An Improved Double Extended Zepp

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N6LF revisits the classic double extended Zepp to improve pattern and SWR bandwidth. He also offers some sage advice about computer modeling.

The sunspot cycle is headed down and the low bands are coming to life for DXing. The next several years will be the time to make those low-band contacts for DXCC. As always, the key to low-band operation is a good antenna. Unfortunately for 7 MHz and down, good antennas don't come in a box ready to assemble. Every location will have a unique set of limitations and opportunities.

One very important difference between this sunspot minima and all past ones is the availability of inexpensive, easy-to-use and powerful antenna modeling software.[1,2] This software allows you to design and optimize an antenna that exactly fits your situation and pocketbook. While cut-and-try experimentation is a very slow way to optimize antennas, modeling is so quick that a wide range of solutions can be investigated easily. The real problem with modeling is generating the will power to stop fooling with the variations and go out and build something!

The following article uses 40 and 80/75-meter double extended Zepps (DEZepp) as examples of what you can accomplish. By adding two small capacitors, made from short lengths of RG-8, in just the right place, the pattern can be improved and the driving-point impedance changed from reactive and narrowband to resistive and wideband. This allows the antenna to be used without a tuner and with an SWR < 1.5:1 over the entire 40-meter band, or with SWR < 2:1 over the entire 75/80-meter band.

A Look at the Classical DEZepp

The classical DEZepp is simply a piece of wire 1.25 λ long, fed at the center, usually with open-wire transmission line and a tuner at the transmitter. The DEZepp displays a useful amount of gain over a dipole of approximately 3 dB. The radiation pattern for a DEZepp designed for 7.15 MHz and suspended 80 feet above ground is shown in **Fig 1**, along with the pattern for a λ/2 dipole at the same height for comparison. The elevation angle is 26°, the peak of the main lobe. The current distribution along the antenna is shown in **Fig 2**.

The DEZepp does indeed provide gain over the dipole, but only over the relatively small angle of approximately 40°. The beamwidth between 3 dB points is 35°. Unless the antenna is pointed directly toward the receiving station, the gain is not usable due to the narrow beam width. In addition to the narrow main lobe, there are significant sidelobes. These are not big enough to be helpful in those directions, but they will also certainly pick up noise and interference. The impedance of the antenna is very reactive, and even when matched at midband does not allow the entire band to be covered without retuning.

For this reason, the DEZepp has traditionally been used with an antenna tuner. This is

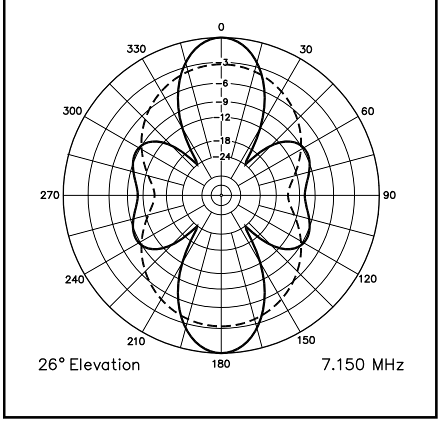

Fig 1—Azimuth pattern of classic double extended Zepp (solid line) at 7.15 MHz, compared with standard dipole (dashed line), both 80 feet high over average ground. Patterns are shown at 26° elevation, where the gain is maximum. The wire runs along the 270° to 90° axis on the graph. Note significant sidelobes for DEZepp.

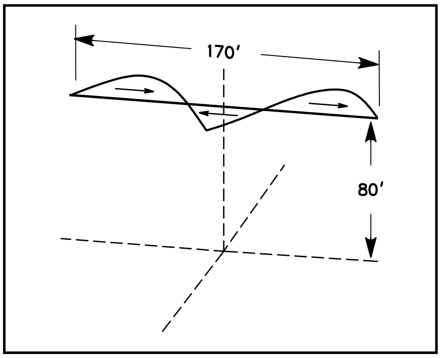

Fig 2—Schematic for classic DEZepp, showing current distribution along antenna. The "bulging out" of the current in the opposite direction near the center of the antenna is responsible for the sidelobes seen in Fig 1.

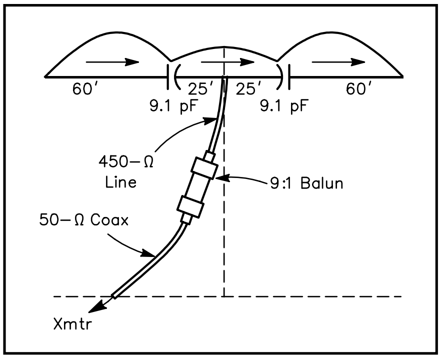

Fig 3—Schematic for modified N6LF DEZepp, with new current distribution. Overall length is 170 feet, with 9.1 pF capacitors placed 25 feet each side of center. Now current distribution doesn't create sidelobes.

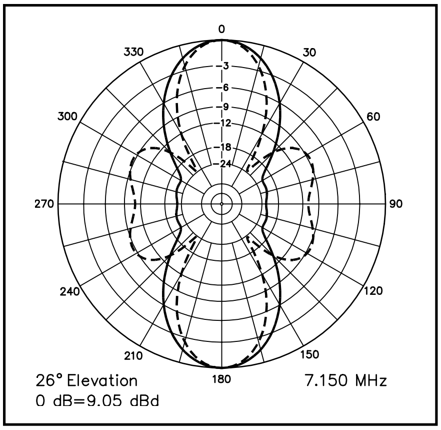

Fig 4—Azimuth pattern for N6LF DEZepp (solid line), compared to classic DEZepp (dashed line). The main lobe for the modified antenna is slightly broader than that of the classic model, and the sidelobes are suppressed better.

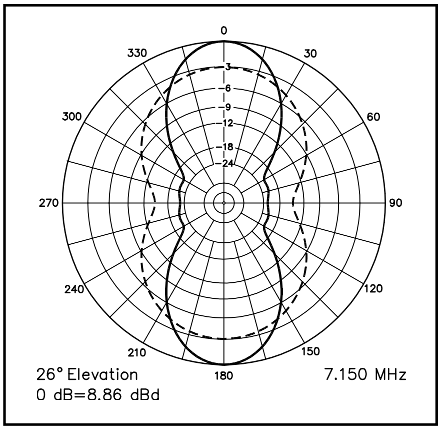

Fig 5—Azimuth pattern for N6LF DEZepp (solid line), compared to dipole (dashed line) at the same height.

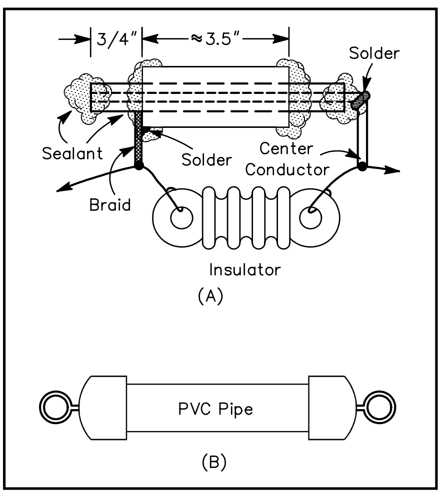

Fig 6—Construction details for series capacitor made from RG-213 coaxial cable. At A, the method used by N6LF is illustrated. At B, a suggested method to seal capacitor better against weather is shown, using a section of PVC pipe with end caps.

not a terrible hardship but it would be nice if the tuner could be eliminated, at least on one band, and a low SWR presented to the transmitter over a whole band.

The gain displayed by the DEZepp is due to the separation between the two current maxima. The small inverted current in the center section subtracts a little from the main lobe and contributes to the sidelobes. The DEZepp is essentially two end-fed collinear dipoles. The transmission line and the center portion of the antenna are the feed system.

It would be very beneficial to suppress the sidelobes and put that energy into a broader main lobe, retaining most of the gain if possible.

A Modified DEZepp

The key to modifying the radiation pattern is to modify the current distribution. One of the simplest ways to do this is to insert a reactance(s) in series with the wire. This could either be an inductor(s) or a capacitor(s). In general, a series capacitor will have a higher Q and therefore less loss. With either choice it is desirable to use as few components as possible.

As an initial trial I decided to use only two capacitors, one on each side of the antenna. I varied the value and position of the capacitors to see what would happen. It quickly became clear that I could tune out the reactance at the feedpoint by adjusting the capacitor value, making the antenna look like a resistor over the entire band. The value of the feed-point resistance could be varied from less than 150 Ω to over 1500 Ω by changing the location of the capacitors and adjusting their values to resonate the antenna. The *AO 6* (Antenna Optimizer) software[1] has the nice feature that it will automatically adjust a variable to tune out reactance. Simultaneously, the pattern was also changing in useful ways.

A number of interesting combinations were created. The one I elected to use is shown in **Fig 3**. The antenna is 170 feet in length. That is a couple of feet shorter than the classic DEZepp, but that also just happens to be all the distance I had between my supporting trees! Two 9.1 pF capacitors are located 25 feet out each side of the center. The antenna is fed with 450-Ω transmission line and a 9:1 three-core Guanella balun[3] used at the transmitter to convert to 50 Ω. The transmission line can be any convenient length and it operates with a very low SWR.

That's all there is to it. The radiation pattern, overlaid with that for a standard DEZepp for comparison, is shown in **Fig 4**. A comparison to a standard dipole is shown in **Fig 5**. The sidelobes are now reduced to below 20 dB. The main lobe is now 43° wide at the 3-dB points, as opposed to 35° for the original DEZepp. The antenna has gain over a dipole for > 50° now. The gain of the main lobe has dropped only 0.2 dB below the original DEZepp.

The reason for the pattern change can be seen in Fig 3, showing the modified current distribution. The main current maxima are still pretty much in the same place, but the current in the center of the antenna now flows in the opposite direction. The resulting pattern is much cleaner.

Experimental Results

I managed to pry myself away from the computer and actually build the antenna. It was made from #14 wire and the capacitors were made from 3.5-inch sections of RG-213, shown in **Fig 6A**. Note that great care should be taken to seal out moisture in these capacitors. The voltage across the capacitor for 1.5 kW will be about 2000 V so any corona will quickly destroy the capacitor. One of the nice features of modeling software is that it gives the current amplitude along the antenna, making it easy to determine the stresses on any series reactances.

I used silicon sealant and then covered both ends with coax seal, finally wrapping it with plastic tape. The solder balls indicated on the drawing are to prevent wicking of moisture through the braid and the stranded center conductor. This is a small but important point if long service out in the weather is expected. An even better way to protect the capacitor would be to enclose it in a short piece of PVC pipe with end caps, as shown in Fig 6B.

Note that all RG-8 type cables do not have exactly the same capacitance per foot and there will also be some end effect adding to the capacitance. I trimmed the capacitor with a capacitance meter. It isn't necessary to be too exact—I checked the effect of varying the capacitance ±10% and the antenna still works fine.

The results proved to be close to those predicted by the computer model. **Fig 7** shows the measured value for SWR across the band. These measurements were made with a Bird directional wattmeter. The worst SWR is 1.35:1 at the low end of the band! With a little adjustment of the antenna length this could have been lowered a bit more, but I figured why bother?

My antenna was oriented to work into Europe. Prior to putting up this antenna I had been using a dipole. I could hear a few Europeans but was unable to work them. Three dB may not seem like much gain but after putting up this antenna I immediately heard many more signals and have been regularly working into Europe with 56/57 reports.

Dick Ives, W7ISV, was sufficiently impressed by the success of the 40-meter version of this antenna to ask me to design a 75-meter version for him. In his location one end of the antenna could only be 60 feet high (< 0.25 λ), and I was concerned about the accuracy of the modeling program, because *MININEC*-based programs are known to be inaccurate for gain and feed-point impedance at low heights. Fortunately, Brian Beezley, K6STI, has a *NEC*-based program called *NEC Wires.*[1] This does model ground accurately and is just the ticket for low antennas. Using this program I designed a new antenna for W7ISV.

Despite the temperatures in mid-December, Dick erected the antenna as shown in **Fig 8**. The series capacitors are 17 pF, and since he isn't interested in CW, Dick adjusted the length for the lowest SWR at the high end of the band. The antenna could have been tuned somewhat lower in frequency and would then provide an SWR < 2:1 over the entire band, as indicated by the dashed line in Fig 8.

This antenna provides wide bandwidth and moderate gain over the entire 75/80-meter band. Not many antennas will give you that with a simple wire structure.

Multiband Operation

When operated with an antenna tuner, one of the advantages of the classical DEZepp is that it is a multiband antenna. Typically a 40-meter DEZepp behaves like a dipole on 75/80 meters and like a long wire on the higher frequency bands. Adding the two series capacitors decouples the ends of the wires on 75/80 meters and a rather poor antenna results. It behaves more like a 30-meter dipole being used on 75/80 meters. For the bands above 40 meters, however, the reactance of the capacitors drops rapidly and the behavior is very much the same as for the normal DEZepp. The price paid for improving operation on 40 meters is the loss of 75/80 meters. Similarly, in the 75/80-meter version, performance on 160 meters is sacrificed.

Some Final Thoughts

The antenna shown here represents a very simple modification of an old idea to suit a particular situation. There are any number of variations that could have produced similar results. Two important lessons were learned during this effort. First, the modeling software is pretty accurate, particularly now that *NEC*-based software is available. The results obtained were very close to that predicted—and this is not the first time I have seen this. Second, the modeling process is a great teacher. It helps you to learn how antennas really work and cuts through many misconceptions. By viewing the current distributions, the associated radiation patterns and driving-point impedances, it becomes much easier to understand which way to modify a design to achieve a desired result. Being able to get results quickly is very helpful also.

There is a whole new world of low-band antennas out there waiting to be created!

Acknowledgments

I want to express my appreciation to Dick Ives for braving the wet and cold to verify the 75/80-meter version of this antenna in time for me to add it to this article. Brian Beezley was very helpful with comments on the design, and of course he created the new *NEC Wires* program. An antenna is only as good as the materials that go into it. I used the beautiful "silky" stranded copper-clad steel wire supplied by Press Jones, N8UG, and also some of his 450-Ω transmission line. It was wonderful to have a wire antenna that didn't try to envelop you (and every nearby bush) like a boa constrictor due to a built-in set in the wire.

Notes

[1] *AO 6.1* and *NEC Wires*, by Brian Beezley, K6STI, 3532 Linda Vista Dr, San Marcos, CA 92069.

[2] *ELNEC 3.0*, by Roy Lewallen, W7EL, PO Box 6658, Beaverton, OR 97007.

[3] Jerry Sevick, W2FMI, *Transmission Line Transformers*, Second Edition (ARRL, Newington, CT, 1990), p 9-28.

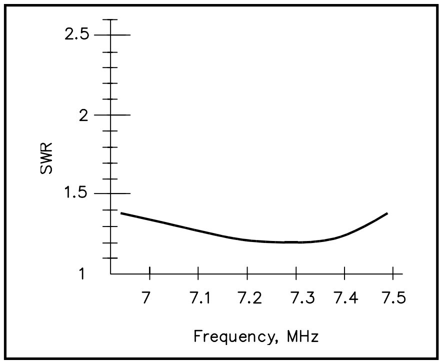

Fig 7—Measured SWR curve across 40-meter band for N6LF DEZepp.

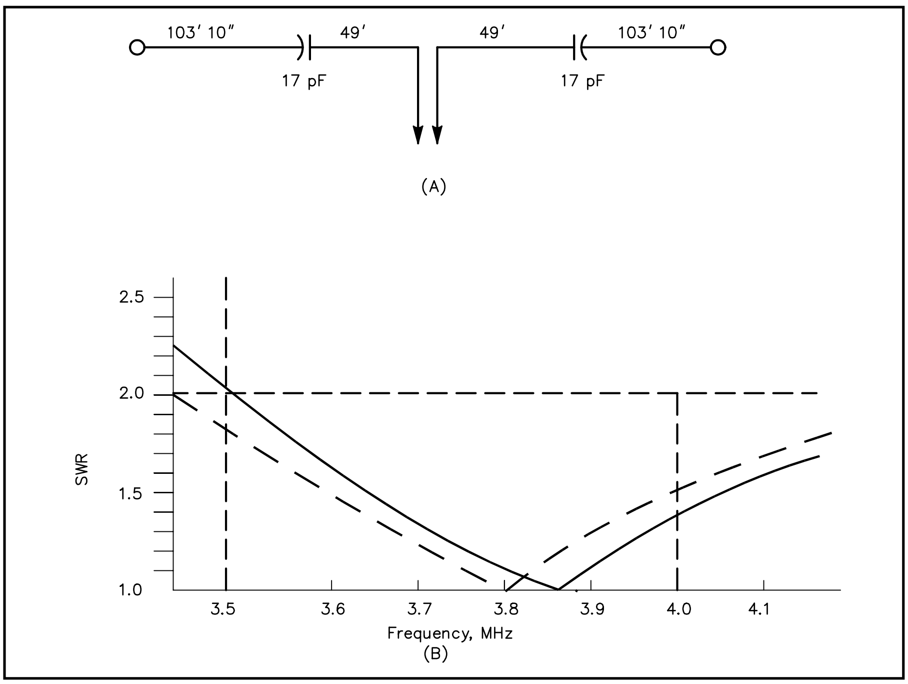

Fig 8—75/80-meter N6LF DEZepp, designed using *NEC Wires*. At A, a schematic is shown for antenna. At B, SWR curve is shown across 75/80-meter band. Solid line shows measured curve for W7ISV antenna, which was pruned to place SWR minimum higher in the band. The dashed curve shows the computed response when SWR minimum is set to 3.8 MHz.

Broadbanding the Half-Square Antenna for 80-Meter DXing

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N6LF discusses a simple way to broadband the classic half-square antenna to operate in both the CW and SSB "DX windows" on 80/75 meters.

The half-square antenna is by nature relatively narrow band.[1] On 80 meters, for example, an SWR below 2:1 can be achieved anywhere in the band, but only over a relatively small range (60 to 100 kHz). The primary reason for using a half-square instead of a dipole is for improved performance on DX contacts.

There are two DX "windows" on 80 meters, 3.500-3.520 MHz and 3.750-3.800 MHz—most CW activity is close to 3.500 and SSB around 3.790 MHz. It is very easy to adjust a normal half-square antenna to have low SWR at either one of these frequencies, but not at both. Practically speaking, any serious DXer will want to be able to use both CW and SSB, so this is a real disadvantage.

It is possible of course to build a matching network of some kind or to use a tuner to load the antenna at both frequencies. However, that may not be as simple as it sounds, because if the SWR is low in one window, it will be very high at the other. It could be 20:1 or more!

The attraction of the half-square is its simplicity. It would be nice to allow operation in both windows while keeping the simplicity. This article shows a way to do that by adding two wires to the classical half-square.

Broadbanding the Half-Square

On 80 meters even a dipole is not a broadband antenna. One trick frequently used to broadband or multiband a dipole is to add additional wires to the dipole to form a fan, as shown in **Fig 1**. The two wires on each side of the feed point have different lengths and are adjusted to produce two resonance points. A variation of this idea works for the half-square. It can provide the desired double resonance and can also provide 3-4 dB of front-to-back ratio if that is desired.

The bi-directional (0 dB front-to-back) version of the half-square is shown in **Fig 2**. The single vertical wires at each end of the antenna have been replaced with two wires, of different lengths (L1 and L2), with the lower ends well separated. Note that the vertical wires are in the plane of the horizontal top wire (L_T). In a bit we will see what happens if the wires are not in this plane. The pattern from this antenna is shown in

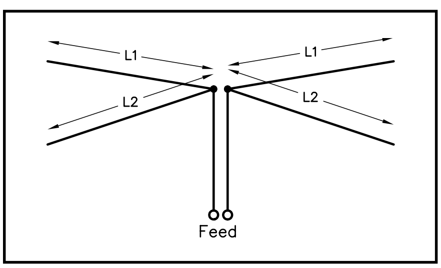

Fig 1—Broadbanding an 80-meter dipole using a fan-shaped pair of unequal-length radiators.

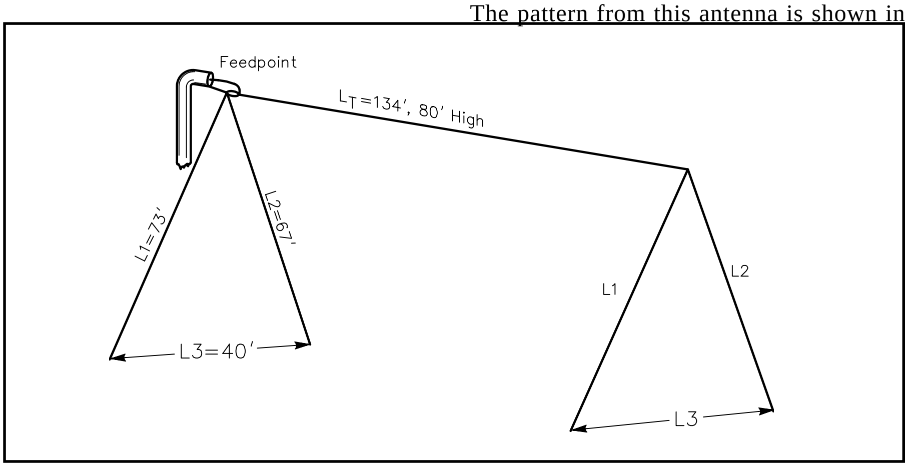

Fig 2—Typical N6LF broadband symmetrical half-square for 80 meters. All wires are in the plane of the horizontal top wire. The vertical wires are spread out 40 feet at the bottom in this case.

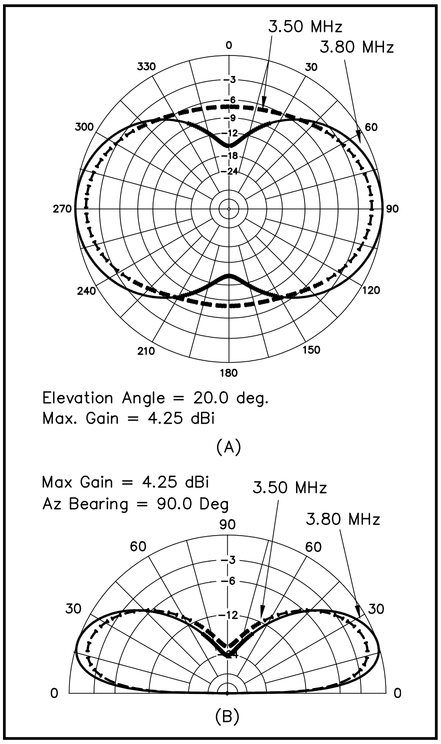

Fig 3—At A, azimuth response of symmetrical broadband 80-meter half-square at 3.8 and 3.5 MHz. At B, elevation response of symmetrical broadband 80-meter half-square at 3.8 and 3.5 MHz.

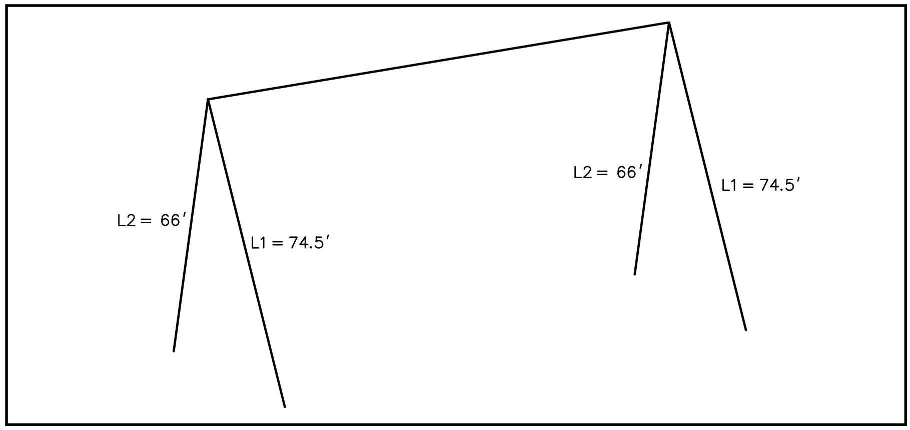

Fig 4—An asymmetrical variation of broadband half-square. Here, the equal-length vertical wires are placed on the same side of a vertical plane cutting through the length of the horizontal top wire.

Fig 3. There is some sacrifice in gain at the lower resonance, but only about 1 dB.

If the vertical wires do not lie in the plane of the top wire, as shown in **Fig 4**, it will still be possible to obtain the double resonance, but the pattern will be affected. As shown in **Fig 5**, the pattern is no longer strictly bi-directional. There can be several dB of front-to-back ratio. The front-to-back ratio improves the gain in one direction; this may be helpful in some situations. More often, however, it is desirable to work long path as well as short path and the bi-directional pattern will be preferred.

Experimental Results

An antenna with the dimensions in Fig 2 was built and the measured SWR is shown in Fig **6**. As expected, there are two resonances, giving acceptable SWR in both the CW and SSB DX windows.

The exact lengths for each wire will depend on the particular installation—the width and height available. If an antenna modeling program such as *EZNEC*[2], *NEC/Wires*[3] or *NEC-WIN*[4] is available, then the antenna can be designed very closely for a particular site, including the ground effects. If the modeling is not available, then it will be necessary to adjust the wire lengths experimentally. Fortunately, all of the adjustments can be made at ground level.

The length of the top wire (L_T) is set during initial construction and can vary from 120 to 150 feet, depending on the space available. The longer lengths will mean that the vertical wires can be made shorter. This allows for lower heights. More detail of this trade-off can be found in Reference 1. There are three other variables: L_1, L_2 and L_3.

The adjustment begins by setting the spacing between the ends of the vertical wires (L_3), then L_3 is adjusted for resonance at 3.790 MHz. Finally, L_1 is adjusted to resonate at 3.510 MHz. L_2 and L_1 are then adjusted one more time. Usually this will be sufficient to place the resonances in the desired locations. If the SWR is not as low as desired, then L_3 can be changed and L_1 and L_2 readjusted. This process should converge rapidly.

Because L_1 and L_2 may need to be either shortened or lengthened, I usually start with extra wire and fold the excess length back on the wire, rather than cutting it off. That way, extra is available to lengthen the wire, if needed.

Conclusion

The narrow bandwidth of the classical half-square antenna can be overcome by adding another set of vertical wires. With a little adjustment, two resonances, with SWR < 2:1 can be achieved. This will allow operation in both the CW and SSB DX windows on 80 meters.

The principle shown here will, of course, also work on other bands. On 160 meters, for example, it would allow a substantial part of the band to be covered without retuning.

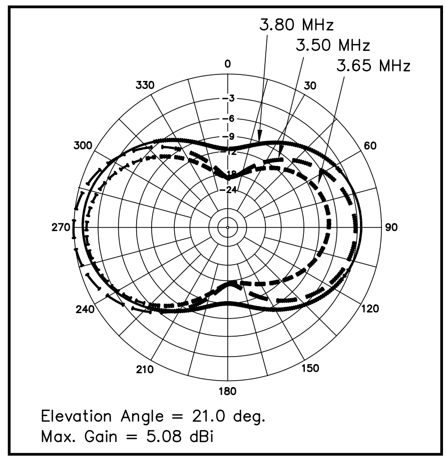

Fig 5—Azimuth response of asymmetrical broadband 80-meter half-square at 3.8, 3.65 and 3.5 MHz, showing how front-to-back ratio changes with frequency.

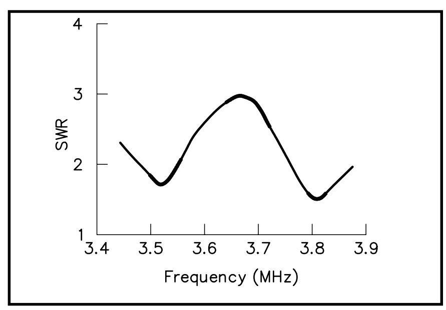

Fig 6—SWR curve versus frequency for symmetrical broadband 80-meter half-square showing characteristic double-resonance.

Notes and References

[1]Severns, Rudy, N6LF, "Using the Half-Square Antenna for Low-Band DXing," elsewhere in this book.

[2]*EZNEC* is available from Roy Lewallen, W7EL, PO Box 6658, Beaverton, OR, 97007.

[3]*NEC/Wires* is available from Brian Beezley, K6STI, 3532 Linda Vista Drive, San Marcos, CA 92069, 619-599-4962.

