matched impedances where construction is less finicky as to achievable gain. I also examined the other available feedpoints on the DR and the rectangle, as the corners and the bottom of the middle wire might offer a better match than the middle wire and would be more accessible for feeding.[7] As you'll see further on, one can always find a convenient feedpoint for 50-ohm coax—regardless of the length of the array. All the antenna parameters are summarized in **Table 2** for antenna lengths of 100 to 220 feet.

The simple rectangle

Before examining the properties of the more complicated double rectangle antennas, I'd like to look at the simple rectangle. This is the open rectangle (**Figure 4**) with the opening at the feedpoint in the middle of one of its vertical sides.

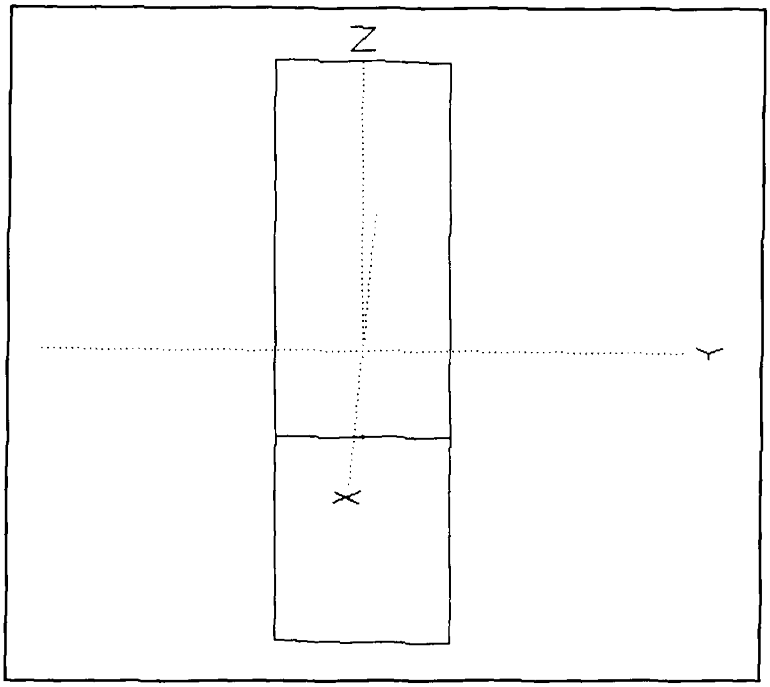

Figure 6. The Hentenna.

The antenna is a full wavelength loop so, to maintain resonance, the height and length are inversely related. As demonstrated by W4RNL, there's an optimum length/height ratio for maximum gain which is frequency dependent.[8] The formula for this ratio is: $R = \sqrt{2\log(100f)}$ (where frequency is in MHz). At a frequency of 3.5 MHz, this ratio calculates out to 3.59.

Table 3
Rectangle: Heights and Gains v. Length and Optimum Length/Height Ratio

Length:	Height:	Gain	Zin	L/H Ratio	
100	46.3	4.02	47.1	2.16	
105	40.8	4.16	36.2	2.57	
110	35.2	4.27	26.8	3.12	
112		4.29			
113	31.8	4.30		3.55	peak L/H ratio
114		4.30			
115	29.6	4.30	19.0	3.89	
116		4.29			
117		4.28			
120	24.0	4.18	12.9	5.00	
125	18.4	3.78	8.5	6.79	

Table 3. Double rectangle antenna data.

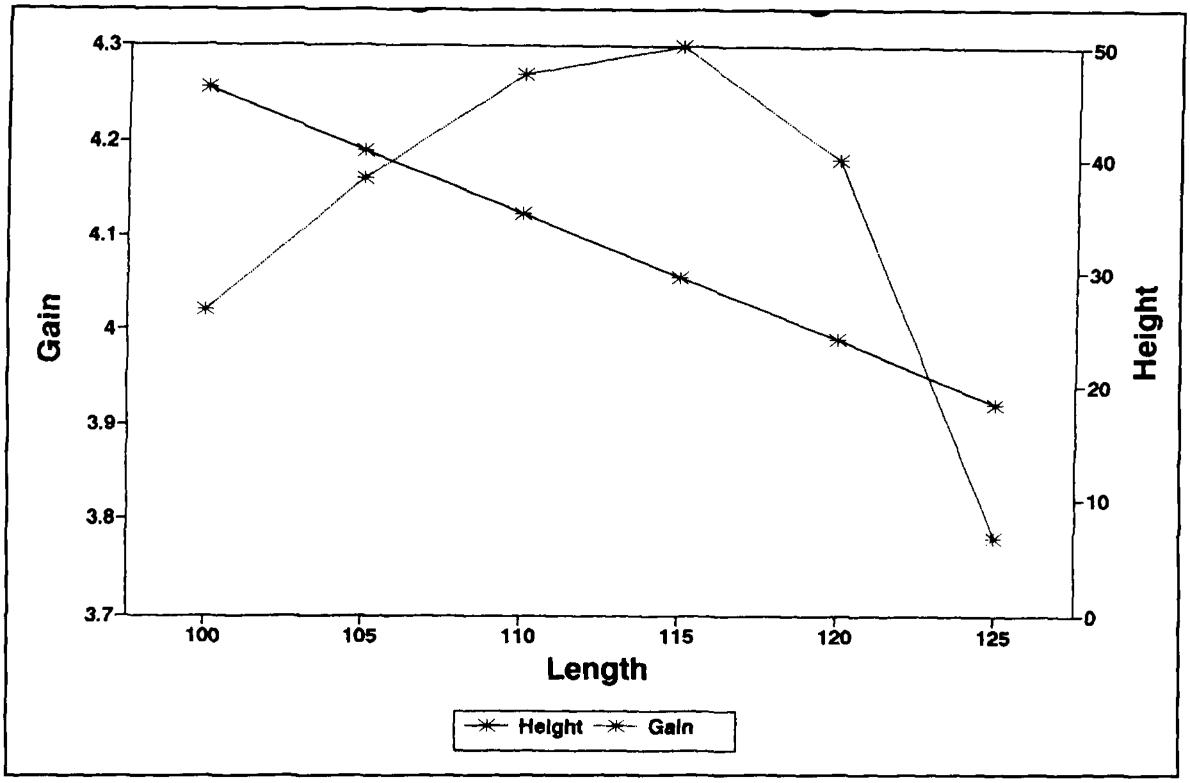

Figure 7. Rectangle: Height and gain versus length.

Table 3 and **Figure 7** show the following information about the rectangle: (1) there is a linear and inverse relationship between the length and the height, the gain peaks at a specific L/H ratio (length of 113 feet and a height of 31.8 feet); and (2) there is a rapid falloff of feedpoint impedances as the rectangle lengthens. At that 113-foot length, which results in the peak gain of 4.30 dBi in free-space, the L/H ratio is 3.55 and is in accord with the formula in the preceding paragraph.

Available feedpoints on the double rectangle

The DR can be fed in four places.[9] These are at the lower corners, denoted in the tables and graphs as "near" and "far" depending on their relative position to the center vertical, and at the center and bottom of the center vertical (marked as "center" and "ctr/bot"). The gains, impedances, and the center wire positions at these feedpoints are found in **Table 2**.

Gain

Figure 8 shows that the gain of the DR rises until an overall length of 210 feet is reached. No further gain is realized beyond this point, so there's no advantage in making it longer. What's interesting in **Figure 8**, and shown in more detail in **Table 2**, is that there appears to be a linear increase in gain between 115 feet and 165 to 170 feet (more about these specific length ranges later), followed by a more rapid increase in the slope of the gain curve until a maximum gain of 5.59 dBi is reached at 210 feet.

Although the best gains are obtained at the center of the center wire, this is the least accessible feedpoint. The corners and the center/bottom are more accessible, at a tradeoff of a few hundredths of a dB in gain.[10]

Feedpoint impedances

Figure 9 and **Table 2** show how the feedpoint impedances in the center and in the near and far corners vary with length. All three feedpoint impedances start off high when the antenna is short (100 feet) and the middle wire is close to the end of the loop (0.9 feet). The changes appear to be sinusoidal. The Zcenter (center feedpoint) reaches a minimum at length 155 feet and starts to rise again until 170 feet when it abruptly inflects and drops linearly. From length 125 feet on, it offers a direct match for 50-ohm coax with minimal SWR. The bottom/center element impedances are approximately 5 ohms greater than those at the center of the center element.

Znear (near feedpoint) begins at 320 ohms,

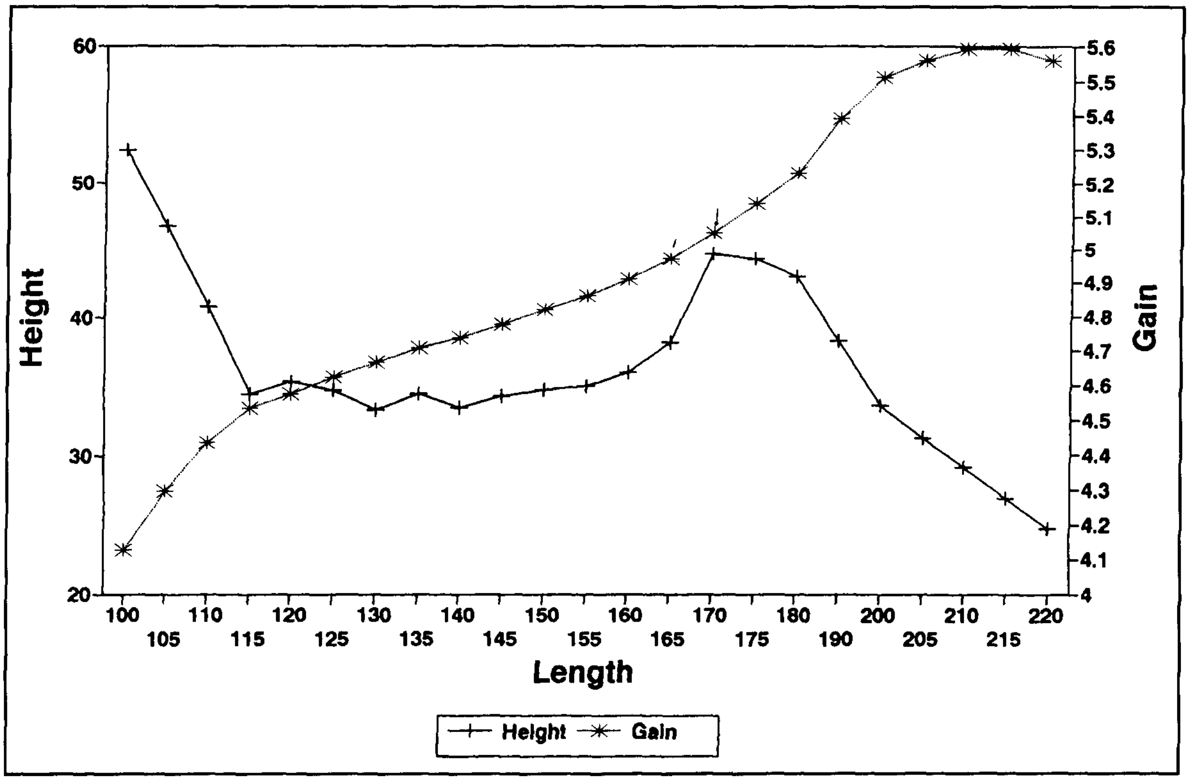

Figure 8. Double rectangle: Height and gain versus length.

falls to a low of 110 ohms at length 117 feet, peaks at over 350 ohms at 160 feet and then falls off. (The closely spaced lengths around 115 to 118 feet and between 165 and 170 feet do not show up on the graph. **Table 2** shows what happens at these inflection points.) Similarly, Zfar (far feedpoint) starts at 85 ohms at 100 feet, reaches a nadir of 25 ohms at 117 feet, and peaks at 94.6 ohms at 175 feet. **Table 2** also shows that at 170 feet, when the antenna becomes a symmetrical DR (spacing = L/2 or 85 feet), the corner impedances become equal and higher than the center wire impedance.

There may be some confusion about the impedances in the 170 to 175-foot range because the antenna can be used either in its symmetrical (SDR) or in the asymmetrical (ADR) form. Look at **Table 2**. At lengths of 168 feet and above, you have two options because the ADR and SDR gains converge. For instance, at a length of 168 feet you can build the antenna asymmetrically, at a spacing of 63 feet, if you want to take advantage of a lower center feedpoint impedance. If you want a higher feedpoint impedance, you can use a spacing of 84 feet, which is half of its length. Gain convergence means the gain (5.01 dBi) will be the same whichever mode you wish to use, symmetrical or asymmetrical.

So, at a length of 168 feet and beyond, the antenna can be designed as an ADR or SDR, depending on your preferences. What you see in **Figure 8** is the impedances modeled for the ADR up to 170 feet and for the SDR between 170 and 220 feet.[11] **Table 2** shows that, at a length of 170 feet, the near and far corner feedpoint impedances are different when the antenna is asymmetrical and equal when the antenna is symmetrical. As for the longer antennas, those greater than 180 feet, the curves in **Figure 9** for the near and far feedpoints may appear different, but they are not. They are identical except for a difference in scaling.

Height

The heights I used to create the curves were the LOWEST heights that resulted in maximum gain. As the DR gets longer, especially at 130 feet and beyond, there's a wide range of heights (and spacings from the end) at which the gain stays constant. **Table 4** shows where you have the most leeway in changing the dimensions for impedance matching while still maintaining peak gain. If you increase the height, and therefore the spacing at a given length, the feedpoint impedance will rise. Refer to **Table 4** for the minimum and maximum feedpoint impedances at peak gain.

Figure 8 shows us that initially, as the DR lengthens, the height decreases. This is similar to the behavior of the simple rectangle shown in **Figure 7**. However, with the double rectan-

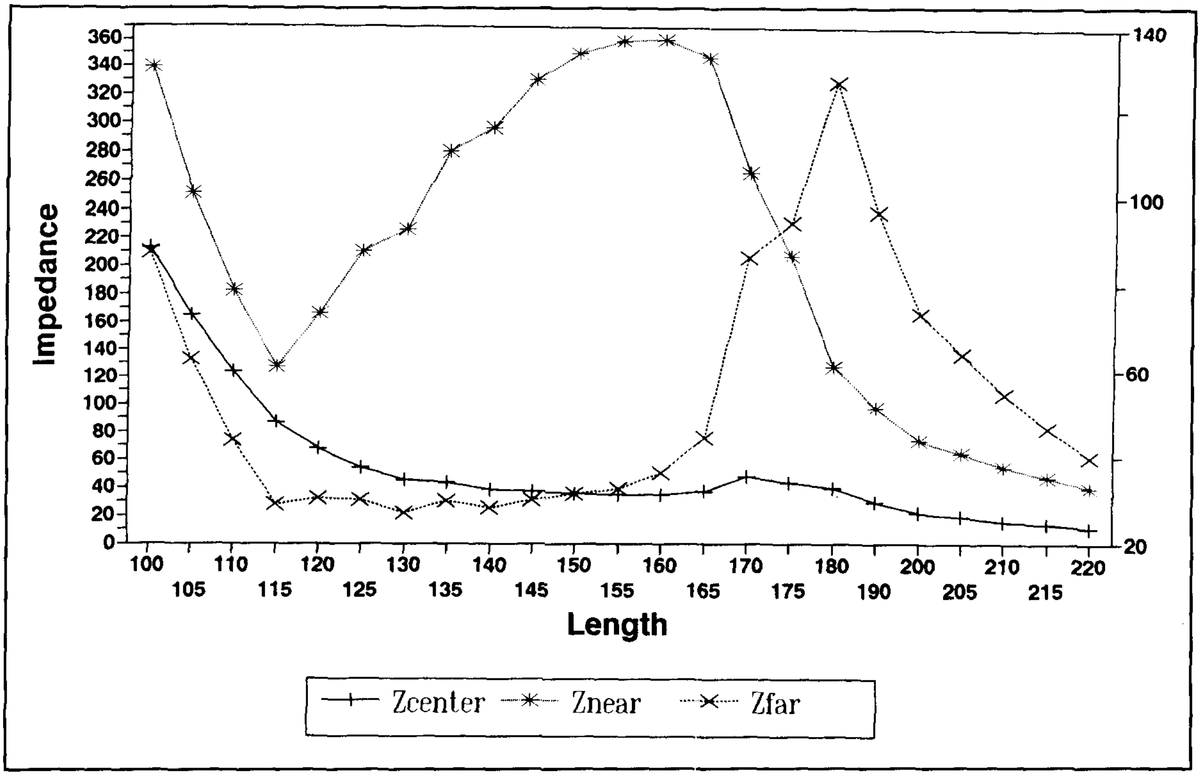

Figure 9. Double rectangle; Impedance: Center, near, far.

gle, the height bottoms out when it reaches a length of 117 feet (height = 32.3 feet) and it remains relatively constant as the length increases until there's an upward inflection when the length reaches 165 to 170 feet. As I mentioned above, this range is where the transformation from asymmetrical to symmetrical takes place. If you look closely at **Table 2**, you'll see that, if the antenna is asymmetrical, the greatest height is reached at a length of 172 feet; if it's symmetrical, the maximum height is at 168 feet. At any rate, after 172 feet, the height falls as the antenna is made longer because there's no longer any room for center wire compensation and tuning must be via changes in dimension.

Middle-wire location

When the middle wire is closest to the near side—at 0.9 feet—the gain of the antenna peaks at a length of 116 to 117 feet. If you lengthen the antenna further at that narrow spacing, the gain drops. From that length on, the middle wire must be moved further towards the center, which is what you see in **Figure 10**. This curve is linear until, near a length of 165 feet, it changes slope more rapidly in the transition zone around 165 to 170 feet and is linear thereafter, once the antenna has become an SDR.

The location of this wire provides the first clue to the transformations that exist at 117 feet and near 168 feet. From 100 to 117 feet,[12] the middle wire results in optimum gain when it is a fraction of a foot from the near end. As best as I could model, using very high segment densities, any movement closer to and away from the end resulted in decreased gain. Beyond 118 feet, as the DR lengthens to 168 through 172 feet, the middle wire begins to approach the geometric center of the antenna and, when that occurs, there's no further reactance compensation. At that SDR point, the array is split into two equal-sized rectangles and the middle wire remains centered as length is increased (at spacing = L/2). The only way to maintain resonance is by decreasing the height. The change in height, which is inversely proportional to the length and which you can see in **Figures 8** and **14**, results in a drop in all feedpoint impedances, as shown in **Tables 2** and **5** (under Zsdr).

Reactance compensation by the middle wire

I found that, as the antenna lengthens beyond 160 feet, the middle wire requires much more movement centerwards to cancel out the reactance caused by lengthening. It becomes progressively more ineffective in its reactance-can-

Table 4 Height and Spacing Tolerances At Maximum Gain								
								Height
L:	Gain	Zmin	Zmax	Spmin	Spmax	Htmin	Htmax	Leeway
100	4.13	213.0		0.9		52.5		
105	4.30	165.0		0.9		46.8		
110	4.44	123.0		0.9		40.8		
115	4.54	86.5		0.9		34.5		
120	4.58	67.7	69.0	6.0	6.7	35.4	36.1	0.7
125	4.63	55.3	63.0	11.1	13.9	34.7	38.0	3.3
130	4.67	45.1	55.1	15.4	19.0	33.3	37.5	4.2
135	4.71	43.4	51.4	22.0	25.0	34.5	38.0	3.5
140	4.74	37.9	51.5	26.6	32.0	33.5	39.6	4.1
145	4.78	36.8	47.0	32.7	37.0	34.3	39.0	4.7
150	4.82	35.5	46.6	38.5	43.5	34.7	40.0	5.3
155	4.86	34.3	48.1	44.2	51.0	35.0	41.8	6.8
160	4.91	34.6	47.9	50.7	58.0	36.0	42.8	6.8
165	4.97	36.8	49.9	58.5	67.0	38.1	44.9	6.8
170	5.05	47.1	53.2	73.5	85.0	44.6	47.9	3.3
175	5.14	44.2	46.1	81.0	87.5	43.9	45.4	1.5
180	5.23	39.7		90.0		42.9		
190	5.39	29.4		95.0		38.2		
200	5.51	21.5		100.0		33.6		
205	5.56	18.3		102.5		31.3		
210	5.59	15.5		105.0		29.1		
215	5.59	13.1		107.5		26.9		
220	5.56	10.9		110.0		24.8		

Table 4. Height and spacing tolerances at maximum gain.

celing job as the antenna approaches symmetry. Or, to put it conversely, changes in middle wire position are more critical when the antenna is short, near the area of the ER to ADR transformation, than when it is long, at the ADR-to-SDR transformation.

