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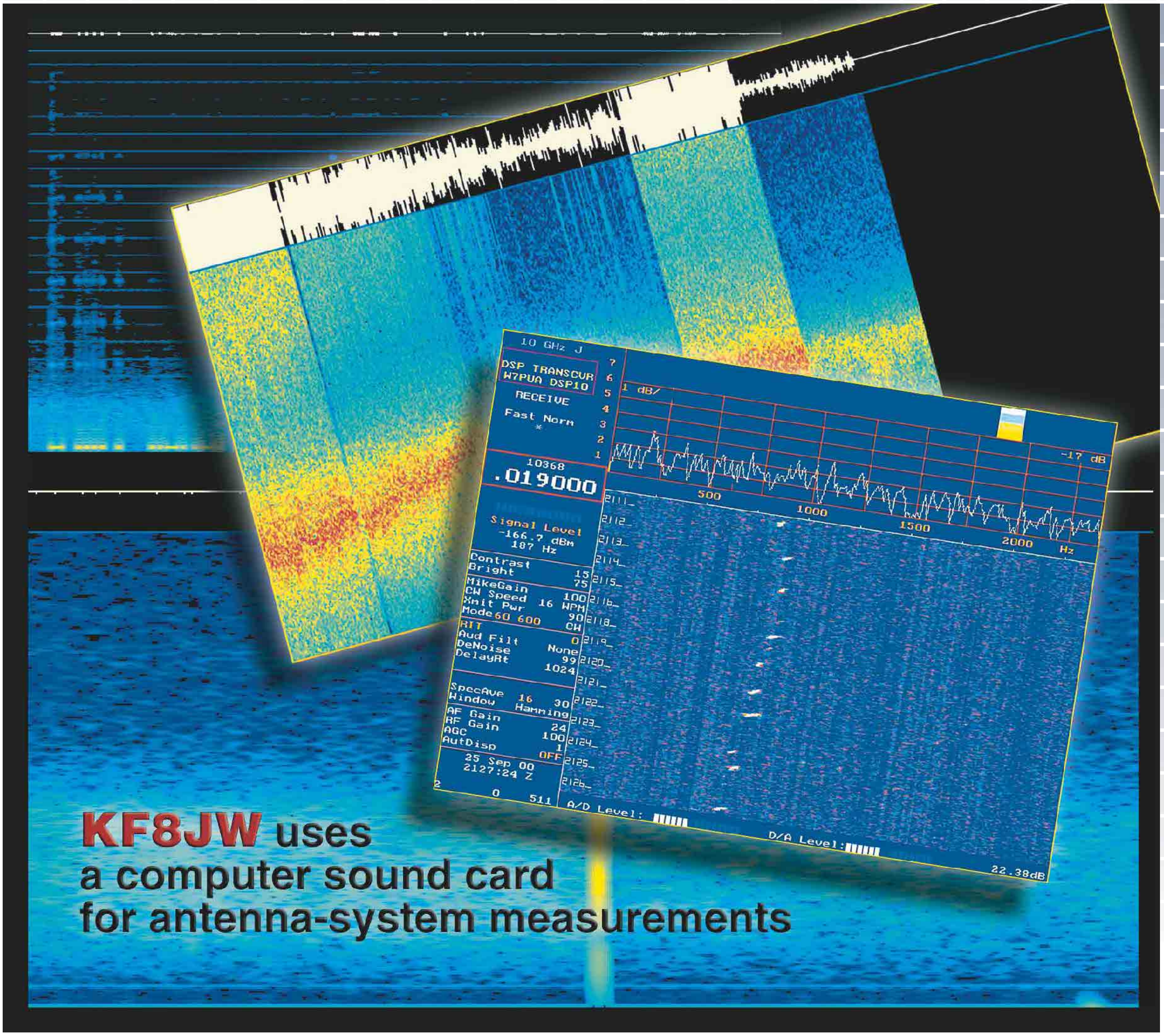
QEX

INCLUDING: COMMUNICATIONS QUARTERLY

Forum for Communications Experimenters

January/February 2002



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QEX

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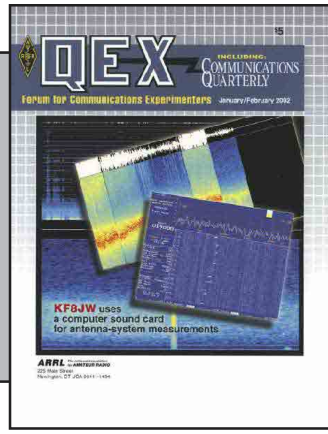

About the Cover

These are charts of Phil Sage's measurements, the article is on p 33.

Features

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THE AMERICAN RADIO RELAY LEAGUE

The American Radio Relay League, Inc, is a noncommercial association of radio amateurs, organized for the promotion of interests in Amateur Radio communication and experimentation, for the establishment of networks to provide communications in the event of disasters or other emergencies, for the advancement of radio art and of the public welfare, for the representation of the radio amateur in legislative matters, and for the maintenance of fraternalism and a high standard of conduct.

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1) provide a medium for the exchange of ideas and information among Amateur Radio experimenters,
2) document advanced technical work in the Amateur Radio field, and
3) support efforts to advance the state of the Amateur Radio art.

All correspondence concerning *QEX* should be addressed to the American Radio Relay League, 225 Main Street, Newington, CT 06111 USA. Envelopes containing manuscripts and letters for publication in *QEX* should be marked Editor, *QEX*.

Both theoretical and practical technical articles are welcomed. Manuscripts should be submitted on IBM or Mac format 3.5-inch diskette in word-processor format, if possible. We can redraw any figures as long as their content is clear. Photos should be glossy, color or black-and-white prints of at least the size they are to appear in *QEX*. Further information for authors can be found on the Web at **www.arrl.org/qex/** or by e-mail to **qex@arrl.org**.

Any opinions expressed in *QEX* are those of the authors, not necessarily those of the Editor or the League. While we strive to ensure all material is technically correct, authors are expected to defend their own assertions. Products mentioned are included for your information only; no endorsement is implied. Readers are cautioned to verify the availability of products before sending money to vendors.

Empirical Outlook

You Asked for It

In response to your feedback, we're making two significant changes. The first will help us produce a better product; the second focuses on getting the word out about *QEX* and on keeping our contributors happy.

Please welcome aboard Raymond Mack, WD5IFS, as Contributing Editor. Ray has graciously volunteered to help us proofread final drafts of published material for errors and omissions. He will also contribute items about new products and services available to experimenters. Ray has written for *QEX* before. First licensed in 1970, Ray's main interest lies in VHF and UHF weak-signal work. He also enjoys experimenting with DSP and digital communications. Having worked in the medical-electronics field for many years, he is now in MPEG software development at Conexant. Ray has also taught part-time in community colleges.

In this digital age, the Internet stands out for its freedom of expression and it even rivals television as an advertising medium. Web publishing often widens the readership of printed periodicals and books. We acknowledge its value and we recognize that references to *QEX* posted on the Web are sometimes the only ways potential readers know we exist.

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As we begin our third decade here at *QEX*, we would like to remind you of certain topics of which we are especially looking for coverage. Those include but are not limited to: digital voice, high-speed networking and Amateur Radio-Internet gateways, remote control systems, so-called software-defined radios, weak-signal and DSP techniques, spread-spectrum, microwave and satellite operations. One area definitely not getting enough coverage is that of antenna radiation-pattern measurement. If you don't see your particular fancy in that list, don't despair. In general, we like to receive any contribution that: (1) solves a problem, or (2) presents a new or better way of doing or explaining things or (3) describes something fun to build or use. We're not starving for articles; but if you or someone you know wants to share what they've learned, please drop me a line.

In This Issue

Andre Jamet, F9HX, finds that super-regenerative receivers are a good way to get going on microwave bands without a lot of complex circuits. Try it: You'll like it! Jack Hardcastle, G3JIR, has some observations about measuring and specifying crystal parameters. They go a long way toward showing how to get the most from your oscillator and filter designs.

Dan Handelsman, N2DT, returns with the second part of his look at rectangular antennas. In this segment, he concentrates on getting the feed-point impedance up while maintaining gain. R. P. Haviland, W4MB, concludes his series on quads. His Part 5 sheds some light on quad variations. Phil Sage, KF8JW, makes use of computer sound cards for comparative antenna measurements and other tests. He shows how digital sound recordings may save you a lot of legwork while you are optimizing the performance of your shack.

Bob Bruninga, WB4APR, describes the design of PCSat, recently launched into orbit from Kodiak, Alaska. He also presents an update on its performance and details of its continuing operation. In *Tech Notes*, Pete Bertini, W1ZJH brings us a piece from Rick Littlefield about a multiband vertical antenna. In *RF*, Zack Lau, W1VT, opens a discussion of 10-GHz DX techniques.—Happy New Year! *Doug Smith, KF6DX,* **kf6dx@arrl.org**

SHF Super-Regenerative Reception

"Give me somewhere to stand and I will move the Earth"—Archimedes

"Give me an oscillator and I'll make a receiver"—F9HX

By André Jamet, F9HX

Super-regenerative reception is not as old as ancient Greek mathematicians, but it is a very old method. It is already employed in professional applications, such as remote control at UHF—433 MHz for example—but it is no longer used extensively by radio amateurs.

In European publications, I have described several receivers for 144, 432 and 1296 MHz, as well for 10 GHz, using the super-regenerative method[1, 2, 3] I present here some further notes on SHF super-regenerative reception.

[1]Notes appear on page 6.

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A Tiny Little Bit of Theory

A lot of articles and books have already been published about super-regeneration theory over more than 70 years, and it is not my intention to revisit them here. I have only a few words to remind you of some of the principles.

A super-regenerative receiver is based on the repeated buildup and decay of oscillation in an oscillator, which is caused to operate intermittently by means of a quenching signal. That signal is supplied from a separate low-frequency oscillator or from a low-frequency relaxation of the oscillator itself. Those two modes are respectively called "separated quenching" and "self-quenching." After each quenching, during the starting and the buildup of oscillation, the device shows a successively positive-resistance behavior, then zero and finally a negative resistance. When the negative resistance is reached, the device is oscillating. Nevertheless, during the period of the exponential buildup of oscillation, it shows a tremendous amplification of up to one million!

When no external signal is present at the input, the amplification applies to the noise present there. You can hear the typical and well-known rushing sound. If a signal at the oscillation frequency is applied, the oscillation is started in advance as the received signal rises above the level of the noise alone. This advance gives an increase of the current in the oscillator, proportional to the signal, but highly amplified.

In the beginning, the electronic tube

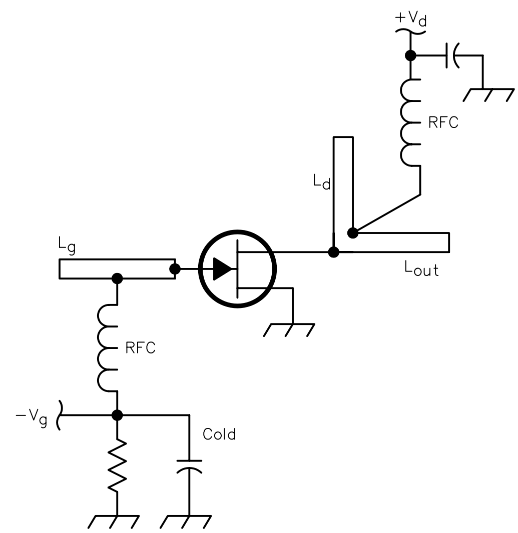

Fig 1—A DRO with gate-bias.

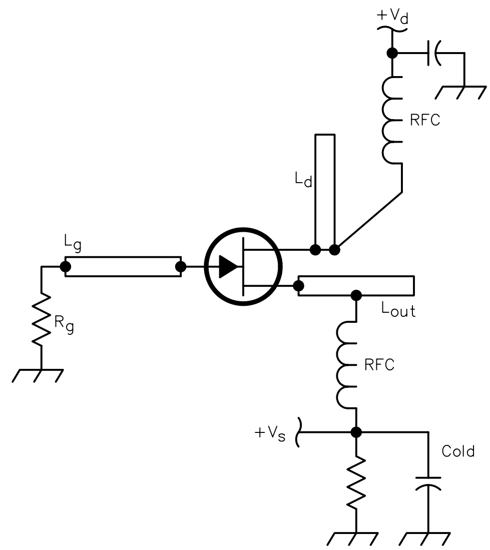

Fig 2— A DRO with source-bias.

was used as the oscillator device; but now, evidently it is the transistor. Nevertheless, other components are able to produce a negative resistance: unijunction transistors, tunnel diodes and so forth. Also—and it was at 10 GHz during the 1940s—a 723 A/B Klystron[4] showed a 150-µV sensitivity!

The final purpose of super-regens is the reception of modulated signals. CW demodulation is very easy as the transistor current increases when a signal is on, since the oscillation starting is advanced. For AM, that current varies as the carrier magnitude. SSB needs a re-established carrier as shown in Reference 1. FM demodulation is obtained by detuning the super-regen to use one slope of its selectivity curve. Owing to the relatively poor selectivity, poor results are usually obtained with NBFM. Nevertheless, at 10 GHz, it is a frequent practice to use a quite large deviation, with Gunn-diode transmitters, for example.

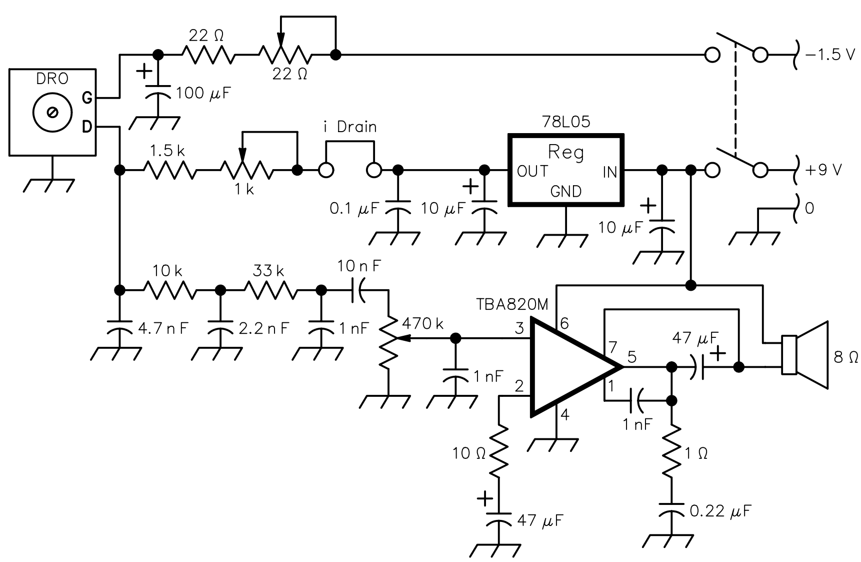

Fig 3—Schematic of a 10-GHz regenerative receiver.

A 10-GHz Device

The use of a dielectric-resonator oscillator (DRO) is a very simple way to make a 10-GHz oscillator. Adding an RC network to the drain circuit provides self-quenching by a low-frequency relaxation oscillation. The audio signal is picked up on the drain resistor. Gate-source voltage of the transistor controls the operating point to get the best sensitivity results. Several configurations are used for DROs as well as all kinds of oscillators.

The resonator may be placed gate-side, drain-side, or between gate and drain. Some experiments showed it is possible to get them working as super-regens fairly easily. It is important, though, to inject the gate-source voltage at an originally "cold" point (RF ground) to avoid DRO modification. Figs 1 and 2 show typical diagrams for two cases. A negative voltage will be injected gate-side or a positive voltage source-side. That voltage is brought through a 0.8-mm hole very close to the cold point.

An audio amplifier is required after the super-regen stage, with a low-pass filter to remove the quenching frequency, which could otherwise disturb the amplifier—even to the point where it is not useful. A regulated power supply is needed for the DRO, as its frequency varies appreciably with supply voltage.

Fig 3 shows the complete diagram. The DRO used requires a negative voltage gate-source so a 1.5-V battery has been added to the two 4.5-V batteries for the positive supply.

Construction

The easiest way to get a 10-GHz DRO is to remove it from a low-noise amplifier/mixer (LNA) used for the reception of satellite TV channels. Since the DRO is used as a local oscillator, its frequency varies according to the received frequency range and the intermediate frequency. Usual frequencies are 9.75, 10.6 and 10.475 GHz. As 3-cm amateur-band traffic is usually passed at about 10.368 GHz, we need to shift the dielectric-resonator frequency.

The best solution is to use a 9.75 GHz DRO. If the frequency shift achieved by the adjustment screw in the LNA is not enough, we must increase the frequency by decreasing the dielectric resonator's height. To do that, we can use sandpaper or, better, a mini-drill with a little millstone. Avoid inhaling the dust.

It is more difficult to shift DROs operating at higher frequencies. We can add a small piece of ceramic taken from another DR or from a capacitor. The quality factor and the temperature coefficient are upset by this modification, but it works.

The DRO is removed from the LNA by sawing out the case, screens and PC board. See Fig 4 and the photographs. That gives a small, shielded box with the original screw for frequency adjustment. After that, the DRO is placed against a waveguide (WR90/R100) in which a slot has been cut. That slot feeds a small piece of Teflon-insulated wire (a 5-mm coaxial cable with the braid removed) connected to the DRO output. The slot allows the DRO to be installed at the optimum place; slides with blocking screws or a

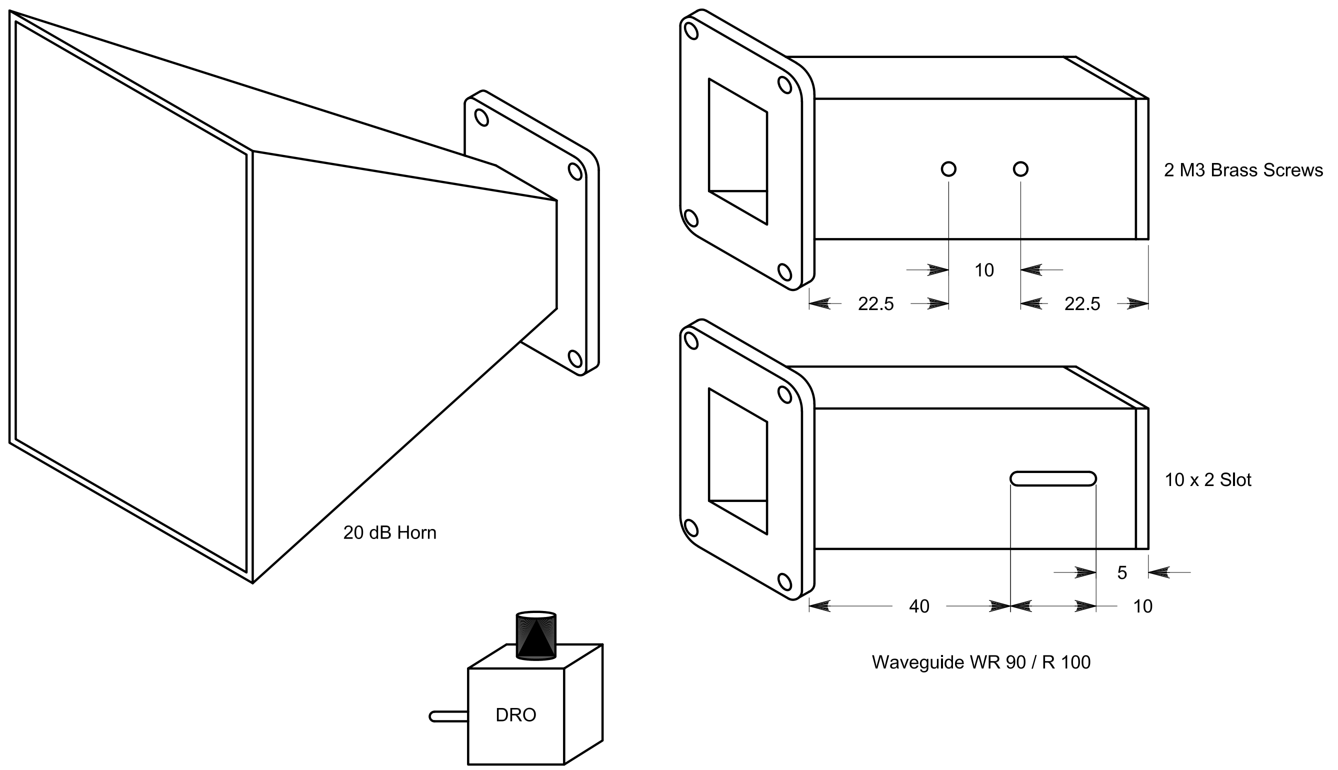

Fig 4—Mechanical design of the 10-GHz regenerative receiver, including wave-guide front and rear views. All dimensions are in millimeters.

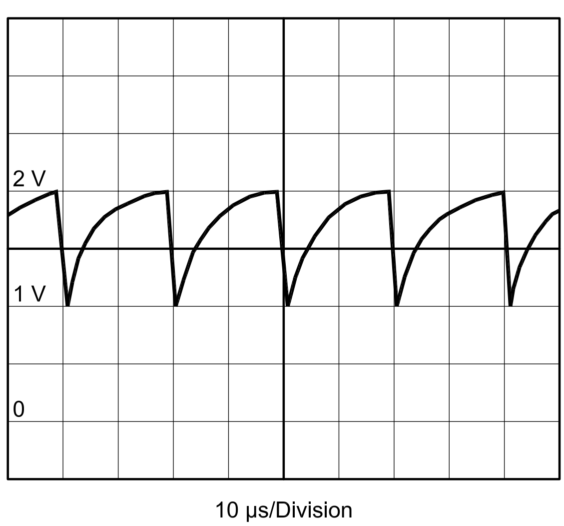

Fig 5—The low-frequency drain-voltage waveform.

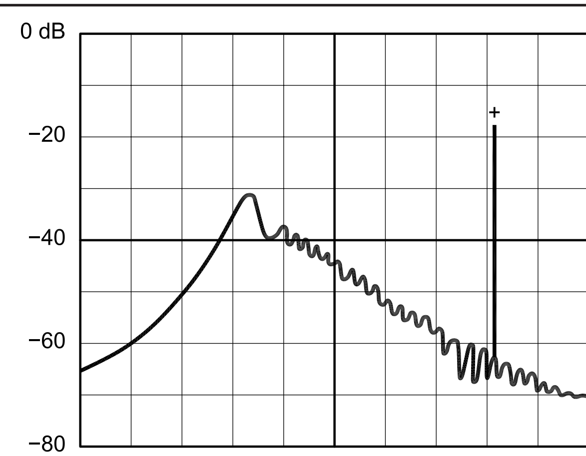

Fig 6—The superregenerative receiver frequency spectrum: scan width 2 MHz/division, scan time 0.5 s/division, bandwidth 10 kHz. "+" indicates the received signal added to the spectrum at the analyzer input.

rubber ring are used to fix it in position.

Two setting screws must be provided to round off the impedance matching. The waveguide is attached to a 20-dB horn made of epoxy-glass.

Measurements

DC, audio and quenching measurements are easy. The drain current is set by both the gate-source voltage and a variable resistor in the drain circuit. Super-regen behavior is insured between 0.8 and 2 mA, depending on the individual DRO characteristics. There is no super-regen action below, but only regeneration. Above, the device is always oscillating, without usable reception. Maximum sensitivity is obtained barely beyond where the hissing appears. Quenching frequency increases—as for any kind of self-quenching super-regen—when the signal is growing and proportional to the drain current. The selected values give a frequency varying from approximately 20 to 200 kHz.

Fig 5 shows the low-frequency waveform at the drain. Bandwidth depends on the quenching frequency (see Reference 4). Selectivity measurements have confirmed that statement and gave from 150 kHz to 2 MHz, depending upon the setting. The measurement was done using two unmodulated DROs as frequency-shift generators. The super-regen's radiated spectrum is shown in Fig 6.

At VHF and UHF, sensitivity measurements have been made showing quite good figures. For example, the MDS (minimum discernable signal) at 144 MHz was 150 nV with a 50-kHz bandwidth; at 432 MHz, 300 nV for 150 kHz; and at 1296 MHz, 300 nV for 500 kHz.

No attempt has been made to measure the actual sensitivity of the SHF, super-regen receiver, because I lack an accurate SHF generator and it is difficult to do measurements with a horn. However, the practical sensitivity seems to be comparable with a single-conversion receiver using a 1N21 diode, a Gunn diode as LO and an IF of about 85 MHz. Compared to a modern receiver with a PHEMT LNA and 3-kHz bandwidth, you can see a world of difference.

On the Air

I have made some experiments to receive various signals: CW, SSB, FM and NBFM. It is amazing to receive stations with such a simple receiver!

I can make a claim for the world distance record at 10 GHz with a super-regenerative receiver: made on September 28, 1997 with my friend F5AYE at a distance of 126 km!

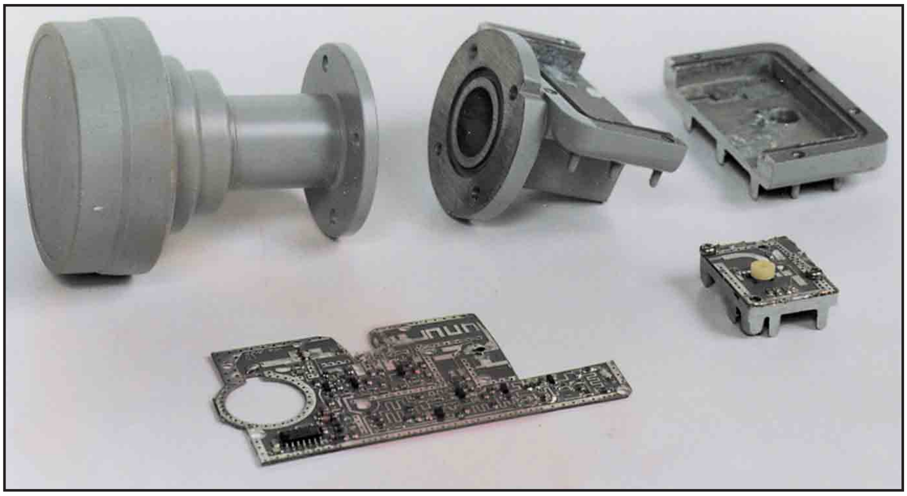

Fig 7—A DRO assembly removed from an LNA.

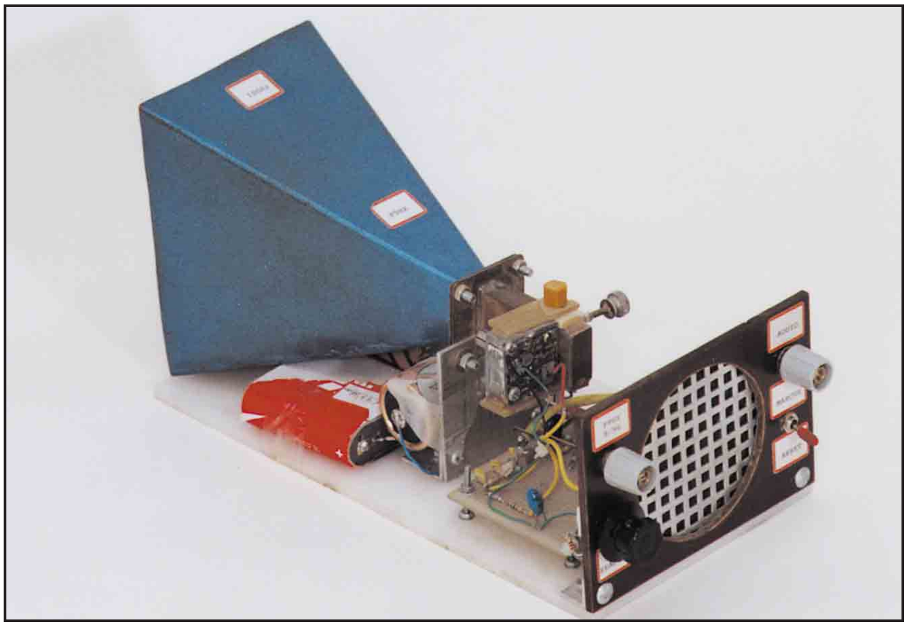

Fig 8—The author's 10-GHz superregenerative receiver.

Conclusion

It is not solely my intention to promote super-regenerative reception at 10 GHz; I want to open the way for much higher frequencies. As I wrote in jest, if I had an SHF oscillator, I would be able to make a receiver, even if it is at 76, 141 or 241 GHz. Now there are frequencies used by radar for automotive applications with Gunn diodes or DROs. So a very large field is open to radio amateurs to work at those frequencies.

Notes

[1] A. Jamet, F9HX, "La superréaction à 144, 432, 1296 MHz et 10 GHz," *Ondes Courtes Informations* (published by Union des Radio-Clubs, 11 rue de Bordeaux 94700 Maisons-Alfort, France), Jun/Jul 1996.

[2] A. Jamet, F9HX, "Un récepteur 10 GHz à superréaction," *Ondes Courtes Informations*, Nov/Dec 1996.

[3] A. Jamet, F9HX, "A 10 GHz Super-Regenerative Receiver," *VHF Communications*, January 1997.

[4] J. R. Whitehead, *Super-Regenerative Receivers*, Cambridge University Press, 1950.

André Jamet, licensed as F9HX in 1947, has published 100 technical articles in professional and Amateur Radio literature. He was previously the General Manager of Coredel-Chloride-France, a subsidiary of Chloride Group P.L.C. until 1988. He is now retired but still works in his lab and on the air.

Quartz Crystal Parameter Measurement

Surplus crystals are abundant, but we must know more than their frequency to use them. Here's a good way to measure their important characteristics.

By Jack Hardcastle, G3JIR

Designing and making your own crystal filters makes sound economic sense only if you need a filter of a frequency or performance specification that is not commercially available, or if you have a source of very cheap crystals and usually the latter reason prevails. Because of the availability of crystals for TV and computer applications, as well as surplus crystals from commercial radiotelephones, there is a wide range of frequencies available; unfortunately, they don't come labeled with their electrical properties. These must be measured before we can determine their suitability for our designs. The graphical evaluation method described here is an aid to discovering these qualities when using simple test equipment.

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Crystal Parameters

Fig 1 shows the well-known electrical-equivalent circuit of a quartz crystal. It comprises a series-resonant LCR combination shunted by the capacitance of the crystal's electrodes, plus circuit strays. This shunt capacitance creates a parallel resonance in addition to the series resonance, and this extra response complicates measurement of the value of the series components. It also plays an important part in determining the frequency response of the final filter. Its influence can be eliminated by resonating it at the series-resonant frequency by means of a shunt inductance. This is the traditional method usually described in textbooks; it can also be used to design filters with a symmetrical frequency response.[1] The alternate technique I shall describe neatly avoids the necessity for this procedure and allows both series and shunt components to be evaluated.

Measuring Series Resonance

The circuit used for measuring series resonance is in Fig 2. As shown, the waveforms of the signals at the input and output of the crystal are monitored on a dual-trace oscilloscope. Series resonance is indicated when the two signals are exactly

[1]Notes appear on page 10.

in-phase. This is a very sensitive indication of series resonance, and the effects of very small frequency changes (a few hertz) can be readily detected.

Firstly, the series resonance of the crystal alone is measured. Let this frequency be f_s. Then the measurement is repeated with *C1* in series with the crystal. Call this frequency f_1. The difference between these two frequencies is related to the product of the crystal's equivalent series capacitance, C_s, and its series-resonant frequency, f_s, by:

$$f_s C_s = 2C1(f_1 - f_s) \qquad \text{(Eq 1)}$$

C_s is sometimes referred to as the *motional capacitance* of the crystal. Plotted on a log-log graph, where the horizontal axis is the added capacitance *C1* and the vertical axis is the frequency shift Δf, this product results in a family of straight diagonal lines, as shown in Fig 3.

If the shift in frequency $f_1 - f_s$ were plotted on the graph for each value of *C1*, a curve similar to the three examples shown would result. The points should be plotted as they are measured, because any that do not conform to this general shape can be rechecked immediately. This ability to detect rogue measurements is the first of the benefits of using this graphical procedure.

Values of *C1* between 500 pF and 5 pF were used in these measurements. While the first three readings follow the diagonal lines very closely, subsequent readings deviate farther right because of the influence of C_p.

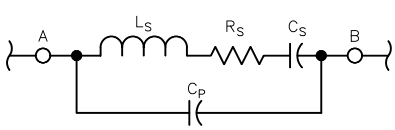

Fig 1—The equivalent circuit of a quartz crystal.

Table 1—Measurements made on a color-TV crystal

For C_p= 10 pF and f_s= 4430.849 kHz

C1 (pF)	C1+C_p (pF)	f_1 (kHz)	$f_1 - f_s$ (Hz)	C_s (pF)
500	510	4430.997	148	0.034
300	310	4431.102	253	0.035
200	210	4431.207	358	0.034
100	110	4431.540	691	0.034
50	60	4432.132	1283	0.035
20	30	4433.401	2552	0.035
10	20	4434.596	3747	0.034
5	15	4435.952	5103	0.035

Appendix: Derivation of Eq 1

Referring to Fig 1, and neglecting R_s in the absence of C_p:

$$f_s = \frac{1}{2\pi\sqrt{L_s C_s}} \qquad \text{(Eq A)}$$

Adding capacitance *C1* in series with the crystal gives a new series resonance at:

$$f_1 = \frac{1}{2\pi\sqrt{L_s \dfrac{C_s C1}{C_s + C1}}} \qquad \text{(Eq B)}$$

Separating L_s and C_s gives:

$$f_1 = \frac{1}{2\pi\sqrt{L_s C_s}} \times \frac{1}{\sqrt{\dfrac{C1}{C_s + C1}}} = \frac{f_s}{\sqrt{\dfrac{C1}{C_s + C1}}} \qquad \text{(Eq C)}$$

Rearranging this and dividing by *C1* gives:

$$f_1 = f_s\sqrt{1 + \frac{C_s}{C1}} \qquad \text{(Eq D)}$$

Since C_s divided by *C1* results in a very small quantity, the binomial approximation can be applied to the square root term:

$$ie\ \sqrt{1+\delta} \approx 1 + 0.5\delta \qquad \text{(Eq E)}$$

Thus:

$$f_1 = f_s\left(1 + \frac{C_s}{2C1}\right) = f_s + \frac{f_s C_s}{2C1} \qquad \text{(Eq F)}$$

Therefore:

$$f_1 - f_s = \frac{f_s C_s}{2C1} \qquad \text{(Eq G)}$$

which gives our expression for the product $f_s C_s$:

$$f_s C_s = 2C1(f_1 - f_s) \qquad \text{(Eq H)}$$

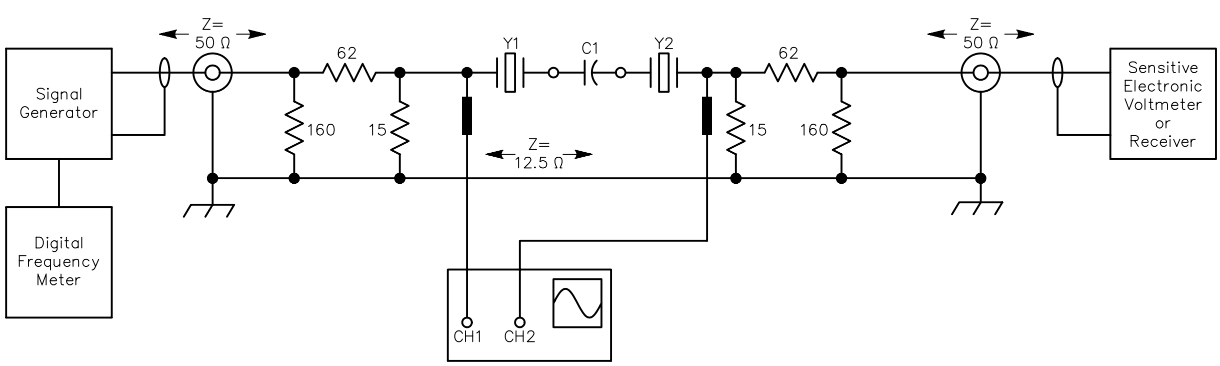

Fig 2—The circuit used for measuring series resonance.

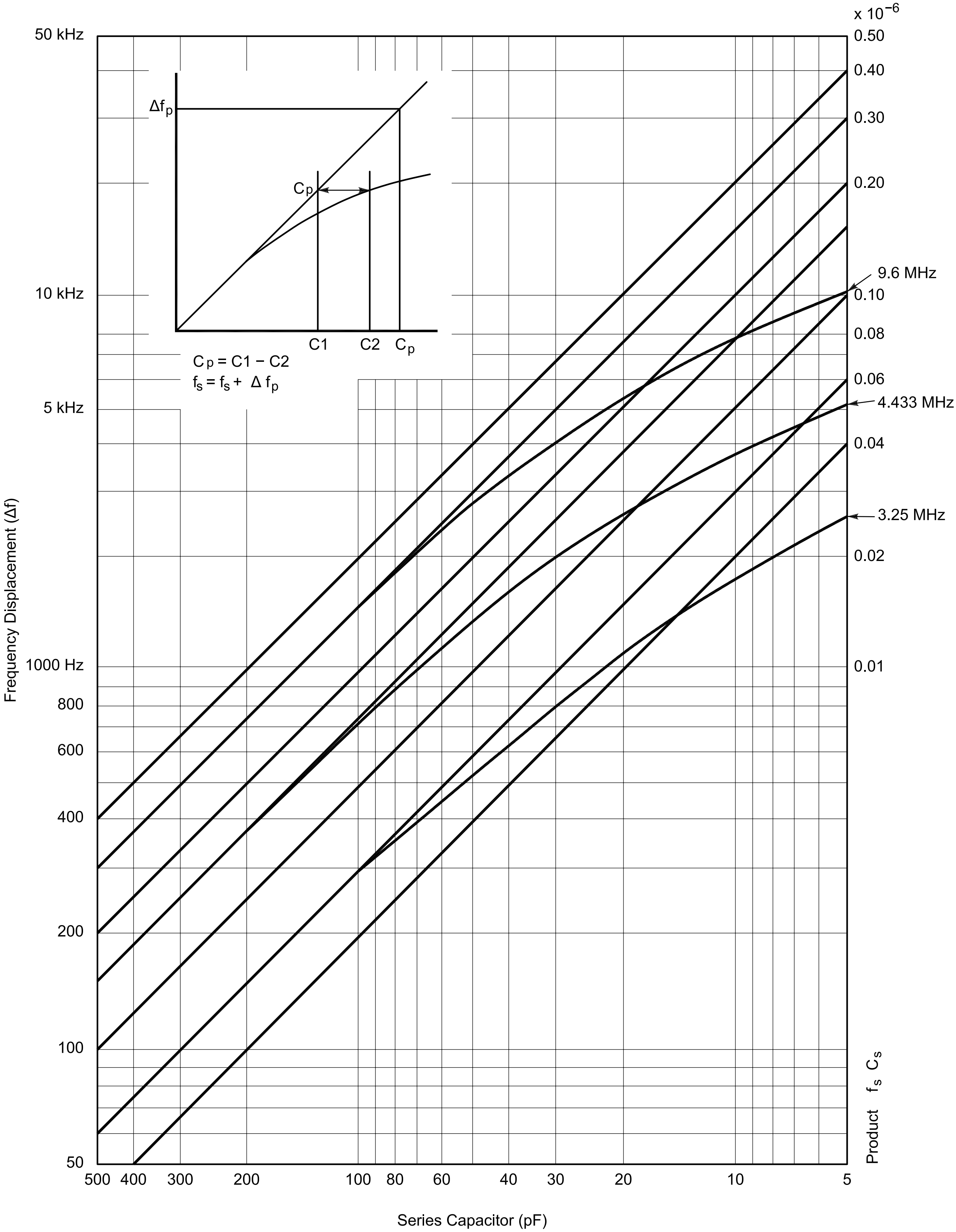

Fig 3—A plot of the various measurement results and values (see text for details).

Ignore these latter readings for the time being and project the diagonal, linking the first three readings to the right-hand edge of the graph. This allows the value of the product $f_s C_s$ to be determined and hence, following division by f_s, the value C_s. The effect of C_p has been neatly avoided.