[4]*NEC-WIN Basic* is available from Paragon Technology, 200 Innovation Blvd, Suite 240, State College, PA 16803, 814-234-3335.

Using the Half-Square Antenna For Low-Band DXing

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Can't put up a really high horizontal antenna for 80-meter DXing? Maybe the vertically polarized "Half-Square" might be the antenna for you.

Antennas widely used by amateurs have a few basic characteristics in common. They provide modest performance and good efficiency, are simple in design, inexpensive to fabricate and very flexible with regard to height, shape and construction materials. There is a very wide range of differences between QTHs, resources and personal circumstances. It is vital that the basic performance of an antenna be preserved even for significant variations in dimensions and materials if it is to be widely useful.

The dipole antenna fits these requirements admirably and is probably the most widely used antenna of all. Unfortunately, on the low frequency bands (80 and 160 meters) it is increasingly difficult to get good DX performance from a dipole due to the problem of getting the antenna high enough (in terms of wavelength). The landmark work by N6BV on HF propagation clearly illustrates this.[1, 2] **Fig 1** shows one of his graphs to illustrate the range of radiation angles most likely to be usable on an 80-meter path from New England to Europe. Over 90% of the time the angles are between 17· and 24·. Other longer paths (and those from different locations) show similar patterns, except that the longer paths have lower peak angles, in the range of 10· to 18·. For DX work on 80 meters, the desirable radiation angles are generally between 10· and 20·.

Also shown in Fig 1 are the radiation patterns for dipoles at 100 feet and 200 feet. At 200 feet the pattern is great, but lowering the antenna to 100 feet reduces radiation at the desired angles significantly. For most hams 100-foot dipoles are not possible and 200-foot dipoles not even a fantasy.

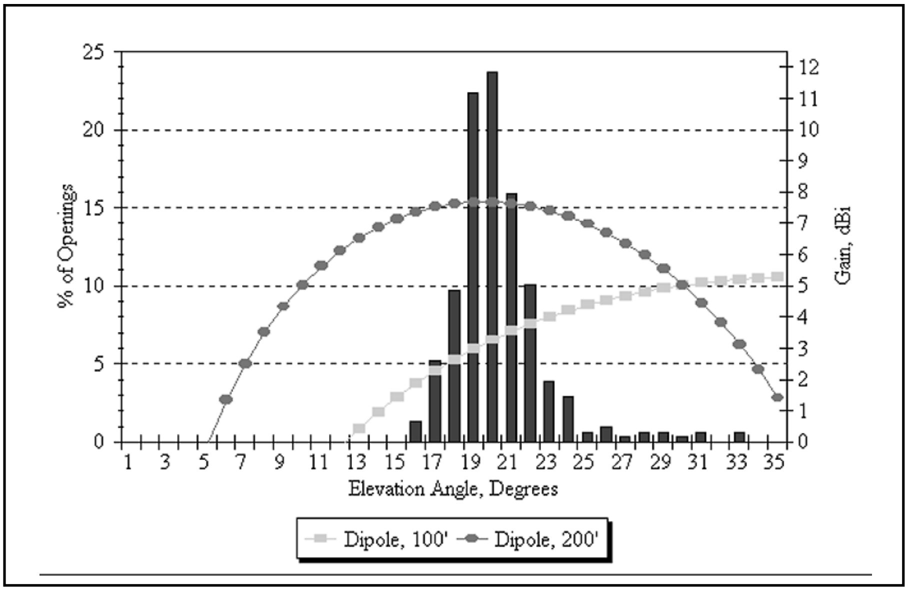

Fig 1—80-meter graph of the percentage of all openings from New England to Europe versus elevation angles, together with overlay of elevation patterns over flat ground for dipoles at two different heights. The 200-foot high dipole clearly covers the necessary elevation angles better than does the 100-foot high dipole. (*From The ARRL Antenna Book, 17th edition, Fig 30.)*

Heights in the range of 40 to 80 feet are much more typical, with the emphasis more towards 40 than 80 feet. This further degrades performance.

Another problem with low dipoles, from a DXing point of view, is that they have great response at high angles. This brings in local and US stations S9+ while you are trying to copy an S3 DX station.

Is there a way to improve on the dipole's DX performance while retaining most of its practical advantages? The answer is "Yes." The half-square antenna can provide 3 to 10 dB of improvement at angles between 10° and 20°, depending on the available height and soil conductivity in the ground reflection zone. In addition, the high-angle radiation can be suppressed. The shape, dimensions and feed point options are also more flexible than previous descriptions have indicated.

The half-square and its close cousin, the "bobtail curtain," have been known to amateurs for nearly 50 years.[3] For the most part, articles describing the half-square have been relatively brief and have not attempted to examine many of the finer points.[4, 5, 6] This very simple antenna has many subtle details and more than a few surprises. You can get very good results without great effort, but it is also possible to obtain very poor performance if moderate care is not taken!

The purpose of this article is to take a careful look at this antenna including:

- Comparison to a dipole at comparable heights, over different grounds.
- The effect of changing shape and dimensions on performance.
- Useful bandwidth, including both impedance and pattern effects.
- Different feed and matching schemes.
- Multiband operation

Modeling Notes

Much of the work presented here was done using computer modeling. Because these antennas are close to ground (in terms of wavelengths) and different parts of the antenna are at different heights, *NEC2* rather than *MININEC* modeling programs were used.[7,8,9] To maximize the accuracy, I included the wire losses and all wires connected at a corner used segment tapering. I assumed real ground, using the high accuracy (Norton-Sommerfeld) ground model. I carefully observed the proscription against grounding wires directly to a real ground. The accuracy of the modeling should be very good.

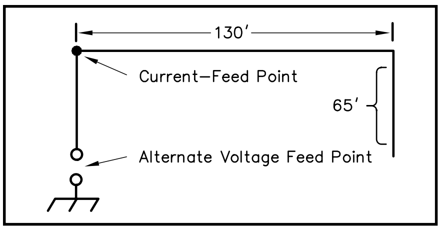

Fig 2—Typical 80-meter half-square, with λ/4-high vertical legs and a λ/2-long horizontal leg. The antenna may be fed at the bottom or at a corner. When fed at a corner, the feed point is a low-impedance, current-feed. When fed at the bottom of one of the wires against a small ground counterpoise, the feed point is a high-impedance, voltage-feed.

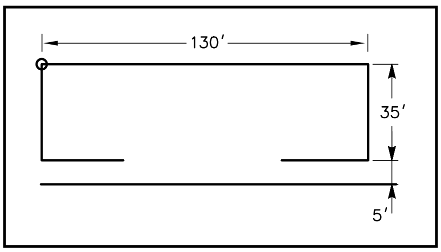

Fig 3—An 80-meter half-square configured for 40-foot high supports. The ends have been bent inward to reresonate the antenna. The performance is compromised surprisingly little.

The Half-Square Antenna

A simple modification to a dipole would be to add two λ/4 vertical wires, one at each end, as shown in **Fig 2**. This is a *half-square antenna*. The antenna can be fed at one corner (low impedance, current fed) or at the lower end of one of the vertical wires (high impedance, voltage fed). Other feed arrangements are also possible.

The "classical" dimensions for this antenna are λ/2 (131 feet at 3.75 MHz) for the top wire and λ/4 (65.5 feet) for the vertical wires. However, there is nothing sacred about these dimensions! You can vary them over a wide range and still obtain nearly the same performance.

This antenna is two λ/4 verticals, spaced λ/2, fed in-phase by the top wire. The current maximums are at the top corners. The theoretical gain over a single vertical, for two in-phase verticals, is 3.8 dBi.[10] An important advantage of this antenna is that it does not require the extensive ground system and feed arrangements that a conventional pair of phased λ/4 verticals would.

Comparison To A Dipole

In the past, one of the things that has turned off potential users of the half-square on 80 and 160 meters is the perceived need for λ/4 verticals. This forces the height to be > 65 feet on 80 meters and > 130 feet on 160 meters. That's not really a problem. If you don't have the height there are several things you can do. For example, just fold the ends in, as shown in **Fig 3**. This compromises the performance surprisingly little.

Let's look at the examples given in Figs 2 and 3, and compare them to dipoles at the same height. For this comparison I have selected two heights, 40 and 80 feet, and average, very good and sea-water grounds. I have also assumed that the lower end of the vertical wires had to be a minimum of 5 feet above ground.

At 40 feet the half-square is really mangled, with only 35 foot high (≈ λ/8) vertical sections. The comparison between this antenna and a dipole of the same height is shown in **Fig 4**. Over average ground the half-square is superior below 32° and at 15° is almost 5 dB better. That is a worthwhile improvement. If you have very good soil conductivity, like parts of the lower Midwest and South, then the half-square will be superior below 38° and at 15° will be nearly 8 dB better. For those fortunate few with saltwater-front property the advantage at 15° is 11 dB! Notice also that above 35°, the response drops off rapidly. This is great for

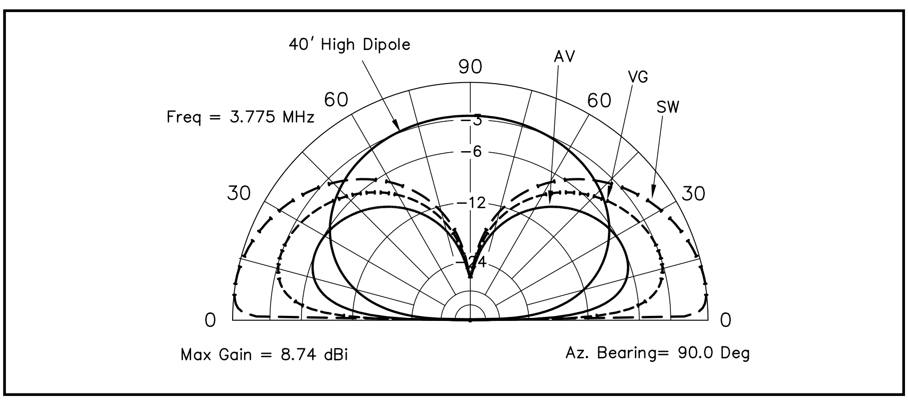

Fig 4—Comparison of 80-meter elevation response of 40-foot high, horizontally polarized dipole over average ground and a 40-foot high, vertically polarized half-square, over three types of ground: average (conductivity σ = 5 mS/m, dielectric constant ε = 13), good (σ = 30 mS/m, ε = 20) and saltwater (σ = 5000 mS/m, ε = 80). The quality of the ground clearly has a profound effect on the low-angle performance of the half-square. However, even over average ground, the half-square outperforms the low dipole below about 32°.

DX but is not good for local work.

If we push both antennas up to 80 feet (**Fig 5**) the differences become smaller and the advantage over average ground is 3 dB at 15°. The message here is that *the lower your dipole and the better your ground, the more you have to gain by switching from a dipole to a half-square*. The half-square antenna looks like a good bet for DXing. However, there are a few other things to consider before replacing your dipole.

Changing the Shape

Just how flexible is the shape? We'll look now at several distortions of practical importance. Some have very little effect but a few are fatal to the gain. Suppose you have either more height and less width than called for in the standard version or more width and less height, as shown in **Fig 6A**.

The effect on gain from this type of dimensional variation is given in **Table 1**. For a top length (L_T) varying between 110 and 150 feet, where the vertical wire lengths (L_V) readjusted to resonate the antenna, the gain changes only by 0.6 dB. For a 1 dB change the range of L_T is 100 to 155 feet, a pretty wide range.

Another variation results if we vary the length of the horizontal top wire and readjust the vertical wires for resonance, while

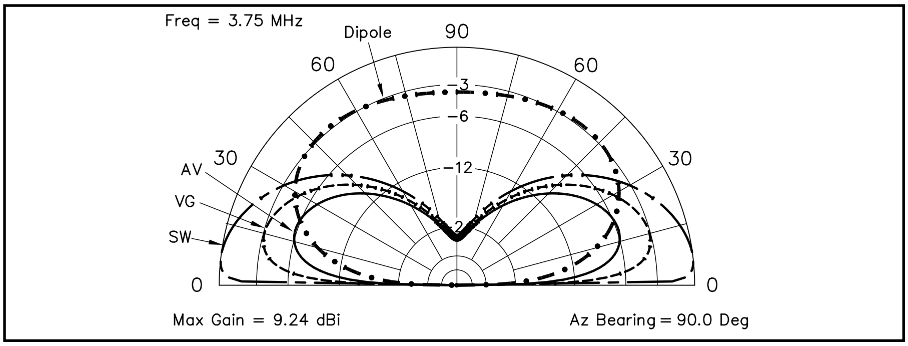

Fig 5—Comparison of 80-meter elevation response of 80-foot high, horizontally polarized dipole over average ground and an 80-foot high, vertically polarized half-square, over same three types of ground as in Fig 4: average, good and saltwater. The greater height of the dipole narrows the gap in performance at low elevation angles, but the half-square is still a superior DX antenna, especially when the ground nearby is saltwater! For local, high-angle contacts, the dipole is definitely the winner, by almost 20 dB when the angle is near 90°.

Table 1

Variation in Gain with Change in Horizontal Length, with Vertical Height Readjusted for Resonance. See Fig 6A.

L_T *(feet)*	L_V *(feet)*	*Gain (dBi)*
100	85.4	2.65
110	79.5	3.15
120	73.7	3.55
130	67.8	3.75
140	61.8	3.65
150	56	3.05
155	53	2.65

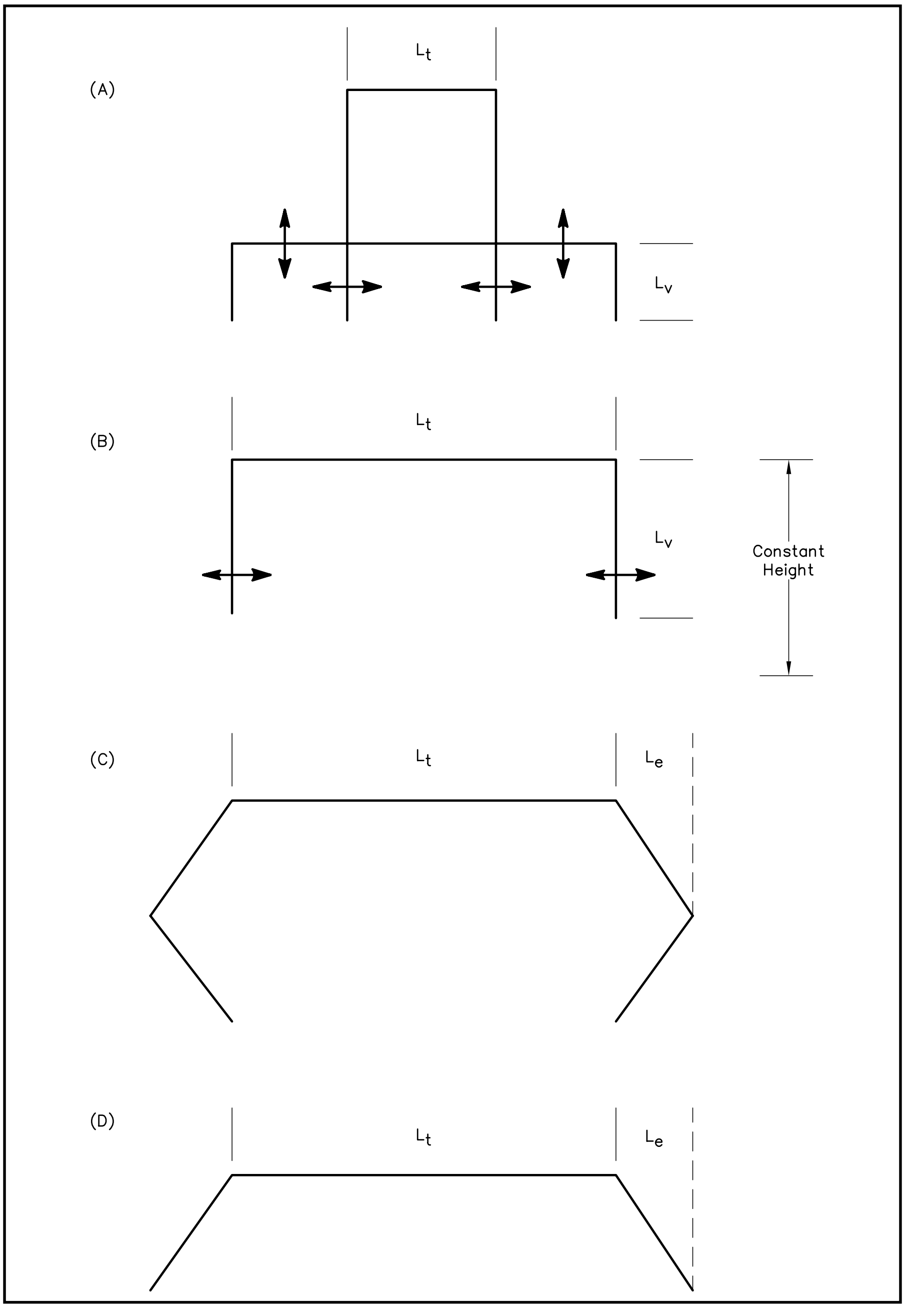

Fig 6—Varying the horizontal and vertical lengths of a half-square. At A, both the horizontal and vertical legs are varied, while keeping the antenna resonant. At B, the height of the horizontal wire is kept constant, while its length and that of the vertical legs is varied to keep the antenna resonant. At C, the length of the horizontal wire is varied and the legs are bent inwards in the shape of "vees." At D, the ends are sloped outwards and the length of the flattop portion is varied. All these symmetrical forms of distortion of the basic half-square shape result in small performance losses.

Table 2

Variation in Gain with Change in Horizontal Length, with Vertical Length Readjusted for Resonance, but Horizontal Wire Kept at Constant Height. See Fig 6B.

L_T (feet)	L_V (feet)	Gain (dBi)
110	78.7	3.15
120	73.9	3.55
130	68	3.75
140	63	3.35
145	60.7	3.05

Table 3

Gain for Half-Square Antenna, Where Ends Are Bent Into V-Shape. See Fig 6C.

Height ⇒ L_T (feet)	H=40' L_e (feet)	H=40' Gain (dBi)	H=60' L_e (feet)	H=60' Gain (dBi)
40	57.6	3.25	52.0	2.75
60	51.4	3.75	45.4	3.35
80	45.2	3.95	76.4	3.65
100	38.6	3.75	61.4	3.85
120	31.7	3.05	44.4	3.65
140	-	-	23	3.05

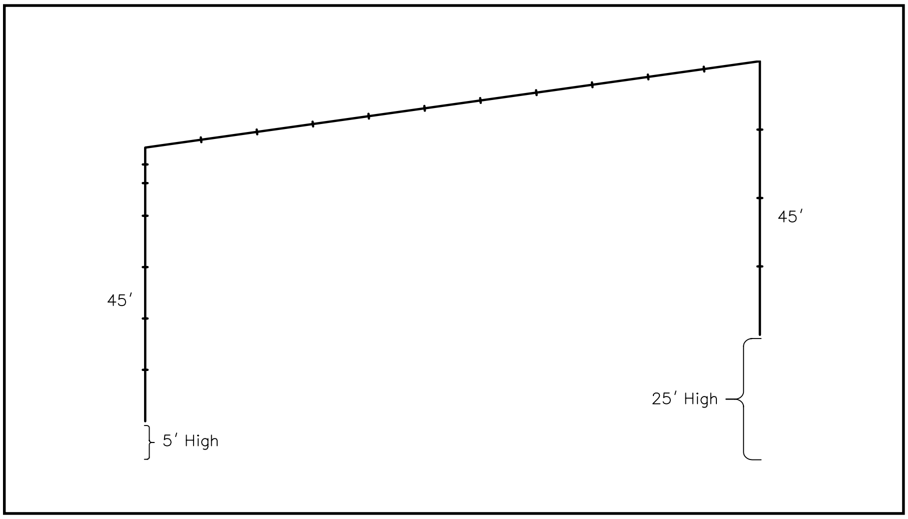

Fig 7—An asymmetrical distortion of the half-square antenna, where the bottom of one leg is purposely made 20 feet higher than the other. This type of distortion does affect the pattern!

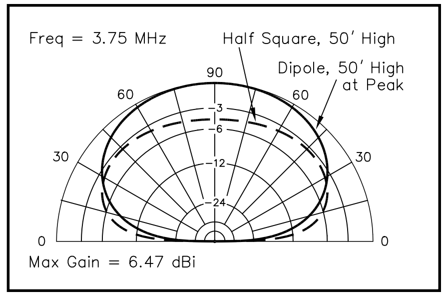

Fig 8—Elevation pattern for the asymmetrical half-square shown in Fig 7, compared with pattern for a 50-foot high dipole. This is over average ground, with a conductivity of 5 mS/m and a dielectric constant of 13. Note that the zenith-angle null has filled in and the peak gain is lower compared to conventional half-square shown in Fig 5 over the same kind of ground.

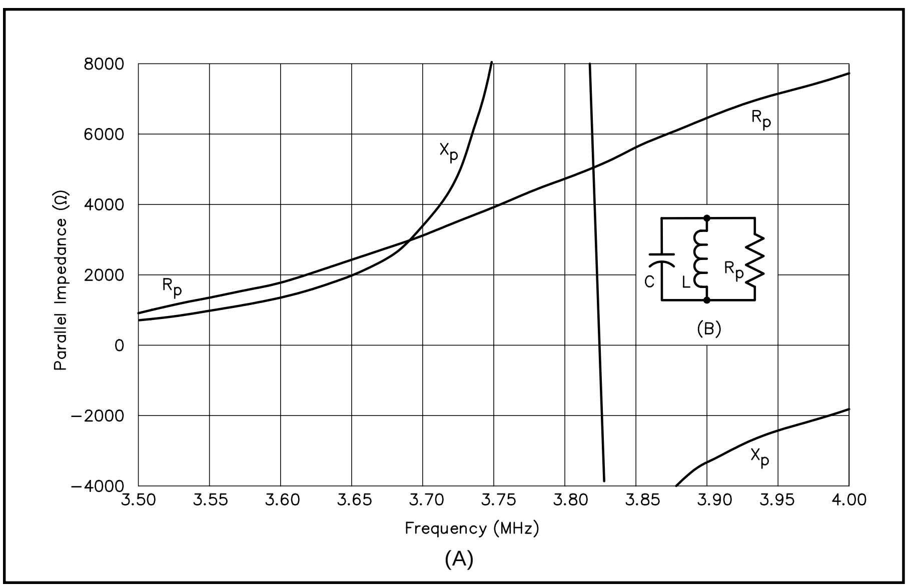

Fig 9—At A, graph of feed point shunt resistance and shunt reactance versus frequency for a half-square with voltage-feed at bottom corner. At B, equivalent parallel circuit of this antenna. This particular half-square is resonant at about 3.820 MHz, where its feed point resistance is about 5000 Ω.

keeping the top at a constant height. See Fig 6B. **Table 2** shows the effect of this variation on the peak gain. For a range of L_T= 110 to 145 feet, the gain changes only 0.65 dB.

The effect of bending the ends into a V shape, as shown in Fig 6C, is given in **Table 3**. The bottom of the antenna is kept at a height of 5 feet and the top height (H) is either 40 or 60 feet. Even this gross deformation has only a relatively small effect on the gain! Sloping the ends outward as shown in Fig 6D and varying the top length also has only a small effect on the gain. While this is good news because it allows you to dimension the antenna to fit different QTHs, not all distortions are so benign.

Suppose the two ends are not of the same height, as illustrated in **Fig 7**, where one end of the half-square is 20 feet higher than the other. The radiation pattern for this antenna is shown in **Fig 8** compared to a dipole at 50 feet. This type of distortion does affect the pattern. The gain drops somewhat and the zenith null goes away. The nulls off the end of the antenna also go away, so that there is some end-fire radiation. In this example the difference in height is fairly extreme at 20

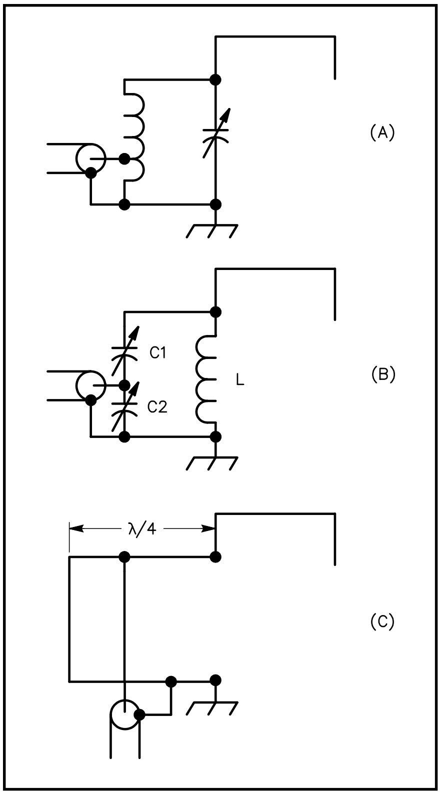

Fig 10—Typical matching networks used for voltage-feeding a half-square antenna.

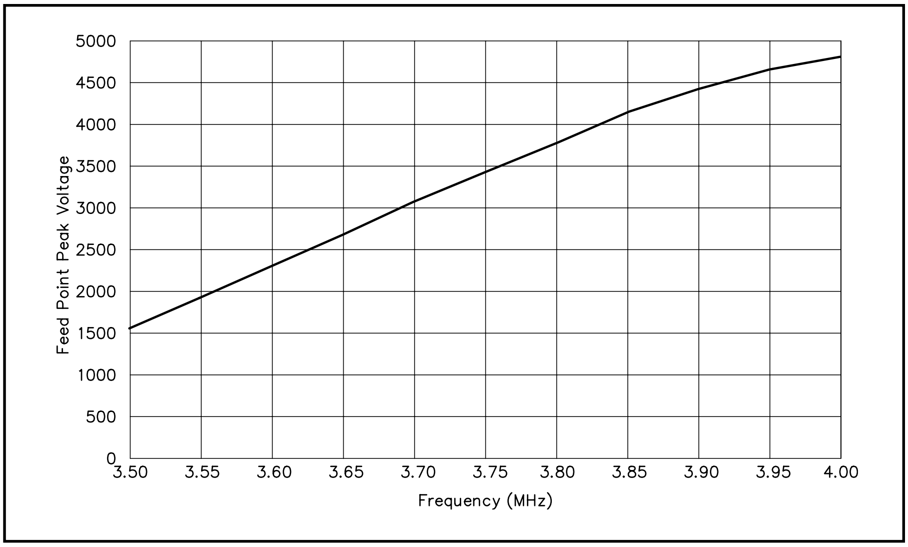

Fig 11—Graph of peak RF voltage at feed point of voltage-fed half-square antenna with 1500 W power.

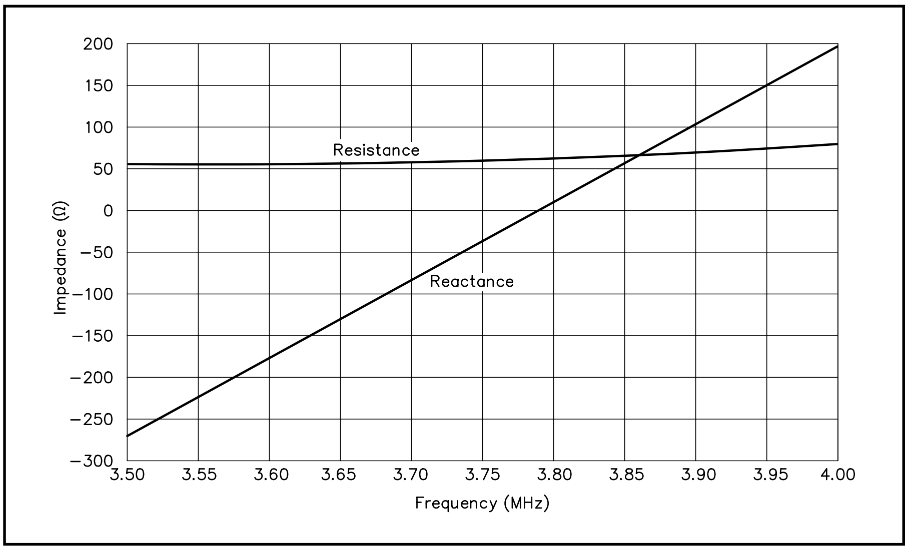

Fig 12—Graph of feed point series resistance and reactance versus frequency for a half-square with current-feed at one corner. Note that the resistive component changes slowly with frequency. This particular antenna is resonant at just under 3.8 MHz.

feet. Small differences of 1 to 5 feet do not affect the pattern seriously.