Analysis

When does the rectangle transform itself into something different?

When you look at **Table 2** and the curves in **Figures 8** and **9**—changes in height with length, changes in the feedpoint impedances, and changes in the slope of the gain—you'll note there are clear-cut transitions that occur at lengths of 117 and 168 feet. It's obvious that the far transition, when the antenna lengthens beyond 168 feet, leads to its becoming a symmetrical DR. At the near transition (**Table 6A**), when you examine the middle wire location, you can see that the middle wire stays close to the near side of the DR until the length reaches 117 feet. It then begins an almost linear rise until the high inflection point, which you see in **Figure 8**.

The middle wire is the key to this antenna. Its function is to maintain resonance as the antenna increases in length. With a simple rectangle, the circumference is basically fixed at or near one wavelength. You can increase the length, but then you must comensurately decrease the height. When the height is decreased, two things occur. First, there is a unique point or height-to-length ratio, seen in **Figure 7**, where gain is maximal. Second, the lower the height, the lower the feedpoint impedance, either at the center or at a corner. There's no point in lengthening the antenna because gain decreases and it becomes harder to feed.

The graphs show that the DR is a glorified rectangle as length increases to near 117 feet. **Figures 11** through **13** illustrate that, at all spacings, the double rectangle acts as a simple rectangle with higher gain. There is a gain peak

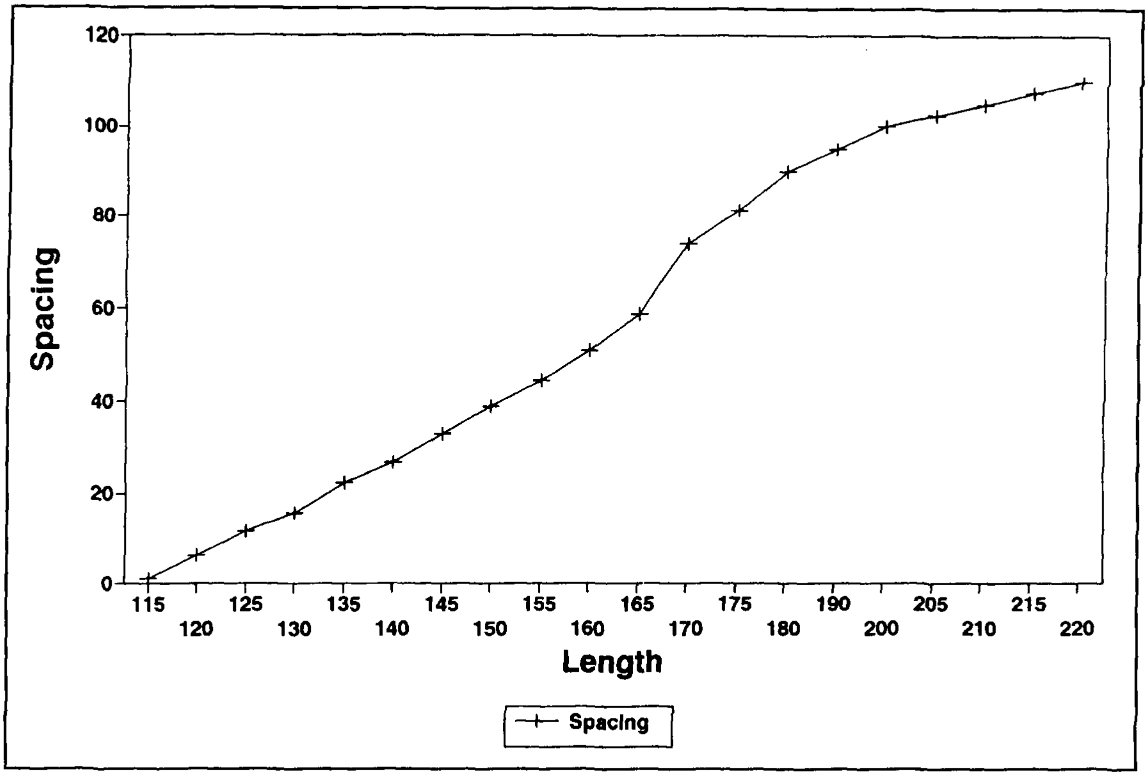

Figure 10. Asymmetric double rectangle: Length versus spacing.

at each length determined by the spacing (**Figure 13**), and the slopes of the antenna heights versus length at different spacings are parallel to each other (**Figure 12**). However, up until the 117-foot length, the height falls as length increases and the middle wire stays close to the end. The changes between 115 and 118 feet presented most clearly in **Table 6A.**

Figure 11 and **Table 1** illustrate what I mean by the term "glorified" rectangle, or what I would now call an enhanced rectangle (ER). The gain, when the DR is longer than 100 feet, exceeds that of any simple rectangular loop because its height and loop size are greater than that of the rectangle. In this ER phase, the small and constant middle wire position out about 1 foot from the near end allows its height to exceed that of a rectangle and allows the gain to peak at the length of 117 feet. What you see in **Table 6A** and **Figures 11** and **12** is that the slope of the L versus H is parallel to and higher than that of the rectangle, and that the gain peak and the subsequent drop off occur later.

Beyond 117 feet, the middle wire must move toward the antenna center to attain the maximum possible gain. The height, seen in **Figure 8**, remains relatively low and stable until a length of 165 feet is reached. Between the lengths of 160 and 170 feet, the height appears to increase sinusoidally and then drops abruptly as the antenna gets longer.

Between 120 and 168 feet, the middle wire effectively compensates for the increasing length, and the antenna is in its "Hentenna" mode. Length, middle wire position, and gain are linearly related. At any spacing, as illustrated in **Figure 12**, the DR has a L/H line parallel to all of the others at other spacings. It is also parallel to that of the simple rectangle, which can be imagined as a DR with a spacing of zero. **Figure 13** shows that, at each specific length, there's a unique spacing which produces the maximum gain for that length.

Furthermore, as the antenna lengthens, there's a less abrupt change in gain with changes in spacing, and there's a broader range of spacings at which maximum gain is maintained. This makes construction easier.

Transformation to an SDR or symmetrical double rectangle

Per W4RNL, the better name for a double loop rectangle where the middle wire is centered is the SDR. This describes its appearance and nature fully without the need to allude to its magnetic radiation properties or calling the center wire a "slot."

When I was first modeling, I noticed the sudden transition at the 170-foot length. Past this length, (**Figure 8**), the loop height begins to drop, as do the feedpoint impedances. The gain keeps increasing, but the middle wire is at or near the center and stays there. Gain becomes asymptotic at around 5.6 dBi at a length of 210 feet.

What I then decided to do was to model backwards. That is, I decreased the length from 170 to 100 feet, while maintaining the middle wire at the center. I then plotted the heights (**Figure 14**) and gains (**Figure 15**) and found that changes were absolutely linear and reciprocal. Height went up and gain went down as the antenna was shortened from 175 to 100 feet. The gain reached that of the optimized 110-foot rectangle when its length was 135 feet. It was bigger, but not better. Most importantly, its gain was very inferior to the DR until the gain curves met at 168 feet. This is where the transformation to SDR took place.

Table 6B shows, in finer detail, the point at which the transformation from asymmetrical DR to symmetrical DR occurs. Antenna heights and feedpoint impedances are rising up to a length of 168 to 170 feet. Afterwards, they both begin the drops that continue from then on. The peak gains of the asymmetrical and symmetrical antennas become equal at 168 feet. So, how is gain increased as the antenna lengthens?

Figures 13, **16**, and **17** reveal that, at each specific length modeled, there is a unique middle wire spacing that results in an optimum gain peak. **Figure 17** demonstrates that, at short lengths and narrow spacings, there's a drop off in gain when the middle wire position is moved to either side of its optimum position. **Figures 16** and **17** show that, as the antenna becomes longer, each such step involves an abrupt shift upwards in gain until the optimum spacing for a specific length is reached.

It's probable that the increase in gain is related to an increase in circumference caused by lengthening. For the middle range of lengths of the DR, **Figure 8** shows that the height remains relatively stable due to the ability of the center wire to cancel out the added inductive reactance. Paradoxically, between 160 and 170 feet, the height actually increases as the length increases.

Up to a length of 125 feet, the gain peaks associated with increases in spacing are very abrupt. **Table 4** (in the column labeled Height Leeway) and **Figure 16** show that, beyond a spacing of 11 feet (when you study a 125-foot ADR) the changes in gain are much less abrupt and that there's a long plateau of maximum gain. From a length of 125 feet, and up to 170 feet where the antenna becomes symmetrical, there's a wide latitude in spacing and height at which peak gain is maintained. As the antenna is made longer, and the spacing moves the center wire further from the near end, the gain rises gradually and the curves become smoother with little or no decrease in gain between optima.

From looking at the linearity of the changes in **Figure 10**, you'd guess that the transitions should be much smoother; gain should go up gradually as the optimum spacing is approached for a given length. Also the drop off in gain for a given spacing should also be smooth and sinusoidal. This is what seems to

Table 5

Symmetrical Double Rectangle Data
Spacing is L/2

Length:	Ht:sdr	Gsdr	Zsdr	Ht:adr	Gadr
100	85.4	3.41	292.0	52.5	4.14
110	80.0	3.65	239.0	40.8	4.44
120	74.8	3.89	193.5	35.3	4.60
130	69.4	4.14	153.7	33.2	4.67
140	64.1	4.38	120.6	33.5	4.74
150	58.9	4.62	93.5	34.7	4.81
160	53.8	4.84	71.7	36.0	4.91
170	48.7	5.05	54.5	47.1	5.05
180	42.9	5.27	41.0	42.9	5.27
190	38.5	5.39	30.6	38.5	
200	34.7	5.52	22.6	34.7	
210	29.2	5.59	16.8	29.2	

Table 5. Symmetrical double rectangle data. Spacing in L/2.

Table 6A Near Transition From ER to ADR				
Length	Spacing	Height	Zin	Gain
100.0	0.9	52.5	213.0	4.13
105.0	0.9	46.8	165.0	4.30
110.0	0.9	40.8	123.0	4.44
115.0	0.9	34.5	86.5	4.54
116.0	0.9	32.2	80.0	4.55
117.0	0.9	31.5	73.8	4.55
118.0	0.9	30.7	67.9	4.54
118.0	4.1	35.5	74.5	4.56

Table 6A. Near transition from ER to ADR.

Table 6B Far Transition from ADR to SDR								
Length	Gadr	Gsdr	Spmin	Spmax	Htmin	Htmax	Zmin	Zmax
165.0	4.97	4.94	58.5	67	38.1	44.9	36.8	50
166.0	4.98	4.97	60.0	72.0	38.5	47	37.1	53.7
167.0	5.00	4.98	64.0	70.0	40.9	45.1	41.2	49.4
168.0	5.01	5.01	63.0	84.0	39.0	48.9	37.5	56.3
169.0	5.03	5.03	68.0	84.5	42.0	48.4	42.6	54.8
170.0	5.05	5.05	73.5	85.0	44.6	47.9	47.1	53.2

Table 6B. Far transition from ADR to SDR.

be happening when the center wire is moved more than 11 feet from the near end. So, why the "sawtooth" stepping in the gain curves when spacing is narrower?

Currents

When you look at the antenna currents in **Table 7**, you see that, among the three verticals at very close spacings, the far wire carries twice the current of the near and center wires. There are phase differences between the near and far verticals that cause interference in the far field which can be additive or subtractive, and result in the "sawtooth" gain curves.[13] As the spacing becomes greater, the phase shifts become less destructive, and the gain curves smooth out so the transitions up to and beyond the gain maxima are minimal. This is shown in **Figure 16**, which models a 125-foot ADR at spacings from 1 to 14 feet. Note that the gain holds at a peak between spacings of 11.1 and 13.9 feet. This results in the 3.8-foot height "leeway" referred to in **Table 4**.

In the ADR phase, with intermediate spacings,[14] the middle and far verticals carry most of the current and there's less of a phase difference between the end verticals. When you reach the SDR stage, the current ratios and the phase angles become symmetrical about the center wire that carries the greater current, and the currents in the three wires approach a binomial distribution.

Another interesting piece of information found in **Table 7** is what happens at very close spacings. At a length of 115 feet, even at a spacing of 0.05 feet or a fraction of an inch, the gain approaches that of a simple rectangle but still exceeds it in spite of the great phase differences in the vertical elements. I threw in the current distribution of the rectangle for comparison; look at a length of 115 feet and a spacing of zero.

Practical construction tips: loop height tolerances

As I alluded to above, there are points on the curve that are critical and others that allow a lot of dimensional leeway. As the loop length increases from 100 to 117 feet, the corresponding heights necessary to achieve maximum gain are very critical. Changes in height of ±0.5 feet result in sharp decreases in gain. However, as

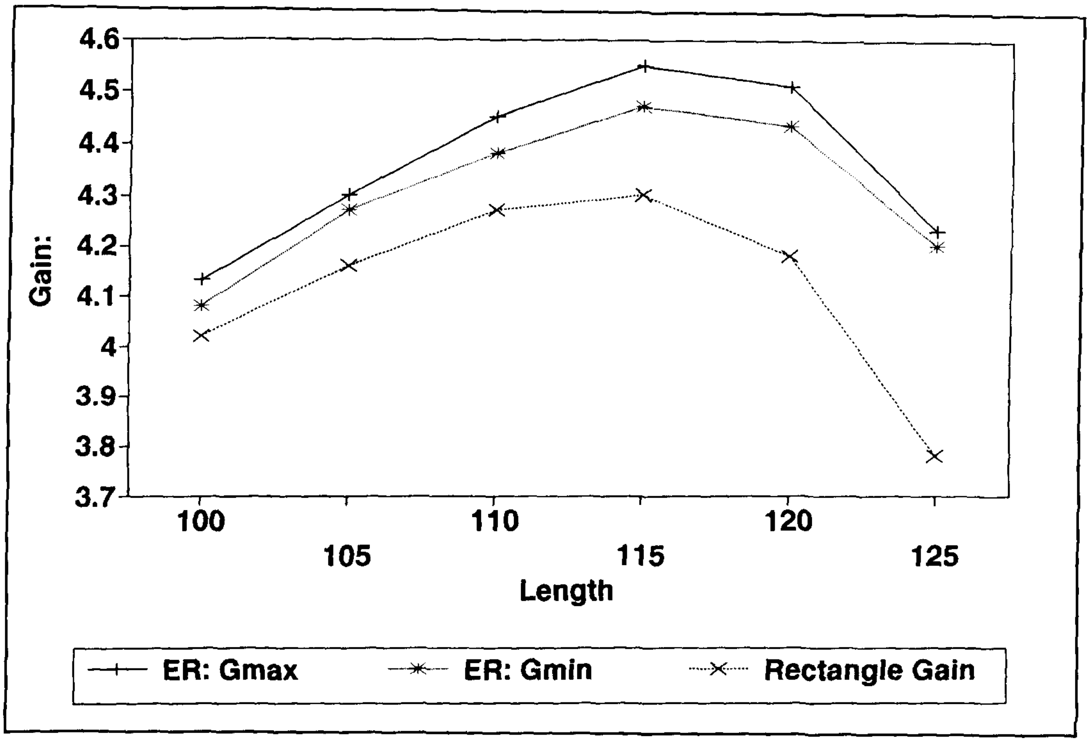

Figure 11. ER and rectangle: Gain comparison.

Figure 11 illustrates, you really can't go wrong if you're slightly off in spacing. In this enhanced rectangle phase, the lowest possible gains (Gmin) are still greater than that of a comparable simple rectangle. The problem in this range lies in obtaining a good match to a feedline because of the high impedances. But this problem is no different from that of the simple rectangle with its very low feedpoint impedances. At 100 to 110 feet, the far corner

Table 7
Currents: Rectangle, ADR and SDR
in the center, near and far verticals with the center fed
Currents are in E-02 amps

Len:	Spc:	Gain	Inear	Phase	Icen	Phase	Ifar	Phase
110	0.9	4.44	0.790	-157.9	0.81	0	1.51	-169.7
115	0.00	4.30	5.280	0.0	Double Rectangle		5.28	0.0
115	0.05	4.33	1.410	-137.9	1.07	0	2.27	-156.2
115	0.60	4.46	1.110	-158.0	1.13	0	2.14	-169.5
115	0.90	4.54	1.090	-160.7	1.16	0	2.15	-171.1
120	6.10	4.59	1.010	-168.7	1.48	0	2.36	-175.9
130	15.40	4.67	1.070	-172.1	2.22	0	3.07	-177.3
140	26.60	4.74	1.010	-175.0	2.64	0	3.28	-179.1
150	38.50	4.82	0.958	-177.6	2.81	0	3.20	-179.1
160	50.70	4.91	0.963	-177.7	2.89	0	3.04	-179.7
170	73.50	5.05	1.020	-177.6	2.13	0	1.77	-178.8
180	90.00	5.23	1.570	-179.5	2.52	0	1.57	-179.4
200	100.00	5.51	2.690	178.5	4.65	0	2.69	178.5
210	105.00	5.59	3.610	178.0	6.46	0	3.61	178.0
220	110.00	5.56	4.950	177.9	9.11	0	4.95	177.9

Table 7. Currents: rectangle, ADR, and SDR in the center, near, and far verticals with the center fed. Currents are in E-02 amps.

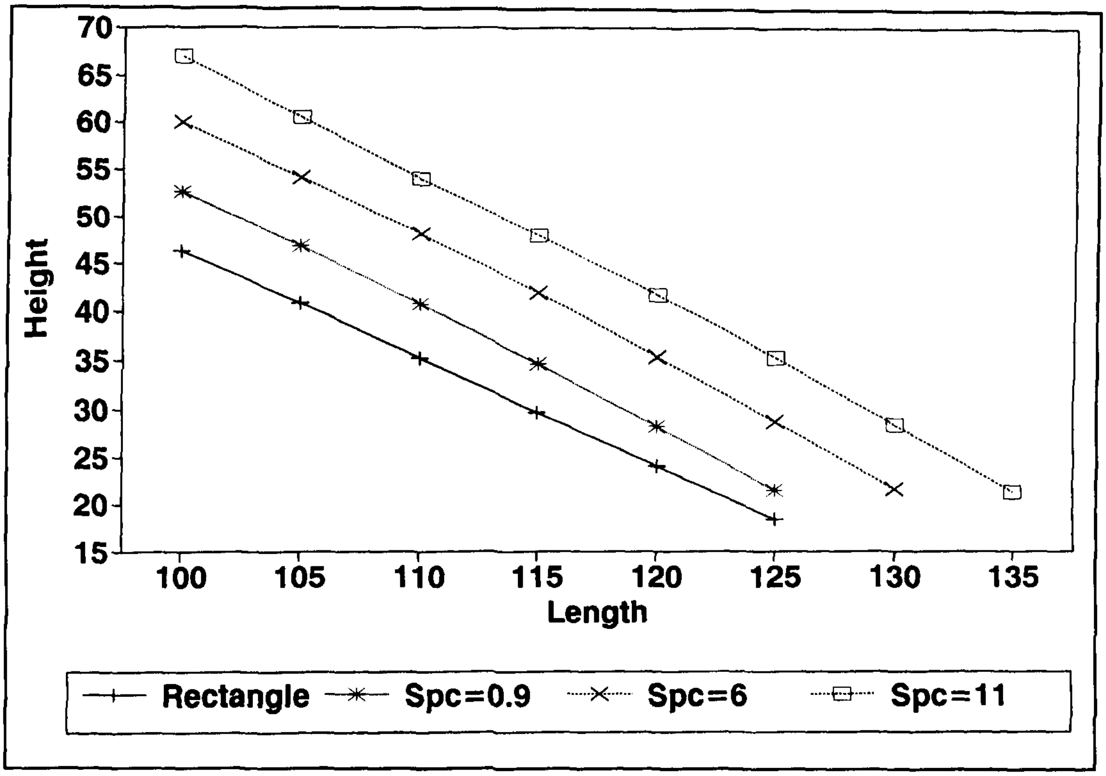

Figure 12. Rectangle-ER-ADR: Heights versus length.

offers a reasonable match to a feedline (see **Table 2** and **Figure 9**). At 115 feet, the center can be fed with a quarter-wave transformer of 75 ohms since, over ground, the feedpoint impedance will be close to 100 ohms.

