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Fall 1999

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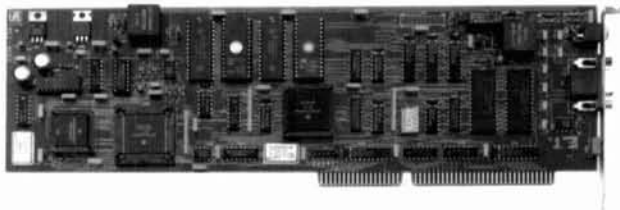




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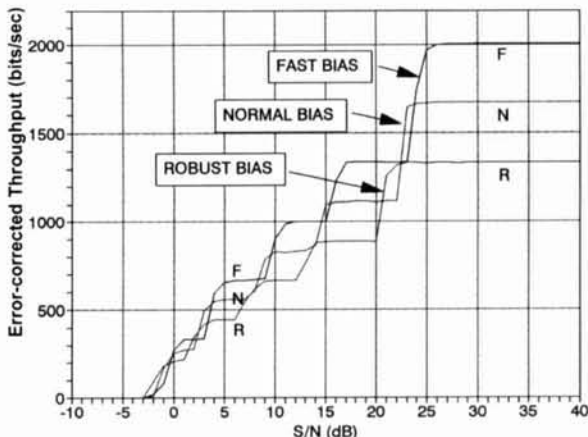
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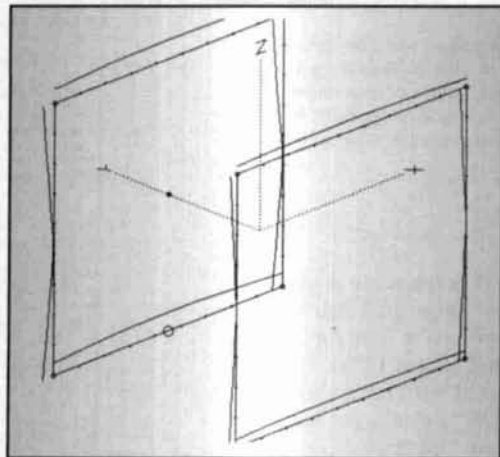
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EDITORIAL

The Thrill of QRP, a Traditional Way to Ham

Recently, at the Hosstraders Hamfest held in Rochester, New Hampshire, I discussed the state of amateur radio with a number of my ham friends. This particular sampling of radio buffs reads every technical ham publication they can get their hands on, feels that building is essential to the continued growth of the hobby, and, oh yes, are all members of a small but powerful group of hams called QRPers.

Who are the QRPers

QRPers are amateur radio operators who prefer to operate low power. These hardy souls get their thrills contacting distant stations while transmitting a minimum number of watts. QRP power levels are 5 watts or less on CW and 10 watts or less on SSB. The difficulty involved in making contacts harkens back to the earliest days of amateur radio—before powerful amplifiers and state-of-the-art rigs boosted the odds. CW is still the favored mode with QRP operators because they believe CW often gets through where voice communications sometimes can't; however, there are those in the fraternity who operate voice.

Most hams who operate QRP are also builders. Designs for radios, antennas, power supplies, etc., abound in various QRP publications like the GQRV Club's *SPRAT*, the QRP Amateur Radio Club International's (QRP ARCI) *QRP Quarterly*, and NorCal's *QRPP*. Online links also offer a host of projects on club web pages and individual ham home pages. Kits like Elecraft's K2 HF Transceiver, designed by N6KR and WA6HHQ, are available for those who want to build their complete rigs or peripherals without the hassle of etching their own boards and tracking down parts. Those who'd rather not build can buy commercial units from companies like Ten-Tec and MFJ.

Portable operation

Loads of QRPers like the option of operating portable. Many of them design tiny rigs in small boxes (remember the metal Sucrets box?) with little power supplies (solar is also a consideration), and portable antennas, so they can operate their rigs anywhere.

While browsing online, I found The Adventure Radio Society. This group combines a love of radio with a love of the outdoors and offers several events for QRPers during the year. One most in keeping with QRP ideals of creativity and low power is "The Trail Friendly Radio Challenge," which encourages members of the QRP community to "develop imaginative equip-

ment and antennas for the outdoor environment." In addition to their web page, the society has an online magazine: *The ARS Sojourner*.

Another QRP operator who knows the meaning of operating portable was featured on the cover of the July, 1999, issue of *QRP Quarterly*. According to the bio on page 6 of the issue, Ed Loranger, WE6W, uses his 10-speed bicycle as his sole means of transportation. Still, he wanted to be able to bring his ham equipment with him wherever he went. After much trial and error, Ed was able to work up a reliable and versatile travel box for the back of his bike that carries his gear and serves as an operating table and antenna stand. WE6W's setup is truly a tribute to the QRPers ingenuity!

QRP's roots in tradition

I asked Rick Littlefield, K1BQT, an early QRP aficionado and QRP Hall of Fame member, what involvement with QRP can do for someone who finds that typical amateur radio operation has lost its thrill. He handed me a page from a manual he was working on, and the sentiments presented there were right on:

Technically speaking, operating QRP means limiting your transmitter power to below 5 watts on CW or below 10 watts PEP on sideband. However, for a growing number of licensed amateurs, QRP symbolizes a return to the basics of radio—with a strong emphasis on operating skills, experimentation, home construction, and fraternity.

Spanning a continent or hopping oceans with less energy than it takes to illuminate a night light is not only exciting; it borders on the miraculous. Yet, QRPers do it every day—often using simple home-built equipment running only microwatts of power. Even staunch QRO contesters have succumbed to the lure and challenge of QRP, revisiting DXCC while running 5 watts or less.

For many, operating QRP restores a sense of personal achievement that's easily lost when high-tech appliances invade the ham shack. Regardless of motivation, this particular segment of the ham radio community is growing steadily and continues to thrive as other techno-fads come and go.

Check it out

I think Rick says it all. If you want to learn more about the low power side of our hobby, direct your online search engine to look for amateur radio links. I found almost 1,700, most of which had additional links for me to explore. One of the best QRP links is that for the QRP ARCI. You can find them at <<http://www.qrparci.org>>. If you're not online, you can still get involved by contacting the QRP ARCI at 117 E. Philip Street, Des Moines, Iowa 50315-4114. **72 de KA1STC**

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TECHNICAL CONVERSATIONS

Checking the Smith Chart

Dear Editor:

In "Tech Notes" in the Summer 1999 issue, Steve Sparks, N5SV, describes a Smith Chart solution to his tower matching problem. I believe there is an error in his procedure.

On pages 102 and 105, he states:

Starting at the antenna end, Z_L , draw a circle of constant impedance using a compass whose center is "0" on the far left of the horizontal resistance axis. The length...is from the worst case...to where it intersects the 1.0 constant resistance circle....

The objective of this step is to determine the shunt capacitance needed to place the transformed impedance (of the load and the shunt capacitance) on the unit resistance circle. But to do this requires travel along a circle of constant conductance. The set circles of constant conductance are all tangential to the "0" on the far left (not centered there) and a radius drawn from "0" does not define either a circle of constant conductance or a circle of constant impedance, as stated above.

Please observe the enclosed **Figure 1**, representing a skeletal Smith Chart. The circle labeled "R=1" is the unit circle of constant resistance. The circle labeled G, whose center is at X1, is a circle of constant conductance through load impedance Z_L . R0 is the radius described by Mr. Sparks, which defines the arc also passing through Z_L . Note that the constant conductance circle centered on X1 intersects the unit resistance circle at P3, while the arc drawn from the "0" point intersects at P2. Generally, the values of both the shunt capacitance and the series inductance will be different from those obtained by using Mr. Sparks' arc. (In his case, the difference is not large, and the use of a variable capacitor masks the error.)

It is not necessary to determine X1. A left-right flip of the chart produces an admittance chart instead of an impedance chart; the constant resistance circles are then constant-conductance circles. Overlaying an admittance chart over an impedance chart allows these two-element matching problems to be solved with ease; photocopying a Smith chart onto a transparency further simplifies the method.

The Smith Chart is an elegant, powerful tool which has been indispensable to engineers for six decades. If repeated matching exercises are anticipated, the purchase of Smith Chart software (such as the modestly priced *MicroSmith* program by Wes Hayward, W7ZOI, available from the ARRL) will both simplify the task and

enlighten the user as to the capabilities of the method. Wideband matching, tank circuit design, tolerance analysis with multiple elements, and more, are readily accomplished with such software.

Garry Shapiro, NI6T
Los Gatos, California

Clarifying the image

Dear Editor:

I appreciated Mr. Gruchalla's article on image scanning and printing (*Communications Quarterly*, Summer 1999, page 9), especially since I am currently in the market for both a scanner and an ink jet printer, and am waiting for adequate quality digital cameras to become affordable. The numbers which he provided on existing analog processes has helped me to sort out what I really need. I would like to clarify two items which appear to be overlooked in the article, however, and to explore their impact on his conclusions.

First, the test pattern used to determine resolution is "digital" in nature. If it were digitized by a system with infinite bandwidth, the result-

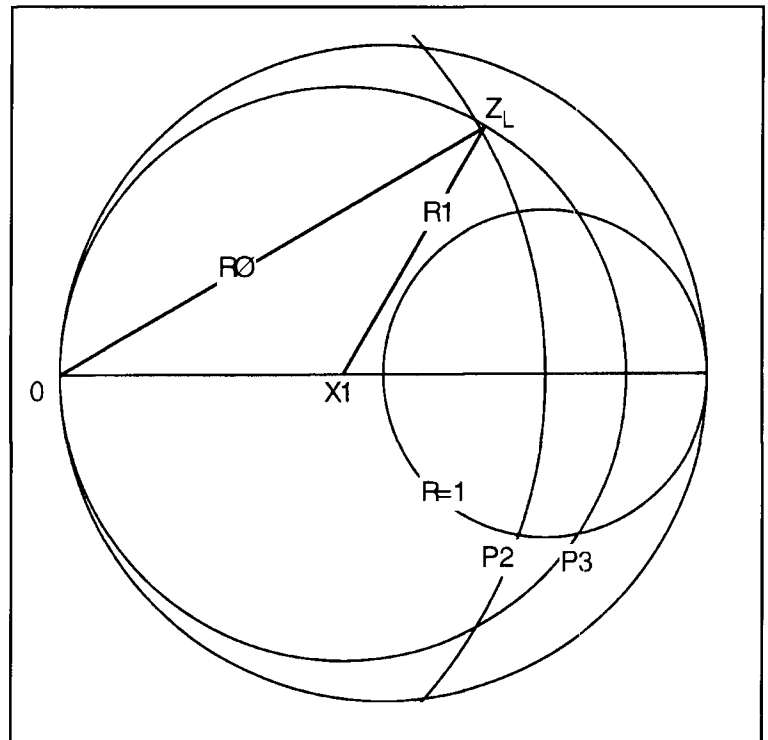


Figure 1. A skeletal Smith Chart.

ing spatial waveform would be a square wave, with infinite harmonics above the frequency determined by the line pair resolution. This is the reason for the artifacts he points out in scanning such an image. In reality, the analog process, like all analog systems, includes some inherent low-pass filtering (although possibly crude) which prevents these sort of artifacts from plaguing it. The resolution limits of the lens is one such filter, the random size and distribution of film grains is another. Were he to photograph the chart and scan the resulting negative or print, the artifacts would be diminished by this low-pass filtering.

When electronic engineers look at digital waveforms we insist on a bandwidth 10 times that of the fundamental if we wish to accurately capture the waveform. It should not be surprising that Mr. Gruchalla wants five times the spatial resolution under similar circumstances. The reason that two times the spatial resolution is adequate for most photographic purposes (besides the low-pass filtering inherent in the analog process which created the image) is that nature rarely provides us with "digital"-type images. Transitions are generally more gradual, requiring less bandwidth.

The second misconception has to do with half-toning and its relation to ink jet printing. The half-tone process used in printing divides each primary color of the image into discrete dots, much like pixels. However, the process relies on spatial limitations and non-linearities of the photographic process to produce dots whose size is a function of intensity, so that each dot carries intensity as well as spatial information. Several of the ink jet printers likewise modulate the dots to produce dots of different sizes depending on the required density (intensity). Such a printer would not need an array of 5 x 5 dots to represent each pixel including intensity. While it is true that the detail structure of any of these processes looks different than an analog photograph under high magnification, the information content can still be identical, and the results indiscernible from the analog counterpart under useful observation conditions. At high magnification, the observer is just comparing one type of artifact (pixels, dithering, etc.) with another (grain structure).

A related point has to do with the interpolated resolution of scanners (or other parts of the digital process). Nyquist's theorem is true in the strictest and most accurate sense only when there exists a reconstruction filter in the decode process. The raw data, while containing all of the information necessary to accurately reproduce the original, does not necessarily represent amplitude levels correctly, and they contain sampling artifacts. A reconstruction filter removes these artifacts and restores the proper

amplitude relationships of the frequency components in the data. When properly done, interpolation of the data points in a scanner produces the same effect, and thus renders a more faithful copy of the original. The primary ingredient here is a low-pass filter.

**Wilton Helm, WT6C
Franktown, Colorado**

Gruchalla responds

Dear Wilton:

Thanks for your comments. My article is simply a brief review of the digital photography process. It was intended to provide a basic insight into the process and limitations. But, admittedly, I did not have the space to thoroughly address the subject (I am pretty sure that Terry considers it more than lengthy enough as it is.)

Your points are well taken; but I believe that I did address them in my piece. Specifically, I believe that I pointed out that the spatial resolving capabilities of most modern lenses far exceed the resolving ability of common films. Therefore, contemporary lenses of average quality do not actually contribute substantially to the analog filtering effect you recite. The image presented to the film, or CCD array, is indeed of quite high spatial frequency content in comparison to the ability to record that content. For film systems, the spatial limitation is primarily in the resolving ability of the film, but the effect is the same. The film provides a smoothing effect of the input image. That is why I based my analyses of scanning of film images on the limitations of the film. But CCD systems are not limited by film limitations, and CCD arrays presently provide much poorer spatial resolution than film. Therefore, if a high-spatial-frequency object is photographed with a CCD camera having a reasonable-quality lens, there will be insufficient spatial filtering to eliminate aliasing.

But you are absolutely correct concerning the smoothing effect. If one were to photograph a test target such as I used in the article (a "square test target" implying abrupt intensity change) and view the result, the sharp edges would be smoothed by the filtering of the resolution limitations of the film. Nevertheless, by using a "perfect" square test target in the article, I could more easily demonstrate scanning limitations by visual example. In reality, the test targets used to measure the spatial frequency response of film (and virtually all other optical elements) have a sinusoidal intensity profile of a specific constant contrast ratio and varying frequency. These are typically analyzed by precision densitometry measurements and mathe-

mathematical processing, such as Fourier transform, to analytically measure performance. But these are very difficult to analyze visually.

The "square" intensity-function test target which I used allows very easy visualization of how the spatial quantization process affects the quantized image. Also, you may have noticed that I used the 1/2-power point of the film response as an analytic measure of its maximum frequency response. (This use of the 1/2-power point of the film is also a departure from the typical photographic art but common in engineering art.) Film actually provides considerable information beyond its 1/2-power point, so it is actually capable of reproducing fine structure well beyond my limitation of the 1/2-power point. The film will provide substantially sharper edges in the image of the square target than the 1/2-power point would predict.

Therefore, my use of the square test target and the 1/2-power point in the analyses are complementary judgment calls—the 1/2-power point under specifies the film frequency capabilities and the square test target over-estimates the film step response. I believe that the aliasing processes that I described using the square test

target can actually be reproduced with real film images of such a target object and contemporary scanners. The film provides insufficient smoothing to eliminate the aliasing.

Indeed, if one were to scan such a perfect square-intensity target, for example scan the actual chip of a large memory, the aliasing effects noted in the article would easily be seen. In such an example, the memory chip is the object, and there is no spatial filtering. If we image the chip onto the scanner, say with 1:1 conjugates to allow us to accurately focus through the window of an E-PROM, the lens would provide some filtering; but I argued that if the lens is of even modest quality by modern standards, the lens spatial resolving ability would be high enough that the lens effects would be negligible in comparison to present scanning technology.

One could of course photograph the memory chip and then scan the resulting image, either on film or on a print, in order to "smooth" the image; however, in the case of this example, that photographic image would not be the origi-

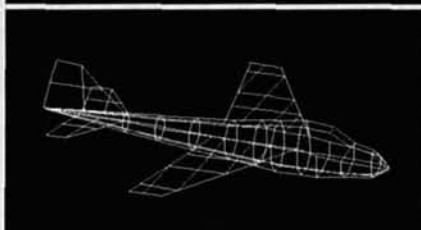
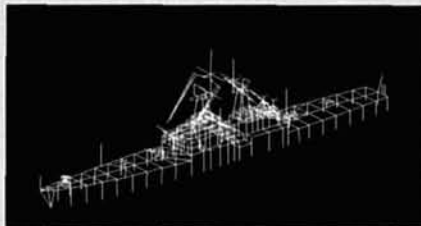
(Continued on page 105)

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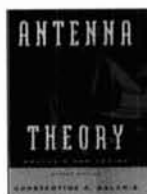


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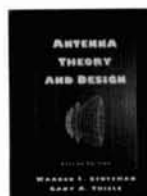


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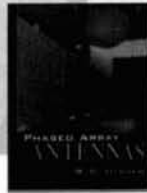


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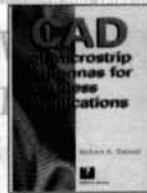


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THE SIMPLE DIODE RF DETECTOR

Operation of the basic diode-detector circuit

The basic diode RF detector in its simplest form consists of only two components: a detector diode and a capacitor. What could be simpler? The diode “detects” the signal and the capacitor stores the “average value” or DC value of the detected signal so it may be measured with some type of DC meter. I specifically did not say that the diode “rectifies” the signal because the word rectify is generally understood to have a slightly different meaning, such as a rectifier in a power-supply circuit. The diode in the RF detector circuit doesn’t generally “rectify” a signal in the same

manner as a rectifier diode rectifies AC power in a power-supply circuit.

With only two components in a diode-detector circuit, it seems that it would be almost impossible to use them incorrectly. That is not quite true. Following my recent *Communications Quarterly* article, “Complex Impedance Measurements,”¹ a reader wrote in to the editor² questioning the detector configuration. This reader wondered whether the diode and capacitor in the forth RF detector circuit should be interchanged to provide a DC ground-return path to the detector capacitor. This was a very

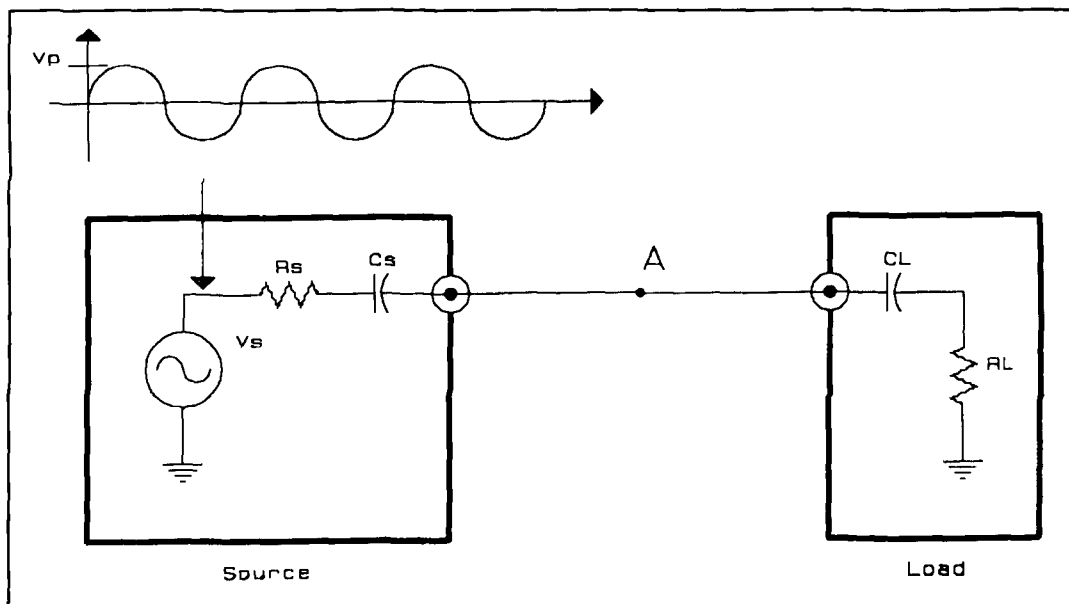


Figure 1. Basic source and load circuit.

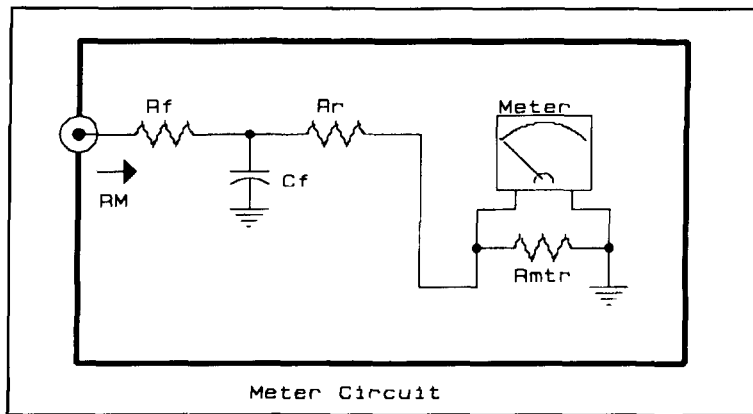


Figure 2. Meter circuit.

good question. The configuration as shown in that article is actually correct. But the operation of the basic diode-detector circuit is not very obvious, even though it only has two parts. In this article, I'll try to explain the operation of the basic diode-detector circuit. This is not a rigorous proof. I won't even use any equations. The intent here is to simply "show" you how the diode detector functions.

Consider the source

First, consider that we have some source driving a load as shown in **Figure 1**. The source voltage is V_S and the source resistance is R_S . A DC blocking capacitor, C_S , is also shown in the source because many RF sources have such a capacitor to prevent any DC component in the source from driving a DC current into the load (DC power of a transmitter final for example), and also to prevent any DC component that may be present in the load from being driven into the source as well. This capacitor is considered very large and effectively a short-circuit at the operating frequency. The load impedance is not necessarily a matched load. If it were, the load voltage would simply be one-half the source voltage V_S . If the load isn't matched to the source, the load voltage will be some other value, but we are only interested in measuring the value of the load voltage.

To make the detection problem a little more interesting, I've included an additional capacitor in the circuit. In **Figure 1**, a load capacitor, C_L , has been included as part of the load. This capacitor may be an actual capacitor used to block any DC component from reaching the load, or it may be an inherent characteristic of the load. For example, if the load were an antenna, this load capacitor would be the antenna capacitance to free space and the load resistance would be the antenna resistance to free space (radiation resistance plus other resis-

tances, such as element resistance, feed-line resistance, and ground-plane resistance). So, the circuit configuration of **Figure 1** is a realistic circuit.

The meter circuit

We'll need some type of meter circuit to "read" the value of the detected signal from our diode-detector circuit. A simple meter circuit is shown in **Figure 2**. The meter is any type of sensitive DC meter. It may be a microamp D'Arsonval meter movement, a digital panel meter, or even a DVM—just some suitable DC meter. The resistance R_{mtr} is the equivalent shunt resistance of the meter. The resistor R_r is included as a range resistor to calibrate the meter as necessary. The capacitor C_f and the resistor R_f make up a simple R-C filter to filter out any RF component of the signal that may be present at the measurement point to ensure that only the DC component of the detected signal is applied to the meter. The DC resistance of the complete meter circuit is simply some value R_M as shown in **Figure 2**. The value of R_M is simply the sum of the resistances R_f , R_r , and R_{mtr} .

Now, in **Figure 1**, suppose we wish to detect the RF signal at node A. If our load were an antenna, this would be the RF voltage applied to the antenna feed point. Note that at node A, there is no DC return path to ground—only capacitors are connected to node A. This is the same situation encountered in the fourth detector circuit in **Reference 1**.

Using a simple diode detector

Suppose we were to use a simple diode detector as shown in **Figure 3**. To simplify this explanation, the diode is defined as ideal having no shunt capacitance, zero forward drop, zero forward resistance, and infinite reverse resistance. This detection configuration is basically a "rectifier" configuration as would be used in rectifying AC power in a power supply. It looks okay, but will it work? After all, it works quite well in AC power circuits. Unfortunately, it will not work as an RF detector. The detected signal will always be zero. But just why this happens isn't very obvious.

The problem is that at circuit node A there is no DC path to ground (as was observed by the reader in the Letter to the Editor² noted earlier). This allows node A to assume any average potential (any DC component of voltage), and there's no circuit element to return the voltage to zero. Suppose the DC component of voltage at node A starts out at zero volts. Now, let the signal start on a first positive cycle. As the sig-

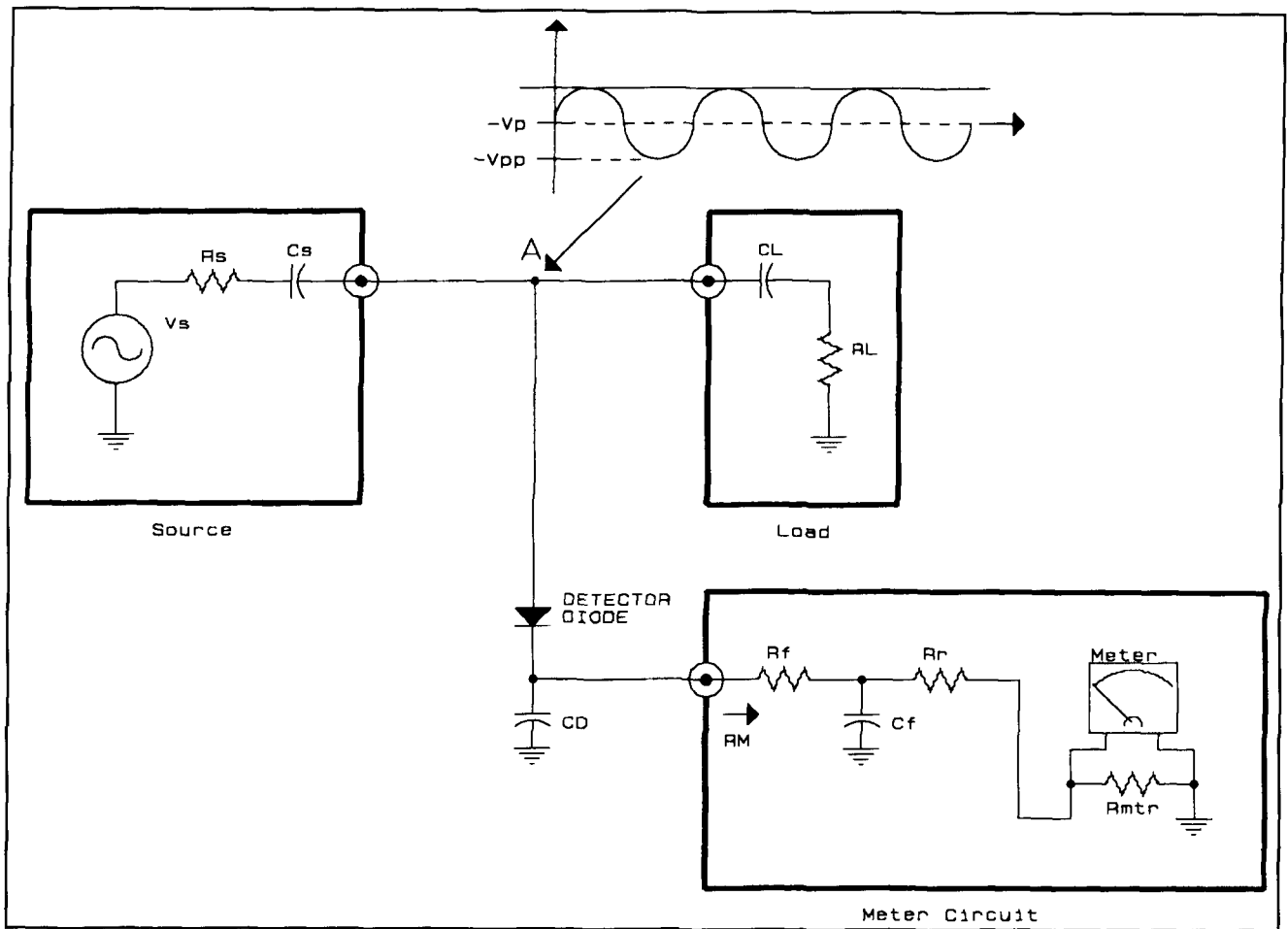


Figure 3. Rectifier-type detector circuit.

nal cycle begins to rise, current is driven through C_S , the detector diode and C_D to ground, and through C_L and R_L to ground. Some current is also driven through the meter circuit to ground. This current flow places a charge on C_S , C_L , and C_D , and this charging continues through much of the positive signal cycle.

When the signal polarity reverses, the detector diode switches off and no current flows in reverse. Whatever DC charge was placed on capacitors C_S and C_L during the positive half cycle of the signal remains there during the first negative half cycle. AC charging of the capacitors also occurs; however, because this doesn't affect the DC charge component, we aren't concerned with these AC components. So, all the DC charge remains on C_S and C_L , but some of the charge on C_D is dissipated in the meter circuit. When the next positive cycle occurs, capacitors C_S and C_L are charged a little more, and some of the charge lost by C_D is restored. C_S and C_L accumulate charge continuously each cycle, but C_D does not. Eventually, the capacitors C_S and C_L are charged to the value of the peak signal voltage.

In other words, the DC component of voltage at node A with respect to ground becomes $-V_{peak}$ after a few cycles of the signal. If the DC level at node A is $-V_{peak}$, and the maximum positive swing of the signal is V_{peak} , the signal value at the anode of the detector diode will never go above zero volts. The signal at node A will swing up from $-V_{peak}$ just to zero volts and then down to $-2V_{peak}$. The resulting signal at node A is shown in **Figure 3**. If the signal at node A never goes above zero volts, the detector diode never conducts, and the signal delivered to the meter circuit is always zero. There may be a small "blip" on the meter when the detector is first connected and the RF signal applied, but that's about all the circuit of **Figure 3** will produce.

The signal as a square wave

It may be a little easier to understand the operation of the detector of **Figure 3** if we look at what happens if the signal is a square wave (see **Figure 4**). Suppose that when we start, all

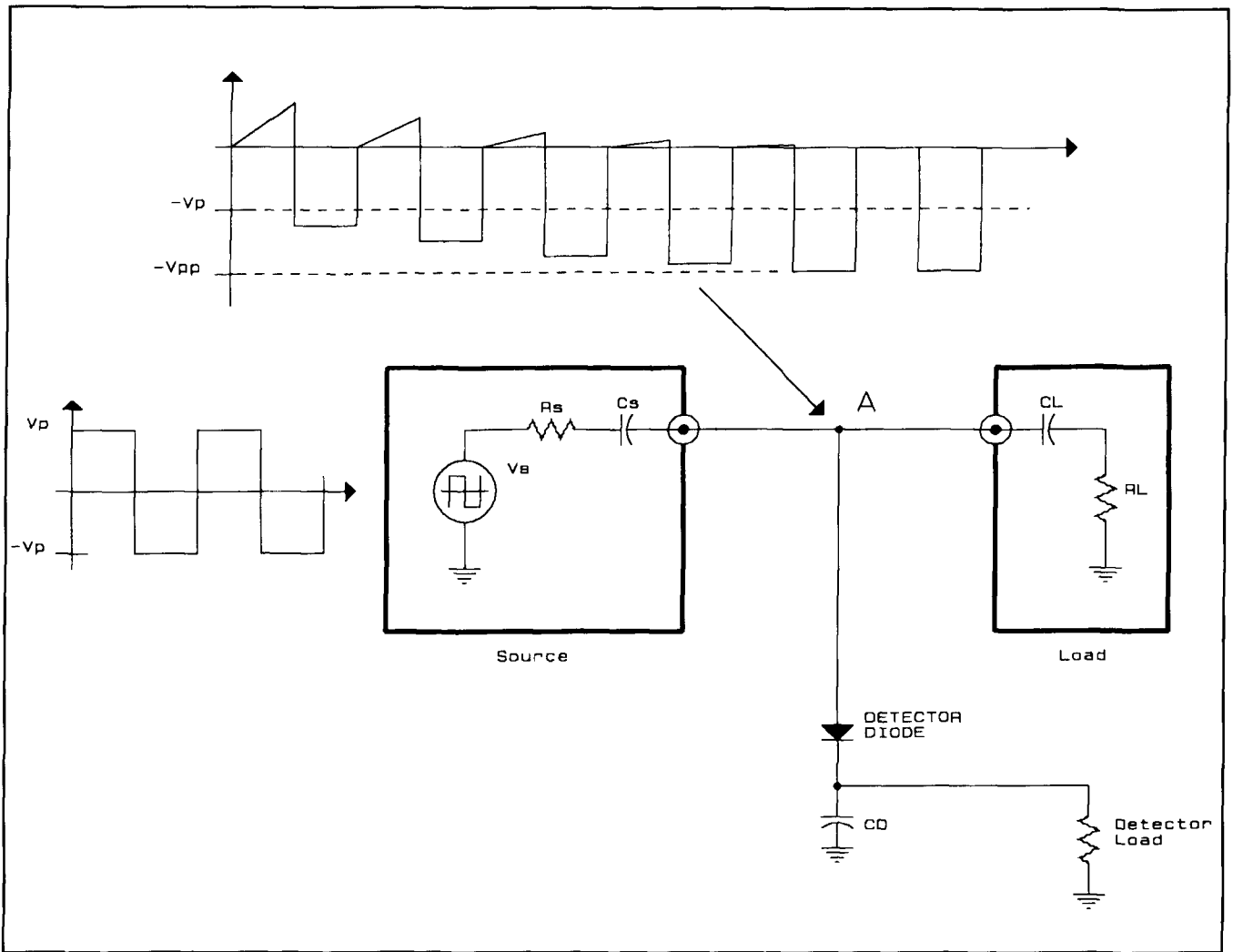


Figure 4. Signal clamping effect.

the capacitors are discharged. Also, suppose that we start the square-wave source on a positive cycle. At the beginning, the source voltage jumps up to its positive peak value, but the voltage at node A follows an R-C charging curve because capacitor C_D is being charged through R_S and the detector diode. Source capacitor C_S is also charged during this initial positive half cycle as is C_L . This charging continues as long as the source voltage remains high. In other words, this charging continues during the entire first positive half cycle.

During the first negative half cycle, the voltage at node A swings below ground, reverse biasing the detector diode. Therefore, no charging of C_D occurs during this negative half cycle. Because the meter resistance R_m is connected across C_D , C_D discharges through R_m during this negative half cycle. However, during the first negative half cycle, the DC charge on source capacitance C_S remains unchanged since there's nothing to charge or discharge this

capacitor. (Again, we are only concerned with the DC components of charge.)

On the second positive half cycle, the source again jumps up to the positive peak, but because some charge remains on C_S , the voltage at node A doesn't jump up as high above ground as it did on the positive portion of the first cycle. As a result, C_D is charged a little less this second cycle than it was on the first cycle, but some additional charge is delivered to C_S . This process continues as shown in **Figure 4** until the DC voltage at node A, the "average" voltage, is at $-V_{peak}$ and the signal at node A swings from ground down to $-2V_{peak}$ and back up just to ground. Since the signal at node A now never goes above ground, the detector diode is never forward biased, so no charge can be delivered to C_D . Because C_D isn't charged each cycle, it eventually discharges to zero (well almost) and the meter reads zero.

It actually takes many more cycles for the steady-state condition to occur; but even if it

takes 10,000 cycles, at HF this is only a few milliseconds. Also, the signal at node A actually rises a little above ground each positive half cycle—just enough to restore whatever charge is lost from C_S due to leakage during each negative half cycle. This may result in a very small average DC voltage on C_D perhaps a few millivolts for a 10- or 20-volt peak RF signal. This is not a detected signal. It's actually an error signal—an error caused by less than perfect circuit elements. However, this error signal does follow the RF signal level somewhat because the bigger the RF signal the bigger this error signal, more or less. However, it's not a very good representation of the RF level as it depends on the losses in the circuit and not on the actual circuit-element values. Still, you may have seen circuits like those of **Figures 3 and 4** actually used to measure RF power.

The detector-circuit configuration of **Figures 3 and 4** is called a “clamp” circuit because the voltage at node A is effectively clamped at ground. It only swings from ground in one direction—negative in this case—but if we were to reverse the diode, the voltage would be clamped positive. We weren't trying to make a clamp circuit, but that's what we ended up with. And that's why the detector configuration of **Figure 3** won't “detect” the RF signal at node A, at least not reliably or accurately.

Making the detector work properly

What can we do to make the detector work properly? One simple “fix” is to add a relatively small value of resistance to ground from node A. This will force the DC level at node A to stay at zero volts. The value of this resistance must be far smaller than the resistance of the metering circuit, perhaps 1,000 times smaller. If the metering circuit resistance is 100 k (typical if a sensitive D'Arsonval meter movement is used as the meter), about a 100-ohm resistor would be needed from node A to ground. This added resistance would result in loss of RF power due to its dissipation and would alter the loading on the source, which could result in poor VSWR. So if we wished to keep everything matched, we'd need to add even more resistive elements to provide impedance matching, which would result in even more power loss. We could also use an inductor to ground from node A, but with capacitors lurking in our circuit, we must be very careful when using inductors. If we want a true wide-bandwidth detector, say HF through UHF or even wider, it's very difficult to find inductors, particularly “big” inductors as would be needed at HF and that “look” like inductors

over that wide a frequency range. Also, the use of such an inductor would needlessly complicate the circuit.

There is a better way. If we look at the waveform that occurs at node A in **Figures 3 and 4**, the average value of the voltage, the DC component, is actually the “detected” signal we're looking for. The detected voltage happens to be negative in this case, but that's simply determined by the way we choose to connect the detector diode, as noted earlier. That DC value is directly related to the magnitude of the load voltage being delivered to the load, the AC RF voltage at node A. Specifically, the DC component of voltage at node A is equal to the peak value of the RF load voltage, and of course the peak value is related to the RMS value, and the RMS value is related to the power delivered to the load. So, if we know the peak value of the AC RF signal at node A, and we know the value of the load impedance (real and reactive parts), we know the RF power delivered to the load. So, our detector circuit of **Figure 3** is actually working properly, we're just looking at the wrong place for the detected signal.

Suppose that in the circuit of **Figure 3** we switch the positions of the detector capacitor C_D and the detector diode (see **Figure 5**). I also flipped the diode over to ground the diode anode to provide a positive average voltage at the diode cathode. Then, the AC signal at node A also appears at node B, but is clamped to ground as described above by the clamping action of the detector diode and charge storage of the detector capacitor C_D . In this circuit configuration, it's very easy to see that because the diode anode is grounded, the voltage at node B can't swing below ground—the diode clamps the waveform at ground. The waveform expected at node B is then as shown in **Figure 5**. Again, the average value or DC component of voltage appearing at node B is the detected value of the signal appearing at node A. All we need to do is filter out the AC RF component and measure the DC component.