If the top height is the same at both ends but the length of the vertical wires is not the same, then a similar pattern distortion can occur. The antenna is very tolerant of *symmetrical* distortions but it is much less accepting of *asymmetrical* distortion.

What if the length of the wires is such that the antenna is not resonant? Depending on the feed arrangement that may or may not matter. We will look at that issue later on, in the section on patterns versus frequency. The half-square antenna, like the dipole, is very flexible in its proportions.

Feed-Point Impedance

There are many different ways to feed the half-square. Traditionally the antenna has been fed either at the end of one of the vertical sections, against ground, or at one of the upper corners as shown in Fig 2.

A typical example of the impedance variation for voltage feed is shown in **Fig 9A**. The impedance generated from the modeling program represents the parallel-equivalent impedance (Fig 9B) when driven at one end. This form is most informative when using a parallel L-C matching network, such as the one shown in **Fig 10**.

In addition to the variation in reactance (X_p), the resistance (R_p) varies from 1200 to 5700 Ω. This very high impedance means that the voltage at the feed point will be quite high. A graph of peak voltage for 1.5 kW drive power is given in **Fig 11**. The feed point voltage will be over 4 kV! This must be kept in mind when designing matching networks. Because of the large range of impedances, simple matching schemes yield relatively narrow SWR bandwidths.

For current feed, the impedance is much lower, as shown in **Fig 12**. The resistive component doesn't change very much but the reactive component does. This is a relatively high-Q antenna ($Q \approx 17$). **Fig 13** shows the SWR variation with frequency for this feed arrangement. Again, the bandwidth is quite narrow. An 80-meter dipole is not particularly wideband either, typically exhibiting an SWR range of about 6:1 over the whole band. A dipole will have less extreme variation in SWR than the half-square.

Patterns Versus Frequency

Impedance is not the only issue when defining the bandwidth of an antenna. The effect on the radiation pattern of changing frequency is also a concern. For an end-fed half-square, the current distribution changes with frequency. For an antenna resonant near 3.75 MHz, the current distribution is nearly symmetrical. However, above and below resonance the current distribution increasingly becomes asymmetrical. In effect,

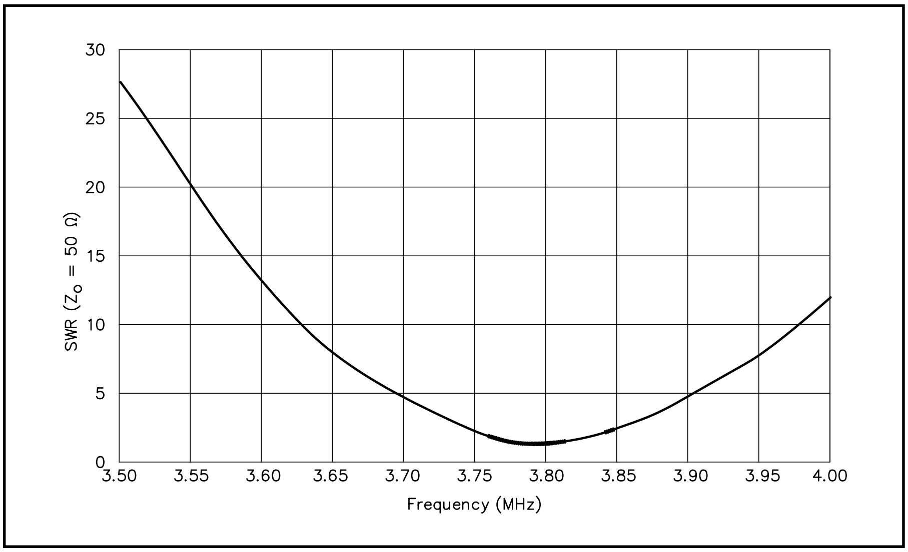

Fig 13—Variation of SWR with frequency for current-fed half-square antenna. The SWR bandwidth is quite narrow.

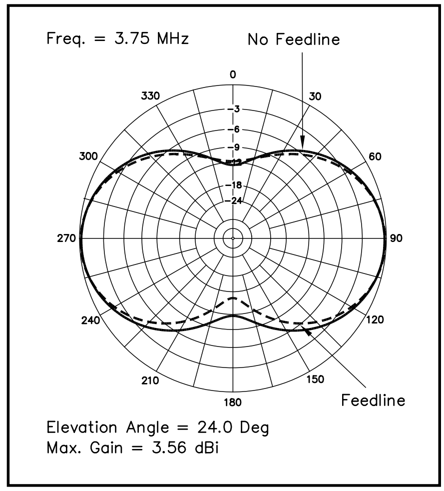

Fig 14—Effect of feed line on azimuth radiation pattern for current-fed half-square antenna. The feed line introduces only small distortions in symmetrical radiation pattern. The coaxial feed line was modeled as being brought out straight for 30 feet from the corner, then brought down close to ground level and led away for 50 feet more, where it was grounded.

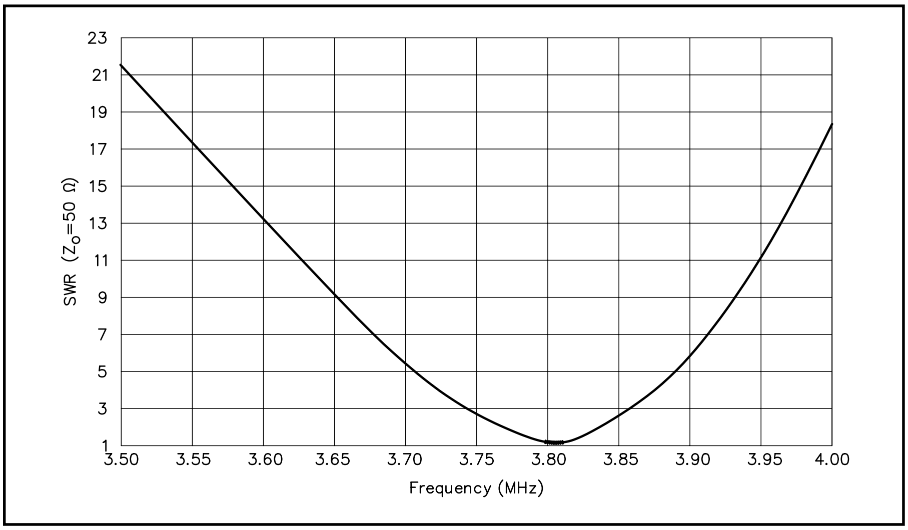

Fig 15—SWR versus frequency for voltage-fed half-square antenna, using matching network shown in Fig 10B, with L = 15 μH, C_1 = 125 pF and C_2 = 855 pF. The SWR bandwidth is less than 100 kHz at the 2:1 SWR points.

the open end of the antenna is constrained to be a voltage maximum but the feed point can behave less as a voltage point and more like a current maxima. This allows the current distribution to become asymmetrical.

The effect is to reduce the gain by −0.4 dB at 3.5 MHz and by −0.6 dB at 4 MHz. The depth of the zenith null is reduced from −20 dB to −10 dB. The side nulls are also reduced. Note that this is exactly what happened when the antenna was made physically asymmetrical. Whether the asymmetry is due to current distribution or mechanical arrangements, the antenna pattern will suffer. In my model, I used four ground wires, 10 feet long. These represent an adequate ground for the antenna when operated not too far from resonance. Even shorter wires could be used.

When corner-feed is used, the asymmetry introduced by off-resonance operation is much less, since both ends of the antenna are open circuits and constrained to be voltage maximums. The resulting gain reduction is only −0.1 dB. It is interesting that the sensitivity of the pattern to changing frequency depends on the feed scheme used!

Of more concern for corner feed is the effect of the transmission line. The usual instruction is to simply feed the antenna using coax, with the shield connected to vertical wire and the center conductor to the top wire. Since the shield of the coax is a conductor, more or less parallel with the radiator, and is in the immediate field of the antenna, you might expect the pattern to be seriously distorted by this practice. This arrangement seems to have very little effect on the pattern!

A number of different feed-line arrangements were modeled. An example of the patterns for one of them is shown in **Fig 14**. The wire, representing the outside of the coax feeding the antenna at the corner, was brought out straight for 30 feet, then brought down close to ground and led away for 50 feet more and grounded. The effect at resonance was barely detectable, as shown in Fig 14. At 3.5 MHz the gain was down by −0.5 dB and at 4 MHz was actually up by +0.1 dB. Other lengths and feed-line arrangements were tried with similar lack of effect. The greatest effect came when the feed-line length was near λ/2. Such lengths should be avoided.

Frankly, this result came as a considerable surprise. There are at least two possible explanations. First, the feed line is connected to a low-voltage point. Second, the feed line is located off the end of the antenna, where the field is canceled to some extent by the phasing of the radiators. Whatever the reason, this is very good news. It means that the antenna can be kept just as simple as a dipole.

Of course, you may use a balun at the feed point if you desire. This might reduce the coupling to the feed line even further but it doesn't appear to be worth the trouble. In fact, if you use an antenna tuner in the shack to operate away from resonance with a very high SWR on the transmission line, a balun at the feed point would take a beating.

Voltage-Feed at One End of Antenna: Matching Schemes

Several straightforward means are available for narrow-band matching. However, broadband matching over the full 80-meter band is much more challenging. Voltage feed with a parallel-resonant circuit and a modest local ground, as shown in Fig 10, is the traditional matching scheme for this antenna. Matching is achieved by resonating the circuit at the desired frequency and tapping down on the inductor in Fig 10A or using a capacitive divider (Fig 10B). It is also possible to use a $1/4\lambda$ transmission-line matching scheme, as shown in Fig 10C.

If the matching network shown in Fig 10B is used with $L = 15\ \mu H$, $C_1 = 125$ pF and $C_2 = 855$ pF, you will obtain the SWR characteristic shown in **Fig 15**. At any single point the SWR can be made very close to 1:1 but the bandwidth for SWR < 2:1 will be very narrow at <100 kHz. Altering the L-C ratio doesn't make very much difference. This antenna has a well-earned reputation for being narrowband. If you only want to DX on phone or CW then that may be acceptable, but most users want to do both.

It is possible to change the capacitors or tune the inductor, either with switches, manual adjustment or a motor drive. However, that level of complexity is unacceptable, especially since we are trying to replace a dipole with something equally simple. It is also possible to design wideband matching networks with multiple elements, but again that approach is relatively complex.

Current-Feed: Matching Schemes

The antenna can be current-fed at points other than the upper corners. Some possibilities are shown in **Fig 16**. As the feed point is moved away from the current maxima, the voltage increases and it becomes necessary to use a balun to decouple the transmission line. For narrowband use or if there is a matching network at the feed point this may be acceptable and may result in a more convenient feed point. As shown in Fig 16A, the feed point can be moved down the vertical wire to a higher impedance point and a 4:1 or 9:1 balun used. If the ends of the antenna are bent back toward the center, then a convenient feed point would be the lower corner, as shown in Fig 16B. By making the ends symmetrical as shown in Fig 16C even better decoupling could be obtained and the symmetry of the antenna is maintained.

Another possibility that has been used in the past is to invert the antenna, as shown in **Fig 17** and feed it at a lower corner. The problem with this approach is that the losses are higher because the current maxima are close to ground. A comparison between a normal half-square and an inverted one, 5 feet over average ground, is made in **Fig 18**. The difference is over 2 dB. For greater height or better ground, the loss would be lower. The killer antenna built by Tom

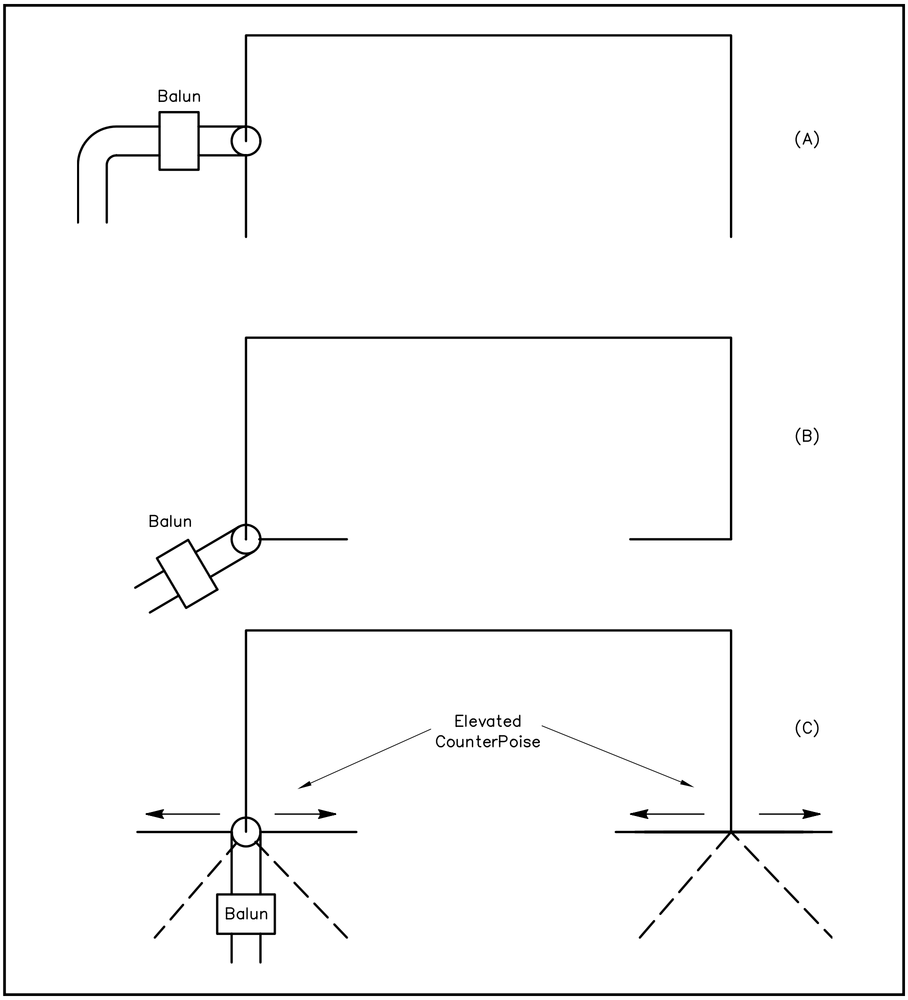

Fig 16—Possible methods for current-feeding of half-square antenna at points other than the upper corners. At A, a balun is used to decouple the feed line from the feed point at the center of one of the vertical legs of the antenna. At B, the ends of the vertical legs are both bent back horizontally to provide a feed point. At C, an elevated counterpoise is used to provide a feed point at the bottom of a vertical leg.

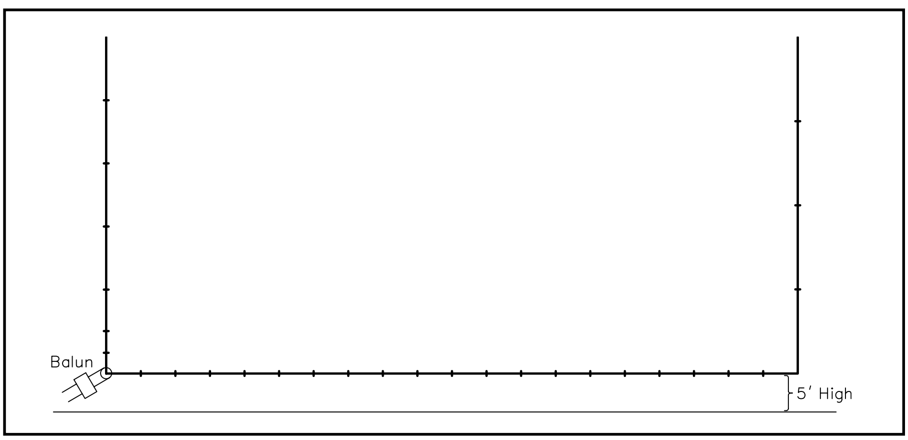

Fig 17—An "inverted half-square" antenna, current-fed at a lower corner. The losses in this configuration are excessive unless the ground under the antenna is exceptionally good, RF-wise.

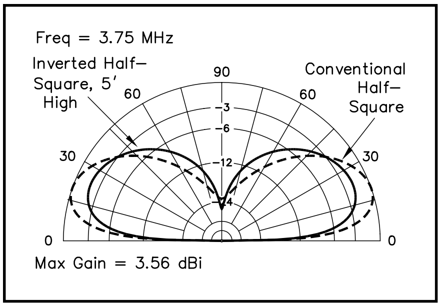

Fig 18—Elevation pattern for a conventional half-square, compared with an "inverted half-square" whose horizontal wire is located 5 feet over average ground. The difference is more than 2 dB.

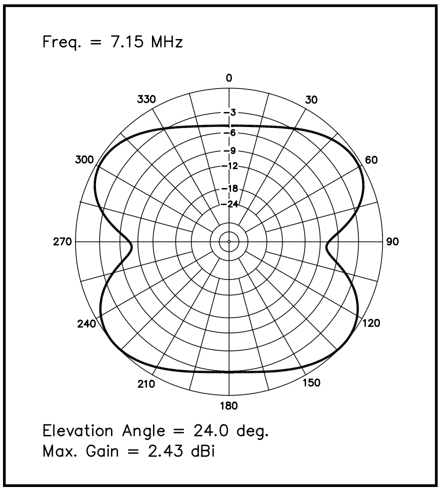

Fig 19—An attempt to load an 80-meter half-square antenna on 7 MHz. The pattern is badly distorted. The half-square is a monoband antenna!

Erdmann, W7DND, used this configuration but it was installed over a saltwater beach.[11] As a consequence the losses were very low and the feed point very conveniently located.

Multiband Operation

An 80-meter half-square can be used on other bands but the pattern and the drive-point impedance will change. A current-fed, 80-meter half-square will have a radiation pattern like that shown in **Fig 19** when driven at 7.15 MHz. On 40 meters the pattern has four lobes and the feed-point impedance is approximately $3300 + j\,1500\ \Omega$. If end-feed is used, the impedance will be in the region of $450 + j\,110\ \Omega$. With end-feed, the pattern will be somewhat asymmetric.

If the antenna is used on 20 meters the pattern will have eight lobes and the impedance at 14.2 MHz will $\approx 1100 + j\,900\ \Omega$. If a tuner is available this antenna can be used at higher frequencies but it will have a multi-lobed pattern typical of a harmonic antenna.

On the higher bands (40 meters and up), the height in wavelengths is greater for a given physical antenna height. Over average ground, the advantage of the half-square over a typical dipole thus becomes smaller and the half-square may even become inferior to the dipole. When the antenna is installed over very good ground or seawater, then the half-square may still be a contender on the higher bands.

Conclusion

The half-square antenna has some definite advantages. It is a simple and effective alternative to a typical dipole on the 80 and 160-meter bands, where the half-square radiates a stronger signal at the low angles most appropriate for DX work. The height and shape of the antenna are quite flexible and can be tailored to fit the needs of a given QTH. As a DX receiving antenna, it has the advantage of discriminating against strong high-angle signals arriving from stations within 1500 miles.

One disadvantage of the half-square is that it is more narrowband than a dipole—for DX work this may not be a serious disadvantage, since the ranges of frequencies for the DX "windows" are quite small. The antenna is also vertically polarized, which means more noise pickup when receiving.

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[7] *EZNEC* is available from Roy Lewallen, W7EL, PO Box 6658, Beaverton, OR 97007.
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[9] *NEC-WIN Basic* is available from Paragon Technology, 200 Innovation Blvd, Suite 240, State College, PA 16803, tel 814-234-3335.
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Monster Quads

By Rudy Severns, N6LF
PO Box 589
Cottage Grove, OR 97424

So you're dreaming about a really big antenna for 40 meters? N6LF tells us about his monster two-element 40-meter quad, with bonus three elements on 20 and 15 meters.

Quads are fascinating antennas. I've been afflicted with the desire to build them for over 40 years. The first one was a two-element job I built while serving with the Army in Germany (DL4ND/DL4SFG) in the 1950s. To this day I'm not sure how I managed to shinny up a 70-foot pole to install the quad, but I was determined to get it working! And work it did!

True madness did not come upon me until I read Lindsay's 1968 article on quads.[1] I promptly built a six-element 20-meter monster (on a 60-foot boom) using the dimensions in the article. While I was at it, I added six elements on 15 meters and 11 elements on 10 meters. During initial testing using a small exciter (about 10 W) the first contact was the Russian Antarctic station, long path. After that I was hooked—and it's been all downhill from there!

If you live in an area where heavy icing is a regular occurrence, this article should be saved for April 1st. At my present QTH in western Oregon we seldom have ice storms. The most severe in the past ten years put about 1/2 inch of ice on my quad. The distortion was alarming but no permanent damage occurred. More ice than that, however, would start to break things. Perhaps it is possible to build quads to stand up to heavy icing, but I doubt it is worth the trouble for antennas of the monster size discussed in this article.

The antennas I describe are large and require significant time, effort and money to implement. The point of this article is to give you some useful ideas and perhaps some inspiration. I have included the dimensions, performance predictions, many mechanical details and some of the mistakes I made along the way. You can, of course, replicate any of these antennas directly but you will get better results if you consider them a starting point and then design an antenna to meet your own particular needs and preferences.

Modeling

Good *NEC-2* and *NEC-4*-based software is now available and is a worthwhile investment for a project of this size. Quads are generally lower-Q antennas than comparable Yagis and therefore somewhat less sensitive to dimensional variations, supporting structures and interlaced multi-band elements. But you will still find the best results can be had only by modeling the complete structure and designing for your particular needs.

I did all my modeling for this article using *GNEC-4*, which is *NEC 4.1*-based.[2] I modeled the WØHTH six-element quad in free space, while my 40/20/15-meter multiband quad is centered at 100 feet, over average ground ($\varepsilon = 13$, $\sigma = 0.005$ S/m) using the Sommerfeld ground model.

While there is remarkably little interaction between elements of different bands, you must take some care to prevent unexpected resonances due to the matching sections and the open-circuited feed lines on those driven elements not in use. I fed each driven element separately and led the feedline to a multi-pole coax relay mounted at the center of the boom. I used the modeling program to select lengths of feedline that did not result in any spurious resonances that would upset the performance on another band. This was not very difficult, but required some attention.

WØHTH Quad

Just for old time's sake. I went back and took a look at the multi-element quad I built in 1968 to Lindsay's dimensions (see **Table 1**). In those days I didn't have a computer on my desk to do antenna modeling, so I just relied on his information. The results had been great, but I was curious to see how modern modeling would compare with Lindsay's experimental work. Based on his

Table 1
Dimensions of 20-Meter WØHTH Six-Element Quad

Element	*Location (ft)*	*1/4 Length (ft)*
Reflector	0	18.04
Driven	12	17.60
Director 1	24	17.28
Director 2	36	17.28
Director 3	48	17.28
Director 4	60	17.32

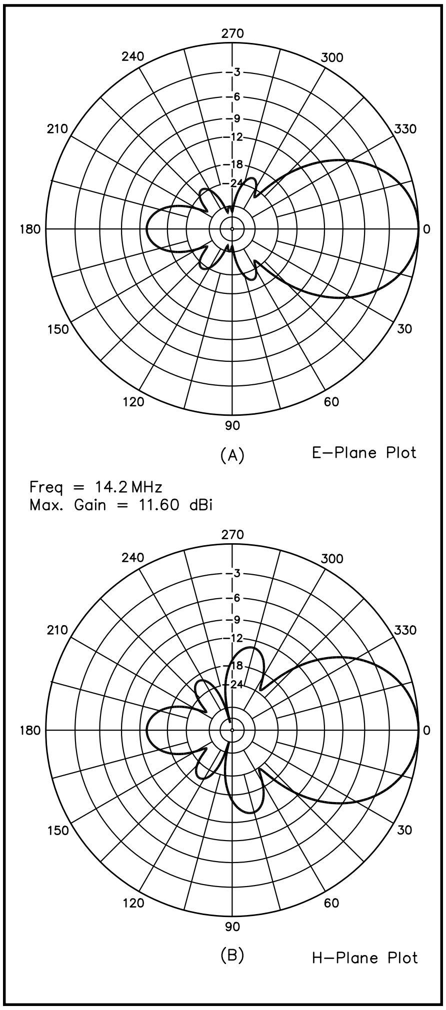

Fig 1—WØHTH 6 element 20-meter quad, free-space radiation patterns. At A, E-plane pattern and at B, H-plane pattern.

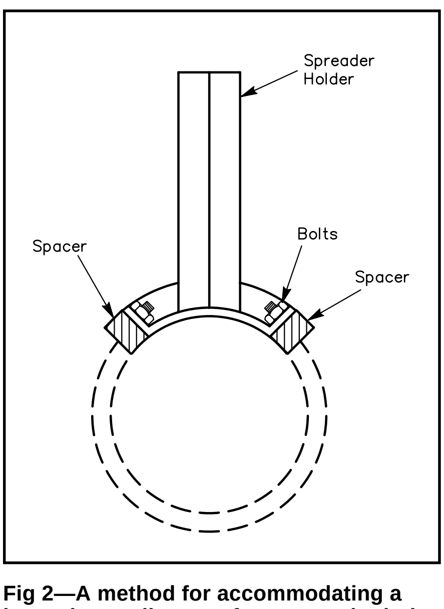

Fig 2—A method for accommodating a larger boom diameter for a spreader hub.

experimental work at 440 MHz, scaled down to 14.2 MHz, Lindsay predicted a forward gain of 13.4 dBi. It is hard to tell exactly what the F/B ratio is from Figure III-3A in his article but it looks to be roughly 15 to 20 dB.

With *NEC-4.1* software the predicted pattern is shown in **Fig 1**. I computed a forward gain of 12.1 dBi and a F/B of 14 dB. This was not too bad, but I suspect that these numbers could be improved with a bit of fiddling with the model. I remember that I adjusted the reflector length slightly for maximum F/B when I built the antenna, and the F/B was quite good.

For 15 meters I scaled the 20-meter element dimensions (retaining the same element spacing) and then adjusted the reflector for maximum F/B and the driven element for resonance. On 10 meters I again scaled the dimensions and adjusted the reflector, but in the true ham spirit of "If a little is good, more should be lots better," I added five more directors. I spaced each 10-meter element by 6 feet.

I made my boom from two 30-foot lengths of 4-inch-OD irrigation pipe. Most commercial spreader hubs are designed for 3-inch, not 4-inch, diameter booms, so to accommodate the larger boom, I placed spacer blocks between the hub sections. See **Fig 2**. This worked well for hubs made from four separate pieces. However, some commercial hubs have one-piece castings and can't be expanded like this. To match to 50-Ω feedline I used λ/4 75-Ω transmission line sections on each band.

With an 11-element quad on a 60-foot boom, I noticed some interesting propagation effects on 10 meters. On several occasions the band dropped out during a transcontinental QSO, with signal strengths dropping from S9+ to just above the noise level. Nonetheless, we were able to continue the QSO for an extended period of time, when for all intents and purposes the band was dead. There is nothing like a big antenna!