When you get to 120 to 130 feet, the antenna can be fed directly at either of the center feedpoints with minimal mismatch. At 130 feet and beyond, the center feedpoints—although they have low impedances in free space—are useful because their impedance rises when the antenna is over ground to close to 50 ohms. However, a builder has lots of leeway in increasing height and spacing to achieve a perfect match, while maintaining peak gain. **Table 4** shows the leeways in both parameters that offer you freedom to play with the dimensions—from a length of 125 feet to one of 170 feet. For example, at a length of 140 feet, there's a usable range of heights between 33.5 to 39.6 feet while maintaining a gain of 4.74 dBi.[15] At a length of 130 feet, heights of 33.2 to 37.5 feet will be usable with appropriate center-wire movement.

If you construct this antenna, I strongly suggest that you go slightly above the minimal heights and spacings for the length you'll be using. You have lots of leeway, if you make the antenna taller and increase the spacing. Just don't make it any smaller as you'll lower the gain.

If you must go with one of the shorter antennas and use a spacing of 0.9 to 1 foot, you must vary the height to resonate the antenna as you would with a simple rectangle. Even though the height may be critical, and you may not land on a gain peak, you'll still have gain substantially above that of the rectangle.

Because the feedpoint impedance increases over ground, which you'll see in the design examples that follow, I believe the antennas in the middle range of lengths from 130 to 170 feet, with free space center wire impedances in the high 30s and 40s will provide an excellent match over ground. The only caveat, as stated above, is to make sure that you go slightly above the minimum dimensions for the length you'll be using.

Design examples

First, you'll notice that the antennas I've modeled were tuned to a frequency of 3.52 MHz. I did so because I like CW, and this puts the resonant frequency right in the middle of the contest frequency range on 80 with acceptable SWR. The bandwidth providing you an SWR of 2:1 is 3.495 to 3.545 MHz.

General tuning suggestions

If you increase the spacing, you'll raise the resonant frequency and vice versa. If you make the antenna taller, you'll lower the resonant fre-

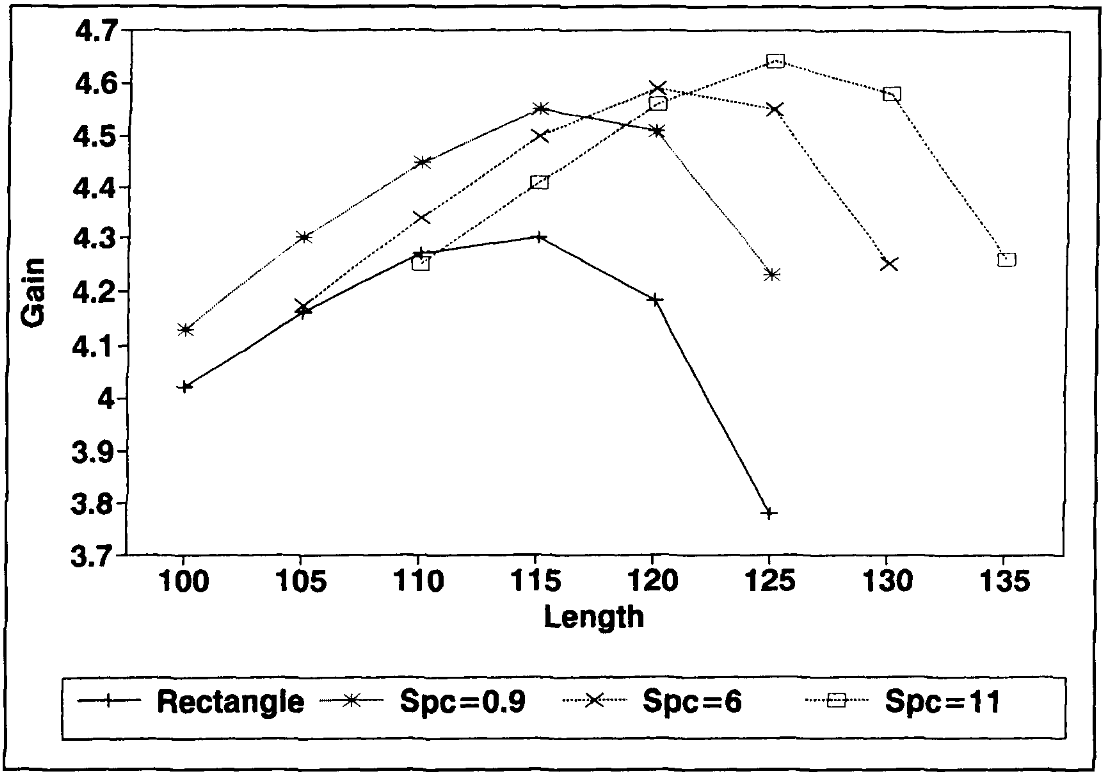

Figure 13. Rectangle-ER-ADR: Gain versus length.

quency. For example, a 140-foot ADR at a height of 33.5 and at a spacing of 26.6 feet will be resonant at 3.52 MHz in free space. If you make the height 34 feet, you'll lower the resonant frequency to 3.51. Over ground, at a height of 46 feet for the lower wire, it will resonate at 3.513 MHz with an input impedance of 51.5 ohms. If you then decrease the spacing to 26 feet, the antenna will be resonant at 3.499 MHz with an impedance of 51.7 ohms and no loss of gain. You can go the other way and resonate it at, say, 3.6 MHz by placing the center wire at 30.3 feet.

The reason for making the antenna slightly taller is to provide the latitude to decrease the spacing and still maintain the maximum gain. You may have to do this in real life because the feedpoint reactance varies with height over ground, and, if it's slightly negative or capacitive, you must give yourself the ability to decrease the spacing without losing gain.

Example 1

Here's one way to tune a DR. Say I want the antenna to be 140 feet long[16] and resonant at 3.520 MHz. I know from the graphs and tables that the minimum height and spacing are 33.5 and 26.6 feet, respectively, in free space, and the center wire input impedance will be 37.8 ohms. The lower wire will be at 46 feet over ground, while the upper will be at about 80 feet.

Just to be on the safe side, I make the antenna 6 inches taller, or 34 feet. Over ground, the feedpoint impedance of this antenna becomes 50.8 + j14.6 ohms. The initial spacing is 26.6 feet, or about 26 feet 7 inches. I then move the center wire to 26 feet 11 inches to compensate for the reactance, and the feedpoint impedance becomes 51 ohms resistive. I now have a perfect match for 50-ohm coax and a gain of 3.44 dBi at a takeoff angle of 18 degrees. You'll note a small discrepancy between this example, which was performed with MININEC, and the same antenna in **Table 8**, which was modeled with NEC-2. The gains are slightly different, as are the spacings. This is of no consequence since, in real life, things will be different anyway. What is important is the "how to" of approaching this problem.

Example 2

Here's another scenario. I'd like to tune the antenna discussed above for 3.550 MHz. If I move the spacing wire or middle vertical to 28.3 feet (28 feet, 4 inches), the antenna will be resonant with an input impedance of 51.3 ohms, so I need to design the antenna for the lowest frequency. This way it will be long and have an inductive reactance at any higher frequency, and I can resonate it simply by moving

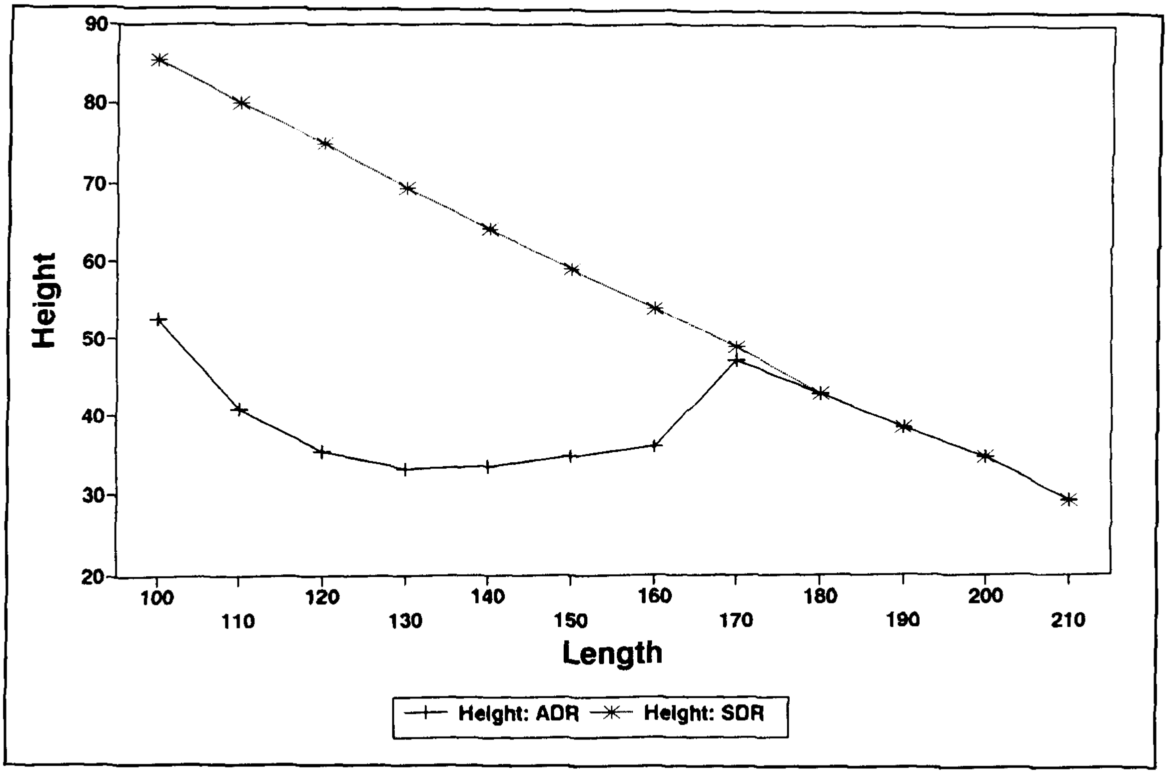

Figure 14. Double rectangle; Height comparison: ADR versus SDR.

the center wire toward the center. I don't want to do the opposite because I may decrease the gain, especially on the shorter antennas of lengths less than 130 feet.

Example 3

Just for the fun of it, I wanted to see how the 140-foot DR with the dimensions of the antenna in **Example 1** would play on 3.8 MHz. Using these dimensions, the input impedance at 3.8 MHz is 40.2 + j353 ohms. A spacing of 38.4 feet makes it resonant with an input impedance of 49.5 ohms. What's interesting about this example is that it's possible to predict where the center wire will fall.

The original dimensions were for a design frequency of 3.520—clearly too large at 3.8 by a factor of 1.08 (3.80/3.52 = 1.08). What, then, is the antenna size which is 1.08 times the size of the original *140-foot version? I'm setting up* an equivalence: a 140-foot antenna, designed for 3.52 but used at 3.80, is equal to what size antenna designed for 3.52 MHz? The answer is 151 feet (1.08 x 140). The spacing for the 150-foot DR is 38.5 feet. Consequently, you should shoot for a match at a point around this spacing. If you plan on using antennas in the linear portion of the spacing curve in **Figure 10** (lengths of 120 to 165 feet), you can estimate the correct spacing directly from the curve. As an added bonus, you'll have the gain of the larger antenna.

In the 1/8-wavelength size, 10-meter model I constructed, I used sections of 1/4-inch aluminum tubing near the area of the horizontals where I expected the center wire to fall, and clamped the center wire to the tubing. Here's how I approached construction of my 140-foot

Table 8 140' Reference ADR at 3.520 MHz							
Length	fs/og	Elev:	Ht:	Sp:	Gain	Angle	R:
140	fs		34	26.6	4.74		37.9
	og	46	34	26.3	3.61	18	49.9
	og	16	34	27.3	3.02	24	76.7
	og	6	34	29.5	2.14	26	101.0

Table 8. 140-foot reference ADR at 3.520 MHz.

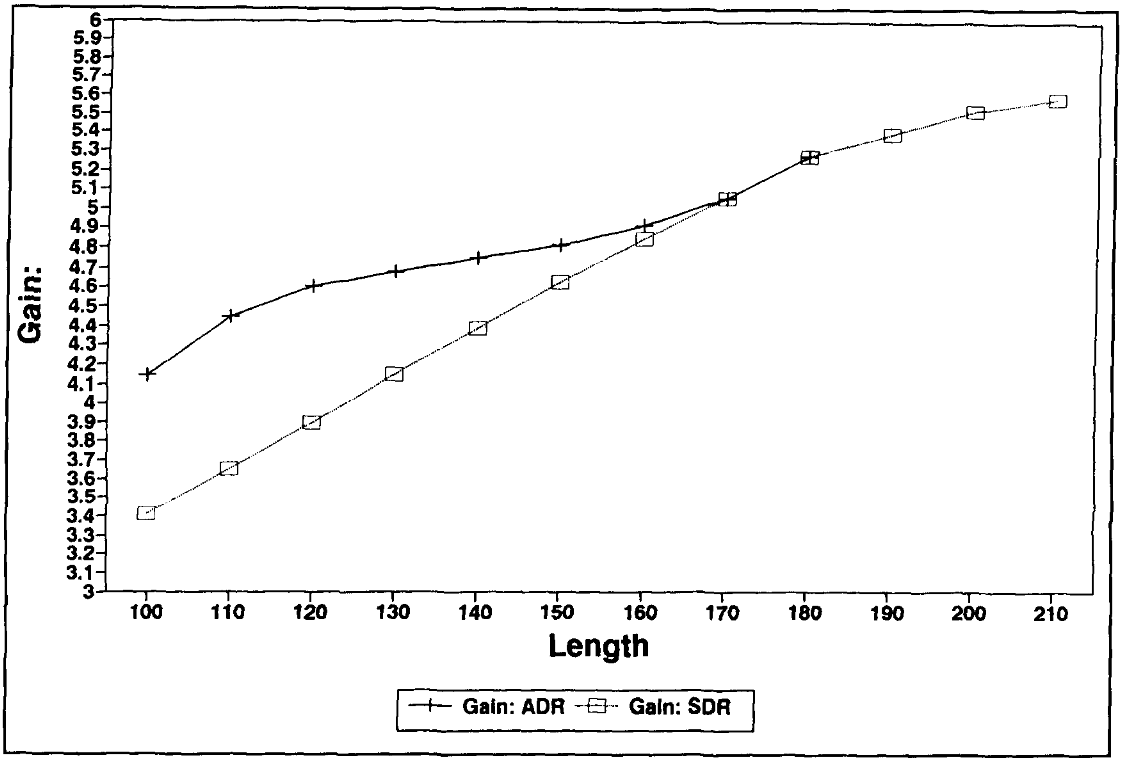

Figure 15. Double rectangle; Gain comparison: ADR and SDR.

ADR. First, as above, I made it 6 inches taller at 34 feet. I then measured off the top and bottom horizontals at 25 feet and terminated them with insulators. I then bridged those points to another set of insulators with 4 feet of bare wire and continued the insulated horizontals. As a result, I bared the horizontals from 25 to 29 feet. I then played with the center vertical, which I connected to the bare horizontals with alligator clips. With an MFJ antenna analyzer feeding the center feedpoint, I moved the center vertical—while lowering and then raising the antenna—until I obtained a feedline match at an SWR of 1.15:1 at 3.520 MHz. In my case, with the lower wire of the antenna at 45 feet, the center wire had to be put at 27 feet 2 inches for the match I desired. At that point, I replaced the bare wire with insulated wires from which I removed the insulation at the points I had measured before, and soldered the center vertical to those points. I then sealed and taped the connections. Of course, you can eliminate all the nonsense with insulators and bridging sections if you elect to make the antenna out of bare wire, say, stranded copperweld.

Antenna height above ground

The advantage of an SCV antenna is that, over real ground, there is useful low-angle gain even at low heights. **Figures 18** and **19** show that, for a 140-foot ADR, the takeoff angle is dependent on the height of the lower wire. As the lower wire drops from 46 to 16 to 6 feet, there's some loss of gain at the angles below 20 degrees. There are wider lobes and less high angle rejection when you lower the antennas, but their ability to provide gain at useful angles for DX is still there. Check **Figure 5** again to see how a lower gain rectangle compares to a dipole. **Table 8** contains the data used to generate the plots shown in **Figures 18** and **19**. The data was generated by NEC-2, as MININEC gives erroneous impedances and VERY optimistic gain figures at heights less than 0.2 wavelengths above ground. You can see that the ADR with a top wire height of 50 and lower wire height of 16 feet still provides substantial low-angle gain. The performance of the one with the upper wire at 40 feet and the lower wire at 6 feet would be marginal when compared to a dipole at 80 feet, but substantially better than a dipole at the ADR height of 40 feet. I wouldn't go below this height because I suspect that the ground losses will mount if the ground is poorer than average[17] and will render the antenna useless.

How high can you go? There's an upper limit above which a detrimental (for DX) high-angle lobe begins to appear.[18] **Figure 20** shows that this secondary lobe makes its appearance at a top/lower wire height of 100/66 feet. The takeoff angle of the main lobe at this height is 16

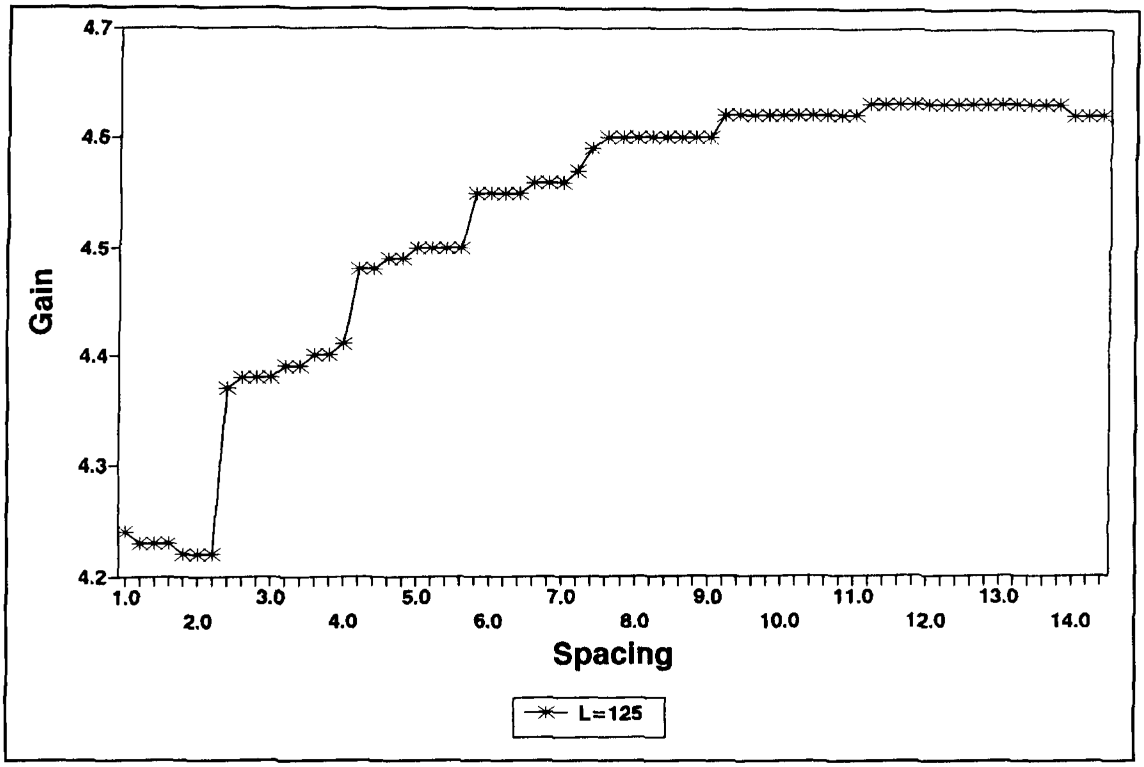

Figure 16. 125-foot ADR: Stepping effect; Gains versus spacing.

degrees. If one goes to a height of 120/86 feet, the secondary lobe becomes prominent and renders the antenna useless from a point of view of high-angle rejection.

Effect of ground on the gain and radiation angle

The antennas were modeled at lengths of 120, 140, 160, 180, and 210 feet and at a lower wire height of 46 feet over average and poor ground to determine what happens to the gain and takeoff angle. The results are shown in **Table 9**. I didn't model poorer terrains, because DXers who live in desert areas or over rocky terrain seem to do better with horizontally polarized antennas.[19]

The antenna over poor ground, as compared with average ground, loses some gain and raises its takeoff angle by two degrees, from 18 to 20 degrees. Overall, this is not significant because the radiation is still superior to any horizontally polarized antenna at the same height. The double rectangles still outperform a dipole at low angles, as shown in **Figure 5** which uses a lower gain simple rectangle for comparison.