The filtering of the meter circuit removes the AC component so that at node C in the meter circuit all that's present is the DC component of the signal appearing at node B, and this DC voltage is then applied to the meter. The DC component at node B is equal to the magnitude of the peak signal level occurring at node A (actually it's equal to one-half the magnitude of the peak-to-peak value of the signal at node A because there's no DC ground-return path at node A to reference DC measurements). So, by measuring the DC voltage at node B, we have a measurement of the peak-to-peak AC RF load voltage at node A. That's exactly what we were looking for! Even though the detector circuit of **Figure 5** may look backwards, and that of

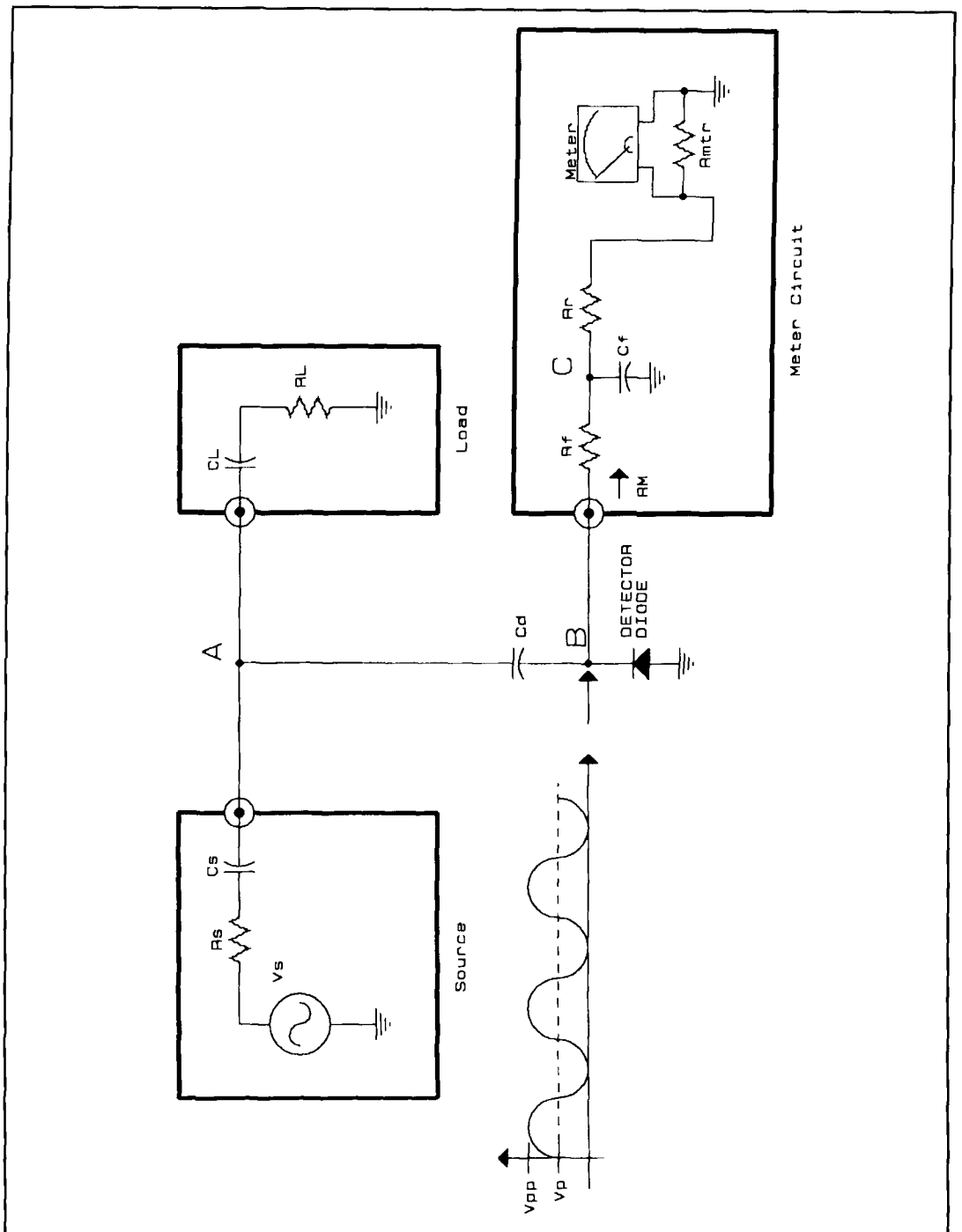


Figure 5. Correct diode-detector circuit.

Figure 3 may seem more obvious, Figure 5 shows the correct detector configuration.

The correct detector configuration

In a circuit such as Figure 5, where there are both source capacitance C_S and load capacitance C_L , the detector capacitor C_D isn't really needed because it's simply in series with the equivalent capacitance of the other two capacitors. But in a general circuit where we wish to

measure the RF signal, we probably won't really know the RF configuration at the point of measurement. If there's any DC component of voltage at the point where we wish to measure the RF signal, such as some resonator circuit in a transmitter for example, we must block that DC voltage and look only at the AC RF component of the signal. Generally, a detector capacitor, C_D , will always be used simply to block any DC component that may be present at the point of measurement. The value of this capaci-

tor isn't too critical. In general, it's only important that its capacitive reactance be much smaller than the resistance R_M of the meter circuit. For example, if R_M is 100 k, a reactance value for C_D of around 100 ohms would be quite adequate.

If we are working at the low end of the HF band, at say 3 MHz, the value of C_D must be greater than about 500 pF. At higher frequencies this capacitance may be smaller. We can make this capacitor as large as we wish. Making it large doesn't significantly increase the load on the point of measurement, because this capacitor is essentially in series with the meter-circuit resistance R_M . The lowest value of resistive loading that can occur at the point of measurement is then simply R_M (ignoring any parasitic effects of the circuit components). Therefore, C_D is generally chosen as some large value, such as 0.001 μ F or even as high as 0.1 μ F.

It's important to choose a reasonable-quality RF capacitor for C_D because any series inductance or shunt resistance will affect the measurement. Almost any RF chip capacitor will work very well, even to several GHz. The detector diode is actually the limiting element. At higher frequencies, the parasitic elements of the diode, such as shunt capacitance and series inductance, seriously affect the detection process. If an RF Schottky diode is used (not a "power" Schottky diode such as used in switch-mode power supplies), very good detection to several GHz is possible.

Note that in the simple diode RF detector it's important to ensure that the terminal of the detector diode opposite the detector capacitor has some DC path to ground (assuming we are metering with respect to ground as is generally the case); however, the detector capacitance doesn't require a ground path. As a result, the metering circuit is connected across the detector diode and not across the capacitor. If we could "float" the metering circuit, we wouldn't have to worry too much about what is grounded. However, in an RF environment, it's almost impossible to truly float anything; there's always some parasitic capacitance to "ground" —even if it's only the capacitance to free space. The best practice is to carefully design circuits with "proper" grounding.

Reviewing the five RF detectors

The circuit from the previous article¹ is included again here as **Figure 6** so we can review the five RF detectors in that circuit. Note that the first, third, and fifth detector diodes corresponding to the signals V_{sd} , V_{XZd} , and V_{Zd} , respectively, all have their anodes connected to ground. These detectors are con-

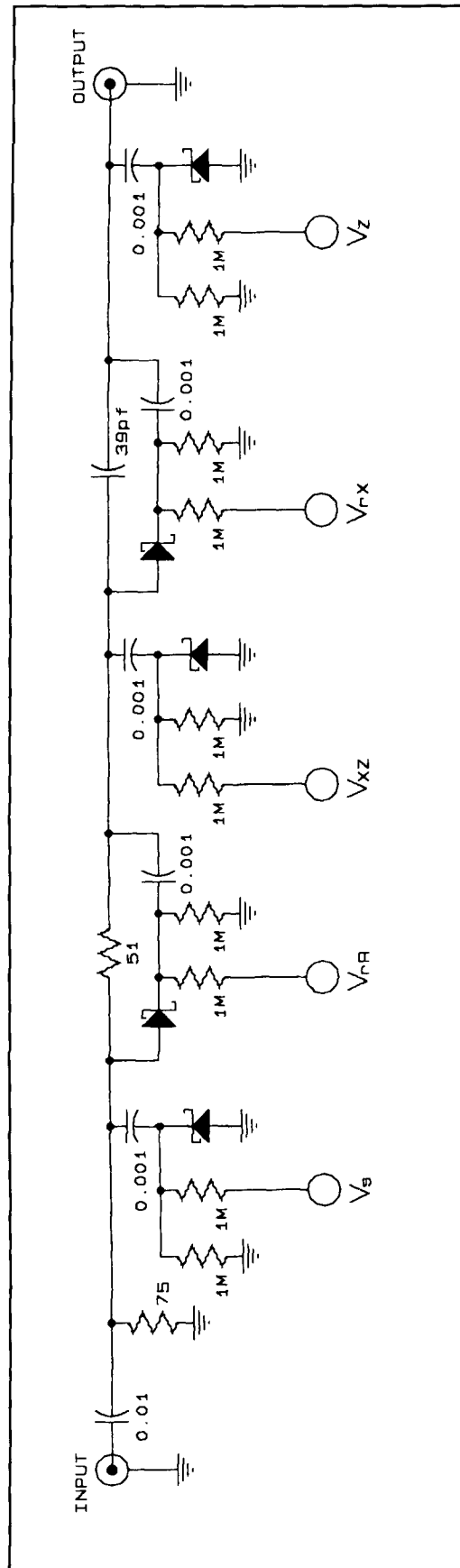


Figure 6. Complex impedance measurement-detector.

nected precisely as in **Figure 5**. The three detectors then detect the RF voltage between ground and the corresponding points in the circuit where their respective detector capacitors are connected. For example, the third detector corresponding to the signal V_{XZd} is detecting the RF voltage between ground and common point between the 51-ohm series resistor and the 39-pF series capacitor.

The second detector, providing the signal V_{rRd} , isn't connected directly to ground. Instead, the detector diode is provided a path to ground through the 75-ohm resistor at the input. This detector is detecting the RF signal appearing across the 51-ohm resistor because one side of that resistor is connected to the detector-diode anode and the other side of the resistor is connected to the detector capacitor. The DC component of the signal at the detector-diode cathode with respect to the diode anode is equal to the peak value of the AC RF signal across the 51-ohm resistor. However, the AC voltage at the diode anode is the AC RF signal at the circuit input. But because we are filtering out all the AC components of the signal and are only interested in the DC components, the added AC component due to the signal across the 75-ohm input resistor doesn't affect the DC measurement.

The important point to note in the second detector in **Figure 6** is that the "detected" signal is the DC component of voltage that appears across the detector diode. If there's no DC voltage across the 75-ohm input resistor, the diode anode will be at DC ground even though it will have some AC voltage at its anode. Since we are only interested in the DC component of the voltage of the detector circuit, the AC voltages aren't a problem as they are simply filtered out.

Similarly, the detector diode for the fourth detector, for the signal V_{rXd} , is provided a DC path to ground through the 51-ohm series resistor and the 75-ohm input resistor. Because this detector circuit is connected across the 39-pF capacitor, the DC component of the detected signal across the detector diode is the peak value of the RF signal appearing across the 39-pF capacitor. The anode of the detector diode is returned to ground through the 51 and 75-ohm resistors. The AC component at the diode anode is a combination of the AC RF voltages across the 51 and 75-ohm resistors, but again

since the AC components are filtered out, these have no effect on the detected DC signal.

Final observations

I have a couple of final observations that may be useful. The anodes of all the detectors in **Figure 6** must be at DC ground to provide accurate measurements of the corresponding RF signals detected. For example, in **Figure 6**, if the input 0.01- μ F capacitor were omitted and the source had some DC component of voltage, a measurement error would be introduced into the reading of the second and fourth detector circuits. This is because the detector-diode anodes wouldn't be properly DC referenced to ground, although the other three detector circuits wouldn't be affected. (Can you see why they wouldn't be affected based on the description above?) Any DC error at the input would add to the detected DC signal of the second and fourth detectors producing an error in the measurement.

The 0.01- μ F input capacitor is included simply to ensure that any DC component of the source is blocked. However, any DC component at the output will not affect detection in any of the detectors because such an output DC component is totally blocked by the circuit capacitances already present. No additional DC blocking capacitors are needed at the output. Finally, in the circuit of **Figure 6**, a small error is introduced by the 1-M resistors to ground in the second and fourth detector circuits. These resistors will result in a very small charging current to the 0.01- μ F input capacitor and the 39-pF series capacitor. However, the total effective resistance to ground due to these resistors is on the order of 250 k to 500 k depending on what type of metering circuit is used. The 75-ohm input resistor is far smaller than this combined resistance (more than 1,000 times smaller), so the DC error caused by these 1-M resistors is very small in the circuit of **Figure 6**.

Hopefully, this brief explanation of the basic diode RF detector has provided some understanding of the operation of this very simple, yet deceptive, and very useful circuit. ■

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1. Michael Gruchalla, P.E., "Complex Impedance Measurements," *Communications Quarterly*, Fall 1998, page 33.
2. Peter Dodd, G3LDO, "Technical Conversations: A Comment on 'Complex Impedance Measurements.'" *Communications Quarterly*, Spring 1999, page 3.

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A 20-Meter QRP Transceiver

By M.A. Chapman, KI6BP

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Fall 1999 issue, pages 15-30.

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IMPEDANCE MATCHING

Interpreting the virtual short circuit

The concepts of impedance matching in the steady-state condition are generally well known. However, the physical wave reflection mechanism through which the steady-state impedance match occurs remains a topic of continued discussion and some misinterpretation throughout much of the amateur radio and engineering communities.

When a transmitter is first energized, there will be an initial forward traveling wave delivered to the input of the transmission line that connects the transmitter to the matching network and the antenna. The level of wave voltage, current, and power in the initial forward traveling wave is a direct function of the characteristic impedance of the transmission line. At this point, no portion of the wave has reached the matching network or antenna and, therefore, no reflections exist within the system. *There is only the initial forward wave traveling in the transmission line.*

This initial forward traveling wave will arrive at the input to the matching network (or tuner). Typically, the matching network will be comprised of lumped circuit components in a T-network configuration. At this point, an initial reflected wave is created that travels backward towards the transmitter. The level of wave voltage, current, and power in the initial reflected wave is entirely dependent upon the initial reflection coefficient seen at the input to the matching network. The portion of the forward traveling wave that is not reflected travels through the matching network to the transmission line connecting the matching network to the antenna.

This initial forward traveling wave will undergo some level of attenuation and phase

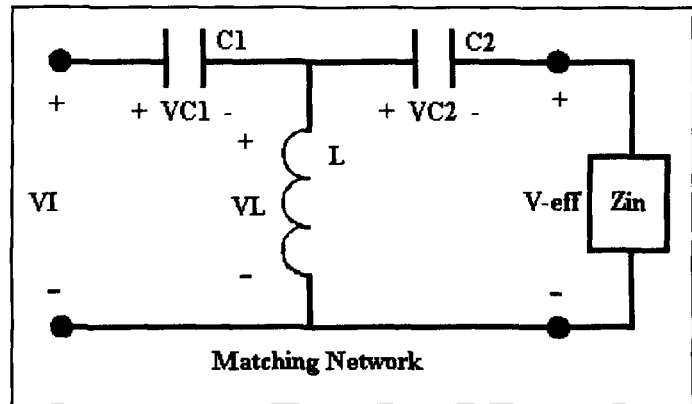


Figure 1. Equivalent circuit of the steady state.

shift due to the transmission line characteristics. When the initial forward traveling wave arrives at the antenna, a portion of it will be reflected back towards the matching network. The amount of wave voltage, current, and power reflected at the antenna is simply a function of the antenna's physical reflection coefficient.

Upon arriving back at the matching network output, a portion of the reflected wave is "re-reflected" back towards the antenna. Assuming the transmitter is still energized, the re-reflected wave will combine with the initial forward traveling wave to become a single forward traveling wave with a different level of forward traveling voltage, current, and power. The level of wave voltage, current, and power reflected for each incident wave arriving at the matching network output is simply a function of the matching network output's physical reflection coefficient. Depending on the value

of the incident voltage and the physical reflection coefficient at the matching network output, the re-reflected wave may combine with the initial forward traveling wave to either increase or decrease the forward traveling power. This process of reflection and re-reflection will continue until the system reaches the steady-state condition.

In the steady state, the total levels of forward traveling voltage and power will either increase or decrease from the initial levels of forward traveling voltage and power. The resulting steady-state forward traveling power in the transmission line is the vector combination of the initial forward traveling power and the re-reflected power developed at the matching network output. The effective net power delivered to the transmission line will be equal to the steady-state forward power minus the steady-state reflected power, both being determined at the input to the transmission line. The effective net power delivered to the transmission line input will be equal to the sum of the power dissipated in the transmission line plus the power delivered to the antenna.

As the system transitions to the steady state, a portion of the reflected wave incident at the matching network output is delivered back through the matching network to its input. When the system reaches the steady-state condition, the total wave delivered back to the matching network input is 180 degrees out of phase with the initial reflected wave at the matching network input, causing a complete cancellation of both waves. This eliminates any backward traveling wave(s) toward the transmitter, resulting in the steady-state impedance at the matching network input being equivalent to the characteristic impedance of the transmission line.

In an article entitled *Examining the Mechanics of Wave Interference in Impedance Matching*,¹ Walter Maxwell, W2DU, describes the impedance matching process in the steady state, including the concept that a "virtual short circuit" exists at the matching network. "*The cancellation of reflections at the matching point simultaneously creates a virtual, totally reflecting open or short circuit to all the reflected waves at the matching point in the network or transmission line.*"

This concept generally leads to the conclusion that all of the wave voltage, current, and power contained in the reflected wave(s) arriving at the matching network output are totally re-reflected back towards the antenna. In fact, *this is not physically the case*, and the concept of the virtual short circuit is really a steady-state equivalent circuit interpretation and not a representation of the true physical impedance matching process. Mr. Maxwell also presents a

more detailed discussion of these steady-state impedance matching concepts in his text *Reflections: Transmission Lines and Antennas*.²

The purpose of this article is to compare the concept of the virtual short circuit presented in **References 1 and 2** with the concept that the steady-state impedance match occurs as a result of reflection cancellation at the matching network input. This reflection cancellation is shown to occur as a direct result of some voltage being delivered back to the matching network input as a result of the reflections from the antenna arriving back at the matching network output. Additionally, it is shown that the voltage re-reflections occurring at the matching network output are a direct function of the impedance seen looking into the matching network output. The analysis and examples presented here include the condition where losses exist in the matching network components.

The concepts presented in this article are also applied to the more general case of a mismatched tuner or matching network. The reflection analysis techniques presented are shown to accurately determine the steady-state input impedance and reflected power at the matching network input under the condition where the network component values are chosen or set to mismatch the matching network input impedance relative to the transmission line characteristic impedance.

Finally, it is shown that, in any practical installation, a conjugate match condition does not physically exist at the connection point between the matching network output and transmission line. In fact, a conjugate match condition is not necessary to achieve a steady-state impedance match.

The steady state and the concept of the virtual short circuit

To provide a complete discussion of the impedance matching process that occurs with a lumped component matching network or tuner, it is necessary to examine the system's initial state and all of the multiple reflections occurring as the system transitions to the steady state. In any transmission line system, there is always an initial and a steady state for each time-independent wave arriving at the matching network input.

Before discussing the full details of the impedance matching process, we will first consider the equivalent circuit of the steady state as illustrated in **Figure 1**. In the steady state, the equivalent load impedance at the matching network output is the steady-state input impedance, Z_{IN} , of the transmission line connecting the matching network to the antenna. Defining

the steady-state input voltage to the matching network as V_I , the total steady-state voltage (V_{eff}) and power (P_{eff}) delivered to the impedance Z_{IN} can be determined using circuit theory.

Let's consider an example to illustrate the concepts of the steady-state condition and the virtual short circuit. All of the calculations discussed in this example are illustrated in a Mathcad document presented in **Appendix A**. Let's assume that the antenna impedance Z_A is equal to $6.2 + j14.4$ ohms. Let's also assume that the transmission line connecting the antenna to the matching network is RG-213 coaxial cable. Finally, let's assume that Z_O is $50 + j0$ ohms, the operating frequency is 21.2 MHz, the matched transmission line attenuation is 1.05 dB/100 feet and the transmission line is 100 feet long. Using simple transmission line concepts,³ the input impedance to the transmission line, Z_{IN} , is found to be $58.32 - j85.85$ ohms. Note that RG-213 coaxial cable has a velocity factor of 0.6594.

Using circuit theory, suitable values for $C1$, $C2$, and L can be determined for matching Z_{IN} to the transmission line's characteristic impedance, Z_O . Assuming lossy matching components, with a capacitor Q of 400 and an inductor Q of 75, the following values for $C1$, $C2$, and L were determined. $C1 = 36.521$ pF, $C2 = 50$ pF and $L = 0.874$ μ H.

Assuming the steady-state voltage delivered to the matching network, V_I , is 70.711 volts, the steady-state voltages at each of the matching network components, and at Z_{IN} , can be determined. Also note that in a 50-ohm system, the voltage V_I represents 100 watts of power. In the steady state, the voltage V_{C1} is 290.71 <89.86 volts, the voltage V_{C2} is 184.62 <62.64 volts, the voltage V_L is 299.01 <76.46 volts, and the voltage V_{eff} is 127.61 <96.69 volts. The steady-state power delivered to $C1$ is 1.03 watts, to $C2$ is 0.57 watts, and to L is 10.24 watts. The steady-state power delivered to the load impedance, Z_{IN} , is 88.17 watts. The total power lost in the matching network is 11.83 watts, which represents a total matching network loss of 0.55 dB.

The 88.17 watts of power delivered to the impedance Z_{IN} is actually the steady-state effective net power delivered to the transmission line. It is not the steady-state total forward power delivered to the transmission line. Additionally, it is not the initial forward power delivered to the transmission line. This second concept will be addressed later. If a wattmeter is located at the input to the transmission line, the difference between the measured steady-state total forward power and the measured steady-state total reflected power will be 88.17 watts. It is now necessary to relate the effective net power delivered to the transmission line to

the multiple reflection and re-reflection process that occurs between the antenna and the matching network output. In establishing this relationship, the concept of the virtual short circuit is developed.

There is an important consideration associated with establishing the correct physical relationship between the steady-state power delivered to Z_{IN} and the multiple reflection process that occurs between the antenna and the matching network. By defining the steady-state load impedance at the matching network output to be Z_{IN} , the multiple reflections that occur between the antenna and the matching network have already been taken into account. In any transmission-line system, the input impedance to the transmission line will be Z_O prior to the point when reflections from the antenna arrive back at the transmission line input. Only after all of the multiple reflections have been taken into account will the input impedance to the transmission line be the steady state impedance Z_{IN} . Beginning with steady-state impedance Z_{IN} to analyze the reflection process, we are actually beginning with the solution and working the problem backwards.

From the steady-state equivalent circuit definition, the power delivered to Z_{IN} is the effective net power delivered to the transmission line connecting the matching network to the antenna. Therefore, the steady-state power delivered to Z_{IN} must be the sum of the power dissipated in the transmission line and the power delivered to the antenna. Any reflection analysis that begins by using the effective net power delivered to Z_{IN} must, under this condition, conclude that none of this power is returned to or delivered back through the matching network. The only apparent conclusion with this approach to a reflection analysis is that the matching network must re-reflect all of the power that is reflected by the antenna. However, total power reflection at any impedance junction will only occur if the reflection

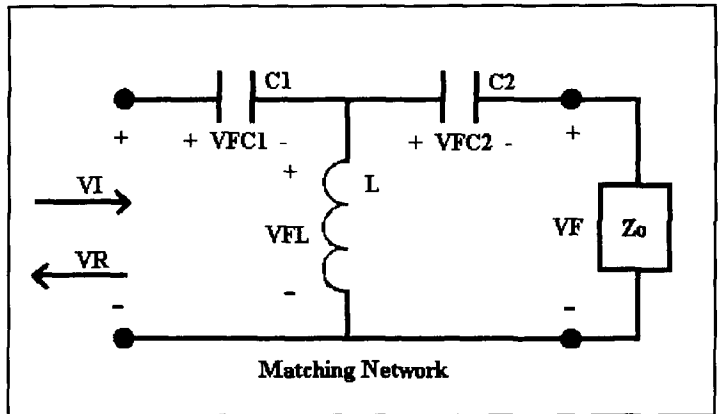


Figure 2. Equivalent circuit of the initial state.

coefficient at the junction has a magnitude of 1. This will never be the case with any practical matching network configuration.

Additionally, by the same steady-state equivalent circuit definition, the total voltage developed across Z_{IN} , or the input to the transmission line, is the voltage V_{eff} , as shown in **Figure 1**. If the voltage V_{eff} is used as the basis for beginning a reflection analysis, the total steady-state voltage developed across the matching network output must still be V_{eff} . With any reflection analysis, the total voltage at the matching network output will be V_{eff} plus the summation of the incident and re-reflected voltages developed at the matching network output due to the reflections from the antenna. Therefore, the summation of the incident and re-reflected voltages developed at the matching network output must be zero. The only mechanism through which this could occur is if the matching network output had an equivalent reflection coefficient of -1 , representing an equivalent short circuit. This would lead to the conclusion that the matching network output presents a virtual short circuit to the waves that are reflected by the antenna.

There is a significant inconsistency to consider between using either the power or voltage delivered to Z_{IN} as the basis for beginning a reflection analysis. If we were to assume that the power delivered to Z_{IN} was the "initial" forward power delivered to the transmission line, we would define the "initial" forward power to be the 88.17 watts determined above. If we were to assume that V_{eff} was the "initial" forward voltage delivered to the transmission line, we would then determine the "initial" forward power delivered to the transmission line to be 325.68 watts ($|V_{eff}|^2/Z_0$). This is an important difference to consider because both were determined using the equivalent steady-state load impedance, Z_{IN} . In reality, neither the above power nor the above voltage can be correctly defined as "initial" values. This will be illustrated further in subsequent sections.

The reflection analysis presented in **References 1 and 2** first assumes that the initial forward power delivered to the transmission line is the effective net power delivered to the transmission line, P_{eff} , which is equal to 88.17 watts. Continuing with the reflection analysis, we then need to determine the total forward power delivered to the transmission line, the power delivered to the antenna, the power reflected at the antenna and the power returned to or arriving back at the matching network output.

The important issue is how to determine the amount of power reflected at the matching network output. The impedance looking back into the matching network output is $71.13 + j77.82$ ohms. Obviously, this impedance is neither a

short circuit, an open circuit, or purely reactive and, therefore, the magnitude of the physical reflection coefficient is not 1. However, if we were to use the calculated reflection coefficient ($|\rho| = 0.56$) to determine the amount of power reflected at the matching network output, we will arrive at an incorrect result for the total power delivered to the antenna. This is a result of assuming an incorrect level of initial forward power delivered to the transmission line.

If we **assume** that the magnitude of the reflection coefficient at the matching network is 1, therefore assuming that all power is reflected back towards the antenna, we will determine the correct total steady-state power delivered to the antenna. Note that this assumption is necessary since we began the analysis using the effective net power delivered to the transmission line. The total steady-state power delivered to the antenna is 41.77 watts. We will also correctly determine the total steady-state forward power delivered to the transmission line (144.4 watts) and the total steady-state reflected power arriving at the matching network output (56.23 Watts). Note that the effective power delivered to the transmission line is 88.17 watts (144.4 to 56.23 watts). Because this analysis yields the correct results, the impedance presented by the matching network output is concluded to be a virtual short circuit. Note again that the calculations used to arrive at these results are presented in **Appendix A**.

Another view of the system power delivery, which supports the concept of the virtual short circuit, follows. The effective net power delivered to Z_{IN} ($P_{eff} = 88.17$ watts), plus the reflected power arriving at the matching network output from the antenna (56.23 watts), is equal to the total forward power delivered to the transmission line (144.4 watts). From this, one could conclude that all of the reflected power arriving at the matching network output is re-reflected back toward the antenna. In reality, however, the effective net power delivered to the transmission line is not the beginning of the power delivery process, it is the end result of the power delivery process.

The above power analysis of the steady-state condition results in determining the correct levels of power delivered to the antenna and dissipated in the transmission line. However, it is very important to note that this analysis begins with the solution (the steady state) to start the analysis and then arrives at the solution as the answer. The assumption necessary to tie this analysis together is that the reflection coefficient at the matching network output has a magnitude of 1. When considering only the steady-state condition, this is the only conclusion that would apparently explain power delivery and reflection between the matching net-

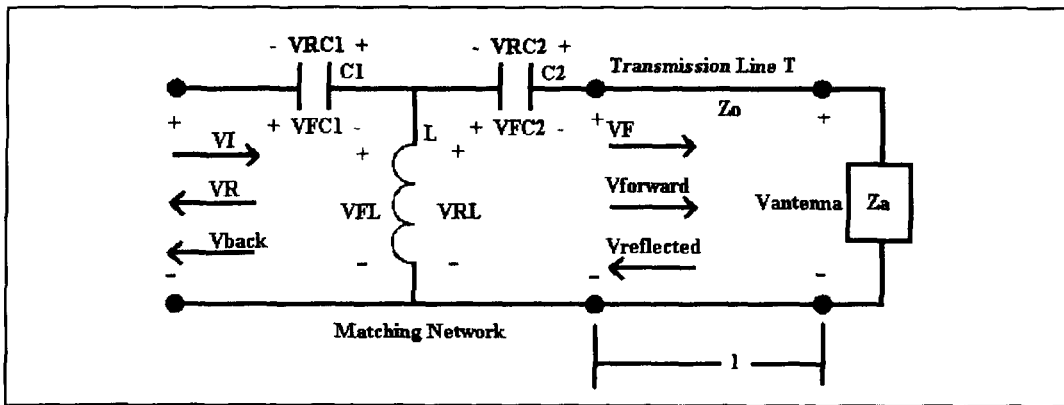


Figure 3. The equivalent circuit of the multiple reflection process.

work and the antenna. However, this entire analysis is derived from the end result of the matching process, not the beginning of the matching process.

The concept of the virtual short circuit and the resulting power analysis is really just an equivalent circuit interpretation or explanation of the steady-state condition. This power analysis is simply a convenient method to determine the steady-state power delivered to the antenna when the steady-state effective net power delivered to the input of the transmission line is known. The analysis does not describe the true physical mechanism through which power is reflected at the matching network output, nor does it describe the correct reflection process through which the steady-state impedance match occurs.

The next section of this article will detail the actual method by which the steady-state impedance match is achieved and the method by which the effective net power, P_{eff} , is delivered to the transmission line. Demonstrating that they can be used to correctly determine the total power loss in the matching network further validates the concepts presented here.

The initial state and the multiple reflection process

To fully describe the actual physical multiple reflection process that occurs between the antenna and the matching network, and then subsequently determine the effective net power delivered to the transmission line and the antenna, we must begin by considering the system's initial state. The equivalent circuit of the initial state is illustrated in **Figure 2**. Note that the initial input impedance to the transmission line is the transmission line's characteristic impedance, Z_0 .

To correctly determine the initial forward voltage (VF) and power delivered to the input

of the transmission line, we must first determine the initial input impedance seen at the matching network input. All of the following calculations are presented in a Mathcad document in **Appendix B**.

The initial input impedance to the matching network is determined to be $188.72 + j25.16$ ohms, which has a reflection coefficient equal to $0.59 + j0.04$ ($|\rho| = 0.587$). Assuming that the initial voltage incident at the matching network, V_I , is 70.711 volts, there will be a reflected voltage, V_R , developed at the matching network input equal to 41.53 <4.26 volts. Voltage V_R is equal to the voltage V_I multiplied by the reflection coefficient determined above. The total initial voltage applied across the input to the matching network will be equal to the vector sum of V_I plus V_R . This total voltage is equal to 112.17 <1.58 volts. The initial power reflected at the matching network input is calculated to be 34.496 watts ($100 * |\rho|^2$).

Using the total initial voltage applied across the input to the matching network (V_I plus V_R), we can determine the initial forward voltage at each of the matching network components and at the input to the transmission line. These voltages are defined as V_{FC1} , V_{FL} , V_{FC2} , and V_F . The initial forward voltage delivered to the transmission line, V_F , is 55.38 <116.23 volts. The initial forward power delivered to the transmission line is 61.34 watts. Note that the initial forward power delivered to the input of the transmission line is not equal to the effective net steady-state power delivered to the transmission line, 88.17 watts, as calculated in the previous section.

To complete the analysis of the multiple reflection and re-reflection process occurring between the antenna and the matching network, we will refer to **Figure 3**. The objective, now that V_F is known, is to determine the steady-state voltage developed at the antenna ($V_{antenna}$), the steady-state voltages developed at the matching network output ($V_{forward}$ and

Vreflected), and, finally, the steady-state voltage delivered back to the matching network input (V_{back}). Once these voltages are known, the power delivered to the antenna and the total steady-state forward, reflected, and effective net power at the input to the transmission line can be determined. Finally, the backward traveling power developed at the matching network input can also be determined.

For the impedance matching network to “work,” this analysis must demonstrate that the steady-state backward traveling power developed at the matching network input is equal in magnitude, but 180 degrees out of phase with the initial power reflected at the matching network input. For this to occur, V_{back} must be the negative of V_{R} . In this case, all power traveling backward towards the transmitter will be canceled, resulting in the steady-state matched condition.

To determine the total power loss occurring within the matching network, the total voltage and current at each matching network component must be determined. At each component within the matching network, the total voltage will be equal to the initial forward voltage at the component plus the vector sum of the steady-state backward voltage developed at the component. In the case of the series capacitors, the total voltage will be equal to the initial forward voltage minus the steady state backward voltage. In the case of parallel inductor, the total voltage will be equal to the initial forward voltage plus the steady-state backward voltage. To understand these statements, it is necessary to review the voltage polarities as illustrated in **Figure 3**.

The initial forward voltage at each component is developed from the initial voltage applied across the matching network input (V_{I} plus V_{R}). The steady-state backward voltage at each of the matching network components is developed from the steady-state voltage applied at the matching network output due to the multiple reflections occurring between the antenna and the matching network. This voltage will be equal to V_{forward} plus $V_{\text{reflected}}$.

The steady-state voltages at the antenna and at the matching network output are determined from the summation of all incident and reflected voltages at each. To determine these voltages, it is necessary to consider the reflection and re-reflection process beginning with the initial forward voltage delivered to the transmission line (V_{F}). The initial voltage, V_{F} , will travel toward the antenna undergoing an appropriate attenuation and phase shift. After arriving at the antenna, a reflected voltage will be created. This reflected voltage will be equal to the incident voltage at the antenna multiplied by the antenna’s reflection coefficient. The

reflected voltage will travel back towards the matching network, also undergoing an appropriate attenuation and phase shift.

Upon arriving at the matching network output, a re-reflected voltage will be created. The level of re-reflected voltage will be equal to the incident voltage multiplied by the reflection coefficient seen at the matching network output. Assuming the transmitter is still energized, the re-reflected voltage will combine with the initial voltage V_{F} to become a new forward traveling voltage. The level of forward traveling power will either increase or decrease depending upon the vector summation of these voltages. This process of reflection and re-reflection will continue until the system reaches the steady-state condition.

It is important to note that the reflection coefficient at the matching network output is not assumed to be -1 . Rather, the physical reflection coefficient of the impedance looking into the matching network is used in all calculations. In this example, the impedance looking into the matching network output is $71.13 + j77.82$ ohms. The reflection coefficient at the matching network output is equal to $0.42 + j0.38$ ($|\rho| = 0.56$).

Using transmission line theory,³ the total voltage at both the antenna and the matching network output are determined considering all of the multiple voltage reflections occurring between the two. The total steady-state voltage developed at the antenna is found to be $40.695 < 69.68$ volts, resulting in a steady-state power of 41.77 watts being delivered to the antenna. The total steady-state forward power delivered to the transmission line is 144.4 watts. The total reflected power at the transmission line input is 56.23 watts. From these levels of forward and reflected powers at the matching network output, the steady-state effective net power delivered to the transmission line is 88.17 watts ($144.4 - 56.23$). These power levels are consistent with the steady-state power levels calculated using the equivalent steady-state circuit analysis. Note that the 88.17 watts of effective net power delivered to the transmission line is a result of the total steady-state forward and reflected power developed at the input of the transmission line.

At this point, it is significant to note that the total forward power developed at the input to the transmission line increased from the initial level of 61.34 watts to the steady-state level of 144.4 watts. In this particular instance, the re-reflected voltage(s) at the matching network output combined with the initial forward traveling voltage such that the total forward traveling voltage increased. As a result, with each re-reflection, the forward traveling power increased. In some instances, the forward trav-

eling power may actually decrease from the initial level as a result of the re-reflections.

The total steady-state voltage developed across the matching network output is the vector sum of the initial forward voltage developed across the network output (VF) plus the sum of all incident (Vreflected) and reflected (Vforward) voltages developed across the network output. This total voltage is 127.61 <96.69 volts. This is identical to the steady-state voltage developed across Z_{IN} as determined using the equivalent steady-state circuit of **Figure 1**.

At the matching network output, the component of the total voltage developed as a result of the multiple reflection process occurring between the matching network and the antenna is the vector sum of the total incident and re-reflected voltages. This voltage is equal to Vforward plus Vreflected. This is the steady-state voltage applied across the matching network output and delivered back to the matching network components.

Knowing this voltage, the backward traveling voltages developed across the matching network components and across the matching network input can be determined. The backward traveling voltage at the matching network input, Vback, is 41.53 <-175.74 volts. Therefore, the backward traveling power delivered to the matching network input is 34.495 watts. The voltage Vback is indeed found to be the negative of the initial reflected voltage VR. The backward-traveling steady-state voltage and power at the matching network input are 180 degrees out of phase with the initial reflected voltage and power. This results in a total cancellation of both voltages resulting in the steady-state matched impedance at the matching network input.

The total voltages developed across each of the matching network components are VC1 = 290.71 <-89.86 volts, VC2 = 184.17 <62.64 volts, and VL = 299.01 <76.46 volts. The power delivered to C1 is 1.03 watts, to C2 it is 0.57 watts, and to L it is 10.24 watts. The total power lost in the matching network is 11.83 watts. This is identical to the calculations for power lost in the network determined using the steady-state equivalent circuit of **Figure 1**.

The above example began by first considering the transmission system's initial state and then determining the system's steady-state conditions by performing a complete reflection analysis. This reflection analysis used the physical reflection coefficients at both the antenna and the matching network to determine the level of reflected voltage. The analysis made no assumptions regarding any of the steady-state conditions. Additionally, the detailed reflection analysis provided a complete description of

how the steady-state match is achieved by illustrating the cancellation of all reflections at the matching network input.

The general impedance case: a mismatched tuner or matching network

Each of the previous examples considered a matching network designed specifically to match the steady-state transmission-line input impedance Z_{IN} to the characteristic impedance Z_0 . To further validate the multiple reflection analysis concepts presented in this article, a more general case of an impedance matching network is considered. In this example, the values of C1, C2, and L are selected such that the steady-state input impedance to the matching network is not matched to Z_0 . As an example, C1 is arbitrarily chosen to be 40 pF, C2 is chosen to be 80 pF, and L is chosen to be 1.0 μ H. The Q of the capacitors is assumed to be 400 and the Q of the inductor is assumed to be 75.

Using the previously calculated Z_{IN} of 58.32 - j85.85 ohms and the equivalent steady-state circuit of **Figure 1**, the steady-state input impedance at the matching network input is now calculated to be 190.55 + j82.78 ohms. This represents a VSWR of 4.574:1 at the matching-network input.

The objective is to verify this input impedance using the multiple reflection process discussed in the previous section and the calculation methods presented in **Appendix B**. To do this, the total voltage and current at the impedance-network input must be determined.