A Two-Element 40-Meter, Three-Element 20 and 15-Meter Quad

In 1989 I built another quad based on Lindsay's article. It had five elements on 20 and 15 meters—and nine elements on 10 meters—on a 50-foot boom. I did not yet have antenna modeling software so again I just used Lindsay's dimensions. The antenna worked very well but it also provided me with a lesson on wind loading and wind strengths on mountaintops in Oregon. The antenna itself stood up very well, but my

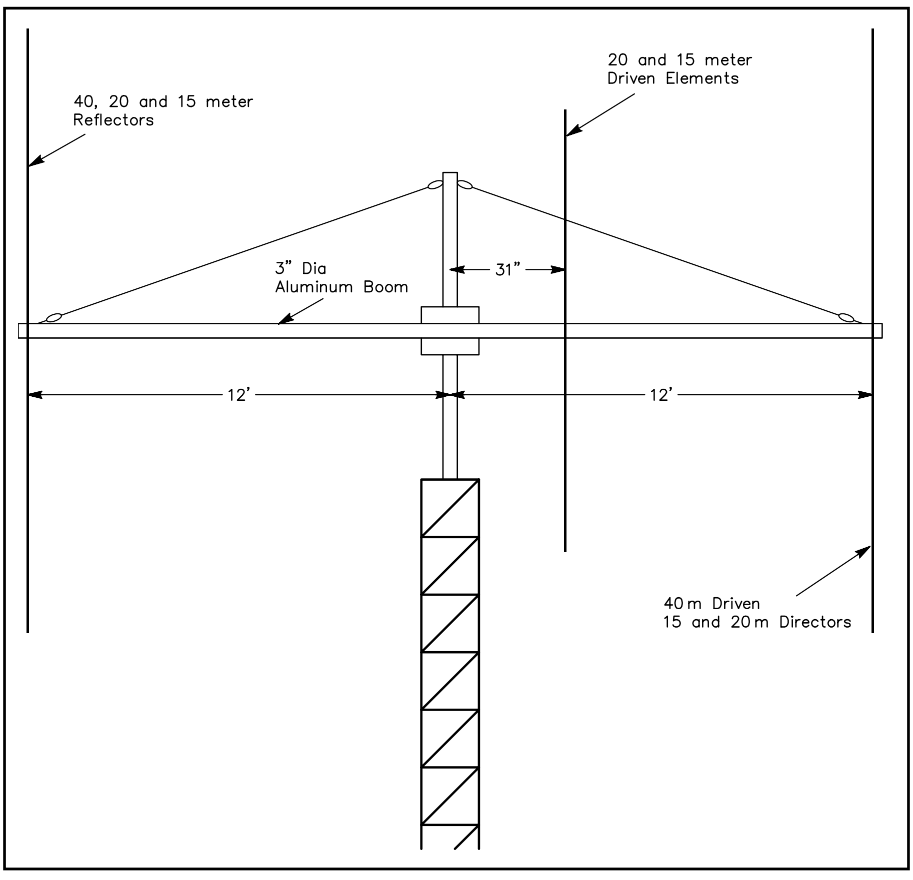

Fig 3—N6LF three-band quad dimensions: two elements on 40 meters; three on 20 and 15 meters.

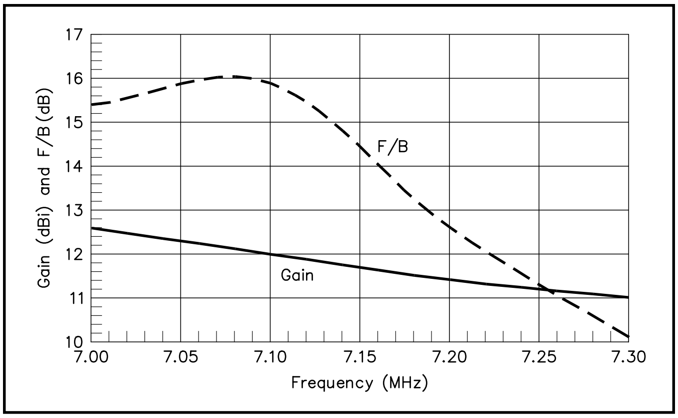

Fig 4—Gain and F/B characteristics on 40 meters.

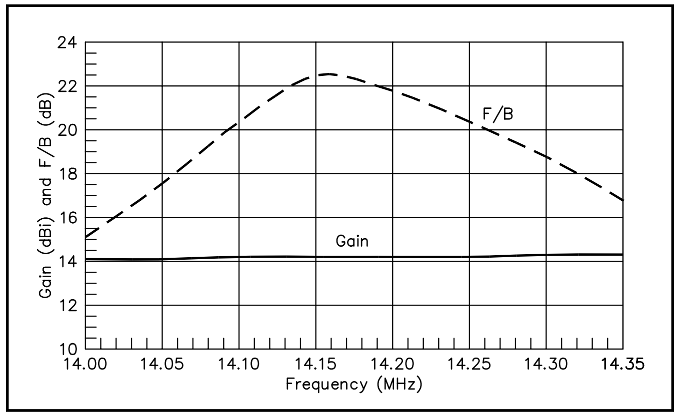

Fig 5—Gain and F/B characteristics on 20 meters.

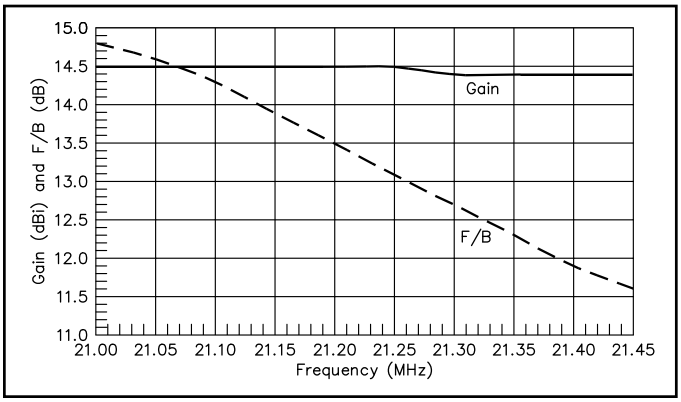

Fig 6—Gain and F/B characteristics on 15 meters.

Table 2
40, 20 and 15-Meter Quad Dimensions

Element	*Location on Boom (ft)*	*¼ Length (ft)*	*Spreader Length (ft)*
40-meter Reflector	0	36.8	26.02
20-meter Reflector	0	18.4	13.01
15-meter Reflector	0	12.5	8.84
20-meter Driven	14.58	17.96	12.70
15-meter Driven	14.58	11.84	8.37
40-meter Driven	24	35	24.75
20-meter Director	24	17	12.02
15-meter Director	24	11.58	8.19

72-foot unguyed tower collapsed in the first real storm that year. The antenna did not take kindly to this!

I replaced this system with a new tower and a monoband six-element 20-meter Yagi. The Yagi worked great but was only good for 20 meters. At that time the sunspot cycle was headed down the tube and it was clear that 40 meters was going to be a big-time DX band for the next several years. However, 20 meters was not going to go away, and there would be some openings on 15 meters even at a sunspot low. So I began to think about a new multi-band quad with full-size 40-meter elements and good performance on 20 meters.

Of course 40-meter elements are twice as long as the 20-meter elements I was accustomed to, so it was pretty intimidating. The wingspan is over 50 feet! I found a source for the spreaders and the hub hardware,[3] and I now had good modeling software to design the antenna, so I went ahead with the project. The antenna has been up since 1993 with no real problems. It has proven to be durable, practical and a very good performer. It is a real killer on 40 meters. **Fig 3**, along with **Table 2**, gives the dimensions and element arrangement of the antenna.

The predicted gains and front-to-back (F/B) ratios for the three bands are given in **Figs 4** to **6**. The 40-meter band is wide enough that it is very difficult to obtain high gain and high F/B over the entire band with a simple two-element array. I chose to emphasize the CW end of the band, and this can be seen in Fig 4. I could have moved everything up in frequency and improved the phone-band performance but that would have meant a poorer F/B in the CW band. In my design, the F/B peaks at 16 dB and is above 15 dB over the entire CW part of the band. The gain peaks just outside the lower band edge and I could have traded a bit of F/B for a little more gain. The old rule that you can't have peak gain and peak F/B at the same frequency definitely applies.

The 20-meter performance is very good. In this case I chose to optimize at roughly midband; Fig 5 shows a minimum F/B of 15 dB over the entire band, with a peak F/B of greater than 22 dB. The gain is also very flat over the entire band. Overall, this is a very nice compromise for a three-element array.

This is a good point to go back to Fig 3 and discuss the choice of boom length and element placement. Normally a two-element array has a boom length of 0.12 to 0.15 λ for best performance. That would have resulted in a 16 to 20-foot boom for 40 meters. However, even at a sunspot low 20 meters is still a workhorse DX band and I wanted to have a really good three-element array on that band. Thus I made the boom a few feet longer to improve the 20-meter performance. The result on 40 meters was to slightly reduce the gain and F/B, but the longer boom had the advantage of presenting an approximately 112-Ω feed-point impedance. This could be easily matched with a 75-Ω (RG-11) λ/4-matching section. The greater spacing also broadbanded the antenna somewhat on 40 meters, which in the end more than compensated for the lower peak gain and F/B.

If you look closely at Fig 3 you will see something unusual. Because the elements in a three-element quad extend well below the boom, the middle element must be moved off center to stay well clear of the tower. In most designs the driven element is moved closer to the reflector. In my case, however, I went the other way because I felt it gave me a better set of compromises. The 20 and 15-meter driven elements are closer to the director. This gave me very nice perfor-

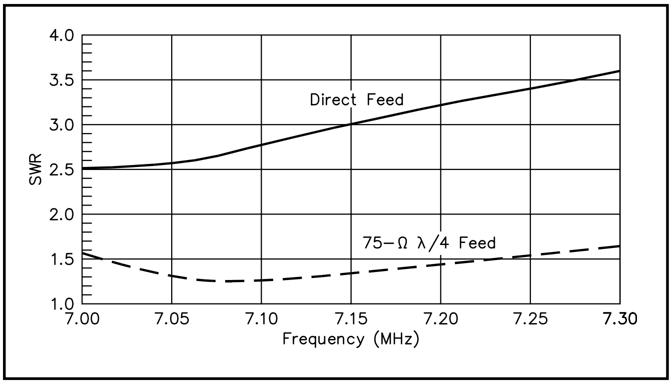

Fig 7—Solid line shows 40-meter 50-Ω SWR with direct feed. The dashed line shows the 40-meter 50-Ω SWR using a quarter-wave 75-Ω matching section.

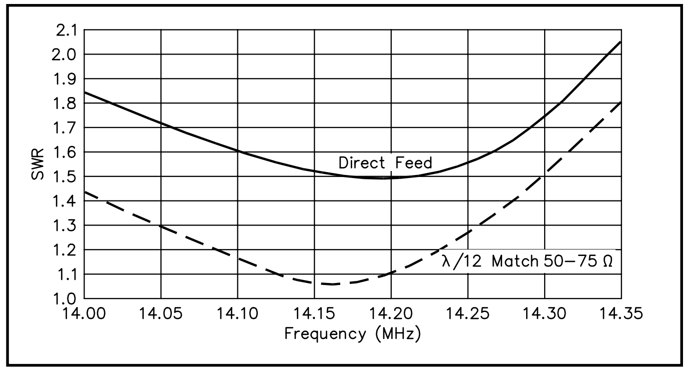

Fig 8—Solid line shows 20-meter 50-Ω SWR with direct feed. Dashed line shows 20-meter 50-Ω SWR with a 50 + 75-Ω series transformer.

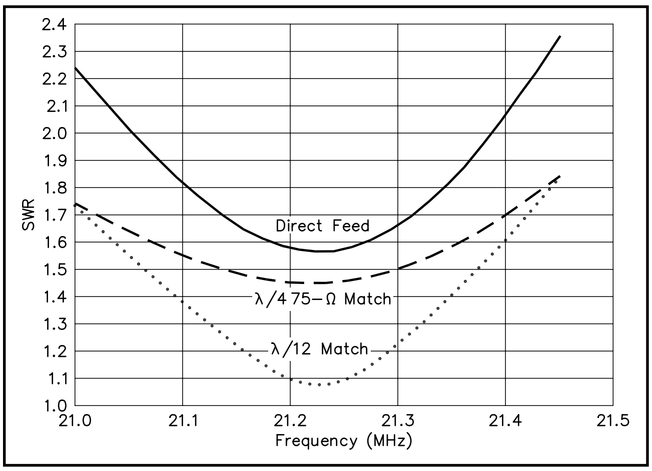

Fig 9—Solid line shows 15-meter 50-Ω SWR with direct feed. Dashed line shows 15-meter 50-Ω SWR with quarter-wave 75-Ω matching section. Dotted line shows 15-meter 50-Ω SWR with a series transformer match.

mance on 20 meters but compromised the F/B on 15 meters. Since I did not expect 15 meters to be a primary DX band during the sunspot minimum I accepted this. This reduced 15-meter performance is shown in Fig 6. The gain is good and very stable over the entire band but the peak F/B is low. I have again deliberately emphasized the CW end of the band just from personal preference.

Figs 7, **8** and **9** show the SWR performance for several matching choices. On 40 meters, if you do no matching, the SWR will be unacceptable (solid line in Fig 7). By adding a λ/4-matching section of 75-Ω line the match is very good over the entire band, as is shown by the dashed line in Fig 7.

On 20 meters you do not have to use a matching section, since the SWR is less than 2:1 over all but the uppermost portion of the band (solid line in Fig 8). However, because I have a nearly 200-foot run of cable, every little bit of loss hurts. I used a twelfth-wave or series-section transformer using 50-Ω (RG-213) and 75-Ω (RG-11) sections.[4,5] The result is shown in Fig 8 as the dashed line.

On 15 meters there are several possible choices. The solid line in Fig 9 shows the SWR for no matching. It is acceptable over most of the band but not at the band edges, especially considering my long feed line. The dashed line in Fig 9 illustrates the effect of a λ/4 75-Ω matching section and the dotted line in Fig 9C shows the effect of a λ/12 matching section. The λ/12 match is better near midband but about the same as the λ/4 section at the band edges. I chose to go with the slightly simpler λ/4 section.

The forgoing discussion illustrates some of the design trade-offs that you must make. It is for this reason I suggested earlier that these designs are more for inspiration than exact replication. You must decide for yourself what the trade-offs should be.

Just for the curious, because of the harmonic relationship between 15 and 40 meters, the 40-meter antenna has a low SWR on 15 meters and can even be operated on that band. There is some gain but the F/B is essentially 0 dB.

Some Mechanical Details

The support hub for the 20/15-meter driven elements was a standard commercial cast-aluminum piece made for 20-meter quads. These hubs are, however, totally inadequate for a 40-meter quad. **Fig 10** is a sketch of the welded hub assemblies (two each) I used for the 40-meter spreaders. These hubs are made from 3/8-inch aluminum plate. I obtained these from the same source as the long spreaders but you could fabricate them yourself.[3]

Wire! A big quad uses a lot of wire. Over the years I have used many different kinds of wire for the elements, ranging from copper house wire, solid Copperweld and stranded Copperweld. In an antenna this large the wire is a key structural element and it must have considerable strength in order to give years of service. Solid or even stranded pure copper wire is unsatisfactory, mainly due to rapid work-hardening from the constant motion of the spreaders as the wind blows. For this antenna I used #13 AWG stranded copperclad steel wire with high-density polyethylene insulation.[6] For some time I used an uninsulated version of this wire but even though I live in a rural area with no pollution, acid rain or salt atmosphere, I found that the wire still corroded. This potentially could weaken the wire and might increase losses. The insulated wire is more than strong enough and shows very little sign of corrosion even after several years. The wire size is also large enough to keep the losses acceptable (≈ 0.2 dB, according to the model).

When I first built this antenna I made some basic errors in the boom diameter and wall thickness, and in the guying (or the lack thereof). When I took down the 20-meter Yagi, I used two sections of that boom for the new quad. The boom tubing was 3 inches in diameter with quite thin walls (≈ 0.060 inch) and I used only a single support guy to each end, as shown in Fig 3. That was not good enough—after a couple of windstorms the boom started to bend sideways. No doubt some better engineering up-front would have told me that!

If you want to use 3-inch thin-wall tubing you must use side guys. There is simply too much mass and wind loading, even though the lever arm is only 12 feet long. Besides side guying, another approach would be to use larger-diameter tubing with a heavier

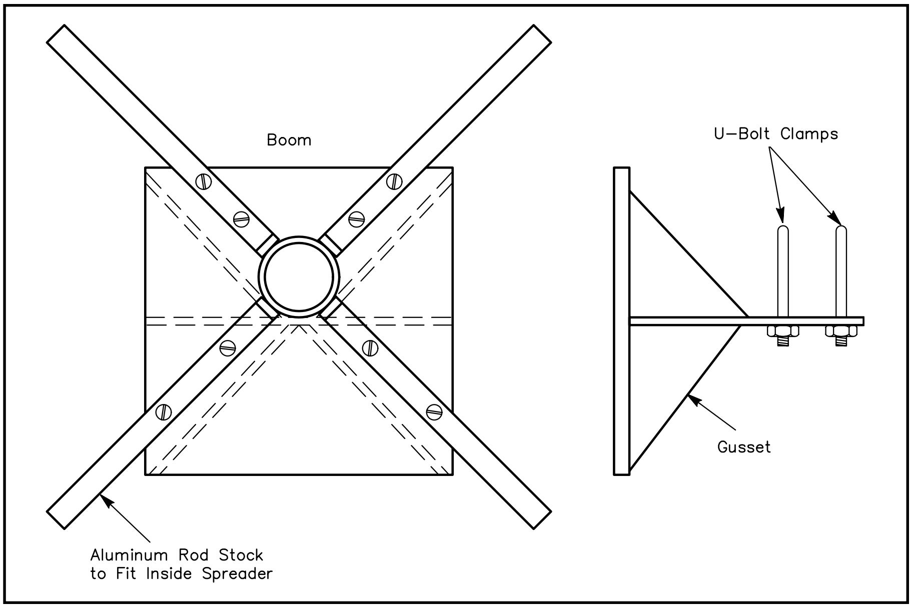

Fig 10—40-meter spreader hub design.

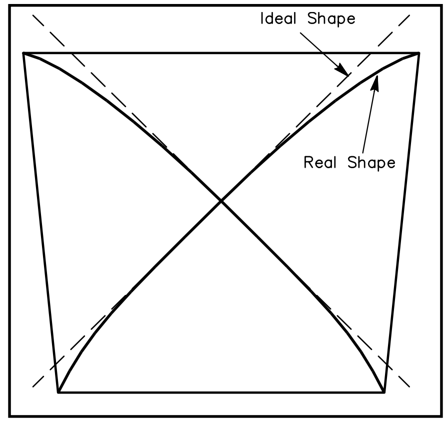

Fig 11—Distortion due to the weight of the spreaders when not anchored at the corners.

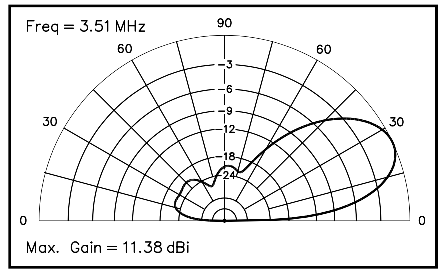

Fig A—80-meter quad elevation pattern at 3.510 MHz, at 100 feet above flat, average ground, including losses in the copper wire elements and Q = 250 inductor loading.

wall. For antennas larger than this example you will certainly have to do both. Since designing this antenna I have obtained copies of Leeson's book[7] as well as articles by Weber[8] and Bonney[9] on the mechanical design of large arrays. These have shown me the error of my ways and I strongly recommend you read them for any new design.

The spreaders for a 40-meter quad are twice as long as the 20-meter ones. They are also much heavier—3 to 5 times heavier. In my past experience with 20-meter quads the droop in the spreaders was very small and I used only a light wire jumper across the corners to keep the wire from sliding through the corner holder. In this antenna the stress on this wire was much higher and one jumper promptly broke, allowing the wire to slide through the corner mounting devices. This in turn allowed the spreaders to droop. The result was distortion in the shape of the loop like that shown in **Fig 11**. The shape is more like a trapezoid than a square. At first I though this was no big deal but a quick check showed the F/B had practically disappeared at the low end of the band.

Modeling the "new" shape showed that in fact the peak F/B had moved up to the high end of the band and the gain was degraded. The lesson is: Solidly anchor the corners of the elements to the spreaders. Realize that there will be a substantial load on this anchor due to the dead weight of the spreaders and the wind loading.

A commercial spreader hub for 20-meter and higher frequency quads usually resembles the one shown in Fig 2. While they are generally pretty reliable, I wanted something more rugged. What I did was to use two hubs, facing each other, trapping the spreader ends between the two faces of the hubs. The result is a much stronger anchor at the base of the spreaders.

Any large array requires a first-class rotator. I have been using an Orion OR-2300 rotator. It has given me more than a little heartburn, but then again I did have practically the first one sold. The manufacturers have been very responsive to problems and I believe the latest version (OR-2800) is a first-class rotator. The average ham rotator won't cut it in this league. With the large mass of the 40-meter spreaders and the heavy-duty hubs at the ends of the array, the moment of inertia is large.

Once you get the array rotating, the rotator has to bring it to a halt again. This can result in high stress on the rotator and also on the top of the tower itself. I can see the whole top of my tower twisting a bit as the rotator applies the brakes. To protect everything I have adopted the policy of using a low rotator speed for small angular changes. For a large change in direction I use a faster speed initially but then slow it down with the speed control as I approach the desired heading.

After the collapse of my old tower I installed an 89-foot motor-driven telescoping model. You can believe I am now a fanatic about keeping the tower down except while actually using it. I don't think the insurance company would be nearly so nice a second time. If a particularly severe storm is expected I will often throw a line over the boom and lash it down to ease the strain on the rotator.

More Madness

Because the present antenna has survived many years of hard use, it's obviously too small. I am in the processes of designing a new antenna, now that the sunspots are back. (By the way do you know how you can tell that Shakespeare was a 160-meter man? Who else would say, "Out, out, damned spot"?)

The new antenna will have three elements on 40 meters, five elements on 15 meters and nine elements on 10 meters. Four of the 10-meter elements will be Yagi-style dipoles, because for single-band elements they are simpler mechanically (not to mention the fact that I have a 10-meter Yagi I can cannibalize). The tentative boom length is 50 feet, which is reasonable in the light of my earlier work. I may also include elements for the 30, 17 and 12-meter bands but that is still to be determined. Perhaps this will be a topic for *The ARRL Antenna Compendium, Vol 7*.

The ultimate madness is on the drawing boards also. A full-size two-element 75/80-meter quad. I intend to tune this behemoth to cover the entire band with a simple relay scheme. See the sidebar for a brief description. Stay tuned for the next installment—coming to you as soon as I can get leave from the asylum!

The Ultimate Insanity

As shown in Ref 11, it is possible to build a full-size, rotary, two-element quad for 75/80 meters. There are two problems to be solved: First, how to tune it remotely so that I can have good performance in at least the two DX windows (3.510 and 3.790 MHz) or better yet, over larger sections of the band. Second, how to solve the mechanical problems imposed by the need for spreaders nearly 50 feet long and boom more than 50 feet long.

Bandspreading the antenna is not just a matter of an acceptable SWR. You also need to keep the gain and F/B as near peak values as possible. If you are going to all the trouble to build this monster there is no reason to compromise! I expect that I'll design the basic quad for the higher end of the band, say 3.850 or 3.790 MHz and then use relays to add in a small amount of inductive loading in both the reflector and the driven elements. If the elements are already near full size then the amount of loading will be small and will introduce very little loss. Of course, the inductors must still be designed for high Q. I will try to optimize the antenna at 3.790 MHz with the loading inductances shorted out with relays and then open the relays for 3.510 MHz operation.

Table A shows the typical dimensions for such an antenna, on a 44-foot boom at 100 feet above average ground. The elevation radiation pattern at 3.510 MHz is given in **Fig A**. Note that the effect of wire and inductor losses are included in this model. In the right location this would be a dominating antenna. By adjusting the loading inductances, this kind of performance could be available at any point in the band.

Because it is not necessary that the entire length of the spreader be insulated, 40-meter fiberglass spreaders could be extended with 2-inch-OD aluminum tubing. In effect, the hub would have a 44+ foot diameter. Modeling work indicates that this large a mass of metal inside the perimeter of the antenna would have little effect on the performance, so long as the longer support guys are broken up with insulators. I'd probably make the support guys from Kevlar or other insulating material.

The hub is designed along the lines of a bicycle wheel and shown conceptually in **Fig B**. Note that Fig B is only for the hub, the fiberglass spreaders are mounted on the ends of the hub arms. Two of these hubs, one on each end of the boom, would be needed. Obviously the boom will have to be guyed to support the weight.

Table A
75/80-Meter Quad Dimensions

Element	*1/4 Length (ft)*	*1/2 Diagonal (ft)*
Driven Element	65.4	47
Reflector	68.2	49

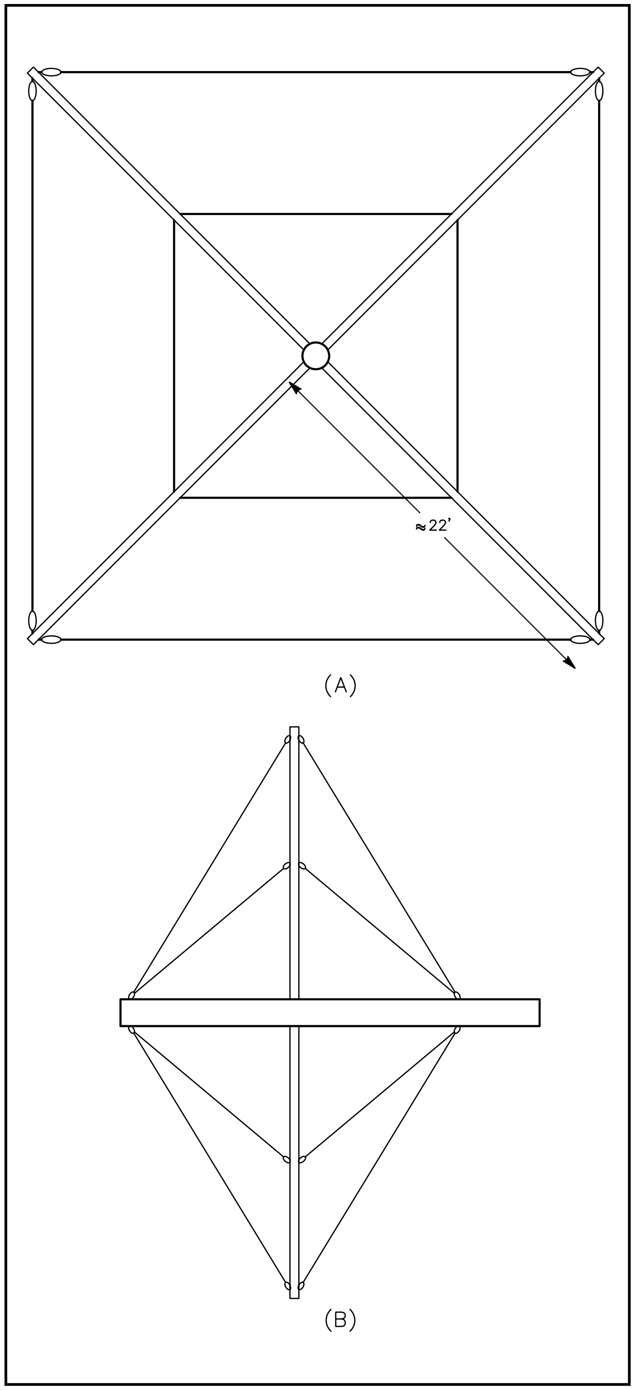

Fig B—80-meter quad conceptual spreader-hub design.

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[1]J. Lindsay, Jr, WØHTH, "Quads and Yagis," *QST*, May 1968, pp 11-19, 150.

[2]*GNEC-4*, Nittany Scientific, Inc, Hollister, CA 95023, 408-634-0573.

[3]Lightning Bolt Antennas, RD 2 Rte 19, Volant, PA 16156, 724-530-7396.

[4]*The ARRL Antenna Book*, 18th Edition (Newington: ARRL, 1997), pp 26-4, 26-5.

[5]D. Emerson, AA7FV, "Try a Twelfth-Wave Transformer," *QST*, Jun 1997, pp 43-44.

[6]The Wireman, Inc, 261 Pittman Road, Landrum, SC 29356, 803-895-4195.

[7]D. Leeson, W6NL (ex-W6QHS) *Physical Design of Yagi Antennas* (Newington: ARRL, 1992).

[8]D. Weber, K5IU, "Determination of Yagi Wind Loads Using the Cross-Flow Principle," *Communications Quarterly*, Spring 1993.

[9]S. Bonney, K5PB, "Practical Application of Wind-Load Standards to Yagi Antennas: Part 1," *QEX*, Jan/Feb 1999, pp 46-50.

[10]Another source for quad parts: The Antenna Mart, PO Box 699, Loganville, GA 30249, 770-466-4353.

[11]J. Devoldere, *Antennas and Techniques for Low-Band DXing,* 2nd Edition (Newington: 1994), Chap 13, Sec. 4.

Short Radials for Ground-Plane Antennas

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Think your elevated radials always have to be full size? N6LF lets you in on some great ideas to lessen the "wingspan" of radials, especially near the beach.