Antenna length: tradeoffs

Figure 21 illustrates the tradeoffs between a 140-foot ADR and a 210-foot SDR. Both are at heights of 46 feet above ground for the lower wire. The gain of the longer antenna is achieved with considerable azimuthal narrowing. With MININEC, the gain of the 140-foot ADR is 3.44 with a beamwidth of 71 degrees, while the gain of the 210-foot SDR is 4.34 dBi at 57 degrees. The vertical lobes peak at 18 degrees.

The choice is up to you. My own opinion is that, if the support structures are spaced wide enough apart, and they happen to line up correctly so the main lobe is pointed directly in the preferred DX direction, you can get away with the extra gain and increased side rejection. If, however, the alignment is not quite perfect, then perhaps you might prefer the lower gain antenna because of its wider beamwidth.

Scaling for 40 and 160 meters

The 80-meter dimensions appear to scale quite nicely to the other low bands of 40 and 160 meters. The scale frequencies were 7.000 and 1.800 MHz. They present a double scaling problem because the frequencies corresponding to 3.520 are 7.040 and 1.760. Again, the reference antenna is the 140-foot ADR for 80 meters in **Table 8.**

40 meters

The design goals for the 40-meter band require a range of takeoff angles from 5 to 20

Table 9 Performance Over Ground Antenna Height: Bottom Wire At 46'									
		Average Ground					Poor Ground		
				ADR - fed at center/center					
Length	Zfs	R	X	G	TO Angle	R	X	G	TO Angle
120	69.6	89.6	-6.3	3.40	18	87.3	-6.8	3.06	20
140	40.0	50.6	-4.3	3.61	18	49.3	-4.0	3.26	20
160	35.7	44.3	-5.8	3.80	18	43.2	-5.4	3.46	20
				SDR - fed at center/center					
180	40.3	47.7	-8	4.16	18	46.8	-7.0	3.85	20
				SDR - fed at a lower corner					
210	54.5	66.2	-14	4.61	18	64.5	-13.0	4.25	20
Average Ground: Dielectric Constant = 13 Conductivity = 5 Poor Ground: Dielectric Constant = 13 Conductivity = 2									

Table 9. Performance over ground. Antenna height: bottom wire at 46 feet.

degrees for optimum DX coverage.[20] The ideal antenna would have a symmetrical vertical lobe centered at about 13 degrees. The antenna is a half-sized 140-foot ADR. Its length is 70 feet, height is 17 feet (a bit large; see above), and the spacing for starters is 13.3 feet. You can follow the steps in **Table 10** to see how it scales and tunes. Initially the antenna is a shade too short in free space. Remember that the 3.520-MHz design frequency would scale to 7.040 and the antenna is therefore capacitive at 7.000, so the input impedance is 39.8 - j47. Moving the cen-

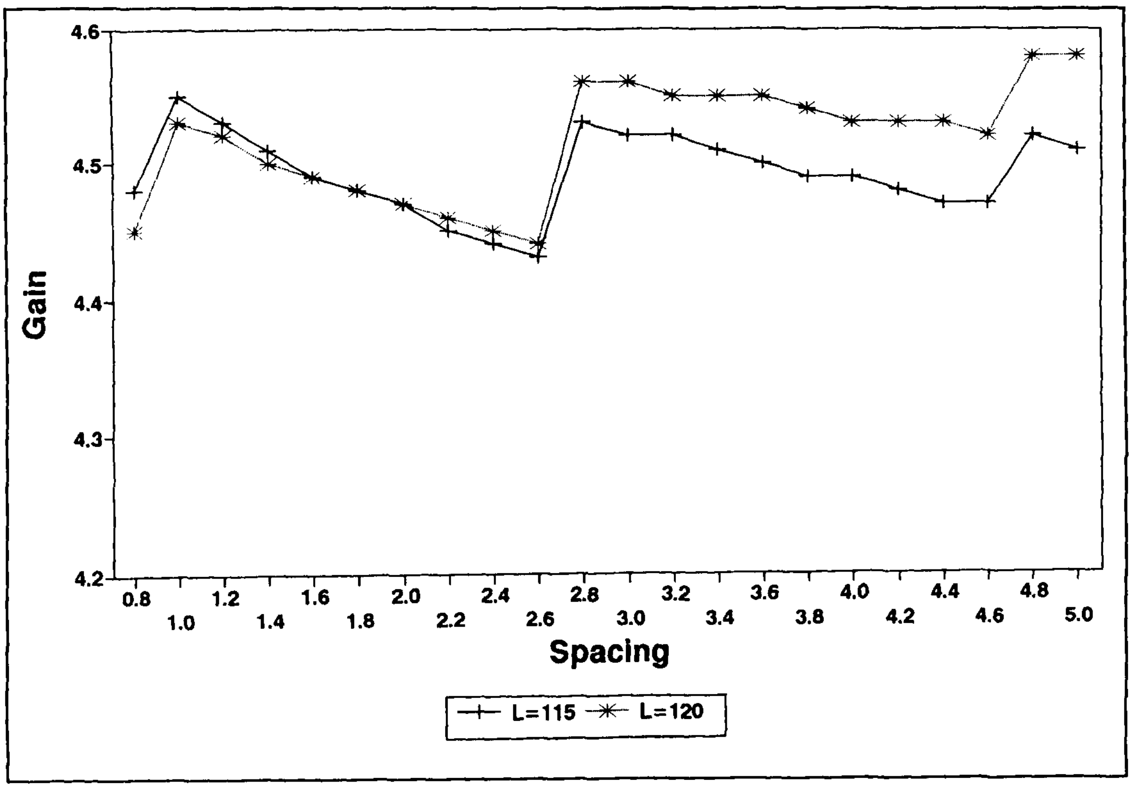

Figure 17. DR: Stepping effect; Gains versus spacing.

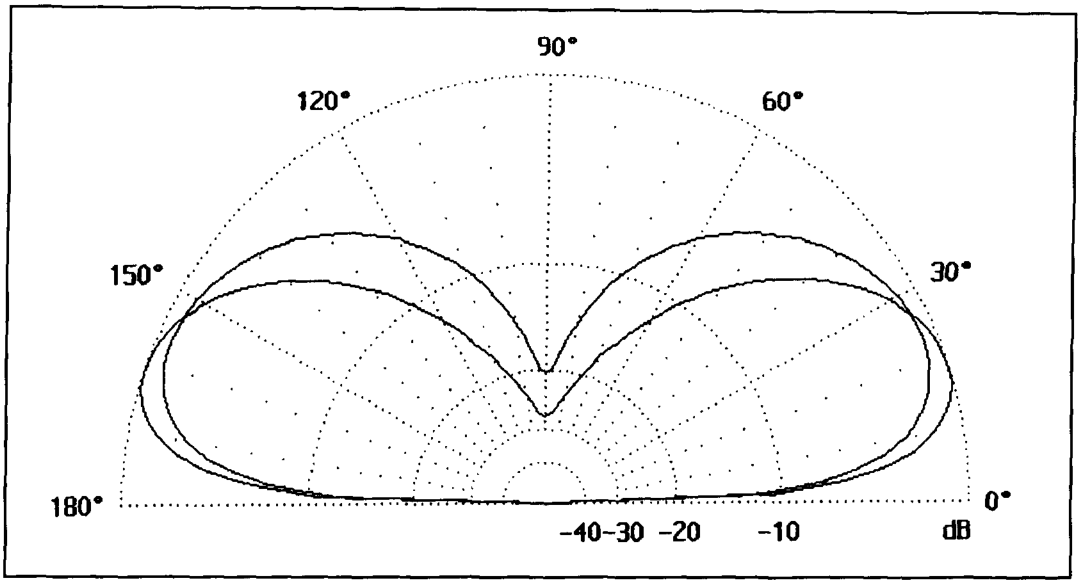

Figure 18. Pattern comparison: 140-foot ADR loops, lower wires at 46 and 16 feet.

ter wire closer to the near end at a spacing of 12.6 feet cancels out the capacitive reactance, and the feedpoint resistance becomes 38.4 ohms. Because we are fairly distant from the near end of the antenna, there's no loss of gain from slightly reducing the spacing; it stays at 4.83 dBi.

At a height of 30 feet above ground for the lower wire, these dimensions yield an antenna that's slightly small; the input impedance is 42.1 - j8.7 ohms. A fractional foot change in the center wire to 12.5 feet cancels out the reactance and the feedpoint resistance becomes 41.8 ohms. The gain is 3.49 dBi at a takeoff angle of 18 degrees.

Because we are fairly well above ground in wavelengths—0.21 to be exact—we expect only a minimal change in the feedpoint impedance if we raise the antenna. This is borne out in **Table 10** by the fact that the feedpoints at lower wire heights of 35 and 40 feet are purely resistive and no changes in spacing are necessary. What we see is that the gain goes up slightly and the takeoff angle goes down to 16 degrees. At the 40-foot height, there is an appearance of a high-angle lobe similar to the one in **Figure 20**, so there's nothing to be gained in raising the antenna further.

160 meters

The 140-foot, 3.520-MHz antenna is doubled in all dimensions. Because the 1.8-MHz target frequency scales to 3.600 MHz, we'd expect the antenna to be long and that the center wire would have to be moved substantially centerward. **Table 11** confirms this with the fact that, in free space, the feedpoint is inductive at 36.9 + j140 ohms. Moving the center wire to 60.6 feet cancels out the reactance, and the feedpoint becomes 40.9 ohms resistive.

We have height problems on 160. The antenna itself is 68 feet high and, if we have 90-foot supports, the lower wire will be at 22 feet. This

Table 10
40 Meter Downscaled ADR at 7.000 MHz

Length	fs/og	Elev:	Ht:	Sp:	Gain	TO Angle	R:	X:
70	fs		17.0	13.3	4.83		39.8	-47.1
	fs		17.0	12.6	4.83		38.4	
	og	30	17.0	12.6	3.49	18	42.1	-8.7
	og	30	17.0	12.5	3.49	18	41.8	
	og	35	17.0	12.5	3.51	16	39.2	
	og	40	17.0	12.5	3.51	16	36.9	

Table 10. 40-meter downscaled ADR at 7.000 MHz.

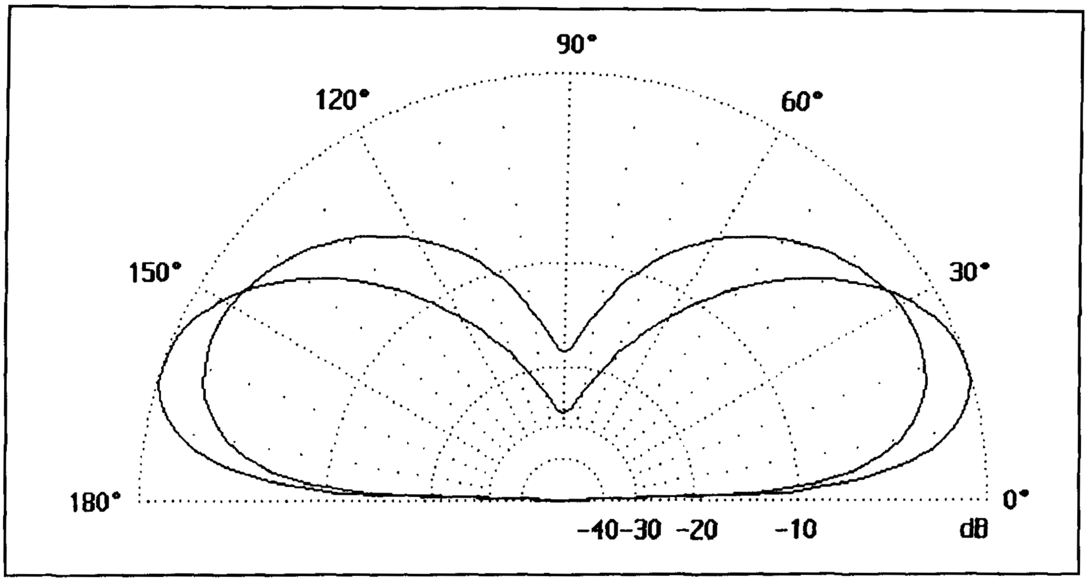

Figure 19. Pattern comparison: 140-foot ADR arrays, lower wires at 46 and 6 feet.

height is only 0.04 wavelength above ground and, as we discussed above, MININEC proved useless. The figures for **Table 11** were all run on NEC-2.[21]

At this height, there are substantial ground effects that make the antenna inductive and which require further movement of the center wire to 63.5 feet to cancel out the reactance. The feedpoint impedance becomes 83.9 ohms and the gain is 3.62 dBi at a takeoff angle of 22 degrees.

If we attempt to lower the antenna further to 12 feet, the center wire must be moved to 66.2 feet. The feedpoint resistance is now 96.2 ohms and the gain is 3.15 dBi at an angle of 24 degrees.

Summary

In essence, the DR on 80 meters lets you choose usable antenna lengths from 100 feet (which is shorter than a dipole) to 210 feet, beyond which there is no increase in gain. Even the shorter antennas, down to 100 feet, have gains in excess of comparable rectangles and have higher feedpoint impedances. While the simple rectangle performs well, the enhanced

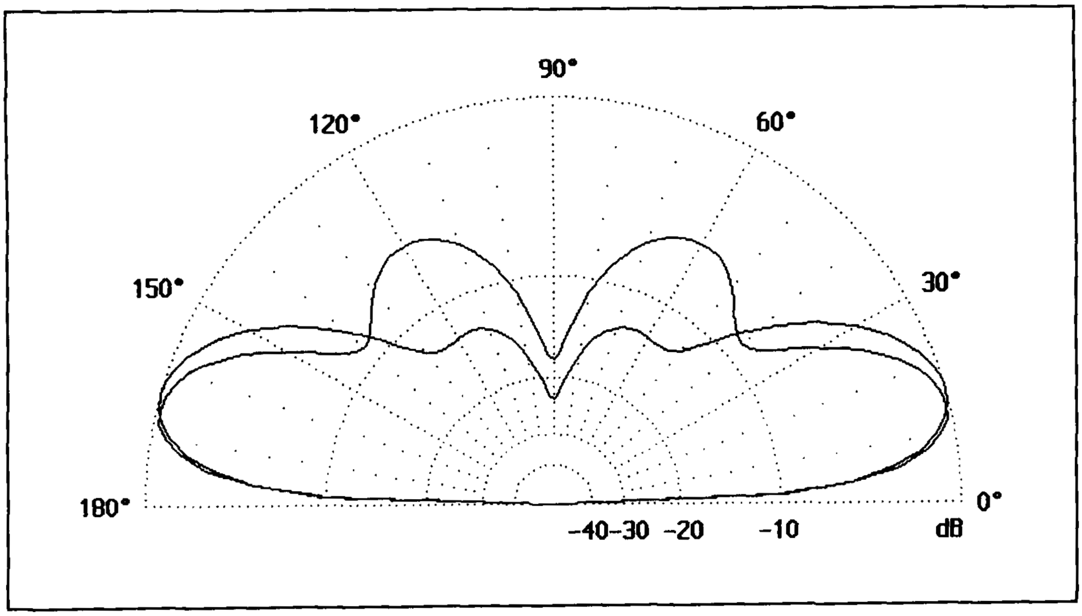

Figure 20. Elevation plot comparison: 140-foot ADRs at 66 and 86-foot heights of lower wires.

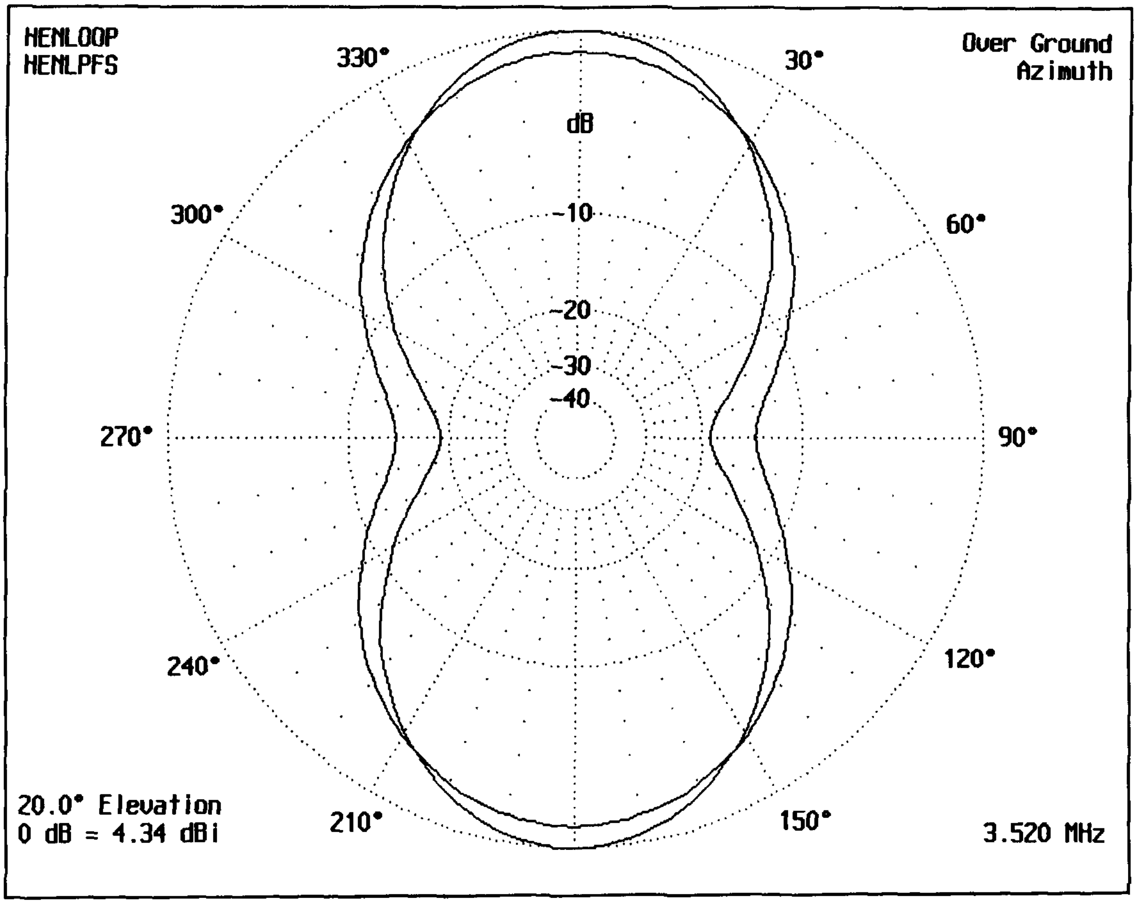

Figure 21. Azimuthal pattern comparison between 140-foot ADR and 210-foot SDR, lower wire height is 46 feet.

rectangle and asymmetrical double rectangles offer more gain at low takeoff angles and better utilize existing space—at a small additional expense in wire. The double rectangle provides more gain than a comparable dipole even when it's much shorter. And, it can be used as a symmetrical double rectangle to provide more gain at lengths from 170 to 210 feet, if space allows.

Furthermore, the DR scales nicely to 40 and 160 meters and has the same advantages as on 80. There are a host of antennas that can be built from the data supplied in this article. One may be just the right candidate for your next 40, 80, or 160-meter project.

REFERENCES/NOTES

1. R. Dean Straw, N6BV, "Ioncap Modeling of Incident Radiation Angles for All Stages of the Sunspot Cycle," *The ARRL Antenna Book*, Volume 15, Table 5, pages 23–25. An antenna that performs at 6 to 31 degrees will cover the entire DX incoming spectrum on 80. The angles from 18 to 31 degrees are best only for the short path W2-to-Europe propagation. Longer paths require takeoff angles of 6 to 18 degrees.

2. L.B. Cebik, W4RNL, "SCV's: A Family Album; Part 1: The Group Picture," *NCJ, September-October* 1998, *pages* 12–16. The four additional parts of this series, one of which is referred to later, will appear in subsequent issues of the *NCJ*. This citation can be read at his web site at: <http://web.utk.edu/~cebik/scv1.html>. The subsequent parts can be reached via this URL.

3. Russel E. Prack, K4RP, "Magnetic Radiators—Low Profile Paired Verticals for HF," *The ARRL Antenna Compendium*, Volume 2, page 39. This is a com-

Table 11
160 Meter Upscaled 140' ADR at 1.800 MHz

Length	fs/og	Elev:	Ht:	Sp:	Gain	TO Angle	R:	X:
280	fs		68	53.2	4.54		36.9	140.0
	fs		68	60.6	4.54		40.9	
	og	22	68	60.6	3.61	22	81.9	48.4
	og	22	68	63.5	3.62	22	83.9	
	og	12	68	66.2	3.15	24	96.2	

Table 11. 160-meter upscaled 140-foot ADR at 1.800 MHz.

plicated antenna involving two parallel loops. It offers a 0.5-dB gain advantage over a single loop and quadruples its feedpoint impedance. The loops must be held in parallel with a trellis. The weight of the trellis, which was made out of pressure-treated wood, when hung at 60 feet, was too much for the weather.