Referring to **Figure 3**, the total steady-state voltage at the matching-network input will be equal to applied voltage V1 plus the initial reflected voltage VR plus the backward traveling voltage Vback. The voltage Vback is the voltage delivered through the network as a result of the multiple reflections incident at the network output. The total current at the matching network input will be equal to the forward traveling current due to the applied voltage V1 minus the backward traveling current due to the voltages VR and Vback.

The calculations necessary to determine the total voltage and current at the matching network input are illustrated in **Appendix C**. These calculations are similar to those presented in **Appendix B** with the added step (Step 10) for determining the steady-state impedance at the network input. Performing a full voltage reflection analysis to determine both VR and Vback, the input impedance at the matching network input is indeed verified as 190.55 +

j82.78 ohms. In this case, V_{back} is not equal to the negative of V_R .

The total reflected power at the matching-network input is determined from the vector sum of V_R plus V_{back} . This total reflected power is 41.11 watts and is consistent with 100 watts of forward incident power at a load with a VSWR equal to 4.574:1. The concepts and reflection analysis techniques presented in this article remain valid in the case where the matching network is intentionally mismatched.

Comments on the existence of the "conjugate" match

In many discussions of impedance matching, the concept of the "conjugate" match is presented.² The significance of the conjugate match condition is straightforward. With any constant voltage source, having a source impedance Z_S , maximum power will be delivered to a load with an impedance of Z_S^* . Unfortunately, most amateur operators will not have any knowledge of their transmitter's source impedance Z_S , nor will they generally have the luxury of creating an impedance equal to Z_S^* at the input to their transmission line.

In the context of the use of tuners and impedance matching networks, the conjugate match condition occurs when the impedance looking into the tuner output is the conjugate of the steady-state input impedance of the transmission line connecting the antenna. In reality, this condition will never exist and is not even necessary for the steady-state impedance match to occur. In the matched examples presented in this article (**Appendices A and B**), the steady-state impedance at the input to the transmission line was $58.3 - j85.8$ ohms, while the impedance looking into the matching network output was $71.1 + j77.8$ ohms. These two impedance values are clearly not conjugates of one another.

In order for a conjugate impedance situation to occur with the use of a tuner, two conditions must exist: First, the impedance looking back towards the transmitter from the tuner input must be exactly the Z_O of the transmission line; and, second, the tuner must have no (0 dB) loss. In practice, these two conditions are not likely to occur. When used in a discussion of tuners or impedance matching networks, the existence of a "conjugate" match is simply a special theoretical condition similar to a discussion involving the existence of lossless transmission lines.

Summary

This article has presented a discussion of the virtual short circuit concept described by

Walter Maxwell.^{1,2} The concept of the virtual short circuit at a matching network or tuner is the result of a steady-state equivalent circuit analysis of the multiple reflections that occur between the matching network and the antenna. The virtual short circuit is an equivalent circuit concept, but it does not represent the actual physical reflection process that occurs between the matching network and the antenna. Additionally, it does not describe the actual mechanism through which the steady-state impedance match occurs. It does however allow determination of the power delivered to the antenna when considering the effective net steady-state power delivered to the transmission line.

This article has also presented a discussion of the actual physical reflection process occurring between the matching network and the antenna. It was shown that re-reflections occurring at the matching network output are a direct function of the physical reflection coefficient at the matching network output. It was also shown that not all of the incident voltage, and hence power, is reflected back towards the antenna. In fact, in order for the steady-state impedance match to occur, it is necessary that some voltage be delivered back through the matching network to its input. Using a detailed voltage reflection analysis, it was illustrated that the steady-state impedance match occurs as the result of a reflection cancellation at the matching network input. The detailed voltage reflection analysis presented in this article correctly determines the following:

- The correct steady-state voltage and power delivered to the antenna.
- The correct steady-state forward and reflected power developed at the matching-network output.
- The correct effective net power delivered to the transmission line connecting the matching network to the antenna.
- The correct total steady-state voltage and impedance developed across the input to the transmission line connecting the matching network to the antenna.
- The correct total voltage and power delivered to each component in the matching network.
- The correct steady-state input impedance at the matching network input.
- The correct reflected power at the matching network input.

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2. M. Walter Maxwell, W2DU, *Reflections: Transmission Lines and Antennas*, American Radio Relay League, Newington, Connecticut, 1990.
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**Appendix A. Steady State Equivalent Circuit Power Analysis.
Reference Figure 1.**

1. Define the known parameters:

Antenna Impedance: $Z_a := 6.2 + j \cdot 14.4$ Ohms
 Operating Frequency: $F := 21.2 \cdot 10^6$ Hz
 Characteristic Impedance: $Z_o := 50$ Ohms
 Transmission Line Length: Length := 100 Feet
 Transmission Line Properties: $\alpha_{dB} := 1.05$ dB / 100 Feet
 $V_p := 0.6594$

$$\lambda := \frac{2.997925 \cdot 10^8}{F} \quad \omega := 2 \cdot \pi \cdot F \quad l := \text{Length} \cdot 0.3048 \text{ meters}$$

$$\alpha := \frac{\alpha_{dB}}{868.59} \cdot 3.28084 \text{ nepers/meter} \quad \beta := \frac{2 \cdot \pi}{V_p \cdot \lambda} \quad \gamma := \alpha + i \cdot \beta$$

2. Calculate the steady state input impedance (Zin) to the transmission line connecting the antenna to the matching network.

$$Z_{in} := Z_o \cdot \frac{\frac{Z_a}{Z_o} + \tanh(\gamma \cdot l)}{1 + \frac{Z_a}{Z_o} \cdot \tanh(\gamma \cdot l)} \quad Z_{in} = 58.32 - 85.85j \quad \text{Ohms}$$

3. Define suitable matching network component values for C1, C2 and L.

$$Q_L := 75 \quad Q_C := 400$$

$$C_1 := 36.520911 \cdot 10^{-12} \text{ F} \quad C_2 := 50 \cdot 10^{-12} \text{ F} \quad L := 0.873927 \cdot 10^{-6} \text{ H}$$

$$R_1 := \frac{1}{\omega \cdot C_1 \cdot Q_C} \quad R_2 := \frac{1}{\omega \cdot C_2 \cdot Q_C} \quad R_L := \frac{\omega \cdot L}{Q_L}$$

$$Z_{C1} := R_1 + \frac{-j}{\omega \cdot C_1} \quad Z_{C2} := R_2 + \frac{-j}{\omega \cdot C_2} \quad Z_L := R_L + j \cdot \omega \cdot L$$

4. Calculate the steady state input impedance to the matching network:

$$Z_1 := Z_{in} + Z_{C2} \quad Z_2 := \frac{1}{\left(\frac{1}{Z_1} + \frac{1}{Z_L}\right)}$$

$$Z_m := Z_2 + Z_{C1}$$

$$Z_m = 50 \quad \text{Ohms}$$

5. Calculate the power delivered to Z_{in} and the power lost in the matching network:

Define the initial voltage driving the matching network input:

$$V_1 := 70.710678 \quad \text{Volts}$$

Calculate the steady state reflection properties at the matching network input:

$$\rho_m := \frac{\frac{Z_m}{Z_0} - 1}{\frac{Z_m}{Z_0} + 1} \quad \rho_m = 0 \quad \text{VSWR} := \frac{1 + |\rho_m|}{1 - |\rho_m|} \quad \text{VSWR} = 1$$

Calculate the reflected and total voltage at the matching network input:

$$V_{\text{reflected}} := \rho_m \cdot V_1 \quad V_{\text{total}} := V_1 + V_{\text{reflected}} \quad V_{\text{total}} = 70.71 \quad \text{Volts}$$

Use simple circuit theory to determine the voltage V_{eff} , delivered to the steady state impedance Z_{in} :

$$I_{\text{total}} := \frac{V_{\text{total}}}{Z_m} \quad V_{C1} := I_{\text{total}} \cdot Z_{C1} \quad V_2 := V_{\text{total}} - V_{C1}$$

$$I_L := \frac{V_2}{Z_L} \quad V_L := I_L \cdot Z_L \quad I_2 := I_{\text{total}} - I_L \quad V_{C2} := I_2 \cdot Z_{C2}$$

$$V_{C1} = 0.73 - 290.71j \quad \text{Volts} \quad |V_{C1}| = 290.71 \quad \arg(V_{C1}) \cdot \frac{180}{\pi} = -89.86$$

$$V_{C2} = 84.84 + 163.97j \quad \text{Volts} \quad |V_{C2}| = 184.62 \quad \arg(V_{C2}) \cdot \frac{180}{\pi} = 62.64$$

$$V_L = 69.98 + 290.71j \quad \text{Volts} \quad |V_L| = 299.01 \quad \arg(V_L) \cdot \frac{180}{\pi} = 76.46$$

$$V_{\text{eff}} := I_2 \cdot Z_{in}$$

$$V_{\text{eff}} = -14.86 + 126.74j \quad \text{Volts} \quad |V_{\text{eff}}| = 127.61 \quad \arg(V_{\text{eff}}) \cdot \frac{180}{\pi} = 96.69$$

Calculate the steady state power delivered to Zin.

$$\theta := \arg(V_{\text{eff}}) - \arg(I_2)$$

$$P_{\text{eff}} := |V_{\text{eff}}| \cdot |I_2| \cdot \cos(\theta) \quad P_{\text{eff}} = 88.17 \quad \text{Watts}$$

Calculate the power lost in each of the matching network components.

$$I_{C1} := \frac{V_{C1}}{Z_{C1}} \quad \theta_{C1} := \arg(Z_{C1}) \quad P_{C1} := |V_{C1}| \cdot |I_{C1}| \cdot \cos(\theta_{C1})$$

$$I_{C2} := \frac{V_{C2}}{Z_{C2}} \quad \theta_{C2} := \arg(Z_{C2}) \quad P_{C2} := |V_{C2}| \cdot |I_{C2}| \cdot \cos(\theta_{C2})$$

$$I_L := \frac{V_L}{Z_L} \quad \theta_L := \arg(Z_L) \quad P_L := |V_L| \cdot |I_L| \cdot \cos(\theta_L)$$

$$P_{C1} = 1.03 \quad \text{Watts} \quad P_{C2} = 0.57 \quad \text{Watts} \quad P_L = 10.24 \quad \text{Watts}$$

Calculate the total power lost in the matching network.

$$P_{\text{lost}} := P_{C1} + P_{C2} + P_L \quad P_{\text{lost}} = 11.83 \quad \text{Watts}$$

6. Calculate the total power delivered to the antenna. Also, calculate the total forward and reflected power at the matching network output as well as the total power lost in the transmission line.

Calculate the antenna reflection properties:

$$\rho_a := \frac{\frac{Z_a}{Z_0} - 1}{\frac{Z_a}{Z_0} + 1} \quad \rho_a = -0.67 + 0.43j \quad |\rho_a| = 0.79$$

Define the reflection properties of the matching network (virtual short circuit):

$$\rho_N := -1 \quad |\rho_N| = 1$$

Determine the total power delivered to the antenna due to the multiple power reflections between the antenna and the matching network.

$$k := \alpha \text{dB} \cdot \frac{\text{Length}}{100.} \quad \alpha := 10^{\frac{-k}{10}}$$

$$PA := \alpha \cdot P_{\text{eff}} \cdot [1 - (|\rho_a|)^2]$$

$$P_{\text{antenna}} := PA + P_{\text{eff}} \cdot [1 - (|\rho_a|)^2] \cdot \sum_{n=1}^{1000} \alpha^{2 \cdot n + 1} \cdot (|\rho_a|)^{2 \cdot n} \cdot (|\rho_N|)^{2 \cdot n}$$

$$P_{\text{antenna}} = 41.77 \quad \text{Watts}$$

Determine the total forward power at the matching network output:

$$P_{\text{forward}} := P_{\text{eff}} + P_{\text{eff}} \cdot \sum_{n=1}^{1000} \alpha^{2 \cdot n} \cdot (|\rho_a|)^{2 \cdot n} \cdot (|\rho_N|)^{2 \cdot n}$$

$$P_{\text{forward}} = 144.4 \quad \text{Watts}$$

Determine the total reflected power at the matching network output:

$$P_{\text{reflected}} := P_{\text{eff}} \cdot \sum_{n=1}^{1000} \alpha^{2 \cdot n} \cdot (|\rho_a|)^{2 \cdot n} \cdot (|\rho_N|)^{2 \cdot n - 2}$$

$$P_{\text{reflected}} = 56.23 \quad \text{Watts}$$

Determine the total power lost in the transmission line:

$$P_{\text{line}} := P_{\text{eff}} - P_{\text{antenna}}$$

$$P_{\text{line}} = 46.39 \quad \text{Watts}$$

**Appendix B. Full Multiple Reflection Voltage and Power Analysis.
Reference Figures 2 and 3.**

1. Define the known parameters:

Antenna Impedance: $Z_a := 6.2 + j \cdot 14.4$ Ohms

Operating Frequency: $F := 21.2 \cdot 10^6$ Hz

Characteristic Impedance: $Z_0 := 50$ Ohms

Transmission Line Length: $\text{Length} := 100$ Feet

Transmission Line Properties: $\alpha_{dB} := 1.05$ dB / 100 Feet

$V_p := 0.6594$

$$\lambda := \frac{2.997925 \cdot 10^8}{F} \quad \omega := 2 \cdot \pi \cdot F \quad l := \text{Length} \cdot 0.3048 \quad \text{meters}$$

$$\alpha := \frac{\alpha_{dB}}{868.59} = 3.28084 \quad \text{nepers/meter} \quad \beta := \frac{2 \cdot \pi}{V_p \cdot \lambda} \quad \gamma := \alpha + j \cdot \beta$$

2. Calculate the steady state input impedance (Zin) to the transmission line connecting the antenna to the matching network.

$$Z_{in} := Z_0 \cdot \frac{\frac{Z_a}{Z_0} + \tanh(\gamma \cdot l)}{1 + \frac{Z_a}{Z_0} \cdot \tanh(\gamma \cdot l)} \quad Z_{in} = 58.32 - 85.85j \quad \text{Ohms}$$

3. Define suitable matching network component values for C1, C2 and L.

$Q_L := 75 \quad Q_C := 400$

$C_1 := 36.520911 \cdot 10^{-12}$ F $C_2 := 50 \cdot 10^{-12}$ F $L := 0.873927 \cdot 10^{-6}$ H

$R_1 := \frac{1}{\omega \cdot C_1 \cdot Q_C}$ $R_2 := \frac{1}{\omega \cdot C_2 \cdot Q_C}$ $R_L := \frac{\omega \cdot L}{Q_L}$

$Z_{C1} := R_1 + \frac{-j}{\omega \cdot C_1}$ $Z_{C2} := R_2 + \frac{-j}{\omega \cdot C_2}$ $Z_L := R_L + j \cdot \omega \cdot L$

4. Calculate the initial input impedance to the matching network:

$Z_1 := Z_0 + Z_{C2}$ $Z_2 := \frac{1}{\left(\frac{1}{Z_1} + \frac{1}{Z_L}\right)}$

$$Z_m := Z_2 + Z_{C1} \qquad Z_m = 188.72 + 25.16j \qquad \text{Ohms}$$

5. Calculate the initial forward voltage and power delivered to the transmission line. Also calculate the initial reflected voltage and power at the matching network input:

Define the initial voltage driving the matching network input:

$$V_I := 70.710678 \text{ Volts}$$

Calculate the initial reflection properties at the matching network input:

$$\rho_m := \frac{\frac{Z_m}{Z_0} - 1}{\frac{Z_m}{Z_0} + 1} \qquad \rho_m = 0.59 + 0.04j \qquad |\rho_m| = 0.59$$

$$\text{VSWR} := \frac{1 + |\rho_m|}{1 - |\rho_m|} \qquad \text{VSWR} = 3.85$$

Determine the initial voltage and power reflected at the matching network input

$$V_R := \rho_m \cdot V_I$$

$$V_R = 41.42 + 3.09j \quad \text{Volts} \qquad |V_R| = 41.53 \qquad \arg(V_R) \cdot \frac{180}{\pi} = 4.26$$

$$I_R := \frac{V_R}{Z_0} \qquad \theta := \arg(Z_0) \qquad P_R := |V_R| \cdot |I_R| \cdot \cos(\theta)$$

$$P_R = 34.5 \quad \text{Watts}$$

Calculate the initial total voltage and current at the matching network input:

$$V_{\text{total}} := V_I + V_R \qquad V_{\text{total}} = 112.13 + 3.09j \quad \text{Volts} \qquad |V_{\text{total}}| = 112.17$$

$$\arg(V_{\text{total}}) \cdot \frac{180}{\pi} = 1.58 \qquad I_{\text{total}} := \frac{V_{\text{total}}}{Z_m}$$

Use simple circuit theory to determine the voltage at each of the matching network components and the initial forward voltage VF, delivered to the transmission line:

$$V_{FC1} := I_{total} \cdot Z_{C1}$$

$$V_2 := V_{total} - V_{FC1}$$

$$I_L := \frac{V_2}{Z_L}$$

$$V_{FL} := I_L \cdot Z_L$$

$$I_2 := I_{total} - I_L$$

$$V_{FC2} := I_2 \cdot Z_{C2}$$

$$V_{FC1} = -12.39 - 120.47j \quad \text{Volts} \quad |V_{FC1}| = 121.11 \quad \arg(V_{FC1}) \cdot \frac{180}{\pi} = -95.87$$

$$V_{FC2} = 149 + 73.88j \quad \text{Volts} \quad |V_{FC2}| = 166.31 \quad \arg(V_{FC2}) \cdot \frac{180}{\pi} = 26.37$$

$$V_{FL} = 124.52 + 123.56j \quad \text{Volts} \quad |V_{FL}| = 175.42 \quad \arg(V_{FL}) \cdot \frac{180}{\pi} = 44.78$$

$$V_F := I_2 \cdot Z_0$$

$$V_F = -24.48 + 49.68j \quad \text{Volts} \quad |V_F| = 55.38 \quad \arg(V_F) \cdot \frac{180}{\pi} = 116.23$$

Calculate the initial forward power delivered to the transmission line.

$$\theta := \arg(V_F) - \arg(I_2) \quad PF := |V_F| \cdot |I_2| \cdot \cos(\theta)$$

$$PF = 61.34 \quad \text{Watts}$$

6. Calculate the total power delivered to the antenna. Also, calculate the total forward and reflected power at the matching network output. This analysis is performed using voltages and the physical reflection properties of the matching network output.

Determine the antenna reflection properties:

$$\rho_a := \frac{\frac{Z_a}{Z_0} - 1}{\frac{Z_a}{Z_0} + 1} \quad \rho_a = -0.67 + 0.43j \quad |\rho_a| = 0.79$$

Determine the reflection properties of the matching network output:

$$Z1 := Z_0 + ZC1 \quad Z2 := \frac{1}{\left(\frac{1}{Z1} + \frac{1}{ZL}\right)} \quad ZN := ZC2 + Z2$$

$$ZN = 71.13 + 77.82j \quad \text{Ohms}$$

$$\rho N := \frac{\frac{ZN}{Z_0} - 1}{\frac{ZN}{Z_0} + 1} \quad \rho N = 0.42 + 0.38j \quad |\rho N| = 0.56$$

Determine the total power delivered to the antenna due to the multiple voltage reflections between the antenna and the matching network.

$$V1 := VF$$

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

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

$$IA := \left(\frac{V1 \cdot e^{-\gamma l}}{Z_0} - \frac{V1 \cdot e^{\gamma l}}{Z_0} \cdot \rho a \cdot e^{-2\gamma l} \right) + \frac{V1 \cdot e^{-\gamma l}}{Z_0} \cdot \sum_{n=1}^{1000} \rho N^n \cdot \rho a^n \cdot e^{-2 \cdot n \cdot \gamma l}$$

$$I_{antenna} := IA - \frac{V1 \cdot e^{\gamma l}}{Z_0} \cdot \sum_{n=1}^{1000} \rho N^n \cdot \rho a^{(n+1)} \cdot e^{-(2n+2)\gamma l}$$

$$V_{antenna} = 14.13 + 38.16j \quad \text{Volts} \quad |V_{antenna}| = 40.7$$

$$\arg(V_{antenna}) \cdot \frac{180}{\pi} = 69.68 \quad \theta := \arg(V_{antenna}) - \arg(I_{antenna})$$

$$P_{antenna} := |V_{antenna}| \cdot |I_{antenna}| \cdot \cos(\theta) \quad P_{antenna} = 41.77 \quad \text{Watts}$$

Determine the total forward power at the matching network output:

$$V_{\text{forward}} := V_1 + V_1 \cdot \sum_{n=1}^{1000} \rho N^n \cdot \rho a^n \cdot e^{-2 \cdot n \cdot \gamma \cdot l} \quad I_{\text{forward}} := \frac{V_{\text{forward}}}{Z_0}$$

$$\theta := \arg(V_{\text{forward}}) - \arg(I_{\text{forward}}) \quad P_{\text{forward}} := |V_{\text{forward}}| \cdot |I_{\text{forward}}| \cdot \cos(\theta)$$

$$P_{\text{forward}} = 144.4 \quad \text{Watts}$$

Determine the total reflected power at the matching network output:

$$V_{\text{reflected}} := (V_1 \cdot \rho a \cdot e^{-2 \cdot \gamma \cdot l}) + V_1 \cdot \sum_{n=1}^{1000} \rho N^n \cdot \rho a^{(n+1)} \cdot e^{-(2 \cdot n + 2) \cdot \gamma \cdot l}$$

$$I_{\text{reflected}} := \frac{V_{\text{reflected}}}{Z_0} \quad \theta := \arg(V_{\text{reflected}}) - \arg(I_{\text{reflected}})$$

$$P_{\text{reflected}} := |V_{\text{reflected}}| \cdot |I_{\text{reflected}}| \cdot \cos(\theta) \quad P_{\text{reflected}} = 56.23 \quad \text{Watts}$$

7. Calculate the effective net power delivered to the transmission line:

$$P_{\text{eff}} := P_{\text{forward}} - P_{\text{reflected}} \quad P_{\text{eff}} = 88.17 \quad \text{Watts}$$

8. Determine the total steady state voltage delivered back to the matching network input due to the total steady state voltage developed at the matching network output.

Calculate the voltage developed at the matching network output due to the multiple reflections.

$$V_A := V_1 \cdot \rho a \cdot e^{-2 \cdot \gamma \cdot l} + V_1 \cdot \sum_{n=1}^{1000} \rho N^n \cdot \rho a^n \cdot e^{-2 \cdot n \cdot \gamma \cdot l}$$

$$V_{\text{network}} := V_A + V_1 \cdot \sum_{n=1}^{1000} \rho N^n \cdot \rho a^{(n+1)} \cdot e^{-(2n+2) \cdot \gamma \cdot l}$$

Calculate the total steady state voltage at the matching network output to in the voltage VF.

$$V_{\text{output}} := V_{\text{network}} + V_F \quad V_{\text{output}} = -14.86 + 126.74j \quad \text{Volts}$$

$$|V_{\text{output}}| = 127.61 \quad \arg(V_{\text{output}}) \cdot \frac{180}{\pi} = 96.69$$

Calculate the voltage, Vback, delivered back to the matching network input. Compare Vback with the initial reflected voltage VR.

$$I_T := \frac{V_{\text{network}}}{Z_N} \quad V_{RC2} := I_T \cdot Z_{C2} \quad V_2 := V_{\text{network}} - V_{RC2}$$

$$I_{L2} := \frac{V_2}{Z_L} \quad V_{RL} := I_{L2} \cdot Z_L \quad I_2 := I_T - I_{L2}$$

$$V_{RC1} := I_2 \cdot Z_{C1} \quad V_{\text{back}} := I_2 \cdot Z_0$$

$$V_{\text{back}} = -41.42 - 3.09j \quad \text{Volts} \quad V_R = 41.42 + 3.09j \quad \text{Volts}$$

$$|V_{\text{back}}| = 41.53 \quad \arg(V_{\text{back}}) \cdot \frac{180}{\pi} = -175.74$$

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

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

$$\theta := \arg(V_{\text{back}}) - \arg(I_2) \quad P_{\text{back}} := |V_{\text{back}}| \cdot |I_2| \cdot \cos(\theta)$$

$$P_{\text{back}} = 34.5 \quad \text{Watts} \quad P_R = 34.5 \quad \text{Watts}$$

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

9. Verify the total power lost in the matching network by calculating the power lost in the individual network components.

$$VC1 := VFC1 - VRC1 \quad IC1 := \frac{VC1}{ZC1} \quad \theta C1 := \arg(ZC1)$$

$$VC2 := VFC2 - VRC2 \quad IC2 := \frac{VC2}{ZC2} \quad \theta C2 := \arg(ZC2)$$

$$VL := VFL + VRL \quad IL := \frac{VL}{ZL} \quad \theta L := \arg(ZL)$$

$$|VC1| = 290.71 \quad \arg(VC1) \cdot \frac{180.}{\pi} = -89.86$$

$$|VC2| = 184.62 \quad \arg(VC2) \cdot \frac{180.}{\pi} = 62.64$$

$$|VL| = 299.01 \quad \arg(VL) \cdot \frac{180.}{\pi} = 76.46$$

$$PC1 := |VC1| \cdot |IC1| \cdot \cos(\theta C1) \quad PC1 = 1.03 \quad \text{Watts}$$

$$PC2 := |VC2| \cdot |IC2| \cdot \cos(\theta C2) \quad PC2 = 0.57 \quad \text{Watts}$$

$$PL := |VL| \cdot |IL| \cdot \cos(\theta L) \quad PL = 10.24 \quad \text{Watts}$$

$$Ptotal := PC1 + PC2 + PL \quad Ptotal = 11.83 \quad \text{Watts}$$

**Appendix C. Full Multiple Reflection Voltage and Power Analysis.
Reference Figures 2 and 3. Matching Network Components set not to
Match the Input Impedance to Zo.**

1. Define the known parameters:

Antenna Impedance: $Z_a := 6.2 + j \cdot 14.4$ Ohms

Operating Frequency: $F := 21.2 \cdot 10^6$ Hz

Characteristic Impedance: $Z_o := 50$ Ohms

Transmission Line Length: $\text{Length} := 100$ Feet

Transmission Line Properties: $\alpha_{dB} := 1.05$ dB / 100 Feet

$V_p := 0.6594$

$$\lambda := \frac{2.997925 \cdot 10^8}{F} \quad \omega := 2 \cdot \pi \cdot F \quad l := \text{Length} \cdot 0.3048 \quad \text{meters}$$

$$\alpha := \frac{\alpha_{dB}}{868.59} \cdot 3.28084 \quad \text{nepers/meter} \quad \beta := \frac{2 \cdot \pi}{V_p \cdot \lambda} \quad \gamma := \alpha + j \cdot \beta$$

2. Calculate the steady state input impedance (Zin) to the transmission line connecting the antenna to the matching network.

$$Z_{in} := Z_o \cdot \frac{\frac{Z_a}{Z_o} + \tanh(\gamma \cdot l)}{1 + \frac{Z_a}{Z_o} \cdot \tanh(\gamma \cdot l)} \quad Z_{in} = 58.32 - 85.85j \quad \text{Ohms}$$

3. Define suitable matching network component values for C1, C2 and L.

$Q_L := 75 \quad Q_C := 400$

$C_1 := 40.0 \cdot 10^{-12}$ F $C_2 := 80 \cdot 10^{-12}$ F $L := 1.0 \cdot 10^{-6}$ H

$R_1 := \frac{1}{\omega \cdot C_1 \cdot Q_C}$ $R_2 := \frac{1}{\omega \cdot C_2 \cdot Q_C}$ $R_L := \frac{\omega \cdot L}{Q_L}$

$Z_{C1} := R_1 + \frac{-j}{\omega \cdot C_1}$ $Z_{C2} := R_2 + \frac{-j}{\omega \cdot C_2}$ $Z_L := R_L + j \cdot \omega \cdot L$

4. Calculate the initial input impedance to the matching network:

$Z_1 := Z_o + Z_{C2}$ $Z_2 := \frac{1}{\left(\frac{1}{Z_1} + \frac{1}{Z_L}\right)}$

$$Z_m := Z_2 + Z_{C1}$$

$$Z_m = 214.74 - 224.39j \quad \text{Ohms}$$

5. Calculate the initial forward voltage and power delivered to the transmission line. Also calculate the initial reflected voltage and power at the matching network input:

Define the initial voltage driving the matching network input:

$$V_I := 70.710678 \text{ Volts}$$

Calculate the initial reflection properties at the matching network input:

$$\rho_m := \frac{\frac{Z_m}{Z_0} - 1}{\frac{Z_m}{Z_0} + 1} \quad \rho_m = 0.78 - 0.19j \quad |\rho_m| = 0.8$$

$$VSWR := \frac{1 + |\rho_m|}{1 - |\rho_m|} \quad VSWR = 9.11$$

Determine the initial voltage and power reflected at the matching network input

$$V_R := \rho_m \cdot V_I$$

$$V_R = 55.17 - 13.17j \quad \text{Volts} \quad |V_R| = 56.72 \quad \arg(V_R) \cdot \frac{180}{\pi} = -13.43$$

$$I_R := \frac{V_R}{Z_0} \quad \theta := \arg(Z_0) \quad P_R := |V_R| \cdot |I_R| \cdot \cos(\theta)$$

$$P_R = 64.34 \quad \text{Watts}$$

Calculate the initial total voltage and current at the matching network input:

$$V_{total} := V_I + V_R \quad V_{total} = 125.88 - 13.17j \quad \text{Volts} \quad |V_{total}| = 126.57$$

$$\arg(V_{total}) \cdot \frac{180}{\pi} = -5.97 \quad I_{total} := \frac{V_{total}}{Z_m}$$

Use simple circuit theory to determine the voltage at each of the matching network components and the initial forward voltage VF, delivered to the transmission line:

$$V_{FC1} := I_{total} \cdot Z_{C1}$$

$$V_2 := V_{total} - V_{FC1}$$

$$I_L := \frac{V_2}{Z_L}$$

$$V_{FL} := I_L \cdot Z_L$$

$$I_2 := I_{total} - I_L$$

$$V_{FC2} := I_2 \cdot Z_{C2}$$

$$V_{FC1} = 49.6 - 58.22j \quad \text{Volts} \quad |V_{FC1}| = 76.48 \quad \arg(V_{FC1}) \cdot \frac{180}{\pi} = -49.57$$

$$V_{FC2} = 78.02 + 3.47j \quad \text{Volts} \quad |V_{FC2}| = 78.1 \quad \arg(V_{FC2}) \cdot \frac{180}{\pi} = 2.55$$

$$V_{FL} = 76.28 + 45.05j \quad \text{Volts} \quad |V_{FL}| = 88.59 \quad \arg(V_{FL}) \cdot \frac{180}{\pi} = 30.56$$

$$V_F := I_2 \cdot Z_0$$

$$V_F = -1.74 + 41.58j \quad \text{Volts} \quad |V_F| = 41.61 \quad \arg(V_F) \cdot \frac{180}{\pi} = 92.4$$

Calculate the initial forward power delivered to the transmission line.

$$\theta := \arg(V_F) - \arg(I_2)$$

$$PF := |V_F| \cdot |I_2| \cdot \cos(\theta)$$

$$PF = 34.63 \quad \text{Watts}$$

6. Calculate the total power delivered to the antenna. Also, calculate the total forward and reflected power at the matching network output. This analysis is performed using voltages and the physical reflection properties of the matching network output.

Determine the antenna reflection properties:

$$\rho_a := \frac{\frac{Z_a}{Z_0} - 1}{\frac{Z_a}{Z_0} + 1}$$

$$\rho_a = -0.67 + 0.43j$$

$$|\rho_a| = 0.79$$

Determine the reflection properties of the matching network output:

$$Z1 := Z_0 + Z_{C1}$$

$$Z2 := \frac{1}{\left(\frac{1}{Z1} + \frac{1}{ZL}\right)}$$

$$Z_N := Z_{C2} + Z2$$

$$Z_N = 169.21 + 204.65j \text{ Ohms}$$

$$\rho_N := \frac{\frac{Z_N}{Z_0} - 1}{\frac{Z_N}{Z_0} + 1}$$

$$\rho_N = 0.76 + 0.23j$$

$$|\rho_N| = 0.79$$

Determine the total power delivered to the antenna due to the multiple voltage reflections between the antenna and the matching network.

$$V1 := VF$$

$$V_A := V1 \cdot e^{-\gamma l} + V1 \cdot e^{\gamma l} \cdot \rho_A \cdot e^{-2\gamma l} + V1 \cdot e^{-\gamma l} \cdot \sum_{n=1}^{1000} \rho_N^n \cdot \rho_A^n \cdot e^{-2n\gamma l}$$

$$V_{\text{antenna}} := V_A + V1 \cdot e^{\gamma l} \cdot \sum_{n=1}^{1000} \rho_N^n \cdot \rho_A^{(n+1)} \cdot e^{-(2n+2)\gamma l}$$

$$I_A := \left(\frac{V1 \cdot e^{-\gamma l}}{Z_0} - \frac{V1 \cdot e^{\gamma l}}{Z_0} \cdot \rho_A \cdot e^{-2\gamma l} \right) + \frac{V1 \cdot e^{-\gamma l}}{Z_0} \cdot \sum_{n=1}^{1000} \rho_N^n \cdot \rho_A^n \cdot e^{-2n\gamma l}$$

$$I_{\text{antenna}} := I_A - \frac{V1 \cdot e^{\gamma l}}{Z_0} \cdot \sum_{n=1}^{1000} \rho_N^n \cdot \rho_A^{(n+1)} \cdot e^{-(2n+2)\gamma l}$$

$$V_{\text{antenna}} = 29.16 + 13.63j \quad \text{Volts} \quad |V_{\text{antenna}}| = 32.18$$

$$\arg(V_{\text{antenna}}) \cdot \frac{180}{\pi} = 25.05$$

$$\theta := \arg(V_{\text{antenna}}) - \arg(I_{\text{antenna}})$$

Determine the total forward power at the matching network output:

$$V_{\text{forward}} := V_1 + V_1 \cdot \sum_{n=1}^{1000} \rho N^n \cdot \rho a^n \cdot e^{-2 \cdot n \cdot \gamma \cdot l} \quad I_{\text{forward}} := \frac{V_{\text{forward}}}{Z_0}$$

$$\theta := \arg(V_{\text{forward}}) - \arg(I_{\text{forward}}) \quad P_{\text{forward}} := |V_{\text{forward}}| \cdot |I_{\text{forward}}| \cdot \cos(\theta)$$

$$P_{\text{forward}} = 90.31 \quad \text{Watts}$$

Determine the total reflected power at the matching network output:

$$V_{\text{reflected}} := (V_1 \cdot \rho a \cdot e^{-2 \cdot \gamma \cdot l}) + V_1 \cdot \sum_{n=1}^{1000} \rho N^n \cdot \rho a^{(n+1)} \cdot e^{-(2 \cdot n + 2) \cdot \gamma \cdot l}$$

$$I_{\text{reflected}} := \frac{V_{\text{reflected}}}{Z_0} \quad \theta := \arg(V_{\text{reflected}}) - \arg(I_{\text{reflected}})$$

$$P_{\text{reflected}} := |V_{\text{reflected}}| \cdot |I_{\text{reflected}}| \cdot \cos(\theta) \quad P_{\text{reflected}} = 35.17 \quad \text{Watts}$$

7. Calculate the effective net power delivered to the transmission line:

$$P_{\text{eff}} := P_{\text{forward}} - P_{\text{reflected}} \quad P_{\text{eff}} = 55.14 \quad \text{Watts}$$

8. Determine the total steady state voltage delivered back to the matching network input due to the total steady state voltage developed at the matching network output.

Calculate the voltage developed at the matching network output due to the multiple reflections.

$$V_A := V_1 \cdot \rho a \cdot e^{-2 \cdot \gamma \cdot l} + V_1 \cdot \sum_{n=1}^{1000} \rho N^n \cdot \rho a^n \cdot e^{-2 \cdot n \cdot \gamma \cdot l}$$

$$V_{\text{network}} := V_A + V_I \cdot \sum_{n=1}^{1000} \rho N^n \cdot \rho a^{(n+1)} \cdot e^{-(2n+2)\gamma l}$$

Calculate the total steady state voltage at the matching network output to in the voltage VF.

$$V_{\text{output}} := V_{\text{network}} + V_F \quad V_{\text{output}} = 62.06 + 79.58j \quad \text{Volts}$$

$$|V_{\text{output}}| = 100.92 \quad \arg(V_{\text{output}}) \cdot \frac{180}{\pi} = 52.05$$

Calculate the voltage, Vback, delivered back to the matching network input. Compare Vback with the initial reflected voltage VR.

$$I_T := \frac{V_{\text{network}}}{Z_N} \quad V_{RC2} := I_T \cdot Z_{C2} \quad V_2 := V_{\text{network}} - V_{RC2}$$

$$I_{L2} := \frac{V_2}{Z_L} \quad V_{RL} := I_{L2} \cdot Z_L \quad I_2 := I_T - I_{L2}$$

$$V_{RC1} := I_2 \cdot Z_{C1} \quad V_{\text{back}} := I_2 \cdot Z_0$$

$$V_{\text{back}} = -10.74 + 22.22j \quad \text{Volts} \quad V_R = 55.17 - 13.17j \quad \text{Volts}$$

$$|V_{\text{back}}| = 24.68 \quad \arg(V_{\text{back}}) \cdot \frac{180}{\pi} = 115.8$$

Note that Vback is NOT the negative of the initial reflected voltage VR. Therefore, there will not be a total cancellation of reflected voltage at the matching network input. There will be some level of steady state reflected voltage and power at the matching network input.

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

$$\theta := \arg(V_{\text{back}}) - \arg(I_2) \quad P_{\text{back}} := |V_{\text{back}}| \cdot |I_2| \cdot \cos(\theta)$$

$$P_{\text{back}} = 12.18 \quad \text{Watts} \quad P_R = 64.34 \quad \text{Watts}$$

9. Verify the total power lost in the matching network by calculating the power lost in the individual network components.

$$VC1 := VFC1 - VRC1 \quad IC1 := \frac{VC1}{ZC1} \quad \theta C1 := \arg(ZC1)$$

$$VC2 := VFC2 - VRC2 \quad IC2 := \frac{VC2}{ZC2} \quad \theta C2 := \arg(ZC2)$$

$$VL := VFL + VRL \quad IL := \frac{VL}{ZL} \quad \theta L := \arg(ZL)$$

$$|VC1| = 104.34 \quad \arg(VC1) \cdot \frac{180}{\pi} = -108.85 \quad PC1 := |VC1| \cdot |IC1| \cdot \cos(\theta C1)$$

$$|VC2| = 91.25 \quad \arg(VC2) \cdot \frac{180}{\pi} = 18.01 \quad PC2 := |VC2| \cdot |IC2| \cdot \cos(\theta C2)$$

$$|VL| = 183.77 \quad \arg(VL) \cdot \frac{180}{\pi} = 35.91 \quad PL := |VL| \cdot |IL| \cdot \cos(\theta L)$$

$$PC1 = 0.15 \quad \text{Watts} \quad PC2 = 0.22 \quad \text{Watts} \quad PL = 3.38 \quad \text{Watts}$$

$$P_{total} := PC1 + PC2 + PL \quad P_{total} = 3.75 \quad \text{Watts}$$

Step 10. Determine the steady state input impedance and the total steady state reflected power at the matching network input

$$V_{input} := VI + VR + V_{back} \quad I_{input} := \frac{VI}{Z_o} - \frac{VR}{Z_o} - \frac{V_{back}}{Z_o}$$

$$Z_{input} := \frac{V_{input}}{I_{input}} \quad Z_{input} = 190.54 + 82.78j \quad \text{Ohms}$$

Reflected Power Calculation:

$$V_{ref} := VR + V_{back} \quad I_{ref} := \frac{V_{ref}}{Z_o} \quad \theta := \arg(V_{ref}) - \arg(I_{ref})$$

$$P_{ref} := |V_{ref}| \cdot |I_{ref}| \cdot \cos(\theta) \quad P_{ref} = 41.11 \quad \text{Watts}$$

PLUMBER'S 40-METER LOOP ANTENNA

A low-profile, easy-to-build design

Do you like to take your radio equipment on vacations? Do you live in a restricted area where it is forbidden to put up an outside antenna? Do you just like to experiment with antennas. If so, you may want to consider this low-profile, easy-to-build loop antenna (**Photo A**).