Dean Straw's (N6BV) article[1] in this book describing the 6Y2A operation and beach-front verticals for DXpeditions shows how useful a vertical or vertical array can be—if you can put it over or adjacent to saltwater. For 20 meters and higher in frequency it is practical to use λ/2 verticals with little or no ground plane. For 40 meters and lower in frequency, however, a λ/2 height becomes prohibitive and a λ/4 ground-plane with elevated radials is a more practical form of vertical. Unfortunately, as you go down in frequency the length of the quarter-wave radials becomes very long (approximately 132 feet on 160 meters), and this takes up a lot of area. In addition, most DXpeditions can't place the radials very high off the ground. This results in a number of wires to trip over or strangle on. And if you have several verticals, the beach really becomes an obstacle course!

One way to reduce the problem is to shorten the radials (leaving the vertical part of the antenna as near a quarter-wave as possible) and use either a loading inductor, a top-loading hat or some combination of the two. The question is, "How much do you lose as you shorten the radials?" I took a look at this using *GNEC-4*, a *NEC-4.1*-based modeling program, and the following is what I discovered. Keep in mind of course that all this assumes *NEC* knows what it's talking about!

A 160-Meter Vertical

I have been planning on a beach-front 160-meter vertical for some property I have on Willapa Bay, WA, so I started with that model. While you probably wouldn't try to construct a full-size quarter-wave 160-meter vertical for a DXpedition, the computed results are very similar no matter what band you use with a beach-front vertical. The antenna I am planning will use four elevated radials and will be made of #13 wire. I was planning to use a wooden A-frame made from three Douglas fir trees (as shown in my *QEX* article[2]) to support the antenna. For modeling, the initial lengths of the radials and the vertical were made equal and adjusted for resonance at 1.840 MHz. I then progressively shortened the radials (keeping the vertical height the same) and re-resonated the antenna with a single series inductor feeding all four radials at the feed point.

An inductor Q of 250 was assumed and with a little care this should be readily achievable. In a salt atmosphere you must put the coil in a sealed enclosure, or by morning the Q will be close to zero. I modeled the base of the antenna at 1 foot and at 10 feet, and for comparison used three types of ground: perfect, seawater ($\varepsilon = 80$, the dielectric constant, and $\sigma = 5.0$ S/m, the conductivity) and average ($\varepsilon = 13$, $\sigma = 0.005$ S/m).

The results are shown in **Figs 1** and **2**. These graphs include both the ground loss and the loss due to the series resistance of the loading inductor. The small wire loss was not included. We can see from Fig 1 the advantage of seawater over average ground: about 4.2 dB more gain for full-length radials. In addition, the peak gain occurs at an elevation angle of 7° for seawater, as opposed to 21° for average ground. As the radial length is reduced the peak gain angle changes very little, but the peak gain goes down. The height of the radials over seawater made very little difference, and the difference between ideal ground and seawater was also very small. The primary difference over seawater is the added loss in the loading inductor.

While Fig 1 shows the peak gain, you can see the variation much better in Fig 2, where the *change* in gain is plotted. Even if you use radials only 40 feet long (0.07 λ!), over seawater the loss is less than 0.2 dB. This is very attractive for DXpeditions. The value of the loading inductor is very nearly the same for all the grounds and heights so that the loss due to the inductor's series resistance is pretty much the same at each radial length. Over average ground, however, the gain reduction is much larger due to in-

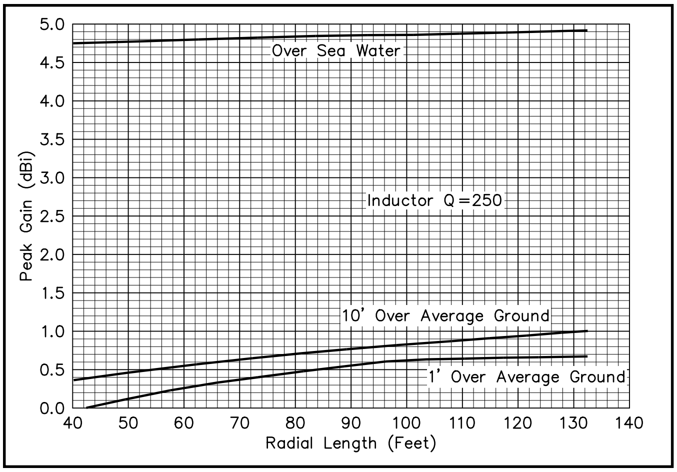

Fig 1—Peak gain for (λ/4-high vertical with four, coil-loaded short radials on 160 meters. The loading coil is assumed to have an unloaded Q of 250. Over seawater, the peak gain doesn't change much, even for quite short radials, while the gain is close to 5 dB less over ground with average conductivity and dielectric constant.

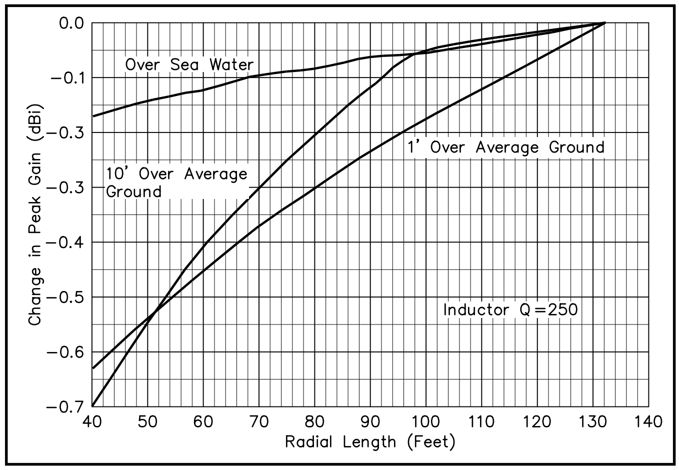

Fig 2—The same data presented in Fig 1, but magnified by showing the change in peak gain versus radial length on 160 meters.

creased ground losses as the radials are shortened.

At some sites the antenna may actually be over seawater, but it is more likely it will be up on the beach adjacent to seawater. How much effect will that have? That depends on two things: the beach's ground characteristics and the distance to the water. If the ground under the antenna is regularly flooded with seawater the conductivity is going to be pretty high. But that may not always be the case, and fairly poor ground characteristics may be encountered—especially on coral islands.

To check this out I modeled the 160-meter antenna site as though it were a circular island located in a sea of saltwater. The island was made up of ground with average conductivity and dielectric constant, and the distance to the saltwater was varied by changing the radius of the island. The results are shown in **Fig 3**. As soon as you move away from the water (that is, you have a larger-diameter island) the peak gain starts to drop and the increased ground loss due to shorter radials shows up.

The message is simple—select a nice salt marsh that is flooded twice a day, or put the antenna out on the reef with water under it! Otherwise, put the antenna as close to the water as practical.

Reducing Losses

We know that the losses due to use of a loading device such as a coil can be reduced by using a higher-Q coil, by moving a loading coil from the base of the antenna to a more optimum location up the vertical radiator, by top-loading or some combination of these.[3,4] For a $^1/_4$-λ vertical with shortened, loaded radials over seawater the ground losses are very small, and even the losses due to base loading of a shortened vertical radiator are small. It is questionable, therefore, whether it is worth the trouble to spend much time trying to minimize the loading loss, except for the case where a vertical's electrical height is much shorter than λ/4.

Unfortunately, short verticals with loading are often used for 80 and 160 meters. On those bands all of the tricks for minimizing losses will have to be used, because shortening the radials as well as the vertical itself can seriously degrade performance, even with a seawater ground.

I looked at modifying the ground plane to see if a more complex radial structure would help. Using eight radials, the difference in ground loss was insignificant. However, the additional radials did reduce the reactance needed to resonate the antenna by almost $^1/_2$. That would reduce the loading coil loss, since a smaller amount of inductance would be needed.

I then looked at tying the ends of the radials together with cross conductors to form a square wagon-wheel shape with four radials. Again, the ground losses were not reduced greatly (≈ 0.2 dB). There appears to be no substitute for long radials if you want that last fraction of a dBi in gain. We really have known this since the 1930s![5,6,7] The reactance, however, was greatly reduced and with the wagon-wheel structure the antenna is resonant—without loading—with a radius of 58 feet, less than half that for normal radials.

Given the fierce corrosion experienced over or near seawater, it would be a good idea to use insulated wire for the radials, some paint on the vertical tubing, conductive joint compound and very careful sealing of all joints and connections.

A Closer Look at Ground Losses

The increase in ground loss with shorter radials is worth a closer look. The additional loss shows up as an increase in feedpoint resistance over that for ideal ground. **Fig 4** is a graph of feedpoint resistance as a function of radial length, without the resistance of the loading inductor. Over seawater the effect of ground loss is very small. It's hard to see it on the graph. Over average ground, however, the effect is very obvious and the loss increases at lower heights.

When generating the data for Fig 3, I noticed that the feedpoint resistance was constant for different values of "island" radius. This is due to the way *NEC* computes impedance, where it takes into account only the first ground characteristic and assumes for this purpose that the ground under the antenna is infinite. For far-field calculations, however, the two ground zones (that is, the ground under the vertical and the seawater surrounding our model island) are taken into account. This means that the ground loss in the model, as reflected in the feedpoint resistance, may be higher than it actually is when close to the water.

NEC can provide a direct calculation of total ground losses using the so-called "RP" card. This card sets the parameters for radiation patterns and can provide a calculation of average gain. An example is given in the Appendix.

Conclusions

If you are lucky enough to be near or on seawater, you can drastically reduce the size of your elevated ground-plane. With a little

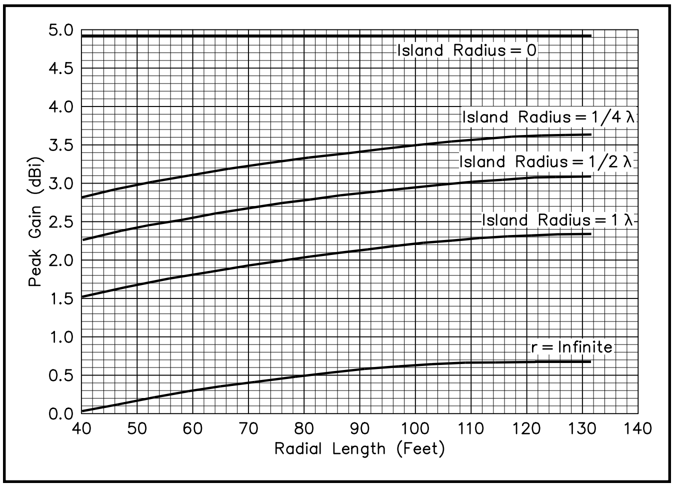

Fig 3—Peak gain for a (λ/4-high vertical with four, coil-loaded short radials on 160 meters, but this time where the antenna is located on a circular island in a saltwater ocean. Three different radial heights are shown over average ground. The radius of the circular island is equal to the length of the shortened radials in this model. Obviously, you should mount your antenna and radials over—or at least as close to—salt water, as you possibly can!

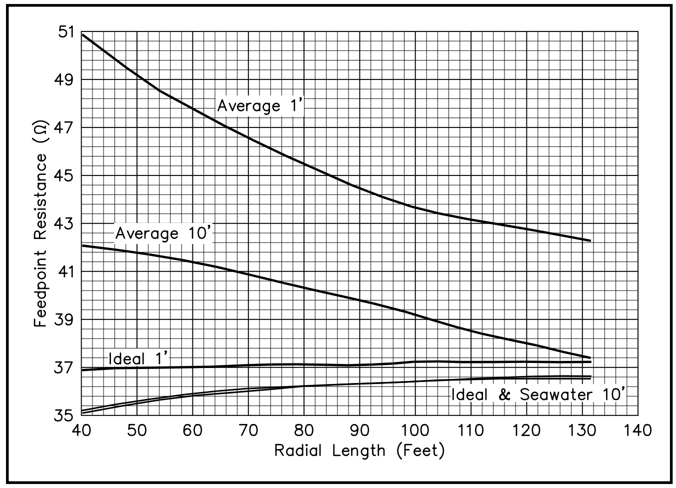

Fig 4—Feed-point resistance as a function of radial length and radial height above ground. Here, the loss resistance of the loading inductor is removed. The effect of ground loss over soil is large compared to that over salt water.

care, loss due to the loading components can be small, on the order of a few tenths of a dB. If you must place the antenna over poorer ground, such as the beach, you should try to get as close to the water as possible and keep the radius of the ground-plane radials greater than λ/8. You can, of course, use a smaller ground plane if you are willing to accept the reduced peak gain.

And here is another important consideration: Even with λ/4 radials, you should decouple the feed line from the antenna with a common-mode choke (balun). As the radials become shorter this becomes even more important, since the voltage between the base of the antenna and ground increases.

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[3]J. Belrose, VE3BLW, "Short Antennas for Mobile Operation," *QST*, Sep 1953; see also subsequent *ARRL Mobile Manuals*.

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[8]GNEC-4 Users Manual, Nittany Scientific, Airline Highway, Suite 361, Hollister, CA 95023.

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Appendix

The average gain for a lossless antenna in free space is 1.0 (or 0 dBi). For a lossy antenna the average gain will be lower by the amount of the total loss. If you model using ideal conductors and lossless loads, then the reduction in average gain directly reflects the ground loss. Inserting the correct parameters for average gain can be a little tricky, however, until you get used to it. For example, when modeling an antenna over ground, instead of averaging over the surface of a sphere, the averaging is done over a hemisphere. Because the total power is radiated into only half as much space, the gain of a lossless antenna will be 2.0 (or 3.01 dBi).

You have to keep track of these things as you go along. As a check on my "card" entries (following the terminology for the FORTRAN-based NEC-4 software) when starting a new analysis, I make the ground perfect and the conductors lossless. I thus should get an average gain very near 1.0 or 2.0, depending whether I'm modeling in free space or over ground. If all is well, then I insert the real ground constants and proceed with the modeling.

The RP card has the following format:[8,9]

RP I1 I2 I3 I4 F1 F2 F3 F4 F5 F6

where:

I1 = Selects the mode of calculation; I1 = 0 for this example

I2 = Number of values of theta at which the field is to be calculated

I3 = Number of values of phi at which the field is to be calculated

I4 = An integer consisting of 4 digits (XNDA), each of which has a different function

X = Controls antenna output format; X = 1 for this example

N = Causes normalized gain to be printed; N = 0 for this example

D = Selects either power gain or directive gain; D = 0 for this example

A = requests calculation of average gain; A = 2 for this example

F1 = Initial theta angle

F2 = Initial phi angle

F3 = Increment for theta

F4 = Increment for phi

F5 and F6 are not needed for this example.

Greek letter θ (theta) is the angle measured from zenith (directly overhead) downward. For a free-space antenna, θ will vary from 0° to 180° and for an antenna over ground, the range is 0° to 90°. Greek letter ϕ (phi) is the angle moving counter-clockwise (viewed at the antenna from the X axis towards the Y axis) rotating around the Z axis. The range of ϕ is 0° to 360°. The number of values for theta (I2) and phi (I3) will be the range selected divided by the increment (F3 or F4) plus 1. The number of values must be an integer.

The number of increments of theta and phi must be large enough to cover the entire field, unless there are known symmetries that can be exploited to reduce the number of calculation points. For example, a free-space antenna will require a sphere, and an antenna over ground will require a hemisphere. For the case of an elevated-radial, ground-plane antenna with four radials, the field will repeat every 90° of phi. It is thus only necessary to compute one quadrant of the hemisphere. The accuracy of the averaging will depend on the number of points over which the gain is averaged. Fewer points mean less accuracy but much faster computation. The way you check your setup is to calculate the average gain for a lossless system, with perfect ground, no wire loss, etc. Under those perfect conditions the average gain ideally will be 1.0 or 2.0. The difference in the actual calculation is the error due to specification of overly coarse steps in the angles. The error in dB, for antennas over ground, can be expressed as error = 10 log (average gain/2). This gives the error directly in dB. Typically, I accept an error of 0.01 dB.

I usually start with 2° increments and go up or down after checking the lossless gain. The number of field points generated with $^1/_2$° increments can be quite large and noticeably slows the computation even on a workstation. This great a resolution will seldom be needed but you should always check an ideal version of the antenna model before proceeding with a real ground and lossy antenna.

In some cases where the field does not vary greatly with either θ or ϕ you can use larger increments in one plane to reduce the computation time. For example, with a four-radial ground-plane antenna, the field variation with ϕ, at a fixed θ, is quite small and usually needs only two values for ϕ—, that is, 0° and 90°. This greatly reduces the computation time. However, for a two-radial antenna, the pattern is asymmetrical and you must use smaller increments for ϕ.

A key point is to recognize that the number of points (I2 and I3) must be adjusted to give total coverage of the desired sector (sphere, hemisphere, or quadrant) when the increments (F2 and F3) are selected. Don't forget to include one extra point for the ends.

My RP card looks like this for one quadrant and 1° increments:

RP 0 91 91 1002 0 0 1 1

With four radials and only a small error this can be reduced to:

RP 0 91 2 1002 0 0 1 90

For an antenna over ground with a pattern symmetrical about the X axis:

RP 0 91 181 1002 0 0 1 1

The average gain will appear at the end of the output file in terms of absolute gain. You can convert it to dB by using 10 log (absolute gain) or 10 log (absolute gain/2).

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THE LAZY-H VERTICAL

A versatile antenna for DX work

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

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

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

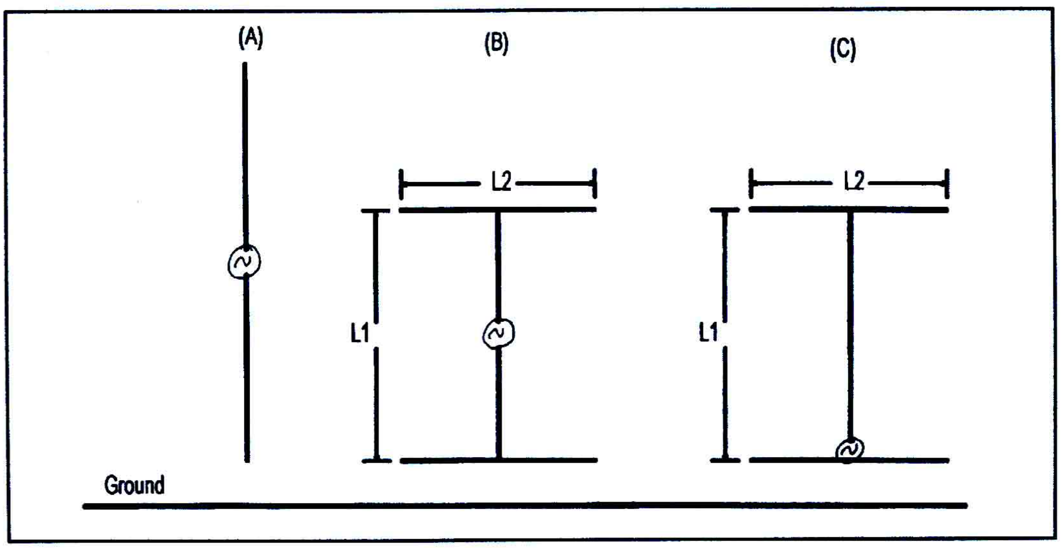

Figure 1. (A) A half-wave vertical dipole. The antenna can be shortened by adding perpendicular wires at the ends (B) and (C).

Table 1. Antenna Comparison at 3.510 MHz

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

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

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

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

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

The half-wave vertical

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

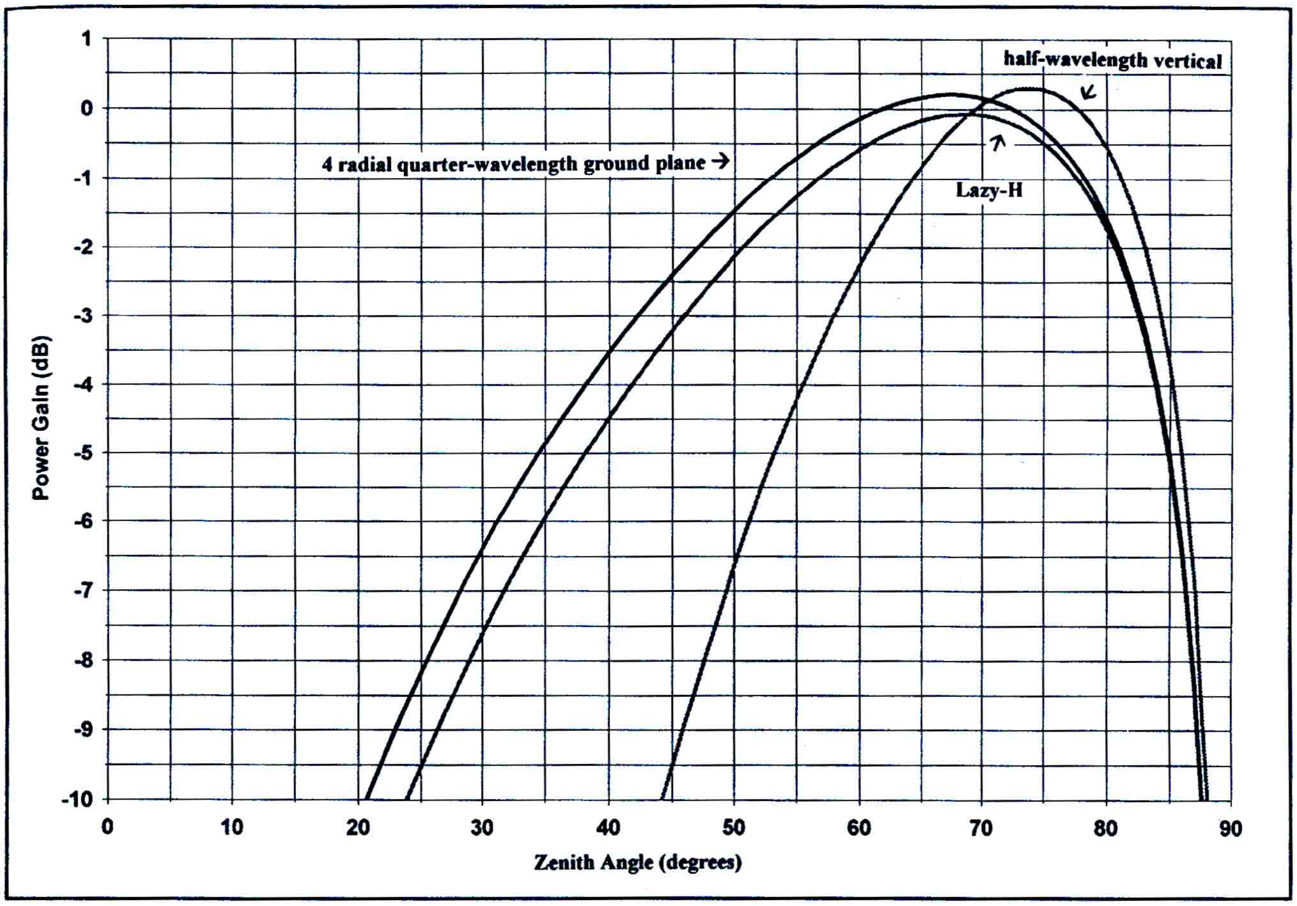

Figure 2. Elevation pattern comparison between 80-meter versions of a half-wave vertical dipole, a quarter-wave antenna with 4 elevated radials, and a lazy-H with L1 equal to a quarter wavelength.

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

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

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

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

There's some interesting information presented in this table:

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

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

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

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

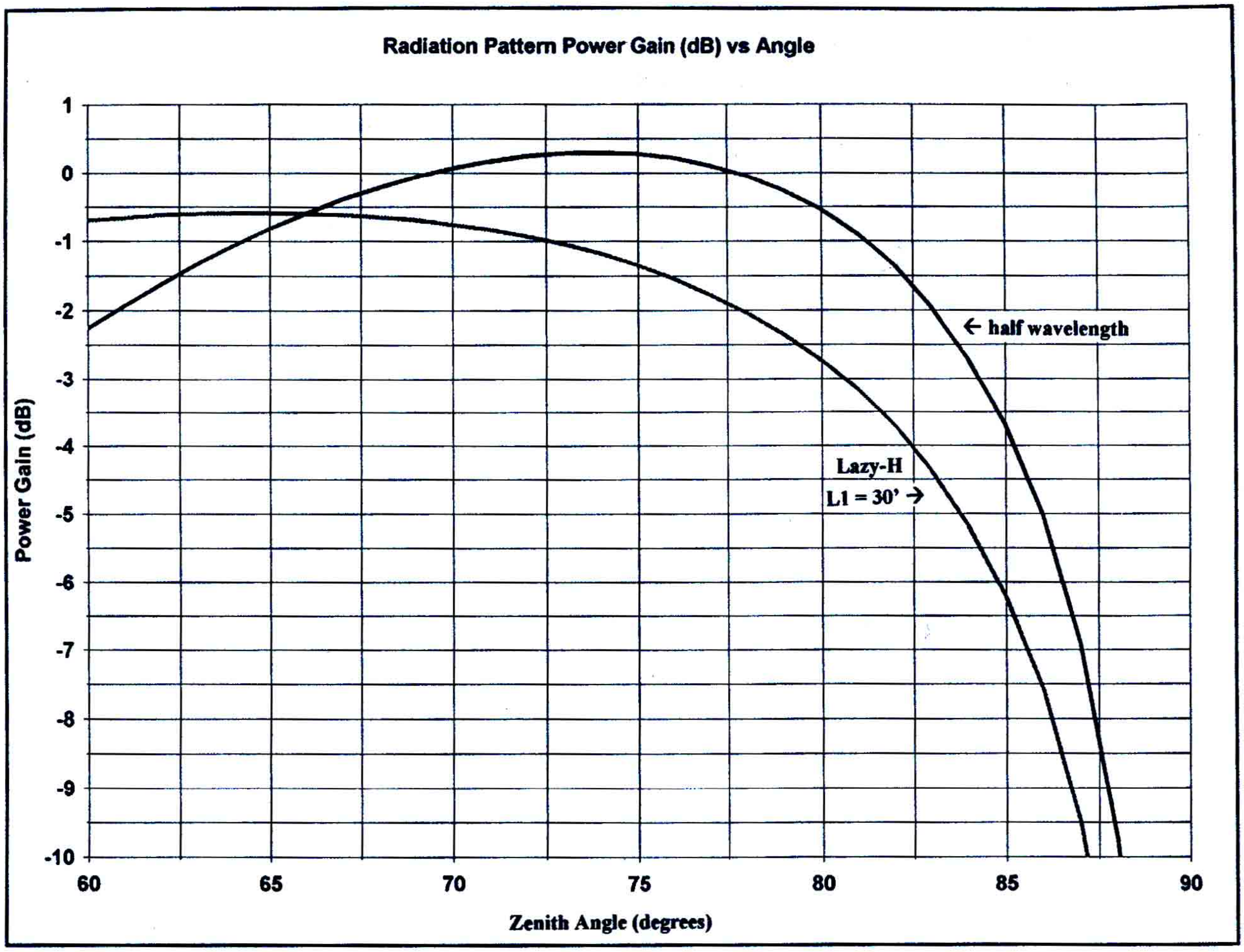

Figure 3. Gain reduction at low angles for L1 = 30 feet, compared with the half wave.

dipole for L1 > 50 feet (≈ 0.18 λ).

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

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

Asymmetric lazy-H antennas

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

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

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

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

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

Feeding the antenna

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

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

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

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

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

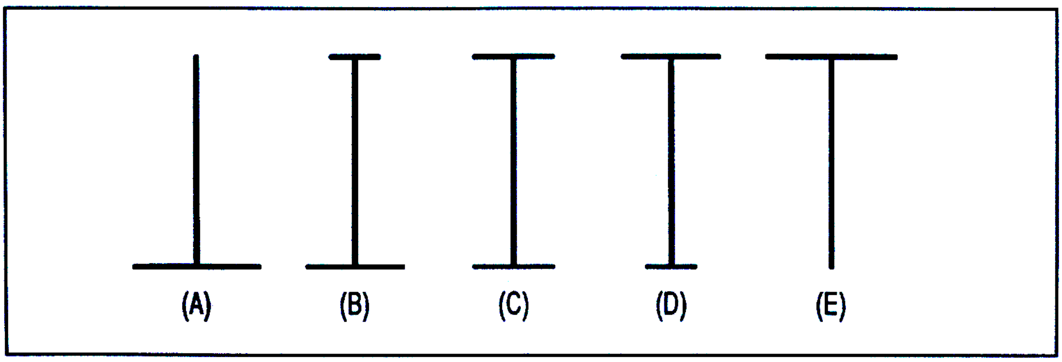

Figure 4. The aysmmetric lazy-H can have a variety of proportions with only a small difference in performance for a given length, L1.