4. Lew Gordon, K4VX, "The Double Magnetic Slot Antenna for 80 Meters," *The ARRL Antenna Compendium*, Volume 4, page 18. This is a double length, when compared to K5RP's, and uses a single loop with a center vertical. He also refers to WØUN's antenna, which is a double/double"—two parallel double loops. This provides marginally more gain at greater complexity.

5. I used Brian Beezley, K6STI's, AO, a MININEC-based program and Nec/Wires, a NEC-2 program for all of the modeling in this article.

6. Shirow Kinoshite, JF6DEA, "The Hentanna—The Japanese 'Miracle Wire' Antenna," *The ARRL Antenna Compendium*, Volume 5, page 66, JH1FCZ was the designer in the 1970s.

7. There are six available feedpoints on a DR: the centers of the two end verticals and the middle element and bases of the same three. Depending on the modeling program you use—MININEC, NEC-2, or NEC-4—the results for the lower corner feedpoint show, respectively, either a few hundredths (3–5) of a dB of loss, about 8 hundredths of a dB of gain, or less than 3 hundredths of a dB of loss as compared to feeding the center of the center vertical. The worst case scenario between any modeling result is a 0.14-dB difference between the highest and lowest figures. In the end, where you place your feedpoint comes down to convenience and impedance matching. Gain is not a factor.

8. L.B. Cebik, W4RNL, MS version of an article in his SCV series to be published in the *NCJ*. It can also be found on his web site at: <http://web.utk.edu/~cebik/svc3.html> as part 3 of "SCV's: A Family Album—The Rectangle."

9. See note 7, above. I left out the data on the centers of the end elements because, if one has to go with an ideal feedpoint, one would go with the center of the center vertical, and these other feedpoints are not favorable from an impedance-matching perspective. This is because the resistance at the center will be too high and the resistance at the far center will be too low. The ideal feedpoints are the center of the center vertical, and, from the standpoint of convenience, the bottom of the center wire and the bottom of the far-end wire. The tradeoffs are in ohms; the base of the center vertical has a higher resistance than the center and the base of the far vertical has a lower resistance. You must pick and choose what's best for your conditions.

10. See notes 7 and 9, above. There may be no tradeoff at all and the gains may be higher in the corners—no one knows. It is impossible to measure such differences, and they should not enter into your decision as to where you feed the antenna. The only statement I would agree with is that the near corner would be very difficult to feed and such a choice is only worthwhile from an educational point of view in this article.

11. Only one set of data for Gain (Gnear) and feedpoint impedance (Znear) is shown for the SDR. Because of its symmetry, the other feedpoint figures will be the same.

12. These are MININEC results, NEC-2 gives the transition point of 120 feet, after which the center wire must begin its movement toward the center. This is of no practical significance. Also, NEC modeling indicates that the gain may actually rise as you decrease the spacing further, but the impedances at such small spacings may be useful only to look at.

13. Cebik, personal communication.

14. I define the ADR or asymmetric double rectangle as any DR with a spacing greater than the 0.9 feet of the ER and less than the length/2 spacing of the SDR.

15. The height leeway with NEC-2 is slightly less—on the order of 3 feet—BUT the gain falls off more gradually, so that even at 6 feet you theoretically lose only 0.02 dB.

16. You probably have noticed that most of the models I use seem to be 140 feet long. This happens to be a half wavelength at 80 meters and is the size of the original Hentenna. This size is marginally larger than that of a dipole, and there's plenty of leeway to make adjustments with the center wire.

17. Cebik, see note 7 above.

18. K4VX, above at page 19. The 100-foot tall version, when compared to the 60-foot version, has slightly more gain below 9 degrees and a narrowing of the lower lobe, but the gain falls off at useful higher angles due to the appearance of the second lobe, which isn't shown.

19. John DeVoldere, ON4UN, "Low-Band DX'ing," 1994, page 8–4, "…signals from area with poor ground conditions (mountainous, desert, etc.) are always generated by horizontal antennas, while from area with fertile good RF ground, we often hear big signals from verticals…."

20. See N6BV above.

21. W4RNL was kind enough to check on my figures with NEC-4, which is a program with limited distribution under a DOD license. I was very suspicious of the gains at very low heights, and still am. The feedpoint impedances and the takeoff angles are in good agreement. At any rate, nobody who has any choice in the matter will use such a low height, but, on 160, who has the choice? Regardless of the gain figures, and even assuming them to be optimistic, we can still say that the antenna will outperform most of the commonly used ones on this band: loaded verticals and inverted Ls. For the curious, here are the differences between my models and W4RNL's.

Medium	Model	Gain	TO angle	Feed Z
Free Space	NEC2	4.54	——	40.9
	NEC4	4.59	——	40.9+j6.6
22' Ave. Gnd	NEC2	3.62	22 deg	83.9
	NEC4	3.70	22	83.3+j5.6
12' Ave. Gnd	NEC2	3.15	24	96.2
	NEC4	3.23	23	95.5+j5.5

PRODUCT INFORMATION

Hitachi's Three-CCD Camera

Hitachi's HV-D15 three-CCD color camera features a single-chip very large scale integration (VLSI) that provides both video processing and encoding and can be remotely controlled from a personal computer.

The 13-bit VLSI digital signal processing (DSP) chip provides higher quality video processing than analog. It incorporates multiple functions like detail, color corrections, noise reduction, color balance, shading, and auto level control. All signal processing from the processor section through the encoder section is accomplished within the single-chip VLSI. The camera's high signal-to-noise ratio (63 dB NTSC, 61 dB PAL) and wide dynamic range are complimented with 10-bit A/D converters as well as its 13-bit internal digital signal processing.

Multiple HV-D15s can be remotely controlled from a PC simultaneously via an RS-232C interface. The camera data can also be transferred to a PC for storage and later recall.

For further flexibility, three application files within the HV-D15 can store user-selected setup information. Switching between application files results in each menu item being reset according to the information previously stored in the file.

For more information on the HV-D15, call (516) 921-7200.

Analog Devices Has 3-Volt, 8-Bit Analog-to-Digital Converter

Analog Devices, Inc. has released the AD9283 100MSPS, 8-bit analog-to-digital converter (ADC). The AD9283 operates at 3 volts and contains an on-board track-and-hold circuit and encode clock.

The encoder input of the AD9283 is TTL/CMOS-compatible and the 8-bit digital outputs can be operated from a +3.0 volt (2.5 to 3.6 volt) supply, allowing users the flexibility to interface with 2.5-volt digital ASICs. A power-down function may be exercised to bring total consumption down to a mere 4.2 mW (in power-down mode, the digital outputs are driven to high-impedance state).

For more information, contact Analog Devices, Inc., Ray Stata Technology Center, 804 Woburn Street, Wilmington, MA 01887; Phone: (781) 937-1428. Visit their Web site at: <http://www.analog.com>.

Edited by Peter Bertini, K1ZJH
Senior Technical Editor

This issue brings an update on solar cycle 23 by Wil Anderson, AA6DD, who first advanced his own predictions for the cycle at this time last year. You'll also find a nifty little squelch for the MFJ-9406 transceiver from Rick Littlefield, K1BQT. Enjoy!

—de K1ZJH

Update on Solar Cycle 23

The continuing story of an interesting cycle

Wil Anderson, AA6DD

Solar cycle 23 continues on a path that agrees with the graphical solution prediction from my article in the winter 1998 issue of *Communications Quarterly* entitled "Predictions for Solar Cycle 23." The 12-month smooth sunspot numbers are on an upward trend (**Table 1**), which will probably lead to the highest sunspot cycle in the 300-year history of solar data. The global warming from very high levels of ultraviolet light radiation from the sun may create problems for sun bathers, but, for hams, it means four years of worldwide of DX on 6 and 10 meters. The monthly sunspot average is now over 100, so 10 meters is now a reliable *F* layer propagation band and 6 meters will experience sporadic *Es*, sporadic F_2 layer, and transequatorial propagation. When the monthly sunspot average reaches a level of 150, 6 meters will become a reliable *F* layer propagation band!

Squelch for the MFJ-9406 Transceiver

Try this simple add-on module

Rick Littlefield, K1BQT

While squelch isn't essential for VHF SSB operation, it takes the "acoustic fatigue factor" out of waiting for band openings. This simple add-on squelch module installs directly above the MFJ-9406 volume control and connects to three easy-to-identify plate-through holes on

Table 1. Smoothed monthly sunspot numbers measured and predicted.

Year	Jan	Feb	Mar	Apr	May	Jun	Jul	Aug	Sep	Oct	Nov	Dec
1996												
Msrd											9.8	10.4
Pred											8	9
1997												
Msrd	10.4	11	13.5	16.5	18.3	20.3	22.6	25.1	28.4	31.9	35.1	39.1
Pred	9	11	12	13	13	15	17	19	21	25	28	34
1998												
Msrd	43.8	48.9	53.4	56.6	59.3	62.4*						
Pred	40	45	52	61	69	78	86	94	104	114	124	132
1999												
Msrd												
Pred	142	151	155	159	161	165	169	175	177	179	186	191
2000												
Msrd												
Pred	193	196	199	202	205	206	207	208	207	206	203	196
2001												
Msrd												
Pred	192	190	190	189	187	186	183	183	181	180	174	170
2002												
Msrd												
Pred	166	161	156	151	146	142	138	133	130	127	126	122

**Estimated.*

Note: The peak can easily run six months behind the prediction, so don't despair!

Parts List

C1	10-μF electrolytic capacitor
R1	22-k resistor
R2	10-k linear pot, 16 or 17-mm miniature chassis mount
R3	33-k resistor
R4	100-k resistor
R5	100-ohm resistor
R6	10-k resistor
Q1	2N7000 switching FET
U1	741 op-amp
D1	1N914 or 1N4148 switching diode

the transceiver's main pc board. Once installed, you can leave your radio on silent sentry duty until something interesting breaks the quiet. The module works equally well with MFJ-9410, MFJ-9406, and MFJ-9402 SSB transceivers.

Description

Referring to the schematic in **Figure 1**, the squelch circuit consists of a simple op-amp comparator (U1) and FET switch (Q1). The inverting input samples a manually adjusted reference level supplied by the squelch control's voltage divider. The non-inverting input samples the receiver's AGC voltage picked up at the radio's RSSI test point. When no signal is present and the squelch is set, U1 output goes high, setting switching-FET Q1 into conduction. This, in turn, pulls the receiver's mute line low to turn off the AF amplifier chip. When a signal appears, or when the squelch control is opened manually, U1 output goes low to set Q1 into non-conduction. This allows the receiver's mute line to restore bias and turn on the AF amplifier chip. Stable Vcc for the squelch module is supplied by the transceiver's 10.25-volt regulator circuit.

Construction and Installation

The pc board is a simple, single-sided layout using readily available parts (see the **Parts List**). Printed circuit artwork is provided in **Figure 2**. If you prefer, pc boards may be purchased from FAR Circuits.* Note that two sets of pads are provided for squelch control R2. The type of miniature squelch pot you use determines which set of holes you select. If your pot has front-mounted tabs, use the front holes. If it has rear-mounted tabs, use the rear holes. When installing components, refer to the parts-placement diagram shown in **Figure 3**.

When drilling the front panel, follow the detail provided in **Figure 4A**. Labeling always

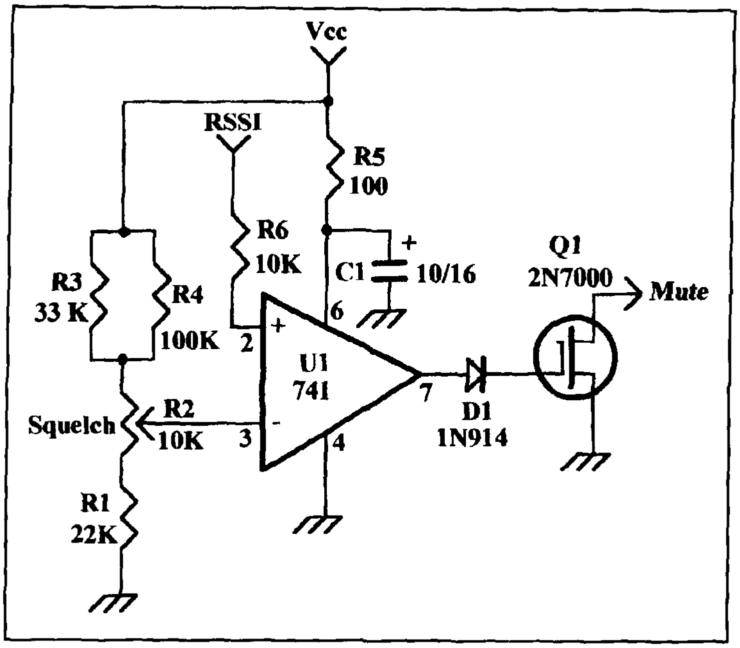

Figure 1. Squelch module schematic.

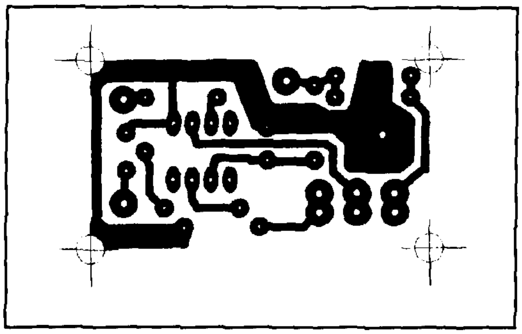

Figure 2. Printed circuit board art.

poses an aesthetics problem when adding controls to commercial equipment. If you can find it, ChartPak™ 10-point Helvetica rub-on lettering provides a reasonably close match to the standard MFJ cabinet lettering.

To mount the module, secure it behind the radio's front panel via the squelch control's shaft bushing. For a second mounting point, replace the main-board mounting nut below the module with a 6-32 threaded stand-off cut to the required length. As an alternative, you could fabricate a simple aluminum mounting bracket and install it with 6-32 hardware. The stand-off (or bracket) also serves as the module's ground return. If you omit this support, be sure to add a ground lead. To interface the module, connect the RSSI, Mute, and Vcc wires to the plate-through holes indicated in **Figure 4B** (*mute* and *RSSI* points are labeled

*Fred Reimers, FAR Circuit, 18N640 Field Court, Dundee, Illinois 60118.

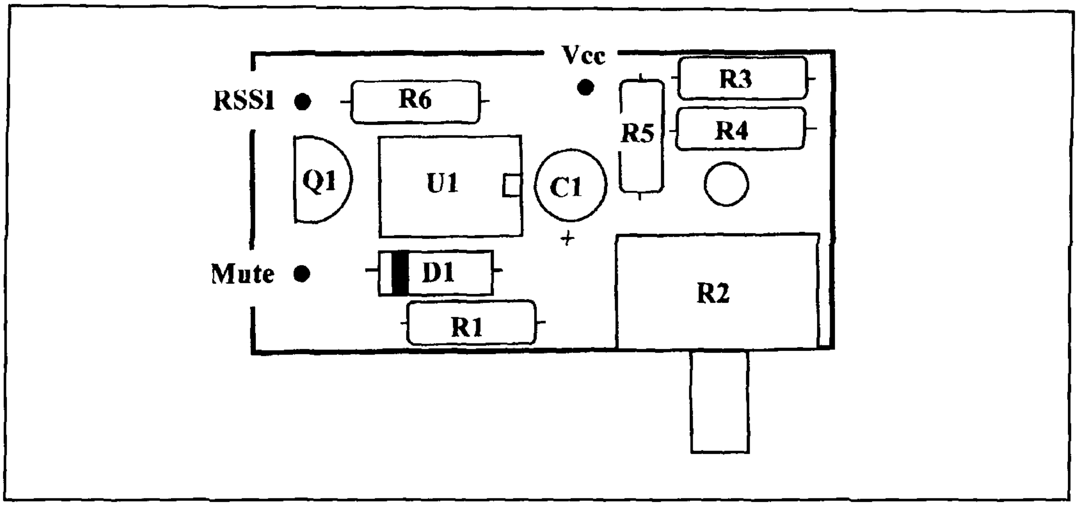

Figure 3. Parts placement diagram.

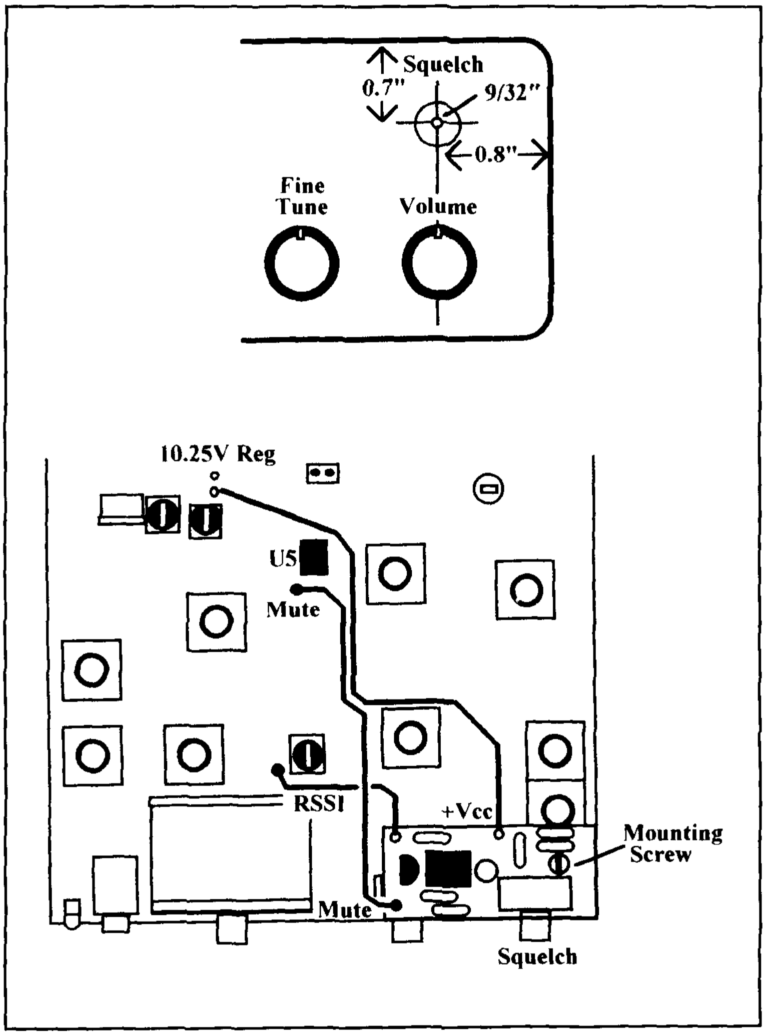

Figure 4. (A) Module connections and (B) panel drilling.

on surface-mount MFJ-9406 pc boards). You should be able to solder these wires in place on the top side of the board, making disassembly of the radio unnecessary.

Operation

Make sure the squelch is working before buttoning up your installation. Control operation is conventional: clockwise for open and counterclockwise for closed. If you plan to use an external receive preamp, confirm that R2 provides sufficient range to cut off the receiver with your preamp on or off. You may move R2's range slightly to reposition the squelch cutoff point by altering the value of R4. Once set, the comparator triggers with only minor changes in AGC level. This characteristic allows even very weak or off-frequency SSB signals to open the squelch. If you "eyeball" the analog tuning dial onto the desired calling frequency, the receiver should detect activity—even if tuning is off by a kHz or two.

Conclusion

Being a laissez-faire DXer, I don't follow the daily index, monitor foreign video, or watch the packet cluster for signs of impending activity. I *do*, however, frequently park the MFJ-9406 on the calling frequency and set the squelch to keep the radio silent while I'm busy doing other things. Here in New England, when the band starts to open up, the calling channels get noisy fast! I owe several 50-MHz DX countries and difficult-to-work U.S. grids to this monitoring approach. ■

Grant Bingeman, *KM5KG*
E-mail: <DrBingo@compuserve.com>

NEGATIVE RESISTANCE ANTENNA ELEMENTS

Obtaining the desired performance

One antenna may have many radiating elements, and its designer may choose to drive more than one of those elements. When this is the case, it's possible for a negative driven element to exist. A careful designer needs to know how to treat negative elements if he expects to obtain the desired antenna performance.