It was noted above that C_p causes the graph to deviate farther to the right. C_p may therefore be measured directly from the graph, since it is the displacement between the curve and the diagonal. Several points may be checked and they will be found to give very similar readings. For the color-burst TV crystals, its average value was found to be 10 pF.

Notice that several crystals were checked on a LF capacitance bridge and their shunt capacitance, on average, was found to be 7 pF, so the measurement test jig must have contributed 3 pF of stray capacitance.

Now modify Eq 1 to include the shunt capacitance. Convincingly consistent values of C_s can be calculated over a wide range of series test capacitors *C1* using Eq 2:

$$C_s = \frac{2(C1 + C_p)(f_1 - f_s)}{f_s} \quad \text{(Eq 2)}$$

An example of this calculation is given in Table 1 for measurements on a UK color-burst (TV) crystal. A remarkably consistent series of values are calculated for C_s, even at very small values of *C1*. Reading $f_s C_s$ from the graph gives a value of 0.155 ∞10[6] and when this is divided by f_s (4, 430,849 Hz), a value of 0.035 pF is given for C_s. This corresponds very satisfactorily to the calculated values in Table 1 and shows that sufficient accuracy for most amateur designs can be achieved using the graph alone. The series capacitors used in these measurements are all silvered-mica types that had been checked on a capacitance bridge.

Parallel Resonance

Applying the graphical method for a third time, a crystal's parallel resonance can be determined. Using the value of C_p as measured above, we project upwards from the horizontal axis to the diagonal line. Using the value of 10 pF, previously obtained, gives us an intersection with the diagonal at a frequency displacement of 7800 Hz. This represents the frequency difference between the series and parallel resonances, so finding the parallel-resonant frequency becomes a matter of simple addition:

$$f_p = f_s + \Delta f \quad \text{(Eq 3)}$$

Since f_s = 4430.849 kHz and Δf = 7800 Hz, f_p = 4438.649 kHz.

Because the parallel resonance is so dependent on circuit strays, this reading (which includes the effect of such strays) is usually found more representative of conditions in a ladder crystal filter than is found by calculation. If you wish to try the calculation, Eq 4 gives the relationship between these quantities; it can be used to check the graphical method.

$$f_p = f_s \sqrt{1 + \frac{C_s}{C_p}}$$

$$= 4{,}430{,}849\sqrt{1 + \frac{0.035 \times 10^{-12}}{10 \times 10^{-12}}} \quad \text{(Eq 4)}$$

$$= 4438.596 \text{ kHz}$$

So f_p = 4438.596 kHz, which differs by only 53 Hz from the value obtained graphically. That is a convincing demonstration of the suitability of the method.

Equivalent Series Resistance

While set up for testing series resonance, the equivalent series resistance, or ESR, should be checked. To do this, insert the crystal into the test set without any series capacitance. Adjust the frequency for maximum output on the millivoltmeter. Record this value. Now remove the crystal and substitute a miniature variable resistor. This is then adjusted until the level measured on the meter is the same as noted previously with the crystal inserted. Remove the resistor and measure it on an ohmmeter. This value is the ESR and is used in Eq 5 to derive the Q of the crystal:

$$Q = \frac{1}{2\pi f_s C_s R_s} \quad \text{(Eq 5)}$$

It should be mentioned that this method of measuring ESR is open to error if harmonics are present in the test signal. If there is any doubt about the purity of the signal-generator output waveform, it should be followed by a low-pass filter. This is because the millivoltmeter can respond to these harmonics when the crystal is not in circuit, if the meter's response extends to a sufficiently high frequency for them to be within its range. Many older instruments do not respond to the harmonics, so this warning will not apply; but if the meter does respond, it will read higher than it would when given a pure signal. In consequence, the variable resistor must be set slightly higher also, so a pessimistic reading of ESR will result. When the crystal is in circuit, it allows only the fundamental to pass. Although crystals have responses at overtone frequencies, these responses do not coincide with the harmonics so the harmonics cannot pass. I stress that this is only a warning of what *could* happen to the unwary; it does not imply any special difficulty in using this method of measuring ESR.

Finally, it is worth buying more crystals than you need for a filter so that you can select those with the highest Q. Any with a Q significantly lower than average should be discarded.

Conclusion

Although this article has been written from the point of view of its application to ladder crystal filter design,[2] the crystal data are equally applicable to crystal-oscillator design. This interrelation has been exploited by Dave Gordon-Smith G3UUR (see Note 1). He measures C_s by placing the crystal in a Colpitts oscillator circuit and measuring the frequency shift when series capacitance is added. If you decide to use his test method, please note that he uses C_s and C_p to denote totally different quantities from the usage in this article. Apart from that, you can take the measured oscillator shift and apply the graphical evaluation method to it, for the measurement of equivalent series capacitance, C_s. It is yet to be proved that the same applies to the shunt capacitance, C_p.

Finally, I would like to acknowledge the contributions made by Lorin Knight, G2DXK, in our discussions during our endeavors to separate the various aberrations present in the graphical method. His input is greatly valued.

I would also like to record the valuable service that *QST* performed during my early years in Amateur Radio. Authors such as Byron Goodman, W1DX, George Grammer, W1DF, Don Norgaard and a great many more who were pushing forward amateur techniques were always read with great interest.

Notes

[1]W. Hayward, W7ZOI, "Refinements in Crystal Ladder Filter Design," *QEX*, June 1995, pp 16-21. This article was reprinted in *QRP Power* (Newington, Connecticut: ARRL, 1996; Order No 210).

[2]J. Hardcastle, G3JIR, "Computer-Aided Ladder Crystal Filter Design," *Radio Communication*, May 1983, pp 414-420.

Jack Hardcastle was born in 1932 and entered electronics at a very early age. His father's experimental broad-

cast receivers inspired his own teenage projects. These coincided with the early post-WW2 period, which was a time of plentiful, cheap war-surplus electronics. By 1950, he had constructed a television receiver, complete with its 6-inch, green CRT. An Amateur Radio license followed in 1953, and with this background, it was logical to work for a local company in Liverpool that manufactured carrier-telephony equipment. There he learned the importance of bandwidth conservation and noise considerations in transmission systems. It also sparked a continuing interest in all types of filters, particularly with respect to developing designs that could be applied by radio amateurs using very modest test gear.

This interest finally resulted in many articles on ladder-crystal-filter design that were published by the RSGB in Radio Communication, *some of which were also reprinted in* QST.

While working in telecommunication, he also pursued part-time study for professional qualifications, which ultimately resulted in IEE membership and Chartered Engineer status. After 10 years, Jack left industry and became an electronic engineer at Liverpool University, leading a small team designing and constructing electronic instrumentation for mechanical-engineering research.

After almost 30 years, he eventually retired in 1993, but still maintains a link to the University as an appointed "Honorary Research Associate." This gives Jack access to the University's libraries and their excellent collection of engineering books and journals.

Aside from design projects, Jack's Amateur Radio interests include 80 and 2-meter ragchewing, with occasional forays into the DX bands in search of IOTA stations.

Outside Amateur Radio, his interests include bird watching, walking and astronomy, enthusiasms that are shared with his wife, Betty, to whom he has been married for almost 40 years. □

The Rectangle Family of Antennas, Part 2: The Asymmetrical Double Rectangle (ADR)

Come learn about an antenna family that offers moderate gain, easy matching and great bandwidths.

By Dan Handelsman, N2DT

In this article, we shall examine the behavior of a class of double-loop rectangles and show how the second loop cures some of the problems endemic with simple rectangles. At the end, we will discuss some practical designs.

The Ills of the Rectangle

In the first article in this series,[1] which analyzed the simple rectangular loop seen in Fig 1, we arrived at some conclusions about the performance of such a loop when we attempted to push its gain beyond that of a square or quad loop. To do so, we found that we had to change its shape to increase the distance between the radiators and decrease their size.

While increasing the inter-radiator distance did increase the gain—to a theoretically maximal but practically unattainable 6+ dBi—the attendant shortening of the radiators brought some undesired effects. Those included a drop in the feed-point resistance (R_{in}), a loss of gain caused by increased antenna losses and a narrowing of the 2:1 SWR bandwidth.

Such antennas, when composed of thin wire and used on the lower HF bands, are loss-limited[2] in that they suffer from a loss in real gain, while their theoretical gain increases as their shape (aspect ratio)[3] is made more extreme. Aspect ratio is defined as the ratio of the inter-radiator distance and the radiator length. As one increases the aspect ratio, the R_{in} decreases. This, in turn, leads to an exponential increase in antenna currents along with the attendant power losses.

Because of the losses, all rectangles have a dimensional point of maximum gain, beyond which the gain falls off. This dimension, at which a decisive point is reached, is related to the currents and the total loop resistance. The loop resistance, in turn, is a function of the wire diameter and its ac resistance.[4]

At the other extreme, when rectangles are constructed with thick elements for the upper HF and VHF/UHF spectrum, the limiting factor in antenna performance becomes the feed-

[1]Notes appear on page 22.

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point resistance. We could, in fact, construct very-high-gain antennas with thick aluminum tubing for 10 meters and above that would have negligible losses, but which would be difficult if not impossible to feed because their extremely low R_{in}s. An example is a 10-meter rectangle, constructed with 1-inch aluminum tubing, a radiator separation or height of 16.4 feet (0.48 λ) and a radiator width of about 1 foot (0.028 λ), having a gain of 5.5 dBi but with an R_{in} of 1.25 Ω. The antenna losses are only 0.05 dB, but any practical matching circuit that steps this resistance up to 50 Ω would introduce far greater losses.

Another problem, regardless of with which end of the frequency spectrum we deal, is that all of the relatively high-gain antennas—those having gains in excess of 4 dBi—are severely limited in bandwidth. For instance on 80 meters, the rectangle with the maximum gain if one uses #10 AWG wire—one with an inter-radiator distance of 115 feet (0.41 λ)—has a bandwidth of slightly under 30 kHz. At the other extreme, on 10 meters, we must back off on the gain considerably to achieve a 1-MHz bandwidth while using 1-inch-diameter aluminum tubing. In Part 1, we determined that bandwidth was proportional to the size of the radiator and to the R_{in}.

The Cure

This article proposes a modification in the design of the simple rectangle that would cure its ills by increasing the R_{in} and bandwidth substantially while further increasing the gain. This modification is the addition of a second loop to the rectangle. We shall examine what happens when this is done, and we shall get into some practical designs at the end of this article.

If one starts with a resonant, single-loop antenna such as the simple rectangle and then conjoins another, smaller, loop onto it, the resulting antenna is an asymmetrical double rectangle (ADR), shown in Figs 2A and B. The "Hentenna" is the prototypical antenna of this class.[5]

This "hen" or weird antenna was conceived as a double loop of 0.5 λ between the two end radiators and with the radiator width set at approximately λ/6. It was fed at the middle wire, which was set about λ/6 from the near wire, and was specifically designed for a feed-point resistance of 50 Ω. As we shall see further on, the Hentenna is just one of a class of ADRs; but it has significant disadvantages in both gain and bandwidth.

Some Conventions

Look again at Fig 2A. The larger rectangular loop is the *primary loop*. The added loop is the *secondary loop*. The ADR results when the secondary loop is smaller than the primary loop. On the other hand, the secondary loop may be made as large as the primary and the antenna then becomes the SDR, or symmetrical double rectangle, seen in Fig 3. The SDR has been called many things: the DMS or double magnetic slot, the H-Double Bay and the Skeleton Slot.[6] At the suggestion of L. B. Cebik, W4RNL, when I wrote an earlier article in *Communications Quarterly*,[7] it was felt that "ADR" and 'SDR" were better, more descriptive and generic terms.

In Fig 2A, the end wires are defined by their relationship to the middle wire. The one nearest the center wire is the "near wire" and the wire at the farthest from it is defined as the "far wire." Since the center wire's position and the antenna's overall length define the size of the two loops, I shall be using another set of terms interchangeably with "primary loop/secondary loop size." This format will be, eg, 120/110/10. These numbers define a rectangle of 120 feet between end radiators, with a primary loop of 110 feet and a secondary loop of 10 feet. One can also think of this antenna as one that is 120 feet long and that has

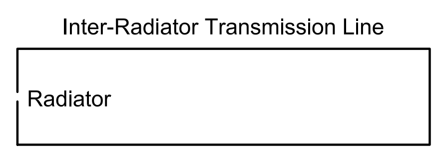
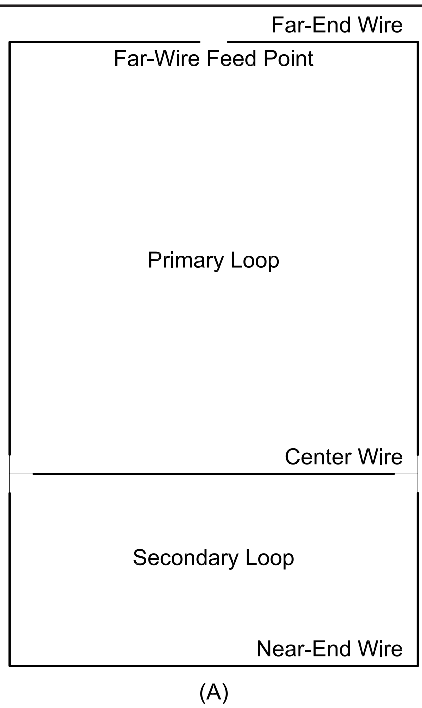


Fig 1—A simple vertically polarized rectangle for 80 meters.



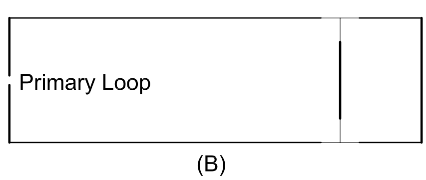


Fig 2—(A) A horizontally polarized ADR for HF/VHF. (B) A vertically polarized ADR for 80 meters.

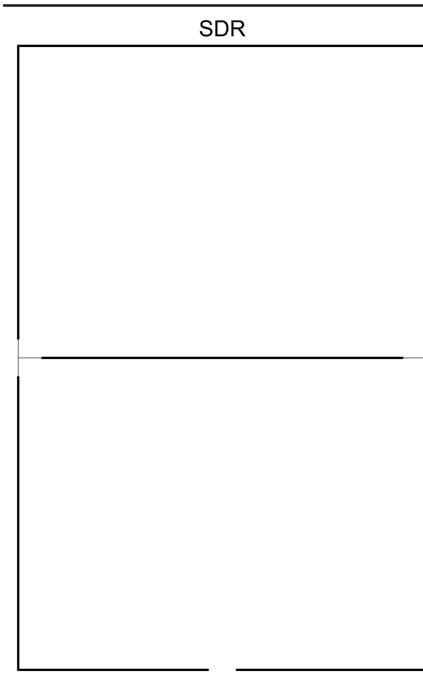


Fig 3—An SDR antenna.

the center wire 10 feet from the near wire—or 110 feet from the far one. I shall also use the term "center wire spacing" as the distance from the near to center wires. This is equivalent to the secondary-loop length.

The orientation of the radiators determines polarization: horizontal as in Fig 2A or vertical as in Figure 2B. Below 30 meters, vertically polarized antennas are preferred and, at 30 meters and above, the horizontally polarized antennas are superior.[8] As for free-space gain, it doesn't matter how they are oriented, and I will be switching back and forth between these modes. All of the further discussions will refer to free-space gain and the antennas will be analyzed at 10 meters (horizontally polarized) and at 80 meters (vertically polarized). The influence of ground on the antenna impedance, pattern or efficiency is ignored. Any exceptions will be noted.

All of the double-loop antennas can be fed at any of their three radiators. Each of the radiators can be fed, in turn, at its center or at an end. From the data that will be presented in this article, it is clear that the preferred feed-point is at the center of the far-end wire,[9] the farthest from the smaller, secondary loop (Fig 2A).

The ADR

The addition of the second loop has the following effects on the starting rectangle: The secondary loop is an impedance step-up transformer that:

1. Significantly decreases the system losses,
2. Significantly increases the bandwidth,
3. Increases the gain, and
4. Makes the antenna easy to tune.

One might suppose that, with the ADR, you may be getting something for nothing because the observed benefits arrive seemingly without any offsetting negative factors. After all, we are increasing R_{in}, gain and bandwidth at the same time, factors that are usually traded against each other in antenna design. Here, the tradeoff is structural. There is no getting around the fact that the design is more complex than that of the simple rectangle.[10] In totality, though, I believe that the benefits far outweigh any structural complexity. Let us now examine each of the benefits in detail.

The Step-Up Transformer Effect

The addition of a small secondary loop to a resonant rectangle results in a very negative system reactance. To resonate the system again, one must then increase the primary loop's perimeter.

We shall start with a resonant square or quad loop as the primary loop and add onto it secondary loops of various inter-radiator lengths, until the secondary loop becomes the same size as the primary. Figs 4 and 5 detail the changes in R_{in} and X_{in} as we do so on 80 meters. Here we begin with a resonant square loop that is 0.2604 λ per side. Looking at Fig 5, it is striking with short secondary loops that the reactance at any of the three feed

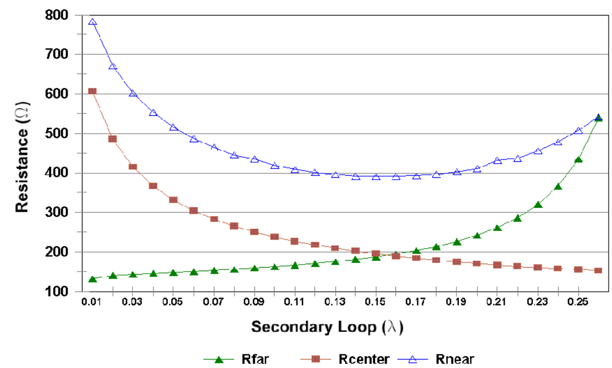


Fig 4—Feed-point resistances (for near, center and far wires) versus secondary-loop λ for an ADR with a primary loop of 0.2604 λ, secondary loops from 0.01 to 0.26 λ and wire diameter of 1×10⁻⁵ λ.

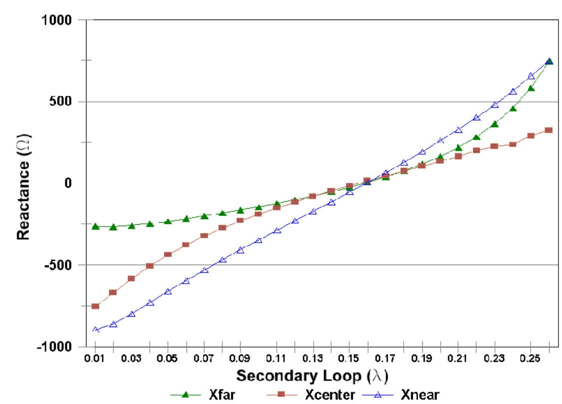


Fig 5—Feed-point reactances (for near, center and far wires) versus secondary-loop λ for an ADR with a primary loop of 0.2604 λ, secondary loops from 0.01 to 0.26 λ and wire diameter of 1λ10⁻⁵ λ.

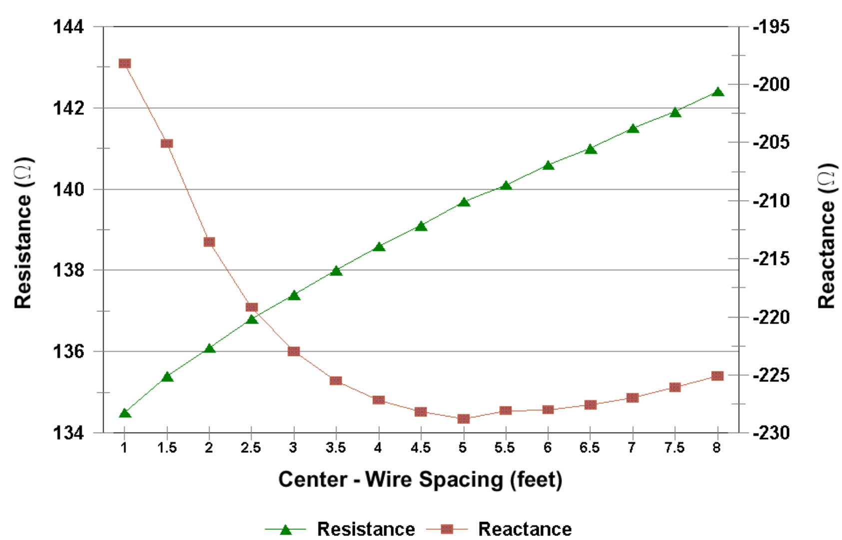


Fig 6—Feed-point resistance and reactance for an 80-meter ADR with secondary-loop spacing varied from 0.0036 to 0.0285 λ and wire diameter of 3×10⁻⁵ λ.

points (near the centers of each radiator) becomes highly negative. Fig 6, again modeling 80-meter ADRs with a starting resonant, square, primary loop, focuses on the reactance at very small secondary loop sizes with greater resolution. The center wire spacings are 1 to 8 feet (0.0036 to 0.028 λ) on 80 meters. The wire diameters used in modeling Figs 5 and 6 were 1×10^{-5} λ and 3×10^{-5} λ, respectively. The difference is practically insignificant but the latter corresponds to the #10 AWG wire I will be using as my standard for the lower HF antennas.

It is important to notice in the bigger picture, Fig 5, that the reactance at all wires starts out very negative with small secondary loops. It increases positively as the secondary loop is made larger. It then crosses the zero line at a certain point (which is dependent on the wire diameter), and then increases up to a positive maximum where the primary and secondary loops become equal in size. This high inductive reactance is the reason why each of the equal-sized loops of the SDRs is always substantially smaller than 1 λ in perimeter.

With respect to Fig 6, which is the higher-resolution picture at very small secondary-loop sizes, we note that the far wire's reactance dips to a maximally negative point before beginning its rise. The significance of this is that, if one chooses a secondary-loop size corresponding to the greatest negative reactance, the perimeter/radiator sizes are maximal when one resonates the system. This point is associated with the widest bandwidth and shall be discussed further in the section entitled "Bandwidth."

Decreasing Antenna Losses

In the first article of this series, we noted that thin-wire rectangles on the low HF bands were loss-limited as to their gains. We established, for instance, that a 115-foot-wide vertically polarized rectangle (with about 30-foot radiators) composed of #10 AWG copper wire was the one with the highest attainable gain on 80 meters. Increasing the radiator separation past 115 feet led to losses that increased more rapidly than the gain and overcame any further gain increase.

Let us now examine what happens to that same rectangle when we add a miniscule 1-foot secondary loop to it. The results are in Table 1.

The "potential" or loss-less gain of both antennas is about 5 dBi. The simple rectangle, on row 1, has 0.67 dB eaten up in losses. Row 2 shows that, if we add on a tiny 1-foot (0.0036 λ) secondary loop, the R_{in} changes little and the X_{in} becomes significantly negative, at –96.2 Ω. There is an insignificant increase in gain.[11] If we now resonate the antenna by increasing its radiator height from 29.68 to 35.19 feet (as in row 3), the R_{in} increases by 37%. The losses have decreased by 0.21 dB and the gain has increased by about an equal amount, 0.23 dB.

What can we conclude about this antenna? *The secondary loop, minuscule as it may be, has led to a significant increase in the radiator size, increased the* R_{in} *significantly and has increased the gain by an amount equal to the decrease in the antenna loss.*

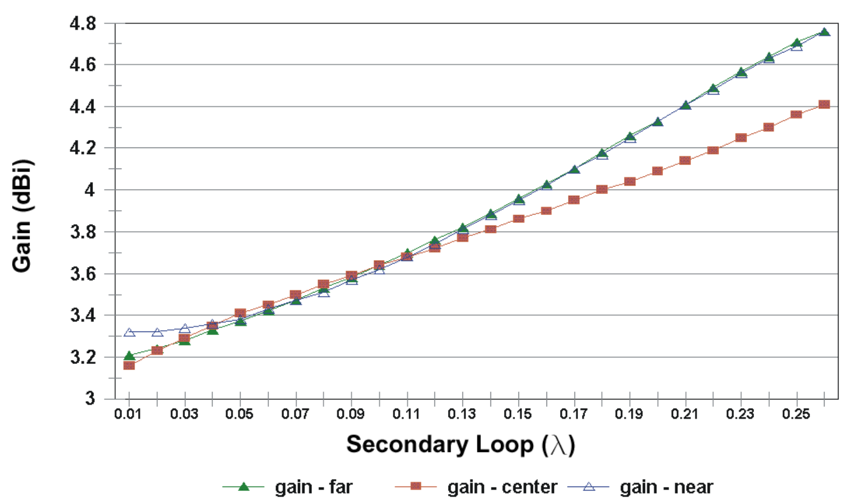


Fig 7—Gain versus secondary-loop λ for an ADR with a primary loop of 0.2604 λ, secondary loops from 0.01 to 0.26 λ and wire diameter of 10^{-5} λ.

Table 1—The Effect of a 1-foot Secondary Loop on a 115-foot Rectangle Antenna

#10 AWG; f = 3.5 MHz

	Primary Loop		*Secondary Loop*					*Potential*
	Width (ft)	*Height (ft)*	*Size (ft)*	*R (Ω)*	*X (Ω)*	*Gain (dBi)*	*Loss (dB)*	*Gain (dB)*
Rectangle	115	29.68	—	18.9	0	4.31	0.67	4.98
ADR	115	29.68	1	18.7	–96.2	4.37	0.60	4.97
	115	35.19	1	25.7	0	4.54	0.46	5.00

Table 2—Two ADR 160s: Dimensions and R_{in} while Holding Gain within 0.01 dB of Maximum (4.90 dBi)

#10 AWG wire; f_{res} = 3.5 MHz

Center-Wire Spacing (ft)	*Radiator Height (ft)*	*1st Loop (ft)*	*2nd Loop (ft)*	*R_{in} (Ω)*
48.2	34.4	111.8	48.2	27.6
59.3	44.7	100.7	59.3	55.5

Gain Increases Caused by the Second Loop or Third Radiator

The gain of the double-loop systems, as seen in Fig 7, increases monotonically with the increase in secondary loop size until the two loops are equal in size. At that point, the antenna becomes an SDR or symmetrical double rectangle. Notice one of the reasons for my preference of an end wire feed point: The gain at the center wire falls off substantially as the secondary loops become larger. This is one major disadvantage of the original Hentenna.

When dealing with loss-less antennas, the "limit" SDR becomes equivalent to three independent, infinitesimally short or "Hertzian" dipoles separated by 0.5-λ sections of transmission line. At this theoretical point, and within limits of modeling capability,[12] the system gain exceeds 7.18 dBi. Therefore, the limit SDR has a gain of somewhat more than 1 dB over that of the limit rectangle.

If one looks at only the gain in real life, ADRs having "lossy" wires reach states of gain equilibrium at any double-loop length.[13] Look at Fig 8 to see what I mean. The antennas referred to, vertically polarized on 80 meters, are DRs that are 140-feet long (0.5 λ) between the end radiators. Notice that the 140/100/40 antenna has about the same real gain as one that is 140/120/20. This is so because the one with the longer primary loop (120 feet) has a higher theoretical gain and higher offsetting losses compared to the one with a 100-foot primary loop.

We can look at this phenomenon from another viewpoint—one that provides us with more insight into how ADRs play. Table 2 shows the dimensional parameters for two 160-foot-long ADRs of two different primary-loop lengths having the same realized gain.

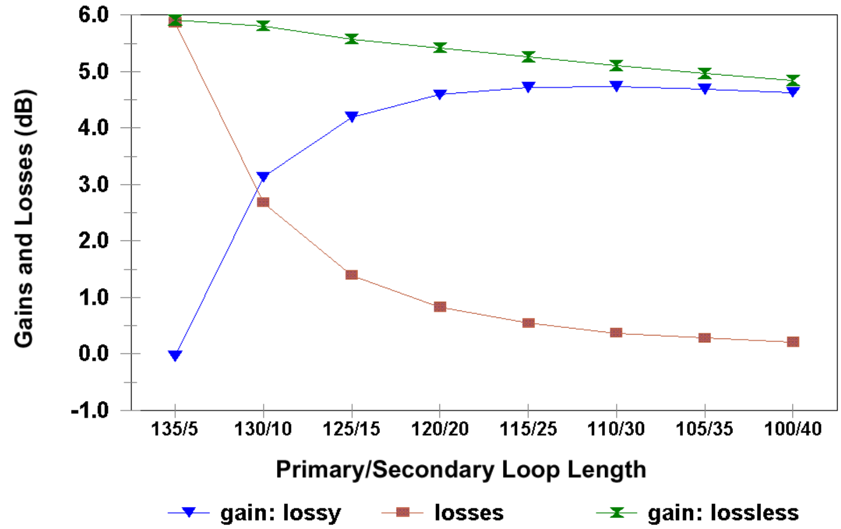


Fig 8—Gains and losses at different primary/secondary loop combinations for a lossy 140-foot ADR.

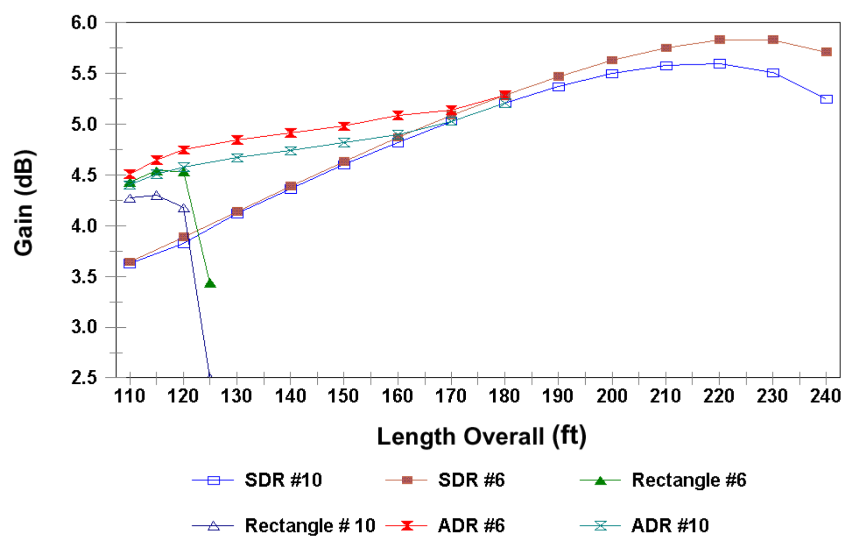


Fig 9—A gain comparison of lossy loops: rectangles, ADRs and SDRs.

Table 3—10 Meter ADRs: R_{in} = 50 Ω; Horizontally Polarized with 13-foot (±) Primary Loops

Overall Height (ft)	*Radiator Width (ft)*	*Spacing (ft)*	*R_{in} (Ω)*	*Gain (dBi)*	*Bandwidth (MHz)*	*1st Loop Height (ft)*
14	6.07	0.90	49.8	4.66	1.14	13.10
15	6.06	1.91	49.9	4.77	1.14	13.09
16	6.00	2.95	49.8	4.86	1.14	13.05
17	5.96	3.98	50.0	4.94	1.13	13.02
18	5.88	5.04	50.2	5.02	1.11	12.96
19	5.77	6.06	50.0	5.11	1.10	12.94
20	5.66	7.08	50.2	5.21	1.10	12.92
21	5.49	8.10	49.9	5.31	1.09	12.90
22	5.33	9.08	50.2	5.44	1.04	12.92
23	5.09	10.05	49.9	5.57	1.00	12.95
SDR 26	4	13.00	50.2	6.31	0.75	13.00

These two antennas have the same gain but there are important differences. The one with the shorter primary loop has a greater radiator height (these are vertically polarized) and a higher R_{in}. We shall also see below, under "Bandwidth," that the antennas with the tallest radiators have wider SWR bandwidths—in fact, the antenna with the greater R_{in} has twice the bandwidth of the other. Please also notice that all antennas within that range of primary/secondary loop sizes will have the same gain, but intermediate R_{in}s, bandwidths and heights.

Now look at Fig 9. Until we reach an antenna of about 170 feet overall length, the gains of the ADRs are higher than those of SDRs. This is so because in ADRs the primary loop is the major determinant of gain, and these have much longer primary loops than the SDRs. At that specific dimensional point—170 feet (0.6 λ) on 80 meters with #10 AWG copper wire—the two equal-sized loops of the SDR begin to predominate from a gain-loss equilibrium standpoint, and all longer antennas with higher gain will have to be SDRs.

Table 3 illustrates another strategy one can use in constructing ADRs. Here we are using thick 1-inch tubing at 10 meters. These antennas are modeled as horizontally polarized with horizontal radiators having a width dimension and with the inter-radiator distance as a height dimension. As a reference, a simple rectangle with the primary loop height of 13 feet has an R_{in} of 40 Ω and a gain of 4.49 dBi. Spacing is the size of the secondary loop. None of these antennas is a Hentenna, since that antenna, while also designed for a 50 Ω R_{in}, is fed at the center wire and, along with different dimensions, has a lower gain and a narrower bandwidth.

Here we have taken the 13-foot primary loop (0.38 λ) of the original rectangle and kept adding longer secondary loops (1 to 13 feet or 0.029 to 0.38 λ) until both loops were the same size (where the ADR becomes a SDR). Even the addition of a 1-foot secondary loop increases the system gain by about 0.17 dB. This is all real gain and not simply loss reduction, since the simple rectangle, constructed with such thick tubing, has almost no losses at the start. However, the secondary loop is already fairly substantial, being about 0.03 λ long, and the extra radiator contributes to the gain.

Another factor to consider in antenna design with such thick wires is that the length of the primary loop determines the R_{in} at the far end wire, no matter how large the secondary loop, until equality is reached. A longer primary rectangle yields a lower R_{in}. Therefore, one can simply choose a primary loop length to select for a target feed-point resistance. In the antennas in Table 3, I targeted an R_{in} of 50 Ω.

Bandwidth

We shall again begin by using a square quad loop for reference. Table 4 gives us a feeling for what happens to a simple vertically polarized quad loop on 80 meters when we add on a short secondary loop of 5 feet (0.018 λ). We can resonate the system by increasing the primary-loop perimeter, either by increasing the radiators' heights or the inter-radiator distance of the primary loop. If we keep the radiator heights the same as the original quad loop's, but increase the width or distance between them; the gain increases by almost 0.5 dB. The losses are negligible at the start and remain so. In spite of the increase in gain, the bandwidth actually increases when compared to the starting quad loop. Look now at the highlighted row for what happens when we resonate the system by making the radiators taller. There is a smaller increase in gain but the bandwidth increases markedly by about 54%.

The next four rows replicate what we did in our 10-meter examples of Table 3. I targeted an R_{in} of about 200 Ω and increased the secondary loop size from about 0.18 to 0.26 λ (51.45 to 72.09 feet) while maintaining the primary loop width (inter-radiator distance) constant at 73.5 feet. The gain

Table 4—Adding Secondary Loops on to a Quad Loop on 80 Meters

Antenna	*Radiator Ht (ft)*	*Width (ft)*	*2nd Loop (ft)*	*R_{in} (Ω)*	*Gain (dB)*	*Loss (dB)*	*BW (kHz)*
Quad	73.5	73.5	0	126.3	3.17	0.09	184
ADR	73.5	84.03	5	138.2	3.65	0.11	196
	84.15	**73.5**	**5**	**196.1**	**3.30**	**0.06**	**283**
	71.4	73.5	51.45	197.1	4.10		232
	68.46	73.5	58.80	196.8	4.25		214
	65.15	73.5	65.15	197.7	4.42		196
	62.2	73.5	72.09	199.4	4.56		175

Table 5—Comparison of Four Sets of Rectangles/ADRs on 80 Meters

3.5 MHz, Wire D = #10 AWG

Antenna	*Width (ft)*	*Radiator Ht (ft)*	*2nd Loop (ft)*	*R_{in} (Ω)*	*Gain (dB)*	*Loss (dB)*	*BW (kHz)*
Rectangle	100	46.28	0	46.7	4.04	0.26	69
ADR	100	55.15	4	68.4	4.21	0.17	101
Rectangle	105	40.82	0	35.8	4.18	0.34	53
ADR	105	48.96	3	51.7	4.38	0.23	76
Rectangle	110	35.29	0	26.6	4.29	0.47	40
ADR	110	42.56	3	38	4.53	0.31	56
Rectangle	115	29.68	0	18.9	4.31	0.67	29
ADR	115	35.91	4	26.6	4.62	0.42	40

increased with secondary-loop size and the bandwidth decreased gradually. However, the bandwidth of the highest-gain antenna, which is almost a SDR having a gain of 4.56 dBi (about 1.4 dB more than where we started), was only marginally narrower than the quad's.

The addition of a small secondary loop exerts the same effect on all other forms of the rectangle. Table 5 shows what happens to four vertically polarized 80-meter rectangles (100 to 115 feet long)[14] when we add on a secondary loop calculated to yield the highest negative reactance and hence the largest loop perimeter. The rectangles, of course, have a zero secondary loop.

When we compare the changes in radiator height, R_{in}, loop perimeter and bandwidth against the inter-radiator distance, the strongest correlation—as we had seen with the simple rectangle—is between R_{in} and bandwidth. They are similar functions. For antennas designed for the same R_{in} but with different-sized secondary loops, the best correlation with bandwidth is the primary loop's perimeter. In Tables 3 and 4, the SDR has the smallest loop perimeter (per loop) and hence the narrowest bandwidth.

The findings in Table 5 can be summarized as follows:

1. As we already know, the gain of the rectangle increases with inter-radiator separation.
2. The addition of the small secondary loop requires an increase in the overall radiator size (height), reduces the antenna-conductor losses and increases the gain of the ADR over that of the parent rectangle.
3. The ADR bandwidth is inversely proportional to radiator separation as it is for its parent rectangle. The bandwidth correlates strongly with the R_{in}.
4. At each rectangle length, the ADR has a significantly greater bandwidth than its parent.

Now recall Table 3, where we modeled thick-wire antennas on 10 meters. Here we see the same phenomenon. The bandwidth decreased as we made the secondary loop larger. This increased the gain but also resulted in a smaller primary-loop perimeter. We can also see the contribution the wire size has made to the bandwidth overall.[15] The 10-meter antenna with the smallest secondary loop (0.9 feet or 0.026 λ) had a bandwidth of 1.14 MHz or 4%. Its 80-meter analog (Table 5), the ADR with the 105-foot (0.37 λ) primary loop and the 3-foot (0.01 λ) secondary loop, had a bandwidth of 76 kHz or slightly over 2%.

Let us now go back to the Hentenna for a moment. It is now clear that with the dimensional constraints necessary to achieve a feed-point resistance of 50 Ω at the middle wire, the radiator size is excessively small. As an example, I have modeled Hentennas with 1-inch tubing on 6 meters (the most popular band on which it is used in Japan) at a center frequency of 52 MHz. The results are quite telling.

For the Hentenna with middle-wire feed, the gain is 5.17 dBi and, most significantly, the bandwidth is only 1.5 MHz. The same antenna, as an ADR fed at the far wire, has a gain of 5.25 dBi and a bandwidth of 3.5 MHz. The radiator sizes and primary-loop perimeters are 0.1453 λ and 1.11 λ for the Hentenna, and 0.1815 λ and 1.38 λ for the end-fed ADR. With the different feed points and the same overall heights, we find minor differences in gain and major differences in bandwidth.

The last point I wish to make with reference to bandwidth is that the examples I have used at both extremes of the HF spectrum involve the widest bands: 80 meters occupies about 14% and 10 meters about 6%. For all of the other bands, we can push the gain far higher and still cover the full bandwidth.[16] For the WARC bands, we can design extremely high-gain antennas, using primary rectangles with higher aspect ratios, since bandwidth is not a big consideration. Here we are limited solely by R_{in}.

Tuning

With these antennas you can "tune without prune." As I pointed out in my *Communications Quarterly* article,[17] all ADRs can be tuned by moving the center wire[18] (changing the ratio of primary/secondary loop sizes). If we move the center wire toward the geometric center of the antenna, we can raise the resonant frequency by decreasing the primary-loop perimeter. The converse, moving the center wire toward the near-end wire, lowers the resonant frequency.

Fig 10 illustrates this effect for an ADR of 130 feet overall length on 80 meters. If the antenna is tuned to 3.5 MHz, moving the center wire from 16 to 36 feet from the near end will resonate the antenna anywhere in the 80-meter band.[19] The change in resonant frequency is almost linear with middle-wire spacing. The only caveat here is that *the antenna must be reasonably asymmetrical to start*; that is, the secondary loop must be small compared to the primary. Why? Because the tuning effect, or the change in reactance with movement of the center wire toward the geometric center, becomes nil as the center wire approaches that center.