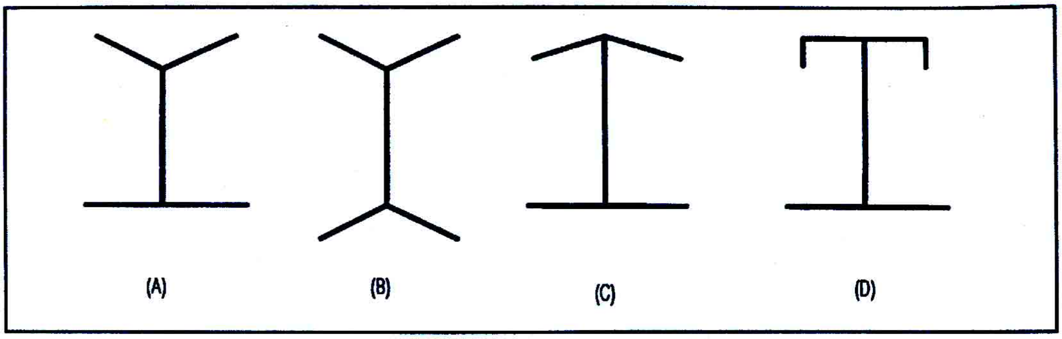

Figure 5. (A)When antenna is suspended between two supports, the top radials will have some droop towards the center. (B)The bottom radials may also droop. (C)If the antenna is suspended from a single support, the top radials may droop down. (D) It is possible to use rigid radials for a portion of the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

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

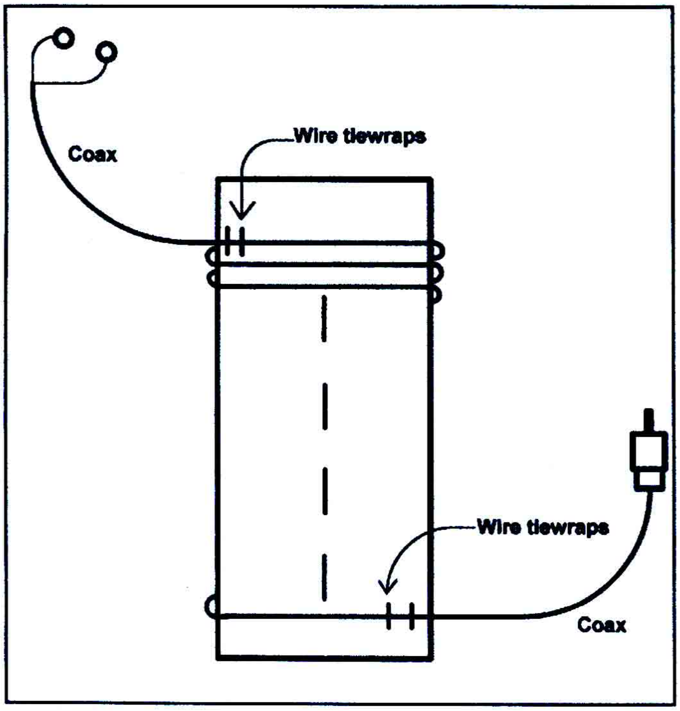

Figure 6. Coaxial choke balun.

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

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

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

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

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

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

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

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

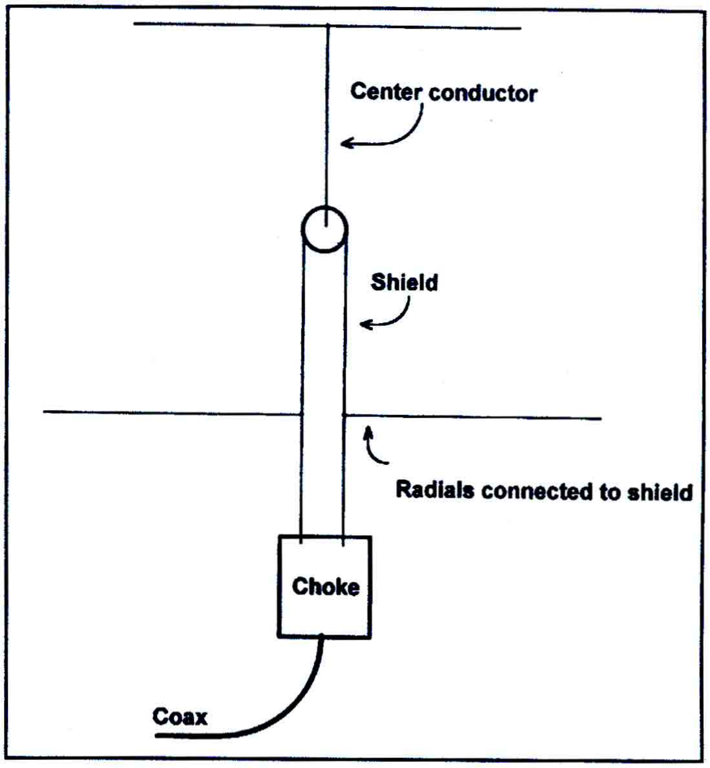

Figure 7. These antennas can be fed at any point on the vertical section (L1) using the method shown here.

A 160-meter lazy-H

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

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

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

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

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

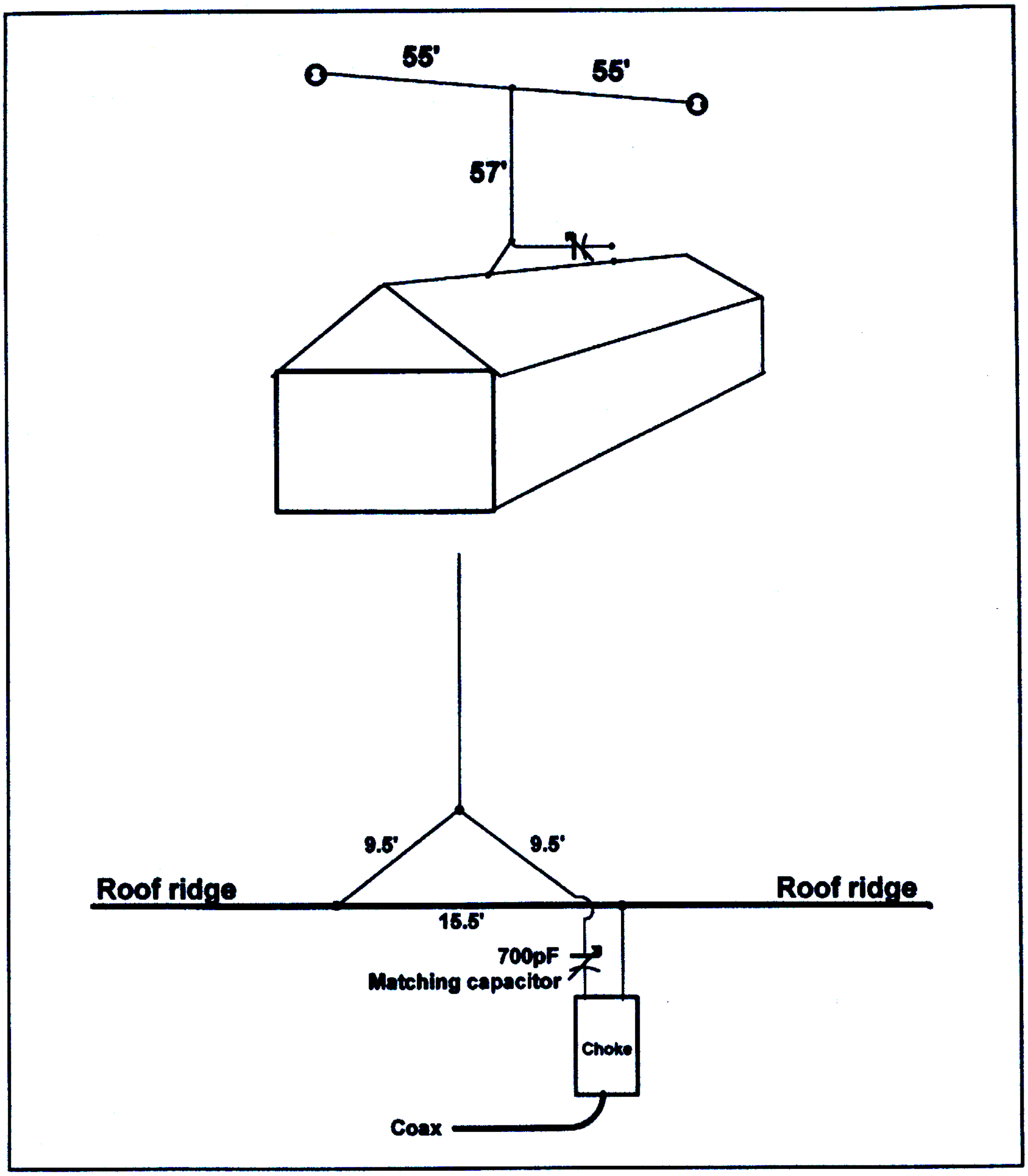

Figure 8. (A) Antenna is matched with a variation of the shunt feed. (B) Final experimental dimensions.

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

An 80-meter symmetrical lazy-H

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

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

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

An 80-meter asymmetrical lazy-H

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

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

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

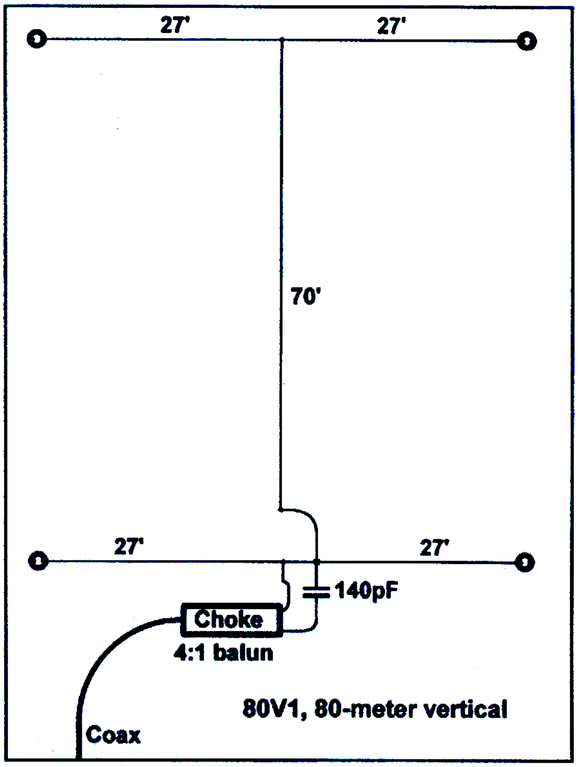

Figure 9. Eighty-meter symmetrical lazy-H.

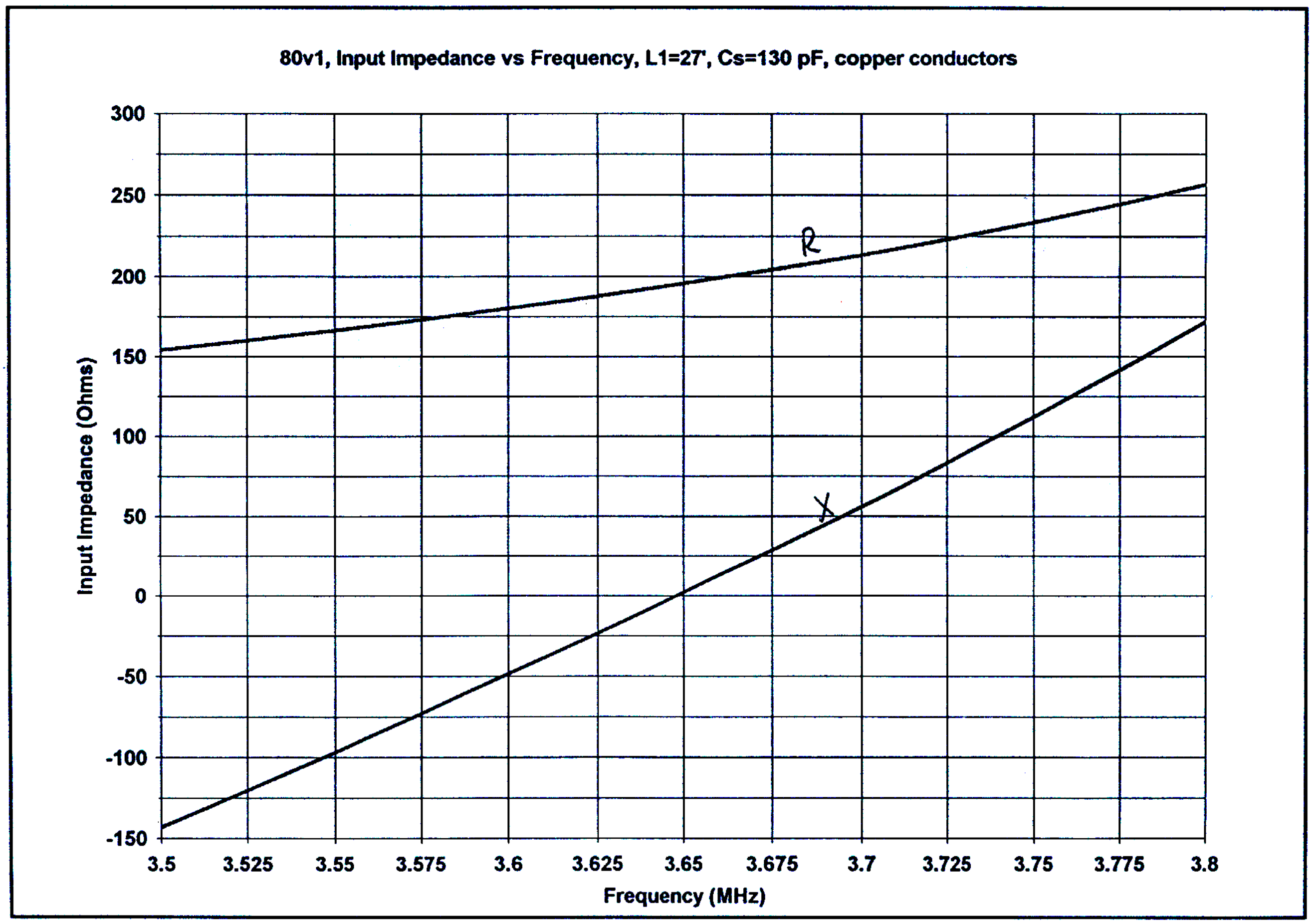

Figure 10. Feedpoint resistance and reactance from 3.5 to 3.8 MHz.

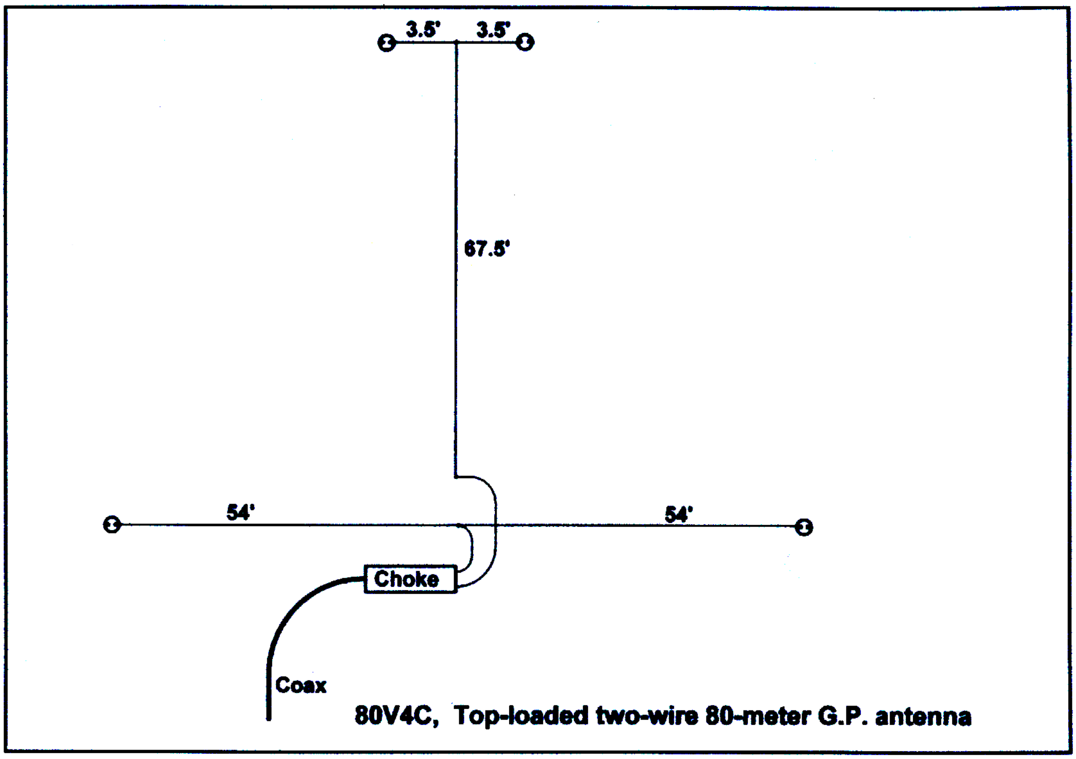

Figure 11. The final version of the 80-meter asymmetrical lazy-H.

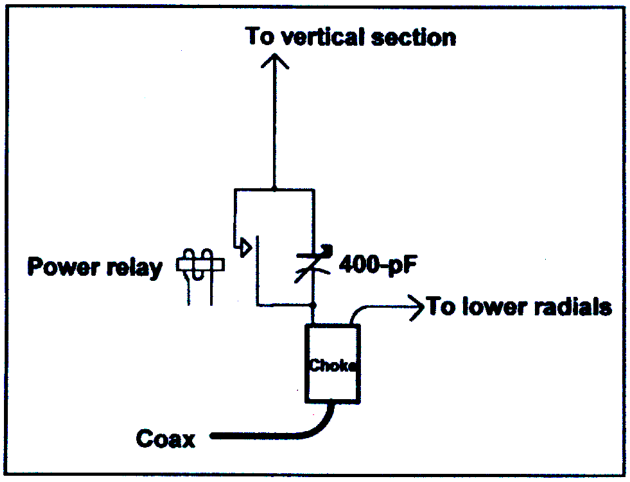

Figure 12. A 400-pF capacitor inserted in series at the feedpoint shorts out the circuit for 3.510-MHz operation, allowing operation on 3.790 MHz.

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

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

Conclusion

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

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2. Al Christman, "Elevated Vertical Antennas for the Low Bands: Varying the Height and Number Of Radials," ARRL Antenna Compendium, Volume 5, 1996, pages 11-18.
3. Smith and Johnson, "Performance of Short Antennas," Proceeding of the IRE, October 1947, pages 1026-1038.
4. R. Littlefield, "Tech Notes: The 2-Meter Discpole Antenna," *Communications Quarterly*, Summer 1996, pages 77-81.
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Single Support Gain Antennas for 80 and 160 Meters

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Introduction

On 80 and 160 meters an antenna with modest gain and good front-to-back (F/B) ratio, along with a steerable pattern, can be very effective for contesting. This sounds like your standard HF Yagi, but unfortunately, for most of us at least, full-size rotary Yagis, at the necessary heights (greater than ½ wavelength), are not an option on the low bands. However, many of us do have a single tall support, usually a tower, from which it may be possible to suspend a vertical array.

The family of vertical arrays made with sloping elements from a single central support has many variations. A typical example is the K1WA[1] array shown in Figure 1. In this array each of the elements is a center-fed sloping dipole. One element is driven at a time with the other elements acting as reflectors. The length of the coax from the switch-box on the tower to the center of an element is adjusted so that when open circuited by the relays in the box, that section of feed line provides inductive loading that tunes each element as a reflector. In this example there is one driven element and four reflectors. Multiple reflectors really don't behave very differently from a single reflector (a little better F/B maybe) so the array is basically just a two element Yagi where the pattern is rotated by changing the element driven.

This theme has many variations: 2 or 3-element parasitic or phased array with vertical elements, straight sloping elements, or elements bent back towards the support. The element lengths may be anywhere from one-eighth wavelength to one-half wavelength, with or without loading as required, and center or end fed.

This article shows a number of typical examples to give you some ideas for your own installation. Details of each of the examples can be found in the references at the end of the article. In particular, John Devoldere's *Low Band DXing*[2] is a goldmine of ideas.

Expectations

Before going into the examples, I would like to indicate what performance can reasonably be expected. Even though there are many, many possibilities, in the end the performance will be quite similar between arrays using the same number of active elements. Most of these antennas will take the form of either a 2-element or 3-element array. Many of examples have 3, 4 or even 5 elements but usually only one element is driven (as shown in Figure 1) and the others are either inactive or act collectively as a reflector—ie, basically a 2-element array. In some examples one element will act as a director and another as a reflector—ie, a 3-element array.

Figure 2 is an excerpt from *The ARRL Antenna Book* 2-element phased array pattern diagram.[3] What is shown is the gain over a single element for a 2-element array with various phasings and spacings. The elements are assumed lossless and the ground perfect. The current amplitude is assumed to be equal in both elements and the height of each element is one-quarter wavelength. Note the tradeoff between gain and F/B—you can't maximize both at once! W4RNL has given an extensive discussion of the possibilities and limitations of 2-element arrays in earlier *NCJ* articles[4] and these are recommended reading.

The greatest gain difference (4.5 dB) is for a spacing of one-quarter wavelength and a phase difference of 135 degrees. The gains shown in Figure 2 are of course idealized. In the real world you won't get quite as much due to conductor losses, which can be substantial in the long wire conductors used on the low bands, and imperfections in tuning, spacing, element shapes, ground system, etc. For example, when you go from perfect ground to average ground (conductivity of 0.005 S/m and dielectric constant of 13), the gain difference for

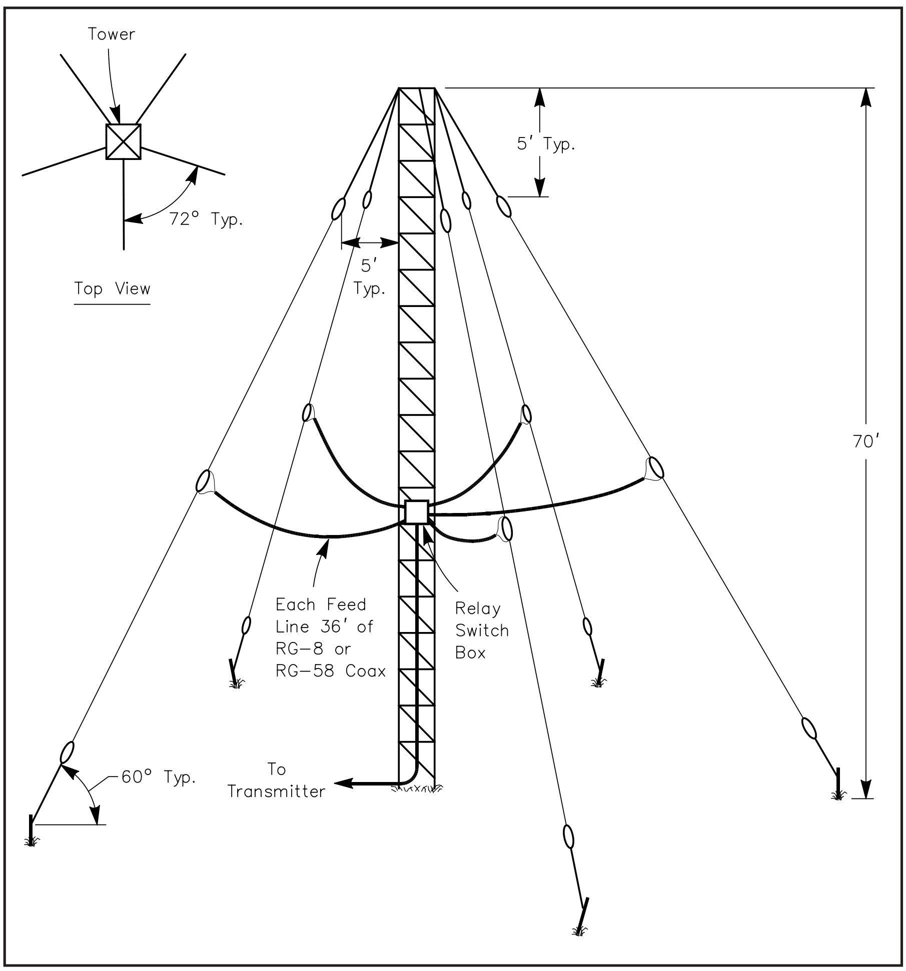


Figure 1—The K1WA Sloper System uses five identical one-half wavelength sloping dipoles spaced uniformly around a tall mast. Each feeder has an electrical length of about 135 degrees.

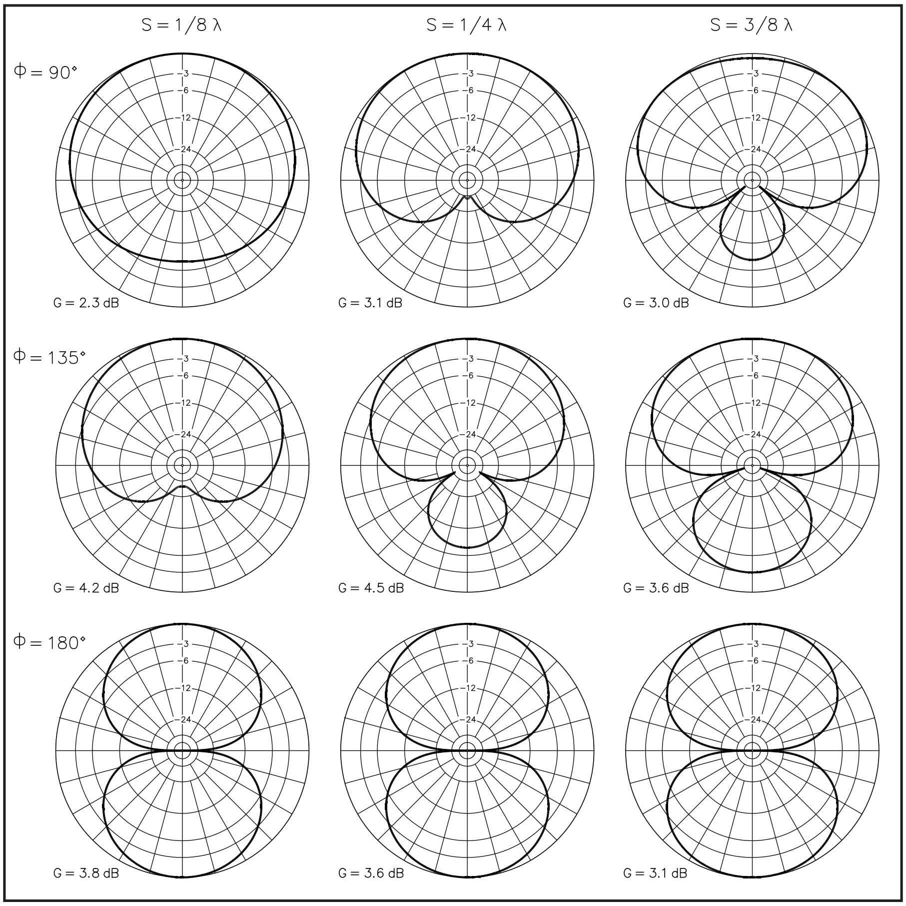


Figure 2—Idealized gain of a 2-element vertical phased array over a single vertical.

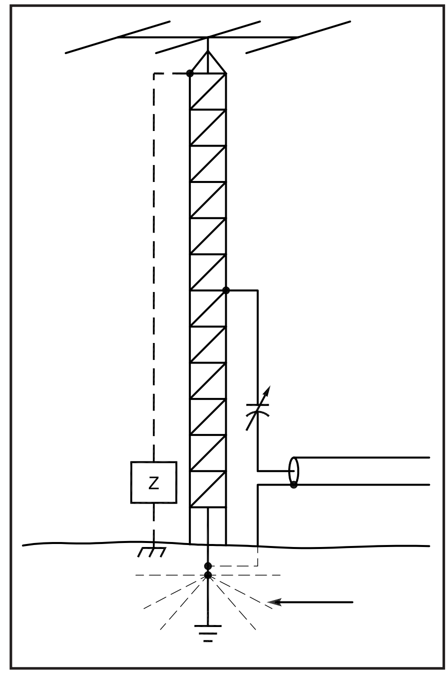


Figure 3—Tower with an HF array and shunt matching and tuning arrangements.

lossless elements drops to 4.3 dB. Adding in a couple of ohms of loss and the gain difference drops another 0.2 dB.