An individual element in a multi-element, multi-source antenna may have a negative input resistance, which should actually be treated as an output. A thorough understanding of negative antenna elements is necessary for successful design and adjustment of phased arrays. The concept of negative resistance is somewhat arcane, but if you consider a negative resistor simply as a source instead of a load, you'll be starting on the right foot.

When an antenna has more than one driven element, it's possible to have more RF power in one element than is actually delivered to the antenna. In this case, the remaining elements have a negative power equivalent to the difference between the total antenna input power and the power in the positive-resistance elements. In other words, it's possible for operating impedances in a phased array to have a negative resistance. This means the negative element is receiving power from the other elements, rather than radiating power. I guess we could call this negative resistance element a collector instead of a radiator.

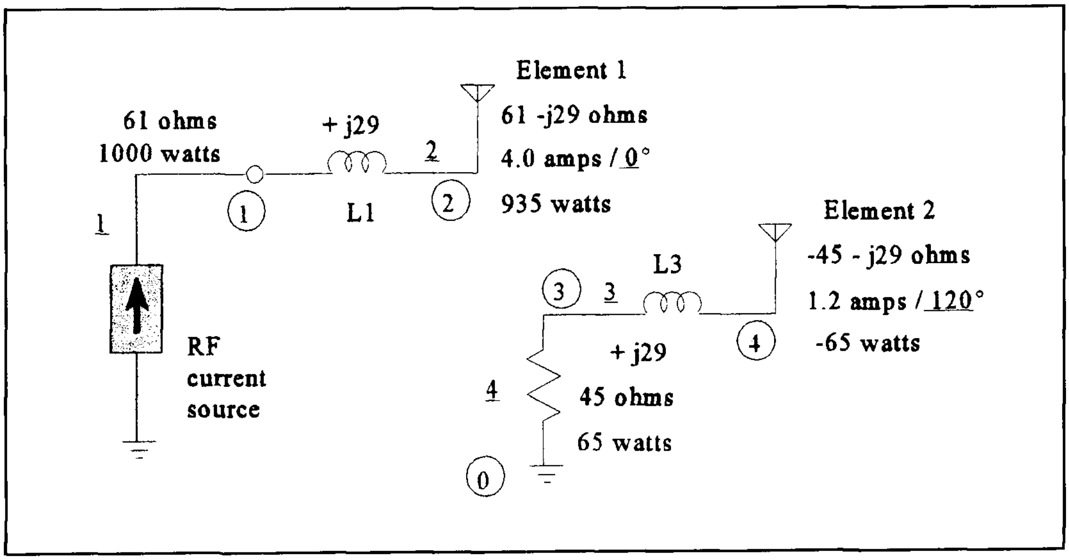

Figure 1. Terminating a negative element with a resistor.

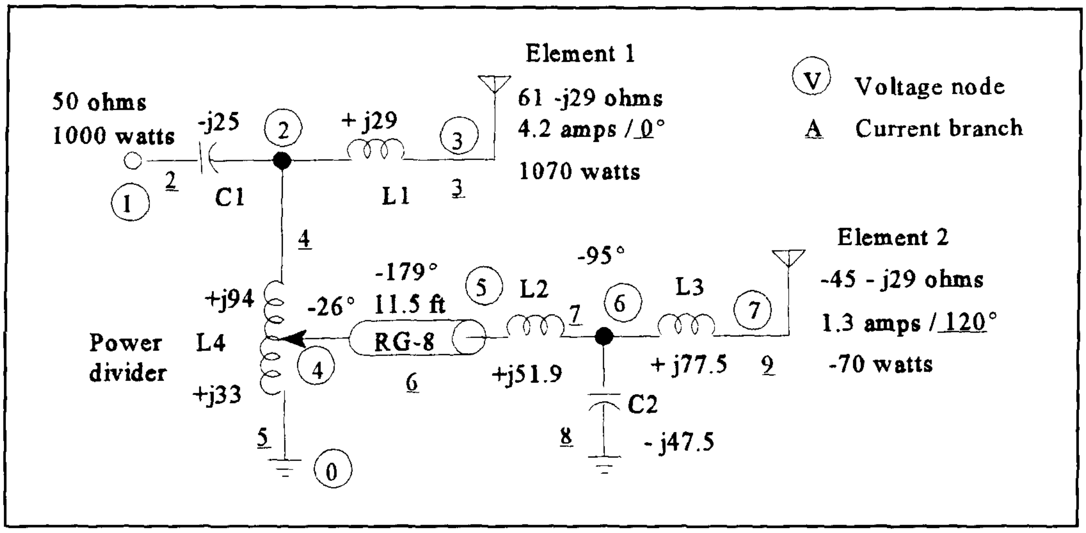

Figure 2. Feeder network for phased array with negative resistance.

There are two ways to treat a negative radiator: you can tune out the reactance of the negative element (resonate it) and terminate with a resistor whose value is equal to the absolute value of the operating resistance (**Figure 1**), or you can match the operating impedance of the negative element to a transmission line and bring the power back to the power divider (**Figure 2**).

This article will show you how to create a negative resistance in a 10-meter ham band antenna consisting of two vertical monopoles erected over an abbreviated ground system. You can build this array and measure its impedance and currents to verify the conclusions presented here. In doing so, you'll learn how to match the impedance of a negative resistance element and how to obtain a low VSWR on the transmission line connected to that element. Negative operating resistances can occur in almost any type of antenna where more than one of its elements are driven.[1]

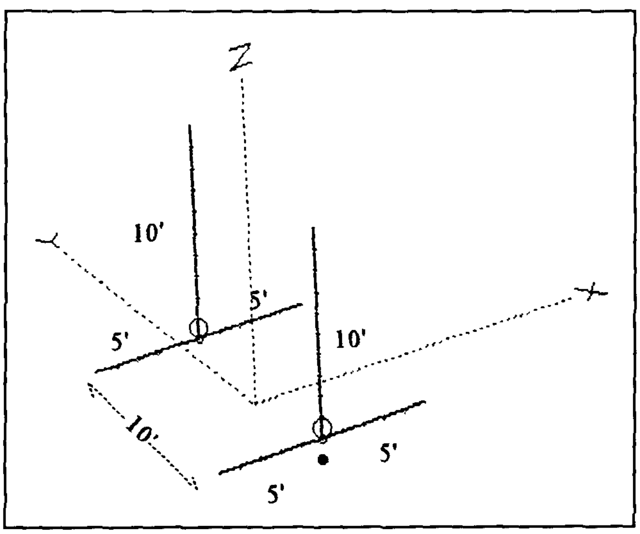

Figure 3. Each monopole is mounted over a two-wire, five-foot radius counterpoise raised a foot above ground.

Negative resistances don't often appear in amateur radio antennas because ham antennas typically consist of only one driven element, while the remaining elements are parasitic and passively tuned.

The example antenna chosen for this article is a pair of 10-foot monopoles spaced 10 feet apart. This is considered a two-element antenna. Each monopole is mounted over a two-wire, five-foot-radius counterpoise raised a foot off the ground, as shown in **Figure 3**. All materials are 1/2-inch type M copper pipe, with an outside diameter of 0.6 inches. The operating frequency is 28 MHz, which has a wavelength of about 35 feet in free space. Thus, the monopoles are about 103 degrees long and the counterpoise wires are about 51 degrees long physically. These wires may look a bit longer electrically, but that subject is beyond the scope of this article.

The example antenna wasn't chosen for its radiation pattern, but rather to highlight the effects of a negative element and the methods used to deal with negative power. In general, antennas with negative elements may tend to have a relatively narrower bandwidth because they have higher circulating currents than antennas that can produce the same pattern without a negative element. Typically, the closer the elements are spaced in an antenna, the greater the likelihood of a negative operating resistance when more than one element is driven. If only one element is driven, you can never have a negative input resistance. If your antenna analysis program tells you that you have a negative input resistance when only one element in your antenna is driven, there's either a mistake in your antenna model or an error in the program.

I used the EZNEC 2.0 method of moments analysis software. The length of the current segments was one foot, the ground conductivity was 5 mS/m, and the dielectric constant of the earth was 13. The self-impedance of these copper monopoles is 56.0-j36.1 ohms, and the mutual impedance between them is 13.0-27.2 ohms (30.1/-64.5°). We could tune out the capacitive self-reactance (the -36.1 ohms) by increasing the counterpoise pipe length to about six feet. However, because the standard length of half-inch copper water pipe in my locale was 10 feet, this wasn't convenient. Besides, I like to keep my antenna input reactance a bit negative because series resonating with a coil is easier than using a capacitor.

In the context of this antenna, the terms pipe, wire, and tubing are synonymous. The schematic diagrams are labeled with voltage nodes (circled) and current branches (underlined). To keep things simple, all discrete network components are assumed to be lossless, unless noted otherwise.

The two-port model of the coupled antenna elements looks like the tee network in **Figure 4**. Sometimes the resistances in this model will be negative. Many canned network analysis programs won't accept negative resistor values. My own program, written 20 years ago, doesn't have this problem.*

We can use the **Figure 5** antenna impedance model in a network analysis program to determine what goes on when we vary the transmission line length, or adjust any of the capacitors and inductors in the feeder system. Remember that what we seek is a low VSWR on the transmission line and the desired relative currents in the antenna elements. When your antenna contains a negative element, intuition can lead to some very inaccurate designs, so a good network analysis program can save you a lot of grief. It's a brute-force means to check your feeder network design. My earlier treatment of phased arrays in the Summer 1998 issue of *Communications Quarterly*[1] contains a detailed description of multi-port antenna modeling when more than two elements are involved.

Where:

Z11 = self-impedance of element 1
Z22 = self-impedance of element 2
Z12 = mutual impedance between both elements
Z1 = operating input impedance to element 1
Z2 = operating input impedance to element 2

In this case Z11 = Z22, since the antenna array is bilaterally symmetrical:

$$Z1 = Z11 + Z12\,(I2/I1) \qquad (1)$$

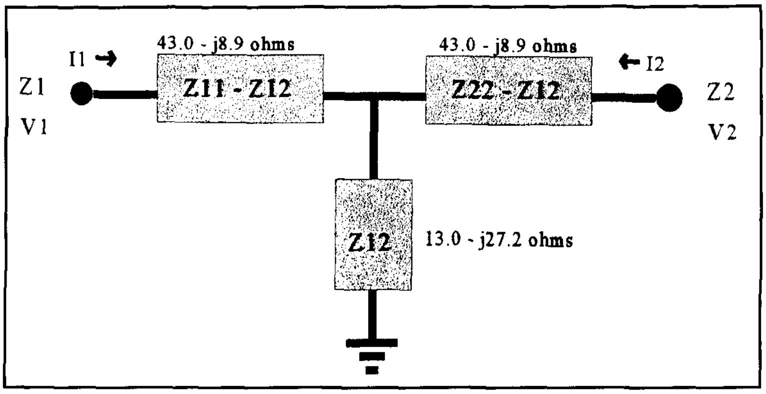

Figure 4. Coupled elements impedance model.

$$Z2 = Z22 + Z12\,(I1/I2) \qquad (2)$$

$$\frac{|I2|}{|I1|} = \frac{|Z12|\cos(\varphi1 - \varphi2 + \varphi12)}{-R22} \qquad (3)$$

Where:

$\varphi1$ = element 1 current phase
$\varphi2$ = element 2 current phase
$\varphi12$ = phase of mutual impedance between elements 1 and 2

In general, the closer the radiating elements are spaced electrically, the more likely you are to have a negative operating resistance when more than one element is driven. This is because the magnitude of the mutual impedance is greater at closer spacings. A larger mutual impedance is more likely to overwhelm the self impedance per **Equations 1** and **2**, when the resistance of the second term of the equations is negative. Of course, this depends on the phase of the element currents and the phase of the mutual impedance.

It should be clear upon inspecting the equations that if the current magnitude ratio is large, this will also make the second term larger. Thus, it's a simple matter to find what current ratio will create a negative operating resistance if you know the self and mutual impedances. In our example antenna, when the Element Two current leads the Element One current by 120 degrees, we have a negative operating resistance for Element Two when the ratio of the element current magnitudes (I2/I1) is greater than 0.54 (**Equation 3**). The general operating impedance equation for multiple elements is described by **Equation 4**.

$$Zi = \Sigma(Ij/Ii)Zij, \text{ for } j=1 \text{ to } n \qquad (4)$$

where

Zi = operating impedance of element i
Ii = complex current in element i

*A copy of the network analysis program used in this article, which can model transformers, lossy transmission lines, negative resistances, lossy inductors and capacitors, and transfer functions, is available for $20 from: LYSIS2, 1908 Paris Avenue, Plano, Texas 75025.

Ij = complex current in element j
Zij = mutual impedance between elements i and j
n = number of elements in the antenna

Equation 5 is the correct formula for determining suitable phase shifts for the transmission lines and networks when Element Two has a negative resistance; **Equation 6** is the formula to use when no negative resistances are involved. Note that they are quite different from each other.

$$\text{negative R: } \theta 1 - \varphi 1 + \varphi 2 + \theta 2^\circ = 180^\circ \quad (5)$$

$$\text{positive R: } \theta 1 - \varphi 1 + \varphi 2 - \theta 2^\circ = 0^\circ \quad (6)$$

where:
$\theta 1$ = the total phase shift of all networks and lines from the common point to Element One
$\theta 2$ = the total phase shift of all networks and lines from the common point to Element Two
$\varphi 1$ = the desired phase of the current in Element One
$\varphi 2$ = the desired phase of the current in Element Two

The specific phase shift numbers for **Figure 2** then become 0-0+120-300=-180 degrees. Of course +180 degrees is the same as –180 degrees in the context of this article. In steady-state systems you can always add or subtract 360 degrees from a phase shift value because the waveform repeats every wavelength.

The feeder system consists of a power divider, a transmission line, and an impedance-matching/phasing network (**Figure 2**). L1 simply resonates the first monopole, and C1 resonates the input to the system. The currents in the two monopoles are unequal in magnitude and have a relative phase of 120 degrees. **Table 1** shows what happens to the input impedance to each monopole as the magnitude ratio of the radiated currents (I9/I3) is varied, and the relative phase is maintained at 120 degrees. The power in each monopole is based on a fixed input power of 1000 watts at the overall input to the feeder networks.

Note that Element Two becomes parasitic (its operating resistance becomes zero) when its current is about 54 percent of Element One's current. In other words, we could tune Element Two as a parasitic reflector if we placed a 32-ohm inductor between its base and its counterpoise (**Figure 6**). But let's push a bit beyond the parasitic case and choose a current ratio of 30 percent, where we have 1070 watts in Element One and –70 watts in Element Two. The vertical pattern is mildly directional (**Figure 7A**).

What happens to that pattern if we drive only Element One and terminate Element Two in 45 + j29 ohms per **Figure 1**? We get the same pattern shape, but it's slightly smaller in size (**Figure 7B**). The maximum gain is reduced by 0.3 dB because of losses in the resistor. We have 1000 watts in Element One, and 66 watts lost in our 45-ohm resistor. The current in the base of Element One is 4.0 amperes at zero degrees, while the current in the base of Element Two is 1.2 amperes at 120 degrees (see **Table 2**, in the **Appendix**). The element currents are slightly lower than those of the ideal situation shown in **Figure 2**.

Table 2 proves that passively terminating a negative antenna element in the negative of its operating impedance is a viable solution if some loss in radiated power can be tolerated. This method is certainly the simplest way to deal with a negative element in a phased array, and it deserves serious consideration because it is much easier to adjust for the desired pattern. In fact, you are well advised to begin tune-up of an array having negative elements with a passive termination on each negative element, even if you plan to graduate to the more complex configuration typified by **Figure 2**.

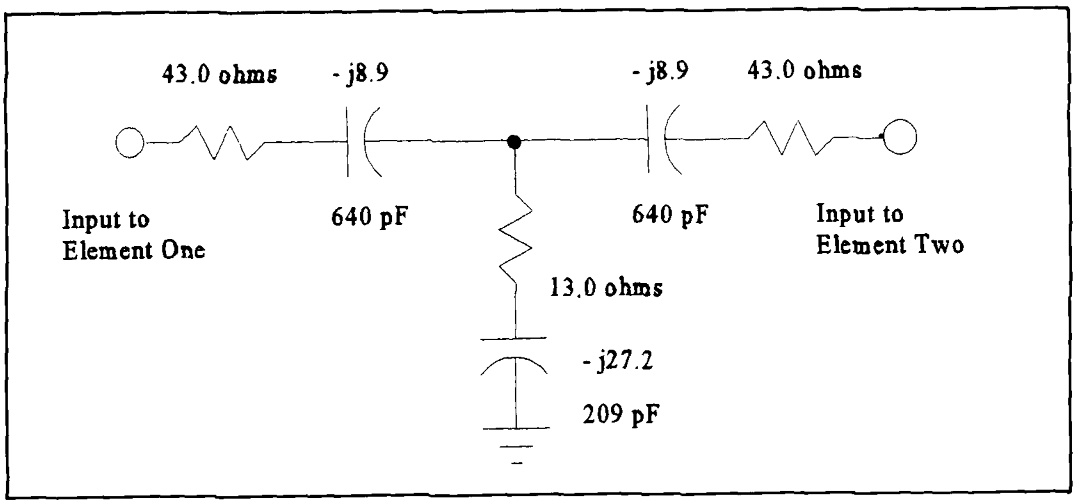

Figure 5. 28-MHz coupled elements impedance model.

Table 1.

19/13 Mag	Phase degrees	Z1 ohms	P1 watts	Z2 ohms	P2 watts
0.3	120	61.1-j28.6	1070	-44.7-j28.9	-70
0.4	120	62.9-j26.1	1052	-19.5-j30.7	-52
0.5	120	64.5-j23.6	1017	-4.4-j31.8	-17
0.6	120	66.2-j21.0	970	5.7-j32.5	30
0.7	120	67.9-j18.3	915	12.9-j33.0	85
0.8	120	69.6-j16.0	856	18.3-j33.4	144
0.9	120	7.13-j13.5	797	22.5-j33.7	203
1.0	120	73.0-j11.0	739	25.8-j34.0	261

Table 1. Monopole pair feedpoint parameters for various current ratios.

We can eliminate this 0.3-dB loss associated with **Figure 1** by replacing Element Two's terminating resistor with a tee network to match the operating impedance to the 50 ohms of a transmission line brought back to a power divider per **Figure 2**. Note that Element One's reactance is tuned out with inductor L1. Element Two's reactance is tuned out by L3, plus a little more inductance is added to L1 as part of the tee network to match the 45 ohms up to 50 ohms at a specific phase shift. This 50 ohms is then presented at the other end of the transmission line to the power divider, L4, which is a tapped coil.

The circled numbers in the various network figures represent the voltage nodes as listed in the network analysis output tables (**Tables 2** through **8**) in the **Appendix**. Node zero is ground. The underlined numbers represent the individual component currents—starting with the current source driving the common point of the networks, which is branch number one. In all the network analysis print-outs listed in the **Appendix**, branch number 1 is the RF current generator or transmitter, connected to node 1.

We need to pay close attention to the phase shifts across the transmission line and the various networks in order to bring the negative power back at the correct phase at the top of the power divider. Note that the total phase shift from the top of the power divider to Element One is zero degrees in **Figure 2**. At first glance, you might assume that you need +120 or –240 degrees of phase shift in the feeder branch to Element Two. However, if we change the **Figure 2** tee network phase shift from –95 to –35 degrees, the results are rather poor (**Figure 8, Table 3**). The relative current in Element Two is 0.6 at 99 degrees, which is a far cry from the desired 0.3 at 120 degrees. The operating impedances are also radically different from the expected values because they are a

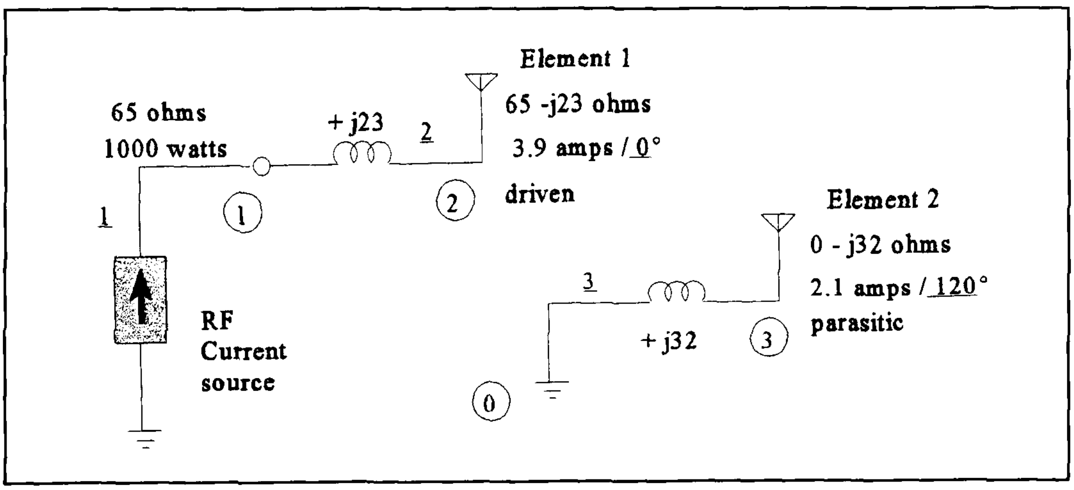

Figure 6. Parasitic array.