If one starts with an antenna that is almost symmetrical, it would then be almost impossible to raise the resonant frequency. SDRs, which are symmetrical at the start, can only be tuned by pruning. If we make the ADRs rea-

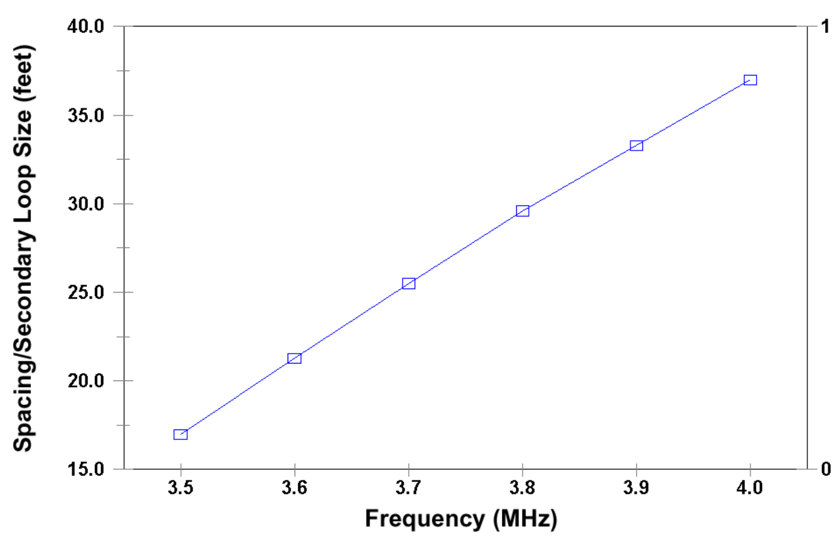


Fig 10—The change of spacing with frequency of a 130-foot ADR.

sonably asymmetrical, moving the center wire can tune out any reactance associated with ground effects and place the resonant frequency exactly where one wants it to be.

The next question that comes to mind is: Can one make an ADR with two center wires to enable an instant QSY from, say, the CW to SSB segments on such a wide band as 80 meters? The answer is "yes" if one uses relays to open and short the appropriate center wires and if one makes some dimensional changes. This will be a subject of a later article.

Beamwidth

Unfortunately, gain with vertically polarized antennas comes with a narrowing of the azimuth beamwidth. The 80-meter ADRs we have discussed have beamwidths of about 60°. Real pattern narrowing with gain (over 6 dBi) fortunately occurs only with the very large SDRs of 200+ feet in length.

The good news, though, is that when one uses horizontally polarized antennas, the lobe compression occurs solely in the elevation lobe and not in the azimuth lobe. The horizontal beamwidth is always close to 90° and that makes for exceptional gain antennas at 30 meters and higher. These have the added advantage of high-elevation-angle QRM suppression.

A Theoretical Description of the ADR

We have seen that the performance of simple rectangles can be easily explained by the effects of the mutual coupling between the elements and by transmission-line theory.[20] Similarly, these modalities suffice to explain the performance of the SDR with three parallel, equally spaced radiators—albeit with some more complexity. I plan to discuss this more in a later article.

The ADR is not explained so easily. There are complex interactions between the three asymmetrically spaced radiators. At this time, I have the inclination but neither the time nor the computing power to deal with this problem. However, some thoughts do come to mind.

When one looks at the impedance curves, R in Figure 4 and X in Fig 5, we can see that the far-wire impedance is reasonably stable. This is not so with the Zs at the closely spaced near and center wires. It seems as if, with narrow center-wire spacing (tiny secondary loops), these two radiators simply act as one but divide the current between them. This is confirmed by the loss-less system gain, which is not much different from that of the parent rectangle. As the secondary-loop size (or center-wire spacing) increases, we begin to see "stacking" gain, Fig 7, of the three radiators. This becomes maximal when the antenna is an SDR, and there is equal separation between the center and the two radiators at either end.

For ADRs where the secondary loop is very small—the near and center wires are close together—one can estimate the input resistance at the center of the far wire by simply assuming that the currents divide equally between the far radiator and the pair at the other end. David Jefferies, G6GPR, and one of his graduate students measured the gains, bandwidth and Z_{in} of a group of ADRs from 1.5 to 2.5 GHz. These were dimensionally structured for an R_{in} of 200 Ω at 1.8 GHz.

He found that the pair of radiators at one end and the far radiator carried about equal currents. He then assumed—solely for the sake of estimation—that the currents divided equally among the two closely spaced wires. Therefore, the current division as estimated was 1/4, 1/4, 1/2. Because, for a constant power input, resistance is proportional to the square of the currents, this indicates a resistance division of 1/16, 1/16, 1/4. Since the ADRs radiators were quite long in these designs, he estimated that the far-radiator resistance was 8/3 × 72 Ω or 192 Ω—a value close to the design resistance of 200 Ω.

Most importantly, his work together with my constructed antennas validates the use of NEC modeling for this particular class of antennas. When the effects of ground are not a factor—antenna height above ground is more than 2 λ—the constructed antennas perform almost exactly as modeled.

I have found it easier to picture what happens when you add a secondary loop by using the following reasoning: Imagine a resonant rectangle of perimeter approximately 1 λ at the desired frequency. To that, physically attach a loop of much smaller perimeter that has a significant capacitive

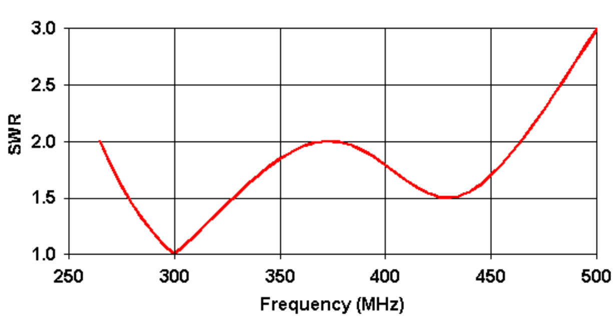


Fig 11—SWR versus frequency of one ADR (designed for 300 MHz) over a frequency range of 260 to 500 MHz.

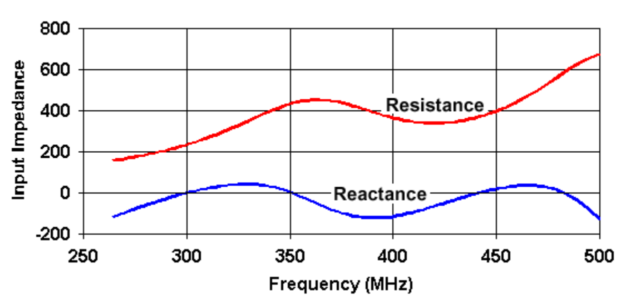


Fig 12—Impedance versus frequency of one ADR (designed for 300 MHz) over a frequency range of 260 to 500 MHz.

Table 6—Wide-Band ADR Designed for 300 MHz

Wide-Band ADR for 300 MHz

Wire Diam.(λ)	*Height Primary(m)*	*Spacing Secondary(m)*	*Width(m)*	*Gain(dBi)*	*R_{in}(Ω)*
0.006	0.38	0.096	0.3332	4.04	234.8

reactance at that frequency. The new antenna therefore acquires a significant negative reactance; it must be resonated by increasing its primary-loop perimeter via either of the two methods we have discussed earlier.

Some Design Examples

Many games can be played with the ADR in increasing the gain, the R_{in} or the bandwidth of a rectangle on various bands. These have been described in *antenneX*.[21] For example, rectangles for the WARC bands can have such extreme aspect ratios (short radiators and large radiator separations) that the low R_{in} renders them difficult to feed. In this case, the addition of a small secondary loop can significantly increase the R_{in}. Similarly, on a wide band such as 10 meters, a simple rectangle will have insufficient bandwidth to cover the full 6%. In this case, the secondary loop can be used to widen the bandwidth and enable full coverage.

With thick-wire ADRs at VHF, I stumbled on an interesting and useful phenomenon, illustrated in Fig 11, which shows the SWR of one ADR designed for 300 MHz over a frequency range of 260 to 500 MHz. The curves for its R and X are shown in Fig 12. This antenna has a tremendous gain and SWR bandwidth. Why is that so? It is because the two loops have contiguous resonances related to the loop perimeters.[22] If we look at Table 6, we can calculate the primary-loop perimeter: 1.2344 λ. The secondary loop perimeter is 0.8584 λ. Based on an f_{res} of 300 MHz for the primary loop, the f_{res} of the secondary loop comes to 431 MHz. Now look again at Fig 11 for the second SWR minimum: It occurs just about where calculated.

When designed for 300 MHz, the ADR described in Table 6 has the following performance characteristics: Its 2:1 SWR bandwidth is from 265 to 464 MHz (199 MHz or 66%). The useable bandwidth, defined as a gain greater than 3 dB and an SWR less than 2:1, is 265 to 399 MHz (134 MHz or 45%). The gain at the low end is 3.43 dBi, the peak gain is 4.42 dBi at 345 MHz and the gain at the upper limit is 3 dBi.[23]

This antenna was constructed with 0.25-inch aluminum tubing. The 2-meter version would employ 0.5-inch tubing. Its dimensions are in Table 6. The overall height is 0.38 λ and we can characterize it as a 0.38/.284/.096 ADR.

The question then arose whether we could scale this antenna to cover three ham bands from 21 to 30 MHz. Of course, we can if we use the scaled tubing diameter of about 3 inches, but this is clearly out of the question. However, L. B. Cebik, W4RNL, came up with a solution. Cebik has been working on the use of a wide-spaced pair of thin-gauge wires to approximate the diameter of much thicker wires. He is using this with quad antennas to increase their performance. His technique is shown in Fig 13 and I used it to come up with the ADR shown in Fig 14.

At 300 MHz, the parameters of my twin-wire ADR are listed in Table 7. It is important to notice that the antenna's loops are composed of copper wires of about 2×10^{-4} λ diameter. The wire diameter is only about 0.2 mm or 0.009 inches. This number was arrived at by scaling the diameter of #10 wire to the design frequency. A reference "Cebik quad" with the same wires and spacing has a gain of 3.34 dBi.

This ADR loop can be scaled to cover 21 to 30 MHz. Its SWR bandwidth, when resonated at 23.8 MHz, is 20.5 to 40 MHz and its usable bandwidth is 9.75 MHz. The SWR curve is in Fig 15. The dimensions, in wavelengths

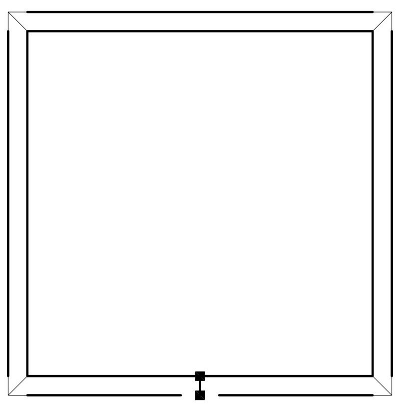


Fig 13—A twin-wire quad loop.

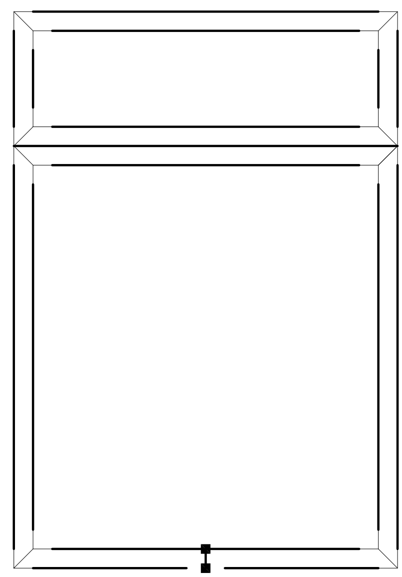


Fig 14—A twin-wire ADR.

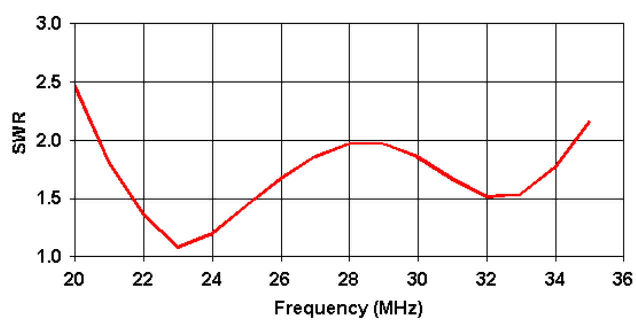


Fig 15—SWR curve for a wide-bandwidth ADR described in the text.

Table 7—VHF ADR Constructed with Twin Thin Wires

Wire spacing = 0.0259 λ or 1 inch, which corresponds to 12 inches at 25.5 MHz (the Center Frequency between 21 and 30 MHz)

					Gain (dBi)			
Height (m)	*Spacing (m)*	*Width (m)*	*Gain (dBi)*	*R_{in} (Ω)*	*Low*	*High*	*Center*	*BW (MHz)*
0.388	0.09	0.36	3.82	254	3.33	3.01	4.21	129

Table 8—HF ADR Composed of Twin Parallel #10 AWG Wire Loops

Wire Spacing is 12 inches; f_{res} = 23.8 MHz; dimensions are for the outside loops

	Height	Primary	Spacing	Width	Gain (dBi)	R_{in} (Ω)	Gain (dBi) Low	High	Center	BW (MHz)
λ	0.373	0.288	0.085	0.36	3.82	300	3.03	3.5	4.11	20.5-30.25
inches	191.4	147.8	43.6	184.7						

and in inches, are found in Table 8. This antenna is constructed with #10 AWG copper wires spaced such that the parallel wires of the loops are separated by exactly 12 inches. Therefore, the interior-loop dimensions are exactly 12 inches smaller than those of the outer loop. Notice the feed-point resistance of 300 Ω, which can be readily matched.

Feeding this antenna involves the following: Open the feed points at the centers of the lower horizontal wires of both loops. Then connect the feed points with a parallel 300-Ω transmission line 12 inches long. Feed the outer wire gap.

This ADR is no more difficult to construct than a triband quad having three concentric elements. Its bandwidth is so great that ground effects on the resonant frequency will be inconsequential. Its gain over the range will be greater than or equal to that of a quad. Fig 16 compares the gain of this ADR over 21 to 29.7 MHz with three optimized "Cebik quads" whose resonant frequencies are centered on 21.225, 24.91 and 28.5 MHz. This ADR is superior at all frequencies except above 29.6 MHz. A single-wire quad loop of any practical wire diameter won't come close.

The wide bandwidth and the high R_{in} of the ADRs make them especially "well-behaved." This is a term that David Jefferies, PhD, G6GPR, uses for antennas that are not sensitive to detuning by the elements, ground or other proximity factors such as other antennas on the same mast or human bodies holding them.

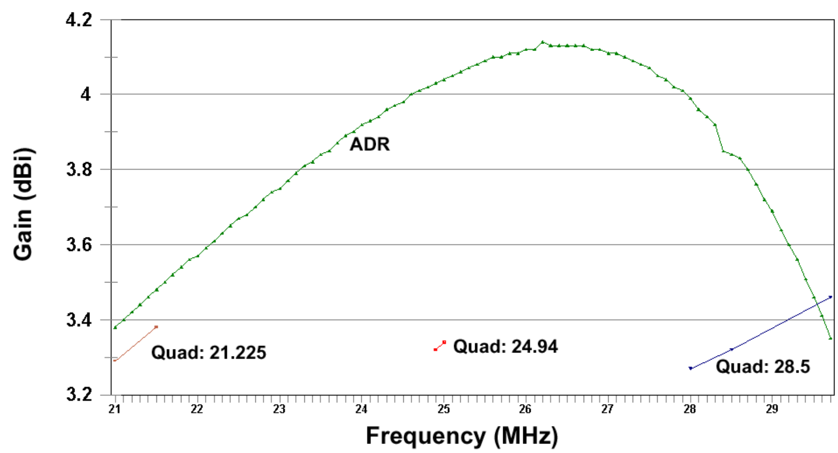


Fig 16—A gain comparison of wide-bandwidth antennas from 21 to 29.7 MHz. All antennas used #10 AWG wire.

Tradeoffs

Certain tradeoffs in ADR design haven't made it into this article. We already know that antennas with larger primary loops, shorter radiators and higher gain have narrower bandwidths. On the other hand, the shorter radiator may be highly advantageous with vertically polarized antennas on the lower HF bands. Here, height above ground is important when one considers the takeoff angle (TOA) and, for a fixed top wire height, a smaller radiator along with a higher bottom wire may actually increase radiation at a lower TOA.[24]

In any case, as I have discussed with reference to simple rectangles, a 30-kHz bandwidth may be more than enough to cover a DX window on 80 meters. Therefore, pushing the gain and making the antenna lower in overall height may not be a disadvantage.

Conclusion

The addition of a secondary loop to any sized rectangle has the following effects: It makes the rectangle highly negatively reactive which, when that reactance is cancelled out, is associated with an increase in the loop perimeter. If one does so by increasing the lengths of the three radiators, the R_{in} increases. If one does it by lengthening the primary loop or the distance between the radiators, the R_{in} stays relatively stable while the gain increases.

Small secondary loops reduce losses and increase the gain by an amount equal to the loss reduction. Making the secondary loop larger—until it equals the size of the primary—results in further increases in gain. Most importantly, the addition of the second loop substantially increases the system bandwidth even as it increases the gain. This all comes at a price of a slight increase in antenna complexity.

The presence of the two loops and their individual resonances can be advantageous in designing extremely broadband antennas that can cover large segments of the HF/VHF spectrum. This is extremely important in creating broadband-loop parasitic arrays; those will be the subjects of a later article.

Lastly, we can create antennas that enable instant QSY between segments of a band as wide as 80 meters. This can be done using two ADR center wires and alternately shorting one and opening the other with relays. This too will be the subject of a future article.

Although I have not gotten into 160-meter designs, there is no reason why an ADR won't perform well on the top band. It will have more gain than any of the common verticals and inverted-Ls and have a wider bandwidth.[25] You just need about 200 feet room for it, or more.

Acknowlegements

I thank L. B. Cebik, W4RNL; Darrel Emerson, AA7FV; David Jefferies, G6GPR; and Grant Bingeman, KM5KG, for their help and infinite patience.

Notes

[1] D. Handelsman, N2DT, "The Rectangle Family of Antennas, Part 1: The Not-So-Simple Rectangle," *QEX,* Mar/Apr 2001, pp 35-46.

[2] See Note 1, p 40.

[3] See Note 1, p 39. The aspect ratio is the ratio of the radiator separation to the radiator size.

[4] See Note 29 (p 46) in the article of Note 1, above.

[5] S. Kinoshita, JF6DEA/KE1EO, "The Hentenna: The Japanese 'Miracle' Wire Antenna," *ARRL Antenna Compendium,* Vol 5, p 66, ARRL 1996. The antenna was developed in 1971 by JE1DEU and JH1FCZ.

[6] P. Dodd, G3LDO, "The HF Skeleton Slot Antenna," *ARRL Antenna Compendium*, Vol 6, p 70, ARRL, 1999. This is a multi-band 10 to 30-meter version of a SDR. The author goes into the history of this antenna in detail. Also see Lew Gordon, K4VX, "The Double Magnetic Slot Antenna For 80 Meters," *ARRL Antenna Compendium*, Vol 4, p 18, ARRL, 1995. The dimensions of this antenna correspond to those of a peak gain SDR composed of #6 AWG wire. The antenna is mixed polarized and fed at a corner. Also, Paul Carr, N4PC, "The H-Double Bay Antenna," p 28, *CQ*, September 1995. This antenna is a horizontally polarized, vertically oriented, 30-meter symmetrical double rectangle which is fed at the center of the lower end.

[7] Dan Handelsman, N2DT, "The Double Rectangle: Three variations on a rectangular theme," *Communications Quarterly*, Winter 1999, pp 67-89.

[8] D. Handelsman, N2DT, "The Double Loop on 30 Meters," *antenneX* (Web publication at **www.antennex.com/**), February 2000. This band is the point in the HF spectrum where—at attainable antenna heights—a horizontally polarized loop will outperform a vertically polarized one at the lowest TOAs.

[9] These antennas produce mixed polarization when fed at the corners.

[10] After all, we have added a third radiator to the system and we expect some return for it.

[11] There is a minuscule decrease in wire losses resulting from a current division between the closely spaced near and center radiators.

[12] All modeling for this article was done with *NEC Win Plus* by Nittany Scientific. This is a version of NEC-2 for *Windows*. Some of the models were kindly checked out by L. B. Cebik with NEC-4. All of the antennas were modeled with segmentation densities designed for convergence of results. The smallest segment size was 0.001 λ for thin-wire models and four times the wire thickness for thick-wire models. A segment density of 50/λ was found reliable and computed absolute gains are believed to be accurate to 0.1 dB or better.

[13] This is exactly what happens with the simple rectangles; see Note 1.

[14] I did some further studies with this 115/110/5 ADR. The feed-point *R* was about 73 Ω at the center wire but with less gain. The near-end wire had an R_{in} of 141 with a f_{res} of 3.498 MHz. The bandwidth at all three wires was identical. At this particular configuration, it may be more useful to feed the antenna at the near wire since matching is easier.

[15] This is due to the higher *Q* of the thicker wire.

[16] D. Handelsman, N2DT, "Loop Antennas Over Ground," *antenneX*, May 2000.

[17] See Note7, p 75.

[18] This is well-known in Japan with the Hentennas.

[19] With ADRs, this appears to be virtually linear. This linearity is lost when the center wire nears the geometric center because much greater movement is necessary to shift the frequency upward.

[20] See Note 1, p 38. I am grateful to Darrel Emerson, AA7FV, for pointing out the way to analyze the SDRs. The mutual impedance and network analyses were simplified by Grant Bingeman, KM5KG's, *RF Designer* software. This can be found at **www.qsl.net/km5kg**.

[21] See Note 10.

[22] Be careful. The overlapping contiguous resonances only occur when you feed the far wire. With these dimensions and if you vary the feed point, the center-wire R_{in} is about 400 Ω and the near-wire *Z* is about 600 Ω (negatively reactive)—it cannot be tuned to resonance with thick wires. Meanwhile, although two distinct resonances are displayed, the central SWR hump is over 4:1 and you lose the great bandwidth. Moreover, the gain at the center is significantly lower than at the far end.

There are still mysteries with the thick-wire ADRs/SDRs. As expected, the loop perimeter increases with increasing wire diameter (as with simple quads). When the wire diameter passes a threshold along with the loop perimeter, the antennas reach a stage where they cannot be resonated by dimensional changes at all. In fact, decreasing the radiator size is accompanied by an increase (positive) in the reactance. In symmetrical antennas such as the SDRs fed at either end wire, this inability to resonate the loops is associated with a pronounced "squint" in the elevation angle in free space. This indicates a propagation delay and a progressive phase shift as the signal moves from the feed point. I would appreciate any suggestions as to how to approach this problem.

[23] We have about 65 MHz of low SWR above the point where the gain drops to 3 dBi. What happens to the gain? The radiators become so greatly separated (in terms o f λ) that there is more end-fire gain (upwards) than broadside gain (along the bore sight).

[24] See Note 5. The performance of vertically polarized ADRs over ground was fully discussed there.

[25] I will be happy to supply the NEC files for any of these antennas to any interested party so that they could be used, with proper scaling, for any band. You can download this package from the ARRL Web **www.arrl.org/qexfiles/**. Look for 0102HANDELSMAN.ZIP.

The Quad Antenna Revisited Part 5: Quad Design Variations

Hams use many kinds of loop and loop-array antennas. Part 5 of this series looks at some of the published variations on quad-loop antenna design. It includes some general thoughts on additional variations.

By R. P. Haviland, W4MB

Supporting the Quad

Every antenna needs a support. If the support happens to be resonant at or near the operating frequency, there can be interaction between the support and the antenna. The "innate perversity principle" seems to apply: The interaction always seems to degrade antenna performance.

As an example of this, the X-Quad curve of Fig 1 shows the gain of a quad loop supported by a metal ×-frame made of two tubes λ/2 long at the design frequency of 37.5 MHz (1.0 on the relative KB scale). For example, this corresponds, very nearly, to a 15-meter quad mounted on metallic

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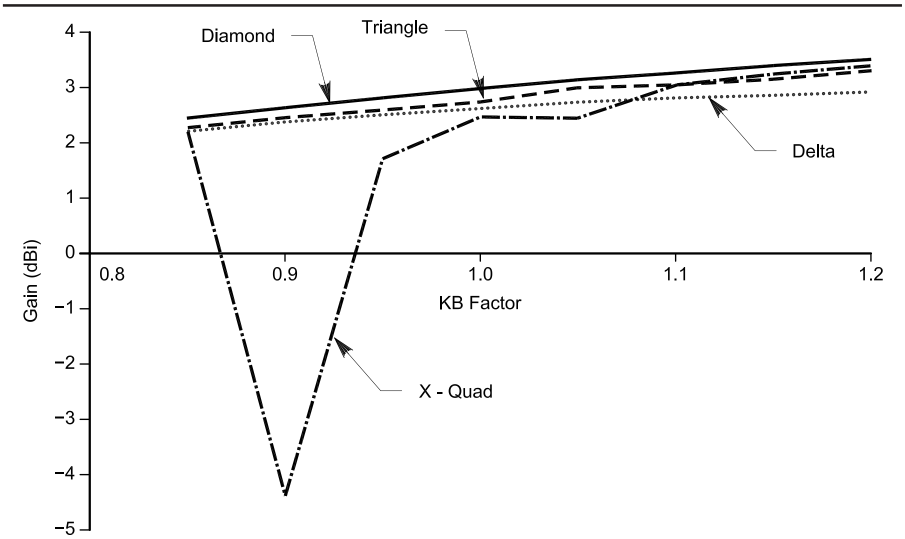

Fig 1—Gains of four loop shapes.

supports for a 20-meter quad. A tube of the size used is resonant at about 0.45 λ, or at 0.9 on the KB scale. As seen in the figure, the ×-frame causes a marked decrease in on-axis gain at this frequency, about 7 dB for the design values used. Physically, the ×-arms acting as dipoles are parasitically excited by the quad loop: they extract energy from the on-axis lobes and radiate this in the plane of the loop. While this may be desirable if nearly omnidirectional (isotropic) radiation is the goal, it can be a real drawback in normal quad design.

I recall having seen (but can't now find) an advertisement using this design, with tuning of the ×-frame as director, radiator or reflector, depending on the purpose of the supported quad loops. Both a broad SWR bandwidth and an increase in gain of one decibel were claimed. I have spent much time on many variations of these possibilities, but I have not been able to confirm the gain increase: Instead, I find a loss of about one decibel or more for any combination tried. Broadening of the SWR bandwidth is possible, but it is usually accompanied by a loss of gain.

On the other hand, I have used the design "on the air" for long periods, in two and three-element quads designed for 10 through 30 meters,

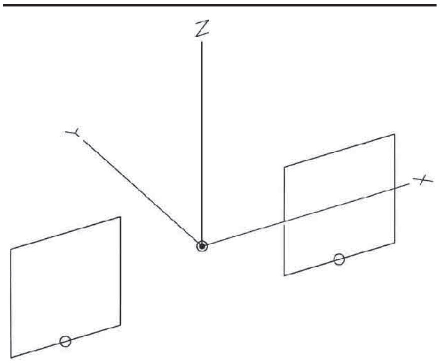

Fig 2—Two quad loops "stacked" side-by-side to increase gain.

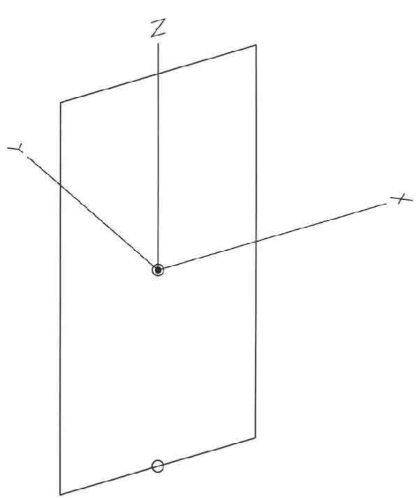

Fig 4—The 3-λ quad, an example of an *N*-wavelength loop with its shape adjusted for lobe reinforcement.

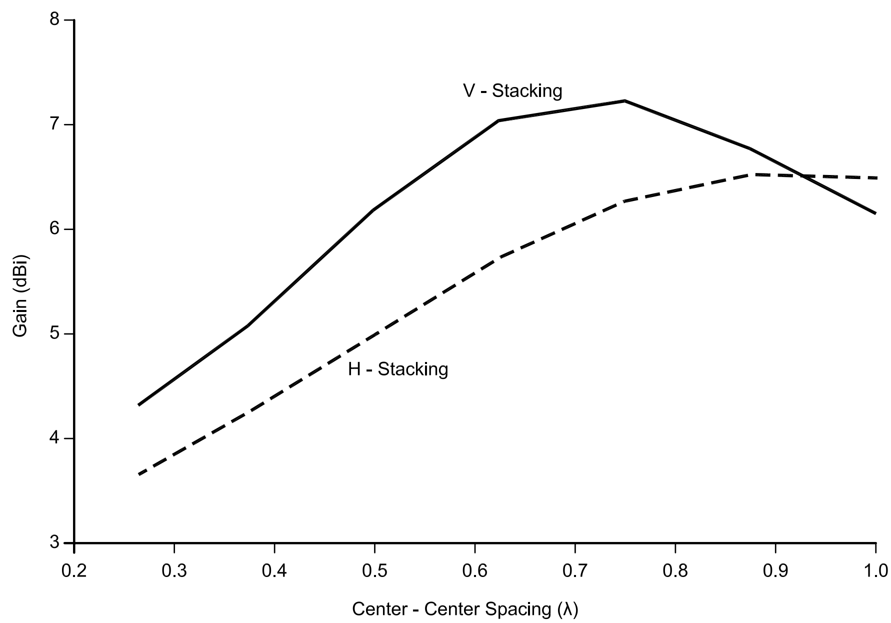

Fig 3—Gain of stacked quads as a function of center-center spacing for 1 λ loops and 1-mm wire diameter.

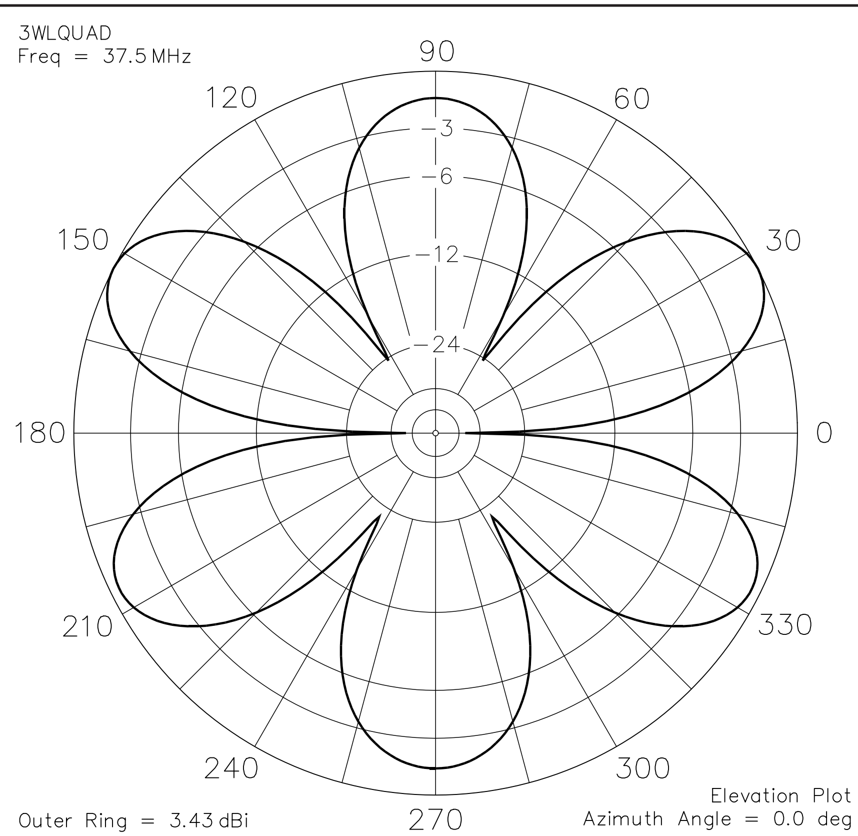

Fig 5—Elevation pattern of a 3-λ quad in free space.

with the supports tuned for 15 meters. The performance of the three-element design (0.45 λ director, 0.47 λ radiator, 1.0 λ reflector) was at least as good as a design with nonconductive arms. However, The performance of the two-element unit was mediocre.

The recommendation here is to use either nonconductive arms (eg, fiberglass) or tune the arms well away from the operating frequency range, by at least 10%. With a multiband design, the geometric mean frequency between adjacent bands is a possibility. If a multi-element design is the goal, it might be worthwhile to spend some time with one of the *NEC* analysis programs. However, be aware that the

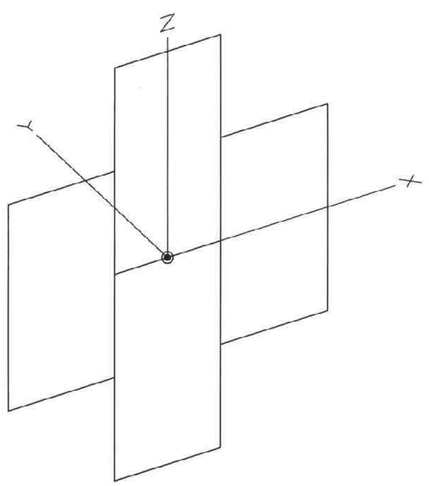

Fig 6—An array of four 3-λ loops.

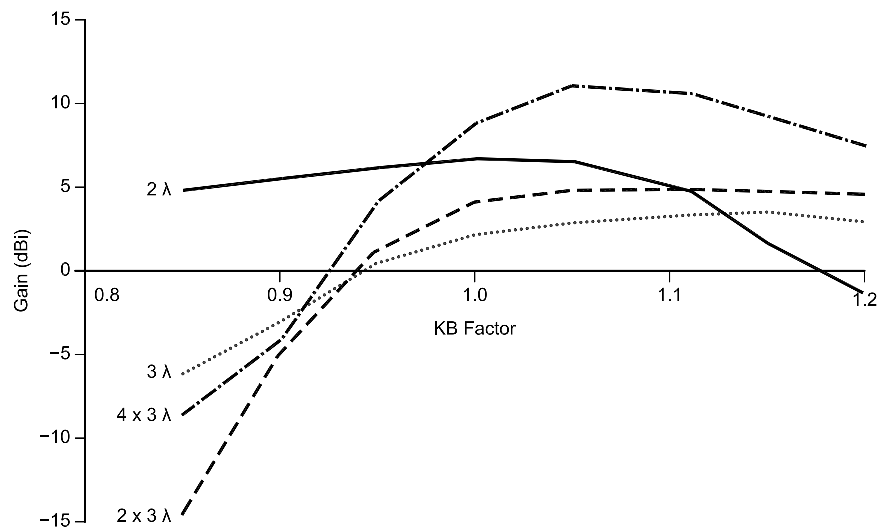

Fig 7—Gain curves for several *N*-wavelength designs (designs shaped for gain).

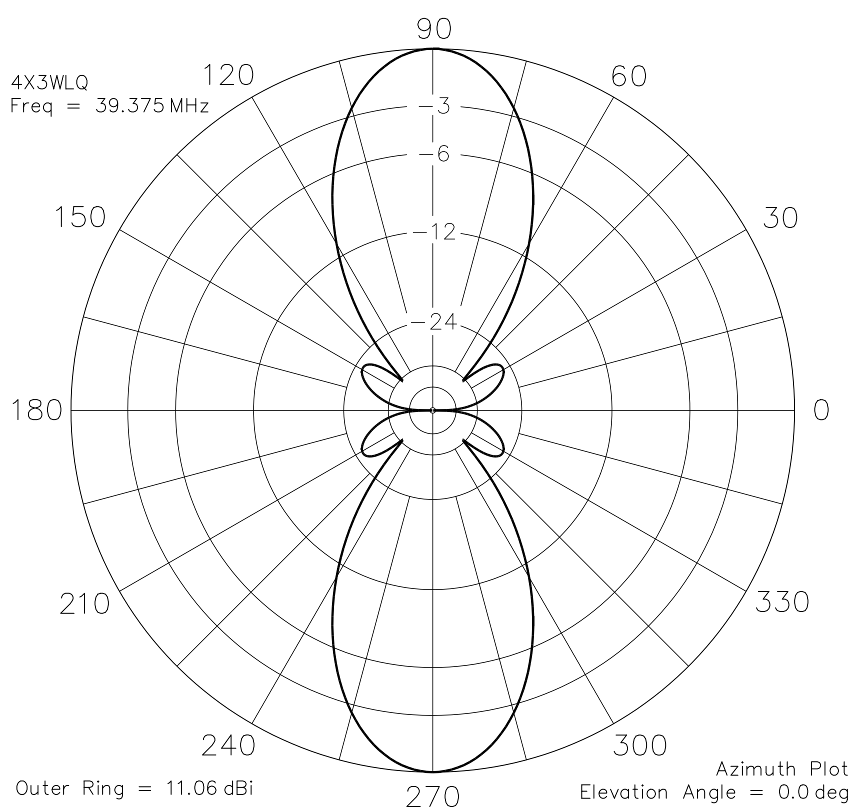

Fig 8—Azimuth pattern for an array of four 3-λ loops as shown in Fig 6.

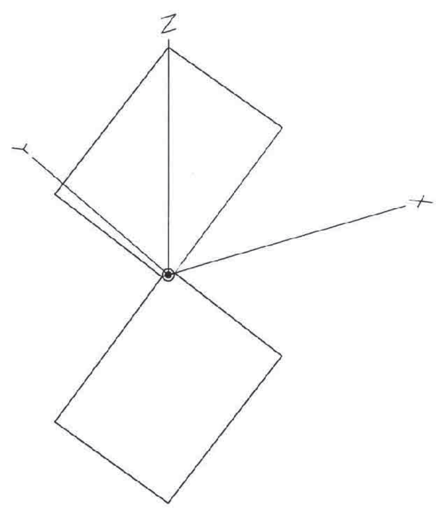

Fig 9—The "Dopple" quad, two stacked loops connected to give a current-feed interconnection.

close spacing of the ×-frame and the loops required in practical cases makes analysis a problem. In particular, the segment lengths must be much less than the spacing between support and antenna. If you try such analysis, remember to increase the number of segments until the results converge.

Some Additional Loop Shapes

Fig 1 includes gain curves for three additional loops. The diamond is the usual type, with equal arms; it is a conventional square quad rotated 45°. Since the loop area is equal to that of a quad loop, the gain is essentially identical: Small differences in gain (and impedance) occur because the distances between equal-current points are not identical.

Triangles (deltas) have equal-length sides for their three arms. There is a reduction in area, as well as greater differences in the separation of equal-current points. These factors reduce the antenna gain, but the reduction is not great, and good antennas can be built with these shapes. Oriented with a horizontal side uppermost, a delta loop reduces the mechanical problems of conventional quads. Typically, there are two side arms of tubing supported by a boom, and the top wire is under enough tension to slightly bow the sides. By contrast, a triangle with a horizontal side at the bottom is suited to 40 through 160-meter operation because it needs only a single elevated support. This support might be an available tree, a mast supporting a Yagi, a house or even a lighthouse on Field Day. (Be careful to stay away from power lines, however).