A PVC "Plumber's Delight"

Old-timers will remember "The Plumber's Delight" antenna—a beam made from water pipe. The frame of my antenna is also made from water pipe, but the plastic PVC type instead of metal (**the PVC material supports the loop conductors**). Having read a number of articles on building small loop antennas, I found that most used copper tubing with motor-driven tuning capacitors. I decided to see if I could come up with a loop that was inexpensive and could be built without any special equipment. Above all, I wanted it to be portable so it could be used for field operations.

My desired specifications were:

1. No more than 6 feet high
2. 50-ohm impedance
3. Very portable
4. Inexpensive
5. Easy to construct
6. Able to handle 100 watts

My loop antenna is patterned after the one described in the July 1993 *QST* article by R. Newkirk called "HONEY, I Shrunk the Antenna."¹ His design used three turns of zip cord wire with three capacitors in the loop to provide a better distribution of current. One of the capacitors was variable. The author also claimed the feed impedance was 50 ohms—one

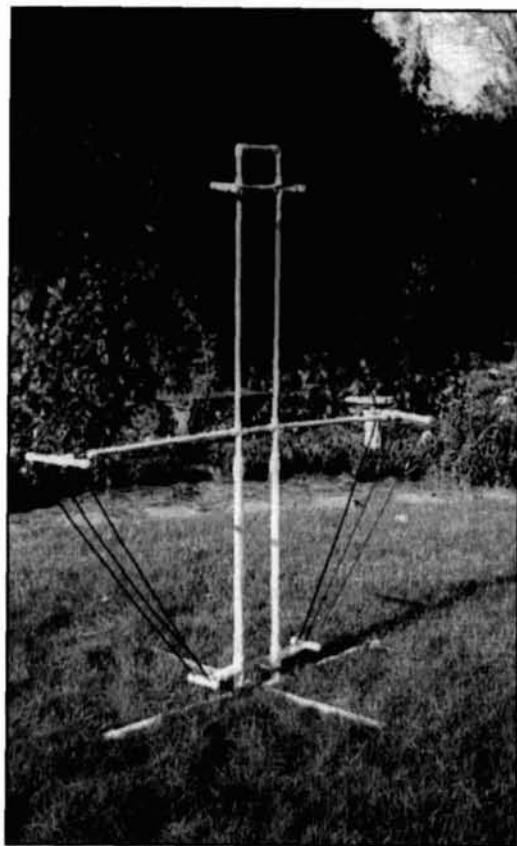


Photo A. The Plumber's 40-Meter Loop Antenna.

of my specifications. Newkirk's design looked like a good one to try to build, and it didn't require copper tubing for the radiator. However, I didn't like the zip cord the author used, and the turns appeared to be close coupled. I decided to loosely couple the turns on my version and note the effect.

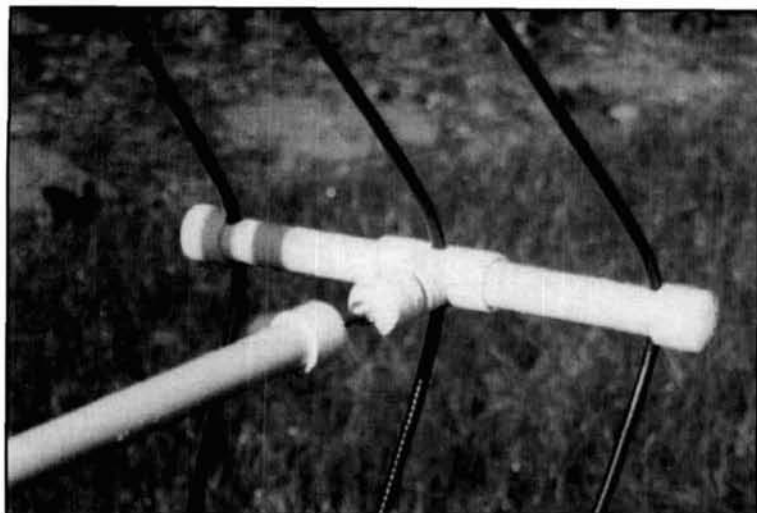


Photo B. The arm assembly.

Apparently the real secret to designing an efficient loop antenna is to keep the DC and RF resistances small in comparison with the radiation resistance, hence the decision made by many to use copper tubing. The variable capacitor is another source of resistance, unless a split stator capacitor is used. Even then, making a good connection to the individual stators could be a problem.

My goal was to make an antenna that did not require a split stator capacitor or other special types. I decided to try using double-sided printed-circuit board material for the fixed capacitors. A 1/16-inch thick pc board with glass epoxy insulation has a capacitance of about 20 pF per square inch. The insulation qualities of the board are very good. I made the variable capacitor from the same material and, by using

copper, was able to obtain a good solder joint. It was motor driven with a low-speed gear drive. The capacitor had a range from about 25 to 80 pF.

My next goal was to make the unit portable for use on field trips. I also wanted the frame to be made of easy to find material, so I chose PVC pipe used for irrigation systems. This lightweight pipe is readily available from hardware stores.

To keep the RF resistance low, I used RG-8/U. Unfortunately, this coax isn't very flexible and was difficult to fold for portability. The next best choice was RG-59/U, which has a diameter of about 1/4 inch. (Only the outer braid is used.)

Checking the design

In about a week's time, I had an antenna built and ready to try. I connected it directly to the output of my Yaesu 820 transmitter without a tuner. I set the transmitter to 7.15 MHz. Crossing my fingers, I pressed the key. I held the power at about 10 watts for safety reasons, as I didn't know how the pc board capacitors would hold up. I set the motor-driven capacitor about mid way. Low and behold, the SWR was less than 2. I really couldn't believe it. I went over the band and found that, with one setting of the variable capacitor, the SWR was never greater than 3. I wondered if this meant I could get rid of the motor drive, the most expensive part of the system.

It really looked like something was wrong. All of the articles I'd read said loop antennas exhibited very sharp tuning because they had a very high Q. I exchanged the variable capacitor

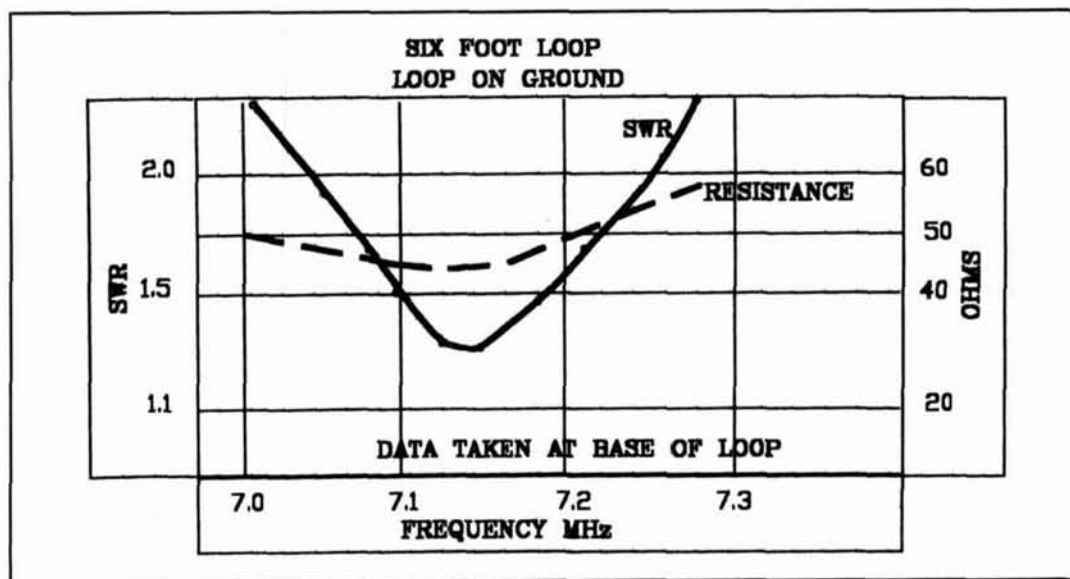


Figure 1. SWR measurements.

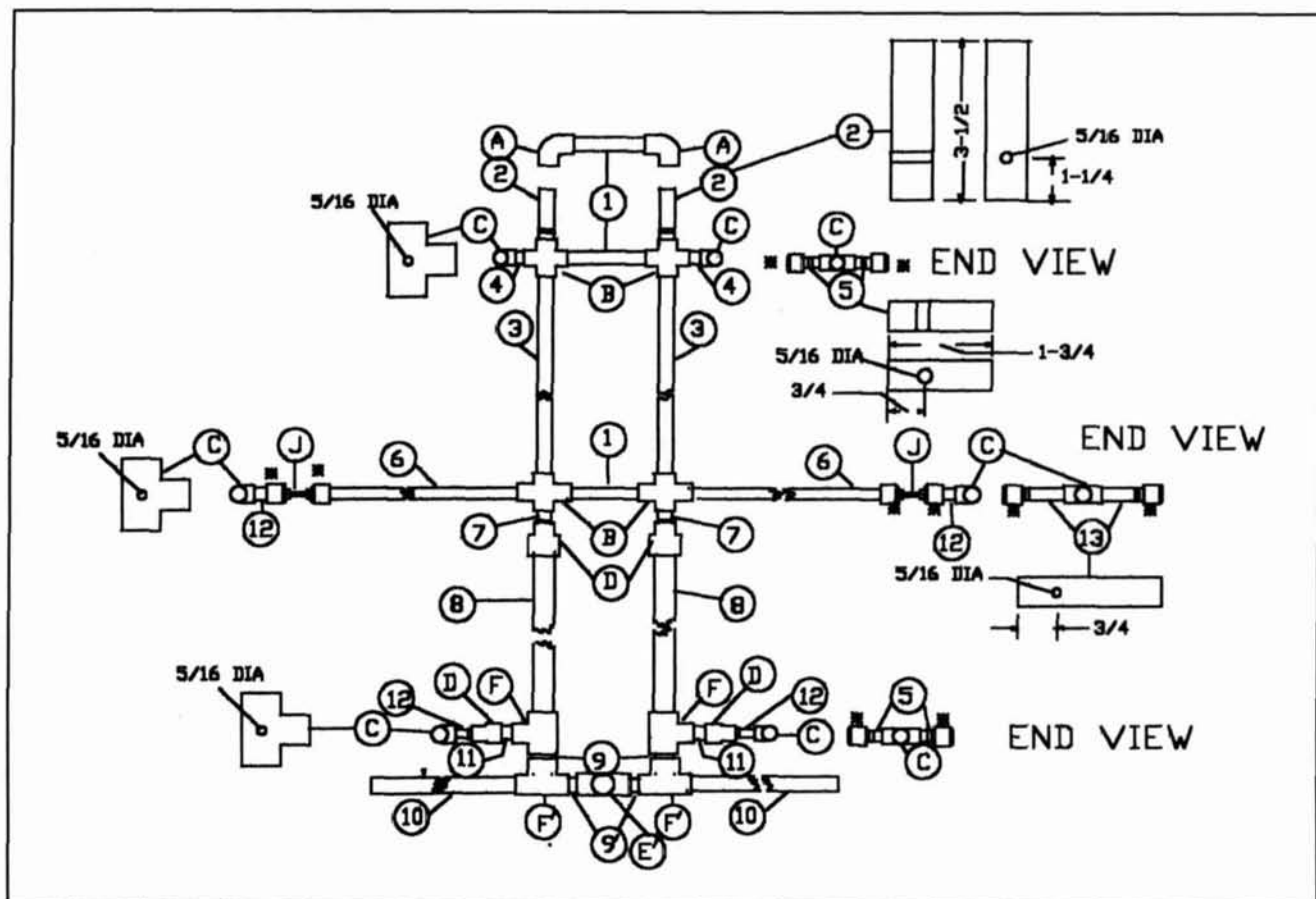


Figure 2. Construction diagram.

for a fixed capacitor made, again, from pc board material.

I got on 40 CW that morning and phoned W6RJY in Santa Rosa, about 100 miles north of my QTH, and asked if he could hear me. The antenna was sitting on the ground in a vertical plane. The transmitter was running 10 watts output. He heard me, so I knew the antenna did radiate. That evening I found another good friend of mine, W6BNB near Santa Rosa, signing from another contact so I gave him a call. W6BNB came back with an S7 report. I was still running 10 watts. The next evening I was in real luck. I heard a KM5 in Texas, gave him a call, and he came back to me. With the transmitter still running 10 watts, he gave me a 339. Not bad for an antenna sitting on the ground. I think the loop was almost broadside to him, which meant he wasn't getting my maximum signal. I concluded the antenna really did work.

Since then, I've worked a station in Pendleton, Oregon, with an S8 report and the same in Laguna Beach, Southern California, with the same report running 10 watts. Both stations were about 350 miles from my QTH. But I must admit, band conditions were very good during this period.

On November 30, I decided to see if I could work SSB on 40. It was a little after noon when I broke into a roundtable with a station in San Bernadino, California, about 300 miles south of here. He gave me an S3, but the other two stations gave me S7/9. They were about 50 miles from here. One station asked me to try putting the loop in a horizontal plane about eight feet off the ground on a ladder to see what would happen. These tests were made with 20 watts output and the results were bad. Their signals were down and mine was almost out.

As a final check, another friend of mine, W6LWR, offered to bring over his MFJ antenna analyzer. We found the SWR went from 2.4 at 7.0 MHz to 1.8 at 7.3 MHz, with the low point being 1.3. At the 50-ohm point the SWR was 1.5, which meant that any solid-state transmitter would have no problem loading this antenna without an antenna tuner (see Figure 1).

Build your own

As I indicated earlier, this antenna is based on one described in the July 1993 *QST*. Three turns of RG-59/U (50 feet) form the loops.

Table 1. Materials chart.

PVC pipe

Item	Description	Length	Qty
1	1/2-inch SCH 40	5-1/4	3
2	"	3/1-2	2
3	"	30	2
4	"	1	2
5	"	1-1/4	8
6	"	23	2
7	"	3-1/2	2
8	1-inch SCH 40	27-1/4	2
9	"	1-3/4	4
10	"	30	4
11	"	2-3/8	2
12	"	2-1/2	4
13	"	4-3/4	4

Connecting hardware

Item	Description	Qty
A	1/2-inch 90	2
B	1/2-inch cross	4
C	1/2-inch T	6
D	1-3/4-inch red	4
D	3/4-1/2-inch red	4
E	1-inch cross	1
F	1-inch T	4
G	1/4-inch carriage bolt, 6-inch long	2
G	1/4-inch nut	4
*	1/2-inch end cap	18

The first step is to build the frame for the antenna (see **Figures 2, 3, and 4** and **Photo B** for assembly information). It's made from heavy-duty schedule 40 PVC pipe. Two sizes—1 inch and 1/2 inch—are used, and two lengths of each are required.

Using the information in **Table 1**, cut the PVC to the required lengths. Some of the lengths require that holes be drilled through them (see **Figure 2**). If one is available, use a PVC cutting tool to cut the pipe; however, a hacksaw will suffice. To get a square cut, let the weight of the saw carry it through the material and don't press down. It will be necessary to clean up the edges if you use a saw.

On the base structure, only the items in **Figure 2** labeled 9', E', and F' are cemented together with PVC cement. When you have all the items prepared you can begin construction.

Continuing to refer to **Figure 2**, add items F and 9 to base item F' (note there are two places). Do not cement in these items. Now add

items E' and 9'. These items *will* be cemented in, but, before doing so, make sure item E' is at right angles to the F' items. Do not cement in anything labeled item 10. These parts must remain removable to allow for portability. No other items will require cementing.

If you've never used this material before, you need to know that the cement sets up in seconds. You have to work fast. The best way to construct a good right angle is to first cement the two 9' pieces that go to item F' (in two places). Insert the items labeled 10 into F'. Item E' also has to two items numbered 10 inserted into it at right angles to the other two.

On a flat surface, join items E' and F' to the other end of item 9' (one end of item 9' has already been cemented to item F'). Set them perpendicular with each other as shown in the **Figure 2**. Mark a line on F' and E' with a permanent marker pen showing where the two go together. Remove the items to be cemented and put cement on the end of item 9' that goes into item E'. Do one side at a time, making sure they are perpendicular to each other. This will be the most difficult part of the entire assembly. Make sure you get it right, or your antenna will have a crooked back. Continue to assemble the rest of the frame.

Make sure items C (six pieces), 2, 5, and 13 have the required hole drilled into them. To fabricate the item J assembly, the end caps must have a 1/4-inch hole drilled in their tops. Place the 1/4-inch carriage bolt head in the cap attached to item 12 and secure it with a 1/4-inch nut. Insert the other end into the cap on item 6, securing with a hex nut or wing nut on that end, too. Run this nut up against the one associated with item 12. Assemble all of the items as shown on the structure drawing in **Figure 2**. Once the structure is completed, you can move on to the capacitors.

I ended up with about 60 pF each for the three capacitances. This required pc board material of about 3 by 1 inches (3 x 1 x 20 x pF). To be precise, my board actually measured 3 x 0.8 inches. To prevent arcing from the copper foil to the other side of the pc board, I cut away the copper from around the edge; 1/32 of an inch does the job. Make three capacitors of the same size as shown **Photo C**.

Using steel wool, polish the copper before tinning. Tin one end on one side and the opposite end on the other side. The tinning need only cover about 3/4 of an inch from the end.

Adding the RG-59/U

You're now ready to put the RG-59/U into the frame. **Figure 3** is an overall view of the assembly. Uncoil 50 feet of RG-59/U and strip the outer covering from one end about 1.5 inch-

es. Push the braid back and remove the center conductor and the insulating material. Bring the braid to its original position and flatten it. Apply solder to the end in preparation for soldering it to one of the pc board capacitors. Feed this end into the hole in item 2, at the top of the structure, in the direction of the item 1 area. Allow about 3 inches of overhang for later connection to one of the capacitors. Take the other end of the 50 feet of RG-59/U and feed it through the center hole on top item C (item C should be mounted into item B at the top). Place item J and its item C assembly (12 and 13) into item 6. The nuts should be next to each other and provide a short length on item J. (Item 6 should be mounted into item B at the center of the structure.) Continue feeding the cable through the center of the item C connected to item J.

Find the item C at the bottom of the structure. Feed the cable through the center hole and then to the right-side hole of item C on the opposite side of the assembly. (Tag this point as one to be cut open later to provide a feed point for the coax cable to the transmitter.) Continue up and around on the right side through each item C until you have returned to the bottom of the structure. At the bottom, cross over to the left hole on the next item C and continue up along the outside all the way around until you reach the bottom again.

Now, cross the bottom and enter the center hole of the opposite item C, continuing up the center holes to the top and through item 2. Tighten the tension on the cables (don't use any of the items labeled J for this tensioning). Just pull it tight by hand around each loop. When done, leave about 3 inches in the item 1 area and cut the cable. Strip back the outer covering and cut out the center conductor as you did before. These ends will be one of the junctions for one of the pc board capacitors. Flatten the copper braid ends and tin them back about 3/4 inches. Solder the two leads to the capacitor, one to each side.

To assist in soldering, I clamped the braid to the copper with a large alligator clip. (You can also lay the entire assembly down on its side to make soldering easier.) I used a 100-watt soldering gun for this purpose, but any heavy-duty soldering iron will work. Try not to overheat the surface because the pc board material will warp. Apply the heat mainly to the braid and the solder to the pc board. A good solder joint is important to keep losses low.

You are now ready to mount the other two capacitors. Two cables should pass through the item 1 area to the other side. Cut the cable so you can fit the 3-inch pc board in between each one. Prepare these capacitors as you did the center capacitor.

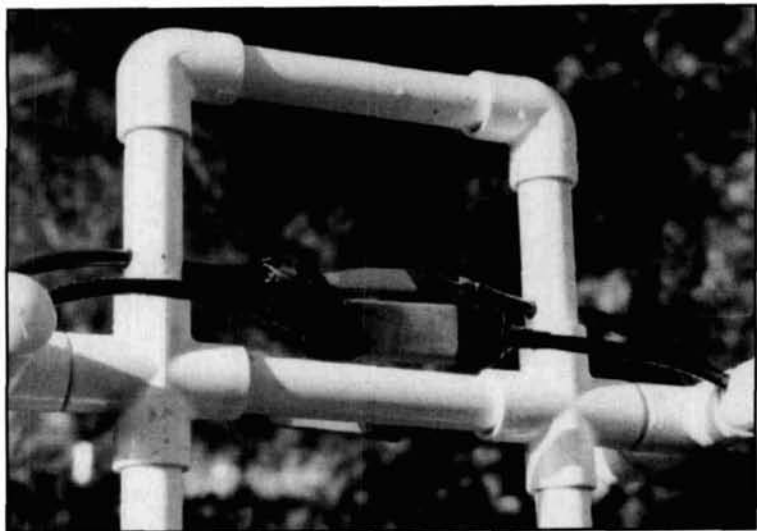


Photo C. The capacitors.

Once you've completed the capacitor mountings, you're ready to make the cable connection to the feed point of the antenna. Three cables pass through this point at the lower part of the antenna. Recall that when you were routing the cable through this area, a tag was placed on the cable to be cut in order to install the SO-239 connector.

The SO-239 connector mounts on a piece of acrylic plastic 3 x 6 x 1/4 inches (Figure 4). The unit is attached at the bottom of the assembly using self-tapping screws. To obtain a low-loss joint, I made an assembly using pc board (copper on one side only) material; one part mounts in front of the SO-239 and the second on the connection pin where you connect the wire (Photo D). This pc board material has a hole to just provide clearance for this pin. To ensure a good connection, I also soldered a wire into the SO-239 center pin and then soldered this to the pc board. The pc board material provides a good surface to connect the wire braid from the cable.

The hole for the SO-239 has a diameter of 5/8 inch. If you don't have a drill this large, use a round file to enlarge the hole. A hole must be made in both the plastic and the pc board material. Of course, you could permanently attach the feed coax at this point and eliminate the SO-239 connector. I fed the antenna with 70 feet of RG-59/U. Losses at this frequency are negligible with this low SWR.

You are now ready to connect the cable to the feed point.

Follow the cable used to start the looping operation down to the bottom (it should be the center conductor and have a tag). The point at which the cable crosses over to the other side is where you cut open the cable to connect the SO-239. I doubt it matters which side you con-

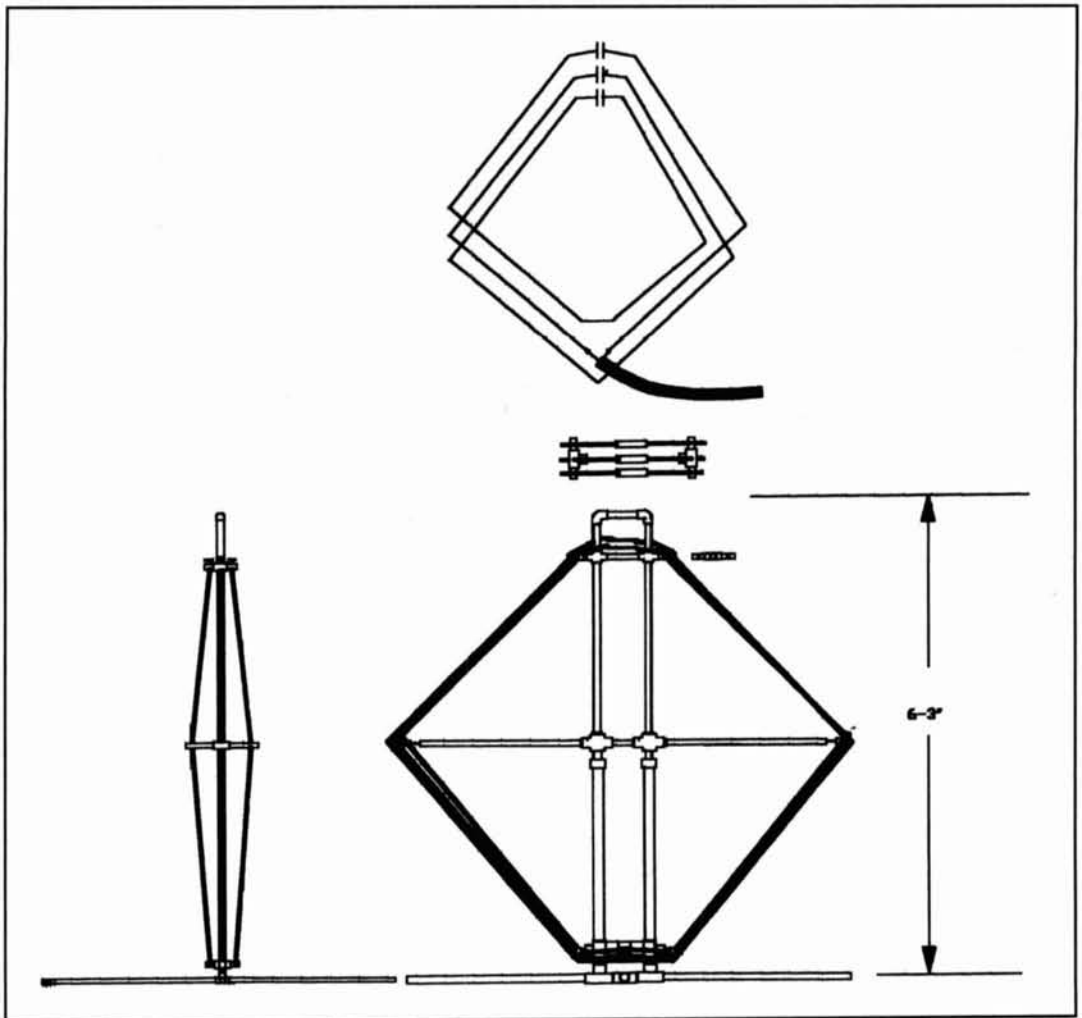


Figure 3. Adding the R-59/U to the assembly.

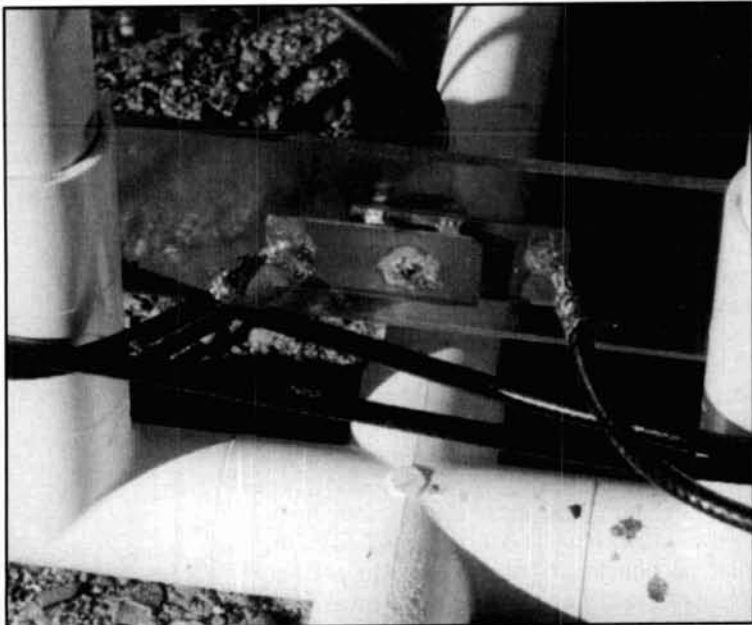


Photo D. The base connection to the coax.

nect to which side, but I connected the first cable from the top to the center pin. The shell goes to the rest of the cable assembly.

The last step is to tighten up the J bolt nuts in order to tighten up the cables. If you followed the instructions, they should go out about a quarter of the way.

Taking it with you

To break down the assembly for carrying, bring the nuts together. This loosens up the cables and allows the arms labeled item 6 in **Figure 2** to be removed from the main assembly. These then fold down to the side. If even a smaller package is desired, the two items numbered 3 and above can be removed and folded down. Also the items numbered 10 can be removed to reduce this area. The total weight of the package is 10 pounds. (**Photo E** shows the collapsed unit.)

You have now completed the assembly and are ready to test the antenna. Remember, no

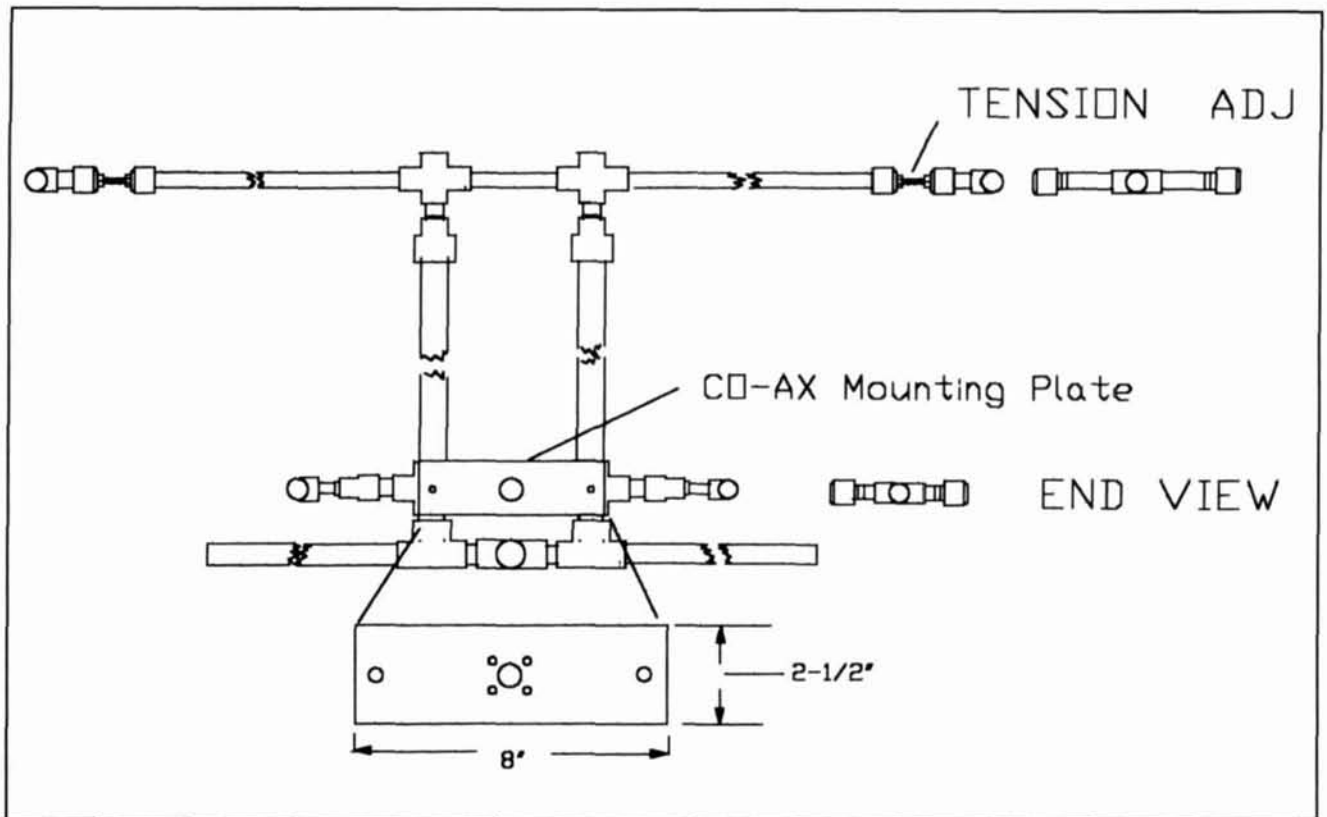


Figure 4. Mounting structure.

tuner is required if the transmitter output is designed for a 50-ohm load impedance with proper filtering.

Testing

The best way to check the antenna is with an SWR meter. Connect the antenna to the transmitter with a 50-ohm cable directly to the 50-ohm output of the transmitter. If you can, set the power to a low value—about 5 or 10 watts. Set the transmitter frequency to 7150 kHz. Key the transmitter, setting the SWR meter to full scale on the forward position. Read the reverse SWR. It should be less than 2. Check the SWR at 7010 and 7290. If the SWR is higher at 7290 than at 7010, it can be shifted down either by adding capacitance to the center pc capacitor or moving in the coax cables. Do this by cutting down the length of the item 13 pieces. To change the capacitance on the center capacitor, solder a wire to the top and bottom plates (about 6 inches) and twist them together to lower the frequency. To move up, the reverse should apply. Check the SWR again and adjust the wire length to give the lowest SWR at about 7150 kHz.

Once the antenna is tuned to the proper frequency, I advise that you weatherproof the

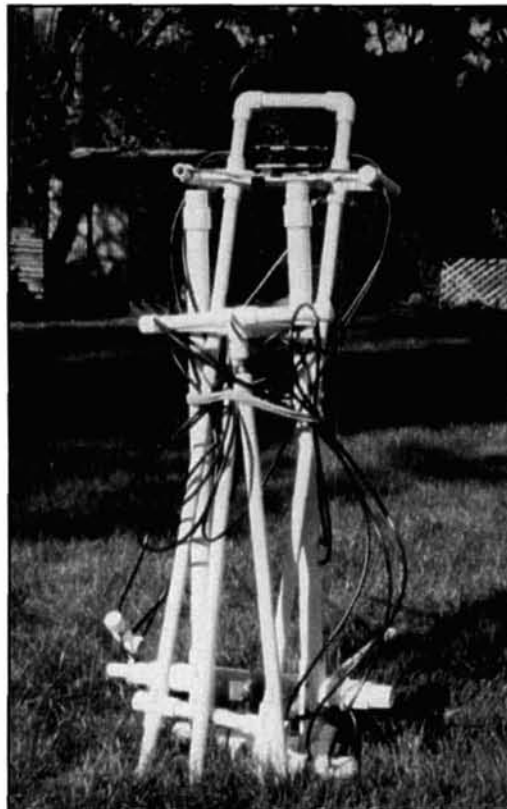


Photo E. The loop collapsed.

capacitors. I coated them with a clear plastic spray and then covered them with black electrical tape. I've noticed very little difference in SWR during rainstorms; however, I wouldn't run 100 watts at this time. You are now ready to try out the loop.

A final check

I compared the loop against a half-wave dipole on transmit and found that the loop is about 2 S units weaker. I have also observed when comparing the loop with a half-wave on receiving that the fade on the signals is deeper on the loop. There are times that it has about the same signal strength, but, usually, when on receive, the loop also is about 2 S units lower. When looking at signals traveling over great

distances, the antennas are about the same. Local checks on the radiation off the side of the antenna indicate about 1 S unit down.

These results aren't too bad for an antenna sitting on the ground among a bunch of pine trees. Remember, this is not an equivalent to a half-wave antenna, and you won't always be able to make a contact; but when the conditions are right, it does a pretty good job. Also remember this antenna could have high-voltage radiation points, so don't use it close to your operating position—especially when running high power.

I'd certainly like to hear from any of you who attempt this project. You can e-mail me at <w6qif@ix.com>. Good QSOing! ■

REFERENCE

I. R. Newkirk, "Honey, I Shrunk the Antenna," *QST*, July 1993, pages 34, 35, and 39.

PRODUCT INFORMATION

Analog Devices Offers Op Amp on 1-mA Supply Current

Analog Devices has released the AD8014, a 400-MHz, low-power, high-performance op amp. The AD8014 is a high-speed amplifier with 400-MHz -3 dB bandwidth, 4000-volt/ μ s slew rate and 24-ns settling time of 0.1 percent. This amplifier has a 60-ns overload recovery, voltage and current noise of 3.5 nV/ \sqrt Hz and 5 pA/ \sqrt Hz, respectively, and distortion of -70 dB THD @ 5 MHz, which makes the AD8014 suitable for wide-band signal processing applications. The 1-mA supply also makes it suitable for any power-sensitive, high-speed system, including battery-operated equipment.

For more information, including samples, on this op amp and other Analog Devices high-speed op amps, visit them at <<http://www.analog.com/highspeedamps>>.

Second Edition of Coaxial Cable Installation Accessories Catalog

Andrew Corporation has released the second edition of its Coaxial Cable Installation Accessories Catalog. The catalog features sections on cable installation accessories, lightning protection products, and value added services. New sections cover installation aids and new services for installers. The pages have short product descriptions, photographs, line

drawings, and products numbers. The appendices contain hints for selecting installation products, lightning strike statistics and definitions, and cable hanger spacing tables.

For further information, call the Andrew Customer Support Center at 1-800-255-1479 and request packet number 424. You can also visit Andrew Corporation's website at <<http://www.andrew.com>>.

Philips ECG Introduces Two New Surge Suppressors

Philips ECG has added the EMF-41 and EMF-65A surge strips to its line of multi-purpose surge suppressors. The new surge suppressors are designed to protect electronic equipment against surges on telephone and AC power lines.

The EMF-65A telecommunications surge suppressor offers protection against voltage spikes on telephone lines as well as AC power line protection in all three modes. The EMF-41 all-purpose surge suppressor is a four-outlet strip that provides basic protection while fitting into tight spaces.

Philips' complete line of surge suppressors is described in ET-2758-2—a new color brochure with complete specifications, photos, and diagrams. ECG products and literature are available from Philips distributors worldwide. To locate your nearest distributor, call 1-800-526-9354.

THE QUAD ANTENNA REVISITED

Part 2: The basic two-element quad

This is the second in a series of articles revising my work on the quad antenna, and it is devoted to the basic two-element quad. The overall goal of publishing the revisions is to improve accuracy, provide data not possible to accrue with earlier analysis methods, and make the results easier to use. A specific goal is to present data in a form that allows easy translation to any design and operating frequency. This is done by using calculations in a specific frequency range and presenting results with “scaling factors” for easy change to other frequencies.

The classic “quad antenna”

The structure of a classic “quad antenna” is shown in **Figure 1**. It is composed of two square loops very nearly one wavelength in perimeter, mounted parallel to each other, and separated by a fraction of a wavelength. Many early designs used twinlead for each conductor, connected to make a two-turn loop; this is rarely done today.

The perimeter is usually denoted by the letter B , the spacing by C , and the radius of the conductor by A . The wire diameter is also used and may be designated by the ratio of length to diameter, $B/2A$. Papers in the scientific journals often use a “thickness factor” instead of $B/2A$ (see **Reference 1** from *The Quad Antenna*),^{1,2} but this is not used here as it introduces an unnecessary calculation step. Dimensions of these quantities may be given in feet, meters, or in wavelengths. A perimeter in wavelengths is designated as kB .