The examples in Figure 2 are for both elements driven. However, driving one element and allowing the other element to be parasitically excited (a Yagi!) is just another way to approximate the correct current amplitude and phasing. In the case of parasitic elements you can't control the phase and amplitude as closely as when both elements are driven independently so again the achievable gains and F/B will be somewhat lower. In exchange, the arrangements for pattern rotation may be considerably simpler in the parasitic array.

For a 2-element parasitic array, a gain of about 3-4 dB over a single vertical would be typical, with a F/B of 10-12 dB for a reflector array. In a 3-element array good F/B (greater than 20 dB) and an additional 1-2 dB of gain are possible. In three element arrays, the element impedance can be low however, especially if short, loaded elements are used. That's okay on 20 meters where the element is made from aluminum tubing, but on 160 meters where the element may be #12 wire and about eight times longer, the losses can be substantial. Wire loss is a basic limiting factor in large wire end-fire arrays. It is perfectly possible to build a 3-element array that has less gain than a 2-element array due to losses. Low impedances also mean that ground loss must be carefully controlled. Care in design and implementation is essential.

General Comments

For most, the available support will be a tower of some height, with probably one or more higher frequency Yagis at the top. Every installation will be different due to different tower heights, top loading due to the HF array, etc. Also, the available space around the tower into which one can stretch sloping elements and support lines will differ. For this reason, each installation becomes a unique design. It is essential to carefully plan and model each installation and then properly adjust it to get the predicted performance. What works great at my place may not be worth much at yours!

Some examples use grounded (or driven against ground) elements in the range of one-quarter wavelength to three-eighths wavelength. It is clear from well-known vertical antenna practice that a good ground system is required to minimize local ground losses. There is the misconception that free floating one-half wavelength elements do not need a ground system. While it is true that these antennas will work relatively well without an extensive ground system, they will work even better with one. The problem is the high electric fields near the ends of the elements that may be close to ground. This leads to losses that can be reduced by the use of a ground screen under the elements.

A key decision is whether to use the tower as an element in the array or just let it be neutral and provide mechanical support only. If you want to excite the tower as part of the array you will usually leave the tower grounded (with a good ground system!) because of the cabling going to the HF antennas, rotors, etc. You can match to the tower using shunt feed as shown in Figure 3. It is not necessary that the tower be resonant but if it is far from resonance then tuning it and getting a proper match may be a bit challenging. The usual means for checking tower resonance and tuning or detuning it is to add a shunt wire from near the top of the tower as indicated by the dashed line in Figure 3. If the tower needs to be tuned or detuned, then an impedance can be inserted in this wire as indicated. It is possible to perform both tuning and matching with the shunt wire. When only a single shunt wire is used, rotating the HF array to different positions may alter the tuning somewhat. Using three wires, symmetrically disposed around the tower, will pretty much suppress this and also provide some additional matching opportunities.

Element shape

Different element shapes can be used in these arrays: vertical elements, a sloping element, like K1WA, or a bent element, like K3LR[5] and K8UR[6] as shown in Figure 4. The sloping element will have both vertical and horizontal current components, in proportion to the slope of the element, which contribute both vertical and horizontally polarized radiation. In an array you will find that the horizontal component is essentially that of a low dipole with lots of high angle radiation. Also it will be noticed that as the element phasing is varied, the total pattern (sum of vertical and horizontal components) does not behave the same as the phased purely vertical elements assumed in Figure 2. The result is an absence of a zenith null in the pattern and some reduction in maximum gain in endfire and broadside patterns. This effect can be seen by comparing the elevation patterns of the K1WA array to an early K3LR[5] array as shown in Figure 5. More discussion on this point can be found in my article in *the ARRL Antenna Compendium Volume 7*.[7]

This observation is not meant to imply that sloping elements should not be used. I know of several amateurs using essentially the K1WA array on 160 meters with very good results. It just may be that mixed polarization is a good thing. The point is to recognize that the two different element shapes will produce different radiation patterns and polarization mix. Some people may want the additional high angle. It doesn't cost much in forward gain and it provides a big signal at short distances. In my case, I choose to suppress the high angle radiation in the array and use a dipole for local and up and down the West Coast.

A bent element can be proportioned, as shown by G3LNP,[8] to null out most of the horizontal component and act much more like a straight vertical element.

Examples of Sloper Arrays

When a guyed tower is driven, one of the simplest ways to add parasitic elements is to convert the guys into elements using strategically placed insulators as shown in Figure 6.[9] The tower is the driven element and the guys act as reflectors. Relays can be placed at the base of each guy (as indicated in the insert in Figure 6) to connect one guy at a time to act as a reflector and rotate the pattern. When the relay contacts are open the guy is non-resonant and transparent to the array. It is also quite possible to cut the active part of the guy to act as a director and then add a enough inductive loading so that it acts as a reflector. To enable a particular guy to act as a director, the relay can simply short out the loading inductor. When the relay is open the guy is a reflector.

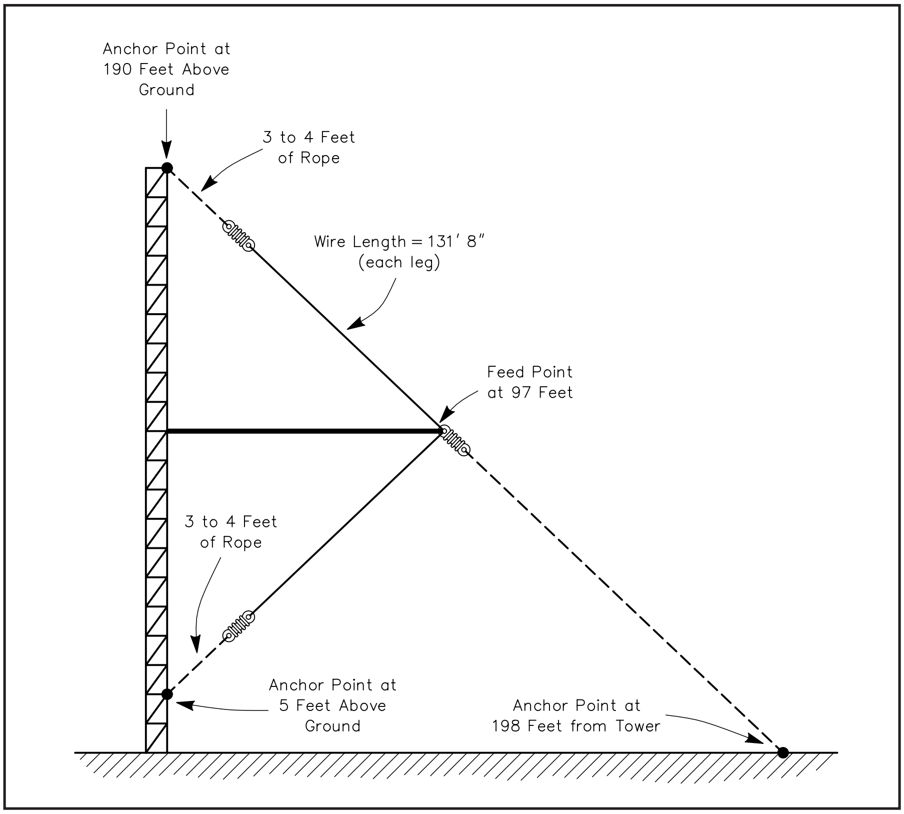


Figure 4—K3LR/K8UR bent element example.

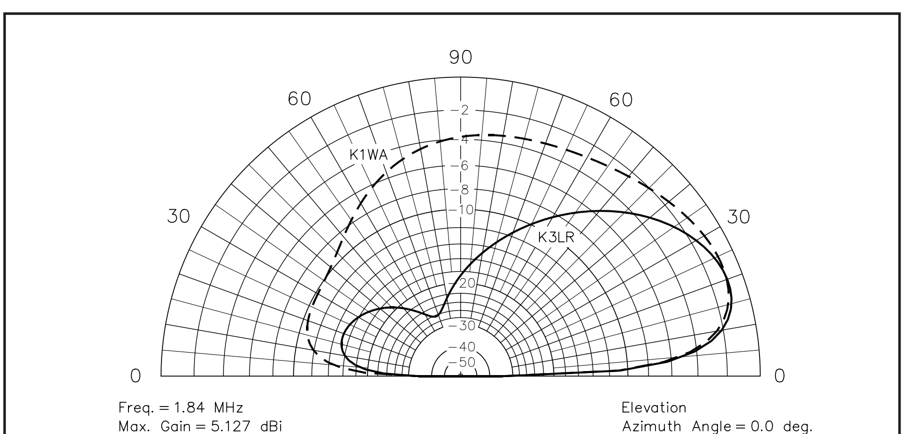


Figure 5—Elevation pattern comparison between straight sloping elements (K1WA) and bent elements (K3LR).

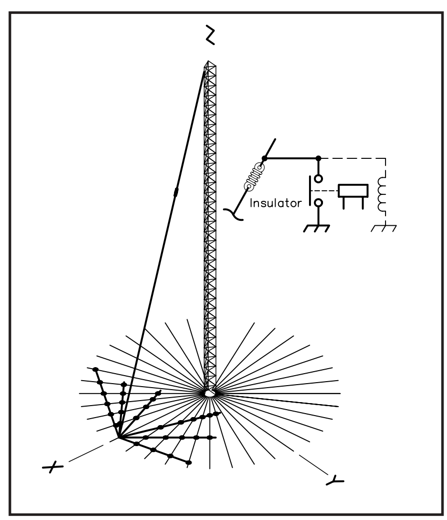
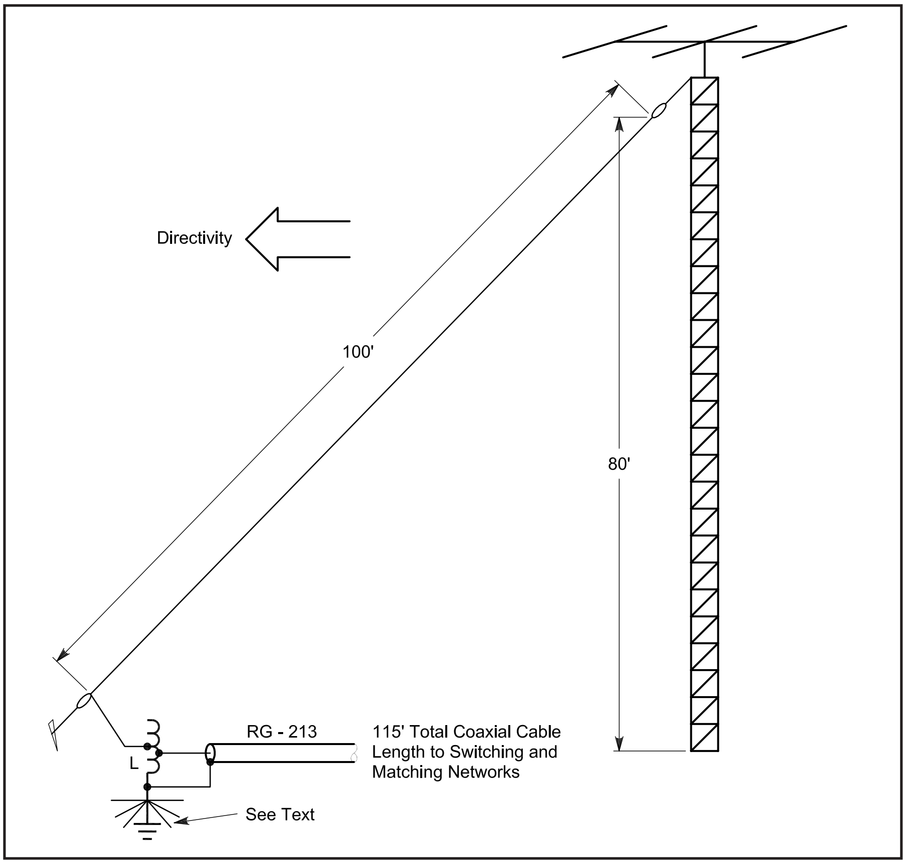


Figure 6—A simple 2-element slant-wire parasitic array.

Figure 7—Single sloping element per 4X4NJ.



4X4NJ has described several arrays[10] for 160 meters, one of which is shown in Figure 7. It is a two element parasitic array and the tower is tuned to act as a reflector. This idea can be extended to multiple elements, spaced around the tower as indicated in Figure 8. Each of the elements is about 100 feet long (0.19 wavelengths at 1.850 MHz) and resonated with a loading coil at the base which also provides a matching opportunity. Because the length of the elements is nearly one-quarter wavelength, the loading coil will be small and not greatly affect efficiency. There are many possible ways to drive the array elements. The tower can be detuned and one element driven with the other elements acting as reflectors like K1WA, or the tower may be driven and the elements tuned as reflectors and directors to form a 3-element parasitic array.

As a phased array, the element phasing can be adjusted to provide several different patterns. However, as indicated by the dashed line in Figure 8, a bent element, with cancellation of the horizontal component, would give better pattern flexibility. Depending on the phasing, this array can be bidirectional endfire, bidirectional broadside, or unidirectional endfire. The unidirectional endfire mode can be adjusted for either maximum gain or F/B. It should be pointed out that because of the relatively close spacing of the elements in most single support sloper arrays, broadside gain is usually modest at around 1-2 dB. Endfire gain can of course be very good if conductor and ground losses can be minimized.

A variation with one-quarter wavelength sloping elements and a driven tower appears in ON4UN's book.[11] Figure 9 shows the array where the elements are made slightly shorter than one-quarter wavelength to act as directors and then converted to reflectors by inserting a small inductance. Note! Each element must be tied into the overall ground system. Also remember that even if you don't use the guys as elements, they must be detuned so that they do not interact with the desired elements. Normally this would be done by breaking up the guys with insulators, or by using non-conductive guys.

The Spitfire array[12] (W1FV/K1VR) shown in Figure 10 is a variation using a driven tower (approximately one-quarter wavelength) and one-half wavelength ungrounded elements for reflector/directors. The changeover from director to reflector is done by connect-

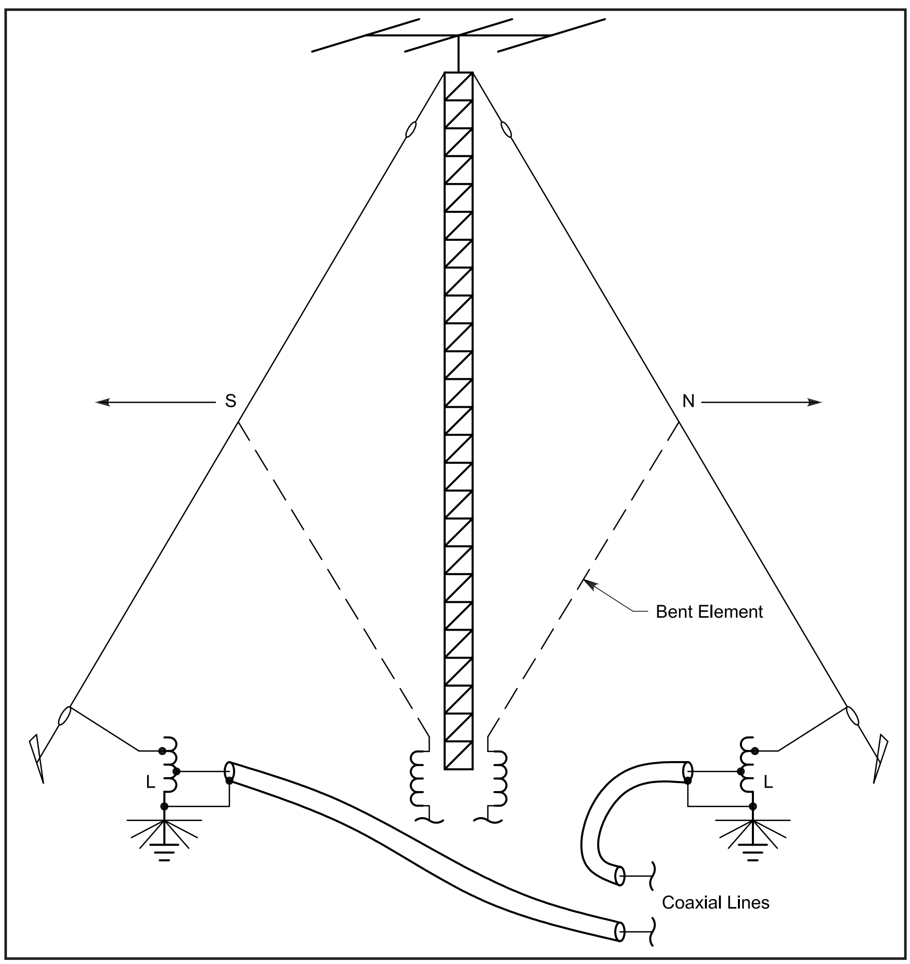


Figure 8—4X4NJ 2-element array.

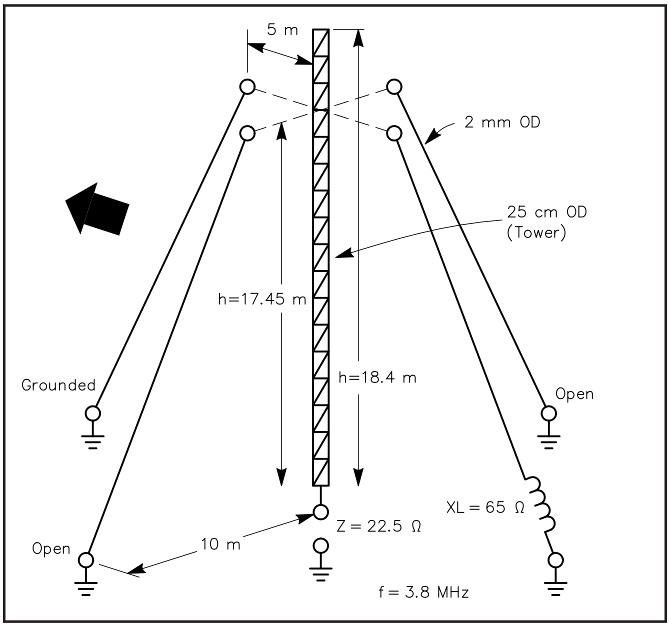


Figure 9—A three element array.

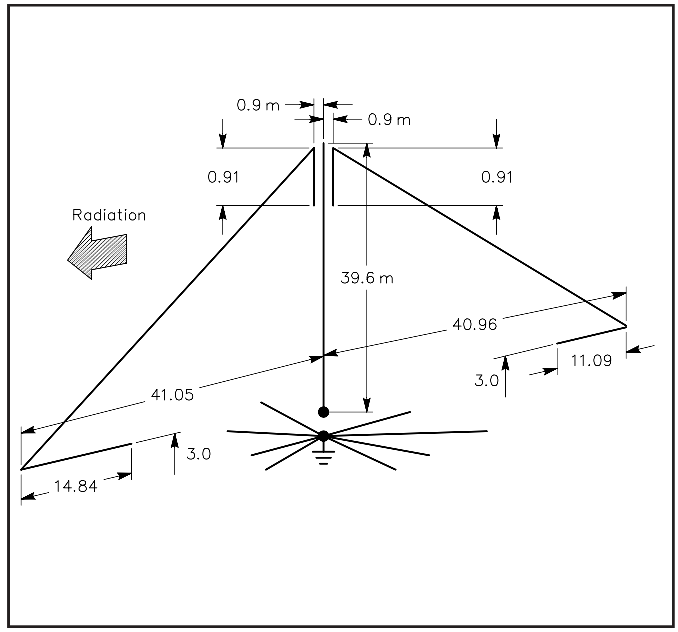


Figure 10—The Spitfire array.

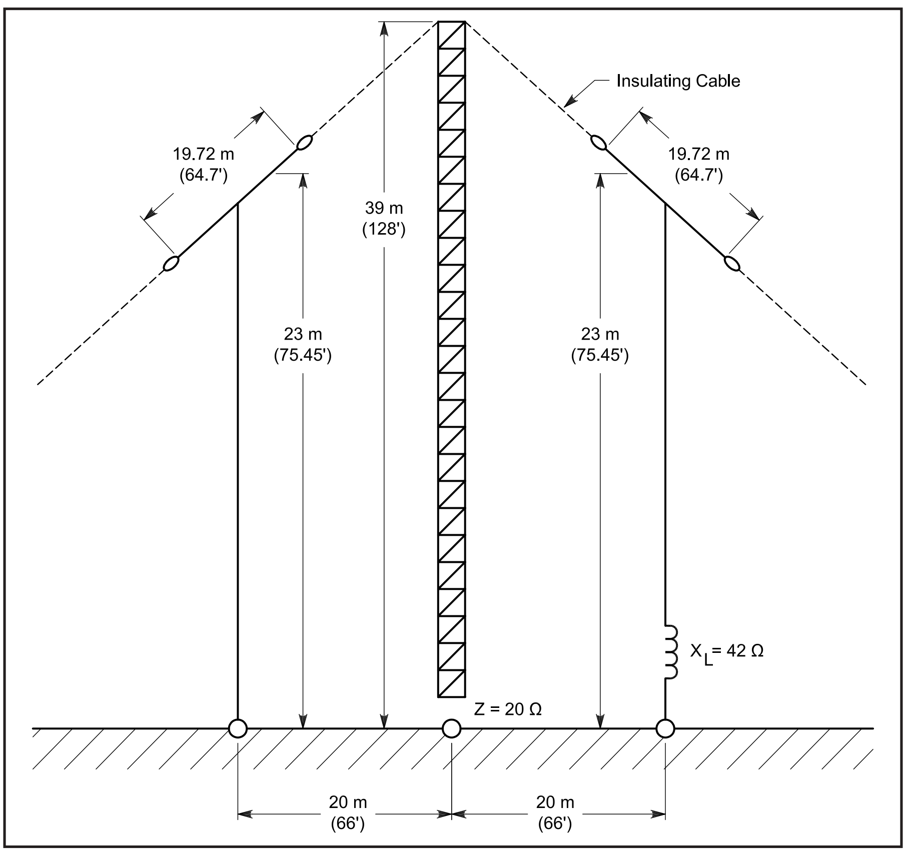


Figure 11—A three element array with vertical top-loaded elements.

ing an additional length of wire to the bottom ends as indicated. One disadvantage is that the relays in this case must be rated for 5 kV or more. A vacuum relay would be typical. Also you have to be careful to decouple the relay coil drive lines from the HV RF on the contacts. Even though the parasitic elements are not directly grounded, it is still important to have a ground screen under the elements due to the very high fields present near the element bottom ends and of course the driven element requires a ground system.

Purely vertical elements can also be used by suspending them from the guy wires and allowing a portion of the guy to act as top loading, as shown in Figure 11.[11]

Comments on Tuning and Adjustment

I have built a number of arrays of this type for 160 meters. I usually begin by carefully modeling the proposed array using *EZNEC* or similar software, being very careful to include conductor losses. Once I think I have a winner I then go out and erect the array. But before adjusting it, I go back to the modeling and model one element at a time, keeping the element lengths the same as the full model, with the other elements either open or absent. What I am looking for is the self-resonant frequency of each element, with the other elements not present, using the dimensions from the complete array model. I then go the actual array and repeat the exercise, only this time adjusting each element to be resonant at the same frequency as the modeling gives for each element in the absence of the other elements. During this part of the tuning process, the other elements are either lowered to the ground or open-circuited so they do not affect the element being adjusted. Resonance can be determined with a dip meter (monitored with a receiver for calibration!). I do this for each element in turn.

Final adjustments for matching should be done with the entire array up. You can also touch up the F/B by placing a source several wavelengths away to minimize the received signal. One word of caution is in order, however. The received signal will be ground wave at a very low angle, minimizing what is not necessarily the same as maximizing F/B at the higher angles more typical of backward lobes.

One point often overlooked in large wire arrays is the effect of insulation on the resonant length of an element. Standard electrical wire insulation can shift the resonance downward 3-4 percent, seriously mistuning a parasitic element. This shows up during tuning by the need to shorten an element by several feet to obtain the desired self-resonant frequency when insulated wire is used in the field but un-insulated wire in the model. This can be a bit disconcerting if you don't expect it.

Conclusion

I think the forgoing discussion makes very clear the wide range of possibilities for creating a directive array with a switchable pattern when a single support of modest height is available. These arrays can be made from simple components: wire, insulators, and sections of transmission line. For the most part they are quite economical. But for all that they can still be very effective and are possible to implement even in modest size lots. Hopefully this discussion has shown you just how flexible the arrangements are and there is probably a solution for almost any situation.

The discussion is just an overview. If you want to build one of the antennas then you should read carefully the references that are full of practical details. In general each installation will be unique and require a design developed for that situation.

Acknowledgment

I would like to thank Mark Perrin, N7MQ, and George Cutsogeorge, W2VJN, for reading and commenting on the draft article. They helped to reduce the confusion quotient drastically.

Notes

[1]*The ARRL Antenna Book*, 19th edition, 2000, p 6-32. This antenna is also in earlier editions.

[2]John Devoldere, ON4UN, *Low Band DXing*, ARRL, 3rd edition. See examples in chapters 11 and 13.

[3]ARRL Antenna Book, 19th edition, 2000, page 8-8

[4]LB Cebik, W4RNL, "Some Notes on Two-Element Horizontal Phased Arrays," in four parts, *NCJ*, Nov/Dec 2001, pp 4-10; Jan/Feb 2002, pp 4-9; Mar/Apr 2002, pp 3-8; and May/Jun 2002, pp 3-8.

[5]Christman, Duffy, and Breakall, The 160-Meter Sloper System at K3LR, *QST* Aug 1994, pp 36-38. See also *The ARRL Antenna Compendium Volume 4*, pp 9-17

[6]D. C. Mitchell, K8UR, The K8UR Low-Band Vertical Array," *CQ*, Dec 1989, pp 42-45.

[7]R. Severns, N6LF, "Getting the Most from Half-Wave Sloper Arrays," *ARRL Antenna Compendium Volume 7*, Fall 2002.

[8]Tony Preedy, G3LNP, "Single Support Directional Wires," *RADCOM*, Aug and Sep 1997, pp 38-39 and 76-78.

[9]John Stanley, K4ERO, "The Tuned Guy Wire—Gain for (Almost) Free," *ARRL Antenna Compendium Volume 4*, pp 27-29.

[10]Riki Kline, 4X4NJ, "Build a 4X Array For 160 Meters," *QST*, Feb 1985, pp 21-23 and 45. Reprinted in *ARRL's More Wire Antenna Classics II*, 1999, pp 6-21 through 6-24.

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[12]John Devoldere, ON4UN, *Low Band DXing*, ARRL, 3rd edition, p 13-50.

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Erwin David, G4LQI, *HF Antenna Collection*, RSGB, 1989, 82-90.

Les Moxon, G6XN, *HF Antennas for all Locations*, RSGB, 2nd edition 1993, pp 227-230.

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NCJ

Rudy Severns, N6LF; n6lf@arrl.net

A 3-Element 160 Meter Vertical Array

In May/June 2008 *NCJ*[1] Al Christman, K3LC, suggested adding a fifth element at the center of a standard 4-square array to improve the pattern. In his scheme the 4-square was transformed into two reversible 3-element arrays at right angles to each other. In the 3-element array, one element was parasitic and two were driven. The remaining two elements were inactive.

After reading the article I realized I had already used a very similar 3-element array on 160 meters, and I can say that it works as advertised. The pattern is very good indeed. My array had one important difference, however. Only the center element was driven; all other elements were parasitic. With this configuration pattern nulls were not quite as deep as the K3LC version, but they were very close. In exchange I had an antenna that eliminated *all* of the transmission lines and phasing networks associated with a driven array. These components were replaced with a small tapped inductor and a SPDT relay at the base of each parasitic element and a simple tapped inductor at the base of the driven element.