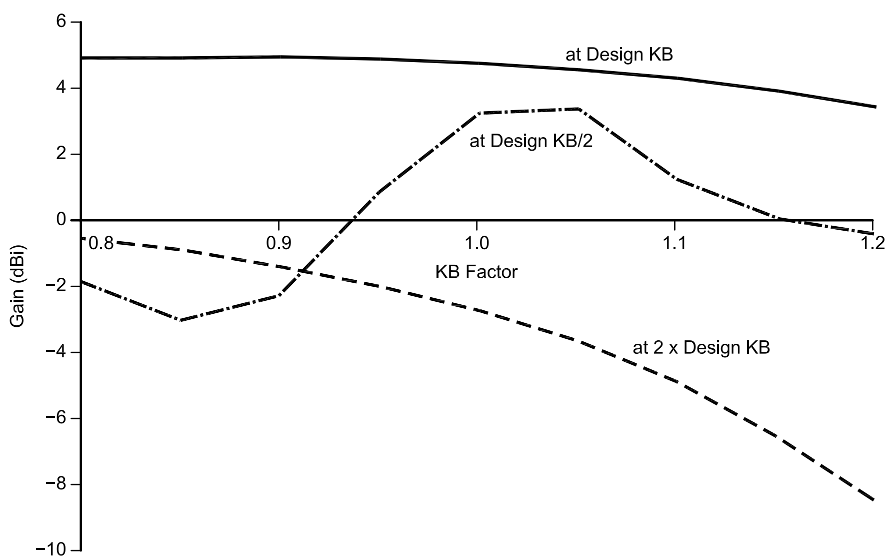

Fig 10—Gains of a Dopple quad at its design wavelength (KB), twice and half this wavelength.

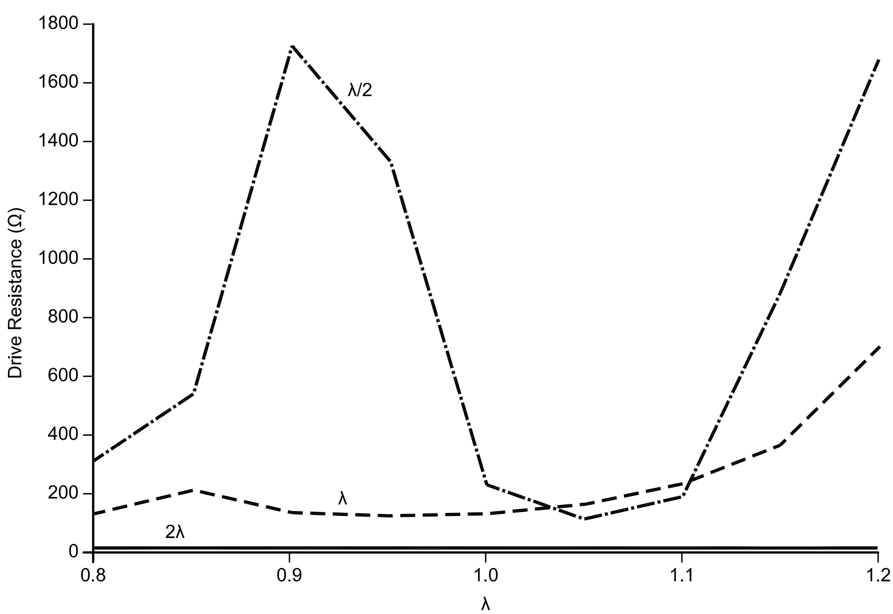

Fig 11—Drive resistance of a Dopple quad at 3 values of wavelength.

Stacking Quads

One way to increase antenna gain is to place two or more antennas side-by-side or stacked as sketched in Fig 2. They are shown horizontally stacked with bottom-center feed. This is a common practice, but the feed point can be at other locations. Side feeding is equivalent to vertical stacking.

Compared to dipoles, the quad's "capture area" or area of equivalent radiation, is larger than its physical area. This means that there is a loss of effective area when the antennas are close together, because the effective areas overlap. This is clearly shown in Fig 3, which depicts the gain versus spacing for horizontal and vertical stacking. The best center-to-center distances are 0.7 and 0.9 λ for vertical and horizontal stacks, respectively. Increasing distance beyond the values shown at first reduces gain, then in-

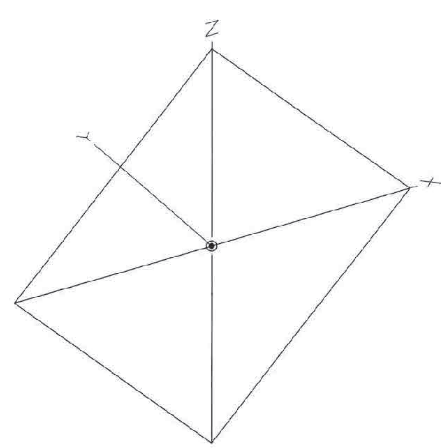

Fig 12—The cross-fed quad.

creases it but it remains less than optimum. At great distances, the gain is exactly twice that of an independent antenna, 3 dB greater. The best spacing distance changes if an array of loops is used rather than a single loop. Best spacing can be determined with one of the *NEC*-based analysis programs.

Combined horizontal and vertical stacking can yield gains somewhat greater than the sum of the element gains plus 10 log(N)dB, where N is the number of elements in the array. We can reasonably approximate the patterns developed as the pattern of an individual element multiplied by an "array factor" developed from the spacing and the phase relationships between elements. (See, for example, Kraus, 1988, for these factors.)

N-Wavelength Quads

One of the characteristics of a conventional square quad loop is that the on-axis gain becomes zero at the second and higher harmonics of the design frequency. There have been many designs to find a way around this, to get the higher gain possible because of the increase in area as measured in wavelengths.

It appears that the first such method was to use a diamond loop opened at the top, operated at a frequency where the loop circumference is 2 λ. In that case, the current relations give vectors that add on-axis. The original designs used two loops perpendicular to each other, fed at the bottom corner and supported by a single pole. The assembly is called a "Bi-Square Beam." Characteristically, the signal is vertically polarized. They were originally used extensively on islands in the Pacific just after WW2. This was okay, but if the ground is poor, ground loss reduces the effective gain compared to that with horizontal polarization. The design is rarely used today, but it deserves to be more popular for island locations and areas such as rice paddies, which are good grounds. The single-pole support and boom-less construction means that the antenna can be built of local material at very low cost.

A 3-λ design requires changing the shape of the loop, usually as shown in Fig 4. The longer top and bottom elements and greater separation give greater gain than a 1-λ loop. The currents in the vertical arms produce strong lobes at some angles, however, as shown in Fig 5.

Three-wavelength loops are usually used in a stacked form, as shown in Fig 6. Note that the loops are current

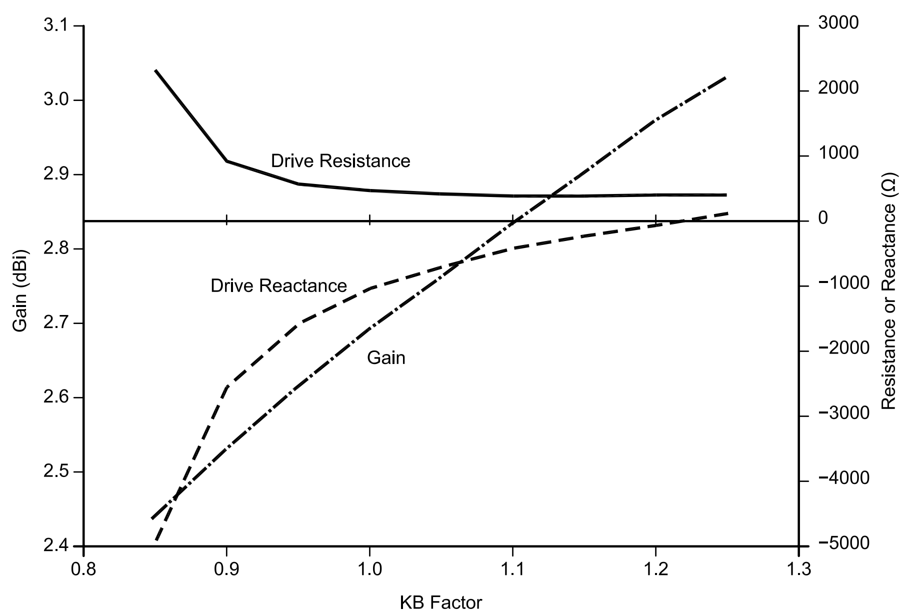

Fig 13—Gains and impedances of a cross-fed quad across range from 0.8 to 1.3 times its design frequency.

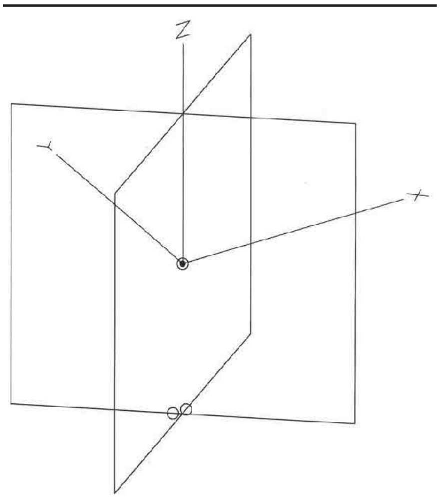

Fig 14—The "Bird-Cage" quad.

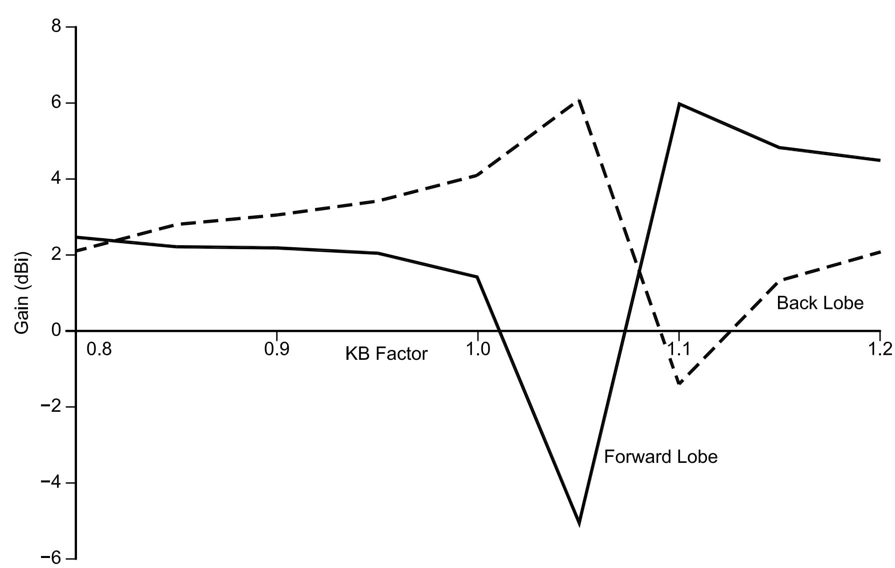

Fig 15—Lobe gains of the Bird-Cage quad.

fed in the two center loops; this form sometimes used alone. In contrast, the outer loops are voltage fed from the center loops. More loops can be added; for example three vertically by five horizontally, or more. (See Kraus 1988 for further information.)

The on-axis gain of this family of loops is shown in Fig 7. Because of the large side lobes, the basic 3-λ loop has less on-axis gain than a 2-λ version. There is also very little gain increase from stacking two vertical loops. However, there is side-lobe cancellation in the four-loop form, which gives the good gain shown. The lobe structure is shown in Fig 8 and it is very good. Used with a reflector screen or a parasitic element, gain will approach 17 dBi. SHF users would find the combination of feed simplicity and good performance worthwhile.

The *Dopple* Quad

The structure of the *Dopple* Quad (*dopple* is German for double) is sketched in Fig 9. It is formed by two 1-λ loops connected together and fed at the connection. Notice that analysis must use a short of three or more segments across the feed point with a current input at the center or a combination of four current feed points, one for each wire. If one of these artifices is not used, the pattern will not be correct.

The dopple quad features on-axis gain at ½, 1 and 2 times the design frequency, f_0. This can be seen by considering the geometry. At f_0, we have two loops stacked vertically. At $f_0/2$, there is 1 λ of wire, in essence a diamond loop with the outer corners pulled toward the center. Half of the area is lost, so the gain is less than for a full-sized quad. At $2f_0$, the loops act as stacked open 2-λ voltage-fed quads. Drive impedance for a center feed point will be high, but it is reduced by the fact that the two loops are connected in parallel.

The gains at the three KB factors of

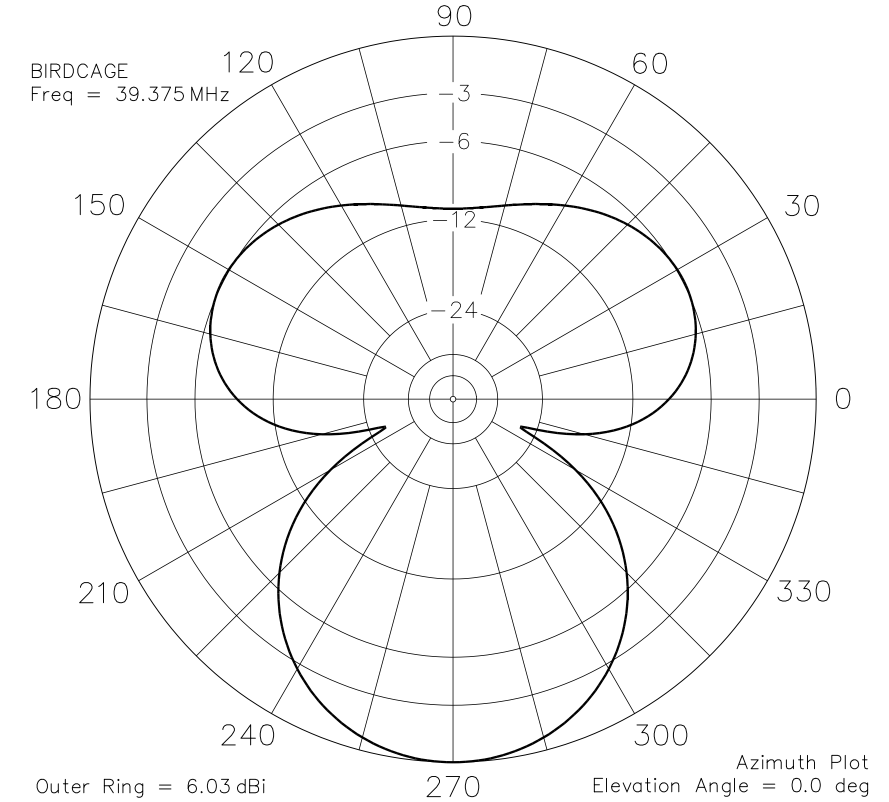

Fig 16—Azimuth plot of a Bird-Cage quad free space.

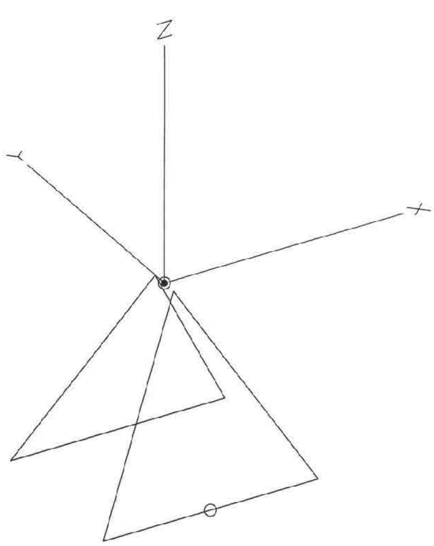

Fig 17—A pyramid antenna.

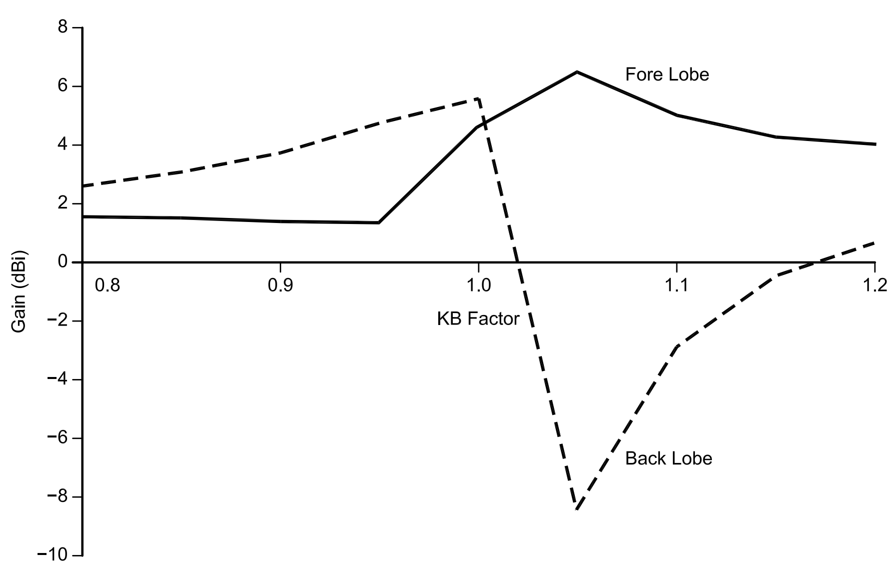

Fig 18—Lobe gains of a pyramid antenna in free space across a range from 0.8 to 1.3 times the design frequency.

0.5, 1 and 2 are shown in Fig 10. The maxima do not occur at exactly one-half or twice the nominal best-gain frequency, f_0. This is characteristic of antennas and such resonators as crystals. The best term to describe their use is "overtone" operation.

Fig 11 shows the variation in drive resistance. It is small at the $f_0/2$ value, because the antenna is small in terms of wavelengths. The $2f_0$ resistance is variable. The drive-point reactance is also variable, so low-loss line and an antenna tuner is needed if the multiple resonances are to be used.

The ½-KB curve illustrates a point in quad design. Bringing the corners near to contact increases the capacitance between them compared to that when they are separated. The added capacitance lowers the resonant frequency below the 1-λ value (in this case). Any addition that increases the capacitance between the high-voltage points reduces the resonant frequency; it is a form of loading. Inductance can also be added at high-current points to secure frequency reduction. For example, loading coils can give 40-meter resonance in a 20-meter-size quad. Gain will be lower than that of a full-sized loop, but worthwhile small designs are possible.

The Cross-Fed Quad

Sometimes called the arm-fed quad, this design uses the support arms as a transmission line, connecting the feed point to the quad loop. This is done by insulating the four aluminum tubes used as support at the center and connecting the outer ends to the quad loop at its corners. The inner end of opposite tubes is used as the feed points. The arrangement is shown in Fig 12 for horizontal polarization. Notice that the other pair of tubes is not connected at the center. Switching the feed to these gives vertical polarization. The arrangement was developed for the citizen's band, where vertical polarization is used for contacts to mobile stations and horizontal is common for other uses such as "skip," often sporadic-E propagation. The common arrangement is an array of a reflector, radiator and two directors.

Because of their length, the arms used as the feed line transform the quad's usual 100 to 150 Ω drive impedance to a high value and introduce reactance. Fig 13 shows the effect. The on-axis gain increases over the KB range shown but is a little lower than for the quad loop itself, because the fed arms radiate as shortened dipoles at right angles to the quad-loop radiation axis. If the antenna is perfectly symmetrical, there is no current on the unconnected support arms. Notice that the feed-point reactance is zero well above the quad-loop resonant point, around frequencies where KB = 1.05. A coiled length of transmission line is used to provide matching to the feed coax in the usual citizen's-band design.

This antenna could be used to pro-

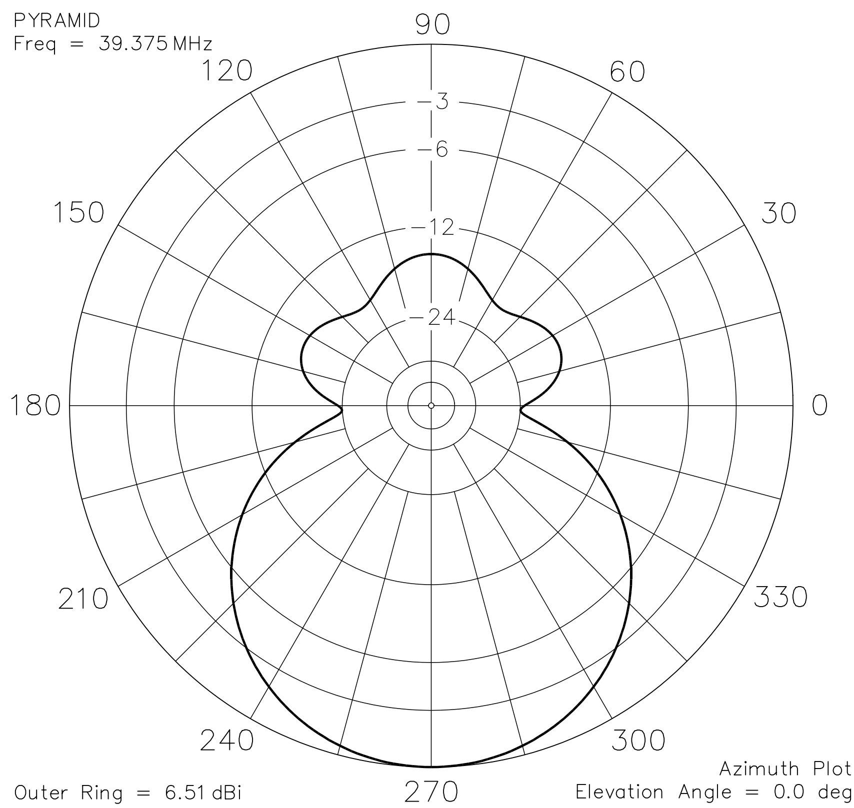

Fig 19—Azimuth pattern of a pyramid antenna in free space.

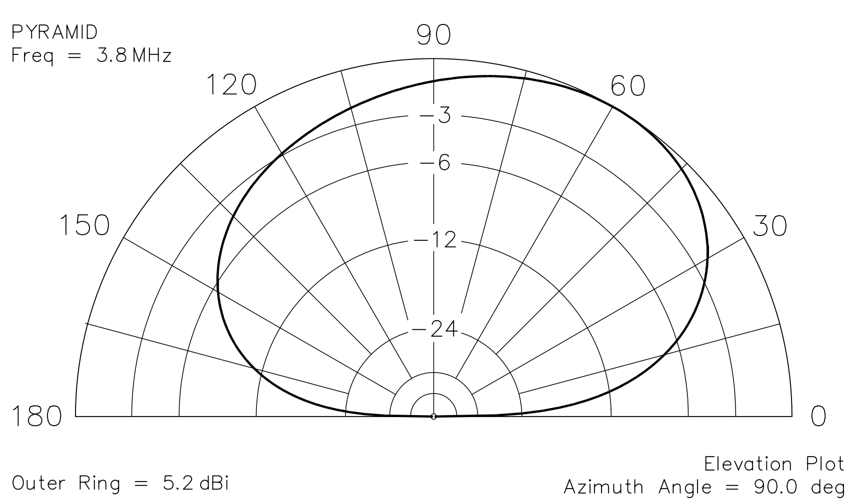

Fig 20—An elevation pattern of a pyramid antenna with 10-foot ground clearance. This plot is at the maximum-gain angle.

vide an interesting and possibly worthwhile experiment in polarization-diversity reception. Normally diversity reception involves the physical separation of antennas. In at least some and probably many propagation conditions, however, the observed fading is polarization loss as the axis of polarization is rotated by the ionosphere. Experimentation is in order.

The Bird-Cage Quad

In addition to changing quad shape by pulling the arms toward the center as in the Bat-Wing quad, they can be pulled at right angles to the plane of the elements. Therefore, resonance points will change some. Drive resistance and gains will decrease, in consequence of the reduced area.

G4ZU used two of these deformed loops in a back-to-back arrangement to create a directional antenna, shown in Fig 14. In the original arrangement, the top and bottom ×-arms were made of aluminum tubing and the vertical elements were wire. Here, all sections are modeled as wire.

A typical performance curve is shown in Fig 15. Gain can reach 6 dB above isotropic. The F/B ratio in this configuration is not especially good, about 11 dB over a narrow band of frequencies. A typical pattern plot is shown in Fig 16. There is no attempt here to find the best design point for this configuration.

The Pyramid Antenna

A pyramid antenna consists of two apex-upward triangular loops positioned close together at the apices and well separated at the bottom, as shown in Fig 17. A major advantage of this arrangement is that the antenna can be supported by a single mast. Additionally, there is some pattern control by choice of feed point. In free space, the antenna is a reasonably good performer, as shown in Figs 18 and 19. Gain reaches just over 6 dB, with an F/B ratio of about 15 dB.

Over reasonably good earth, the antenna gives good performance, even if the lower part is only 10 feet above ground as in this example. A typical vertical-plane pattern is shown in Fig 20, with maximum radiation at 58° elevation. The signal is at isotropic level at an elevation of 19°. The F/B performance of the antenna is not especially great, usually less than 10 dB.

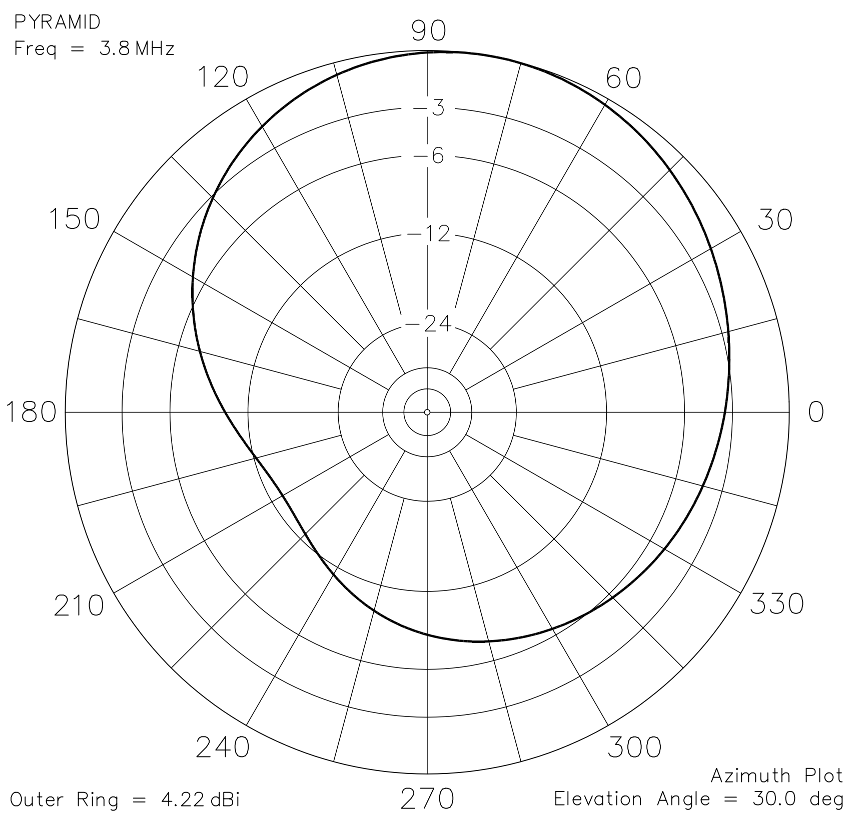

Fig 21—Azimuth plot of the pyramid antenna at 30° elevation, bottom-corner fed. Notice how this pulls the pattern to one side.

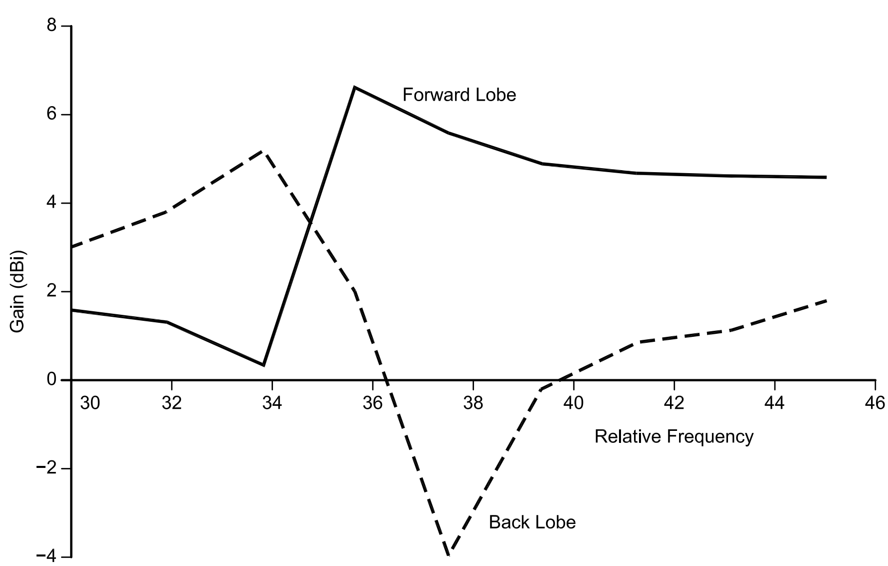

Fig 22—Lobe gains of a "Quagi," a quad loop plus a dipole.

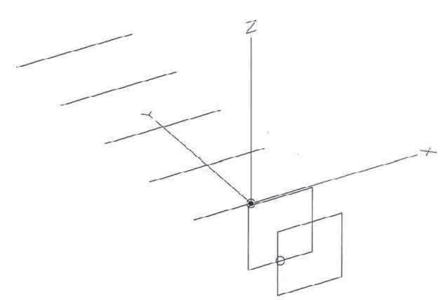

Fig 23—A Quagi with a quad radiator and reflector plus five dipoles tuned as directors.

An attractive feature of the pyramid antenna is that the pattern can be pulled to the side by moving the feed point to a corner, as shown in Fig 21. Very good coverage in all directions can be secured by arranging a switched feed to all four corners. Some three S-units of signal-interference improvement is possible.

As sketched here, the antenna would require a 90-foot tower on 80 meters. Lower tower heights are possible, and experimentation with these loop sizes should be rewarding in terms of low-band DX worked.

The general guideline for these antennas is to keep the loop length at, or close to, one wavelength. This can be done by changing the triangle shape and by increasing the bottom separation. However, the average separation of the two loops should be in the range of 0.1 to 0.2 λ. In the example, the loop tops are shown separated. This is easily done with a short boom clamped to the tower. This is not necessary, but helps to keep the average separation more nearly constant.

The Quagi

Early experimenters with quads found that the performance is essentially unchanged if the shape of the directors is changed from loops to disks or even to rods. The important factor is that the shape displays a resonance just above the operating frequency. (See Appel-Hansen, 1972, for an early discussion). The finding led to the combination of a two-element quad with a set of Yagi style directors called the Quagi.

Fig 22 shows the lobe gains of a single 1-λ loop spaced 0.2 λ from a λ/2 rod, a Yagi element. The variation follows the same pattern as for two loops, with maximum forward gain at one frequency, and best F/B ratio at a higher one. The F/B ratio is not especially high with this combination of lengths and spacing, about 10 dB maximum, but it can be improved by changing the relative size of the rod.

This is done for the antenna of Fig 23, with two 1-λ-perimeter loops and five rods of 0.45 λ. The reflector loop space is 0.2 λ, with the director rods spaced 0.3 λ. Lobe performance is shown in Fig 24. Gain is nearly 13 dBi (free space), but the F/B is not especially good, less than 16 dBi maximum. It also shows a change that indicates that the element tuning could be improved, by change in element size and spacing. (Unfortunately, all of the common antenna-optimization programs are set up for Yagi elements only, so this can take some effort.)

The pattern of this antenna is shown in Fig 25 at two frequencies. The main lobe beamwidth is only 33°. As always occurs with multi-element antennas, there are minor lobes that change size as frequency is changed. Some control of these is possible by changing the element size and spacing, but they will

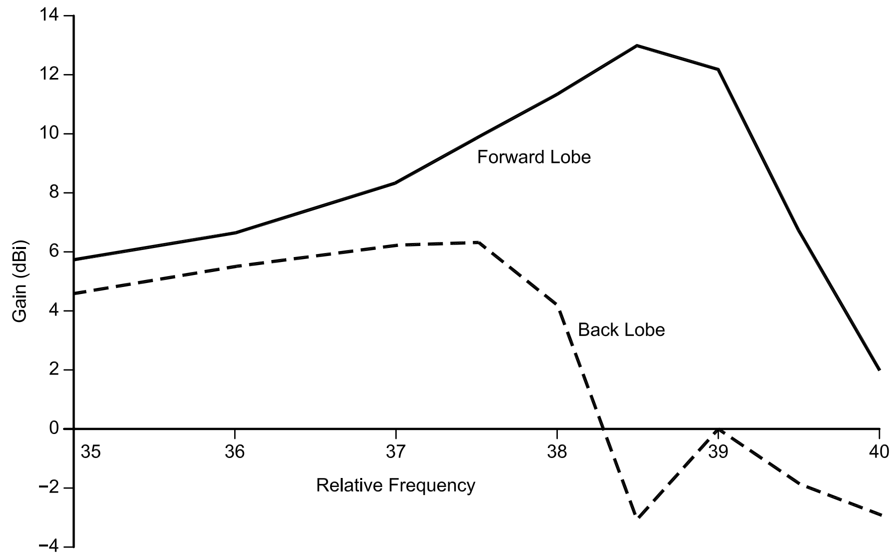

Fig 24—Lobe gains of the seven-element Quagi.

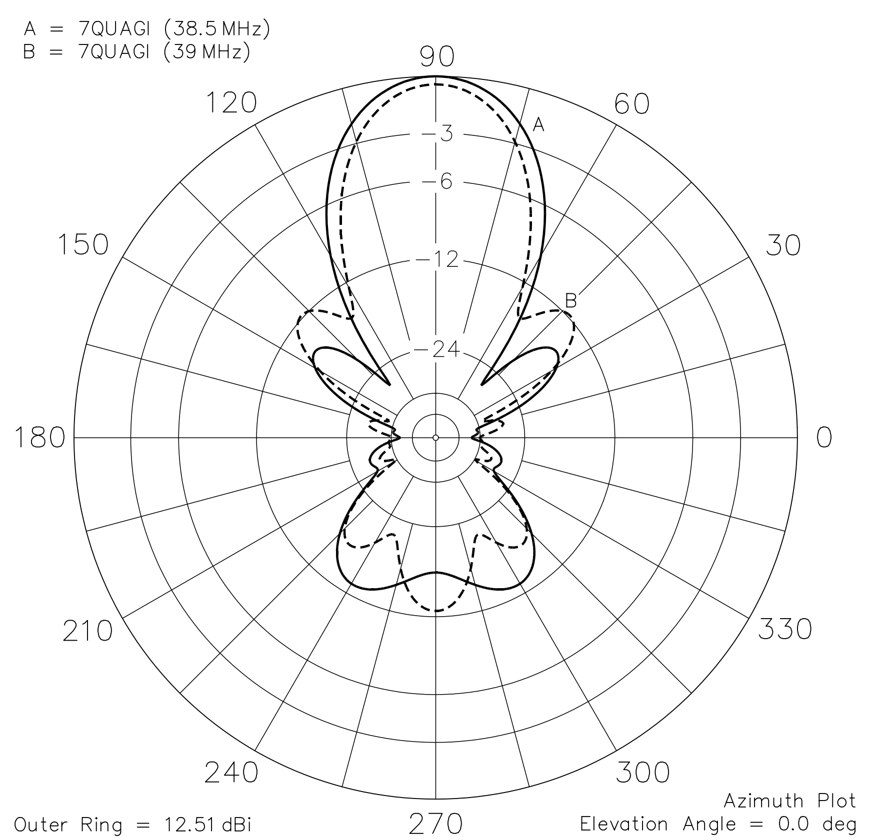

Fig 25—Typical patterns of the seven-element Yagi. At A, 1.026 times nominal design frequency; at B, 1.040 times nominal design frequency

always be present at some level.

Old timers familiar with 20 meters may remember an outstanding G-land signal from Folkstone. The antenna used had two quad loops, plus two triband trap directors taken from a conventional multiband Yagi. I think more experimentation is needed.

Closing Comments

The antennas described here by no means cover the complete range of possible shapes. Any geometric figure with a perimeter approximating 1 λ is a possibility. Providing the shape has a reasonable area, it will have useful gain in some direction and will not be too difficult to drive. Experimentation is in order, preferably first in one of the *NEC*-based analysis programs, then on the air.

As a low-band possibility, consider joining tower guys by means of a horizontal wire some 8 to 10 feet above ground and fed at some point. Also try clamping some split ferrite interference filters around the guys just below the horizontal junction. This can be modeled by introducing a reactance load in the guy model.

References

J. Appel-Hansen, "The Loop Antenna with Director Arrays of Loops and Rods," *IEEE Transactions on Antennas and Propagation*, July 1972.

H. K. Bourne, ZL1OI, "A Minature Quad Loop Antenna for 15/20 Meters," CQ, Mar 1978. The 20-meter element is 13 feet 9 inches on each side. Capacitive loading is provided by a 12-foot wire spaced 5 inches from the loop sides. It is center connected to the loop-side center. The 15-meter element is full size.

J. W. Fisher, W8JKN, "A Log-Periodic Quad Array," *CQ*, Feb 1977. An experimental antenna uses four loops. For its 6-MHz lowest frequency, the largest loop perimeter is 167.5 feet. The first two elements are spaced 10 feet apart. The ratio of spacing and element size is 0.85. Design data for 11 elements covering 2-30 MHz is included.

R. P. Haviland, W4MB, "The Quad Antenna Revisited, Part 4: Effects of Ground on Quad Loops," *QEX*, Mar 2001, pp 47-54.

"The Quad Antenna Revisited," *Communications Quarterly*, Pt 1, Summer 1999, pp 43-73; Pt 2, Fall 1999, pp 65-85; Pt 3, *QEX/Communications Quarterly*, Nov/Dec 2000, pp 10-19. A multitude of *EZNEC* description files used in this series is available for download from the ARRL Web **www.arrl.org/files/*qex*/**. Look for QUADS.ZIP.

J. D. Kraus, *Antennas* (New York: McGraw-Hill, 1988).

K. Kunde, "Try a 'Dopplequad' Beam Antenna for 2 Meters," *QST*, Feb 1985, p 28.

J. Ross Anderson, "Meet The Curtain-Quad Antenna," *QST*, Nov 1984, p 48.

J. P. Tyskewicz, W1HXU, "A Low Profile, Three-band Quad, MK IV," *CQ*, Dec 1976. This combines reduced height and a bottom-center loading coil with diamond loops. A 20-meter loop is 12 feet high by 24 feet wide.

R. P. Haviland was first licensed as W9CAK in 1931. He is a retired Professional Engineer specializing in space communication and Fellow of the IEEE who holds a BSEE, BIS and AAS. He was a project engineer for the first rocket to transmit from beyond the ionosphere.

Out of the Box: New Products

NEW RF POWER MOSFET's

Advanced Power Technology has announced four new devices in their line of RF power MOSFETs: the ARF462A/ARF462B and ARF464A/ARF464B. Unlike previous parts in this line, these were designed with linear as well as nonlinear applications in mind. The ARF462 is rated at 150 W output up to 60 MHz. The ARF464 is rated at 100 W output up to 100 MHz. The ARF460 through ARF464 are all rated for linear RF operation.

These parts share two important mechanical features with the other devices in the line. First, they have the source connected to the heat sink tab on the TO-247 plastic package. This allows the part to be attached directly to the heat sink without an insulator. The inductance in the source lead is reduced, so greater bandwidth is possible. The second mechanical feature is that the A- and B-suffix parts have connections that are mirror images of each other. This allows better layout of push-pull amplifiers by putting the input leads on one side of the source connections and the output leads on the other. I appreciate the significantly greater ease of mounting the TO-247 versus a RF package like the 211-11. It is easier to get one hole drilled and tapped in the correct place than two.

These devices were designed for better operation in high-SWR applications. Previous parts in the line had a design V_{dd} of 50 V and a breakdown voltage of 125 V. The ARF462 and ARF464 have a design V_{dd} of 65 V and 200 V breakdown rating. This allows additional headroom in 50 V systems. They have also increased the safe operating area for these devices. This makes the parts more rugged for instances when someone forgets to put the load in that RF heater or leaves the feedline disconnected from the amplifier.

The preliminary data sheet gives data for nonlinear applications, but does not address IMD performance. These devices are primarily intended for high-power, single-frequency industrial RF generation, so they have not yet been characterized for communication applications.

The die of these parts is roughly twice the size of the MRF 150, so the gate capacitance is also about doubled. The gate bond-wire inductance is about four times larger. The real advantage of these parts over the MRF series is the significantly cheaper package and the increased mismatch handling.