Commonly, one loop is excited by current source at the bottom, indicated by the small cir-

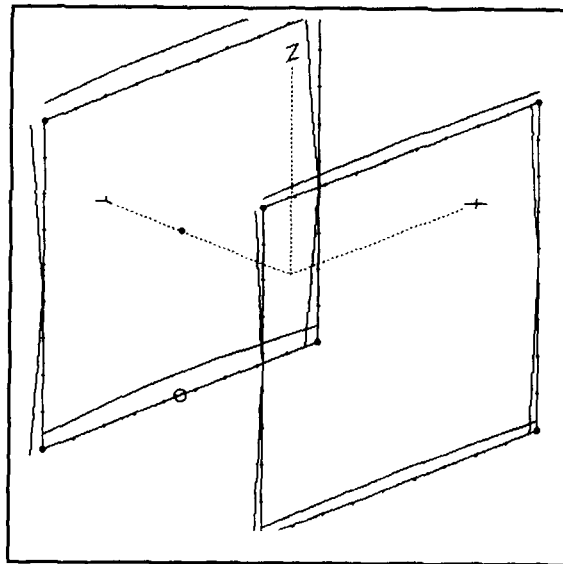


Figure 1. Schematic of basic two-element quad.

cle in **Figure 1**. Because the two loops are coupled by magnetic and electric fields, current flows in both wires, as shown in **Figure 1** and in detail in **Figure 2**. Here the part of the curve to the left is the current on the excited loop. This machine-generated plot does not show that there is a continuous current flow around each loop. Note that the current on the second or parasitic element is only about one third that of the driven element at the same part of the loops. Note also that the current on the two elements never goes to zero, and that there is a variation of phase along the driven element that does not appear on the parasitic. There will be differences in details of the current distribution

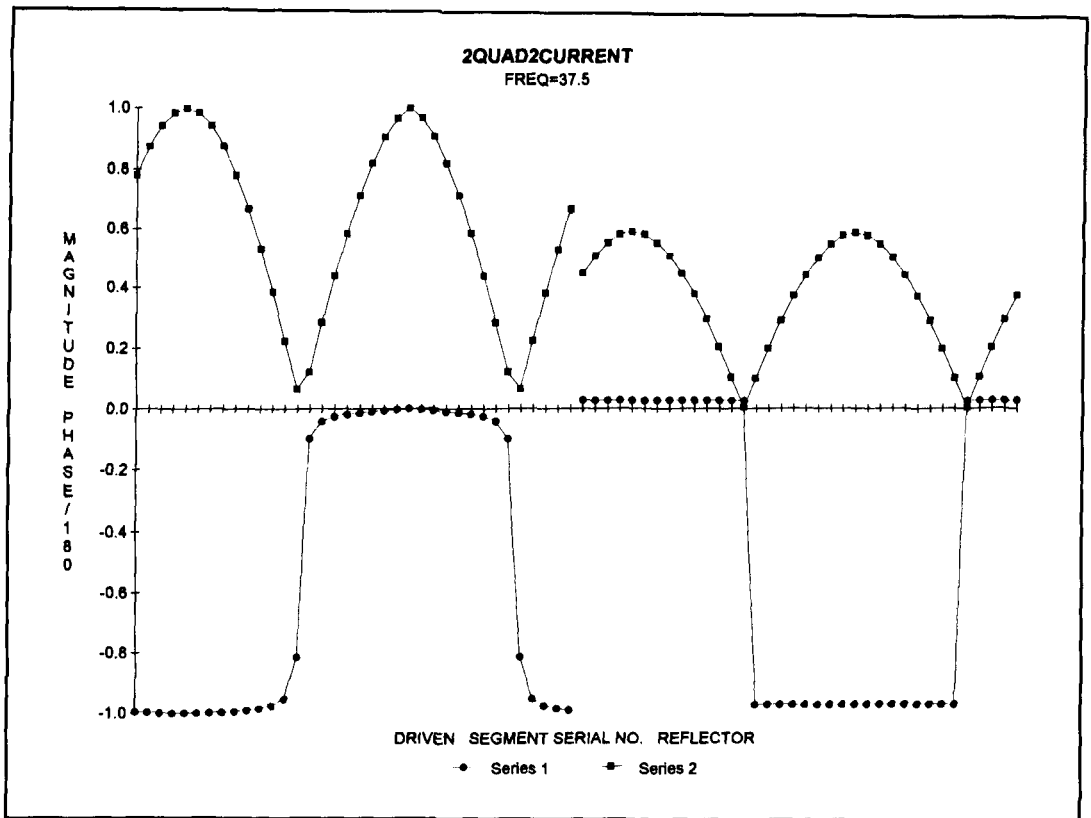


Figure 2. Current on 2QUAD elements.

2-QUAD MUTUAL CONDUCTANCE AND LOBE INTENSITY				
L/D=8000=C/2A				
SPACING KB WAVELENGTHS	CONDUCTANCE MILLIMHOS	ANGLE DEGREES	TOWARDS DB	AWAY DB
0.01	65.4	-90.1	1.51	7.12
0.02	25.7	-89.3	0.58	6.87
0.05	11	-88.8	0.82	6.93
0.075	8.15	-89.8	1.45	6.96
0.1	6.86	-92.3	2.25	6.93
0.15	5.65	-102.2	3.88	6.6
0.2	4.9	-117.4	5.05	5.99
0.25	4.12	-134.9	5.28	5.72
0.3	3.35	-151.5	5.93	4.6
0.35	2.71	-166	5.81	4.08
0.4	2.26	-178.8	5.4	3.76
0.45	1.94	169.3	4.71	3.64
0.5	1.73	157.8	3.72	3.72
0.6	1.51	132.5	1.03	4.26
0.7	1.43	102.2	1.2	4.77
0.8	1.35	66.3	3.92	4.67
1	1.03	-5.4	3.9	3.9
1.5	0.712	-179.2	3.78	3.78
2	0.54	3.8	3.67	3.67

Table 1. Mutual impedance and gain of 2QUAD versus element spacing.

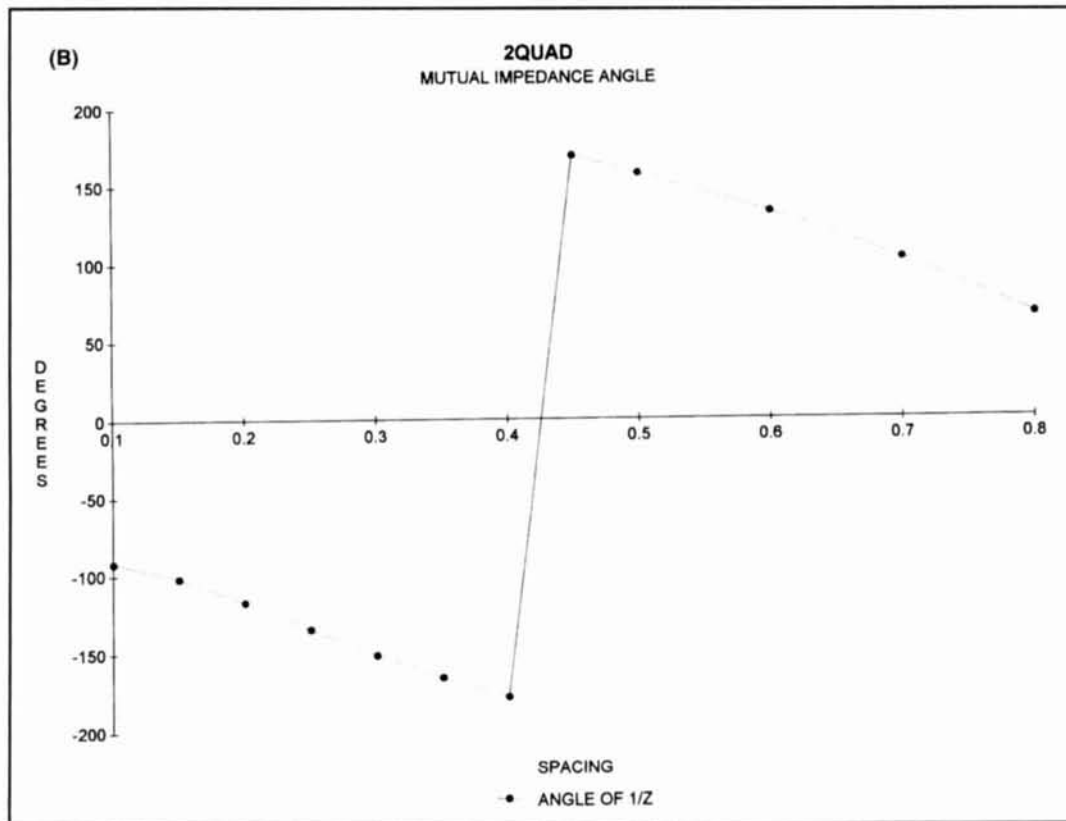
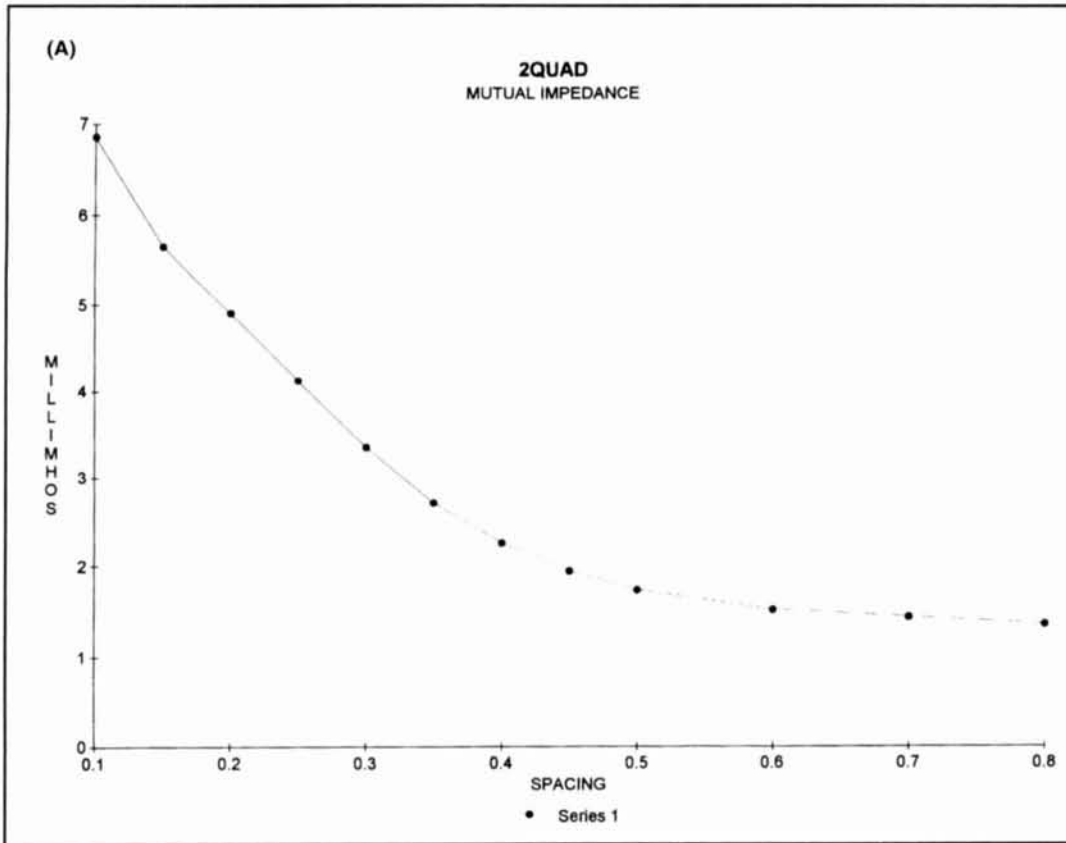


Figure 3. Mutual impedance: (A) magnitude, (B) angle.

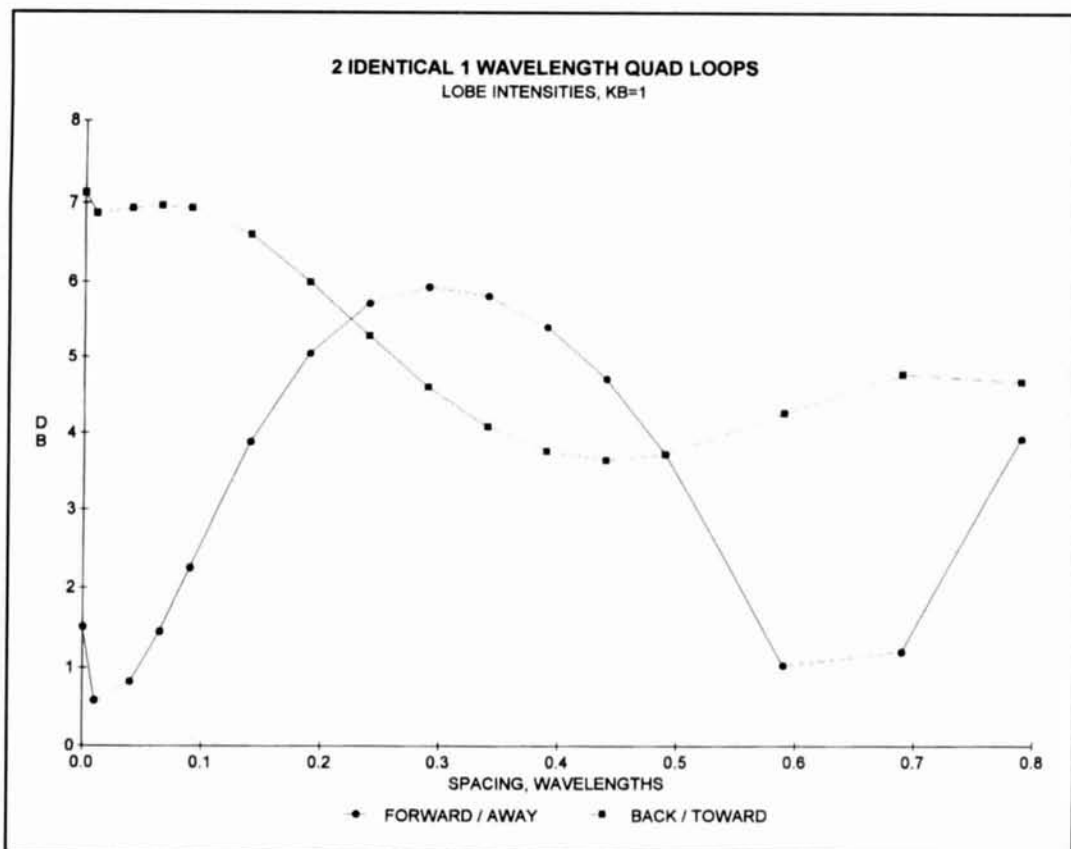


Figure 4. Effect of spacing on identical element lobes.

Twenty Years of Quads

It has been nearly 20 years since I began studying the quad antenna family in detail, and just over ten years since the first publication of the study results. During this time, there have been developments in the field of antenna analysis that provide much improved data—calculated values which come much closer to values found in test. There has also been a great increase in computer capacity and speed. Evaluations that took 10 or 12 hours at the start of study now take seconds. And some techniques of transfer from “antennas on paper” to hardware antennas in the real world have appeared, which make it easier to get the design performance.

These factors have added up to the point that it seems desirable to revise the results of the my studies, published in the book *The Quad Antenna*.* The changes are also being prepared as this series of articles.

*Available from CQ Communications, Inc., 25 Newbridge Rd., Hicksville, NY 11801.

as wire size, loop perimeter, and spacings change, but the general pattern is similar.

Mutual impedances

Each loop has a self impedance, Z_{11} and Z_{22} —the value given by tables and curves in Part 1 of this series. The impedance between the two, the mutual impedance, is defined by the relation $Z_{12} = Z_{21} = E_1/I_2$ (Z_{12} may not be equal to Z_{21} if ionization or rectification is present).

The NEC-calculated values of mutual impedance are given in **Table 1**. The behavior is easier to see in graphical form, as shown in **Figure 3**, with **A** giving the impedance magnitude and **B** the phase angle, over the usual range of loop separation. Very nearly, the mutual impedance magnitude decreases inversely with separation distance, and the phase angle decreases directly with distance at these separations.

The impedances allow for calculation of the current on each loop. The magnitude and phase can then be combined with equations or tables of “pattern factor” to provide a close approximation of the pattern developed by any combination of loop sizes and separations. However, today it is just as fast and more accurate to develop the patterns directly with one of

the "Method of Moments" programs; NEC is used here.

The lobe structure of the radiation pattern

Because the currents on the two elements are essentially out of phase, the two form an end-fed antenna array. While the complete pattern is of importance, it is often possible to concentrate on just two values: the lobe intensity on the axis of the loops and in the direction "away" from the parasitic element, and that "toward" the parasitic, both as seen from the excited element. In the usual two-element quad, the "away" pattern is the "forward" lobe.

However, remember that this forward and back designation is arbitrary. This is illustrated in **Figure 4**, which shows the two intensities for identical loops as a function of loop separation, with k_B equals one wavelength. At very close spacing, the parasitic acts as a director, the maximum radiation being toward it. It then changes to a reflector, with maximum radiation directed away from the parasitic. There is alternation between director and reflector as spacing increases. (This alternation will be obvious later, when multi-element arrays are studied.) The two lobes are identical at very large spacings, at the 3.1 dB value of an isolated loop.

In addition to the spacing, the size of the reflector can be changed. **Figure 5** shows the effect of this on the "away" or forward lobe gain. For any parasitic size, there is a spacing giving forward maximum gain. A parasitic with k_B around 1.1 wavelengths provides a greatest gain of around 8.0 dB at a spacing of 0.15 wavelength.

Changing the wire size of the parasitic conductor appears to have an effect on lobe structure. However, on examination, this is due to the change in resonant frequency of the loop with wire size. If adjustment is made for this, the wire size effect is negligible. The most important factors in the lobe structure are parasitic perimeter and loop spacing. Note that the value of k_B becomes the major scaling factor, specifically for size and operating frequency.

The total effect of these factors can be observed in **Figure 6**, showing the family of patterns for spacings between 0.1 and 0.4 wavelengths, and for parasitic perimeters between $k_B=0.85$ and 1.2 wavelengths. This family gives a good view of what can be expected with any two-element quad, plus some hints on further gain increase by adding elements. I will talk more about this in a future article.

A review of the patterns and data in this family shows:

- Maximum gain is secured with the parasitic operating as a reflector;

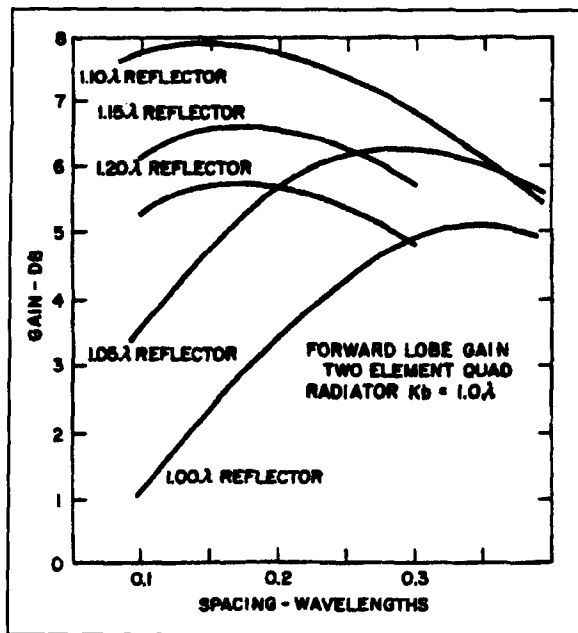


Figure 5. 2QUAD patterns. Rows-Element size, k_B : 0.85 to 1.2 wavelengths. Columns-Element spacing: 0.1 to 0.4 wavelength.

- The gain is nearly as great with parasitic director;
- The minimum back lobe occurs with a reflector;
- The back lobe with a director is much poorer (larger).

The result is that the classic two-element quad is always shown as a radiator-reflector combination. This design was based on early experimental work, later confirmed by analysis.

The family of plots shows that change in spacing is much less important than change in parasitic size. They also indicate that forward gain does not vary enormously as parasitic size is changed. The away lobe change is less than 5 dB of the size range covered—less than one S-unit. This is in contrast to the range of the back lobe, over 25 dB, or more than 5 S-units. Basically, these values mean that the quad will produce good performance for forward lobe signals over a wide range of frequencies, but that there will be considerable difference in response to back lobe interfering signals as frequency is changed. This can be an important factor on, say, 20 meters, where propagation can be good in all directions, but an unimportant factor on 10 meters unless the interference is via ground wave.

Details of lobe structure of two-element quads

To look at the performance details of the lobes, let us concentrate on the front and back

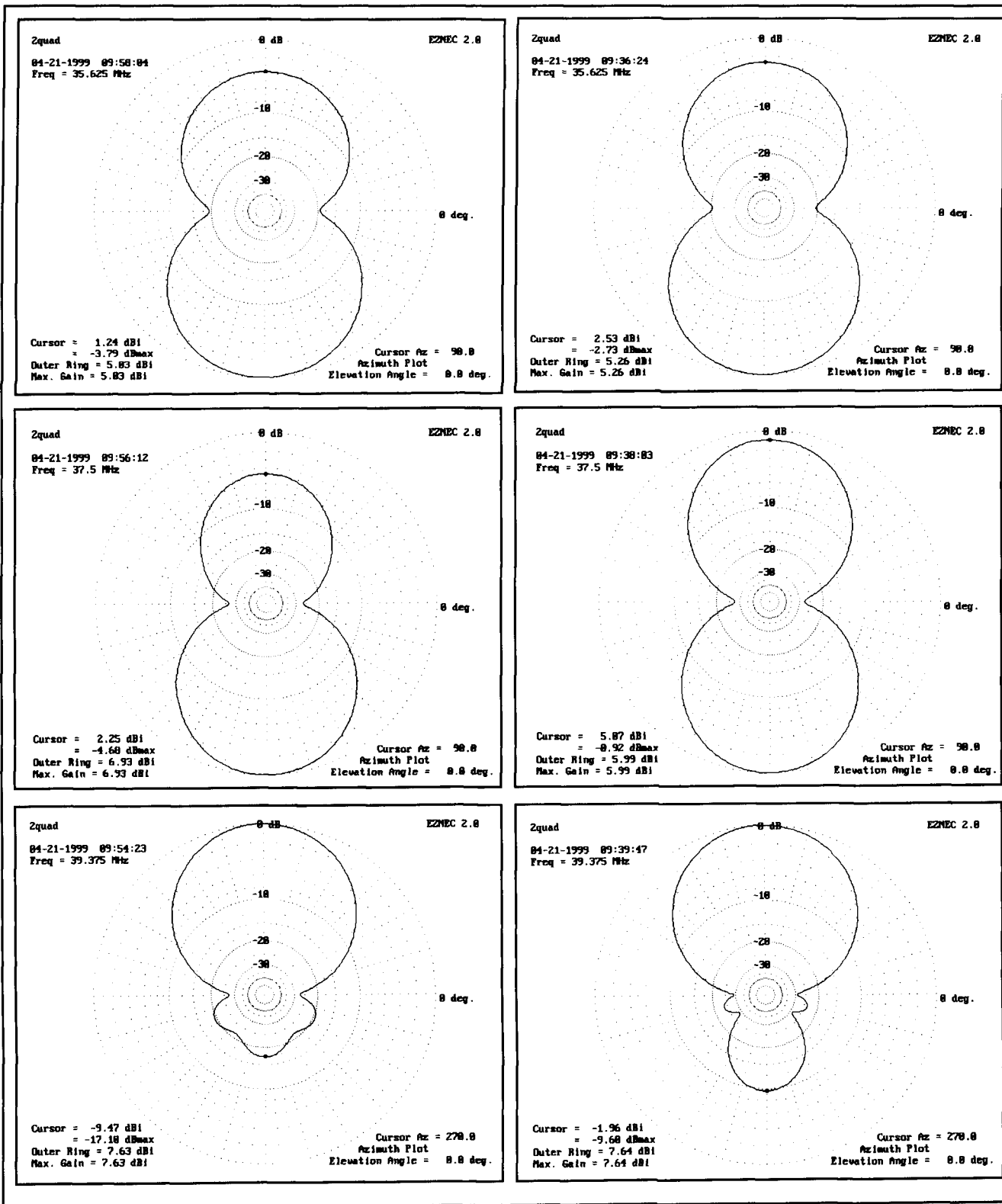
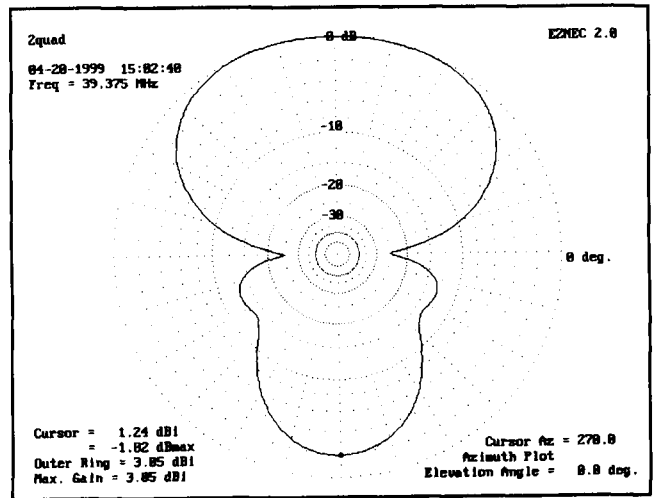
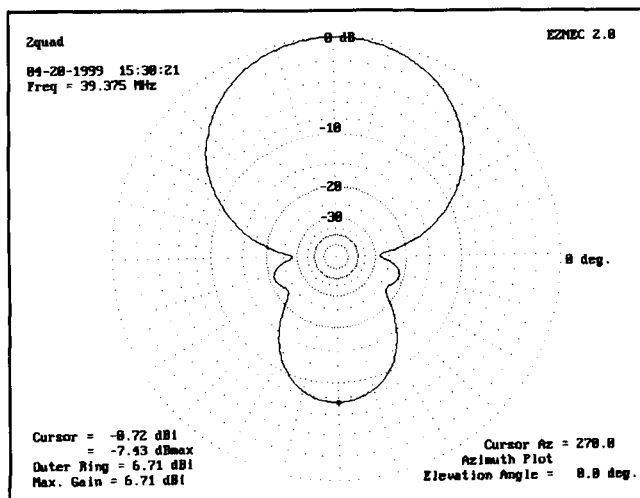
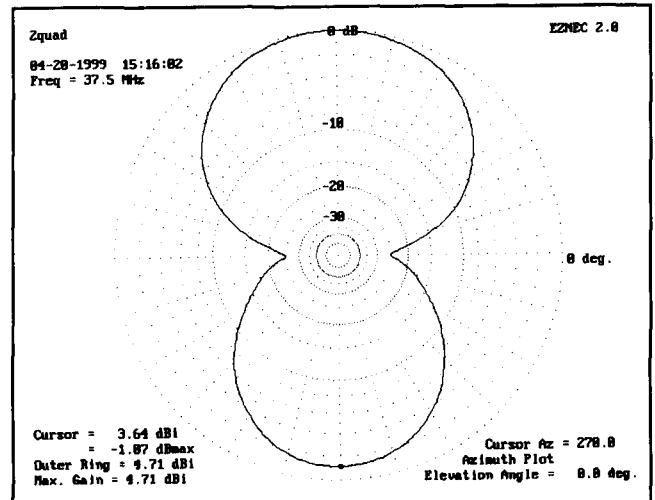
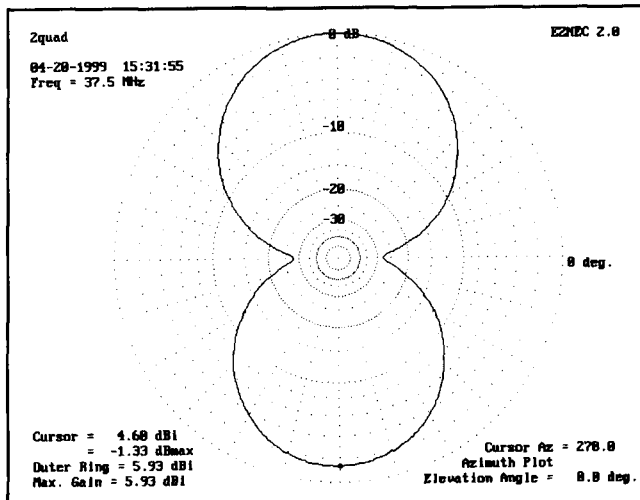
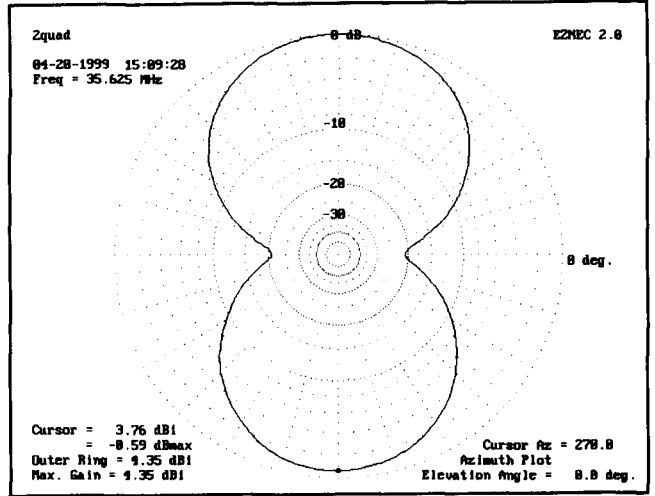
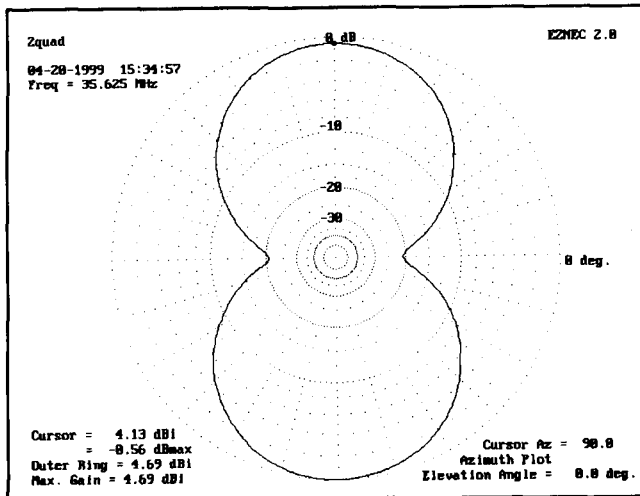


Figure 6. 2QUAD gains, element size and spacing varying.

lobe values only. We can do this by plotting the two gain values against the reactance of the parasitic element, as shown first in Figure 7. Here a plot using parasitic size in wavelengths or its self-resonant frequency could have been

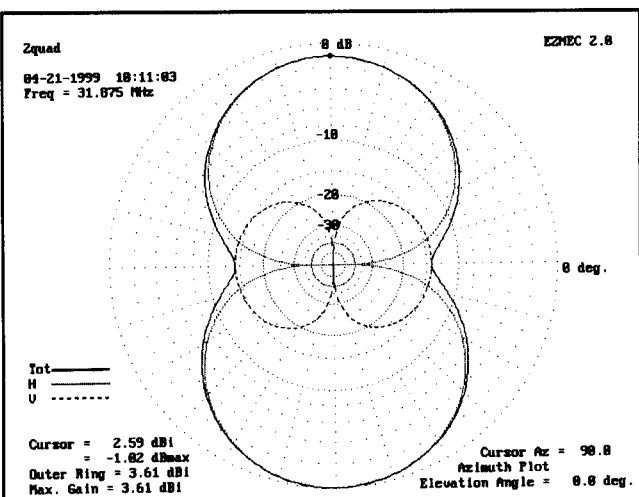
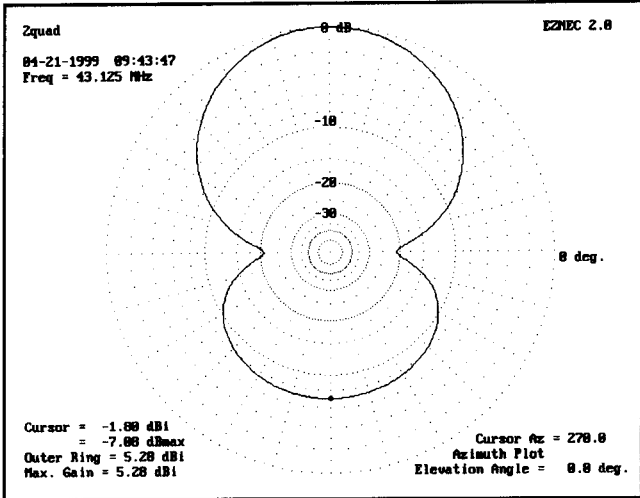
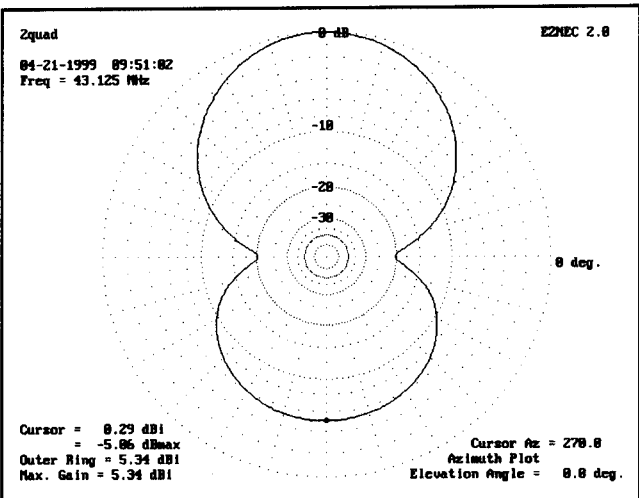
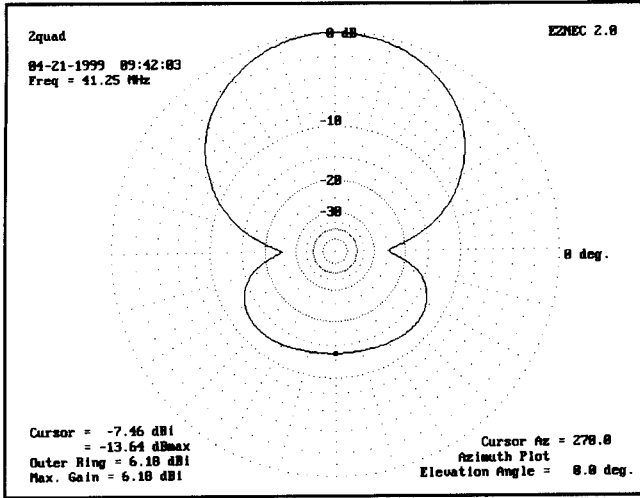
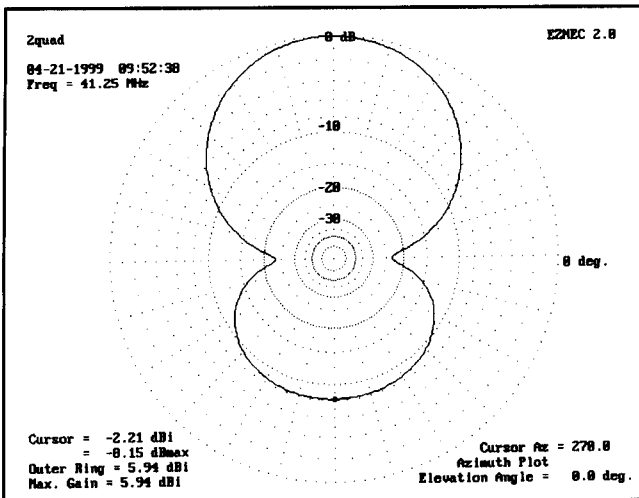
used. The main difficulty with physical size is that the parasitic effect then also varies with conductor size. The reactance value does vary with both loop size and conductor diameter, but the effect on the lobe structure is the same for



all elements of a given reactance (at least over the usual range of parasitic element sizes).

There are several ways of controlling the reactance present. Changing loop size is probably the simplest, but we can also add reactance

directly. The original way of doing this was by adding a section of transmission line to the parasitic element, "tuning" by changing its length for best gain or F/B. One or more lumped reactances, such as coils or capacitors, can also be



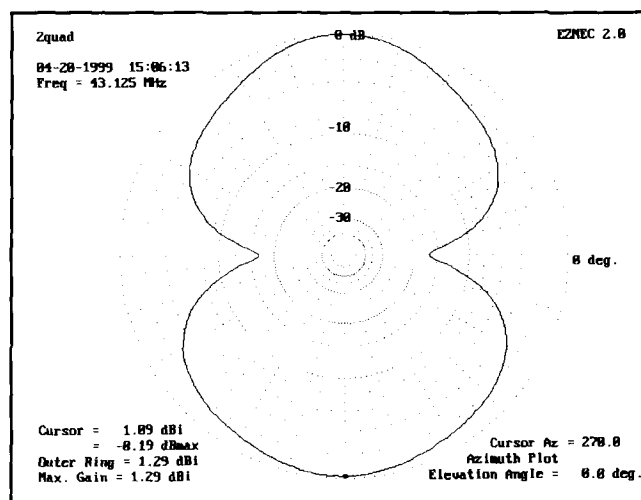
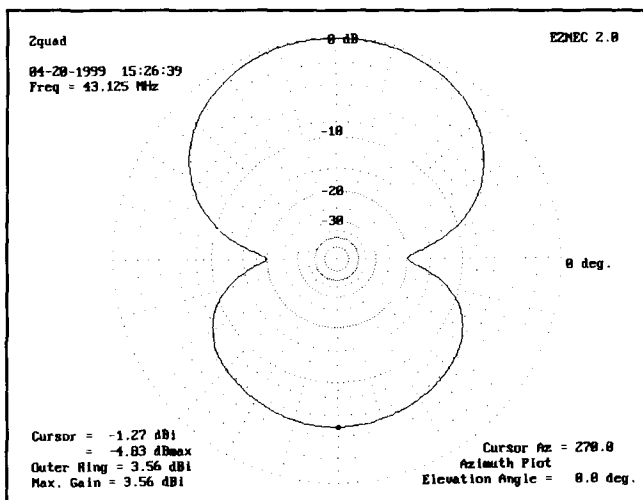
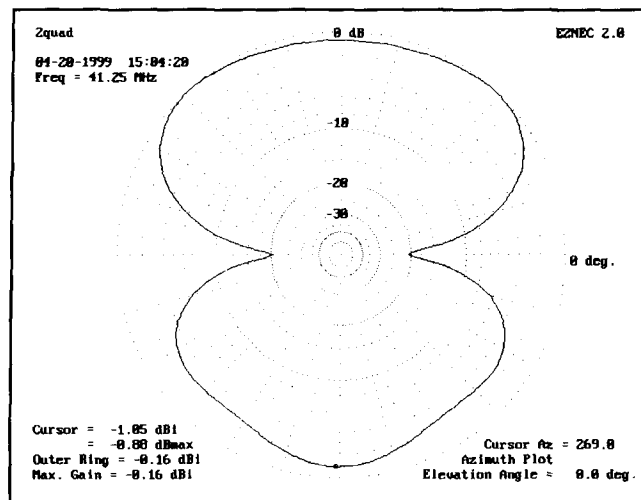
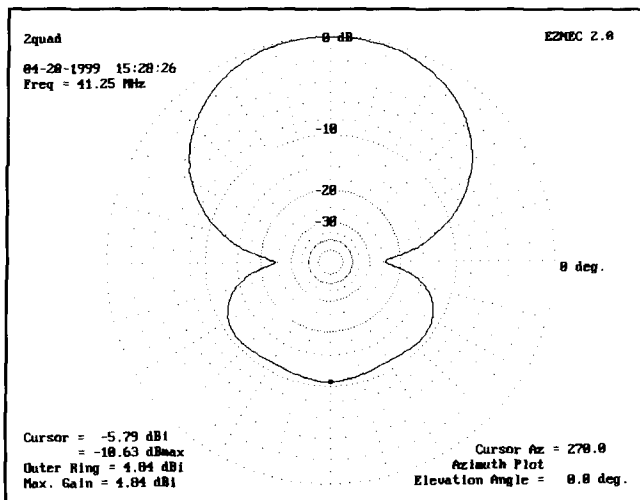
occur at the same frequency; this is characteristic of all quads. Another factor is clearly evident when **Figure 7** is considered together with **Figures 8, 9, and 10**: there is a spacing which provides the best F/B ratio. It is also noticeable that, while the gain is lower as the spacing increases, the gain changes less rapidly at wide spacings. The F/B ratio also changes less rapidly with frequency as spacing increases. These factors are important when good performance over a range of frequencies is needed: a Quad is inherently a broad-band antenna, and comparisons not reported here show that it is much more so than of the normal Yagi design.