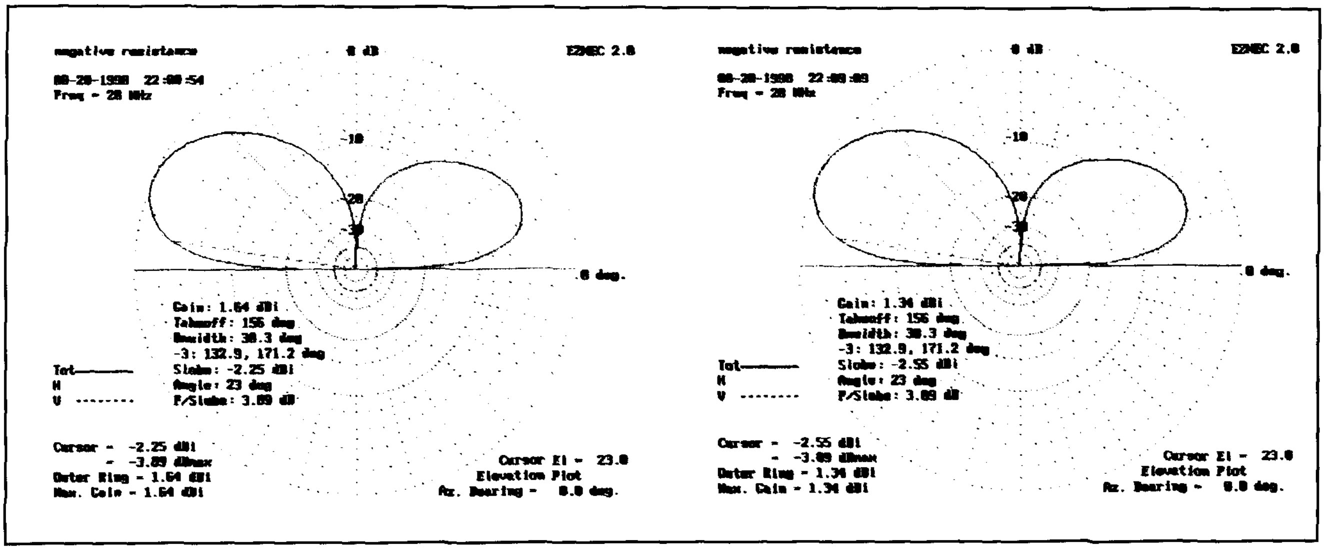

Figure 7. (A) Vertical pattern. (B) Pattern when driving Element One and terminating Element Two in 45+j29 ohms.

function of the element currents per **Equation 4**. And the actual phase shifts across the RG-8 and the networks also differ from the expected values because nothing within the feeder system is matched anymore. So what we thought was a –35-degree tee network, for example, is actually something else.

Referring to **Equation 5**, the correct design phase shift from the top of the power divider to Element Two is about –300 degrees, which is the same as +60 degrees. This 60 degree phase shift consists of about –34 degrees across the power divider, -171 degrees across the transmission line (velocity factor = 0.66), and –95 degrees across the tee network. We want a relative phase between element currents of 120 degrees, which is the same as –240 degrees. According to my lumped-parameter network analysis program, the network of **Figure 2** attached to the antenna impedance model of **Figure 4** produces the desired currents in the two radiating elements with low VSWR on the transmission line (**Table 4**).

As you can see in **Table 4**, the input impedance to the feeder network of **Figure 2** is quite good, about 50-j1 ohms. The RG-8 impedance match is also good, about –47 + j5 ohms at either end. Don't let the negative resistance bother you; it simply means that power is flowing from right to left instead of the conventional left to right on the drawing.

Table 4 verifies **Equation 5**, even though intuition suggests that the relative phase shift between element currents produced by the networks of **Figure 2** should be 60 instead of 120 degrees. So what's really happening here? Let's take another look at the phase shift selection equations, and compare the negative resistance case (**Equation 5**) to the positive case (**Equation 6**). For the case of a negative

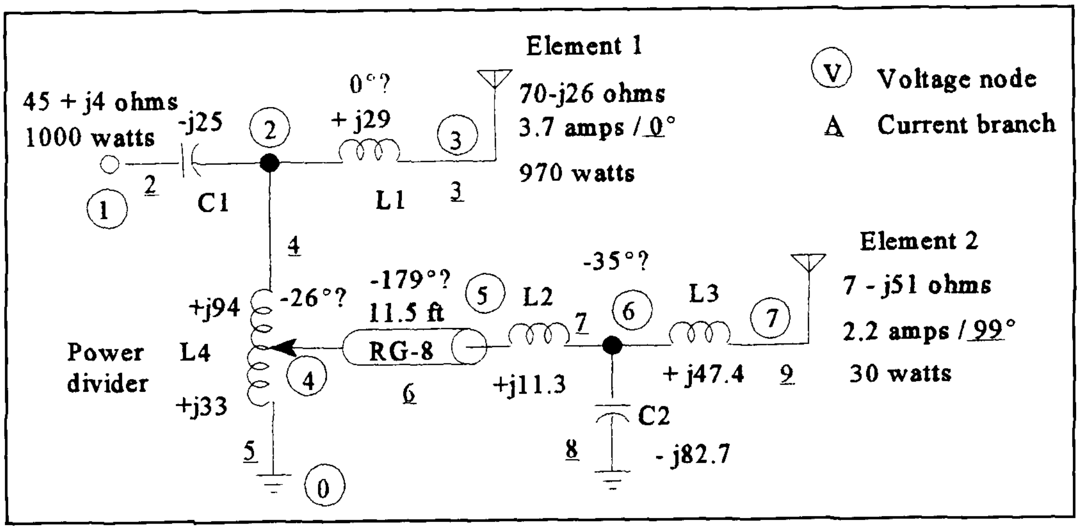

Figure 8. Feeder network for phased array with negative resistance.

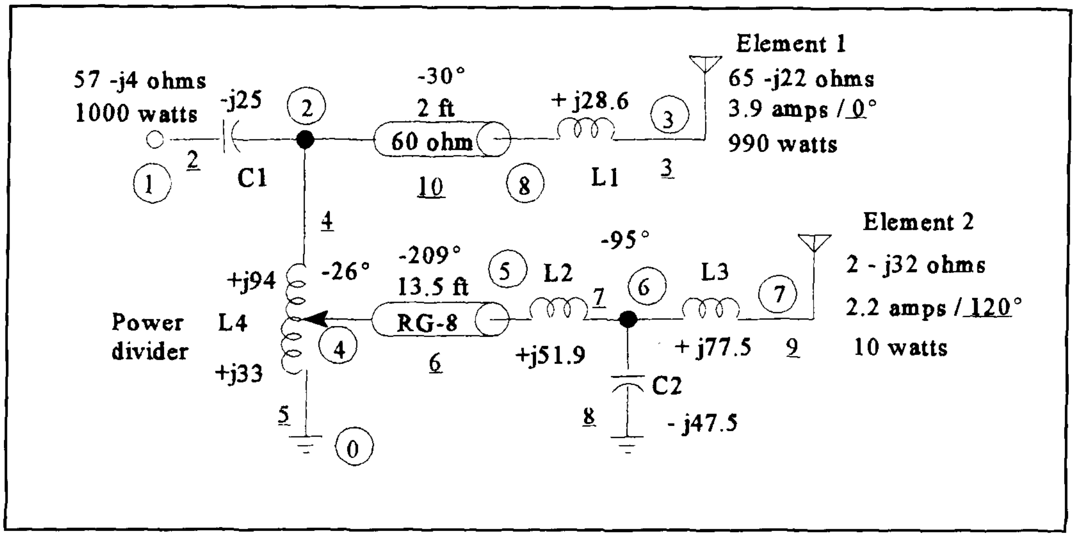

Figure 9. Feeder network with additional line lengths.

resistance, we change the sign of the negative element's feeder phase shift (θ2) to account for the fact that the power is flowing "backwards." The 180-degree term is added because we want the returned power to subtract from the forward power to produce a net input power of 1000 watts. This is similar to the way you read a reflectometer. In physics, energy is conserved, and you don't get something for nothing. So we're not putting in 1000 watts and getting out 1070 watts. If you add up the power in the **Figure 1** antenna model resistors, you'll see in all cases that it comes to 1000 watts (see **Appendix**).

As another test of our phase shift selection method, let's add equal amounts of phase shift to each element's feeder branch. Intuition tells us that if we "balance the equation" by doing the same to each feeder branch, the relative current phase in the radiating elements will remain the same. For example, if we add two feet of 60-ohm transmission line in Branch One of **Figure 2** between L4 and L1, and two feet of RG-8 to the existing 50-ohm line in Branch Two, we introduce an additional -30 degrees of phase shift to each element, assuming a velocity factor of 0.66 (**Figure 9**). Do we still get the desired element currents and a good impedance match on the RG-8?

The answer is a resounding *"NO!"* (see **Table 5**). The relative phase of the current in Element Two is okay, but the magnitude is now 0.55 instead of the desired 0.3. In fact, Element Two no longer has a negative resistance. The operating impedance has changed significantly, and the VSWR on the RG-8 is quite high (in this case the coax won't melt, since the power in Element Two is now only 10 watts). But the point is that everything has changed dramatically from the desired operating conditions. I also want you to take note of the fact that the input impedance to the feeder network is only a bit off. If you were relying on a change in this input impedance to tell you that something was wrong with your antenna, you'd never know, would you?

Remember that we're dealing with a negative power in Element Two, and the simple fact is that if we add two feet of transmission line to Element One, then we must *subtract* two feet from Element Two. This is because the power is traveling in opposite directions on the two transmission lines. In other words, if we were to add a –60-degree tee network at the base of Element One, then our existing –95-degree tee network at the base of Element Two would have to be changed to –95 + 60 = -35 degrees. Similarly, if we add –30 degrees of transmission line to Element One's feeder, we must either add +30 degrees of phase shift to Element Two's tee networks (-95 + 30 = -65_), or reduce the length of Element Two's transmission line by 30 degrees.

If, instead, we subtract two feet of RG–8 from the original 11.5-foot length (**Figure 9** with 9.5 feet of RG-8), our antenna behaves the way we want it to, the VSWR on the transmission lines becomes reasonable, the monopole operating impedances and currents are close to the design values, etc. (**Table 6**). Again, we could have achieved the same effect by leaving the 11.5-foot length of RG-8 intact, and compensated for the additional –30 degrees of phase shift in feeder one by adding +30 degrees to the feeder two tee network, making it –65 degrees (**Figure 10, Table 7**).

Now you might ask, "Why go to all the bother of bringing back the negative element's power to the antenna input, as it requires several additional components and only improves

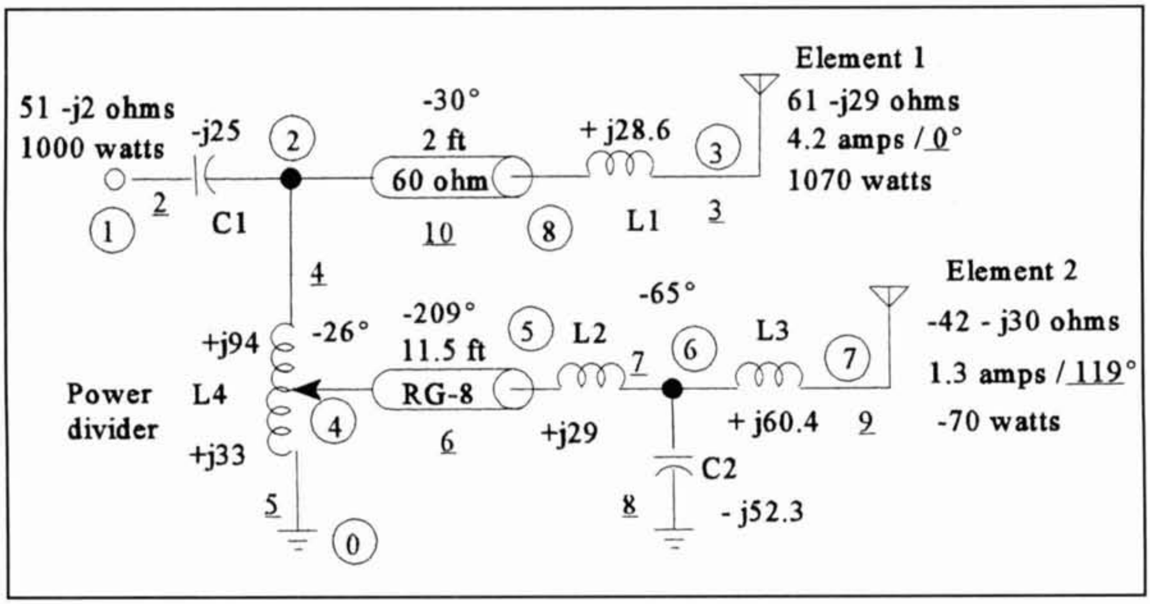

Figure 10. Feeder network with Line 1 inserted and tee phase shift reduced.

the radiated field intensity by 0.3 dB?" In fact, because we now have transmission line losses and extra component losses to worry about, we may actually end up with **less** radiated power compared to the simple coil and resistor termination of the negative element in **Figure 1**. If we assume all coils and capacitors have a Q of 200, and the RG-8 loss is 1.1 dB/100 feet at 28 MHz, then **Table 8** tells us that we only lose 11 watts in the feeder network, which is 0.05 dB. Thus, the complete network of **Figure 2** has a 0.25-dB gain compared to the simpler network of **Figure 1**.

Also keep in mind that there may be times when you have a lot more power in the negative element than the seven percent in our example dual monopole array. Thus, the loss in a simple resistor termination may be high enough to force you to bring the negative power back to the power divider if you expect any gain from your antenna. If you don't know how to bring that power back to the antenna input properly, you'll never get the desired radiation pattern shape and size, and you may even damage your network components. Phased array adjustment can be tricky, and it's easy to get lost in the woods. But if you have an understanding of **Equations 5** and **6**, at least you'll be heading in the right direction.

I've used **Equation 5** successfully in a 200-kW, four-tower array with **two** negative towers, and I have every confidence that this equation will work for you as well. But because there are a number of antenna designers who are still using **Equation 6** instead of **Equation 5** to accommodate a negative element, I've compiled the rather lengthy **Appendix**. It's still possible to obtain the desired currents in an antenna having negative elements, if you're willing to settle for significant impedance mismatches within the feeder system. However, if you want to do it right, use **Equation 5**, unless you really like flying by the seat of your pants!

REFERENCES

1. Grant Bingeman," Phased Array Adjustment," *Communications Quarterly*, Summer 1998.

Appendix

Computer Printouts of Network Analysis Program LYSIS2

Legend:
Node zero is ground

I = Current Source
R = Resistor
C = Capacitor
L = Inductor
Z = Impedance
X = Transmission Line
M = Mutual Inductance
K = Coupling Coefficient
T = Transformer
G = Transconductance

Table 2.

Input Parameters

Frequency:	28000.0 kHz
Resistance:	61.0 ohms
Reactance:	-0.4 ohms
Current:	4.0 amps RMS
Voltage:	247.0 volts RMS
VSWR:	1.22
Power:	1000.0 watts

Node	Voltage-to-Ground Polar		Cartesian		
1	247.0	-0.4	247.0	-1.8	Network Components
2	272.8	-25.1	247.0	-115.8	
3	54.0	-59.9	27.1	-46.7	
4	63.7	-27.9	56.3	-29.8	
5	110.9	-42.2	82.1	-74.5	Antenna Model
6	108.1	-47.6	72.9	-79.8	
7	136.8	-57.8	72.9	-115.8	
8	97.4	-73.2	28.1	-93.3	

Branch		Nodes Fr	To	Branch-Voltage Mag	Phase	Branch-Current Mag	Phase	From-Impedance R ohms	X ohms	To-Impedance R ohms	X ohms
2	L	1	2	114.0	90.0	4.05	0.0	61.0	-0.4	61.0	-28.6
3	L	3	4	33.8	-149.9	1.20	120.1	-45.0	0.0	-45.0	-28.1
4	R	3	0	54.0	-59.9	1.20	-59.9	45.0	0.0	0.0	0.0
5	R	4	5	51.6	120.1	1.20	120.1	-45.0	-28.1	-88.0	-28.1
6	C	5	6	10.7	30.1	1.20	120.1	-88.0	-28.1	-88.0	-19.3
7	R	2	7	174.1	0.0	4.05	0.0	61.0	-28.6	18.0	-28.6
8	C	7	6	36.0	-90.0	4.05	0.0	18.0	-28.6	18.0	-19.7
9	R	6	8	46.8	16.8	3.60	16.8	13.0	-27.1	0.0	-27.1
10	C	8	0	97.4	-73.2	3.60	16.8	0.0	-27.1	0.0	0.0

Branch		Nodes Fr	To	Branch Reactance	Node From-Power	Node To-Power	Branch Power	
2	L	1	2	28.1	1000.0	1000.0	0.0	
3	L	3	4	28.1	-64.8	-64.8	0.0	
4	R	3	0	45.0	64.8	0.0	64.8	65 watts lost
5	R	4	5	43.0	-64.8	-126.7	61.9	total of
6	C	5	6	0.0	-126.7	-126.7	0.0	935 watts
7	R	2	7	43.0	1000.0	295.2	704.8	radiated
8	C	7	6	0.0	295.2	295.2	0.0	
9	R	6	8	13.0	168.5	0.0	168.5	
10	C	8	0	0.0	0.0	0.0	0.0	

Transfer Function		Mag	Phase	dB
1	I3/I2	0.296	120.1	-10.6

Table 2. Computer analysis of Figure 1 network including coupled antenna model with passive termination of negative element.

Table 3.

Input Parameters

Frequency:	28000.0 kHz
Resistance:	45.0 ohms
Reactance:	4.0 ohms
Current:	4.7 amps RMS
Voltage:	212.9 volts RMS
VSWR:	1.14
Power:	1000.0 watts

Node	Voltage-to-Ground Polar		Cartesian		
1	212.9	5.1	212.1	18.8	Network Components
2	261.4	35.8	212.1	152.8	
3	278.7	13.7	270.7	66.2	
4	18.0	-0.8	18.0	-0.3	
5	20.2	-174.3	-20.1	-2.0	
6	15.9	104.3	-3.9	15.4	
7	110.8	50.1	71.1	85.0	
8	135.1	7.2	134.1	17.0	Antenna Model
9	120.1	1.9	120.0	4.0	
10	140.5	-9.5	138.5	-23.3	
11	108.3	-23.7	99.1	-43.6	

Branch		Nodes Fr	To	Branch-Voltage Mag	Phase	Branch-Current Mag	Phase	From-Impedance R ohms	X ohms	To-Impedance R ohms	X ohms
2	C	1	2	134.0	-90.0	4.71	0.0	45.0	4.0	45.0	32.4
3	L	2	3	104.5	124.1	3.71	34.1	70.4	2.0	70.4	-26.1
4	L	2	4	247.2	38.3	2.65	-51.7	4.3	98.5	4.3	5.3
5	L	4	0	18.0	-0.8	0.54	-90.8	0.0	33.4	0.0	0.0
6	X	4	5	0.0	0.0	2.26	-43.1	5.9	5.4	5.9	6.7
7	L	5	6	23.8	-132.9	2.25	137.1	5.9	6.7	5.9	-3.8
8	C	6	0	15.9	104.3	0.19	-165.7	0.0	-83.6	0.0	0.0
9	L	6	7	102.3	-137.2	2.15	132.8	6.5	-3.5	6.5	-51.0
10	R	7	8	92.6	132.8	2.15	132.8	6.5	-51.0	-36.5	-51.0
11	C	8	9	19.1	42.8	2.15	132.8	-36.5	-51.0	-36.5	42.1
12	R	3	10	159.6	34.1	3.71	34.1	70.4	-26.1	27.4	-26.1
13	C	10	9	33.0	-55.9	3.71	34.1	27.4	-26.1	27.4	-17.2
14	R	9	11	52.0	66.3	4.00	66.3	13.0	-27.1	0.0	-27.1
15	C	11	0	108.3	-23.7	4.00	66.3	0.0	-27.1	0.0	0.0

Branch		Nodes Fr	To	Branch Reactance	Node From-Power	Node To-Power	Branch Power	
2	C	1	2	0.0	1000.0	1000.0	0.0	Network Components
3	L	2	3	28.1	969.9	969.9	0.0	
4	L	2	4	93.2	30.1	30.1	0.0	
5	L	4	0	33.4	0.0	0.0	0.0	
7	L	5	6	10.6	30.1	30.1	0.0	
8	C	6	0	0.0	0.0	0.0	0.0	
9	L	6	7	47.5	30.1	30.1	0.0	
10	R	7	8	43.0	30.1	-169.4	199.5	Antenna Model
11	C	8	9	0.0	-169.4	-169.4	0.0	
12	R	3	10	43.0	969.9	377.3	592.6	
13	C	10	9	0.0	377.3	377.3	0.0	
14	R	9	11	13.0	207.9	0.0	207.9	
15	C	11	0	0.0	0.0	0.0	0.0	

Transfer Function		Mag	Phase	dB
1	19/13	0.580	98.7	-4.7

Table 3. Computer analysis of Figure 8 network including coupled antenna model incorrectly using positive design equation for negative case.