The large input capacitance requires a low resistance in parallel to minimize the impedance change versus frequency in untuned applications. A broadband amplifier would likely require a value lower than the 25 ý that was used to characterize the device for the datasheet. Additionally, the input impedance changes from capacitive to inductive around 50 MHz because of the inductance of the source and gate bond wires.

Production quantities will be available in the fourth quarter of 2001. APT's plastic-packaged RF power devices are lower in price than conventional ceramic-packaged devices. In quantities of 1000 pieces, the unit price for ARF462 is $24; the ARF464 is $18. They are available through the factory and all authorized APT distributors including Richardson Electronics. In October, Richardson did not list these parts as stock, but they should be available by early 2002. Using other APT parts as a gauge, these are likely to be in the $35 to $40 range in single quantities. Preliminary data sheets are available to assist designers, describing the features and benefits of the new MOSFETs. Data sheets and *SPICE* models may be downloaded from APT's Web site at **www.advancedpower.com** or obtained from the factory.—*Contributing Editor, Ray Mack, WD5IFS;* **rmack@arrl.org**

Sound-Card Antenna Measurements and Other Useful Techniques

Here is a technique to measure antenna and receiver performance and record the results, even when you are not present at the measurement point.

By Philip T. Sage, KF8JW

The recent proliferation of sound-card hardware and software has offered the amateur radio operator a variety of new exciting programs and modes like PSK31. In 1965, J. W. Cooley and J. W. Tukey revived the fast Fourier transform (FFT).[1] Now there are many versions of FFT-recording software that allow both visual display and quantitative measurements, from which comparisons of antenna-system performance can be made.[2, 3, 4] In many cases, these measurements can be made using equipment that may already reside in your shack, perhaps along with a temporary reference antenna. Additional tests can readily be performed to isolate noise-generating RFI equipment and to minimize their effects in the shack.

[1]Notes appear on page 45.

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While we all tweak, tune and sometimes boast about how good our antenna systems are, it is difficult to really measure minor improvements in how well the antenna receives a station of importance. If a method could be developed that gave reasonable results, then a lot of weekend tweaking on the rooftop and tower could be refined and measured improvements recorded. There are a host of well-documented processes and practical techniques available to the amateur operator for improving antenna systems and feed lines. The intent of this article is to focus on a method for comparison testing of improvements.

Antenna Performance Comparisons

An antenna is simply a transducer or coupler between a suitable feed line and the environment surrounding it. To be consistent in comparing different antennas, *The ARRL Antenna Book* suggests that the environment surrounding the antenna be standardized.[5] Ideally, measurements should be made with the measured antenna so far removed from any objects causing environmental affects that it is literally in outer space—a very impractical situation.[6] A less-involved comparison is of more interest to many operators. We are not so much interested in absolute antenna characteristics, especially if one is on a small plot where the demand for a reference antenna in the best of cases is unrealistic.

Unfortunately, many amateurs do not know how to evaluate performance scientifically or compare one antenna with another. Typically, we put up one antenna and try it; then we erect another to see how it goes. This is obviously not a great method because changing band conditions or different S-meter characteristics influence the returned signal reports.

Many times, the difference between two antennas or between two different locations for identical antennas amounts to only a few decibels. This is difficult for S meters to discern. Expanding upon the approach outlined for indoor antenna evaluation in *The ARRL Antenna Book*,[7] very little by way of test equipment is needed to actively compare outdoor antennas. Differences of less than a couple decibels are difficult to hear and require the use of a good voltmeter. Now one can use a sound card at the receiver output with the AGC turned off. To compare two antennas, switching the coaxial line from one to the other is necessary. Nothing special is needed: A simple coax switch or routing through an existing antenna tuner via the direct paths will work. *The ARRL Antenna Book* shows even a simple toggle or slide switch will provide more than 40 dB of isolation at HF provided you don't transmit. Whatever difference shows up in the received signal is the difference in performance between the two antennas in the direction of that signal.

On ionospherically propagated signals, there will be continual fading; for a valid comparison, it will be necessary to take the average of the difference between the two signals.

Occasionally, the inferior antenna will deliver the stronger signal; but in the end, the law of averages will put the better antenna ahead. Feed systems should be identical, if possible, to eliminate differences caused by feed-line differences; however, if you are intent on evaluating differences from antenna to received audio overall, any reasonable feed-line system will work. The test chosen is slightly different for each set of circumstances, though.

In one situation, you are testing the difference between two antennas, all other things being equal; in the other, you are testing two different antennas, feed lines, and impedance mismatches. Both provide meaningful information from which the pursuit of improvement will provide benefits. SWR on each should be below 2:1 to balance the feed-line impedance if you are using identical feed lines and only comparing the antennas. This limits any interaction with the receiver.

So how can we use the new technology to make some basic relative measurements of antenna system performance with reasonable accuracy, and empower those with sound cards worldwide to do the same? Whatever technique is used, it must be capable of canceling, minimizing or incorporating on-the-air effects such as atmospheric noise, band conditions and so forth. Utilization of the frequency domain versus the time domain is seldom affordable for most amateur operators; however, using the new advances in sound-card technology and available software, several useful techniques are now within reach of most operators.

In theory, if the measurement between two antennas can be made nearly simultaneously with the same equipment, effects of band conditions and measurement-device errors can be effectively canceled. This is done by subtracting one result from the other, netting the relative difference between the two systems. By making repeated measurements with different stations as signal sources, one can obtain a relative ranking between the antenna under test, a reference antenna and any improvements you may have made or may contemplate. Plain atmospheric noise can provide some interesting differences between system performances.

Example

Let's say we establish some input-signal reference that we call 0 dB. Now antenna 1 measures a signal on 7.020 MHz at –30 dB that includes receiver noise, an error, atmospheric noise and transmitted signal strength.

Antenna 2 measures the same signal on 7.020 MHz at –21 dB including the same receiver noise, error, atmospheric noise, additional cable loss and transmitted signal strength. Because the signals are sampled at the same time—by subtraction, –21 – (–30)—9 dB is the difference in relative antenna system performance detected on the signal desired at the shack. In this case, the difference is between the antennas and the feed-line performance.[8]

Test Setup Hardware and Software

Sound-card coupling to your receiver is identical to that needed for PSK31 and other modes. I highly recommend installing audio transformers (RadioShack #273-1374) or equivalent to reduce ground-loop noise. As some operators have reported, a direct connection will work; but the 60-Hz ground-loop noise and

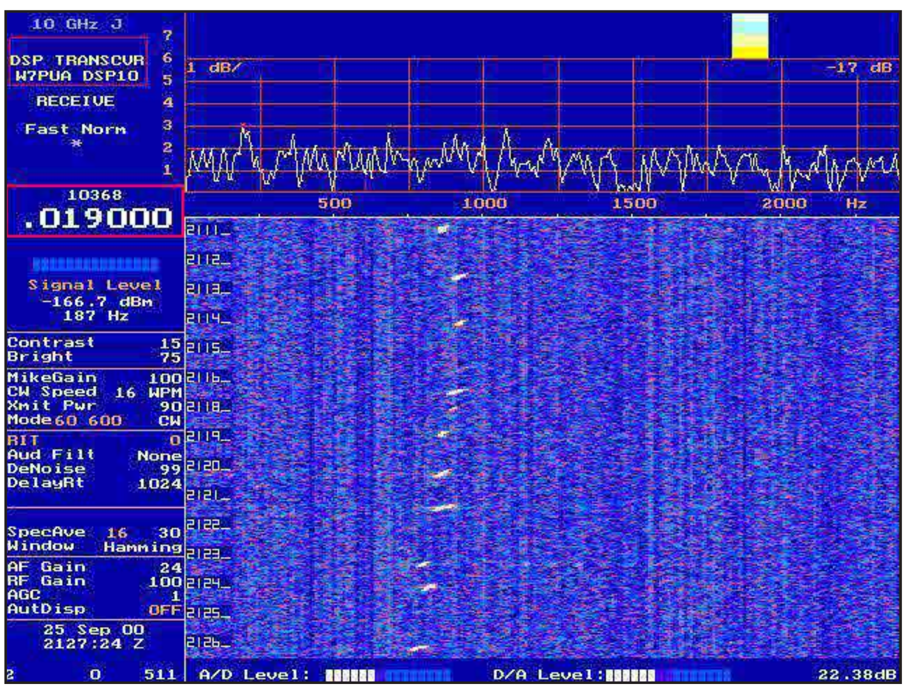


Fig 1—This is a screen capture from Mike Reed, KD7TS's DSP-10 software defined radio (SDR). Mike wrote: "The signal I am receiving is from the 10-GHz beacon maintained by NU7Z, and is located about 15 miles north of the airport. The path to the beacon is totally blocked, (via Mountain) so the only way to see the signal is via rain, aircraft and sometimes tropospheric scatter."

harmonics can be seen, even if they are at –60 dB. See PSK31 Web sites for further details on how to connect your receiver to your sound card.[9]

As mentioned, for antenna comparison, the equipment consists of two or more antenna systems, a coax switch, a sound card, a receiver/transceiver and a PC running fast enough to support the sound card. For RFI research and testing, only the receiver and soundcard are needed. Some form of timing device—a watch or egg timer—is also helpful. One of the many FFT software packages available must also be running on your PC. I used Richard Horne's software for most of my measurements; however, Mike Cook's and others work just fine. For example, Fig 1 is a screen print showing 10-GHz airplane reflections received at the location of Mike Reed, KD7TS.[10] Mike used a DSP-10 software-defined radio (SDR) to capture the data. Even the DSP-10 can be used for this work at higher frequencies with a transverter.

A reference antenna can be employed to measure before-and-after differences caused by modification of our antenna of choice. The reference antenna I used for HF comparison was a random-length wire connected to an antenna tuner via the matching transformer for long wires. I compared it with a G5RV Jr., 40-meter doublet at about 50 feet and a Gap Titan DX vertical mounted near the ground. Feed lines on the G5RV and the Gap Titan DX are lengths of RG-213 or RG-8U foam depending on what I had available. Similarly, I could have made up identical feed lines and compared antenna against antenna in their current locations. All antennas are connected to an MFJ Versa Tuner II using the direct selections, as against tuner-applied paths in the Versa Tuner unit, although any suitable coax-switching device will work.

The random-length wire was selected due to its relative ease of construction and installation. If you need a reference antenna, this will work in a pinch. A single-band dipole is perhaps better. Some care at your location must be taken to ensure that large objects do not influence relative differences between antennas on your test site. The method described can also yield the difference between your poorly placed system and a better-placed system. The random-length wire reference antenna is useful for gauging changes made to your permanent antenna if you do not have a second antenna already erected.

Comparison Methods

The FFT is a very useful tool for examining the frequency content of a given signal. I use it to advantage below. To be brief, the conversion from a time-domain wave, like a sinusoid, to the frequency domain is accomplished using the Fourier transform. A CW signal, for example, consists of nearly a single frequency. Given a perfect sine wave $x(t) = A\sin(\omega t)$, where $\omega = 2\pi f$, and A is the magnitude of the CW wave at its peak, $X(f)$ represents the Fourier transform of $x(t)$ at frequency f.[11] The *fast* Fourier transform (FFT) is an algorithm that calculates $X(f)$ quickly on today's PCs.

Note: An antenna system in this article shall consist of the antenna, feed line, connectors, and all inherent losses due to installation location, ground effects and so on, although I will reference them differently by type. A switch is used to effectively allow both antenna systems to be connected to the same receiver at almost the same time. By continuously recording data to a file and throwing the coax switch at a predetermined time from the start of recording, toggled back and forth between antenna systems, a data set is recorded in a brief period of time that includes both antenna systems. This is considerably easier than using a voltmeter or oscilloscope to read the audio output of your receiver. Of course, extreme care should be taken to disable accidental transmitting. I establish a log file and adhere to a naming convention for saved data sets, as it is real easy to build up a lot of data files and get confused.

By taking enough samples by toggling the coax switch, and analyzing sufficiently close to the switch point, one can effectively cancel out the random effects of changing band noise and conditions. CD-quality sound cards are capable of sampling at 44.1 kHz, so generating data is not a problem.

The receiver AGC should be turned *off*. If it is not, the AGC action will mask gain differences between test antennas. Consult your radio manual or the manufacturer to learn how to disable the AGC. Many receivers have this feature without modification and some do not. Once complete, I suggest you experiment by switching a couple of times and reviewing the results to ensure it is off.

Sound-Card Initialization

First, connect an audio source to the input jack of your sound card. Either the **LINE IN** or **MIC** inputs may be used

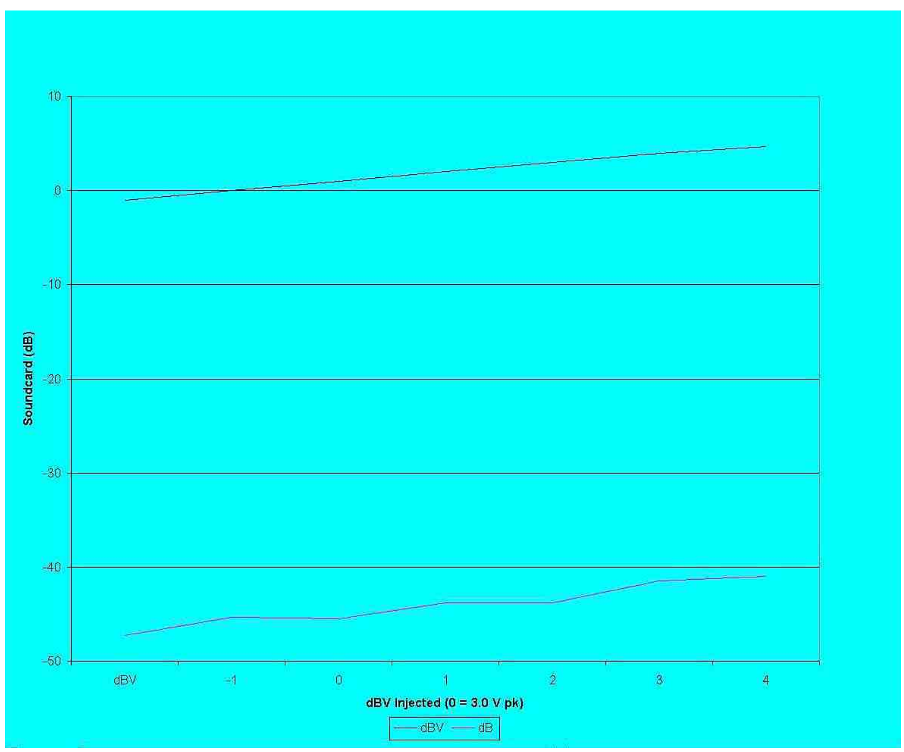


Fig 2—A chart of injected signal strength (dBV) compared to that read by the software.

depending on the signal level expected; however, for most radios, **LINE IN** is the correct choice. It is always best to turn off your computer before making these connections.

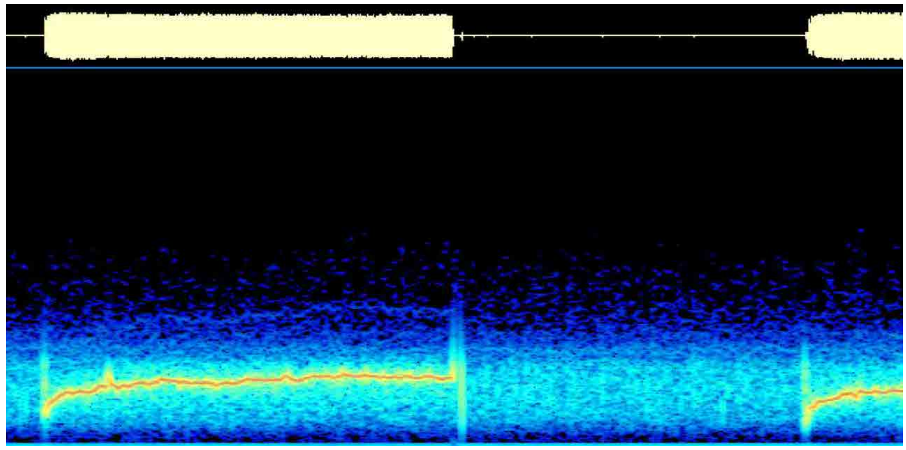
For software running in a Windows environment, you must instruct your sound card to process audio signals. *Windows 95* provides a separate volume-control program for this purpose. Choose "File - Volume Control" from the *Spectrogram* menu (or function key F6). This will start the volume-control program. Then choose "Options - Properties" from the recording control menu. Then mark the "Recording" button and under "Show the following volume controls" choose "Line-In," and "Microphone."

Now when you click "OK," you will see a Recording Control allowing you to adjust the volume for line-in and microphone recording. Select either the line-in or microphone input here, depending on the audio source that you intend to use. It is best not to choose both inputs simultaneously, but to choose one or the other.

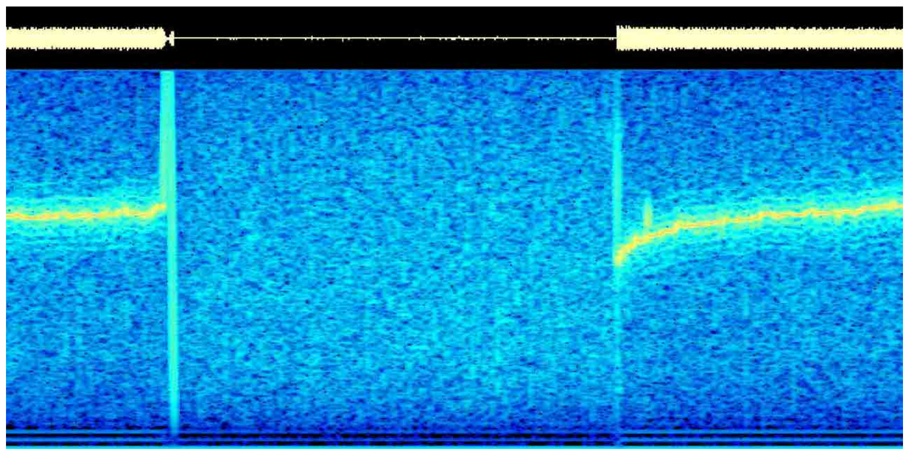
Occasionally the Windows recording control will also have an active menu item called "Options - Advanced Controls." If this menu item is available, it should also be selected. This selection will make available an "Advanced" button on the recording control for turning automatic gain control (AGC) on or off. AGC in software is also to be turned off. An RF generator tuned to the receiver, or a strong signal switched into the receiver input, will impose a nice step change into the receiver. Simple inspection will yield the time constant of most AGC systems.

Most sound cards offer 16-bit resolution and sampling rates of 44.1 kHz or 48 kHz, although they will also operate at lower quality settings for less-demanding circumstances. Most sound cards utilize a 16-bit A/D converter to produce a signed 16-bit number (one sign bit and 15 data bits). Therefore, the maximum digital output amplitude of the sound card is $(2^{15}) - 1 = 32767$.

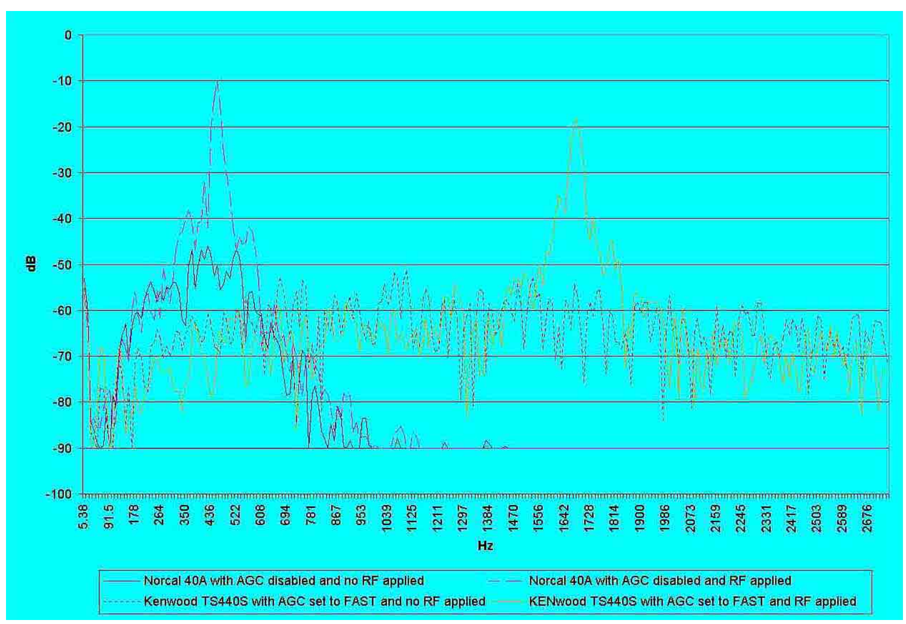
CD digital employs a sampling rate of 44.1 kHz and a 16-bit dynamic range. That is, 44,100 samples per second, each one describing the waveform amplitude at that moment in time with a 16-bit number—16-bits offering 65,536 steps from which to choose. That works out at around 176 kB/s, 10.5 MB/min or 630 MB/hour when recording in stereo. The most common file format used to store digital audio on PCs is .WAV. Therefore it is important to record only what you



(A)



(B)



(C)

Fig 3—At right, (A) is from a Norcal 40A CW transceiver with its AGC disabled. A fixed RF signal is switched in and out of the receiver circuit to confirm the AGC is not varying the receiver gain. (B) is from a Kenwood TS440S with the AGC set to FAST. Note the subtle differences in background noise and noise floor at the transition from RF on-to-off and off-to-on. If the AGC is on, even fast, it will change the gain of the receiver, regulating to the higher-level signal, attempting to maintain constant audio output. The comparison test at (C) illustrates graphically why AGC circuits should be switched off or disabled (see text and Note 20).

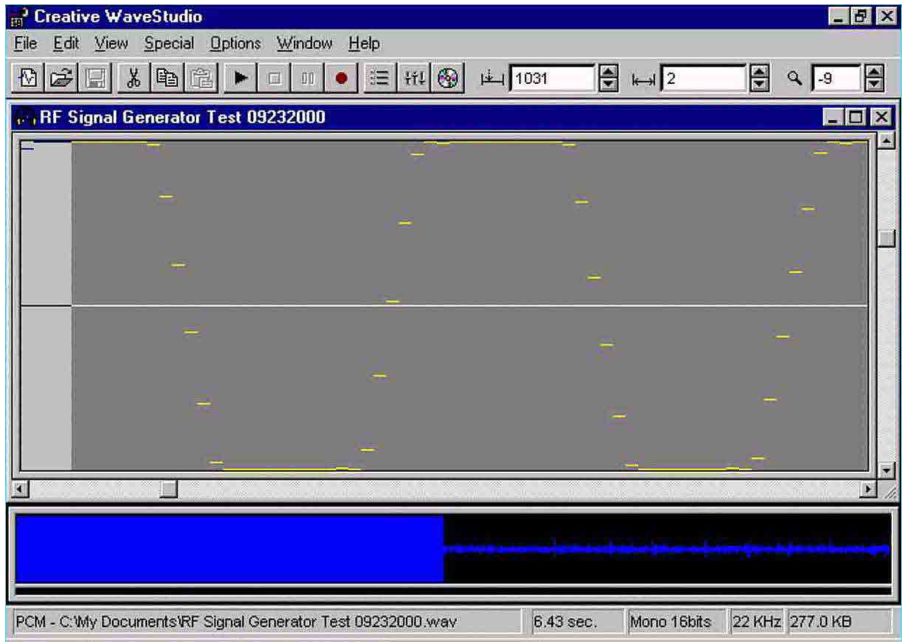
intend to analyze and to keep it short.

The *Spectrogram* program (see Note 3) uses this maximum value as the decibel reference. If *S* is the digital output of the A/D converter on the sound card in counts, then the decibel value is 20 log (*S*/32767). There is no way of knowing the conversion factor for the input network of every flavor of sound card. If you need to know this conversion, you can simply calibrate your system by injecting known levels of audio signals, then measuring the output on the spectrograph display or another suitable program. *Spectrogram* and other software are more suitable for making relative comparisons between signals than for making absolute measurements.

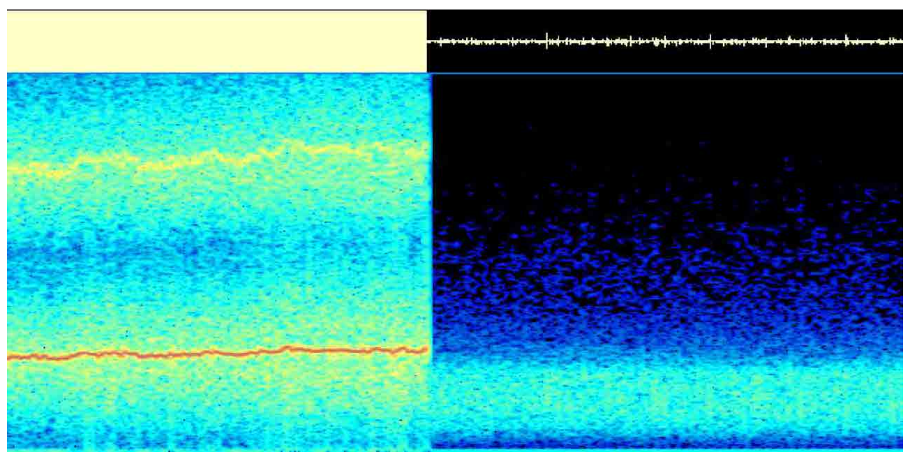
By not touching the receiver and switching with a coax switch, the gain of the receiver stays constant across a few seconds. As long as the A/D converter has a linear relationship to the input signal level, then a decibel comparison between two signals will be accurate even when their absolute values are not known. This is the basis for the "difference method."

The time base on the sound card/computer can affect the sampling rate, and thus the accuracy of the frequency domain information somewhat. I measured the input frequency of the audio generator and found the software-sampled frequency to be within 1% or less. I also injected an audio signal at around 950 Hz, with steps from –2.0 dBV to +4.7 dBV. The sound volume control was set to about 80%. The result was linear. Dependent on what software you choose, you may want to verify the input scaling also.

I recommend leaving the volume control at 100%, and reducing receiver RF gain or audio volume, as this provides more signal level to the software and may improve accuracy. If you have access to an RF signal generator, you can inject a known reference signal into your receiver's antenna jack as a double check. Switch this reference signal into the recorded data at the beginning and the end of each data set to ensure the test setup has not drifted during data collection. You can also use it to calibrate the receiver/soundcard system to a known RF input (see Fig 2). However, because the difference method effectively cancels out all drift issues except those that may exist in your receiver/sound card across a couple of seconds and all unknown gain issues, you can omit this step if you like. The AGC circuit, as mentioned, needs to be turned off. As with



(A)



(B)

Fig 4—(A) clearly shows that the signal is clipped. (B) is a view of a .WAV file with the line-input slider varied from 100% (left) to 50% (right) while injecting about 12 mV at 7.040 MHz into a receiver.

Table 1—Tabulated Results for Fig 3C

Receiver/Test Conditions	*Summation of dB*Hz*	*Average dB per 10.77 Hz Bin*	*Average dB per spectra**
Norcal AGC Off No RF	–69420.2	–66.463	–0.711
Norcal AGC Off w/RF	–63241.4	–59.764	–0.639
KW AGC on FAST No RF	–179996	–64.921	–0.254
KW AGC on FAST w/ RF	–180045	–65.302	–0.255

*Due to the crystal CW filter in the Norcal 40A, data was evaluated from 0-1007 Hz. For the Kenwood TS-440S, with no filters selected, the data was evaluated for 0-2750 Hz. The evaluation method is Average dB per Spectra = (Average dB per 10.77 Hz Bin)×10.77/D, where *D* = 1007 for the Norcal Spectra and *D* = 2750 for the Kenwood spectra.

any measurement, employing more time, care and equipment increases the potential accuracy.

AGC Concerns:
A More Thorough Discussion

The AGC circuit can obscure test results because as the input signal strength changes, the AGC regulates the audio to near-constant level for listening comfort. To prevent the AGC circuit from changing the gain of the receiver, it must be disabled for a more than casual comparison test. To prove that AGC was not varying the receiver gain, I devised a simple test.

Most AGC circuits work off the level at the last IF of the receiver. Using a servo loop, they regulate the audio level to a fixed value so that the operator does not experience undue stress on his or her ears when a strong signal suddenly appears. To accurately measure differences between two unknowns, the AGC must be disabled or turned off. This forces the receiver gain to be constant. Other circuits that dynamically affect the receiver gain must also be turned off. On some radios, this step is very easy. On the DSP-10, the AGC is all inside the DSP software. When you turn it off (set delay to zero), the receiver gain becomes fixed.[12]

To measure that the AGC is off, I choose to simply apply a strong signal, and measure the resultant change in output power. If the AGC is on and functioning, a strong RF signal induced into the receiver should force the AGC to lower the gain and output the same audio power to the speaker.

If the AGC is truly off, the test should yield a rather dramatic increase in audio output when the strong RF signal is applied. Using base receiver noise as a difference platform, I performed several tests with different transceivers. The results and method are presented below, where I choose a Norcal 40A[13] with narrow CW filters, AGC disabled, and a Kenwood TS-440S with no filters selected to help demonstrate the test.

The comparison test in Fig 3 illustrates graphically why AGC circuits should be switched off or disabled. The output level of the RF oscillator was not varied. The Norcal 40A curves are essentially on top of each other, one with an RF peak and one without. The measured "with RF" curve is louder across the graph than the "without RF" curve. With the AGC off, this is what you would expect. The Kenwood, with its AGC on, has reduced the audio gain to maintain nearly the same level of audio. Notice the difference in the noise floor: The noise level is reduced in the presence of a strong signal. The noise level with RF applied should never be less than the level without RF, if the AGC is *off*. If the AGC is not switched off or disabled, it will mask the true decibel differences you seek.

Mathematically, you can further verify your AGC is off. This step is not required for casual comparison, but verification is considered prudent for precise measurements. The data are in evenly spaced frequency-band buckets or *bins* with each bin containing the value (in decibels) for its portion of the spectrum. Therefore, we can multiply each bin's decibel value by its bandwidth and summing the results for all bins to approximate the area under the spectral curve in decibel-hertz. Integrating across the spectrum (by simple summation in spreadsheet) yields the

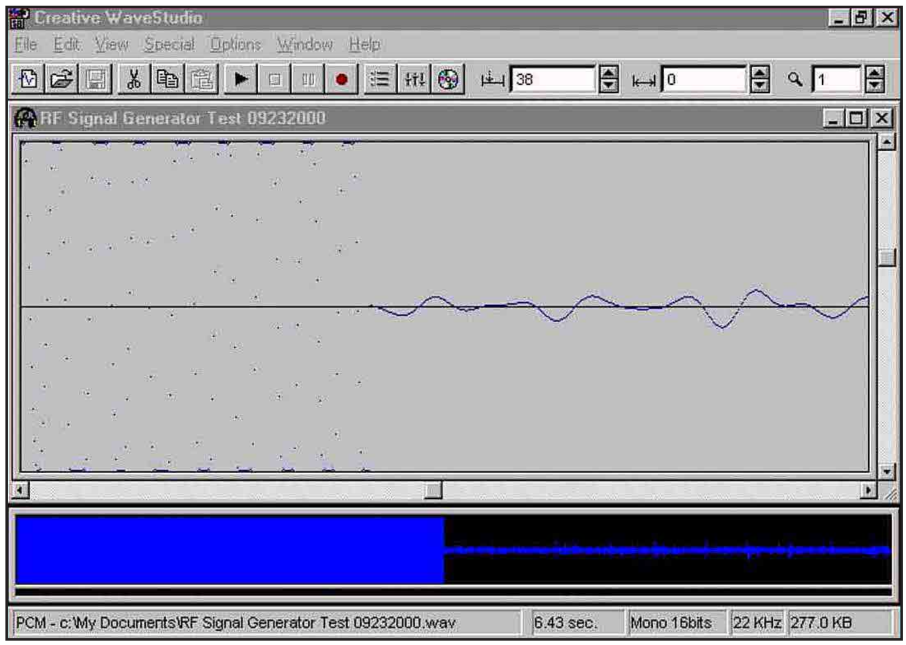


Fig 5—A view of a .WAV file with the line-input slider varied from 100% (left) to 50% (right).

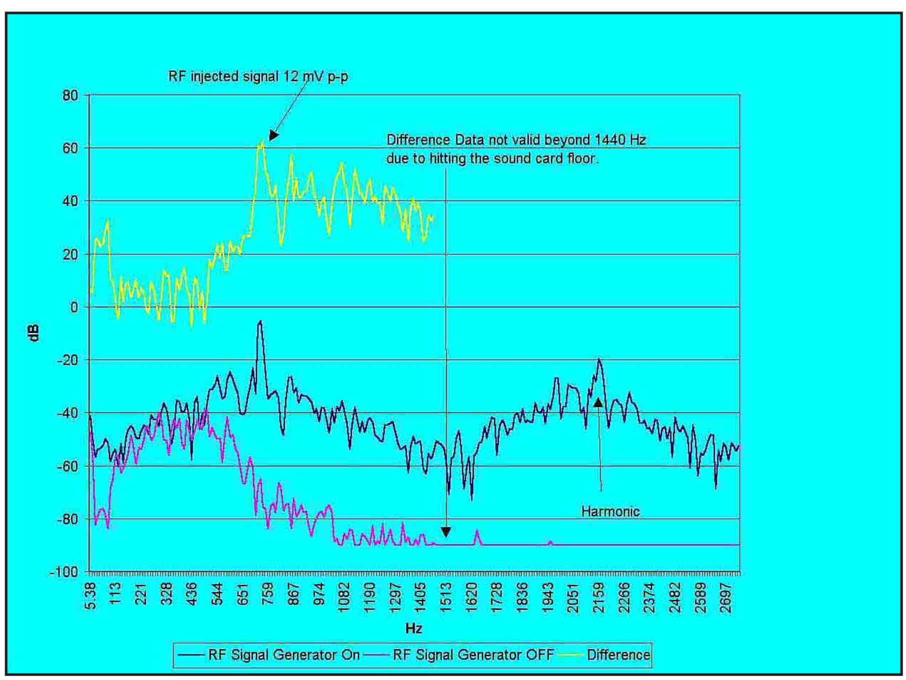


Fig 6—Several pitfalls to avoid in measurments. In particular, the difference is only valid as long as both signals are within the sound card's capabilities. (See Note 20.)

total power of the audio signal from your receiver (see Table 1)

If the AGC is on and working, the difference between the condition with RF applied and that with no RF applied should be near 0 dB-Hz. This is so because the AGC circuit adjusts the gain to prevent strong signals from overloading the receiver's audio circuit and hurting your ears.

Clearly there are no stark differences in the data for the Kenwood with its AGC on. In fact, the difference in spectral power for the test was essentially zero (–0.001 dB). The NorCal 40A, however, does have the AGC disabled, and the average difference is (–0.711 – –0.639) or –0.072 dB.

Other Avoidable Pitfalls

Other pitfalls to avoid include audio clipping, in Fig 4. Audio clipping can be generated either by applying a peak-to-peak voltage to the sound card input that exceeds its permissible range or by increasing the software gain too much. Both produce the same undesirable results. In theory, a square wave is made up of the sums of many sine waves at frequencies ranging from its fundamental to infinity. As more of the peaks are clipped from a sine wave, it more closely resembles a square wave and contains more high frequencies, which can appear in the FFT of the clipped signal.

In Fig 4A, a screen capture of *Creative Wave Studio*[14] clearly shows that the signal is clipped, so it is not usable because it contains harmonics that were not in the original signal. You can prevent clipping by varying the sound control's line-input slider control, as shown in Fig 4B. For any signal you choose to sample, quickly check and set the incoming level to prevent clipping. Once they're set, do not vary the levels during the tests. It's also a good idea to record the settings for future reference.

Spectrogram amplifies the displayed signal shown at the top, but the FFT calculations are based on the unamplified signal.[15] In short, if no clipping is shown on *Spectrogram*, there is none. If the display is clipped, the signal *might* be clipped. Also, notice the faint "ghost" trace near the top of the left-hand spectrum that does not continue when the clipping is removed.

There are several ways to reduce audio clipping. Simply turning down the volume control of your receiver should allow you to prevent the receiver from overdriving the sound card. In software, there are a few items to check.

In *Windows 9x*, the volume control has a level bar graph for the line-in selection under the recording properties. Keeping this level in the green or yellow areas will prevent clipping. In *Spectrogram*, inspection of the time waveform trace at the top of the scroll display will show you whether you may be clipping. It does not guarantee it, though. Richard Horne, *Spectrogram's* author, uses the raw ADC signal. In *Spectrogram*, however, signals that are shown within range are not clipped. The time waveform at the top is amplified so that smaller levels of signals are more easily readable. Other FFT programs may interact differently with the sound card.

The signal shown in Fig 4B is clearly maximized in the left-hand time waveform. Take care to ensure a proper reading is taken—as with any measurement. The source here is my Heathkit RF oscillator, Model IG 5280. It wobbles a bit, but puts out constant amplitude. Here the RF gain control is at minimum and the unit is connected to the MFJ Versa Tuner II Coax-1 input, fed di-

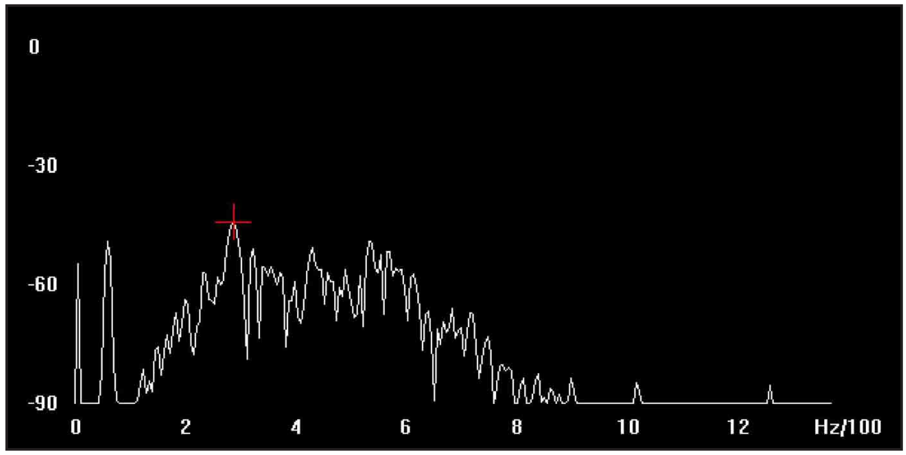


Fig 7—Simple FFT analysis of a sampled signal yields a large amount of information. The red crosshair is at 289 Hz on top of the CW signal, as seen by the soundcard from the receiver at 15.635 s at –47 dB. The Approximate Noise Level of combined receiver noise, atmospheric noise and other sources measures about –54 to –62 dB within ±100 Hz of the detected signal. This was sampled quickly using the cursor to display the level of signal at various frequencies. The signal above noise level is then –47 dB – (–54 dB + –62 dB) / 2 = 11 dB. Exporting data in a text-file format and then importing it for manipulations in a standard spreadsheet application can make more detailed comparisons easier to accomplish.

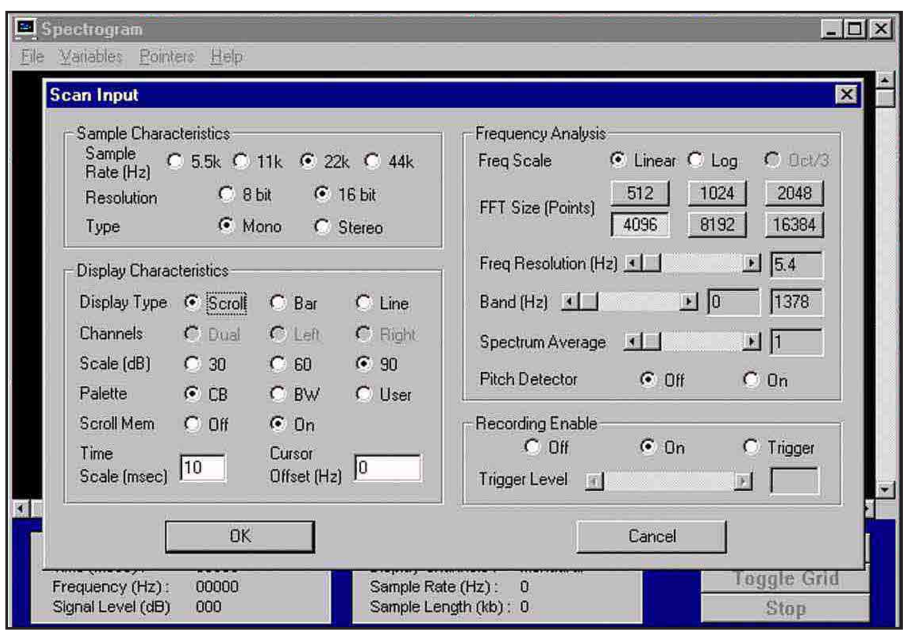


Fig 8—A screen capture of *Spectagram's* Scan Input menu showing the author's settings.

rectly through to my receiver.