For two-element quads, these four figures show that:

- Maximum gain is found with a spacing of 0.1 wavelength and parasitic reactance of 50 to 75 ohms. For a fairly small conductor of B/2A around 8000, maximum gain corresponds to $kB = 1.04$, which is also the ratio of resonant frequency to nominal frequency

used. A 2:1 change in operating frequency is not impossible using these methods and will be considered in a later part of this series.

A factor very noticeable in **Figure 7** is that maximum gain and maximum F/B ratio do not



(39/37.5 MHz for the calculation example); F/B is around 10 dB.

- Maximum F/B is found for a separation of 0.15 wavelengths and a parasitic reactance of +25 ohms, corresponding to $kB = 1.07$ —the calculation frequency being 40.125 MHz.

- Minimum change in gain is found for a spacing of 0.2 wavelength; a 1-dB change covers the kB range of 1.02 to 1.07. A spacing of 0.25 is nearly the same.

- The F/B ratio range is also best at a spacing of 0.2 wavelengths, exceeding 15 dB for the kB range of 1.05 to 1.10. At the spacing for best F/B, the kB range for F/B exceeding 15 dB is only 1.06 to 1.08.

The factors of gain/, F/B ratio, and bandwidth mean that, if one is selected as the most important, and the design values selected to give best values for it, the other values are set and must be accepted. This usually means that quad antenna design will depend on compro-

mise among the factors, to provide “pretty good” values of all factors.

Drive impedance of the two-element quad

The drive point resistance and reactance for the four spacings considered is shown in **Figures 11 through 14**. Two general patterns are noted: drive resistance is low below about $kB=1.05$, and the lower resistances are associated with negative drive reactance. Drive reactances are not essentially straight lines, as found for the self-reactance of an isolated loop close to resonance.

The variation in drive resistance is much greater at close spacings. This means the SWR variation is also greater. Whether this is important or not depends on the feed method, discussed in part later, and more completely in Chapter 9 of *The Quad Antenna Book*.² As stat-

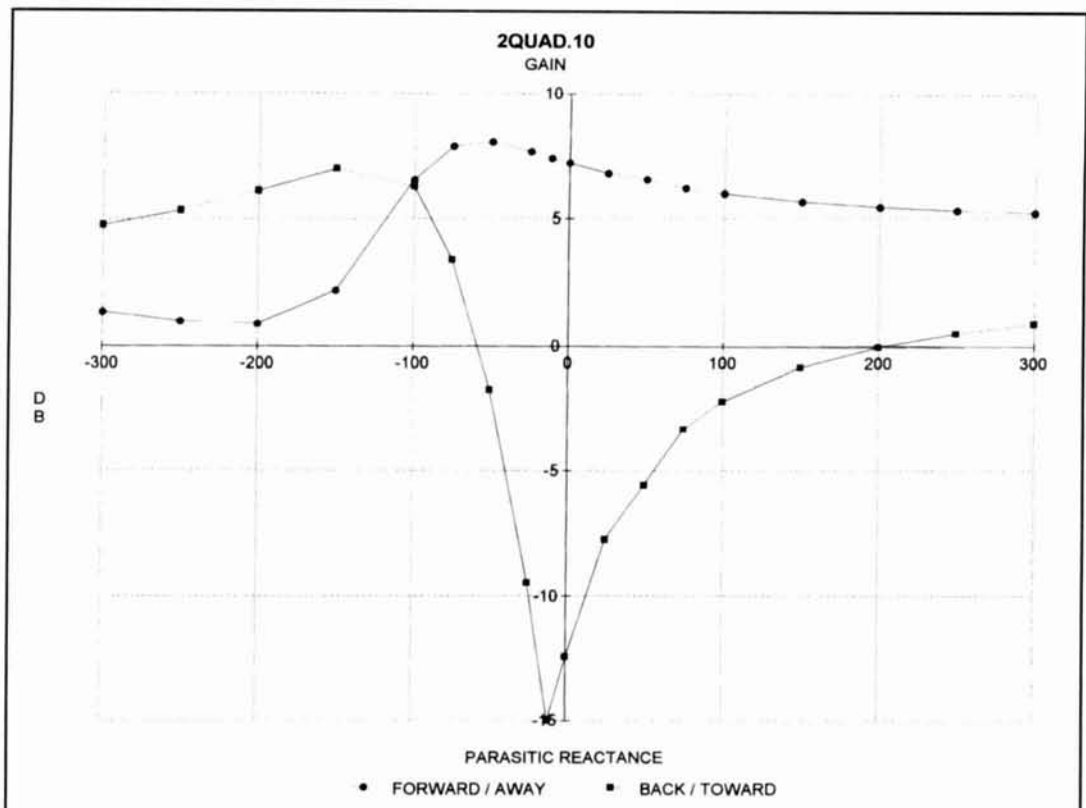


Figure 7. 2QUAD lobe intensities, spacing-0.10 wavelength.

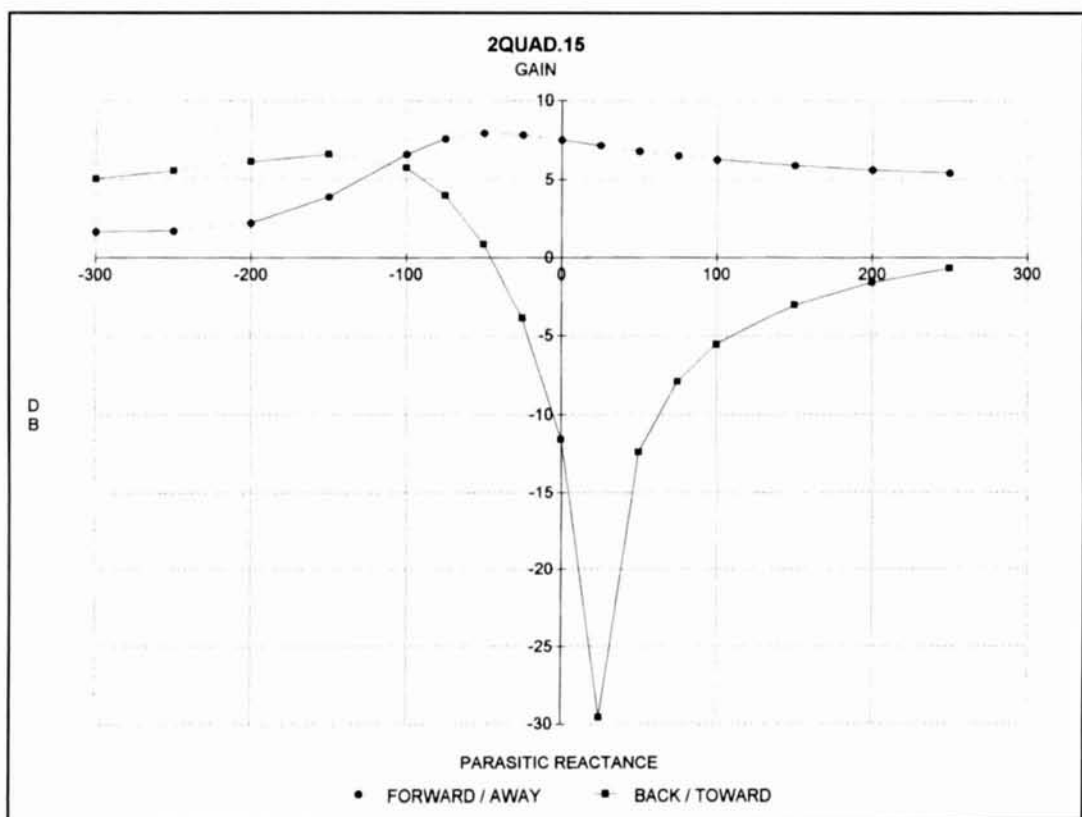


Figure 8. 2QUAD lobe intensities, spacing-0.15 wavelength.

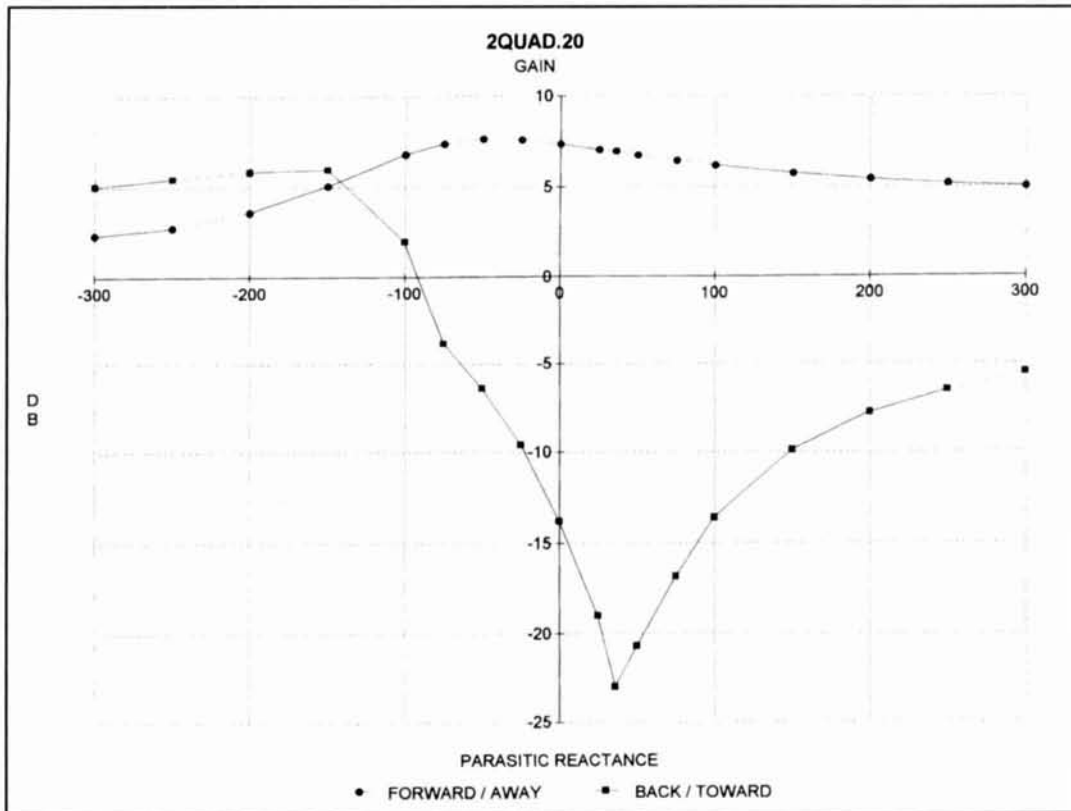


Figure 9. 2QUAD lobe intensities, spacing-0.20 wavelength.

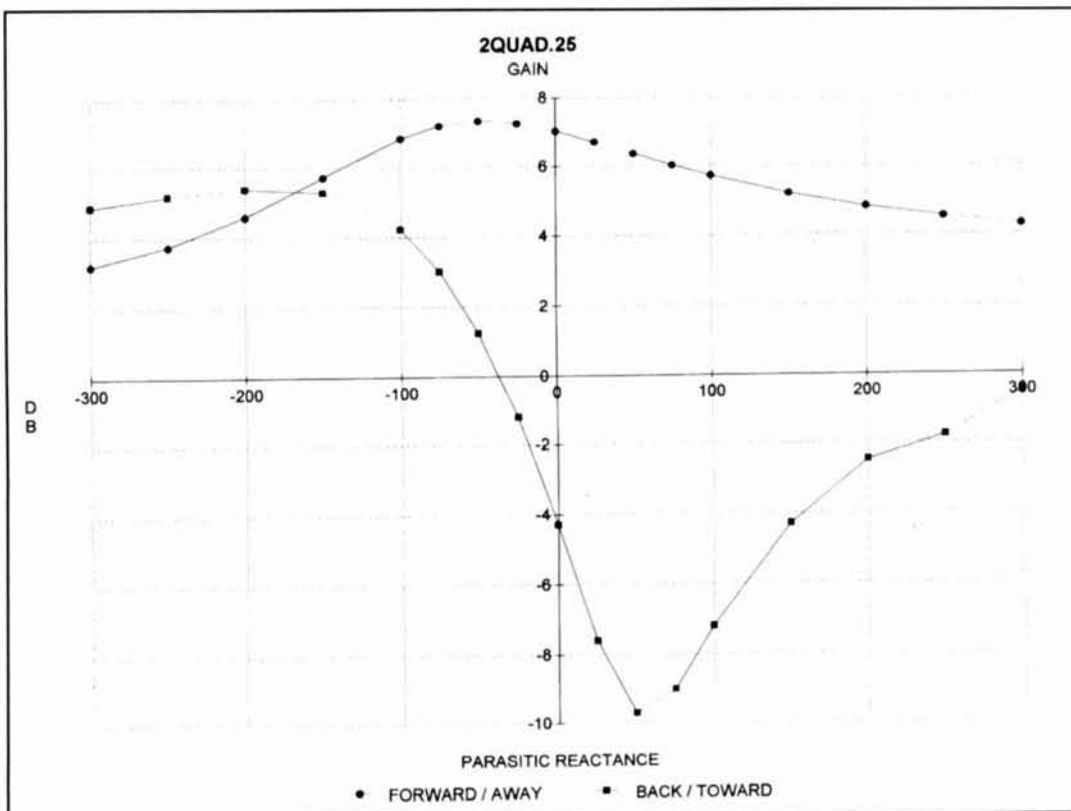


Figure 10. 2QUAD lobe intensities, spacing-0.25 wavelength.

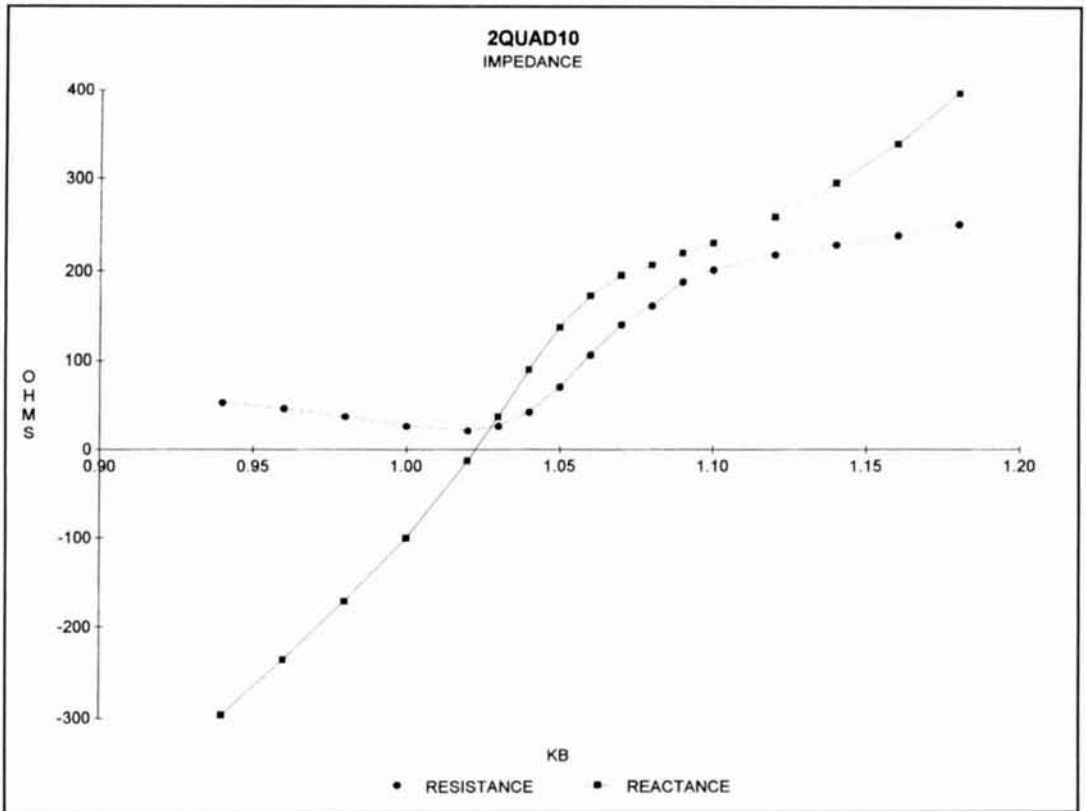


Figure 11. 2QUAD drive impedance, spacing-0.10 wavelength.

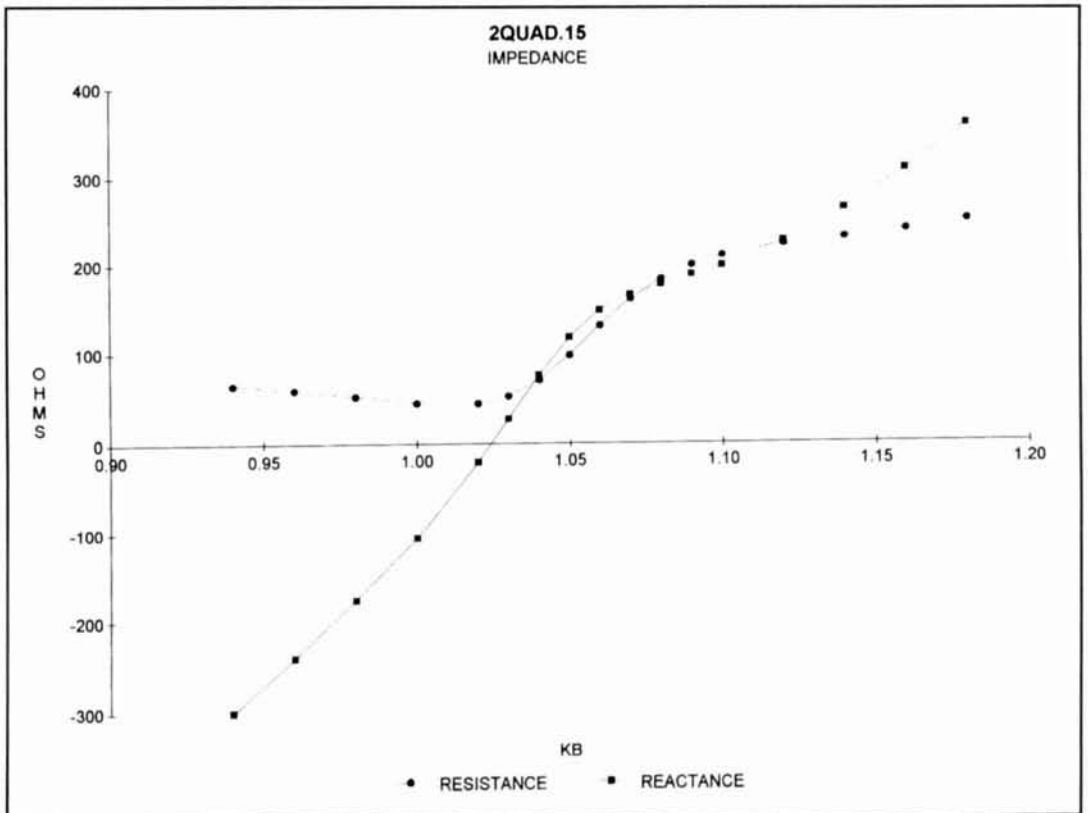


Figure 12. 2QUAD drive impedance, spacing-0.15 wavelength.

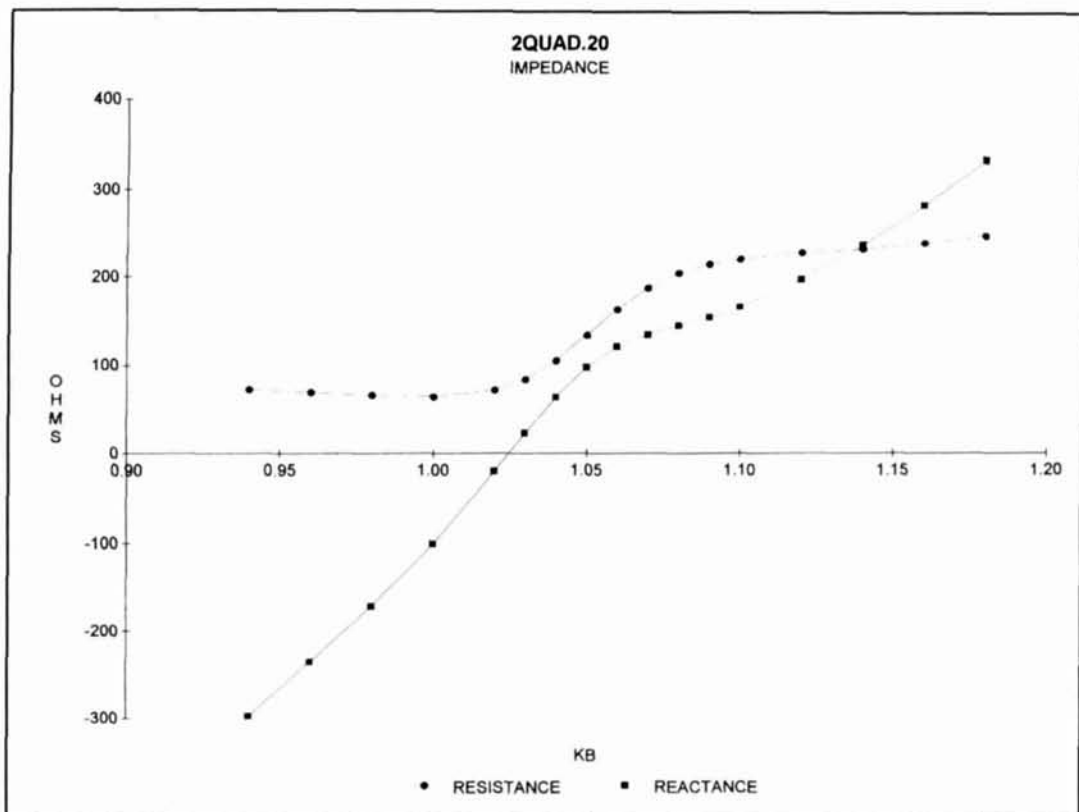


Figure 13. 2QUAD drive impedance, spacing-0.20 wavelength.

ed there, it is unimportant if low-loss feed line and a good "antenna tuner" are used.

Note that there is very little variation in the zero impedance point as spacing is changed.

Summary of quad family general results

Figure 15 shows the forward gain of the 2QUAD antenna as a function of size, kB , for the four spacings studied. Gains for intermediate spacings can be estimated by interpolation, although there is little to be gained in forward lobe intensity. The curves emphasize that maximum gain is provided by a spacing of 0.1

wavelength (but this also gives the narrowest bandwidth).

Figure 16 is the same type of plot for the back lobe. Best F/B is for a spacing of 0.15, but best low-backlobe bandwidth is for the spacing of 0.2. The rapid variation in F/B ratio indicates that intermediate spacings should be investigated when high F/B and wide bandwidth is needed. Comparison of the two curves indicates that a reasonable compromise between gain and F/B ratio requires kB in the range of 1.05 to 1.10.

Figures 17 and 18 summarize the results for drive impedance. A reasonable match for 50 or 75-ohm line with no other matching can be secured with 0.1 spacing and kB around 1.04; however the bandwidth will be rather small,

PROTOTYPE 2-QUADS FOR DESIGN SCALING						
Copper $C/2A=8000$						
GOAL	kB	SPACING	GAIN	F/B	Xiso	Fdes
	WaveLengths		DB	DB	Ohms	
MAX GAIN	1.04	0.15	7.73	8.1	-35	39
MAX F/B	1.07	0.15	6.9	24.4	45	40.125
COMPROMISE	1.075	0.2	6.7	20.21	58	40.3125
EXCITER	1.1	0.15	2.28	5.55	125	41.25

Table 2. Prototype design values by goal.

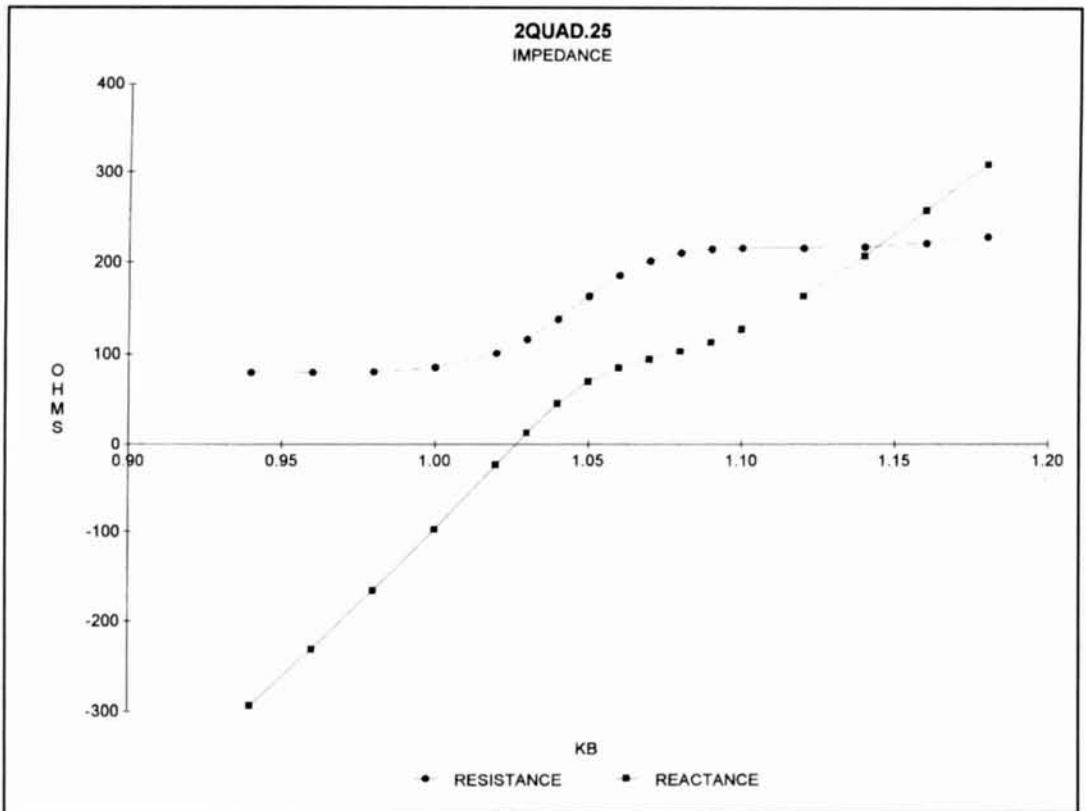


Figure 14. 2QUAD drive impedance, spacing-0.25 wavelength.

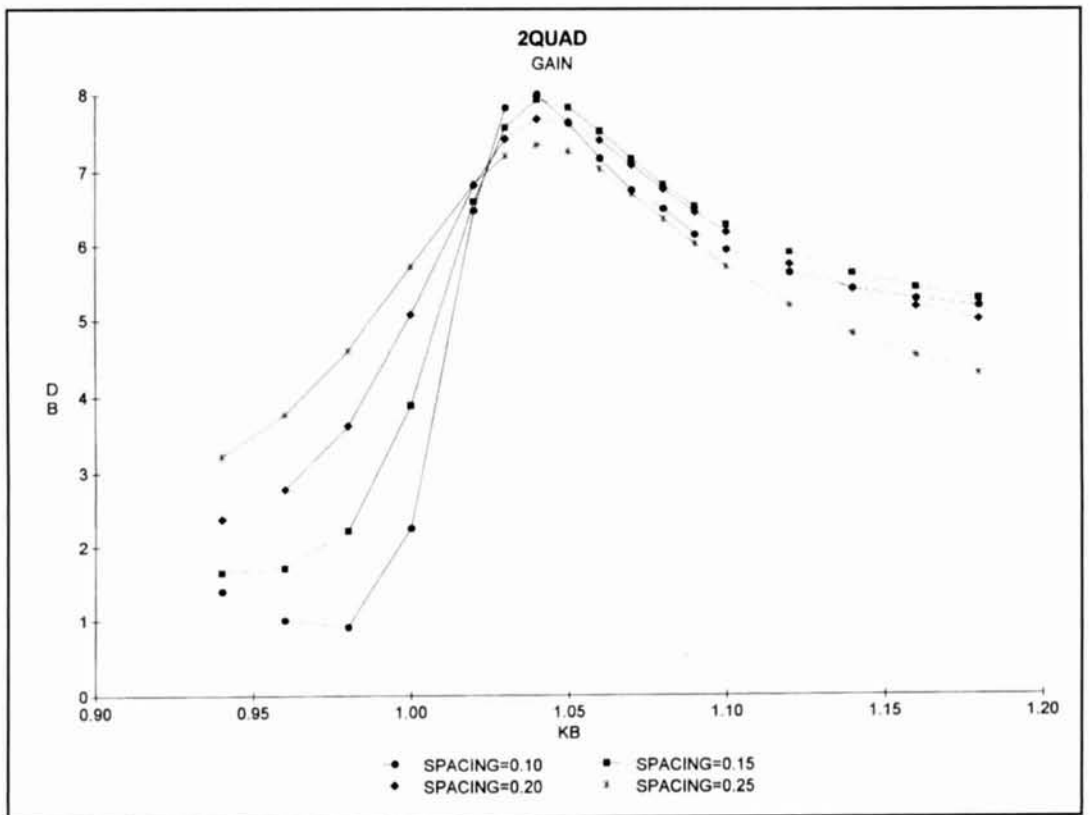


Figure 15. Comparison-2QUAD forward lobe (gain).

DESIGN DATA FOR 2 ELEMENT QUADS									
BAND	GOAL	CORNERS +/- FEET	LENGTH FEET	SPACING FEET	WIRE # CU	GAIN DB	MAX F/B DB	R IN OHMS	X IN + OHMS
40 METER									
7.1 MC	MAX G	18.02	144.18	21.25	12	7.64	10.4	88	100
	MAX F/B	18.54	145.21	22.25	12	6.44	18	182	177
	COMP	18.63	149.02	29.81	12	6.43	17.5	204	149
	BIG RAD	19.06	152.49	22.87	12	5.82	9.1	214	221
20 METER									
14.15 MC	MAX G	9.04	72.34	10.85	12	7.73	8.1	77	84
	MAX F/B	9.3	74.42	11.16	12	6.92	25	170	170
	COMP	9.35	74.78	14.96	18	6.38	17.2	205	150
	BIG RAD	9.56	76.51	11.48	14	5.98	10.1	214	211
15 METER									
21.1 MC	MAX G	6.04	48.29	7.24	12	7.76	7.12	72	74
	MAX F/B	6.21	49.68	7.45	18	6.75	20.93	176	173
	COMP	6.24	49.91	9.98	18	6.55	19.02	203	144
	BIG RAD	6.38	51.05	7.67	18	5.93	9.9	214	212
10 METER									
28.5 MC	MAX G	4.49	35.92	5.39	18	7.65	8.2	79	84
	MAX F/B	4.64	37.13	5.57	18	6.72	19.4	182	176
	COMP	4.64	37.13	7.43	18	6.67	20.1	200	141
	BIG RAD	4.75	37.99	5.7	18	6.05	10.6	213	206

Table 3. Design values by band and goal.

so this combination would be best for a single-frequency design. If matching is important, try wider spacing with a 50:200-ohm balun.

A specific design— The Orr Quad

In his first quad book,³ Bill Orr, W6SAI, gives design data for many quads, all based on experimental work. The performance of a two-element quad scaled to place the design at 37.5 MHz is shown in **Figures 19** through **21**.

Figure 19 shows that the gain is reasonably constant over 36 to 37.8 MHz. The indicated F/B ratio is about 20 dB, a maximum of about 4 S-units, and about 2 units or better over the range 36.7 to 37.4 MHz. The plot with smaller steps in **Figure 20** shows the lobe intensity variation with frequency.

Figure 21 shows the impedance variation with frequency, and **Figure 22** the resulting SWR with 75-ohm feed line. The 2:1 SWR bandwidth is about 0.5 MHz, or about 1.3 percent, a little less than 200 kHz on 20 meters. A different matching method or an antenna tuner would allow use over the entire band, with good forward lobe performance and acceptable F/B at a selected frequency.

Very nearly, this design was optimized for maximum gain. Orr recommends use of a short

adjustable stub, adjusted for best gain or best F/B, as needed.

Steps in antenna scaling and matching

Suppose we decide that the prototype 2QUAD at 0.2 wavelength spacing is a satisfactory design, and that we want to obtain the good backlobe performance in **Figure 9** at a specified frequency. Using the reactance versus frequency curve, **Figure 13** of Part 1 of the series, for wire size used, shows that the best F/B is at 40.5 MHz. This occurs at $k_B = 1.08$. Because the reference frequency is $k_B = 1.0$, the antenna is too small to provide the desired performance at the reference frequency. Accordingly, all dimensions must be multiplied by 1.08. At the reference frequency of 37.5 MHz, the loop corners must be at ± 1.08 wavelengths, the spacing at 0.216 wavelength, and the wire size at 0.864 mm. (A lot of time in antenna construction can be saved by using rulers and tapes calibrated in both feet/inches and meters/millimeters.)

To convert these values to any other frequency, multiply the dimensions by the ratio $37.5/f_{new}$. For example, for a design frequency of 14.2 MHz, the dimension multiplier is $37.5/14.2=2.64$. Loop corners are at

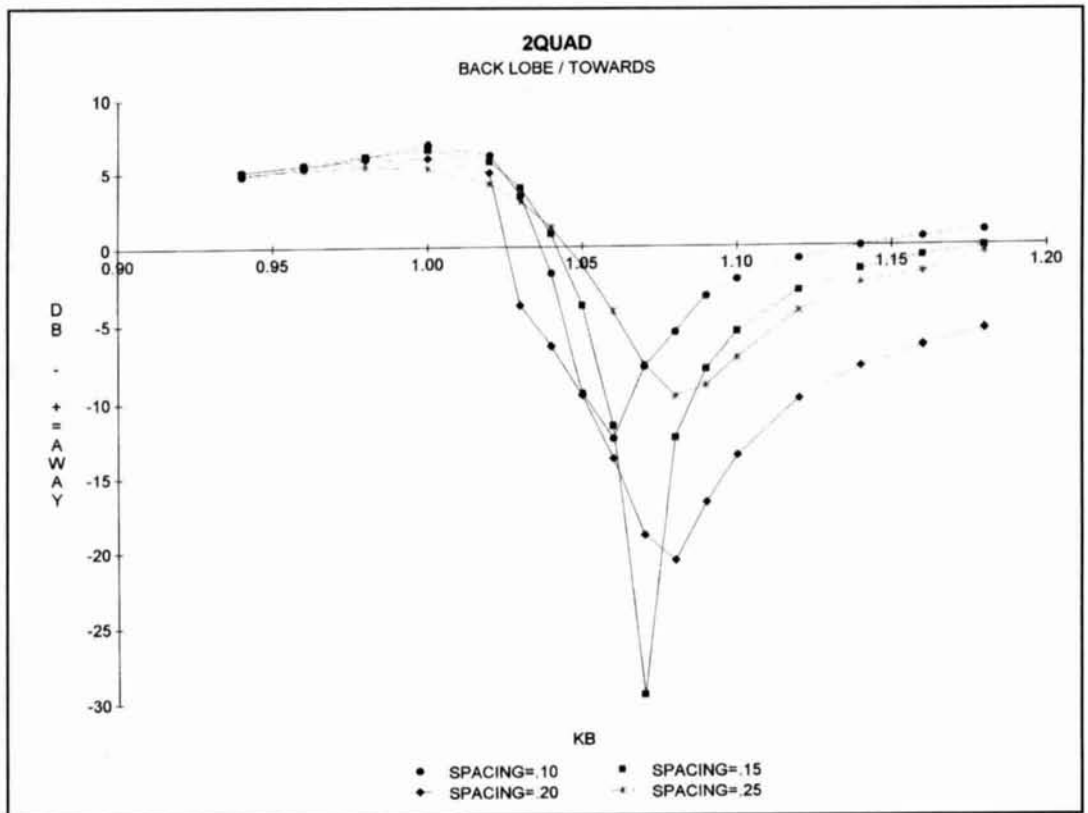


Figure 16. Comparison-2QUAD back lobe.

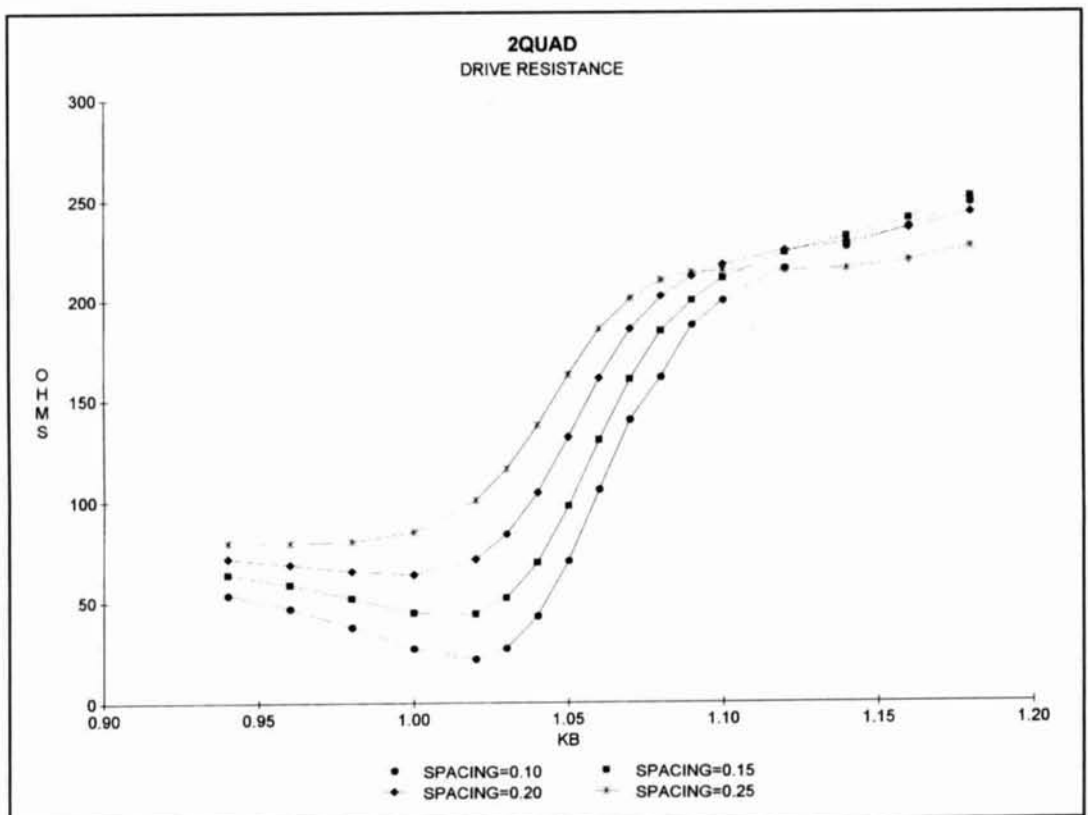


Figure 17. 2QUAD drive resistance.