Table 4.

Input Parameters

Frequency:	28000.0 kHz
Resistance:	50.4 ohms
Reactance:	-0.9 ohms
Current:	4.5 amps RMS
Voltage:	224.5 volts RMS
VSWR:	1.02
Power:	1000.0 watts

Node	**Voltage-to-Ground**				
	Polar		**Cartesian**		
1	224.5	-1.0	224.4	-4.0	Network Components
2	255.7	28.7	224.4	122.6	
3	282.9	4.1	282.2	20.0	
4	57.6	54.9	33.1	47.1	
5	57.8	-126.8	-34.6	-46.3	
6	80.3	-76.6	18.6	-78.1	
7	68.9	1.0	68.9	1.2	
8	116.1	-13.3	112.9	-26.7	Antenna Model
9	113.0	-18.5	107.2	-35.8	
10	142.8	-28.5	125.4	-68.2	
11	101.9	-44.1	73.1	-70.9	

Branch		**Nodes**		**Branch-Voltage**		**Branch-Current**		**From-Impedance**		**To-Impedance**	
		Fr	**To**	**Mag**	**Phase**	**Mag**	**Phase**	**R ohms**	**X ohms**	**R ohms**	**X ohms**
2	C	1	2	126.6	-90.0	4.46	0.0	50.4	-0.9	50.4	27.5
3	L	2	3	117.8	119.4	4.18	29.4	61.1	-0.8	61.1	-28.9
4	L	2	4	205.7	21.5	2.21	-68.5	-14.3	115.0	-14.3	21.8
5	L	4	0	57.6	54.9	1.72	-35.1	0.0	33.4	0.0	0.0
6	X	4	5	0.0	0.0	1.22	-119.5	-47.0	4.7	-47.3	4.8
7	L	5	6	62.0	149.1	1.22	59.1	-47.3	4.8	-47.3	-46.2
8	C	6	0	80.3	-76.6	1.70	13.4	0.0	-47.4	0.0	0.0
9	L	6	7	93.9	-122.4	1.21	147.6	-47.4	46.2	-47.4	-31.2
10	R	7	8	52.2	147.6	1.21	147.6	-47.4	-31.2	-90.4	-31.2
11	C	8	9	10.8	57.6	1.21	147.6	-90.4	-31.2	-90.4	-22.3
12	R	3	10	179.9	29.4	4.18	29.4	61.1	-28.9	18.1	-28.9
13	C	10	9	37.2	-60.6	4.18	29.4	18.1	-28.9	18.1	-20.0
14	R	9	11	48.9	45.9	3.76	45.9	13.0	-27.1	0.0	-27.1
15	C	11	0	101.9	-44.1	3.76	45.9	0.0	-27.1	0.0	0.0

Branch		**Nodes**		**Branch Reactance**	**Node From-Power**	**Node To-Power**	**Branch Power**	
		Fr	**To**					
2	C	1	2	0.0	1000.0	1000.0	0.0	Network Components
3	L	2	3	28.1	1069.8	1069.8	0.0	
4	L	2	4	93.2	-69.8	-69.8	0.0	
5	L	4	0	33.4	0.0	0.0	0.0	
7	L	5	6	51.0	-69.8	-69.8	0.0	
8	C	6	0	0.0	0.0	0.0	0.0	
9	L	6	7	77.4	-69.8	-69.8	0.0	
10	R	7	8	43.0	-69.8	-133.1	63.3	Antenna Model
11	C	8	9	0.0	-133.1	-133.1	0.0	
12	R	3	10	43.0	1069.8	317.3	752.6	total radiated
13	C	10	9	0.0	317.3	317.3	0.0	power = 1000 watts
14	R	9	11	13.0	184.1	0.0	184.1	
15	C	11	0	0.0	0.0	0.0	0.0	

Transfer Function		**Mag**	**Phase**	**dB**
1	19/13	0.290	118.3	-10.8

Table 4. Computer analysis of Figure 2 feeder network including coupled antenna model.

Table 5.

Input Parameters

Frequency:	28000.0 kHz
Resistance:	56.7 ohms
Reactance:	-4.0 ohms
Current:	4.2 amps RMS
Voltage:	238.8 volts RMS
VSWR:	1.16
Power:	1000.0 watts

Node	Voltage-to-Ground Polar		Cartesian		
1	238.8	-4.1	238.2	-16.9	Network Components
2	259.3	23.3	238.2	102.4	
3	269.2	-28.3	237.2	-127.5	
4	91.1	21.6	84.7	33.5	
5	101.9	-159.5	-95.5	-35.7	
6	97.3	-161.7	-92.4	-30.6	
7	68.9	23.2	63.3	27.1	
8	255.8	-4.3	255.1	-19.3	
9	112.3	-31.9	95.4	-59.4	Antenna Model
10	101.8	-40.4	77.5	-66.0	
11	123.3	-54.3	71.9	-100.1	
12	91.8	-66.1	37.2	-83.9	

Branch		Nodes Fr	To	Branch-Voltage Mag	Phase	Branch-Current Mag	Phase	From-Impedance R ohms	X ohms	To-Impedance R ohms	X ohms
2	C	1	2	119.3	-90.0	4.20	0.0	56.7	-4.0	56.7	24.4
3	X	2	8	0.0	0.0	3.83	25.5	67.6	-2.6	65.4	5.8
4	L	8	3	109.7	80.6	3.90	-9.4	65.4	5.8	65.4	-22.3
5	L	2	4	168.3	24.2	1.80	-65.8	2.3	143.6	2.3	50.4
6	L	4	0	91.1	21.6	2.72	-68.4	0.0	33.4	0.0	0.0
7	X	4	5	0.0	0.0	0.92	106.6	8.6	-98.1	533.8	686.1
8	L	5	6	6.0	-121.6	0.12	148.4	533.8	686.1	533.8	635.0
9	C	6	0	97.3	-161.7	2.05	-71.7	0.0	-47.4	0.0	0.0
10	L	6	7	166.0	-159.7	2.15	110.3	1.6	45.3	1.6	-32.1
11	R	7	9	92.2	110.3	2.15	110.3	1.6	-32.1	-41.4	-32.1
12	C	9	10	19.1	20.3	2.15	110.3	-41.4	-32.1	-41.4	-23.2
13	R	3	11	167.5	-9.4	3.90	-9.4	65.4	-22.3	22.4	-22.3
14	C	11	10	34.6	-99.4	3.90	-9.4	22.4	-22.3	22.4	-13.5
15	R	10	12	44.1	23.9	3.39	23.9	13.0	-27.1	0.0	-27.1
16	C	12	0	91.8	-66.1	3.39	23.9	0.0	-27.1	0.0	0.0

Branch		Nodes Fr	To	Branch Reactance	Node From-Power	Node To-Power	Branch Power	
2	C	1	2	0.0	1000.0	1000.0	0.0	Network Components
4	L	8	3	28.1	992.7	992.7	0.0	
5	L	2	4	93.2	7.3	7.3	0.0	
6	L	4	0	33.4	0.0	0.0	0.0	
8	L	5	6	51.0	7.3	7.3	0.0	
9	C	6	0	0.0	0.0	0.0	0.0	
10	L	6	7	77.4	7.3	7.3	0.0	
11	R	7	9	43.0	7.3	-190.5	197.8	Antenna Model
12	C	9	10	0.0	-190.5	-190.5	0.0	
13	R	3	11	43.0	992.7	339.9	652.7	
14	C	11	10	0.0	339.9	339.9	0.0	
15	R	10	12	13.0	149.4	0.0	149.4	
16	C	12	0	0.0	0.0	0.0	0.0	

Transfer Function		Mag	Phase	dB
1	I10 / I4	0.551	119.7	-5.2

Table 5. Computer analysis of Figure 9 network including coupled antenna model with 13.5 feet of RG-8.

Table 6.

Input Parameters

Frequency: 28000.0 kHz
Resistance: 50.4 ohms
Reactance: -1.8 ohms
Current: 4.5 amps RMS
Voltage: 224.7 volts RMS
VSWR: 1.04
Power: 1000.0 watts

Node	**Voltage-to-Ground** Polar		Cartesian		
1	224.7	-2.0	224.6	-8.0	Network Components
2	253.9	27.8	224.6	118.5	
3	282.5	-27.3	251.1	-129.4	
4	56.7	54.1	33.2	45.9	
5	60.9	-158.2	-56.5	-22.6	
6	79.1	-111.7	-29.3	-73.5	
7	67.2	-27.0	59.9	-30.6	Antenna Model
8	256.2	-2.7	255.9	-12.1	
9	115.9	-43.7	83.7	-80.1	
10	112.4	-49.2	73.5	-85.0	
11	141.7	-59.5	72.0	-122.0	
12	101.3	-74.8	26.5	-97.8	

Branch		**Nodes** Fr	To	**Branch-Voltage** Mag	Phase	**Branch-Current** Mag	Phase	**From-Impedance** R ohms	X ohms	**To-Impedance** R ohms	X ohms
2	C	1	2	126.5	-90.0	4.45	0.0	50.4	-1.8	50.4	26.6
3	X	2	8	0.0	0.0	4.21	29.2	60.3	-1.4	61.4	-0.4
4	L	8	3	117.4	87.6	4.17	-2.4	61.4	-0.4	61.4	-28.5
5	L	2	4	204.7	20.8	2.20	-69.2	-14.2	114.8	-14.2	21.6
6	L	4	0	56.7	54.1	1.70	-35.9	0.0	33.4	0.0	0.0
7	X	4	5	0.0	0.0	1.21	-119.4	-46.4	5.3	-53.5	5.9
8	L	5	6	57.7	118.2	1.13	28.2	-53.5	5.9	-53.5	-45.1
9	C	6	0	79.1	-111.7	1.67	-21.7	0.0	-47.4	0.0	0.0
10	L	6	7	99.0	-154.3	1.28	115.7	-41.9	45.6	-41.9	-31.8
11	R	7	9	55.0	115.7	1.28	115.7	-41.9	-31.8	-84.9	-31.8
12	C	9	10	11.4	25.7	1.28	115.7	-84.9	-31.8	-84.9	-22.9
13	R	3	11	179.3	-2.4	4.17	-2.4	61.4	-28.5	18.4	-28.5
14	C	11	10	37.0	-92.4	4.17	-2.4	18.4	-28.5	18.4	-19.6
15	R	10	12	48.6	15.2	3.74	15.2	13.0	-27.1	0.0	-27.1
16	C	12	0	101.3	-74.8	3.74	15.2	0.0	-27.1	0.0	0.0

Branch		**Nodes** Fr	To	**Branch Reactance**	**Node From-Power**	**Node To-Power**	**Branch Power**	
2	C	1	2	0.0	1000.0	1000.0	0.0	Network Components
4	L	8	3	28.1	1068.4	1068.4	0.0	
5	L	2	4	93.2	-68.4	-68.4	0.0	
6	L	4	0	33.4	0.0	0.0	0.0	
8	L	5	6	51.0	-68.4	-68.4	0.0	
9	C	6	0	0.0	0.0	0.0	0.0	
10	L	6	7	77.4	-68.4	-68.4	0.0	
11	R	7	9	43.0	-68.4	-138.7	70.3	Antenna Model
12	C	9	10	0.0	-138.7	-138.7	0.0	
13	R	3	11	43.0	1068.4	320.7	747.7	
14	C	11	10	0.0	320.7	320.7	0.0	
15	R	10	12	13.0	182.1	0.0	182.1	
16	C	12	0	0.0	0.0	0.0	0.0	

Transfer Function		**Mag**	**Phase**	**dB**
1	I10 / I4	0.307	118.1	-10.3

Table 6. Computer analysis of Figure 9 network including coupled antenna model with 9.5 feet of RG-8.

Table 7.

Input Parameters

Frequency: 28000.0 kHz
Resistance: 50.6 ohms
Reactance: -1.9 ohms
Current: 4.4 amps RMS
Voltage: 225.1 volts RMS
VSWR: 1.04
Power: 1000.0 watts

Node	Voltage-to-Ground Polar		Cartesian		
1	225.1	-2.2	224.9	-8.6	Network Components
2	253.9	27.6	224.9	117.8	
3	282.3	-27.5	250.3	-130.5	
4	58.7	53.0	35.3	46.9	
5	58.8	-128.5	-36.6	-46.0	
6	65.6	-98.5	-9.7	-64.9	
7	66.4	-28.0	58.6	-31.2	
8	256.0	-2.9	255.6	-13.1	
9	115.3	-44.1	82.8	-80.3	Antenna Model
10	112.0	-49.6	72.6	-85.3	
11	141.4	-59.9	71.0	-122.3	
12	101.0	-75.2	25.8	-97.6	

Branch		Nodes Fr	To	Branch-Voltage Mag	Phase	Branch-Current Mag	Phase	From-Impedance R ohms	X ohms	To-Impedance R ohms	X ohms
2	C	1	2	126.4	-90.0	4.45	0.0	50.6	-1.9	50.6	26.5
3	X	2	8	0.0	0.0	4.21	28.9	60.3	-1.3	61.3	-0.3
4	L	8	3	117.5	87.4	4.17	-2.6	61.3	-0.3	61.3	-28.5
5	L	2	4	202.4	20.5	2.17	-69.5	-14.6	116.0	-14.6	22.8
6	L	4	0	58.7	53.0	1.76	-37.0	0.0	33.4	0.0	0.0
7	X	4	5	0.0	0.0	1.17	-123.4	-50.1	3.1	-50.3	3.1
8	L	5	6	32.9	145.0	1.17	55.0	-50.3	3.1	-50.3	-25.0
9	C	6	0	65.6	-98.5	1.25	-8.5	0.0	-52.6	0.0	0.0
10	L	6	7	76.1	-153.7	1.27	116.3	-42.3	29.4	-42.3	-30.4
11	R	7	9	54.7	116.3	1.27	116.3	-42.3	-30.4	-85.3	-30.4
12	C	9	10	11.3	26.3	1.27	116.3	-85.3	-30.4	-85.3	-21.6
13	R	3	11	179.5	-2.6	4.17	-2.6	61.3	-28.5	18.3	-28.5
14	C	11	10	37.1	-92.6	4.17	-2.6	18.3	-28.5	18.3	-19.6
15	R	10	12	48.5	14.8	3.73	14.8	13.0	-27.1	0.0	-27.1
16	C	12	0	101.0	-75.2	3.73	14.8	0.0	-27.1	0.0	0.0

Branch		Nodes Fr	To	Branch Reactance	Node From-Power	Node To-Power	Branch Power	
2	C	1	2	0.0	1000.0	1000.0	0.0	Network Components
4	L	8	3	28.1	1068.6	1068.6	0.0	
5	L	2	4	93.2	-68.6	-68.6	0.0	
6	L	4	0	33.4	0.0	0.0	0.0	
8	L	5	6	28.1	-68.6	-68.6	0.0	
9	C	6	0	0.0	0.0	0.0	0.0	
10	L	6	7	59.8	-68.6	-68.6	0.0	
11	R	7	9	43.0	-68.6	-138.2	69.7	Antenna Model
12	C	9	10	0.0	-138.2	-138.2	0.0	
13	R	3	11	43.0	1068.6	319.2	749.4	
14	C	11	10	0.0	319.2	319.2	0.0	
15	R	10	12	13.0	180.9	0.0	180.9	
16	C	12	0	0.0	0.0	0.0	0.0	

Transfer Function		Mag	Phase	dB
1	I10 / I 4	0.305	118.9	-10.3

Table 7. Computer analysis of Figure 10 network including coupled antenna model 2 feet 60 ohm line, 11.5 feet RG-8, tee network phase shift –65°.

Table 8.

Input Parameters

Frequency:	28000.0 kHz
Resistance:	50.5 ohms
Reactance:	-0.9 ohms
Current:	4.5 amps RMS
Voltage:	224.7 volts RMS
VSWR:	1.02
Power:	1000.0 watts

Node	Voltage-to-Ground Polar		Cartesian		
1	224.7	-1.0	224.7	-3.9	Network Components
2	255.4	28.7	224.0	122.6	
3	281.7	4.2	280.9	20.5	
4	56.6	53.7	33.5	45.6	
5	57.7	-127.7	-35.3	-45.6	
6	77.1	-77.3	16.9	-75.2	
7	69.8	2.5	69.7	3.1	
8	116.3	-13.0	113.4	-26.1	Antenna Model
9	112.9	-18.1	107.4	-35.1	
10	142.3	-28.2	125.5	-67.2	
11	101.8	-43.8	73.5	-70.4	

Branch		Nodes Fr	To	Branch-Voltage Mag	Phase	Branch-Current Mag	Phase	From-Impedance R ohms	X ohms	To-Impedance R ohms	X ohms
2	C	1	2	126.5	-89.7	4.45	0.0	50.5	-0.9	50.3	27.5
3	L	2	3	116.8	119.1	4.15	29.4	61.5	-0.8	61.4	-28.9
4	L	2	4	205.5	22.0	2.20	-67.7	-12.9	115.1	-13.4	21.9
5	L	4	0	56.6	53.7	1.69	-36.0	0.2	33.4	0.0	0.0
6	X	4	5	0.0	0.0	1.17	-117.1	-47.6	7.7	-48.5	7.2
7	L	5	6	60.0	150.5	1.18	60.7	-48.5	7.2	-48.7	-43.8
8	C	6	0	77.1	-77.3	1.63	12.4	0.2	-47.4	0.0	0.0
9	L	6	7	94.4	-124.0	1.22	146.3	-45.8	43.6	-46.2	-33.8
10	R	7	8	52.5	146.3	1.22	146.3	-46.2	-33.8	-89.2	-33.8
11	C	8	9	10.8	56.3	1.22	146.3	-89.2	-33.8	-89.2	-24.9
12	R	3	10	178.5	29.4	4.15	29.4	61.4	-28.9	18.4	-28.9
13	C	10	9	36.9	-60.6	4.15	29.4	18.4	-28.9	18.4	-20.1
14	R	9	11	48.9	46.2	3.76	46.2	13.0	-27.1	0.0	-27.1
15	C	11	0	101.8	-43.8	3.76	46.2	0.0	-27.1	0.0	0.0

Branch		Nodes Fr	To	Branch Reactance	Node From-Power	Node To-Power	Branch Power	
2	C	1	2	0.0	1000.0	997.2	2.8	Network Components
3	L	2	3	28.1	1060.0	1057.6	2.4	
4	L	2	4	93.2	-62.8	-65.1	2.3	
5	L	4	0	33.4	0.5	0.0	0.5	
7	L	5	6	51.0	-67.1	-67.5	0.4	
8	C	6	0	0.0	0.6	0.0	0.6	
9	L	6	7	77.4	-68.1	-68.7	0.6	
10	R	7	8	43.0	-68.7	-132.7	64.0	Antenna Model
11	C	8	9	0.0	-132.7	-132.7	0.0	
12	R	3	10	43.0	1057.6	316.6	741.0	
13	C	10	9	0.0	316.6	316.6	· 0.0	
14	R	9	11	13.0	183.9	0.0	183.9	
15	C	11	0	0.0	0.0	0.0	0.0	

Transfer Function		Mag	Phase	dB
1	19/13	0.294	116.8	-10.6

Table 8. Computer analysis of Figure 2 network with lossy components.

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