Fig 5 shows how a reduction in the *Windows* volume control brings the signal back in line and eliminates potential software clipping. Other software programs may operate differently, particularly those running under *DOS*.

Lastly, there must be enough input signal to ensure the FFT does not "bottom out." The difference method is only valid if both data sets are wholly between 0 and –90 dB. By varying the Sound control Line-Input-Slider control you can remove any potential audio clipping the sound card and software are introducing. For any signal you choose to sample, a quick check and setup of the incoming level to prevent clipping should be performed. Once set, it should not be varied for the duration of the difference test. I suggest recording these settings for future reference.

Fig 6 illustrates some of the pitfalls to avoid by demonstrating the undesirable results obtained. Notice that the difference is only valid as long as both signals are within range of the sound card's capabilities.

Method for Data Acquisition

Once your test setup is wired and ready, it is time to take a few on-air samples, and begin the process of analyzing the difference. I'll describe steps for *Spectrogram* running under a *Windows 9x* environment. Other systems and software would be similar.

The recorded signal when analyzed at specific points in time yields detailed data of interest. Among others, I chose to record a CW QSO in progress on 40 meters. I highly recommend either CW or PSK31 for comparison purposes. An idling PSK signal makes a great poor-man's two-tone generator for some interesting receiver tests.

Pick a signal that appears solid and consistent both audibly and visually. If you choose to use a CW signal, look for a slow sending rate. A little practice is needed to switch coax in the center of a CW dash for precise measurements; but less-precise measurements yield acceptable results. A PSK idling signal is a bonus, but is not found routinely on all bands, at all times.

Simple conversion of a sampled signal, processed via FFT, yields a large amount of information (see Fig 7). Again, before the test data is taken, carefully check that receiver AGC is off and that any expected signal will produce no audio clipping on any of the antennas you will switch to the receiver. For pure antenna-difference measurements, excluding environmental effects generated by placement, the feed lines should be identical.

Carefully consider the data directory structure and setup the test-data directories before you start taking data. Make a simple checklist of what data you expect to take and where you will store it; it may help later. Record the antenna sequence to be switched, so it may be referenced later. A stopwatch running with a helper indicating when to switch, eg, every three seconds, will help compress the amount of data recorded, easing the analysis burden later.

Data Collection

I'll describe detailed steps for the *Spectrogram* program. Start *Spectrogram* in the Scan Input mode, without selecting the record option. Tune across the band to find a stable signal to sample (CW or PSK31). Once you find a signal you want, restart *Spectrogram* in the Scan Input mode, enabling the record-to-file option. Enter your planned filename, start recording and switch antennas at the desired times. Close the record feature by pressing the stop button on the screen.

The Recording-Enable option allows you to switch recording on and off while you are scanning audio input. If you select Recording Enable On, you will be prompted for a file name. Then while scanning, you can click the Save and Stop-Save buttons to control recording. Notice that it may take a few seconds before recording can be started again after you have stopped it. Each time you restart recording, the new data segment is added to the end of the wave file you have specified. This feature allows you to record interesting events without the need to record continuously for long periods. Please notice that use of the Recording-Enable feature requires that your hard drive be continuously powered. Some computers can operate in a power-saving mode where the hard-drive power is switched off if the drive is not accessed in some fixed amount of time (usually 10 to 20 minutes). If you attempt to Quick Save after the hard drive power has been turned off, the time delay required to turn the drive on and bring it up to operating speed will disrupt the timing of the *Spectrogram* program. This delay will affect the data display and recording. If you intend to record after an extended period of scanning, then you must disable your computer's power-saving mode so that the drive runs continuously.[16]

I used the settings shown in Fig 8. Other settings vary the data saved, so care in selection is important. The settings shown best fit the characteristics of my Norcal CW radio; others will work for receivers with wider IF filters, but will lower the frequency resolution. *Spectrogram* does a nice job making the selection process user-friendly.

Analyze a Data Set

Data are next selected by opening the file you just created with the Analyze-File option, playing it fully and then selecting the switched transitions. Here is where a helper—who times the switch points—can pay off. The file displayed is the total file recorded. Having written down both the switch sequence and time, you can now select the appropriate data set you want. By clicking on the correct time, as indicated by the readout, you select that data sample as a starting data point. In scroll mode, the switch points may be obvious; but having the time ensures you don't have to repeat the measurement.

The "law of averages" helps to minimize effects of random noise. I performed random single samples, compared them against others using the built in Spectrum-Average function and found no difference on a selected data set after averaging. Use the Spectrum-Average slider to determine the number of sequential spectrum measurements that are averaged together before display on either the scrolling *Spectrogram* display or the scope display. Averaging is particularly useful in recovering weak periodic signals from a noisy background, but it is probably not of much interest in analysis of speech or other rapidly varying signals. You can choose averaging of one (no averaging) to 128 spectrum measurements.

It is important to understand how the spectral-averaging function works.

Table 2—Spectrogram 5.0 Data

44100 Hz Sampling Rate
2.69 Hz Frequency Resolution
15640 msec Event Time

Num	*Hz*	*dB*
0	1.35	–51.0
1	4.04	–54.3
2	6.73	–59.3
3	9.42	–58.3
4	12.11	–57.9
5	14.80	–61.2
6	17.50	–59.8
7	20.19	–66.4

Carefully select analysis points that are not too close to the transition. Here is how it works. If you are sampling in the time domain at 22 kHz (about 45 µs per sample), the time required to fill an FFT buffer of 4096 samples is 0.186 s. With spectral averaging set to 10 samples, *Spectrogram* will average 10 FFT samples taken at the time base indicated in the setup screen. For example, the screen in Fig 8 shows the Time Scale set to 10 ms.

Say you put your cursor at 15.000 s. The first FFT slice will be calculated using the time domain recording from 14.824 s to 15.000 s, or roughly 0.186 s of time-domain data. The next FFT slice will be calculated on the data set from 14.834 s to 15.000 s. This is two of the 10 FFT slices averaged with spectral average set to 10.

For the difference method to work, you must select a point at least 0.186 s after the transition if you are going to use the spectral averaging feature to smooth out random noise. This prevents any data taken before the switch was thrown from being incorrectly selected. I sampled most data at 44.1 kHz, so the time I needed to select after the switch point is roughly 0.093 s.

Using the Log-Data function, you can further expand your analysis. For quick comparisons, the Line display and cursor capabilities are sufficient. Log Data exports data to delimited text files (see the truncated set in Table 2). The data sets with the settings shown conveniently contain 255 samples. This aligns well with my spreadsheet's maximum charting capability and allows easy transition to enhanced graphical review of multiple data sets.

Importing this into a spreadsheet is easy, since most new spreadsheets recognize text files and will help you import them. Once you have imported the two sets of data from either side of a switch point, you can extract the difference easily. By cutting the decibel data from one, pasting it into another, then subtracting the difference between the two data sets, you have the difference between the two data sets in discrete bins.

Averaging the difference data set across the sampled spectrum provides

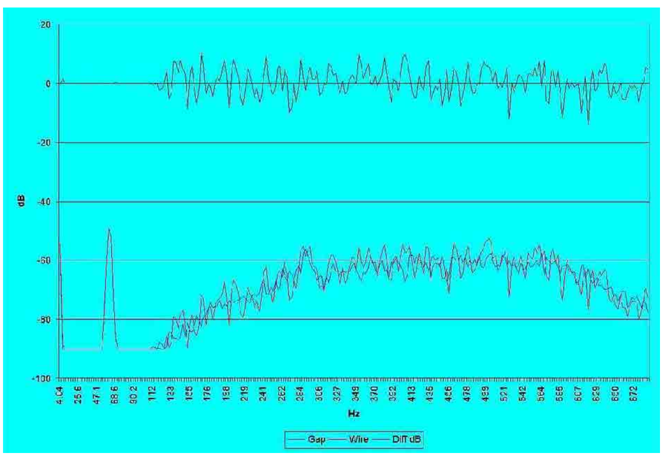


Fig 9—Test Results show nearly identical responses when overlaid as shown above. The wire antenna clearly has much more noise than the Gap Titan DX. The Gap clearly pulls out the CW signal at 289 Hz as the highest peak, the random-length wire has peaks of noise higher than the desired signal. (See Note 20.)

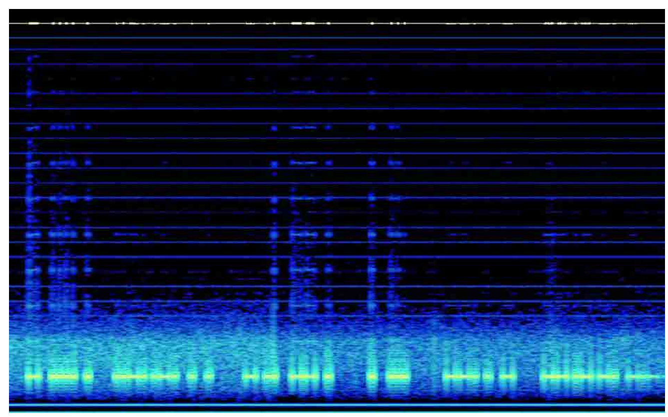


Fig 10—The spectrum of a switch transition between Gap Titan DX and G5RV Jr system.

Fig 11—The received FFT scroll for a sequence of G5RV, Gap Titan and random wire. Notice the subtle, but detectable, difference in noise floor. Also, there are some detectable harmonics on the CW signal only detectable on the G5RV, and they are around –40 dB below the main carrier. The harmonics are more clearly visible when on the G5RV Jr than on the GAP system.

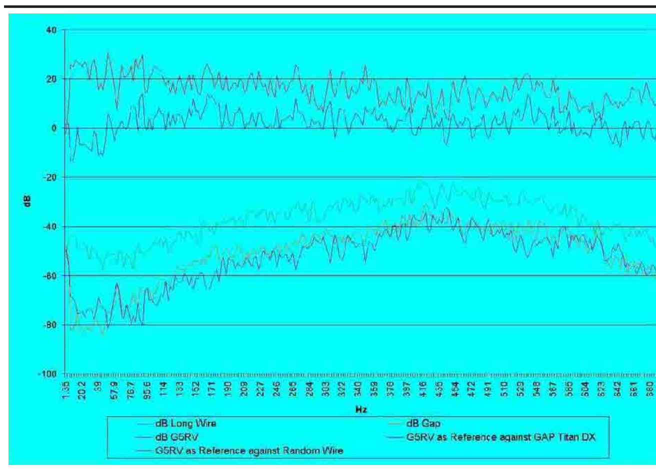


Fig 12—The final analysis showed the current system used as the reference has the least noise of systems tested. Additional checks on the random-length wire indicated that the data was displayed as clipped, but was not clipped. (See Note 20.)

the average signal + noise of the set. Carefully selecting the signal you recorded and the frequency bin containing it allows a quick calculation of signal – (signal plus noise) as it relates to the difference between the two systems under test.

Additional averaging, by taking multiple data sets from several different stations and locales will allow a more comprehensive evaluation of the difference between the two systems.

On-Air Receiver Tests

Using an idling RTTY signal, several quick comparisons where made between systems. About 10.5 s of data were stored to a text file for each sample, imported into my spreadsheet and "differenced" against the quietest system (see Figs 9, 10, 11, 12). "Differencing" can be performed against either a reference antenna or a shielded dummy load. If you have a dummy load, this allows a further difference method using the receiver's no-input performance as a base reference.

The final analysis (refer to Table 3) showed that the system used as a reference was the lowest noise receiving antenna system tested. Additional checks on the random-length wire indicated data was displayed as clipped but was not actually clipped.

Atmospheric Noise as a Test Input

Using atmospheric noise received for comparison between antennas and dummy loads allows for some interesting analysis of antenna-system noise pickup and rejection capabilities.

Carefully review the periods used to ensure equal test conditions are present. For instance, the horizontal swath shown just below center in Fig 13 is an annoying growling noise that worked its way up the band during the test. The sequence was initially rejected, but later included to demonstrate that care must be taken once the sample is taken to declare it usable. By taking samples very close to the switch point, even error caused by changes in atmospheric noise on a hectic night can be minimized; however, selecting a period of just plain old noise is much better.

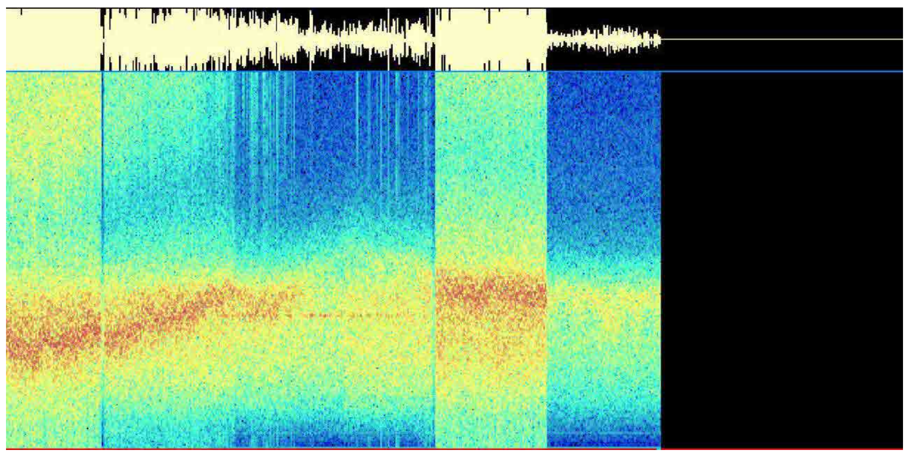


Fig 13—An the FFT scroll display from *Spectrogram*. The sequence from left to right is random wire, Gap Titan DX, G5RV, Gap Titan DX, random wire, dummy load (50 Ω). See text for details.

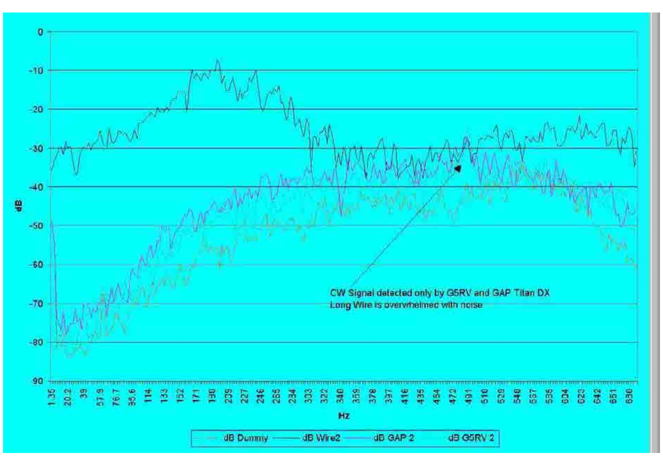


Fig 14—A comparison of three antenna systems against a dummy load, with atmospheric noise as the signal generator. This view shows the dummy load as just one more antenna system. (See Note 20.)

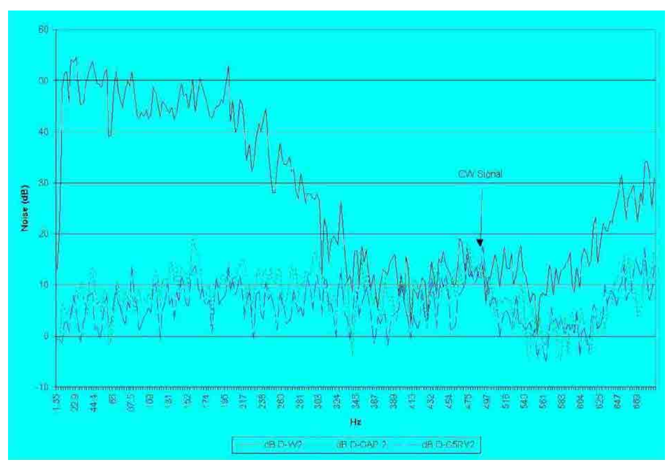


Fig 15—The same comparison, with the antenna responses plotted as difference from the dummy-load response. (See Note 20.)

Table 3—Noise Comparison

G5RV with RG-213, as Reference	*Difference Averaged across 255 2.69-Hz Bins (Positive is more noisy)*
GAP Titan DX – RG-8U.	2.25 dB
Random Long Wire – Versa Tuner II Balun	15.89 dB

The top portion of the view shown in Fig 13 is the time waveform; the bottom is the FFT scroll display from *Spectrogram*. The sequence left to right is random-length wire, Gap Titan DX, G5RV, Gap Titan DX, random-length wire, dummy load. Notice the intensity of the noise signal apparently overloading the input of the sound card on the random-length wire. The test was performed on a night when 40 meters was dead because of extremely high atmospheric noise. High atmospheric noise is not a requirement; it just helps to better illustrate the issue.

The receiver RF-gain setting was high enough to allow reasonable measurement on the G5RV Jr. in the center of the figure. In the middle of all that noise is actually a brave CW operator's signal. The results of each transition were analyzed, and plots of the later transitions in the frequency domain are presented below.

At the conclusion of the sequence on the far right, is a plot with the receiver connected to a dummy load. This was done to determine receiver and power-supply induced noise, as a reference for this version of the difference method. The data sets track well, except for the random-length wire. Subtle differences that averaged 2.4 dB were detected between the systems of G5RV and the system with a Titan DX. Notice that this is for the system: including feed cable, mounting location and all that goes with losses generated by many issues. It does not necessarily represent the capabilities of either antenna in optimal circumstances. Replay was used to ensure that the data-set selection point was taken during a dash. Figs 14 and 15 show two ways of viewing the data.

Differences between more obvious data sets can be inferred directly from the FFT plots of several software packages. This is a fast method to examine larger differences. Table 4 is one from *Spectrogram*. When using atmospheric noise as a source, several data sets must be taken to average out differences from high noise levels. The results help draw conclusions about the noise pick-up capability of your test antennas.

Certainly, a lot of attention over the years has been focused on other characteristics of antenna performance. In Table 4, I have calculated the S/(S+N) ratio in decibels, but it is interesting to note the apparent difference in noise susceptibility among the different systems.

Again, please notice that the above tabulation is for antenna systems, not just the antenna alone. The Gap and G5RV include a coaxial feed line—in this case, different lengths and types—and of course the long wire has none. There is relatively a small difference between the received signals, but surprisingly there is a large difference in the S/(S+N).

A comparison of what you have installed, as I have done, gives solid information not necessarily about the antenna design, but about system performance overall. The next step is to get out the Noise Bridge and other test equipment. I recommend that you use multiple signals. They should be analyzed before you make a decision on altering your antenna.

Applying the Difference Method

Noise- and RFI-Reduction Techniques

A useful application of this method is to test for RFI. By switching suspect devices on and then off, you create different conditions that generate data for the difference method. I decided to work on improving the receive capability of my 40-meter CW QRP rig. I built a Norcal 40A kit from Wilderness Radio. Late one evening, I noticed that when I switched off the power supply, the rig received for about six seconds before

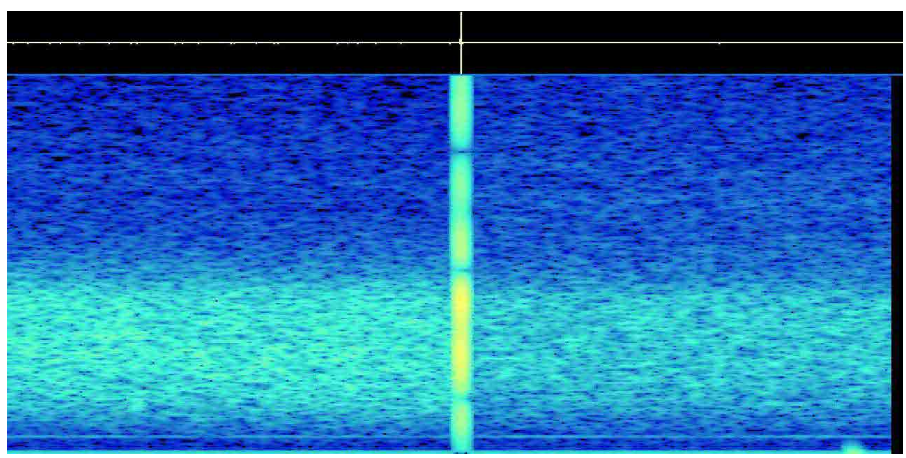


Fig 16—This is 6 s of data extracted from the *Spectrogram* program. The central vertical dividing line is noise generated by switching off the ac power to the 13.8 V dc power supply. The left-hand side (power on) clearly shows more noise than the right-hand side (power off).

Table 4—Data Inferred from *Spectrogram's* FFT Line Plots

Calculation	*Dummy Load (dB*Hz/2.69)*	*Long Wire (dB*Hz/2.69)*	*Gap Titan DX (dB*Hz/2.69)*	*G5RV-Jr (dB*Hz/2.69)*	*G5RV-GAP Difference*	*G5RV-Wire Difference*
Summation of Spectral Power (dB)/2.69 Hz Bucket	–13518	–6575	–12033	–11415	—	—
Average Spectral Power (dB) per 2.69 Hz = (S+N) Ave	–52.8	–25.68	–47.00	–44.59	—	—
Signal at 488Hz – Dummy Load Reference (dB)	N/a	–11.4	–13.9	–11.7	2.2	2.5
S–((S+N)Ave) (dB)	N/a	–15.7*	8.1	3.48	4.61	23.82

*Actual signal was below noise floor or spectrum.

the power supply capacitor drained and shut down the rig.

This was an accident, but I noticed the *Spectrogram* (still recording) seemed to get considerably quieter while running on what charge was left on the filter capacitor. I had discovered that the inexpensive power supply was a good RF noise generator. The noise came through on the received audio, via the supply voltage, as a small ripple. My multimeter measured the ripple at only about 1.4 mV ac. The oscilloscope showed about 5 mV of noise with occasional isolated peaks at around 8 mV. The oscilloscope was ac-coupled, sweep was set to 200 msec and 20 mV/ div.

Joe Bottiglieri, AA1GW, recently compared some power supplies for *QST*.[17] His review demonstrated that some generate more noise than others. He displayed several output spectral plots of various supplies under load. Many of them generate noise right in the middle of several amateur bands, at levels that can make a difference even to casual radio operators.

So, I set out to determine how bad my QRP power supply really was. Comparing against commercially available units that Joe reviewed, the measurements I made appeared favorable, so why such a visual/audible difference?

I used a simple test setup with my Norcal 40A transceiver connected to a homebrew 50-Ω dummy load and switched the power supply on/off. An additional 12-V dc battery on the power bus allowed more off time, and I sampled the received noise in both cases with the soundcard and my computer.

The results were surprisingly poor. Fig 16 is an example of what that looked like on *Spectrogram*. The central vertical line divides noise generated by the ac supply before and after it was switched off.

By extracting data sets from *Spectrogram* representing before and after the power off moment, I was able to import these into a spreadsheet and examine the differences mathematically.

Spectrogram puts FFT data into frequency bins. These bins can then be analyzed for content. Multiple data sets were extracted from before and after, then averaged across the spectrum to minimize random noise not from the

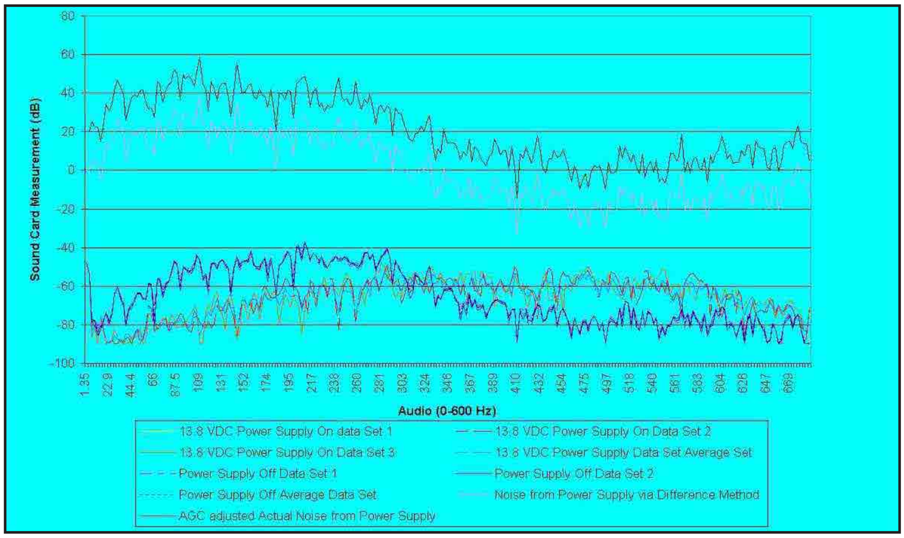


Fig 17—Noise (in decibels) induced in reciever audio from a 13.8-V dc power supply. The light line commencing at 0 dB is the effective noise generated by the 13.8-V dc power supply as calculated by the difference method. The dark line starting at 20 dB is the AGC adjusted estimated value showing almost 42 dB of noise coming from the radio power supply. (See Note 20.)

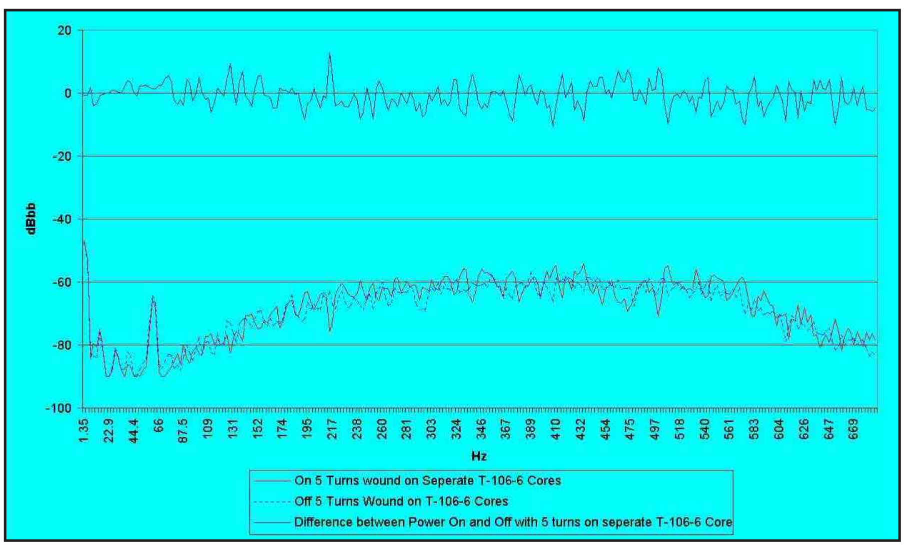


Fig 18—Test data for the power supply with each lead having five turns around a separate T106-6 powdered-iron toroid core. (See Note 20.)

Table 5—Differences between Noise Suppression Tests/Methods

Suppression	*Average of Difference (dB)*	*Standard Deviation of Difference (dB)*	*Approximate Absolute Difference (dB) – AGC Shifting**
None	0.579	16.97	42*
5 turns on T106-6 each lead	–1.11	5.14	12
10 turns on T106-6 each lead	–2.10	4.64	4

*The toroid I thought was counterwound actually wasn't counterwound. As a result, it increased noise from the power supply as shown above. Due to the late night winding error, it is excluded from the results. Winding 10 turns proved best of those tried. It lowered power supply induced noise well enough to allow the receiver noise to drop below the sound-card measurement noise floor below about 170 Hz.

power supply. Typically, I used 10 samples if I did not extract individual sets at random. The two averaged bin sets were then differentiated via simple subtraction to establish the true contribution from the power supply.

Further analysis sums the total of all the blocks, evenly spaced in frequency, and compares the before and after data sets. This comparison indicates average audio power (decibel-hertz) across the sampled spectrum. Because I had not disabled the AGC on the Norcal, it was still active. In the data presentation, I made one bold assumption to overcome the AGC effects. Assuming the audio noise will not decrease with applied RFI, I estimated the amount of AGC correction and shifted the difference accordingly. This demonstrates how some radios may provide generally accurate results even if the AGC cannot be disabled. Use extreme care with any assumptions.

The AGC in the NorCal 40A regulated audio output power to the same level in the spectrum. Because of change in level of noise injected by the power supply, the noise-spectrum signature changed. The difference between the two data sets shows the 13.8-V dc power supply is generating approximately –42 dBr of broadband noise into the received audio sent to the sound card (see Fig 17).

A few data sets are included behind the averaged data set to show the extracted data in various forms. Because it is unlikely that the noise was reduced (negative difference) with the RFI-generating supply on, the data were shifted up to cancel out the apparent effect of the AGC circuit (my ever-so-bold assumption). For most tests, the AGC should be off; but here, it was left on to demonstrate the affect of having it add or subtract gain from the receiver. This would mask a true difference calculation, if left uncorrected.

The interference was ultimately found to be coming down the power supply leads. William Orr, W6SAI, wrote, "The unwanted signal may detected by any active device (tube, transistor or IC) in the audio circuitry and the resultant interference voltage is amplified as a normal audio signal by the rest of the device. Audio RFI is heard regardless of the settings of the amplifier controls or the frequency of the desired reception. It seemingly 'blankets' the receiver."[18]

To suppress the noise, I wound both the positive and negative power supply leads (five turns each) through T106-6 toroid cores. The pair of toroids made a

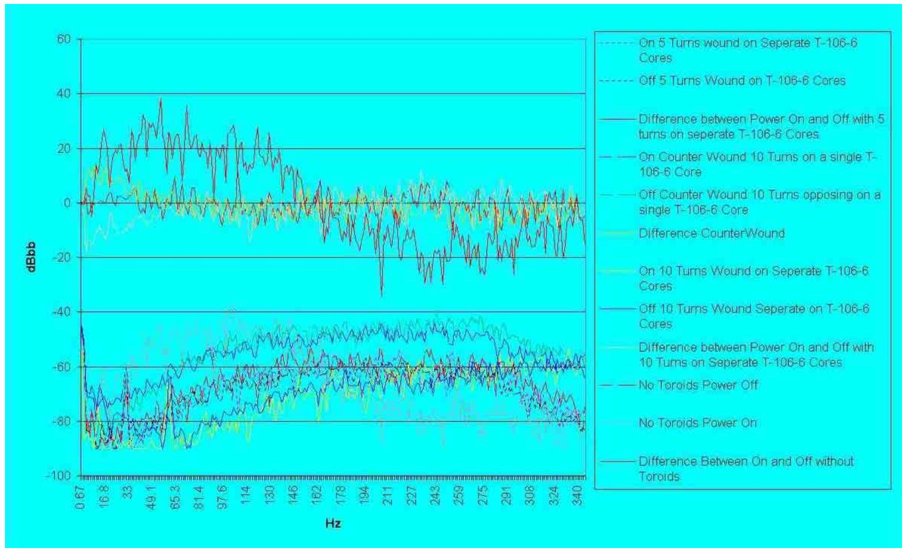


Fig 19—A summary of various measurements and cures used to decouple power-supply noise from the transceiver. Also see the note with Table 5. (See Note 20.)

large improvement as shown in Fig 18. I kept experimenting until finally settling on 10 turns for each lead (Fig 19), each wound on its own toroid, a T106-6 I had left over from an unfinished project. There are other great RFI-reduction references available to the radio amateur.[19] The differences between noise-suppression tests and methods are summarized as shown in Table 5.

The harmonics of 60 hertz (spikes on the flat, bottom trace) were intentionally recorded in Fig 19 to demonstrate what happens if you develop a ground loop on the audio circuit. Many operators use a simple audio transformer for isolation. When an audio circuit is properly isolated as described, the noise floor of the measurement is the noise floor of the sound card, with no harmonics. Some PSK31 Web sites only suggest using an isolating transformer on the transmitted signal coming out of the sound card to your transceiver. I think being able to "see" several decibels better when your weak-signal PSK31 contact is centered over an integral multiple of 60 Hz is worth the investment.

The final proof is in the listening. Now I can comfortably listen to a weak CW signal with the receiver's RF gain at about 55% of its prior setting.

Conclusion

Antenna-comparison measurements can be made accurately if care is taken in the setup of equipment. This method allows for comparison of two antennas and it quantifies the differences between the two. Rather than trying to solicit responses from far-away stations, or infer a signal difference you thought you heard, you can actually measure the difference between two systems. Because it is a difference and because you are keeping known gain factors constant, the procedure produces the desired comparison between the systems under test.

The method is very useful when you cannot be present at the measurement point while switching between systems. It also can be used effectively to identify, quantify and measure a reduction in RFI in the shack from a variety of sources, or to measure antenna improvement against an unchanged reference antenna. Finally, the measurement of atmospheric noise pick-up may yield some interesting insight into why some antenna designs perform better.

Philip Sage holds two degrees from Cleveland State University: one in Physics and another in Electrical Engineering. He was the Manager of Maintenance and Technical Services and Finishing at LTV Steel's Cleveland works until they recently closed, leaving him unemployed.

Phil was first licensed as KA8UHH. He has operated as F/KF8JW/P from France and annually operates an improving Field Day station with K8CAV. Three of Phil's four children participate each Kid's day when he uses Amateur Radio to contact other children and parents.

Notes

[1] J. W. Cooley and J. W. Tukey, "An Algorithm for Machine Calculation of Complex Fourier Series," *Math Computations*, Vol 19, April 1965, p 297.

[2] M. Cook, AF9Y, An example of the spectral map display is included in the article by Steve Ford, WB8IMY, "A Conversation with Mike Cook, AF9Y," *QST*, Jan 1988,

pp 56-57. More information about AF9Y's spectral display program FFTDSP is available at his Web site **www.webcom.com/~af9y/**.

[3]R. Horne, *Spectrogram*, a multipurpose scrolling FFT display program with logging and analytical capabilities. Richard's program *Spectrogram* is available at **www.visualizationsoftware.com/gram.html**.

[4]R. Larkin, W7PUA, "The DSP-10—An All Mode 2-Meter Transceiver using DSP IF and PC Front Panel," *QST* Sep, Oct and Nov 1999. Program *UHFA.CFG.* Contact the author at **boblark@proaxis.com**. Download available at the DSP-10 home-page at **www.proaxis.com/~boblark/dsp10.htm**.

[5]*The ARRL Antenna Book* (Newington: ARRL). I have both the 15th edition (published in 1988) and the most current edition, which includes several useful software tools for optimization of transmission lines and other useful antenna related calculations. For more information, visit the ARRL Web site **www.arrl.org/catalog/**. (The current edition is the 19th, ISBN: 0-87259-804-7, ARRL Order No. 8047, $30.—*Ed.*)

[6]*The ARRL Antenna Book*, p 27-31, "Antenna Measurements."

[7]*The ARRL Antenna Book*, p 6-3, "Indoor Antennas—Empiricism."

[8]By using identical feed lines and matching antenna-input impedances you can eliminate the effects of these components, and accurately compare one antenna to another.

[9]PSK31 Websites: best for overall information is K4ABT's at **www.packetradio.com/**. Others with useful links to information and diagrams are the "Official PSK31 WWW Homepage" at **aintel.bi.ehu.es/psk31.html** or the Digipan Site at **members.home.com/hteller/digipan/**.

[10]M. Reed, DSP-10 Reflector Post 10/2/2000 and subsequent e-mail conversations. See Mike's Web site for many very interesting screen saves at **www.qsl.net/kd7ts/dspindex.html**. (This site requires membership in QSL.net for access.—*Ed.*)

[11]H. Stark and F. Tuteur, *Modern Electrical Communications Theory and Systems* (Englewood Cliffs, New Jersey: Prentice Hall Inc), p 62.

[12]R. Larkin e-mail on DSP-10 AGC circuit description received 9/30/2000. Contact Bob at **boblark@proaxis.com** or see the DSP-10 Web site for further details on this radio.

[13]Wilderness Radio, the NorCal 40A 40-meter transceiver; **www.fix.net/jparker/wild.html**.

[14]Creative Wave Studio was supplied with the AWE 64 Soundblaster Soundcard. Creative, their full line of products and technical support can be contacted at **www.soundblaster.com**.

[15]Richard Horne E-mail correspondence 9/23/2000.

[16]*Spectrogram* help files, Richard Horne.

[17]J. Bottiglieri, AA1GW, recently compared several power supplies in "Product Review: *QST* Compares: Switching Power Supplies," *QST,* Jan 2000, pp 70-73. The article compares several commercially available units.

[18]W. I. Orr, W6SAI, *Interference Handbook—How to locate and cure RFI: Radio Frequency Interference* (copyright 1981, Radio Publications Inc., 925 Sherwood Dr, Lake Bluff IL, 60044, USA; ISBN 0-933616-01-5), p 173.

[19]E. Hare, W1RFI, Editor, *The ARRL RFI Book* (Newington: ARRL, 1998, ISBN: 0-87259-683-4, ARRL Order No. 6834, $20.00).

[20]The many curves in the graphics of this article are much easier to distinguish in color, so I've placed them in a zip file for readers. You can download this package (4 MB) from the ARRL Web **www.arrl.org/qexfiles/**. Look for 1x02sage.zip. *—Bob Schetgen, KU7G, QEX Managing Editor*

An APRS Satellite for Mobile/Handheld Communications

There's a new dedicated APRS satellite in the sky. Come learn what it can do for you.

By Bob Bruninga, WB4APR

This paper describes the design of a low-cost easy-to-build APRS satellite to meet the needs of mobile and handheld amateur satellite users. This APRS Satellite Mission is for worldwide real-time message and position/status data exchange between users. It is in contrast to the mission and design of all existing amateur PACSATs that concentrate on message store-and-forward. Further, it incorporates the Internet as part of its design instead of trying to compete with it. Although we are working on such a satellite at the Naval Academy, we have been urging a number of other satellite owners and designers over the last seven years to accommodate these concepts into their designs as well. The SUNSAT team has been especially receptive of such suggestions and has conducted a number of on-orbit experiments to validate this design. The following list of mission objectives form the basis for this design:

1. Handheld/mobile live digital tracking and QSOs within the footprint
2. Worldwide handheld/mobile position/status reporting (via the Internet)
3. Handheld/mobile message uplink to the satellite (then to Internet)
4. Handheld/mobile message downlink delivery from the Internet
5. Nationwide bulletin delivery to all users
6. Low-power GPS tracking of buoys, telemetry devices, wildlife and so on
7. Other UI (unnumbered information frame) digipeating applications (to be determined)
8. Worldwide one-line E-mail

All of these mission objectives can be met with just a simple hardware TNC on orbit acting as a UI digipeater. In addition, with the sophistication and added I/O capabilities of recent TNCs designed for APRS, the TNC itself can be the command and control system. Thus, no additional on-orbit CPUs are required. Not only is the satellite hardware simple, but it can be easily reproduced by other satellite builders to help form a constellation of these relay satellites. All of the satellites would operate on the same frequency to give mobile users extended access beyond what is possible with one satellite alone. This concept of a Builders Channel for similar-mission space-

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wb4apr@amsat.org

craft was presented at a recent AMSAT Symposium.[1]

Background

Ham radio is on the move. Satellite wireless is the leading edge of technology. In ham radio, it should be a major driver for future amateur-satellite missions. In recent years, we have seen *many* hints at the future of amateur mobile and handheld satellite communications:

- Growing popularity of UI digipeating via MIR through 1999
- Continuing high popularity of AO-27 for handheld FM voice communication
- Activation of UO-14 for FM voice-repeater mode in February 2000
- Experimental UI digipeating via a 9600-baud packet Satellite
- FM voice repeating via SUNSAT SO-35 throughout 1999
- Activation of SUNSAT SO-35 for UI digipeating and APRS
- Recent introduction of integrated TNC/radios
- Dayton 2000 introduction of the upgraded Kenwood TH-D7 data H-T!

[1]Notes appear on page 53.