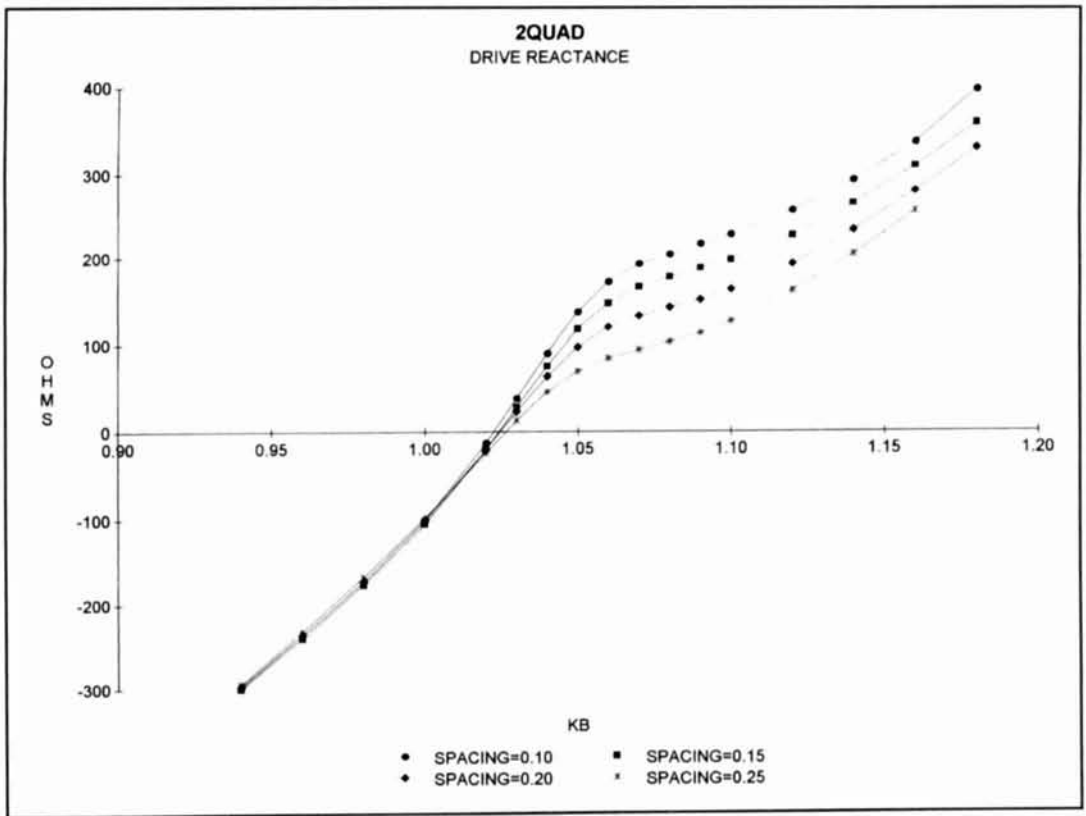


Figure 18. 2QUAD drive reactance.

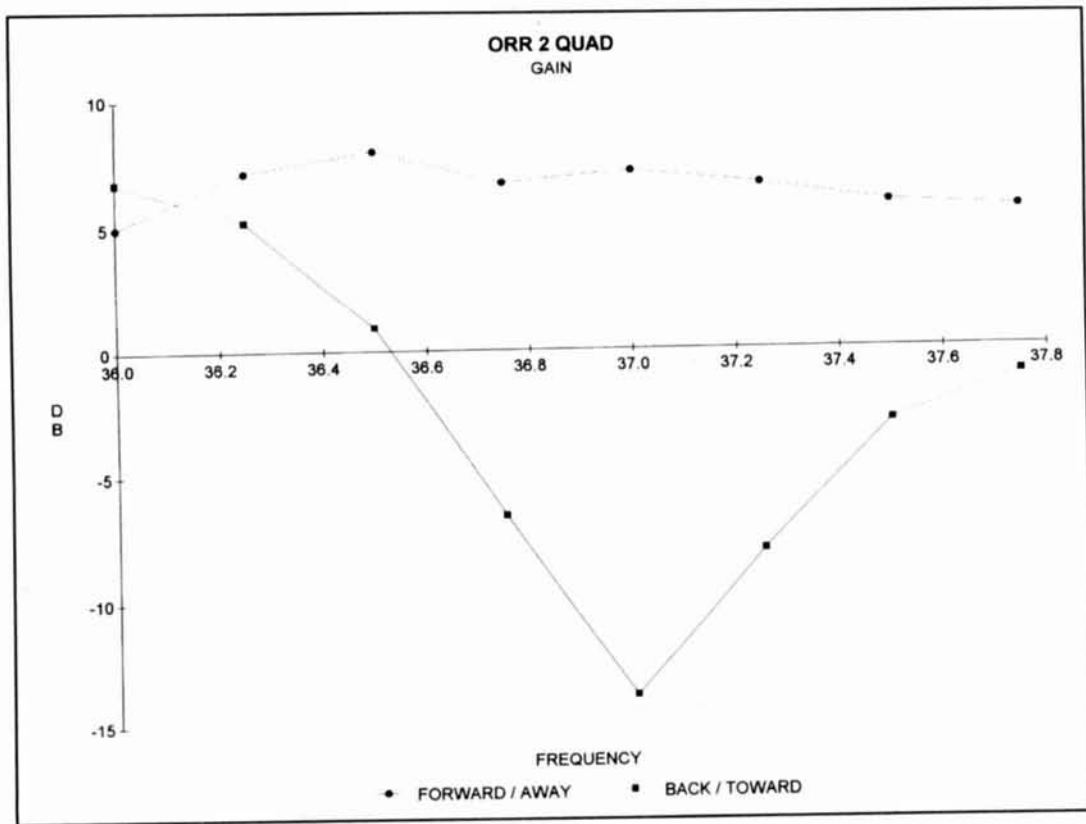


Figure 19. Orr 2QUAD lobe intensities.

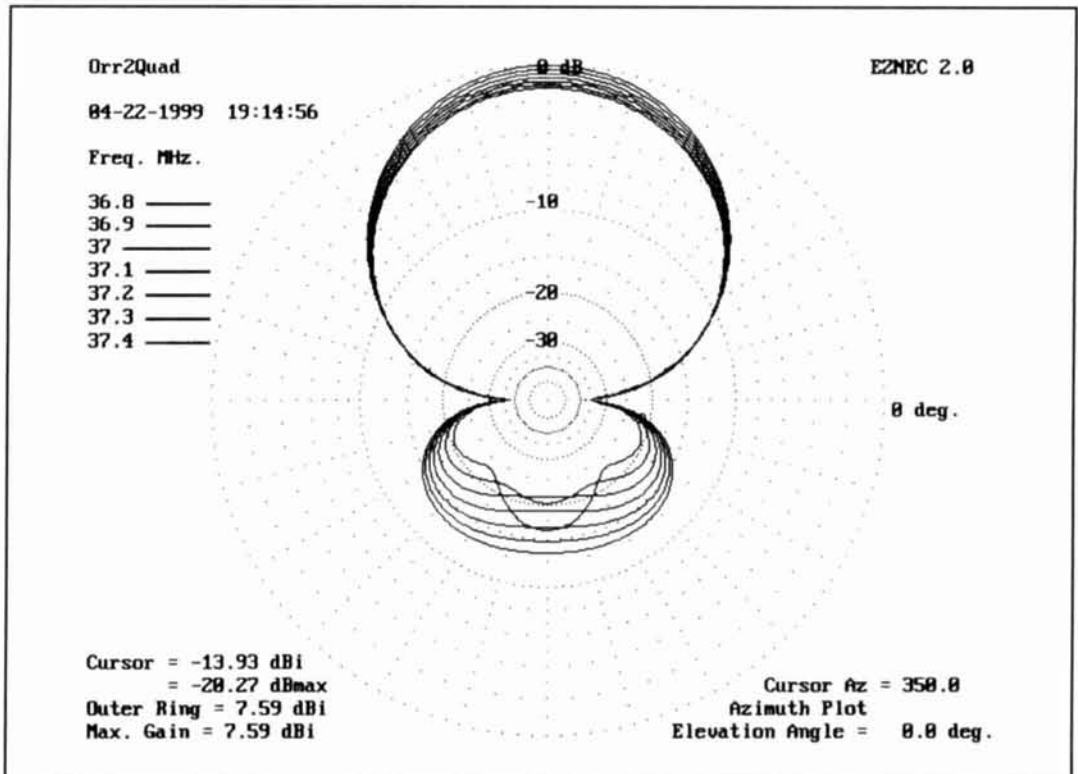


Figure 20. Orr 2QUAD family of patterns.

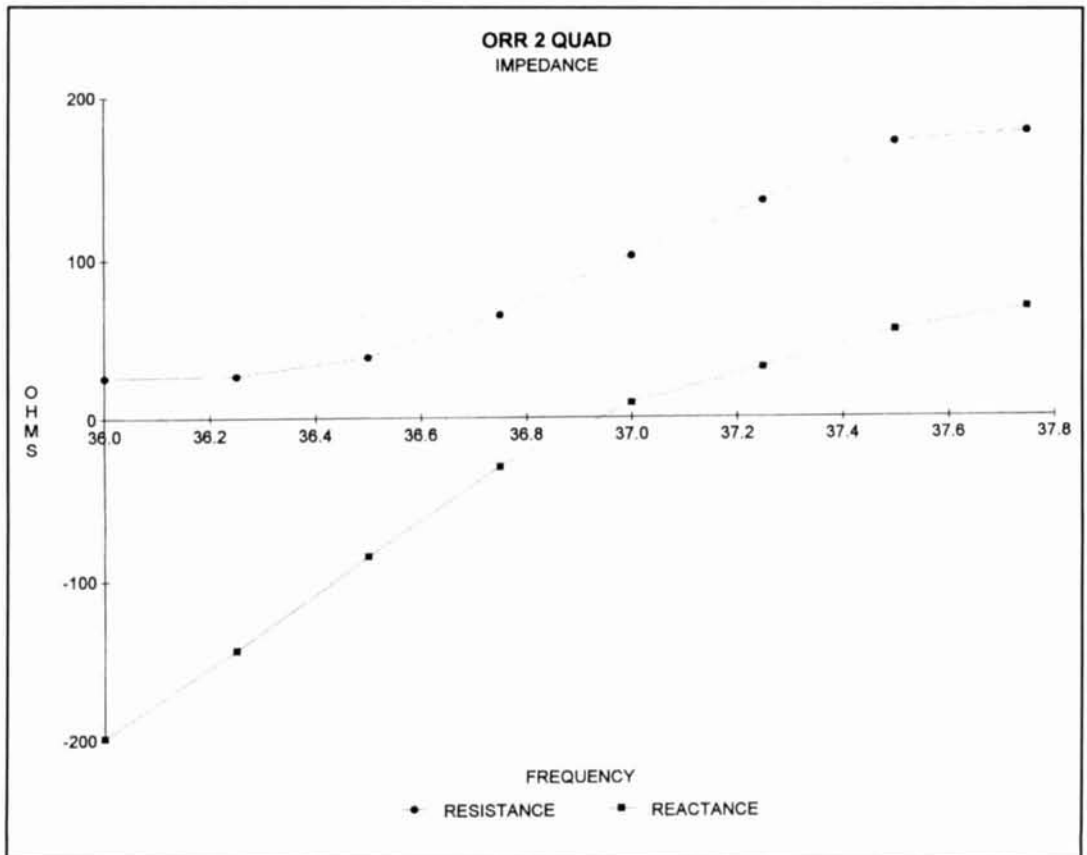


Figure 21. Orr 2QUAD lobe drive resistance and reactance.

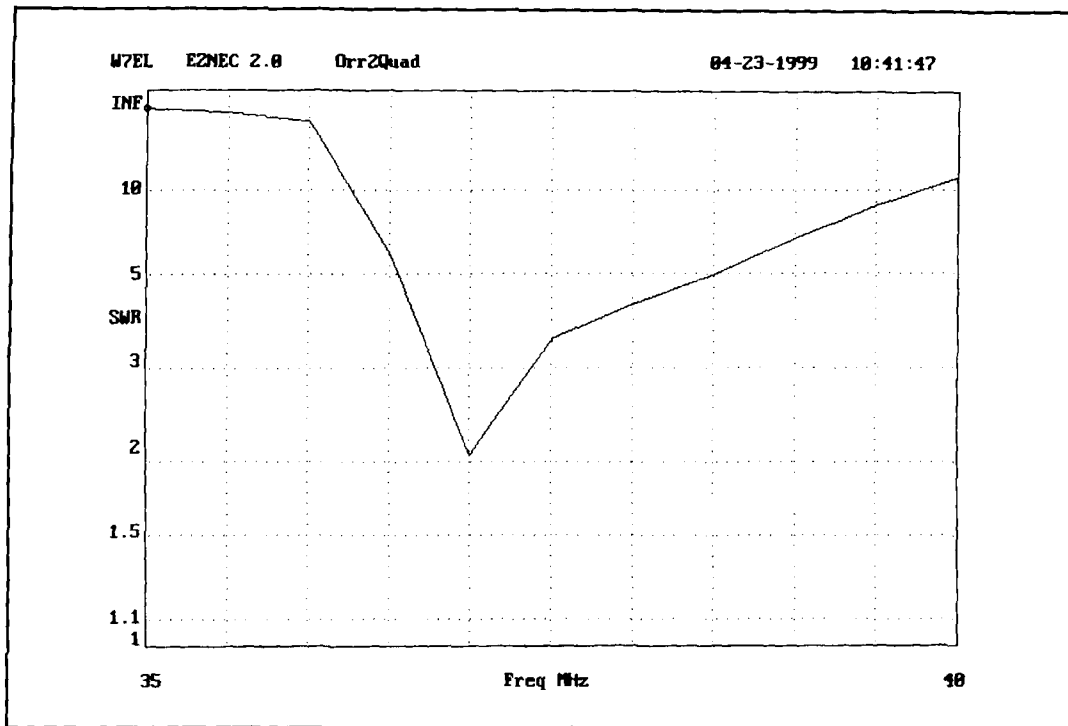


Figure 22. Orr 2QUAD lobe SWR with 75-ohm feedline.

(± 1.08) $\times 2.64$ meters = ± 2.85 , for a total loop wire length of 22.81 meters or 74.8 feet. Spacing is (1.08×1.6) $\times 2.64 = 4.562$ meters or 15 feet. Wire size would be 89.8 mils, or #11 (#12 should be okay).

We could have gone directly from the prototype frequency of 40.5 to 14.2 MHz. For example, the dimension multiplier is 40.5/14.2 or 2.85, which is the \pm value of the loop corners, as above. In the same way, the spacing is 2.85×1.6 , or 4.56 meters.

Returning to the reference design, suppose we decide not to change the spacing in the scaling to the reference frequency, but keep it at 0.2 wavelength. A look at the curves shown earlier indicates that the performance change should be small. Using these values in a NEC program provides a forward gain of 6.82 dB, a F/B of 21.8, and a drive impedance $197 + j153$.

Now suppose we would like to feed this antenna with coax and have a matched condition. If we get rid of the 153 ohms of reactance, this can be done with a 4:1 balun by placing an open stub across the feed point. *The ARRL Antenna Handbook*,⁴ gives the process for determining the stub length needed.

An alternative is to change the size of the radiator. The data for this is the reactance versus size curve in **Figure 13** of the first part of this series. Interpolating for wire size, we estimate that the 150-ohm reactance reduction requires a change in radiator size from 1.08 wavelengths to 1.025 wavelengths. Placing this

size in NEC gives a new drive resistance of $170 + j1.7$ ohms at 37.5 MHz, $145 - j15$ at 37 MHz, and $183 + j16$ at 38 MHz. At 37.5, the forward gain is 6.75 dB and the F/B is 22.8 dB.

Figure 23 shows the calculated SWR curve and **Figure 24** the pattern family.

The Big Radiator Quad

Figure 8 of Part 1 of this series showed that the gain of an isolated square loop increased as the size of the loop increased, up to about $kB = 1.5$. This means that some increase in the gain of the two-element quad can be expected if the radiator loop is made larger than that needed for resonance or for zero drive reactance. There are, of course, penalties for such operation. Both the drive resistance and drive reactance increase, and the mechanical problems increase as loop size increases.

Quite a few studies, by myself and others, reported in the original quad antenna book,² indicate that a reasonable design choice is to make the radiator and reflector the same size, choosing this on the basis of the reflector giving the desired design balance. Looking forward to a later part of this series dealing with multi-element quads, these studies indicate that loops of $kB = 1.1$ or even 1.2 are a good size for the radiator-reflector combination. We may call such a pair "the exciter quad," used to excite a set of directors, which may be quad loops

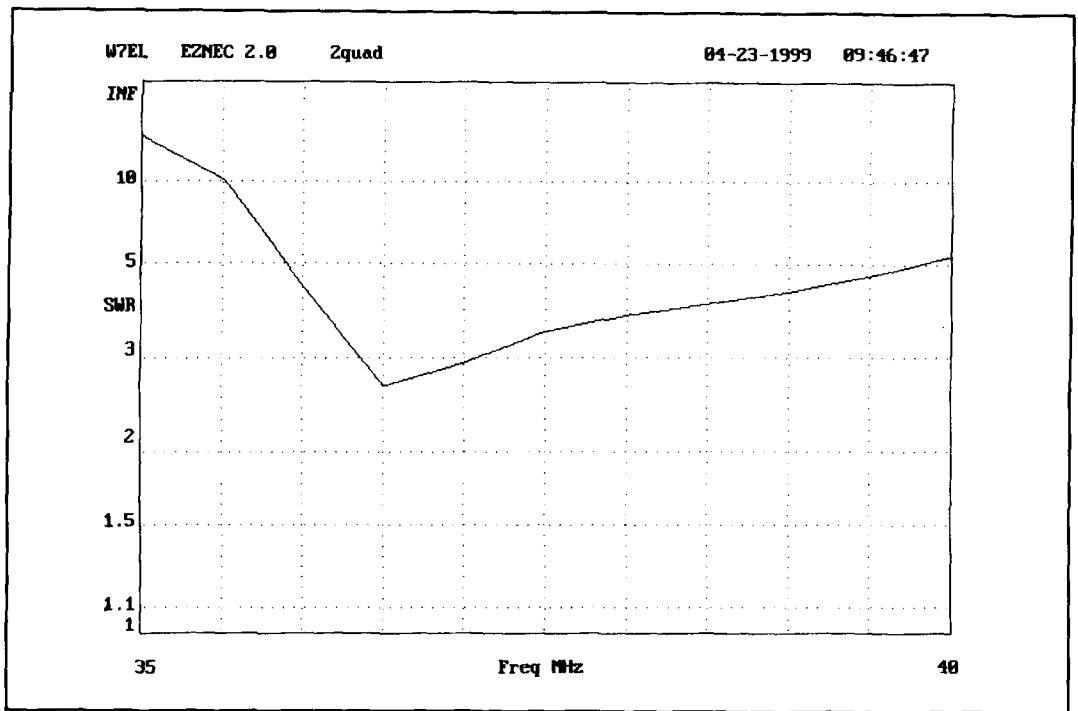


Figure 23. SWR of scaled 2QUAD with smaller radiator.

or Yagi-type dipole elements. We leave most of this work for later, but I do include here a set of 1.1-wavelength loops as an example of the change from usual two-element quad design.

Example designs for common bands

It is handy to have a set of tables which shows designs for specific bands. Here the bands at 40, 20, 15, and 10 meters are covered. The specific frequencies chosen are 7.1, 14.15, 21.2, and 28.5 MHz—the intent is to place the chosen maximum design performance goal at these frequencies. Four design goals are used: maximum gain; maximum F/B ratio; a compromise to balance gain, F/B, bandwidth, and drive resistance; and the 1.1 wavelength exciter quad.

Table 2 shows the resulting design values of kB and spacing plus the general characteristics of the four designs. Here Xiso is the reactance of an isolated loop of the indicated kB size, and Fdes is the nominal or design frequency of that loop. The resonant frequency is given by $1.053 \cdot F_{des} / kB$. (Note that a small change in dimensions can induce a considerable change in F/B ratio.)

Scaling these values to the new frequencies involves the two steps covered above. The first is to scale all dimensions by the ratio, F_{des} / F_{new} . Because this changes the conductor size, the performance must be recalculated with the

selected conductor—typically #12 for 40 meters, #14 for 20 meters, and #18 for 15/10 meters, assuming copperweld is used. However, a larger conductor may give a small increase in gain due to lower loss, but #18 has been found satisfactory for 20/15/10 meters.

If the design goal requires the best possible F/B at the new design frequency, change in reflector size will probably be necessary and is easily done with one of the analysis programs in a few tries. The step is not necessary for gain. If omitted, the frequency for best F/B will be off a small amount.

Table 3 gives dimensions and major performance parameters for the four bands selected, and for the four design goals. Antennas built to these values will have good performance at the design frequency, with the caveat that physical factors, such as conductor supports, mast, etc. may move the frequency of best F/B. A later article will cover methods of measuring the changes and compensating for them. Part 3 will be devoted to multi-element design.

Concluding remarks

The two-element quad antenna is possibly the most “cost-effective” antenna design for amateur HF through VHF use. It is economical of materials: wire is much less expensive than aluminum tubing. The antenna will have good gain, excellent bandwidth, and can have

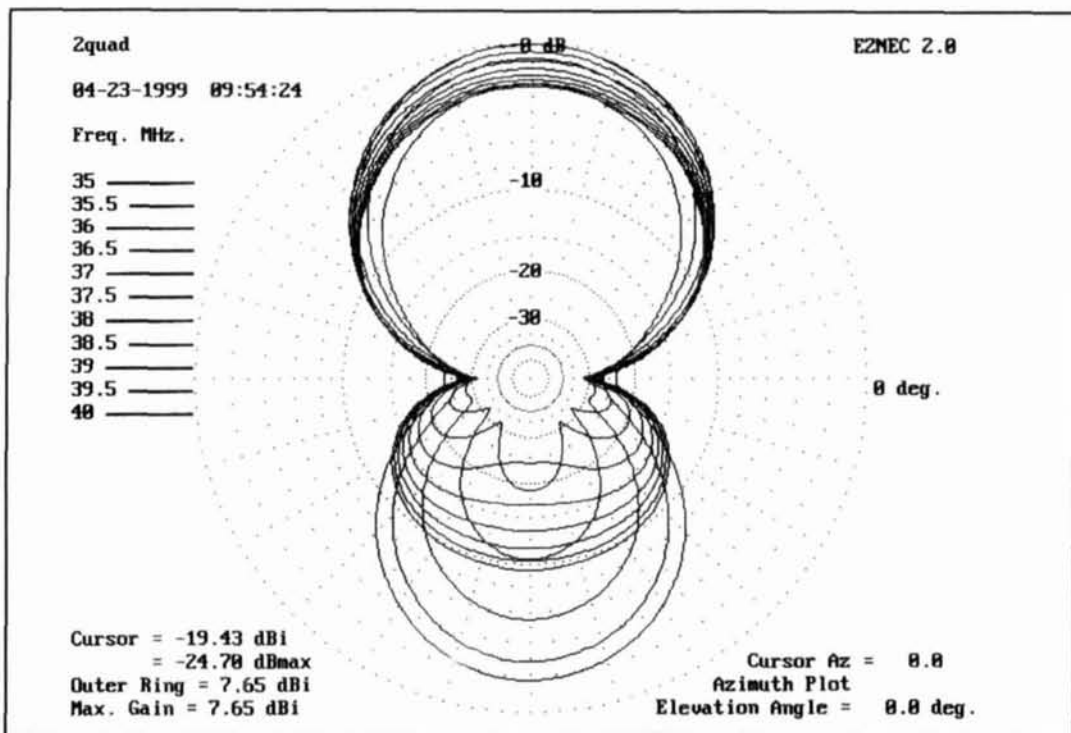


Figure 24. Patterns of small radiator scaled 2QUAD.

good F/B ratio. The turning radius is less than that of a Yagi of the same gain. In addition, the quad has better characteristics with respect to precipitation static due to rain, dust, and dry snow storms.

The usual objection to quads is based on reports of failure in wind and ice storms. These problems arise from inadequate mechanical design and can be prevented by using the tec-

niques of the original series of quad articles, or from Reference 2. ■

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LeCroy has published a four-color, 18-page application note that provides information on video standards and the use of analog oscilloscopes for video signal measurement. Discussed are the fast screen update capability, alias-free operation, and application-specific techniques for measuring video signals using analog scopes.

An introduction to video measurements found in the application note includes discussions of video basics, video triggers, pedestal clamping, delayed traces, and subcarrier horizontal phase adjustment. A section on display technology looks at analog scope displays and how analog, storage, and color oscilloscope technology can be combined.

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MICROWAVE SPECTRUM ANALYZER ON A BUDGET

*A practical approach to microwave
signal analysis for the home
experimenter*

Test equipment commonly found in the amateur workshop usually isn't suited for microwave experimentation. The most important piece of equipment for microwave work is a spectrum analyzer. And, while older commercial units are available, they have several shortcomings: they are costly, they may depend upon components that aren't readily

available, and repairs can be expensive. However, there is an alternative. *Ham Radio* magazine described a low-cost spectrum analyzer designed by Murray Barlow, WA2PZO, and built by Robert Richardson, W4UCH,¹ that offered effective spectrum analysis at reasonable cost. The design used cable TV tuners, a 66-MHz IF strip, and sweep circuitry to inter-

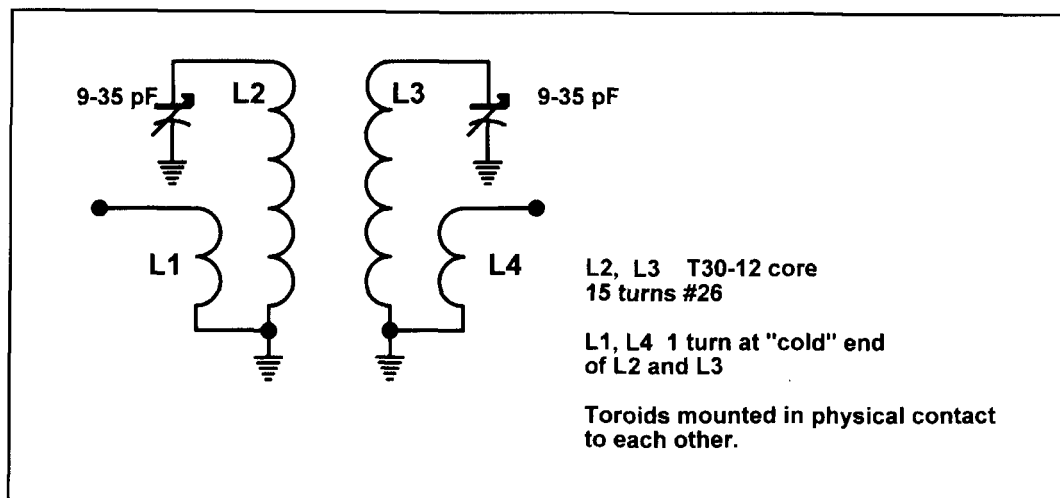


Figure 1. Double-tuned circuit.

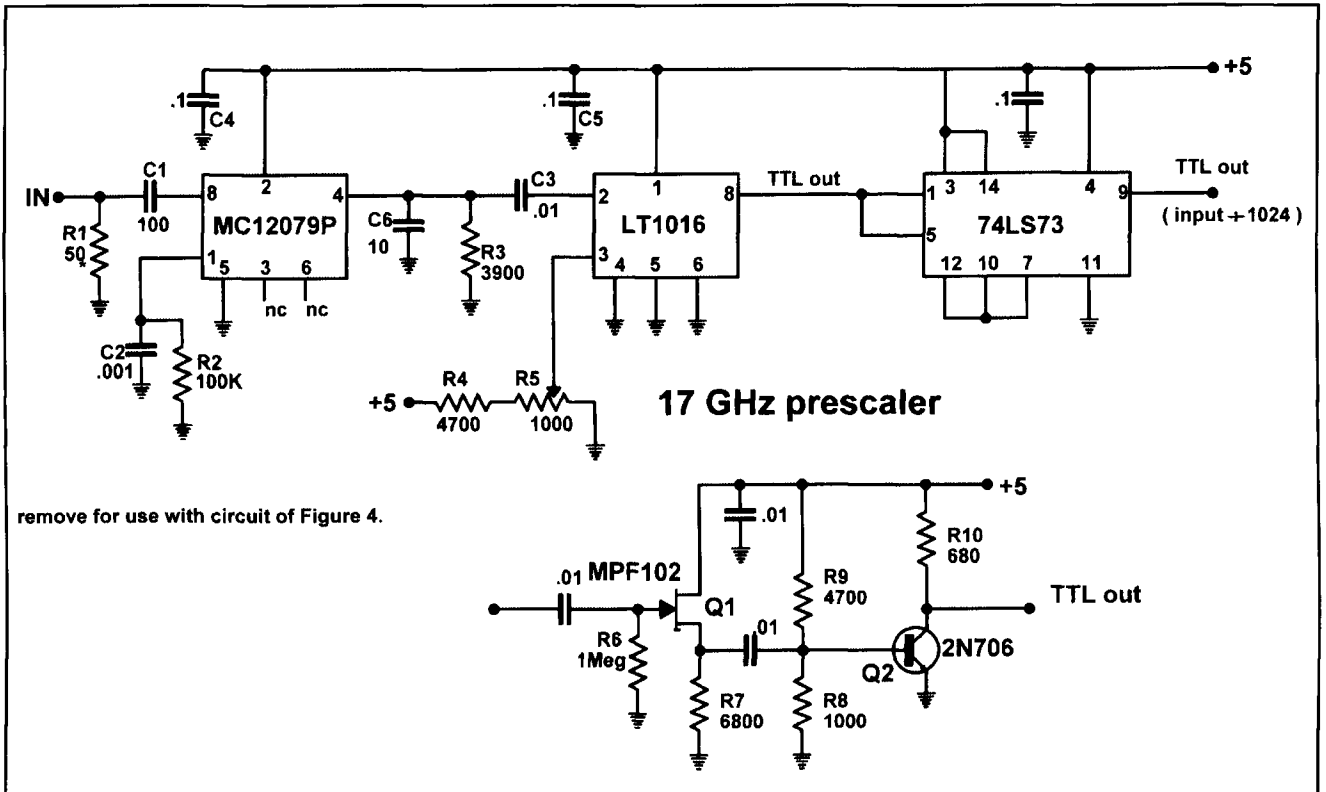


Figure 2. Prescaler built around a MC12079.

face with an oscilloscope. An option permitted coverage of the 1 to 1.75-GHz range. I'll show how to extend that range to higher frequencies.

The schematic of the IF strip of WA2PZO's analyzer shows that the 66-MHz IF stage uses a balun input. I added a double-tuned circuit

(Figure 1) to provide enough selectivity to attenuate image responses. The filter is built "surface-mount" fashion on a small section of printed circuit board, with the two toroids mounted in physical proximity. The board is mounted in the enclosure housing the IF strip.

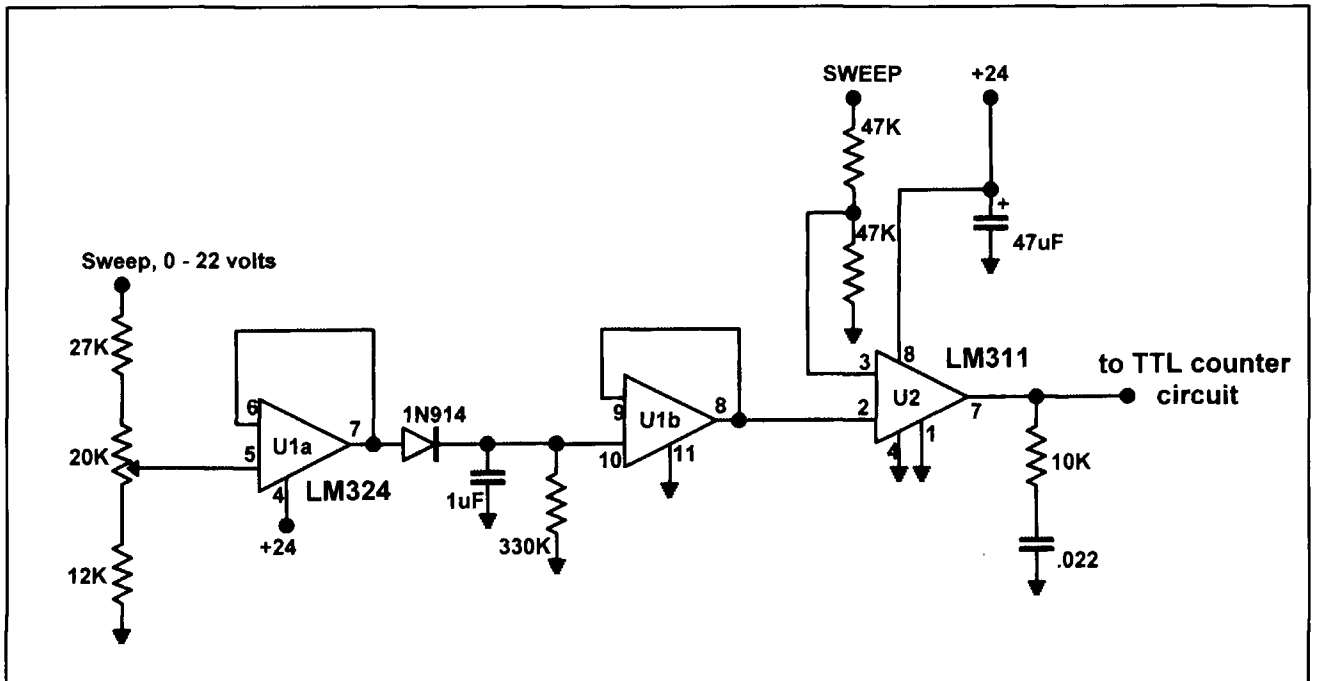


Figure 3. Peak detector.

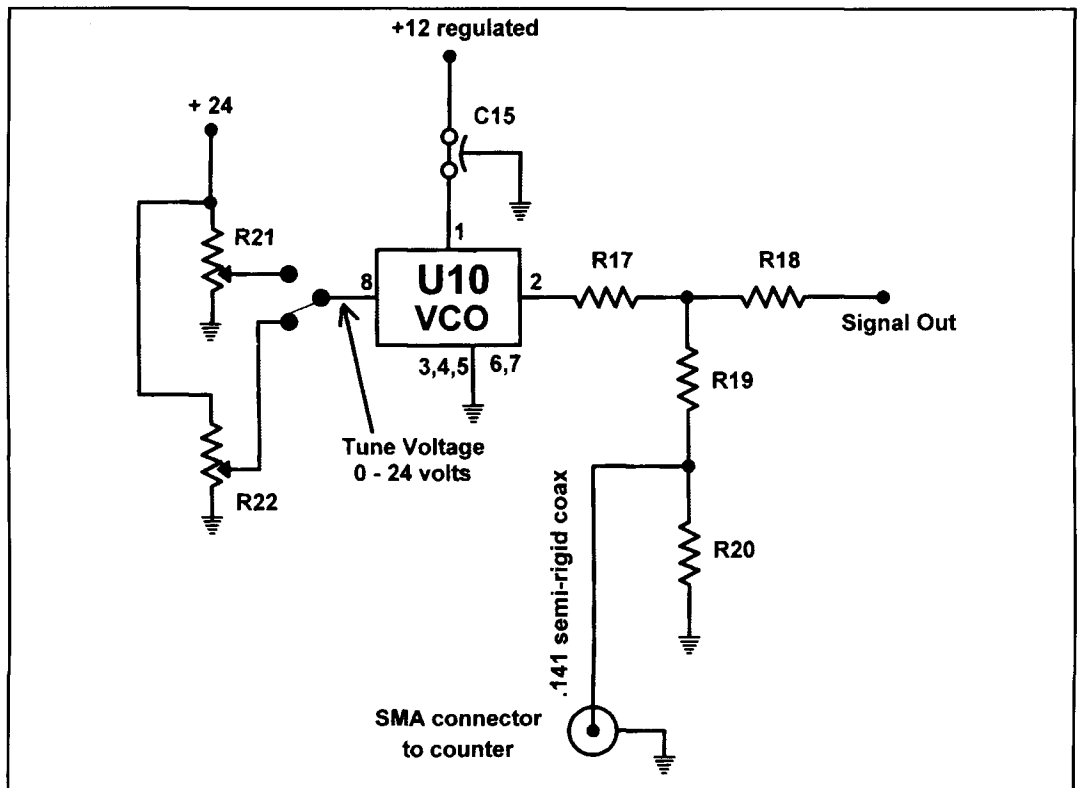


Figure 4. Simple generator capable of 940-2300 MHz.

Once the board is placed in the enclosure, images can be tuned out.

The next problem is to determine the frequency to which the analyzer is tuned. The tuner provides a sample port (RCA jack) to allow the use of a prescaler. A number of prescaler chips are available for operation to 3 GHz or higher. **Figure 2** shows a circuit built around the MC12079. This device contains a very sensitive on-chip preamp, and divides by 256 in this application. An additional divide-by-four circuit yields a final divide ratio of 1024; this can be fed to the counter circuit found in **Reference 2**.

The prescaler and counter

The prescaler was assembled on 0.06-inch G10 epoxy circuit board. The chip comes in an 8-pin DIP package. The input trace must be kept very short. I used an SMA receptacle on the enclosure, made of 0.039-inch circuit board. This was attached to the tuner using a 1-1/2 inch length of RG-174 coax with an RCA fitting. This may be a bit unorthodox, but it works. Note that pin 1 of the MC12079 should be biased through a 100-k resistor to either V_{cc} or ground to prevent oscillation.

At first the tuner was unable to drive the prescaler at frequencies near 1600 MHz. I traced the problem to the 100-ohm carbon film

resistor used to sample the RF. It had long leads, one of which was connected to ground and the other to the output jack. Replacing it with a carbon composition resistor solved the problem.

Once the prescaler and counter were operating, I had to determine where the readout was sampling along the frequency continuum. **References 2** and **3** assume a center frequency can be determined by calibration. I wasn't able to get that approach to work and devised the circuit shown in **Figure 3**. This is a peak detector that permits the comparator, U2, to start the frequency sample at a point which can be adjusted up to the approximate apex of the sweep waveform. When adjusted, a signal appearing on screen at the far right is reliably indicated by the readout—regardless of the settings of the other sweep controls.

The counter is designed to read out the LO frequency. By using preloading counter chips, as shown in **Reference 3**, the actual signal frequency can be displayed. This will work for the 0.9 to 1.75-GHz tuner. Otherwise, it is necessary to subtract 66 from the readout to determine the frequency.