Fig 1—Charles Richards, W4HFZ's mobile APRS satellite capability (also includes HF). With an APRS satellite, he can send and receive brief text messages anywhere on the planet a few times a day.

The potential of two-way satellite handheld text messaging (national paging) was serendipitously demonstrated at the Dayton Hamvention during a parking-lot demonstration of the SUNSAT down-link. Due to a scheduling error, there was no success at the expected time so the H-T was placed in a pants pocket and forgotten. Minutes later, the tell-tale beeping of the TH-D7 alerted me to an incoming APRS message and on inspection, it was a bulletin from SUNSAT. Thus, amateur satellite message delivery to an unattended, obscured handheld transceiver was demonstrated.

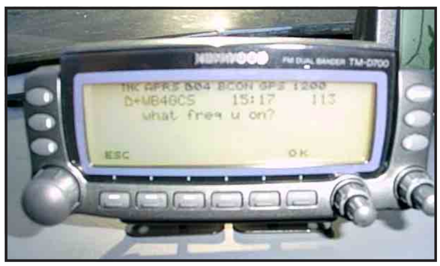


Fig 2—The front panel of the TM-D700 showing an incoming 15 byte message. (Messages can be longer, up to 64 bytes.)

ASTARS

To give this APRS satellite communications system a name, we call it ASTARS, for APRS Satellite Tracking and Reporting System, which has evolved through a number of existing and previous satellite communications experiments. First was 1200-Baud PSK ASTARS, which we called TRAKNET[2] at the 1998 and 1999 AMSAT conferences using AO-16, LO-19 and IO-26. It is a very viable capability for stations with PSK TNCs or using KA2UPWs sound-card up-link capability.[3] Nonetheless, it never became popular due to the rarity of PSK modems amongst most APRS operators.

Satellite packet experiments using 1200-baud AFSK ASTARS, however, which *any* TNC can do, was demonstrated many times during experiments with the *MIR* space station[4] packet system and SAREX.[5] These experiments culminated in the June 1999 week-long experiment via *MIR,* which used the new Kenwood TH-D7 with built-in 1200 and 9600-baud TNCs to demonstrate two-way self-contained APRS communications via *MIR* at 1200 baud. During this test,[6] over 55 stations conducted two-way handheld-to-hand-held message communications.

Recently we have been able to experiment with 9600-baud ASTARS using two satellites and the new Kenwood 1200/9600 baud APRS data mobile radio, the TM-D700A.[7] This dual-band data radio with built-in TNCs and front-panel APRS displays made it possible to send and receive very short APRS-style communications via any 9600-baud PACSAT with digipeat enabled. Thus, the TM-D700 radio is an off-the-shelf satellite data terminal ready for ASTARS, and it needs *no* PC or other accessory. Kenwood also followed suit with 9600-baud upgrades to the TH-D7(G) H-T with its internal front-panel displays. Alinco sells another integrated TNC/Radio called the DR-135, which can also have both 1200 and 9600-baud built-in, though it needs an external computer to display the APRS data.

PCsat in Space Now . . .

PCsat was launched at 0240z 30 Sept 2001 from the Kodiak Alaska Launch Complex. After some initial checkout, it was enabled for users on 3 October providing a new service to the mobile satellite user. All indications are that it is healthy and all systems on board are fully functional except for one –Z solar panel.

Users from around the world report good success hitting the satellite with only a 5-W H-T and a full size whip. With its 800-km orbit, *PCsat* has relatively long 12 to 15 minute passes and covers an entire continent simultaneously. The success of this up-link, however, is completely dependent on collisions and QRM on the frequency. Although we cannot control the QRM from non-amateur interlopers on the 2-meter band in some countries, we can ask all users to respect other users and limit their up-links to the intended mission of this satellite.

Because of the unique mission to provide amateur satellite communicaitons for travelers with only an H-T, *PCsat* has taken a bold step in the Amateur Satellite service by publishing and asking users to follow the operating guidelines in a written User Service Agreement (**www.ew.usna.edu/pcsat**). In summary it says:

PCsat is open to radio amateurs worldwide and anyone is welcome to use *PCsat* as long as they share the asset with others and follow the guidance in the User Service Agreement and *PCsat* bulletins. In general, this means operating with a low-duty-cycle commensurate with user precedence. Of course, emergency and priority traffic have precedence, and so do demonstrations. These guidelines are lessened during the period 2300 to 0800 local time everywhere to allow for experimentation as needed.

The downlink from *PCsat* as received around the world is fed into the worldwide APRS network and can be viewed on **pcsat.aprs.org**. We are only beginning to tap the potential of this mobile satellite service for ham radio travelers everywhere. Stay tuned—*de WB4APR, Bob*

The Internet

Unlike previous amateur satellite designs, an APRS satellite can capitalize on the connectivity of the Internet rather than competing with it. The Internet makes it possible to link together multiple disparate down-link sites, which allows a tremendous gain in reliability through space and time diversity reception. Normally, each station requires a down-link receiver and only hears packets within its own footprint. The Internet allows a few stations, called SAT-Gates (satellite I-Gates) to combine all packets heard into the existing worldwide APRS infrastructure (APRServe)[8] for delivery to any APRS operator, anywhere.

APRS Messages

Satellite operators unfamiliar with APRS messages should understand that an APRS message is a single *line* of text. Most messages stand alone, but they are occasionally strung together if the information won't fit on one line. Fig 2 is a photo of a very brief 15-byte message received on a TM-D700 radio. Messages from mobiles are usually quite brief, as they must be entered via the DTMF pad. Nevertheless, longer messages—up to 64 bytes—are routinely displayed.

Table 1—Decoded "One-Line" Mobile E-mail Message

Date: Mon, 7 Feb 2000 07:58:09 -0500 (EST)
From: WB4APR-9@unknown.net
To: wb4apr@amsat.org
Subject: APRS Message from WB4APR-9

---

testing delivery via pacsat from my van en route to work.

Message received by MacAPRS I-Gate station WU2Z
Located in NO BRUNSWICK, NJ
APRS path = WB4APR-9>APK101,SUNSAT*:

E-Mail

Similarly, the APRS messaging system can send and receive standard e-mail messages worldwide via the established worldwide APRServe Internet system. This capability is limited but very useful. The first limitation is that messages are *only one line* and the one line *includes* the full e-mail address. This forces *brevity*! Secondly, although e-mail can be originated under the control of the ham sending it, e-mail replies back from the Internet are only allowed via special I-Gates. Their operators have volunteered to screen such traffic for third-party legality before returning it to the RF path. To demonstrate, I transmitted an e-mail from my D700 mobile en route to work. The *single* packet entered into the D700 was simply: "E-mail **wb4apr@amsat** testing delivery via pacsat from my van en route to work." Yet, it was received by my fixed e-mail system after being SAT-Gated to APRServe and from there, picked up by the e-mail engine at WU2Z and shipped out as regular e-mail. Table 1 shows what I received.

User Ground Station Equipment

To design an APRS satellite, we must fully understand the capabilities of the users' mobile stations. Table 2 below shows the up-link power and receive-antenna gains for all participating stations in the ASTARS system. The column labeled "RX STBY dBi" is for users who are unfamiliar with and whose stations are not optimized for satellite reception: for example, someone hiking with an H-T in his pocket or a mobile parked under trees.

Although a wide range of power and receiver gains is involved, these values are what form the basis of the APRS satellite design and the architecture of the overall ASTARS system.

Requirements/Constraints/ Design Drivers

To design a satellite to meet the H-T/mobile communications objective and the Internet links as well, there are a number of factors involved in selecting the frequency band, antenna types, and baud rates for each of the mission objectives. First, there are a number of boundary conditions or assumptions:

- Optimum ALOHA channel efficiency is about 20% due to collisions
- VHF links have a 9-dB advantage over UHF links (using omnidirectional antennas)
- 1200-baud AFSK has a 7-dB advantage (measured) over 9600-baud FSK
- TR delays render 9600 only twice as fast as 1200 for APRS bursts
- UHF up-links require wide-band satellite receivers to handle Doppler-shifted signals (–4 dB)
- UHF down-links require user tuning during passes (not desired)

With these design drivers as a guide, the following are some of the obvious first-order alignments of hardware requirements. From these, then, we need to determine the optimum tradeoffs to arrive at our final design.

- Message delivery to an H-T in standby mode requires the best-possible down-link (1200-baud VHF). The I-Gate up-link is relatively unconstrained.
- Message receipt from an H-T requires best possible up-link (1200-baud VHF)

Table 2—Ground-Station Equipment

	ERP UHF (W)	*ERP VHF (W)*	*RX UHF dBi*	*RX VHF dBi*	*RX STBY dBi*	*Applications*
User Stations						
H-Ts	3	5	3	3	–6	Sailboats, hikers, wilderness
Mobiles	70	100	5	5	–6	Remote travelers, boats
Home Stations	700	1000	13	13		Not intended for up-link
Network Stations						
I-Gate Receiver			7	5		Omni Internet receive site
Message Node	70	100				Internet to user up-link site
Command Station	700	1000	13	13		US Naval Academy

- Down-link to the Internet is relatively unconstrained
- Continent-wide bulletin delivery requires the existing 144.39-MHz frequency over the USA and 1200 baud. The same service for Europe will require a common European frequency, too.
- H-T/mobile real-time messaging requires the same up-links, down-links and baud rates
- GPS H-T/mobile tracking is relatively unconstrained
- Low-power GPS tracking devices require the best up-link possible (1200-baud VHF), and the up-link must not be used by any other satellites, to avoid unintentional interference to other systems
- Other UI digipeating applications should be cross-band full-duplex circuits and should use same up/down-link baud rates
- It is desirable to spread up-links among multiple receivers to minimize collisions
- Synchronization of same-band down-link transmissions is desirable to maximize the available half-duplex satellite receive time
- Redundancy and backups are desirable
- Bundling of packets in bursts amortizes individual transmit delays
- UHF down-links are of little value due to poor link budget and Doppler effects
- The KISS principle should reign (Keep it Simple, Stupid)

Hardware Alignment to Requirements

Using these criteria, we arrived at a design using two KPC-9612 dual-port TNCs. These TNCs have all the latest APRS generic digipeating advantages and can even cross-route packets between ports. By using standard off-the-shelf TNC hardware and firmware, we have minimized risk on orbit, relying on the dependability of identical hardware in use all across the country for terrestrial APRS. Thus, the firmware is proven.

Each dual-port KPC-9612 can cross-relay from either of its two inputs to its two outputs. Since we only have two transmitters on VHF for best down-link budget, we must output both the 1200 and 9600-baud channels to the same transmitter, one for each TNC as shown in Fig 3. With four ports, we need a single VHF half-duplex channel in the ITU Satellite Subband and one other possible VHF up-link. The UHF up-links are more readily available and are not expected to be an issue. Notice that the other VHF down-link over the North American continent will use 144.39 MHz for down-linking occasional bulletins or directed messages to distant travelers. Thus, they can receive urgent messages from the satellite at any time, while also monitoring

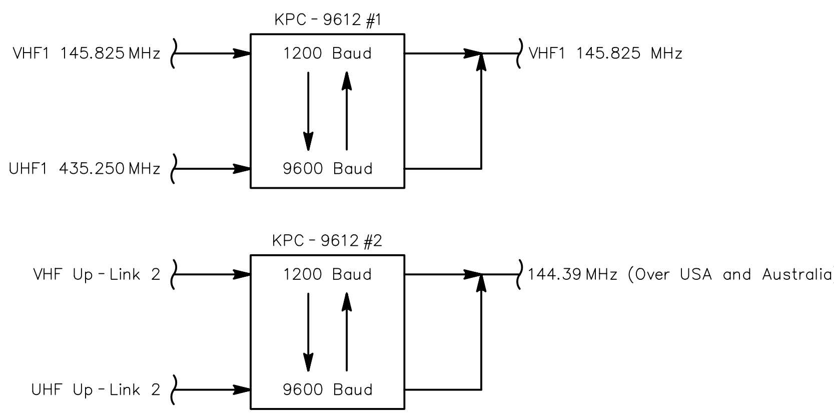


Fig 3—Satellite TNC allocation.

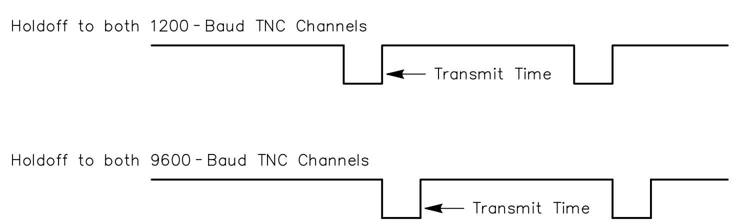


Fig 4—Satellite channel timing.

Table 3—Channel Usage and Mission Scenario

Mission Element	*Uplink*	*TNC Path*	*Downlink*
H-T Up-link of messages or positions to Internet	VHF1	UIDIGI	VHF1
Live H-T to H-T footprint QSOs	VHF1	UIDIGI	VHF1
Live H-T to mobile cross-links	VHF1	MYgate	VHF1@9600
Live Mobile to H-T cross-links	UHF1@9600	MYgate	VHF1
Mobile up-link of messages or positions to Internet	UHF1@9600	UIDIGI	VHF1@9600
Live mobile-to-mobile footprint communication	UHF1@9600	UIDIGI	VHF1@9600
Voice relay	UHF1	bypass	VHF1
Command and control	ALL	MYRemote	VHF1
Other UI applications	TBD	UIDIGI	VHF1
Low power trackers	VHF2	UIDIGI	144.39 MHz USA
Nationwide message delivery	UHF2@9600	MYgate	144.39 MHz USA
Nationwide bulletin delivery	UHF2@9600	MYgate	144.39 MHz USA

the terrestrial channel when in range of the terrestrial network.

To maximize receive (up-link) time, a cycle timer drives the channel-busy inputs of each of the four TNC channels. By holding off both transmitters for *N* seconds and then allowing them to both transmit simultaneously, we minimize the VHF transmit time and thus maximize VHF receive time. We also bundle multiple packets into one burst, thus amortizing the transmit delays across several packets for further savings. UHF receive time is unaffected. This channel synchronizing will be done with a two-step timing circuit as shown in Fig 4. When each of these signals goes low (allowing the transmitter to key) any packets pending will be transmitted. If there are none pending, no transmission occurs.

Failsafe Reset

Since we are using commercial off-the-shelf Kantronics TNCs as our only on-orbit CPU or command processor, we must have a way to reset them in case of a lockup condition. First, we get special ROMS from Kantronics with all of our default parameters burned in. Second, we integrate each TNC with a failsafe circuit. These circuits monitor the PTT of each TNC and as long as a transition occurs at least once a minute, the TNC is assumed to be operating correctly and the TNC remains powered up. If there are no transmissions for over one minute, then a one-shot timer removes power from the TNC for one second to allow for a complete power up reset of the TNC.

The one-minute baseline is established as the transmit rate for our telemetry packets. Thus, as long as the TNC is operating normally and transmitting packets, then it will not be reset. There will also be an unpublished backdoor backup reset command.

Telemetry

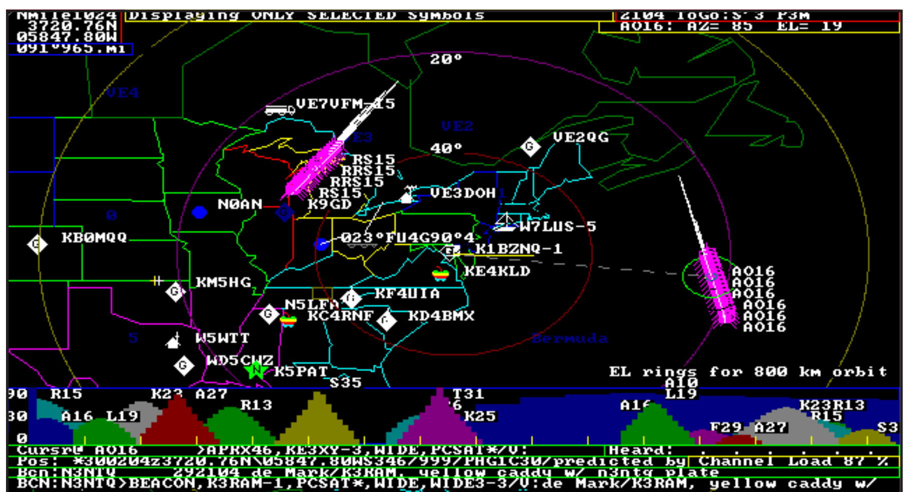
Based on the APRS telemetry formats that we established back in 1995, Kantronics has added at least four channels of analog telemetry to all of their recent TNCs. To make this usable on our satellite, we have added a 16-channel hardware multiplexer to permit reading as many as 16 values of telemetry per TNC. These APRS telemetry formats were used on STENSAT and more recently adopted for APRS modes on SUNSAT, too.

Each APRS telemetry packet will have five analog values: four telemetry channels, and the fifth is the value of the 5-V reference.

Link Budget

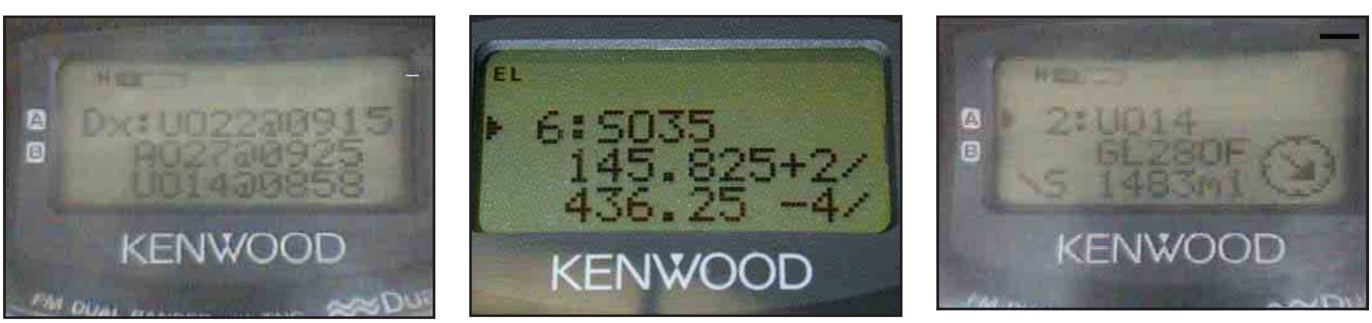
The primary driver of this APRS satellite design was to deliver messages to handheld transceivers and mobiles with only whip antennas. To do this, we will have a down-link that is at least 12 dB stronger than most existing digital satellites. We do this by taking advantage of the 9-dB link improvement of 2 meters compared to 70 cm and we use a 2-W transmitter. Further, our satellite will operate at a low transmit duty cycle. This is unlike all existing PACSATs, which are required to operate with low power budgets so they can keep their transmitters on 100% of the time whether the satellite is in use or not. Because the worldwide ham-radio population only covers 10% of the Earth's surface and considering the low duty cycle of the ALOHA style of APRS operations, less than 4% of our average power budget is required for each transmitter.

Similarly, to conserve link budgets and bandwidth, we reserve the 2-meter up-links for only the low-power stations. These include users with handheld radios, stand-alone low-power tracking devices, data collection buoys or remote weather stations such as the one built by Ronald Ross, KE6JAB, in Antarctica.[9] The mobiles and SAT-Gates, which have 35 to



All satellites are shown on the map as moving objects. Across the bottom of the screen the next 2.5 hours of satellite passes are shown in a graphic showing the maximum elevation of the pass.

Fig 5—The page at www.ew.usna.edu/~bruninga/satinfo.html.



(A) (B) (C)

Fig 6—These screen shots show what the TH-D7 will capture and display about the satellites while monitoring the terrestrial network if an *APRSdata.exe* station is in range. At (A) the DX-spot list shows that there are three satellites UO22, AO27 and UO14 coming up in the next 80 minutes and when they are available. The next two screens (B and C) show when the satellite is in view. They show the range, azimuth frequency, Doppler shift and distance to the satellite.

50-W transmitters, will be asked to operate only on the UHF up-link frequencies, where they can afford the more difficult link budget. Thus, we also have the further advantage of having spread out the user base over four up-link channels to minimize collisions.

Channel Usage and Mission Scenario

Table 3 maps the mission objectives into the various up-links and down-links on the satellite. It matches strengths and weaknesses of each mission area to the available link budgets and hardware.

Notice the advantage of incorporating the single North-American-continent-wide coordinated APRS frequency into the down-link frequency plan. Although this frequency is in use by over 2000 users full-time including over 600 wide-area digipeaters, it is an established universal frequency where *all* APRS operators can be found. Normally, mobiles cruise with their radios tuned to this frequency whether they are in range of the terrestrial network or not. Actually, although 90% of the US ham population is within range of this terrestrial infrastructure, 80% of the landmass is not.

Because of the shared use of 144.39 MHz with the thousands of existing users, this down-link on 144.39 will *only* be used for the special applications consistent with the national significance of this channel. Such applications might be:

- Getting an emergency or priority message to an existing APRS mobile regardless of location
- Infrequent bulletins of national interest.
- Low-power, high-profile tracking of special devices, for example, the Olympic Torch

Due to the low duty-cycle channel statistics of an ALOHA TDMA channel like APRS, even though the channel is in full use by thousands of users, it is still clear more than 50% of the time, as heard by any mobile anywhere at any instant.

Operations Scenario

To develop a viable satellite communication system that can communicate between mobiles and to/from the worldwide APRServe system within these limitations, the following operating scenario is recommended.

- Mobile APRS stations use a -3 SSID (secondary station ID) so the SAT-Gates know to deliver such traffic via the satellite and not to the home or terrestrial stations.
- The mobile up-link objective is two position/status successes per pass. This can be met with a one-minute rate.
- Mobile up-link of a few outgoing messages per pass using the built-in Kenwood one-minute message rate is about right.
- SAT-Gate will transmit return messages during the central three minutes, when the satellite is at its closest distance to an intended station.
- Mobiles with beam antennas can indicate to the SAT-Gate to try longer in the pass by including "QRZ" in their STATUS, or by indicating CUSTOM MSG 3, which is easier to set from the DTMF pad.

SAT-Gate Operations

The mobile-to-mobile and H-T to H-T communication missions work without any special considerations on the satellite or on the ground. Yet, the more useful application is sending and receiving messages to any other APRS station, worldwide, as packets are received by the SAT-Gates that monitor the satellite down-link frequency and every packet heard is fed into the APRServe system. These SAT-Gates perform the following functions:

- Monitor both down-links and feed *all* packets into the Internet
- Maintain a track on all call signs heard via satellite
- Monitor the Internet and capture messages for these call signs
- Deliver these messages at a "fair" rate under these conditions:

1. The satellite is within 1400 km of the mobile (above 30° elevation)
2. It sees "QRZ" in the Mobile's STATUS text or CUSTOM-3
3. Deliver these messages until seen in the down-link three times

Omni No-Track SAT-Gates

Setting up a SAT-Gate is trivial, requiring nothing more than a normal packet station and an omnidirectional antenna. Any APRS station can do it with existing software that contains the built-in I-Gate capabilities. Even if the station does not have horizon-to-horizon coverage, they are only contributing their packets to the same stream as all the other I-Gate receivers, so any station can help. Unlike any previous amateur satellite activity, we use the Internet to combine the outputs from a dozen such stations nationwide and the result is over a 99.96% chance of capturing every packet over the USA! Even if only four stations at any one time have the bird in view of their station and even if they only have a 60% chance of decoding each packet, their combined probability is 98%. Nevertheless, if the original packet is replicated *twice*, then this probability becomes 99.96%—a certainty!

Base-Station Operations

Since the APRS satellites are shared assets with limited bandwidth, we only want to encourage this message system for use by mobiles, who have no other means to communicate from distant locations. For this reason, we do not encourage base-station operations other than SAT-Gates or for direct contact with a mobile if needed. A *Mic-E* style packet from the D700 is only nine bytes long, compared to a typical WinAPRS 80-byte position report.

Base-station transmissions are strongly discouraged. Stations that use software other than Mic-E should operate only in APRS compressed mode and minimize their STATUS text. Beams on the up-link are also undesirable for this application. To use this mode, we must keep each station's ERP to 50 W with an omnidirectional antenna—or equivalent—to give everyone equal access.

Satellite Tracking and Pass Predictions

With dedicated APRS satellites, we will have less of a problem with QRM and congestion on the up-links than we have had with our experiments on some of the shared-access birds. In any case, it is nice to know when a satellite will be in view to carry your message traffic. To this end, I have added satellite tracking to APRSdos in the form of *APRStk.exe*. When run within an existing APRSdos file structure (so you get all the maps and other built-in-data), it presents the satellite predictions on the APRS map and will auto-tune the Kenwood radios—including Doppler effects. It is available as a zipped file containing a complete system for download from: **ftp://tapr.org/aprssig/dosstuff/APRSdos/aprstk.zip**

Distributing Live SAT Tracking Data to Mobiles

Another version of the same APRSdos derivative is called *APRSdata.exe*. It has the unique feature that it can distribute (via the terrestrial network) sufficient pass information so that other travelers are aware of pass times long before they drive out into the wilderness. Fortunately, we have an excellent way to display this special satellite-pass information directly on the mobiles' radios. Not only can *APRSdata.exe* post data about the satellites in view as objects to the local 144.39-MHz network for mobiles, but it can also transmit the schedule to the TH-D7 or TM-D700's DX-spot list for future reference. Thus, mobile satellite users can get the pass information they need without lugging along a laptop (see Fig 6). This information is perfect for aiming a handheld antenna. For more details on this resource for non-PC distribution of satellite information, visit **www.ew.usna.edu/~bruninga/satinfo.html**.

The power of this on-line, real-time delivery of current satellite-pass data to mobiles and handheld users without the need for a laptop is, in itself, a brand new opportunity for Amateur Radio. Already we have expanded it to hundreds of other data screens that we can push to these radio displays. We call them Tiny Web Pages.[10] Although the application is beyond the scope of this paper, remember that we can deliver Tiny Web Pages to any H-T or mobile anywhere on the planet with the combined resources of the existing APRS infrastructure and the future APRS amateur satellites.

Conclusion

The time is ripe for extending our amateur satellite digital communications services to mobile and handheld users. Since packet was first introduced on the Space Shuttle mission STS-35, there have been numerous experiments to test and validate the capability for using UI packet digipeating for real-time digital communications between users. Instead, the mainstream use of satellite packet matured into the PACSAT store-and-forward protocol to meet the immediate need and higher demand for efficient worldwide bulk-message communication. Yet more recently, the Internet has also matured as a global resource for exchanging data worldwide. This obviates much of the appeal of such an amateur constellation of store-and-forward satellites.

Now, however, we have a unique opportunity to join the advantages of the Internet and amateur satellites. Together, they are a means of tying together ground stations throughout the world where the infrastructure exists to extend worldwide amateur communications to mobiles in areas where it doesn't exist. We need not start such a global system from scratch. The APRS protocol and worldwide Internet infrastructure provide a means of packaging, delivering and displaying this type of real time traffic to users both on satellite down-links and worldwide via the Internet.

Finally, the introduction of the Kenwood and Alinco integrated TNC/radio combinations give us off-the-shelf mobile and handheld satellite communications terminals to all users. We hope to use UI digipeating APRS satellites to bring all of these pieces together into the most powerful and far-reaching Amateur Radio satellite project to date.

Bob Bruninga was first licensed in 1962. He graduated from Georgia Institute of Technology in 1970 with a BSEE. He earned an MSEE from the Navy post-graduate school in 1971. Bob served 20 years in the Navy as a Combat Systems and Communication Engineer. He is a registered Professional Engineer. He now serves as the Chief Engineer for the US Naval Academy Satellite Ground Station. Bob wrote the first Amateur Radio RTTY BBS software in 1978 and the first dual-port VHF/HF packet BBS software for a Commodore Vic-20 computer in 1983. Bob developed APRS in 1992 and presented it at the ARRL Digital Communications Conference in Teaneck, New Jersey.

Notes

[1] R. Bruninga, WB4APR, "AMSAT Builders Channel for HT and Mobile Satellite Communications," *Proceedings* of the 17th Annual AMSAT Symposium, pp 191-195.

[2] **www.ew.usna.edu/~bruninga/traknet.html**

[3] **members.aol.com/dquagliana/upw/index.html**

[4] **www.ew.usna.edu/~bruninga/mirex.html**

[5] **www.ew.usna.edu/~bruninga/sarex.html**

[6] **www.ew.usna.edul/~bruninga/mir-ht.txt**

[7] **www.ew.usna.edu/~bruninga/TM-D700A.gif**

[8] **www.ew.usna.edu/~bruninga/aprs.html**

[9] R. Ross, KE6JAB, "Return to Antartica: A Digital Satellite Journal," *AMSAT Journal*, March-April 2000, pp 6-12.

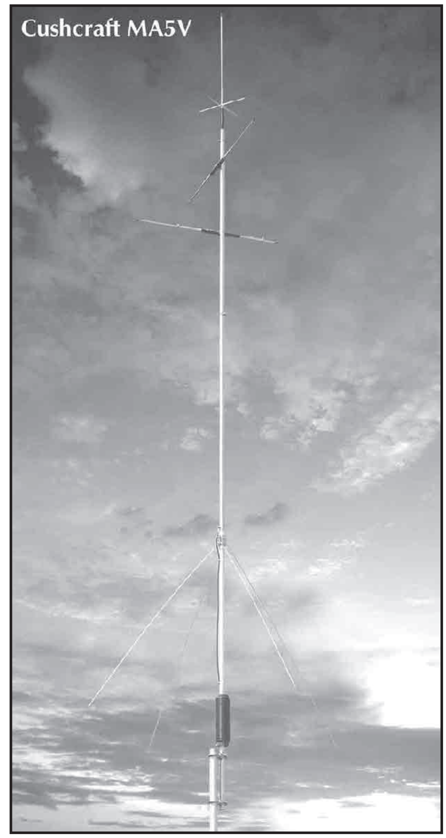
[10] R. Bruninga, WB4APR, "APRS Tiny Web Pages," *Proceedings* of the ARRL/TAPR Digital Communications Conference, 2000, pp 13-20.

Tech Notes

Conducted by Peter Bertini, K1ZJH

This Tech Note is from Rick Littlefield, K1BQT, of Cushcraft Corporation. Rick describes a compact ground-independent vertical that offers low SWR on six bands without the use of external matching circuitry or "matching boxes." Instead, the impedance transformations are performed by the antenna's multiresonant element using off-center-fed dipole (OCFD) matching techniques, resulting in a simple, low-cost, lightweight design that can be installed just about anywhere. Rick may be reached at ***k1bqt@aol.com****, or at 109A McDaniel Shore Dr, Barrington, NH 03825.*

A Compact Six-Band Off-Center Fed Vertical Dipole



OCFD impedance matching is based on the principle that the driving impedance of a half-wavelength dipole is lowest at mid-element and increases progressively as the feed point is moved toward either end. As shown in Fig 1, the rate of change is very gradual near the center and increases exponentially toward either end. *EZNEC* predicts that it is possible to obtain a nonreactive drive point ($X = 0$) anywhere between the antenna's center and the final 10% of either leg.

OCFD matching techniques may be applied to any dipole, including low-impedance Yagi driven elements and portable antennas using lumped loading. When a dipole is shortened and one or more forms of loading are applied to restore resonance at the original frequency, the driving impedance between the loading components becomes lower at all points (Fig 2). Greater percentages of loading yield lower impedances.

Off-center feed represents only one

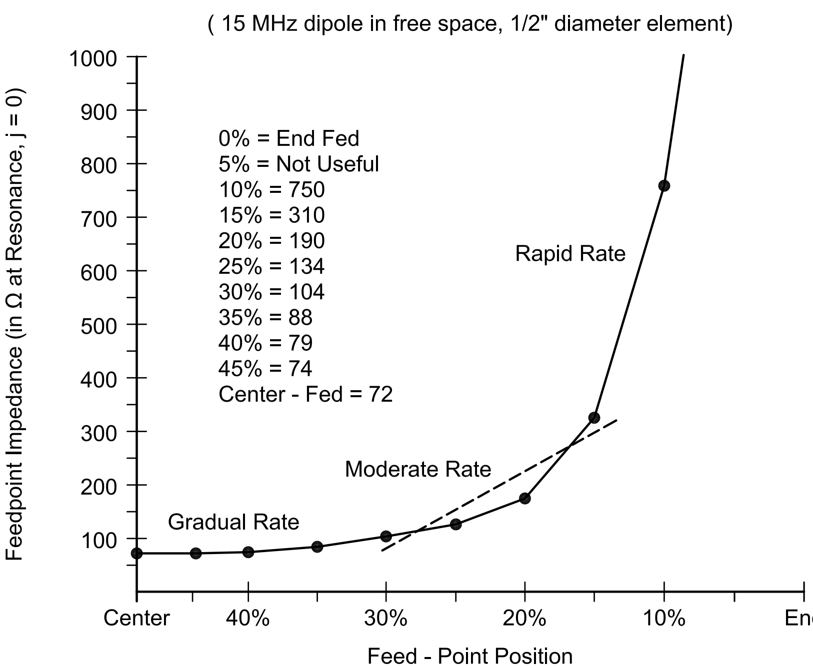


Fig 1—Driving impedance increases gradually near the dipole center, but accelerates exponentially near either end.

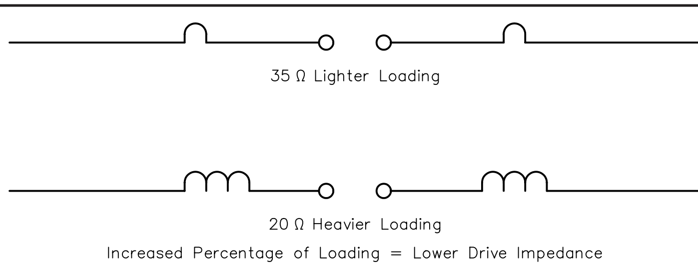


Fig 2—A higher percentage of loading yields lower driving impedances at all points along the element between the loading components.

of many ways to match a low-impedance antenna to 50-Ω line. Other more traditional choices include broadband transformers, hairpin, gamma, and T matches (Fig 3). In some applications, off-center feed may be a low-cost alternative to other methods.

It is relatively simple to combine two or more elements that are resonant at different frequencies as long as they all have the same drive-point impedance. However, if each radiator has a different percentage of loading and presents a different driving impedance, the task of combining becomes more complex. Such is the case shown in Fig 4, where the lowest-frequency radiator uses considerable loading to achieve resonance within the antenna's size constraint, while the highest-frequency radiator requires no loading at all.

To resolve this dilemma, I established a fixed length for one dipole leg and used it in common on all bands. I then reconfigured the other leg as two frequency-selective resonators—one tuned for the highest-frequency band of interest and the other for the lowest. The result was a dual-band antenna where the lower-frequency radiator functioned as an OCFD.

A Dual-Band Antenna

The specifics of the revised configuration are illustrated in Fig 5, where the common element consists of four drooping radials fanned at 45°. On six meters, this structure functions as a set of independent, λ/4 counterpoise wires. On 20 meters, it functions more as a single capacitively loaded wire with an electrical length of roughly 0.11 λ (equivalent to approximately 86 inches of wire at 14 MHz).

This common leg provides the essential compensating mechanism needed to achieve multiband operation. Because the leg's electrical length is different for each band, it follows that the electrical location of the feed point is also different. Therefore, the higher-frequency unloaded radiator is fed at its center where the driving impedance is relatively low. At the same time, lower-frequency loaded radiator is fed well off center, where the driving impedance is relatively high. When the 20-meter OCFD radiator is proportioned to yield 50 Ω at its feed point, the antenna will exhibit favorable SWR on both bands.

Although SWR has little impact on an antenna's capability to radiate RF, achieving low SWR at mid-band becomes crucial when a loaded radiator with compressed bandwidth is fed as part of a 50-Ω system. For the Amateur Radio Service, we typically define "usable bandwidth" as that frequency span where an antenna's SWR measures 2:1 or less. As shown in Fig 6, lower minimum-SWR levels at mid-band yield a wider usable bandwidth.

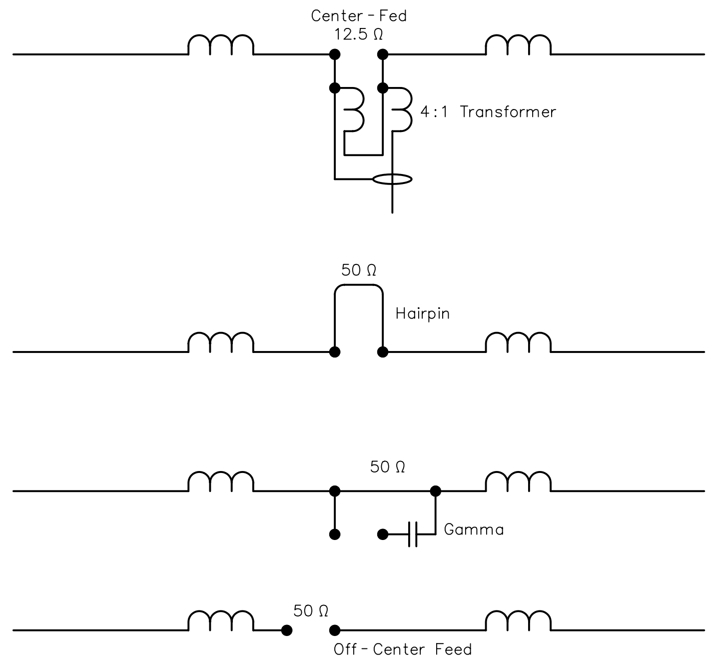


Fig 3—Typical strategies for matching low-impedance loaded elements.

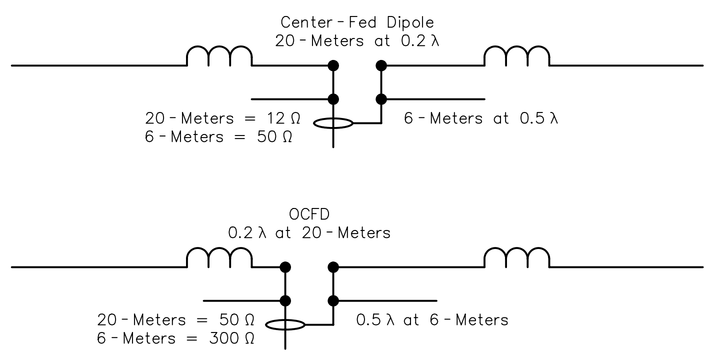


Fig 4—When one element is loaded and one is not, the driving impedances are dissimilar and the two elements cannot be readily combined at a common feed point without compensation.

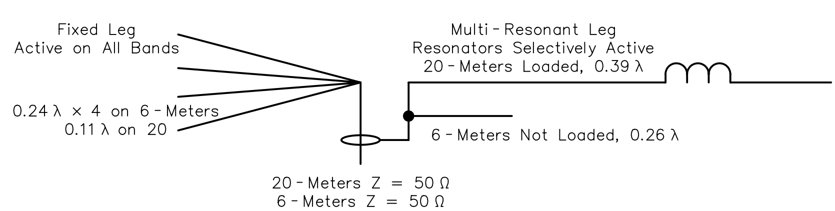


Fig 5—When one dipole leg is common, the element presents a different electrical feed point on each band.

Tuning the dual-band antenna for minimum SWR on six meters is easy—it takes only a simple length adjustment. However, optimizing the loaded 20-meter OCFD radiator is more involved because factors such as percentage of loading, positions of the loading components and inductor Q all influence the feed-point impedance. While a pencil-and-paper plan for a prototype resonator may be derived through calculation or even by a well-placed "SWAG," modeling the design on *EZNEC* provides a better method for approximating the initial configuration. Once this model is converted into aluminum, the strategies outlined in Fig 7 may be used to fine-tune it for operation in the real world.

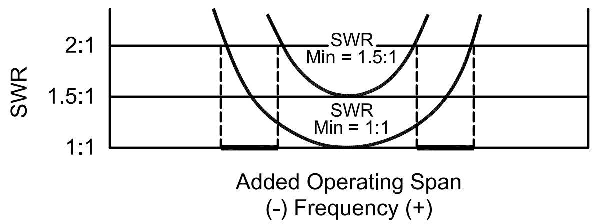


Fig 6—Optimizing a loaded antenna for minimum SWR improves the 2:1 SWR bandwidth.

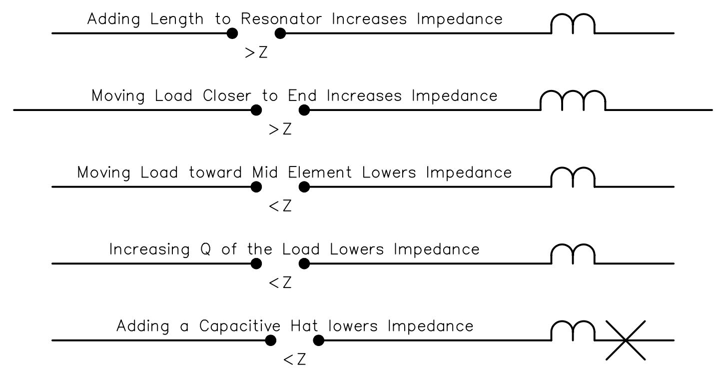


Fig 7—Strategies to adjust the feed-point impedance for individual resonators.

A Multiband Antenna

The six and 20-meter dual-band antenna I've described provides a solid platform for building a more complex multiband design. For the added bands, I elected to install new resonators at strategic points along the existing 20-meter leg, as illustrated in Fig 8.

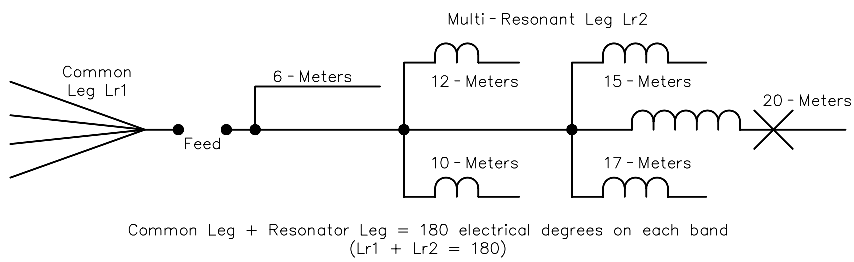


Fig 8—Added resonators permit coverage of all amateur bands between six and 20 meters.

For each new band, the electrical length of the common leg plus the band resonator must approximate 180°. The antenna's upper frequency limit is realized when the fanned leg becomes fully segmented (electrically) into independent $\lambda/4$ wires. The lower-frequency limit occurs when the fanned element becomes too short electrically to support a workable impedance transformation at the feed point. The concept of a fanned common element is nothing new. Similar appendages are found on many elevated-feed verticals under the mantle of "ground radials," "counterpoise" and "feedline decoupling rods." However, when analyzing the function of these structures over a wide frequency range, I find it more useful to view them as one leg of an OCFD.

For structural simplicity, I wound the 10 and 12-meter resonator inductors on opposing ends of a single elongated fiberglass form and installed a 90° mounting clip at the center. I then installed telescopic end sections to provide tunable resonator tips. The inboard resonator length (for the pair) was adjusted by the positioning of the form on the 20-meter leg. This approach was repeated for the 15 and 17-meter resonator pair, but the assembly was positioned higher to increase the element length from the feed point to the coil. It was also rotated 90° in the horizontal plane. With

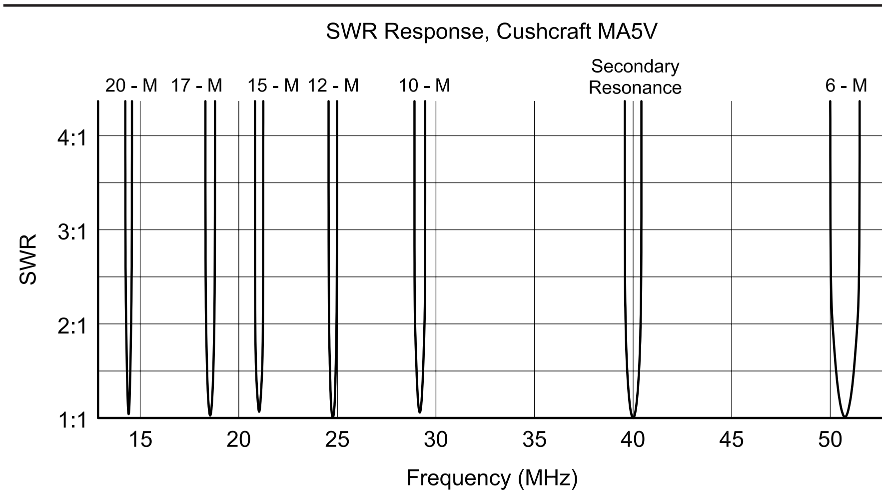


Fig 9—The completed prototype yielded low mid-band SWR readings on all six amateur bands between 14 and 54 MHz.

this configuration, every loading coil and resonator tip is oriented either 90° or 180° off-axis from all others, a strategy used to minimize stray coupling and reduce resonator interaction between bands. During optimization, each resonator mounting position, loading inductance and tip length was manipulated to proportion the antenna for best midband SWR using the techniques illustrated earlier. SWR performance for the completed prototype when mounted 35 feet above ground is shown in Fig 9.

Special Design Considerations

One primary disadvantage of the asymmetrical OCFD element is its tendency to interact with the feed system. To prevent unwanted common-mode currents, a high-impedance wound coaxial choke was installed where the feedline departs the element (in this case, in line with the tips of the fanned leg rather than at the feed point). In addition, because the fanned leg extended below the feed point, a short length of insulated mast was installed to prevent conductive coupling to the tower or support mast (Fig 10).

Coil losses play a key role in determining the efficiency of low-impedance loaded radiators (a topic for an upcoming Tech Note). While considerations such as encapsulation method, mechanical structure and cost may preclude specifying silver-plated #8 wire on an air-wound form, it is nonetheless important to use the largest-gauge wire and best shape factor practical for the antenna's intended power level. Also, in keeping with loading-coil research conducted by Jerry Sevick, W2FMI, coils should be well-separated from adjacent aluminum tubing to prevent inadvertent Q degradation. Failure to maintain sufficient spacing frequently degrades the Q of otherwise well-designed loading inductors.

Finally, because loaded radiators exhibit bandwidth compression, it is helpful to increase the apparent diameter of the radiator as much as possible, this as a strategy to preserve operating span. For this design, the element's length-to-diameter ratio was improved (decreased) by fanning the shared leg, appending resonators perpendicular to the main leg and using relatively large element tubing. Some added bandwidth may also be recovered by using the padding effect derived from normal feedline losses. Toward that end, it may prove more beneficial to use RG-58 for a 50-foot coax run and live with 1 dB of cable loss rather than install LMR-400 and lose 50 kHz of band coverage.

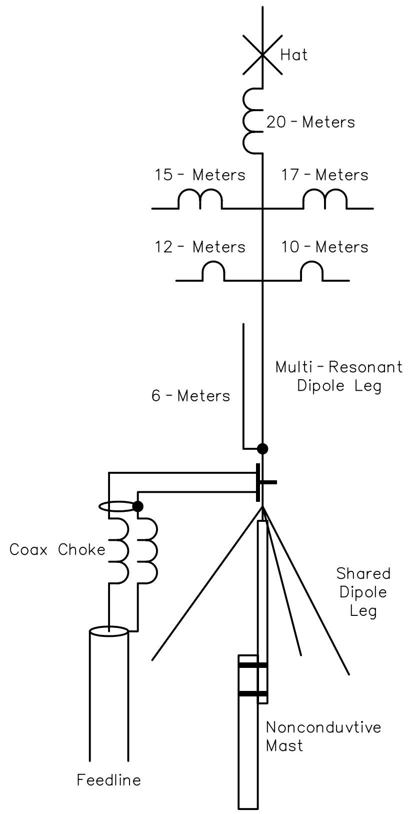


Fig 10—A coax-wound common-mode choke is essential to decouple the antenna from the feed system. Coupling length between the drooping leg and the support structure is reduced by a of nonconductive mast.

Conclusion

This article outlines a unique approach for constructing a compact multiband vertical antenna that requires no external matching. The design discussed here has been thoroughly tested and is presently being implemented in at least one commercially manufactured product. I encourage amateurs to explore and experiment with other potential applications. Thanks to Roy Lewallen, W7EL, and Steven Best, VE9SRB, for their supportive technical review and many helpful suggestions.

References

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R. Straw, N6BV Ed, *The ARRL Antenna Book*, CD version 2.0, (Newington, Connecticut: American Radio Relay League, Newington, 2000).

R. Littlefield, K1BQT, "Build A 20-Meter DX-Pole Antenna," *Communications Quarterly*, Spring 1997, pp 98-102.

R. Littlefield, K1BQT, "An Ultra-Light Center-Fed Vertical Antenna for 20 Meters," *Communications Quarterly*, Winter 1994, pp 89-94.

J. Sevick, W2FMI, *Building and Using Baluns and Ununs*, (CQ Communications, Inc, 1994) pp 111-119.

RF

By Zack Lau, W1VT

How to Work 10-GHz DX

Part 1: Location, Location, Location

The latest generation of commercial equipment for the amateur 3-cm band has made it quite practical to work 10 GHz DX, even if you don't have a friend in the surplus business. It is now practical to buy solid-state amplifiers with output levels of 2 W or more; that's high power on 10 GHz. Two watts is enough to routinely make 200-km troposcatter contacts. There is no need for high mountaintop locations or unusual propagation conditions if you can illuminate the troposphere with enough RF. Hams with experience on 2-meter SSB may wonder why 2 W is as effective on 10 GHz as 100 W or more on 2 meters. The answer is the practicality of high-gain antennas—a 34-dBi 10-GHz dish can be as small as two feet in diameter. A similarly sized 2-meter antenna has about 10 dBi of gain. However, this only applies if the antenna is high off the ground—a 3-meter-high 2-meter Yagi may have no more ERP than an isotropic radiator, if you only consider the low radiation angles useful for long-distance troposcatter. Thus, a rover station that is too far away to make VHF contacts may still be able to make microwave contacts, if the operators are skilled.

One of the most important skills is the choice of the proper location. Illuminating the troposphere requires a good view of the horizon. At 10 GHz, it also requires that no nearby trees be in the way. Trees are a double whammy—they attenuate the signal and increase the background noise picked up by the receiver. These two requirements aren't easy to obtain on Connecticut hilltops—nearly all good hilltops have obstructing trees. Thus, I've actually had better luck at the beach than on Connecticut hilltops. This is true even when working inland into Vermont and New Hampshire hilltops. The absence of trees easily offsets the lack of height at 10 GHz. 2 meters is much different, as the attenuation from a few trees is negligible.

Another issue is timing—two stations listening or transmitting at the same time does not work. One technique is to use one-minute sequences—one station transmits on the even minutes and the other station transmits on the odd-numbered minutes. This time interval seems to work well if both stations have good frequency calibration. It may be useful to use a clock with a second hand, although I've had no problems with a digital wristwatch accurately set with a GPS receiver. I thought my Garmin GPS 45 XL would be suitable, but it generates noticeable interference on 2 meters, unless the RF circuitry is disabled. Similarly, the inexpensive Garmin *e*trex only reads time to the nearest minute, making it unsuitable as well. Garmin receivers since the 45XL can provide Maidenhead or grid-square output. This is handy for determining whether you are actually in a rare grid square.

The most common technique is to use

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2-meter SSB to set up contacts. The obvious difficulty is that you need a good 2-meter station—25 W and a small beam isn't enough to match a 10-GHz station with a couple of watts and a small dish. However, a power source to power a 160-W amplifier for many hours is quite heavy. This may significantly reduce the number of practical operating locations. Ideally, one station can turn on a carrier and allow the other station to find and peak up on it. The 2-meter liaison is then used to notify the other station, so that both stations can optimize aiming and receiver tuning. This allows the maximum SNR to be obtained before attempting a contact. In some cases, it is even practical to retransmit the received 10-GHz signal. This will allow the transmitting station to peak the dish alignment.

Good frequency calibration can be difficult at 10 GHz. An error of 10 kHz is common. I've had good luck using a marker generator phase locked to a good 10-MHz "ovenized" oscillator, using a circuit developed by WA6CGR.[1] By tuning in a known 10368.000-MHz signal, I can easily determine the conversion error of my transverter. I initially attempted to lock a 96.0019-MHz fifth-overtone crystal to produce a harmonic on 10368.200 MHz, but discovered that the reference frequency was too low for the time constant of the crystal—the loop wouldn't lock. John Stephensen, KD6OZH, recommends a minimum frequency of 50 kHz.[2] A 5 or 10-MHz ovenized oscillator is optimized for stability—a short-term stability of better than 1×10^{-8} isn't unusual. This is just 100 Hz at 10 GHz. Thus, once you figure out the frequency of your ovenized oscillator, it should be very accurate over the entire contest weekend. However, quartz oscillators age—they drift with time and need to be set to the exact frequency. A popular method is to use the GPS satellites as a reference for the exact frequency.

I don't know of any affordable GPS receivers that provide an RF output standard—say a convenient 5 or 10-MHz output. Typically, you buy a GPS engine, a special purpose GPS receiver that provides an accurate 1-pulse-per-second output. This can then be phase locked to an ovenized quartz oscillator.[3, 4, 5] Some Trimble receivers also provide a 1-pps output, but they tend to be too costly to be dedicated to this application. Alternatively, you may be able to use on-the-air stations as frequency references.

[1]Notes appear on page 60.

The goal is to be able to set your transmitter to a commonly accepted frequency. Since 10-GHz DX is typically far from the band edges, it really doesn't matter if the actual frequency differs, by a few kilohertz, from the commonly accepted international standard, as long as everyone sets their equipment to the *same* frequency.

A rubidium standard would be even more accurate, but these typically run about $300 to $500—even as surplus. Good quartz standards can often be purchased for $50 to $200. Actually, a rubidium standard is more costly than the price would indicate—rubidium standards are much more complex and prone to failure. It is rare to hear of quartz standards failing, even after decades of service. It was previously possible to obtain excellent quartz standards out of old HP frequency counters quite cheaply, until Brooks Shera described how to phase lock them to GPS satellites for even more stability.

Even if you have a rubidium or GPS-locked system, I wouldn't count on better than one kilohertz of frequency accuracy on CW. A good operator knows that precise frequency spotting is problematic. Many VHF radios implement CW poorly—the peak in the audio passband may not correspond to the actual frequency offset. Operators forget to disable RIT controls. Even on SSB, there may be an unexpected frequency offset, as the local oscillator may be sensitive to load or voltage changes. Hunting around for a signal is always a good idea, unless you are lucky and hear the station immediately. Contacts become easier as you become familiar with your equipment, as well as the gear used by other stations.

What Antenna Should I Use for 10-GHz DX?

Horns are good antennas for beginners. Sadly, they are impractical when you need more than about 23 dBi of gain. A big horn that actually works becomes very long. While you could fold or bend the horn, this gets away from the idea of simple, high-performance antennas. I suggest an 18-inch DSS dish or a two-foot conventional parabolic dish. A DSS dish is cheap, while a conventional dish is easier to use. You don't want too big of a dish—big dishes have very sharp beamwidths that make pointing difficult. Unlike EME, you rarely have convenient targets. Imagine how much tougher EME would if the moon were invisible to the eye. An antenna diameter of about two feet seems a good compromise between gain and difficulty of antenna pointing—many people have had frustrating experiences attempting to use four-foot dishes. The –3-dB beamwidth of a four-foot 10-GHz dish is just 1.5°. A two-foot dish has a 3° beamwidth on 10 GHz. By comparison, a huge 20-dBi Yagi has a 3-dB beamwidth of 20°.

How Accurately Can You Point Your Dish?

A popular way of pointing a dish is to use a setting circle. A setting circle is a large compass rose that indicates the direction of a pointer. I made my first ones by pasting paper patterns for plotting antenna gain to wooden disks.[6] Currently, I make them out of sheet aluminum. Covering the paper with plastic laminate offers some protection against rain. Paul Wade, W1GHZ, recommends obtaining metal disks from old computer hard drives. It is time-consuming but useful to mark the patterns with the headings of popular operating locations. An eight-inch diameter setting circle has a circumference of 25 inches. An angle of 1.5° is just 0.10 inches of that 25-inch circumference! This is just one source of error. Your setting circle may not be perfectly concentric with the axis of rotation, adding some error. How accurately is your dish feed positioned? Moving the feed off to one side will slightly move the heading of the dish. A dish that is warped due to mishandling may have its pattern skewed by several degrees. Thus, it is quite possible for the dish to look like it is pointed in a slightly different direction, compared to the signal peak.

The beam headings themselves are another source of inaccuracy. At 50 km, using the centers of six-character grid squares at 40° latitude introduces a possible error of 7°. The error worsens toward the equator, where the squares are larger. Thus, while Garmin GPS receivers provide grid-square or Maidenhead readout, you may need more accuracy for short distance dish pointing. You might also consider the program used—a program that assumes a perfectly spherical earth will not be as accurate as one that correctly uses an oblate spheroid.

Garmin GPS receivers can be used to accurately calculate bearing and distance, but entering waypoints is rather cumbersome and time consuming. In addition, Garmins do not calculate the reverse bearing. The heading from

FN31RG to FM08JL is 243.65°. The actual reverse bearing is 59.37°, not even close to 64°. A four-degree error is quite serious, if you insist on using a big dish. The inaccuracy from using six-digit Maidenhead locators for the 651-km path from Hammonasset beach to Reddish Peak is less than 1°, quite acceptable with the recommended 18-inch or two-foot dish. I find it much faster to use a Palm handheld computer to calculate bearings with six-digit Maidenhead locators in the field. Six-digit grid squares work well if you are on a mountaintop—short distance contacts are usually loud enough to overcome the pointing inaccuracy, while longer distances reduce the inaccuracy acceptably.

Ideally, a strong beacon many miles away can be used for dish calibration. Dish pointing can be quite accurate if you are moving just a few degrees from a known signal source. Alternately, you can work someone from a known location and point your dish. However, neither may be available.

(This discussion has been split into two parts. Part 2 will discuss 10-GHz antenna hardware in the next issue of *QEX*.—*Ed*)

Notes

[1]D. Glawson, WA6CGR, "A Universal Phase Lock Loop System for Microwave Use," *Proceedings of Microwave Update '94*, pp 69-76.

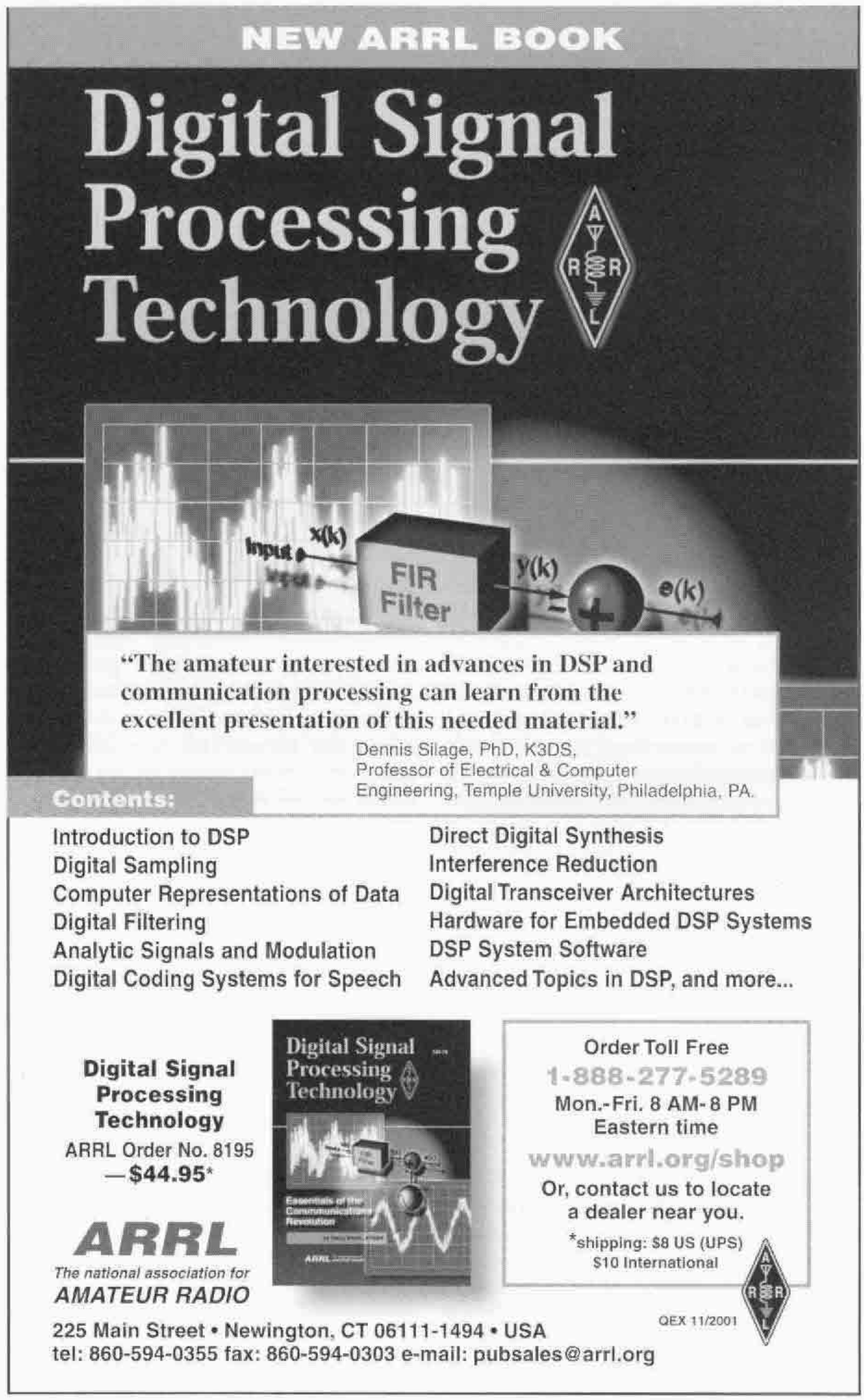
[2]J. Stephensen, KD6OZH, "A Stable, Low-Noise Crystal Oscillator for Microwave and Millimeter-Wave Transverters," *QEX*, Nov/Dec 1999, pp 11-17.

[3]B. Shera, W5OJM, "A GPS-Based Frequency Standard," *QST*, July 1998, pp 37-44.

[4]**www.rt66.com/~shera/index_fs.htm**

[5]**www.tapr.org** is also a good source of information—look at the Totally Accurate Clock project in the kits and projects section.

[6]Antenna Pattern Worksheets, #1360, $3. ARRL publications are available from your local ARRL dealer or directly from the ARRL. Check out the full ARRL publications line at **www.arrl.org/shop/**.



Letters to the Editor

The *Q* of Single-Layer, Air-Core Coils: A Mathematical Analysis (Sep/Oct 2001)

I am not a regular reader of *QEX* but an acquaintance sent me a copy of this article. I wish to offer some thoughts in response.

I base my discussion on the following premise: For a wire of constant length and thickness, a coil of maximum inductance may be constructed. In other words, for a given inductance, a minimum wire length exists. From this it follows that when an inductor is built using the least wire, its Q is at its highest.

We can find the proper relationship from Mr. Murphy's Eq 1,

$$L_\mu = \frac{d^2 n^2}{18d + 40l} \qquad \text{(Eq 1)}$$

Where d is the diameter, n the number of turns, and l the end-to-end length of the solenoid. The wire length is $l_d = \pi n d$. Now we have $l_d/\pi = nd$ and

$$\left(\frac{l_d}{\pi}\right)^2 = n^2 d^2 \qquad \text{(Eq 2)}$$

That is the term in the numerator of Eq 1. It is directly proportional to the square of the wire length.

If the wire length and the inductance are constants, then the denominator of Eq 1 must also be a constant. A fraction is largest when its denominator is smallest. In this case, that occurs when the addends are equal, or $18d = 40l$. Then, the ratio $l/d = 18/40 = 0.45$. This is the construction condition of the solenoid with the largest inductance for wire length l_d.

Substituting $d = l/0.45$ into Eq 1, we get

$$L_\mu = \frac{l \times n^2}{16.2} = \frac{S_W n^3}{16.2} \qquad \text{(Eq 3)}$$

Where S_W is the winding pitch. The number of turns may then be expressed as

$$n = 2.53\left(\frac{L}{S_W}\right)^{\frac{1}{3}} \qquad \text{(Eq 4)}$$

On p 36, you show in Fig 4 a coil of 34.7 µH and the values $n = 26.994$, $S_W = 0.125$ inch, $d = 3.00$ inches, $l = 3.375$ inches, $f = 1.8$ MHz and $Q_t = 549.64$. For this coil, $l_d = \pi n d = 254.5$ inches. Although Mr. Murphy cites Terman in agreement with my $l/d = 0.45$, why does he use $l/d \approx 1.125$ for this coil, then?

I calculate for $L_\mu = 34.7$ µH and wire diameter $W = 0.063$ inch the following values:

$$n = 2.53\left(\frac{34.7}{0.125}\right)^{\frac{1}{3}} = 16.5$$

$$l = nS_W = 2.06 \text{ inches} \qquad \text{(Eq 5)}$$

$$d = \frac{2.06}{0.45} = 4.58 \text{ inches}$$

The Q of this coil is 595 as against Murphy's $Q = 557$.

My examinations reveal that the highest coil Q is produced not with $S_W = 2W$ but with $S_W = 2.5W$. Murphy's Eq 8 sets the limit at $S_W = 2.22W$ and Eq 3 isn't valid any more. Eq 7 is also doubtful since with $S_W = 1.0W$, the turns are almost touching one another and it is only that thicker wire is used. The coil's self capacitance gets larger, but its Q is not necessarily lower than with $S_W = 1.43W$.

Lastly, I present the following law of nature that can also be used to verify air-coil results:

$$L_\mu = \frac{\left(\mu_0 n^2 A\right)}{l_f}$$

where

$$\mu_0 = 4\pi \times 10^{-7} \frac{Vs}{A}, \qquad \text{(Eq 6)}$$

$$A = \frac{\pi d^2}{4}$$

and l_f is the length of the shortest magnetic field line. With all the equations above, air-core coils may be calculated for their numbers of turns, inductance and Q. These equations also apply to coils on printed-circuit boards.

I would be pleased to hear from your readers, but they will have to use regular mail since I do not have e-mail or fax.—*Dr. Lothar König, Ingeborg-Bachmann-Strasse 1, 01219 Dresden, Germany*

VE3ERP responds:

In response to Dr. König's letter: With reference to the spacing of turns, there are several differing opinions (see my equation sources), ranging from about 1/0.70 to 1/0.45 times the conductor diameter. I used these factors in Eqs 7 and 8. Since I am not the originator of any of the other equations in the article, I am not qualified to offer an opinion on Dr. Konig's comments. I can only suggest that Dr. Konig refer to the original equations from my list of sources.

Referring to the value of l/d being 1.125:1 in Fig 4, a glance at Fig 3 will verify that for a coil δiameter between 2.94 and 3.12, the l/d ratio is somewhere between 1.22:1 and 0.98:1. Fig 4 is an example of interim results, from which the user selects a diameter (in this case 3.00) and Fig 5 shows results using this choice.—*George Murphy, Box 759 275 Victoria St East, Alliston, ON L0M 1A0, Canada;* **ve3erp@encode.com**

The Art of Making and Measuring Low-Frequency Inductors (Sep/Oct 2001)

Some old techniques still work just as well as they ever did. Many were abandoned commercially with rising labor costs and the introduction of inexpensive vacuum tubes.

"Basket-weave" makes the number of support dowels odd and winds the turns under and over, greatly mitigating the increase of R_{ac} caused by the proximity effect. It also substantially reduces distributed capacitance because of increased wire spacing over most of each turn. There is little change in L if seven or more supports are used. Use string ties at the crossing points to support and space the wires; use lots of Q dope to make them self-supporting when the dowels are removed.

The "pancake helical" form of inductor is also good, especially when combined with basket-weave winding. Unfortunately, it also cuts a larger area as an antenna in its own right. Some very good Tesla coils have used a conical, basket-weave coil.

References to "Litz" wire should be in quotes unless explained that it is a short form for Litzendraht, a proper name. The original Litzendraht paper gave calculations for design of a wire weave in which each wire was the same length in each possible position in the bundle, and in an optimally short length of the total for minimum R_{ac}. That is almost like making shield braid, but caution: Don't try to use actual shield braid. The wires *must* be individually insulated.

Randomly bundling wires usually approaches the Litzendraht requirements. That is best achieved by laying out the straight lengths required, stripping and soldering the ends to assure equal length, then winding. A wrap with a few threads of silk or Rayon will help hold it all together. Don't twist the thing into a firm bundle: That loses the randomness. Avoid numbers of wires that give an exactly circular bundle, such as 7 and 19. See any stranded-wire table where those combinations are listed.

On VLF, antenna-tuning coils having high Qs yield incredibly narrow bandwidths. That is okay for fixed-frequency operation of slow CW, but it can be a pain if you use the same arrangement for receiving [other frequencies] unless you make it variable. A few turns arranged as a "variometer" at the low-voltage end make the coil tunable without much *Q* degradation. Don't try for a 10:1 inductance ratio, though—as in the old Atwater-Kent broadcast tuners—or you will greatly sacrifice *Q*.

A final note on variable capacitors at these frequencies: Don't use the bearings as the ground return. To maintain high *Q*, use a short length of wire or braid connecting rotor to frame. The same recommendation applies to a variometer coil for a tunable inductor. See the variable capacitor in old Hammond organs, used for the "noise-less" swell-pedal circuit.

Both inductor articles are nice. The authors are to be congratulated.—*Chan Shaw, WA6EWY, 17350 Firma Ct, Granada Hills, CA 91344-1902;* **72570.216@Compuserve.com**

The authors respond:

We know "basket weave" well, but it is complex to build. Our purpose was to describe an easy way to realize high-*Q* coils. Paying attention to the proximity effect and increasing the winding pitch, we can obtain values of up to 800. Why complicate our lives?

Thanks for your clarifications on the wire. We have used "Litz" wire of 42×0.18 mm as typical, since it is easy to find thanks to its use in 100 to 200-kHz switching-regulator transformers.

The final *Q* of an antenna system is usually determined by ground resistance. We have a ground resistance of about 80 Ω and it's not useful to dissipate additional power in the coil (80 Ω with R_{GND} = 20 Ω and I_{ANT} = 2 A).

We agree with you: It's not so easy to find and implement a high-*Q* variable capacitor. Many thanks for your suggestions and comments.—*Paolo Antoniazzi, IW2ACD and Marco Arecco, IK2WAQ;* **Paolo.ANTONIAZZI@st.com**, **MARCO.ARECCO@st.com**

A Two-Meter Reflective SP4T PIN-Diode Switch (Nov/Dec 2001)

I got the copies of *QEX* with my article inside. Thank you very much indeed. I compared my last comments and corrections with the printed article and I found only one mistake. The right form of Eq 3 is:

$$V_{neg} = -(2Rx1 \bullet I_f) + V_f' \quad \text{(Eq 7)}$$

I read some other articles inside *QEX*. It was the first time I had read the magazine. I must say that it is very good. That means it has a very good technical level. These articles are readable by a wide amateur forum. You can find a little bit of everything, from theoretical [discussions] to practical experiments. I very much like your small number of pointed articles rather than many incomplete articles, which would be for nothing. I see that there is direct and strong feedback between readers and authors inside "Letters to the Editor."

The major reason why I sent my article to *QEX* is the following. I have written several articles such as: "Diode Double-Balanced Mixers," "Norton's HF Pre-Amplifier, 1.8 - 30 MHz," "Switchable Band-Pass Filters for 1.8-28 MHz," "Audio CW/SSB Filter" and so forth for *Amaterske Radio* (Czech Amateur Radio magazine). [The magazine] had 95% technical articles and 5% advertising, but the current situation is the opposite—95% advertising—I don't like it.

The article from K5AM, "HF Circuits for a Homebrew Transceiver," is especially excellent—not excellent, but really superb! I wish you many happy readers and best regards.—*Pavel Zanek, OK1DNZ, Slovenska 518, Chrudim, Czech Republic, 537 05;* **Zanek.pavel@worldonline.cz**

HF Circuits for a Homebrew Transceiver (Nov/Dec 2001)

A correction: In Fig 7 (p 29), a 1000-pF capacitor is incorrectly shown connected to the tube's plate. It should be connected only to the ring—*Doug Smith, KF6DX,* QEX *Editor*

A Spreadsheet for Remote Antenna Impedance Measurement (Sep/Oct 2001)

Pat Wintheiser, WØOPW, and author Ron Barker, G4JNH, have informed us of a typographic error in the denominator of Eq 8. It was erroneously shown as:

$$R_{IN} = \frac{R_a\left(1+\tan^2\theta\right)}{1+\dfrac{R_a}{Z_0}\tan\theta} \quad \text{(Eq 8)}$$

The correct equation is:

$$R_{IN} = \frac{R_a\left(1+\tan^2\theta\right)}{1+\left(\dfrac{R_a}{Z_0}\tan\theta\right)^2} \quad \text{(Eq 9)}$$

RF: A Simple 10-Meter Satellite Turnstile Antenna (Nov/Dec 2001)

Some clarifications: Fig 1 isn't an accurate representation of the two-source model. Attempts to model it indicate that the currents are unbalanced and the phase error is significant. Proper phasing with a single 50-Ω source may be obtained by changing L1 and L3 to 0.48-µH inductors with 1 Ω of series resistance and L2 and L4 to 0.775-µH inductors with 6.5 Ω of series resistance, according to an *EZNEC* model. Fig 2C and 2D are more obvious if drawn to scale: Shorten the horizontal lines to represent the sloping wires. An *EZNEC* file of the modified design is available for download at **www.arrl.org/qexfiles/**. Look for RF1101.ZIP.—*Zack Lau, W1VT, QEX Contributing Editor;* **zlau@arrl.org**

Upcoming Conferences

Long-Island Mobile Amateur Radio Club Hamfair

On Sunday, February 24, 2002 a 9 AM, the doors of Levittown Hall will open to the public for the annual Winter Long-Island Mobile Amateur Radio Club Hamfair and Electronics Show.

Over 100 tables will be piled high with Amateur Radio parts and equipment, from antique Morse code keys to the latest in transceiver technology. Computer equipment and wireless communications devices of all types (and vintages) in all price ranges will be in evidence. This is the place to be to purchase everything from a call-sign badge to an antenna, a handheld radio, accessory (battery, case etc) or full HF station.

This is the place to make up for what Santa didn't bring you! Come down and cruise the aisles. There will be a food truck providing hot coffee and sandwiches. If you want to get an Amateur Radio license or upgrade, there will be an ARRL VE Exam session for all classes of license starting promptly at 10 AM. (The FCC exam fee is $10; please bring a photo ID, your original and a copy of your license and an CSCE if applicable.) A VHF tune-up clinic will be available to test your radio and there will be information about Amateur Radio from the ARRL and LIMARC.

Check out the Web page at **www.limarc.org** for more information, vendor-reservation forms and directions. Talk-in will be on LIMARC's W2VL repeater, on 146.85 MHz. General admission is $6; children 11 and under get in free. Hamfair tables are $25 and include one admission. Tables will be available in advance only. There is special close-in parking and a drop-off area available for vendors. Doors open to vendors at 8 AM. For more information contact LIMARC Winter Hamfest Chairman Rich Rosner, N2STU, at **hamfest@limarc.org** or tel 631-563-1859.

Ham Radio University 2002

On Sunday, January 20, 2002 at 9 AM, the doors of the Babylon Town Hall Annex on Phelps Lane will open for the third annual Ham Radio University. Ham Radio University 2002 is a day of education about Amateur Radio. This year the forums have been expanded and new forums have been added. There will be 20 one-hour presentations with special forums geared to the non-ham as well as the experienced ham radio operator. The focus will be "hands on" with many demonstrations.

Come join us at the Babylon Town Hall Annex, Phelps Lane, North Babylon, Long Island. You can learn about a range of subjects: satellite communications, low-power operating using radios as small as a tuna tin, how to put up an emergency antenna, the latest information on advanced digital communications, finding and fixing radio frequency interference problems, purchasing your first station equipment or H-T. There will be a special "Ask the Experts" forum.

There will also be a Special Event Station set up and operational on HF. HRU 2002 is a cooperative effort between over 20 clubs and organizations in the New York City / Long Island area. It is also the American Radio Relay League, New York City / Long Island Section Convention. Along with the forums, there will be tables set up with information about the different groups, ham-radio classes, exam-session schedules, public service and other activities. Join us for a day of fun and education for the whole family!

Admission is open to all. A donation of $2 per person will help us (complimentary refreshments are available). For more information contact: Phil Lewis, N2MUN, Chairman (tel 631-226-0698; **N2MUN@optonline.net**) or ARRL NLI Section Manager, George Tranos, N2GA (tel 631-286-7562; **N2GA@arrl.org**). Check out the Web page for a full list of forums, directions and talk-in information: **www.arrlhudson.org/nli/hru2002.htm**.

Next Issue in *QEX/Communications Quarterly*

Steve Hageman returns with another timely tidbit from the Hagtronics labs. If you have been following his recent work, you certainly won't want to miss this one.

Valentin Trainotti, LU1ACM, has some observations about dipoles and monopoles. His article contains some good basic information and a few new ideas.

Out of the Box: New Products

NEW CASCODE RF AMPLIFIER MMIC: MBC13916

Motorola has announced a new cascode MMIC amplifier based on a silicon germanium-carbon process. This is a four-pin part in an SOT-343R package. It is used in a similar fashion to a traditional MMIC. All biasing is done internally. The bypass for the top transistor is also provided internally, so the user only needs to provide a ground and input/output matching. This part is a significantly improved version of the MRFIC0916. The SiGe/C process gives much improved noise performance over the previous silicon device. This device would be a good alternative to traditional MMICs in LNA designs where NF is a primary design consideration. A preliminary data sheet is available at **www.motorola.com/wireless-semi**.

The MBC13916 General Purpose RF Cascode Amplifier is available at a suggested list price of $0.23 in 10,000-piece quantities. Samples and small quantities of the device are available from stock.

MBC13916 Features

Usable Frequency Range = 100 to 2500 MHz
19 dB Typical Gain at 900 MHz
NFmin (Device Level) = 0.9 dB @ 900 MHz
NFmin (Device Level) = 1.9 dB @ 1.9 GHz
2.5 dBm Typical Output Power at 1.0 dB Gain Compression at 900 MHz, VCC = 2.7 V
45 dB Typical Reverse Isolation (Device Level) at 900 MHz, VCC = 2.7 V
4.7 mA Typical Bias Current at VCC = 2.7 V
2.7 to 5.0 V Supply

More Motorola News

While researching the MBC 13196, I discovered that Motorola has finished the sale of its RF-power-semiconductor business to M/A-COM. M/A-COM is supporting the most popular devices in the line including the TMOS power-FET line, the HF and L-band bipolar transistors. M/A-COM also supplies numerous old Siliconix power devices. These previously Motorola devices are available through Richardson Electronics and RF Parts. See **www.macom.com**.—*Contributing Editor, Ray Mack, WD5IFS;* **rmack@arrl.org**

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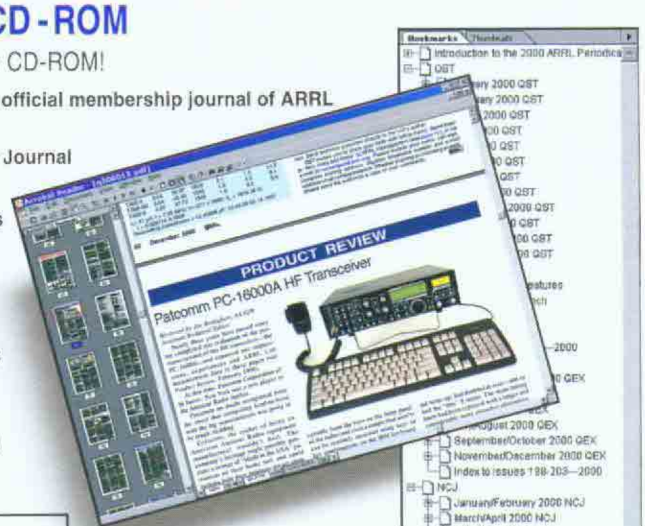

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