A simple signal generator

There's another approach which is desirable for units using this tuner as a fixed IF. By con-

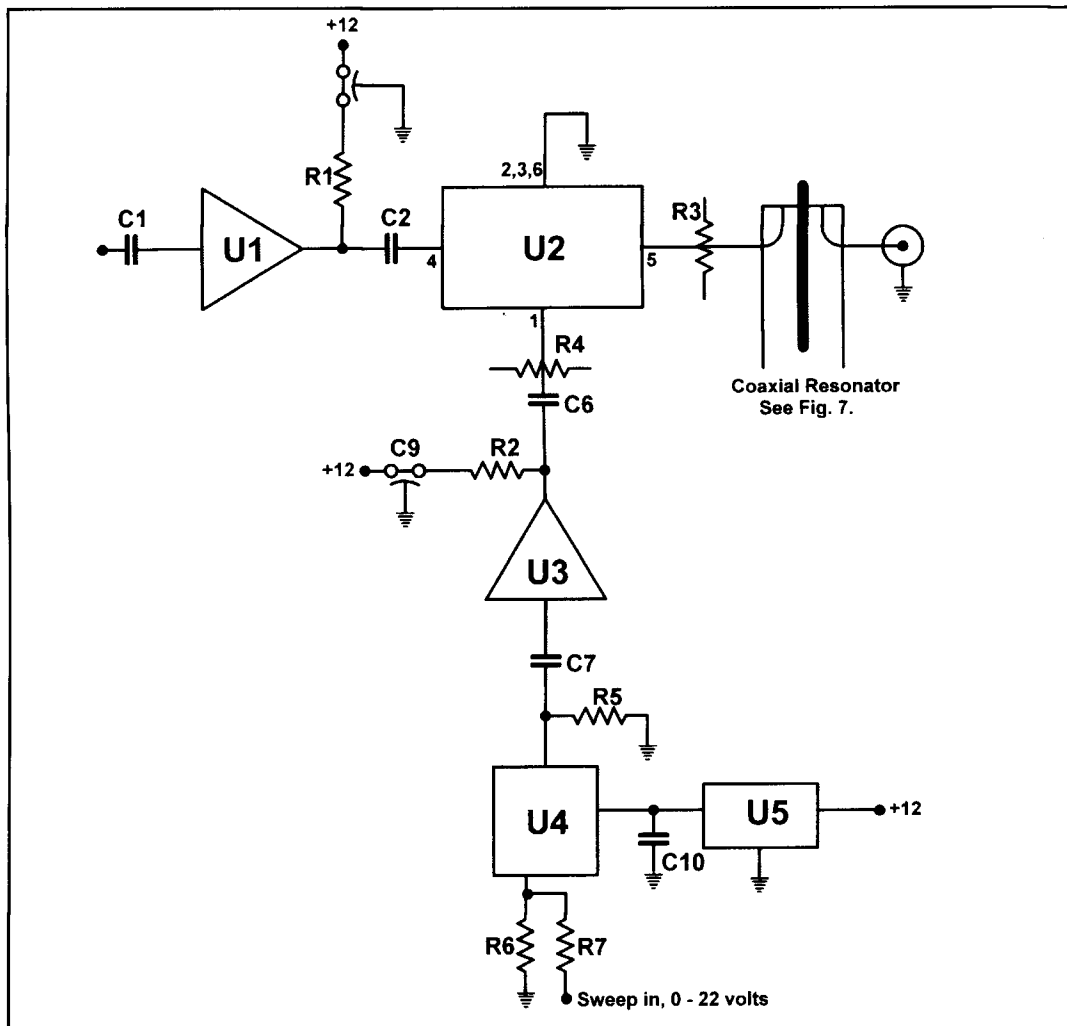


Figure 5. Method for reaching higher frequencies.

necting a signal generator to the counter, with low level injection to the converter input, you can read out the frequency at any point on the screen with accuracy. I constructed a simple generator using the MiniCircuits POS 2120W VCO* that tunes from 940 to 2300 MHz, as detailed in **Figure 4**. When placed in the vicinity of the analyzer, it produces a low-level trace on the scope display. At this point, it's only necessary to move the generator blip to coincide with the unknown signal blip to readout the frequency.

I used the sweep board from Science Workshop.** I found it helpful to look at the sweep output on a scope while varying the controls to better understand what I was seeing on the screen. Many settings of the rate, width, and tuning yield a sweep of constant amplitude and a trace with no signals. It's possible to adjust the controls so the entire sweep range is available and the entire tuner range is displayed.

Modification of the sweep circuitry to incorporate Joe Carr's DAC⁴ would also be beneficial, but I won't discuss that here.

I wasn't able to obtain cable tuners above 1800 MHz. It would be nice to display the entire range of 100 to 1000 MHz while aligning multiplier strips, and it would also be desirable to raise the upper frequency limit as well. However, it is possible to extend the range of the analyzer at reasonable cost.

Achieving higher frequencies

Figure 5 demonstrates how to achieve higher frequencies. First, you need to find a swept frequency source at a reasonable price. Backward-wave oscillators were used in older commercial oscillators and YIG oscillators are used in more modern equipment. Both offer wide bandwidth, but for a price! Microwave VCOs offer a bandwidth of 1 GHz or more when tuned with a voltage from 0 to 20 volts. Some of these devices are reasonably priced, have output lev-

*MiniCircuits, P.O. Box 7128, Branson, Missouri 65615. Phone: (417)335-5945.

**Science Workshop, Box 310, Bethpage, New York 11714.

els of 7 to 10 dBm, and are available with ranges to 4 GHz and higher. Wireless Radio's SM2-1750 VCO*** has a range that spans 1300 to 2400 MHz for a 0 to 22-volt input. With an IF at 1200 MHz, you would be able to get the analyzer to display from either 100 to 1100 MHz, or from 2500 to 3600 MHz.

A MiniCircuits RMS 30 DBM is surface mounted on 0.032 Duroid material, as shown in **Figure 5**. A 1200-MHz coaxial filter allows only the 1200-MHz IF signal to pass. MiniCircuits advises you to use a 3-dB pad at the output if you don't use a diplexer. Another 3-dB pad at the LO port reduces LO drive to 7 dBm while minimizing variations from the nominal 50-ohm termination impedance. A MCL MAR-6 device provides about 10 or 12 dB gain at the RF port.

Printed circuit material is placed around the sides of the mixer assembly. The VCO is housed in a separate enclosure and is connected to mixer via a short piece of 0.08-inch solid coax. I used a Wireless Radio surface-mount VCO based on the wide tuning range and rea-

sonable cost. It's not buffered, so I use an ERA2 to raise the output level and provide isolation between the output transistor and mixer. A 51-ohm chip resistor to ground, with a 4.7-pF chip cap to the ERA2 enables the VCO to provide a fairly constant output level. These VCOs are sensitive to impedance variations in the load as the frequency changes. The ERA2 impedance varies with frequency, but is probably better than the DBM in that regard; it's also needed to raise the output level.

The VCO is mounted on 0.032-inch Duroid material. A number of wires were passed through small holes drilled around the VCO module and soldered on both sides. The VCOs are surface-mount devices, and it's difficult to solder them using an ordinary soldering iron. Inadequate contact with the ground plane will adversely affect performance. The high-frequency model (**Figure 5**) would not work properly when mounted in this fashion. I then cut a square hole in the Duroid which fit the VCO exactly. Use of conducting traces to meet the tuning, power, and output ports, and using thin strips of hobby brass material soldered across

***Wireless Radio, P.O. Box 452, Dexter, Michigan 48130.

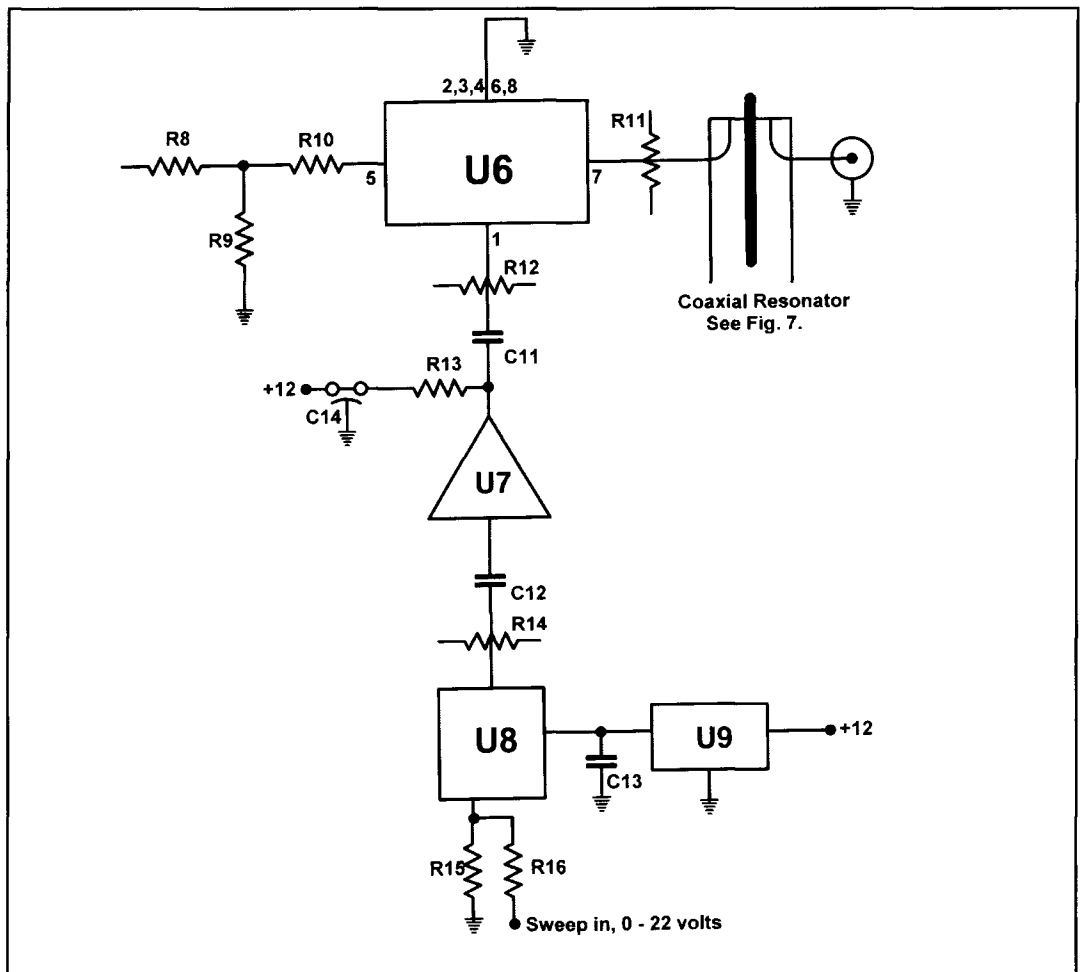


Figure 6. Converter using Wireless Radio model SM3-3000 VCO.

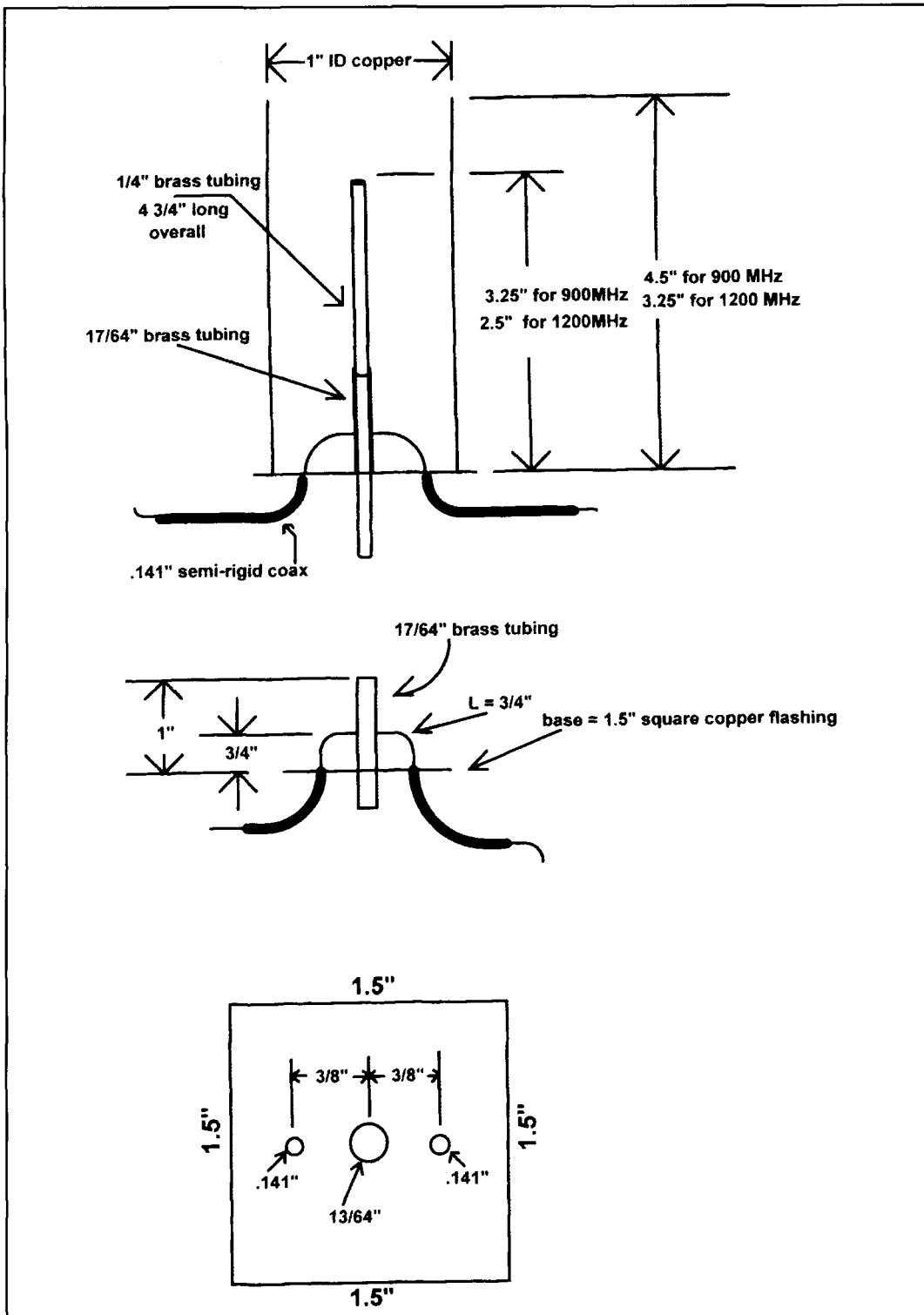


Figure 7. Coaxial filters.

the bottom of the VCO from the ground plane, improved the performance.

A second converter

A second converter (**Figure 6**) uses a Wireless Radio model SM3-3000 VCO cover-

ing from 2650 to 3700 MHz. A MiniCircuits SKY 5G mixer was needed because of the higher frequency ranges. A 900-MHz IF yielded a display range of 1750 to 2800 MHz. The image frequency is 3650 to 4600 MHz. These units are usable "as is" if you know the range of frequencies you're dealing with. If neces-

sary, a highpass or lowpass filter may be used ahead of the converter.

The interface of the VCO to the ERA2 buffer presented some problems. A MiniCircuits PAT3 pad seemed to work best, with a capacitor of 6.8 pF. A bit of experimentation is needed to achieve a constant output over the frequency range. This circuit worked for me, but injection is inadequate over a small range of frequencies. Although this hasn't been a problem, more work may be needed. The VCO should be mounted close to the DBM to keep the interconnecting trace short.

The coaxial filters (Figure 7) are made from 1-inch ID copper water hammer arresters. They are cheap at hardware stores, as are the hobby brass tubes used in the rest of the filter. The center tube extends 1/4 wavelength into the assembly. I got this idea from Reference 5; Reference 6 describes the theory.

The converter outputs are connected to the microwave tuner supplied by Science Workshop. The sweep is removed from the tuning port. For a 900-MHz IF, the tuning port is tied to ground; for a 1200-MHz IF, a positive 5.9 volts is applied. The correct tuner frequency may be verified by checking the LO frequency, which should be at 1266 MHz.

Calibrating the frequency display

The easiest way to calibrate the frequency display is to set a signal generator to display a

low-level signal at the frequency of interest, and sample the generator frequency with a counter. Figure 4 details a simple generator circuit using a new MiniCircuits VCO that covers from 940 to 2400 MHz. Operating the generator in the vicinity of the analyzer will provide a low-level signal on the display. The counter circuit described earlier provides the frequency readout. Additional higher-frequency VCOs may be acquired to extend the range.

These circuits work well for the alignment of multiplier strips, the detection of unwanted RF energy, etc. While lacking the features of commercial units, they do work well and are inexpensive. Recent advancements in microwave VCOs and lower-cost mixers make this a very sensible and very affordable project. ■

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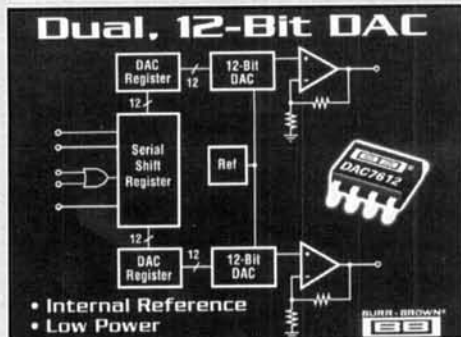
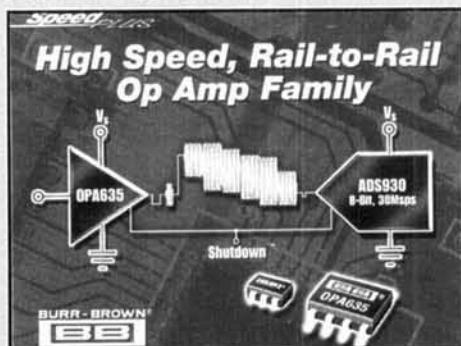
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TECHNICAL CONVERSATIONS

(from page 5)

nal object. But even in such a photograph, if the memory chip were photographed at a magnification providing an image with spatial frequency at the 1/2-power point of the film, the fine structure of the film image would be sufficient to produce aliasing when scanned with a contemporary scanner as reviewed in the article.

The intent of the article was to demonstrate how the scanner, and CCD, spatial quantization affect the results when the object scanned is of sufficient spatial frequency to challenge the limitations of the scanning system. At this point in the technology, I believe that film does indeed challenge contemporary scanning technology. I combined the concept of the square test target and the specified 1/2-power-point spatial resolution of the film to develop a type of maximum requirement for scanning resolution based only on the limitations of the film (or photographic paper) and totally independent of the actual object photographed. I believe this a reasonable estimate of the scanning resolution needed to preserve virtually all the analog intensity information in contemporary film images regardless of the subject.

Similarly, I believe that I addressed briefly the film grain structure. The typical grain size and spacing are well beyond the resolving ability of even the highest-resolution scanners, by perhaps as much as two orders of magnitude. Film resolving ability is limited by other parameters, such as scattering and dispersion of light in the emulsion. Therefore, the actual film or paper grain structure has a negligible effect on the scanned result.

As you point out, and as did I, I believe, most things we may photograph are comparatively smooth analog things (logs, dogs, cats, rats, etc.). But what if we choose to photograph a memory chip, or a high-contrast brick building at a distance? In those cases the objects are not smooth analog subjects, but indeed are very digital in nature. And the aliasing I demonstrated with the square test targets can be reproduced with these real photographic objects. Again, the focus of the article was from the perspective of the limitations of the optical materials and not the subjects, or how the subjects may be filtered to accommodate the scanning technology. The focus of the article was what scanning resolution is needed to preserve the full resolving capability of the analog photographic media independent of the subject being photographed.

Also, I believe that I did comment very briefly on the enhanced printer technologies, such as the HP Photo REt and the improved

Photo REt II technology, but specifically limited my discussion to the simple ink-jet “dot matrix” half-tone systems. These are by far the most common. Also, it is much easier to visualize the dot-matrix half-tone process than something such as the Photo REt process. Once the basic process is visualized, one can more easily comprehend the more complex processes.

Also, there is considerable debate as to whether the modulated dot technologies, such as the HP Photo REt II, are visually equal to the micro-dot technology, such as used by Epson and Canon. Personally, I do not like the HP output. To me, it has a very obvious graininess caused by the fewer dots per unit length, regardless of their precision size modulation and spatial dithering. A very obvious example of this is the “banding” seen in images with very slow changing density across the image of comparatively bright subjects, such as clouds for example, and a mottling of continuous-tone features such as skin tones. It is true that over the entire image, the dot modulation does reproduce the average intensity, but the graininess is obvious, at least to me.

The Epson and Canon 1440 DPI systems, and even the 720 DPI systems for most images, also faithfully reproduce the intensity information, but with substantially less grain (assuming the system is working properly, which is not always the case). For example, I see virtually no mottling with these systems, whereas I do with the HP. So, based on my personal preference, I cannot agree with you that the modulated-dot print technologies are equal to the micro-dot technologies. But that is only my subjective personal preference. I use both systems in my work, but I use the micro-dot technology for my personal digital photos. I chose not to make these comparisons in the article both in the interest of space, and because I felt it inappropriate to criticize a particular manufacturer’s technology.

But remember, the purpose of my article was to explore what is needed to capture all the information in an analog photographic subject—not what is simply “good enough” if viewed at a great enough distance. In other words, how good must be the scanning and printing technology be to produce a digital print which cannot be distinguished from a true analog photograph, even when the subject (not the medium) is viewed at high magnification?

Finally, the issue of interpolation is quite controversial. The bottom line is that interpolation adds information that is not there. One cannot take the Fourier transform of the image and then “replace” or “improve” the sidebands synthetically to improve the image to make it a more faithful representation of the object unless one knows what the side bands must be. That is

a function of the object. But that object information was not recorded, and in general we have no way of recapturing that information. If we could, then we would have a higher-resolution image anyway.

One may guess at some object characteristics; for example, that the object was a generally a smooth analog subject, such as logs, dogs, cats, rats, etc., and synthetically interpolate based on some corresponding estimating function (i.e., a specific reconstruction filter), but that is just a guess. Some a-priori knowledge of the object is needed. What if the object were a memory chip? Since the memory chip is for most practical purposes a true spatially digital object, if we were to interpolate the digitized image to increase the apparent scanning resolution based on a reconstruction filter based on a guess of a smooth subject, the result would be nothing like the original object.

But this does not mean that interpolation is not useful. For example in audio CD players, such interpolation, or over sampling as it is termed, allows much simpler and less expensive filters to be utilized to reconstruct the audio signal. This over sampling, or audio interpolation, provides no improvement to the faithfulness of the reproduced representation of the original audio "image." It simply simplifies (and reduces the cost) of reconstructing the audio image. The quality of the audio image is fundamentally limited by the digitizing precision (number of bits), not the amount of over sampling in reproduction. The over sampling simply makes the hardware easier to build, more stable and robust, and lower in cost, all without degrading the audio quality limited by the digitizing depth.

Nevertheless, here too I cannot agree with you that interpolation of a scanned image, "when properly done," will produce a more faithful copy. I argue it will not. It will produce a smoother image, not a more faithful image—is a smoothed image of a memory chip a more faithful representation than an aliased one? But of course one will look better, regardless of whether it is a faithful representation. That is what the article was about: what do you need to totally faithfully reproduce the analog characteristics of the scanned subject—a film image, for example?

What I think all this means is that each individual must research the various technologies available in the digital photography art and use that most appropriate to their needs (photos of dogs or computer chips). My intent in the article was to provoke the reader to think about the process. I seem to have done that. But of course I welcome any additional comments or suggestions you may have. Perhaps we could actually "do some science" together to explore the limi-

tations of the contemporary scanning art and analog limitations of photographic media, such as film and CCDs?

Mike Gruchalla
Editorial Review Board
Albuquerque, New Mexico

The objective value of interpolation

Dear Mike:

Thanks for the reply. As with the original article, the reply was enlightening, and I appreciate hearing from you. I think we are pretty much in agreement, and you obviously have considerably more direct experience with this. My attempt was to illuminate a couple of areas that other readers might have missed which you had touched on very briefly, if at all.

I still believe that some interpolation has objective value, in the manner you indicated with audio CDs. Digital video systems can sometimes show artifacts of the digitization and presentation processes, which are not in the original, like a 16-bit D/A on a CD that is not followed by a low-pass filter. The over-sampled lower bit converter, as you pointed out, takes less filtering to accurately reconstruct. If a video system lacks the filtering, then interpolation is a useful way of improving the situation. In neither case will the interpolation restore artifacts caused by aliasing because of lack of input filtering, but it can render a more accurate version of what actually is available, especially on an output device lacking in smoothing capacity.

An interesting variation of this is oversampling on the digitization process, where the dot size is larger than the DPI. Obviously some redundant information is being collected, but with proper processing, the results can exceed that predicted by the dot size.

Again, thanks for your reply. I got something I hadn't anticipated—additional insights into the process. I am at present looking for both a scanner and an ink jet printer for a variety of tasks including photographic.

Wilton Helm, WT6C
Franktown, Colorado

The value of interpolation

Dear Wilton:

You are absolutely right that interpolation can have real value—in making prints to be viewed at reasonable distance. By interpolating, the "pixelization" is reduced. No new data is added, but sharp pixel edges are fuzzied up a

bit to allow the pixel information to run together. A somewhat fuzzy picture (photographers call that “soft focus”) is much more pleasing to view than a sharply pixelized image.

However, I believe that one should apply interpolation only when making a print to be viewed. It should not be used for archiving information. The data for archiving should be as accurate and unmolested as the digitizing and recording technology allows. You can always fiddle with the raw data in the archive later as new tools become available. However, if you fiddle with the raw data and then archive it, you probably cannot un-fiddle it later. Hence the thrust of my article—just how good does scanning (and CCD) technology need to be to provide raw, “un-fiddled with” data that is “just as good as” the real analog source?

As for scanners and printers, I’ll give you a couple of ideas. I looked at a bunch of scanners. I decided that I wanted at least a 600 x 600 unit and preferably 1200 x 1200. By the way, make sure the 600 x 600 or whatever is the true “optical” resolution, not “effective” (read interpolated) resolution. If you look at CompUSA and BestBuy, some times you can find a lost-leader sale where the scanner is actually free. A while back, CompUSA had a 300x300 parallel-port unit for sale for less than the rebate. Wasn’t much of a scanner, but certainly worth the price. And something like that is a good place to start if you have no idea what you need (or want). But these “low-end” units actually work surprisingly well. I ended up getting an HP6270A with the document feeder (didn’t need the feeder, but the unit was on sale for a good price). This is a 1200 x 1200 unit and it was lower in cost (much) than the other 1200 x 1200 units at the time. As scanners go, this one is a bit pricey, about \$300, but that is in line with the high-res units. Works great!

As for printers, good luck! I must have looked at every printer made, twice. Some thoughts. I presume you want color. Color lasers are out! They cost a lot, are expensive to run, and produce poor results. Also, stay away from the weird technologies—dye sublimation, wax transfer, etc. Some of these work quiet well, but they cost quite a bit to operate and the prints may not be very durable. That

pretty much leaves the ink jet printers. As I mentioned in my previous note, I just do not like the HP images (personal preference). The Epson, Canon, Lexmark, etc. models with the 7-ink system all seem to be about equal as far as resolution. It is pretty easy to get 720 x 720 dpi now and may be even 1440 x 1440. The 720 x 720 seems to be just about all you need, but there is still just a bit of pixelization. Also there are compromise units with 1440 x 720. These are okay too. The trick is finding something that will actually keep working and does not consume an entire set of cartridges on a single print.

In general the printer guys should give the printers away—the consumables eat you alive. Some of the printers have the print head in the machine (Epson) and some in the ink cartridge (HP and Canon). I prefer the head-in-cartridge configuration; it’s easy to “replace” the head when (there is no “if” here) it gets clogged. I ended up getting a Canon BJC-5000 1440 x 720 unit (this is now obsolete, replaced by the BJC-5100 I think—the same basic machine). This is the only machine that I could go anywhere in town and find it working where I could print a demo. Virtually all other machines were screwed up in one way or another. I have had virtually no trouble with it at all, I even let it sit for about three months and then it printed perfect pix. I believe that the newer unit (5100 as I recall) is about \$150 more or less; that’s about the ballpark of all the units. And this unit will do 11 x 17 prints to boot—great for B-size drawings.

You need to go to every place near you that sells printers and do some demos yourself. If possible, take along your own image file to print; everyone here was quite happy to let me use my own demo picture. Then you can compare the resolution and color (accuracy and saturation) of all the different machines (at least the ones you can find that are working!) To do this comparison, you probably should purchase some of the high-resolution paper (not glossy, just high res). Good luck!

Mike Gruchalla
Editorial Review Board
Albuquerque, New Mexico

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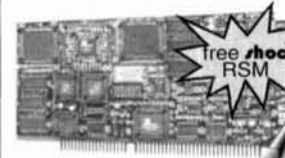
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Correction

Dear Editor:

Here is a correction and addendum for the "The Quad Antenna Revisited: Part 1" from the Summer 1999 issue.

First, in **Figure 4**, the shape factor equation should read

$$SF=2*B/S$$

where B is the perimeter.

Second, the drawings in **Figure 37** and **Figure 38** are interchanged.

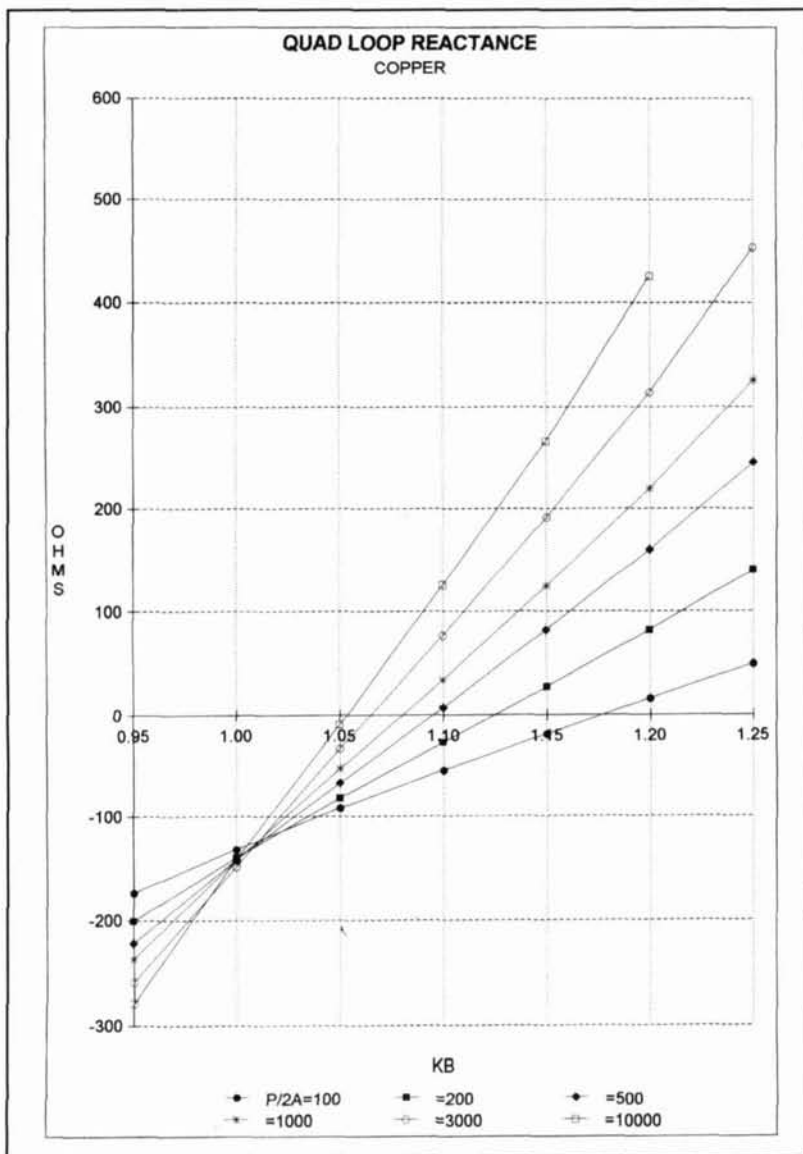
Third, the enclosed **Figure 13** replaces the original.

Finally, an addendum. For those who prefer to work with quad dimensions in feet rather than wavelengths, the resonant frequency equation on page 59 becomes:

$$B/f=1437.6-185.6*X+20.7*X^2 \text{ feet/MHz}$$

where X is the log of the wire thickness factor. For TF=10000, about #12 wire in 20 meter loop, B/f is 1026. For accuracy, the thickness factor must be considered. This is especially important where good F/B ratio is needed.

R.P. Haviland, W4MB
Daytona Beach, Florida



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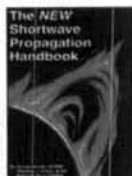
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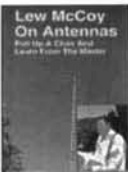
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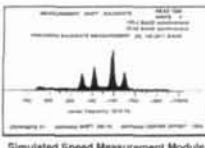
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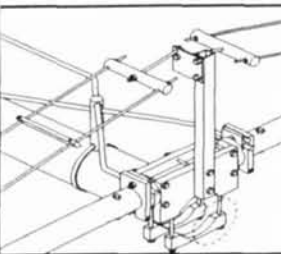
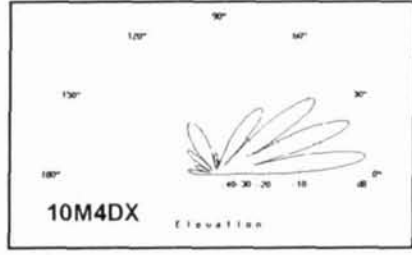
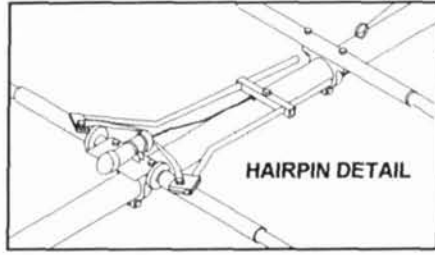
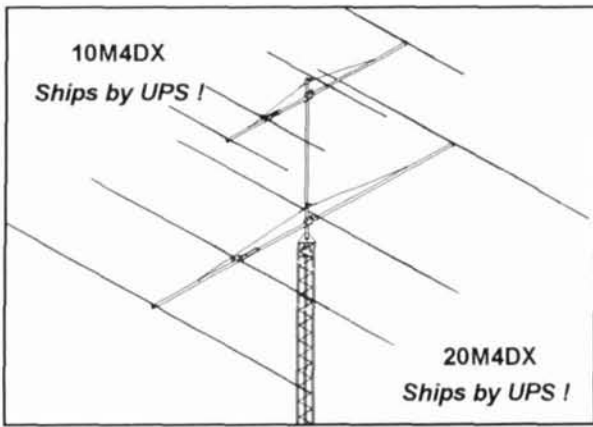
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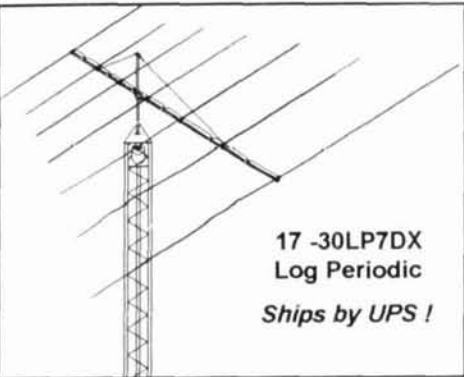
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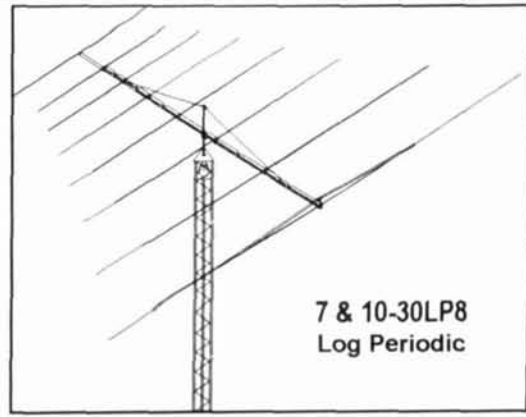
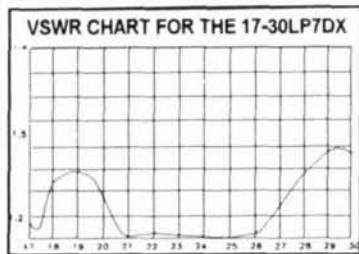
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20M4DX	34' 8"	7.4 dBd	38 lbs. / UPS
17M3DX	19' 2"	6.3 dBd	30 lbs. / UPS
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MODEL	BOOM LENGTH	GAIN IN FREE SPACE	WT. / SHIP
6-10LP5	30'	5 dBd @ 6-10	86 lbs. / TRK
7 & 10-30LP8	30'	6 dBd @ 10-30 / 3 dBd @ 7	86 lbs. / TRK
10-30LP8	29'	6 dBd @ 10-30	75 lbs. / TRK
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SS-10	7	10	1 1/2 x 6 x 9	3.2
SS-12	10	12	1 1/2 x 6 x 9	3.4
SS-18	15	18	1 1/2 x 6 x 9	3.6
SS-25	20	25	2 1/2 x 7 x 9 1/2	4.2
SS-30	25	30	3 1/2 x 7 x 9 1/2	5.0

DESKTOP SWITCHING POWER SUPPLIES WITH VOLT AND AMP METERS

MODEL	CONT. (Amps)	ICS	SIZE (inches)	Wt.(lbs.)
SS-25M*	20	25	2 1/2 x 7 x 9 1/2	4.2
SS-30M*	25	30	3 1/2 x 7 x 9 1/2	5.0

RACKMOUNT SWITCHING POWER SUPPLIES

MODEL	CONT. (Amps)	ICS	SIZE (inches)	Wt.(lbs.)
SRM-25	20	25	3 1/2 x 19 x 9 1/2	6.5
SRM-30	25	30	3 1/2 x 19 x 9 1/2	7.0

WITH SEPARATE VOLT & AMP METERS

MODEL	CONT. (Amps)	ICS	SIZE (inches)	Wt.(lbs.)
SRM-25M	20	25	3 1/2 x 19 x 9 1/2	6.5
SRM-30M	25	30	3 1/2 x 19 x 9 1/2	7.0

2 ea SWITCHING POWER SUPPLIES ON ONE RACK PANEL

MODEL	CONT. (Amps)	ICS	SIZE (inches)	Wt.(lbs.)
SRM-25-2	20	25	3 1/2 x 19 x 9 1/2	10.5
SRM-30-2	25	30	3 1/2 x 19 x 9 1/2	11.0

WITH SEPARATE VOLT & AMP METERS

MODEL	CONT. (Amps)	ICS	SIZE (inches)	Wt.(lbs.)
SRM-25M-2	20	25	3 1/2 x 19 x 9 1/2	10.5
SRM-30M-2	25	30	3 1/2 x 19 x 9 1/2	11.0

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SS-18GX
SS-12EFJ
SS-18EFJ
SS-10-EFJ-98, SS-12-EFJ-98, SS-18-EFJ-98
SS-12MC
SS-10MG, SS-12MG
SS-101F, SS-121F
SS-10TK
SS-12TK OR SS-18TK
SS-10SM/GTX
SS-10SM/GTX, SS-12SM/GTX, SS-18SM/GTX
SS-10RA
SS-12RA
SS-18RA
SS-10SMU, SS-12SMU, SS-18SMU
SS-10V, SS-12V, SS-18V