Incorporating this idea into a 4-square and eliminating the transmission lines and phasing networks should reduce losses substantially and save a lot of time, not to mention the expense of phasing networks and transmission lines. The new arrangement is also much easier to adjust for peak performance than a standard 4-square or even K3LC's version. There is, of course, the disadvantage of an additional element in the case of a modified 4-square, but if you already have four elements in place, the center element does not have to be self-supporting. It could be a wire suspended from the other four elements. A number of similar arrays use a tower as the driven element, with the parasitic elements suspended from the tower.[3, 4]

What follows describes my 160 meter array "as built." The approach used is very flexible, and there are many different ways it could be implemented to suit a particular situation. You could use the 3-element array just as I did or modify a 4-square.

The N6LF 160 Meter Array

Figure 1 is a sketch of the 160 meter 3-element array. Each element is 80 feet of 4-inch aluminum irrigation pipe, top loaded with two 40-foot lengths of #12 wire sloping downward at about 45°. The length of each loading wire was adjusted so the elements — *without* base loading — were individually self-resonant at about 2.0 MHz. This made it possible to adjust the final resonant frequencies by adding a small tapped inductor (about 5 µH) in series with the base of each element.

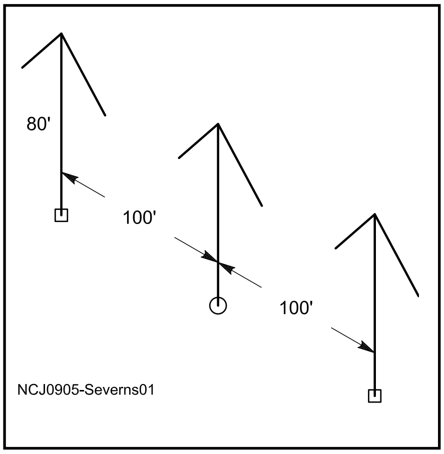

Figure 1 — A *NEC* view of the N6LF 3-element 160 meter array

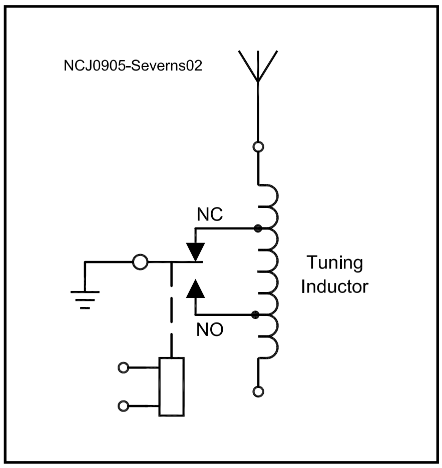

Figure 2 — Tapped base inductor with a relay to change from reflector to director operation

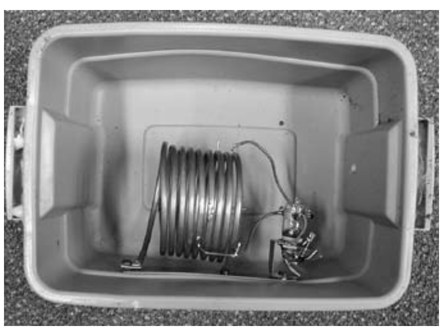

Figure 3 — Relay and base inductor example

For director operation a tap point was selected that made the element resonant at 1.95 MHz; for reflector operation a tap was selected for resonance at 1.8 MHz. The self-resonant frequencies of the parasitic elements were adjusted with the other two elements *open circuited*. After both parasitic elements were tuned, one was set to be a reflector and the other a director. At this point the driven element was tuned to resonance (1.83 MHz in my case) and matched to the feed line by varying the tap on the base coil.

The change from director to reflector was done using a SPST relay and two taps on the base inductors for the parasitic elements (see Figure 2). The inductors were made a bit larger than the minimum required size so the two values of inductance required could be reached simply by moving the taps. The bottom end of the coil is not connected to anything. This was just a matter of convenience during adjustment.

To match the feed line to the driven element I also used a small inductor with two taps, but no relay was needed. One tap was connected directly to ground and used to resonate. The other tap was adjusted for minimum SWR on the input feed line — very close to 1:1 at 1.830 MHz. Figure 3 shows an example of a base inductor and relay; I used a vacuum there only because it was handy. A simple open contact relay would be fine for this application, as long as the contacts can carry the current. One small trick was to invert the NO and NC contacts between the two parasitic elements so that with no power to the relay, the antenna would fire east. With power applied it would fire west. In my case I powered the relay via the coaxial feed line, using RF chokes and capacitors to isolate the RF from the dc voltage for the relay.

Radiation Patterns

Figures 4 and 5 depict the radiation patterns derived from *NEC*. The predicted parasitic element currents assume 1 A at

[1]Notes appear on page 15.

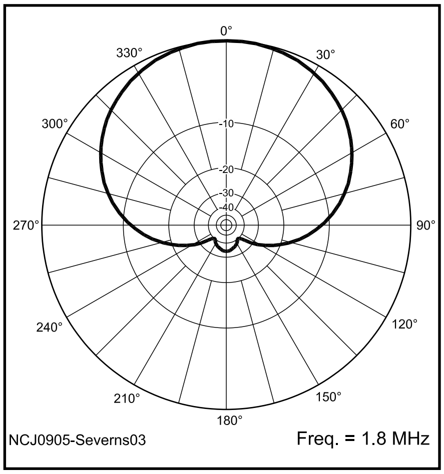

Figure 4 — N6LF array azimuth pattern at 20° elevation

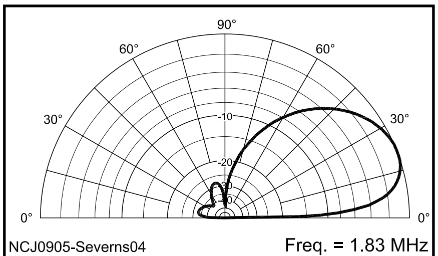

Figure 5 — N6LF array elevation pattern at 0° **azimuth**

0° degrees in the driven element. Table 1 offers a comparison to those for the K3LC array. Both the patterns and the currents in the N6LF array are close to those specified by K3LC.

Joe Johnson, K3RR, has suggested another way to tune the elements.[4] Instead of reducing the top loading so the elements are self-resonant above the desired frequency and using an inductor to resonate, he suggests using a little extra top loading so that the elements are self-resonant below the desired frequency. He then resonates the elements again using series capacitors. This might prove more efficient and, if part of the capacitance is variable, make adjustment very easy. A relay then would be used to short out a portion of the capacitance to switch from director to reflector.

Building Your Own Version

If you're not replicating the antenna as I've described it, then you'll have to determine in advance the proper element height and top loading using modeling software such as *EZNEC*.[2] When you do this you will find that the achievable current amplitudes and phases and the resulting pattern will depend on the height and loading of the verticals, as well as on element spacing.

When all the elements in an array are driven you can have any combination of phase and amplitude for the element currents — at least in principle. When some of the elements are parasitic, however, there are built-in limitations to the achievable element current phases and amplitudes. For example, I found I could much more closely approximate the K3LC element currents with 80-foot top-loaded elements than I could with full quarter-wave (130-foot) elements. This is not to say that the taller elements wouldn't work, but operating as parasitics I could not get as good a pattern because I could not achieve the desired current phases and amplitudes as closely. This was fine from my point of view, since I would much rather put up 80-foot elements than 130-foot elements. The final efficiency was still quite high even with the shorter elements, although I had to be very aggressive with my ground system.

It's been suggested that element length doesn't matter, and all you need to do is tune a parasitic element to resonate at the desired frequency.[5] That is not the case, however. Obtaining a better pattern in a parasitic array by using shorter loaded elements is nothing new. For example, a 2-element Moxon style Yagi — where the ends of the elements are bent toward each other — can have a substantially better front-to-back (F/B) than the same antenna with full-sized elements. In a recent talk, Tom Schiller, N6BT — who has a great deal of experience with parasitic arrays — discussed the utility of using shorter, loaded elements in a parasitic array.[6]

Table 1
Element current amplitude and phasing in the N6LF and K3LC arrays

	N6LF	**K3LC**
Director	0.43 A @ –128°	0.5 A @ –130°
Reflector	0.59 A @ +127°	0.5 A @ +130°

Conclusions

Overall, Al's idea to improve the pattern of a 4-square by converting it into two 3-element arrays at right angles works just fine. There are many variations on this theme to fit different situations.

Notes

[1]A. Christman, K3LC, "Modifying the 4-Square Array for Improved Front-to-Back Ratio," *NCJ*, May/June 2008, pp 5-6
[2]R. Lewallen, **www.eznec.com**
[3]J. Devoldere, ON4UN, *Low-Band DXing*, (4th ed), ARRL, pp 13-30 through 13-35
[4]J. Johnson, K3RR, private communication
[5]J. Devoldere, ON4UN, *Low-Band DXing*, (4th ed), ARRL, p 13-28
[6]T. Shiller, N6BT, Pacificon 08, Antenna Forum, Oct 17-19, 2008, **www.pacificon.org**

Antennas with Gain and Bandwidth for 80 and 160 Meters

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On 80 and 160 meters an antenna with even modest gain can give you a very real edge in a contest. Unfortunately, the long wavelength (λ/4 is 70 feet at 3.510 MHz and 134 feet at 1.83 MHz) associated with these bands makes gain antennas very large.

An additional problem is the width of the 80-meter band. It's tough to design an efficient antenna that will work over more than a small portion of the band without retuning. Phased arrays of λ/4 verticals work great but require a great deal of effort, real estate and money to bring on line.

For most of us simple wire arrays, such as the half-square and bobtail curtain, are more practical. When compared to a ground-plane antenna over average ground they have gains of 2.1 dB and 4.6 dB respectively. While useful, both of these arrays have quite narrow SWR bandwidths, typically <100 kHz for SWR <2:1 on 80 meters. While it is possible to make these antennas resonant at multiple points within a band,[1] the SWR between these points will still be high. Various schemes for switching in and out tuner components have also been used. It would be better if we could keep the antenna really simple and still have the gain and bandwidth. Another problem with the bobtail curtain is that it is a full wavelength wide (approximately 280 feet on 80 meters), limiting its use to those with large lots.

The Bruce Array

There is another simple array that has been mostly forgotten by hams. The Bruce array has been around since the '20s.[2-6] This antenna has appeared in the *ARRL Antenna Book* since the first edition in 1939, but the section on the Bruce array has been abbreviated over time leaving out a number of interesting ideas.

A few variations of the Bruce array are shown in **Figure 1**. It is simply a wire one or more wavelengths long, folded so that the currents in the vertical portions are in phase, contributing to radiation and currents in the horizontal portions that tend to cancel. Note that the wire lengths of each side of the squares are 1.05 × λ/4. The square loops in the Bruce behave very much like quad loops—they also have to be made longer for resonance. This is a bi-directional broadside vertical array with all the elements in phase with more or less equal currents. This antenna offers a number of advantages:

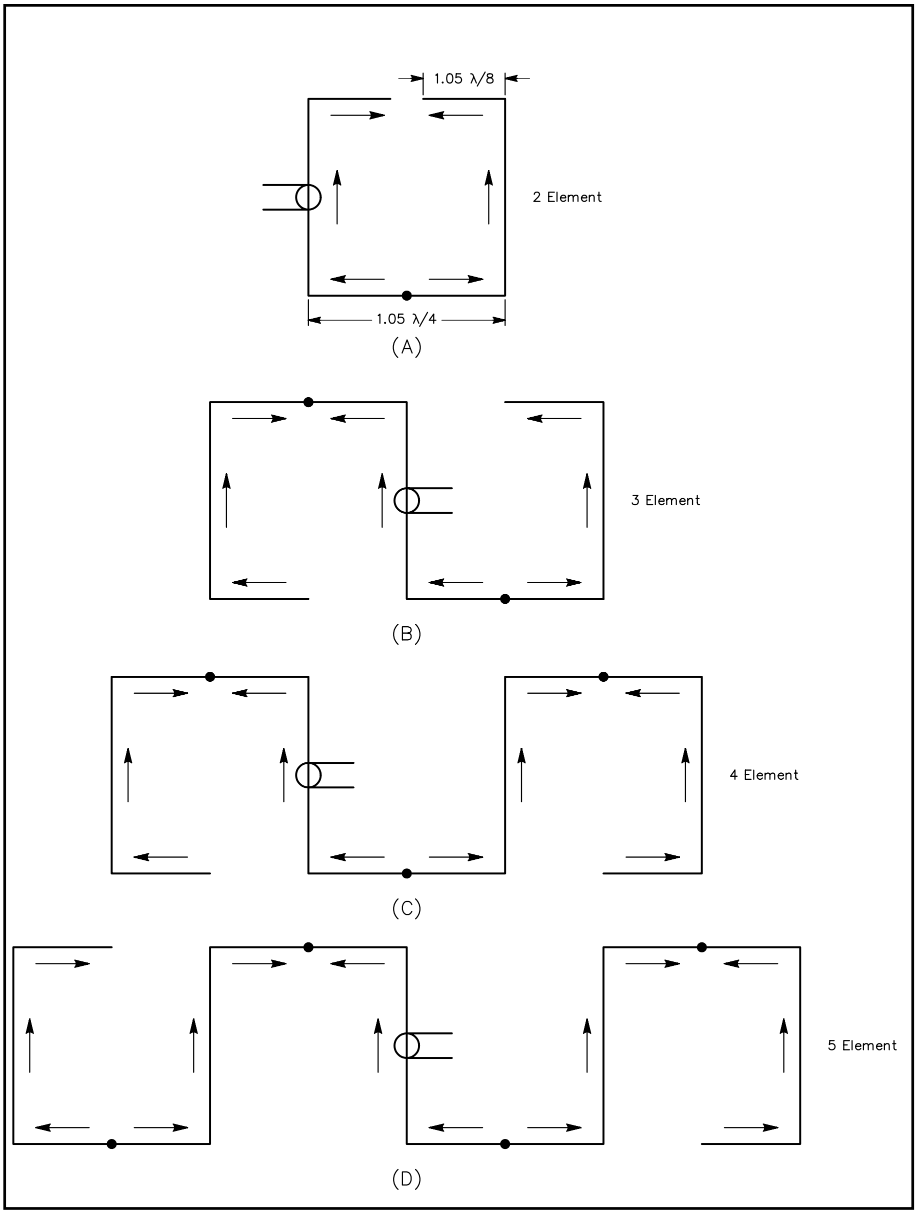

Figure 1—Bruce arrays with 2 to 5 elements. The feed points are nominal. See the text for other feed arrangements.

- It is only λ/4 high.
- The size can be adjusted to fit the space available.
- It provides substantially greater SWR bandwidth than either the half-square or the bobtail curtain.
- It can be fed at several different points to suit a given installation.
- No ground system is required.

One comment on ground system requirements. Half-squares, bobtail curtains, Bruce arrays and other nominally ground-independent vertical antennas can all be operated without the usual ground system associated with single verticals. That is not to imply that an extensive ground system under these antennas would not reduce ground losses to at least some degree. As I show later (in **Figure 7**) an extensive ground system can be employed under

[1]Notes appear on page 52.

the Bruce array if you have the space and patience to install one.

Figure 2 is an overlay of the free-space patterns for 2-, 3-, 4- and 5-element Bruce arrays. As you would expect, the wider the array the greater the gain. **Figure 3** shows a pattern comparison between a 4-element Bruce ($^3/_4$-λ wide) and a 3-element Bobtail curtain (1-λ wide). The Bruce has just as much gain but is a full λ/4 shorter (130 feet on 160-meters!).

As you make the Bruce wider (adding more elements) the gain increases, the pattern narrows and side lobes begin to appear. In general more than five elements are not worth the trouble—the pattern is already narrow and the sidelobes are starting to become significant. If you really want even more gain (approximately 3 dB), hang a Bruce reflector about λ/8 behind the main array. Alternatively you could space the second Bruce λ/4 away and drive it with a 90° phase shift. This would produce a unidirectional pattern that could be switched 180°. Of course this is getting away from the idea of

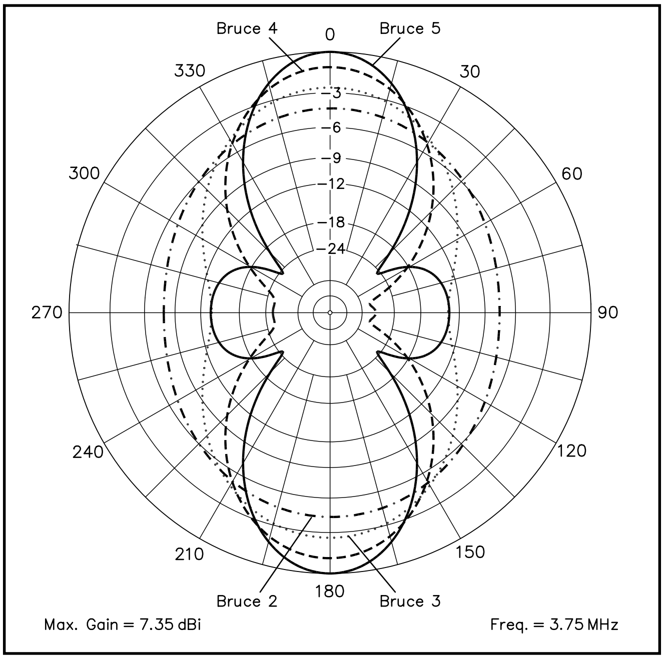

Figure 2—Comparison of free-space radiation patterns for 2 through 5 elements.

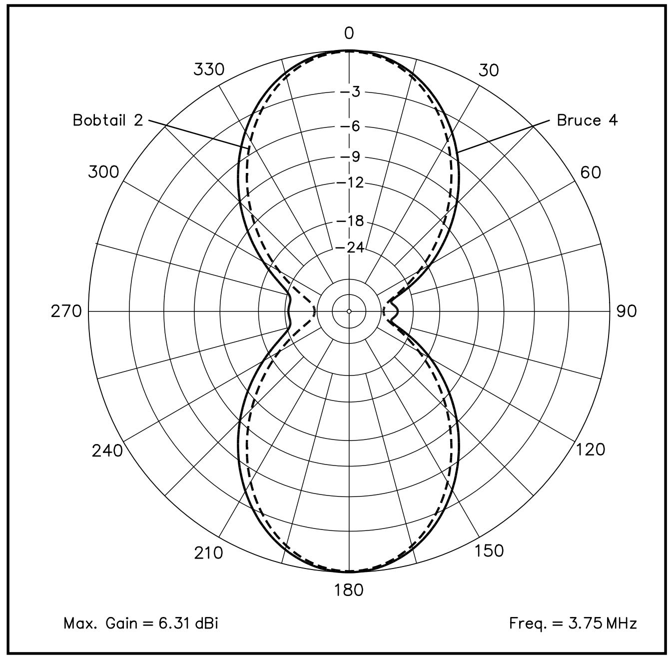

Figure 3—Comparison of free-space radiation patterns between a 4-element Bruce array and a bobtail curtain array.

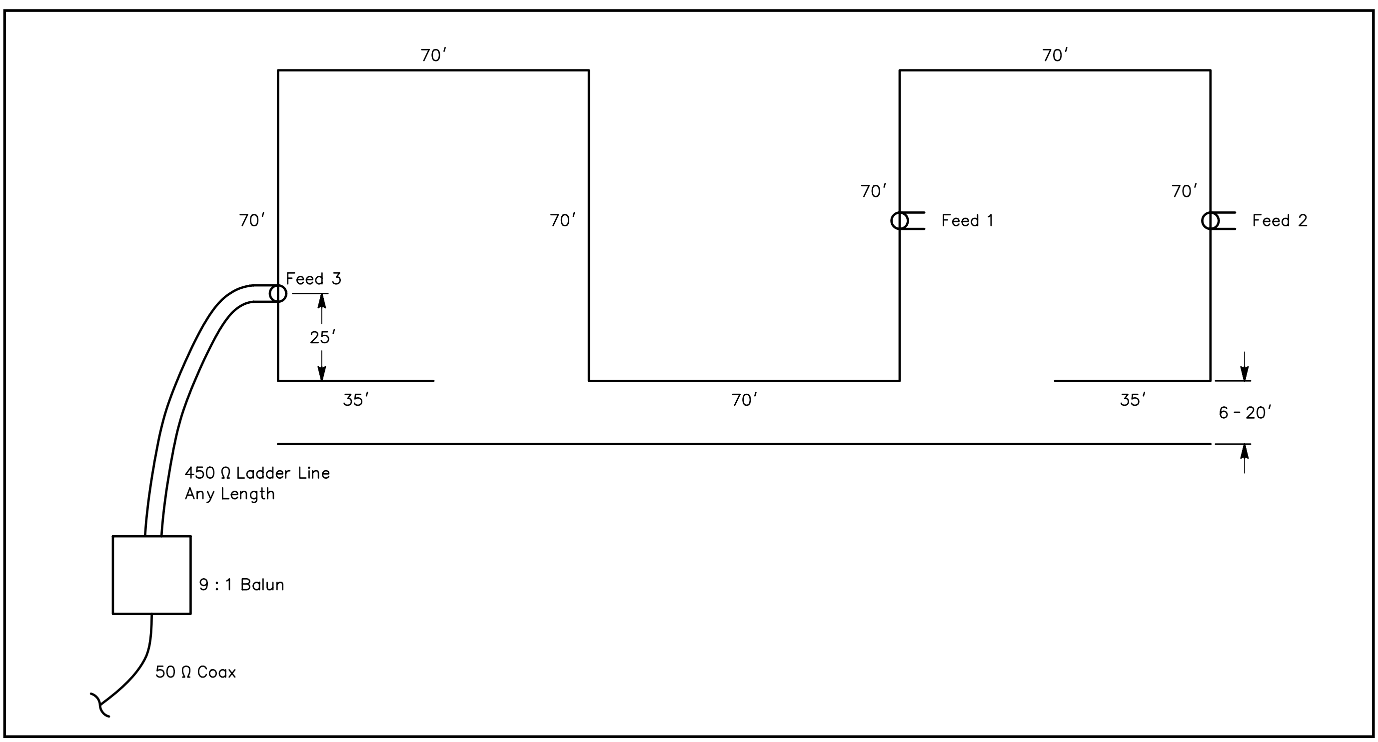

Figure 4—The 80-meter Bruce array employed at N6LF. Alternate feed points are indicated.

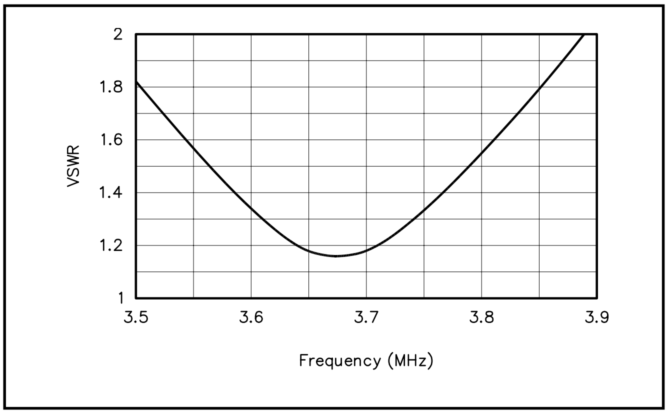

Figure 5—The SWR plot for the N6LF Bruce array.

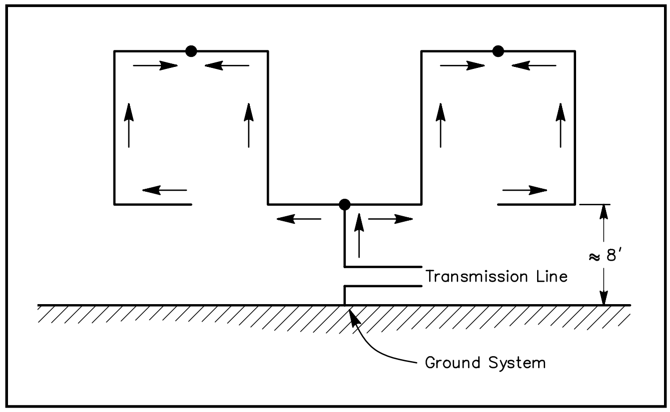

Figure 6—An example of driving a Bruce array against a ground system. This feed scheme produces a very symmetrical pattern with deep nulls off the ends if the array itself is symmetrical.

simplicity that is the basic advantage of the simple version of the Bruce.

I have used the 4-element Bruce array shown in **Figure 4** to good effect. As indicated, the array can be fed at several different points. (I've only shown a few of these.) The impedance at feed-1 and feed-2 is close to 370 Ω—a good match for #16 × 1-inch ladder line.

I chose to feed my antenna at feed-3, slightly off-center from a current maximum. At this point the input impedance is about 450 Ω. This works very nicely using #18 × 1-inch ladder line down to the ground where I connect a 9:1 balun and use 50-Ω coax for the run into the shack. The ladder-line can be any length as it is operated with low SWR.

Figure 5 shows a typical SWR plot that has a 2:1 SWR bandwidth >400 kHz. This covers most of the 75-/80-meter band. The actual bandwidth in a given installation will depend to some extent on the ground characteristics and the height above ground of the bottom of the array.

The gain of this antenna, when compared to a λ/4 vertical with 8 elevated λ/4 radials, is about 4.6 dB—very worthwhile indeed. The pattern is bi-directional with a –3 dB beamwidth of 55°. When fed at one of the inner vertical elements the pattern is very symmetrical. Feeding at one of the outside vertical sections, as I have done, introduces some asymmetry in the pattern but the small side lobe that appears is still 15 dB or more down from the main lobe.

Like the half-square and the bobtail curtain, the Bruce antenna has deep nulls off the ends and is relatively insensitive to the presence of a metal tower off the ends. If you space the outside elements 10 feet or more away from the tower you can use a tower (or towers!) as supports without degrading the pattern greatly. In my case I used a very tall (100 feet to the support point) fir tree at one end and a 95-foot pole at the other end.

One of the nice things about the Bruce antenna is that there are several other ways it can be fed. For example, if you already have a vertical with a ground system you can simply hang the Bruce over the ground system and feed it as you did the vertical (see **Figure 6**). The feedpoint impedance will be 200 to 400 Ω and may be reactive. This method of feed was used in the original versions of the Bruce array but they seem to have been forgotten by hams.

An alternative feed arrangement would be to use an elevated radial system as shown in **Figure 7**. A minimum of two radials are needed, but you could use more (just as you would for a ground-plane vertical). The dimensions shown in Figure 7 are for phone band (75-meter) operation.

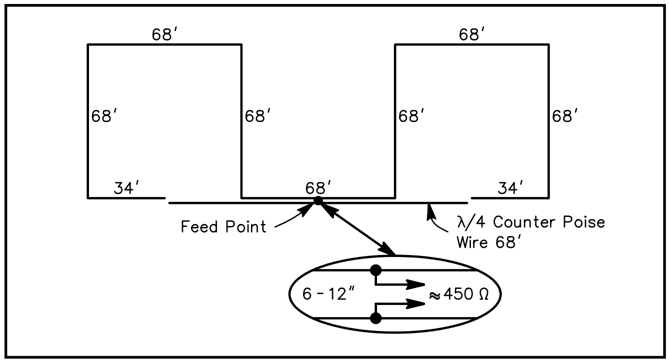

Figure 7—A 75-meter Bruce array driven against an elevated radial system. As few as two radials can be used. More radials will reduce ground losses somewhat.

Conclusion

If you have a couple of supports from which to hang an array, then you should give the Bruce array some consideration. It is very simple and flexible and is one of those antennas that just seems to "want to work."

I noticed that the dimensions are not critical. If you have some height but not enough width, you don't have to make the bays square—you can make the vertical sections taller and the horizontal sections shorter. Conversely, if you have plenty of width but not enough height, you can use shorter vertical sections and longer horizontal sections. Variations of up to ±20% in the height-to-width ratio have little effect on the gain and general performance.

Good luck! I'll listen for your thunderous 80-meter signal in the next contest.

Notes

1 *ARRL Antenna Book*, 18th Edition, pp 6-13 - 6-16.
2 Oswald, *Transatlantic Telephone Service*, Bell System Technical Journal, 1930, p 287.
3 *Admiralty Handbook of Wireless Telegraphy*, 1931, pp 820-821.
4 E.J. Sterba, *Directional Transmitting Systems*, IRE Proceedings, Volume 19, Number 7, July 1931, p 1202.
5 *The "Radio" Antenna Handbook*, 1936, pp 57-58.
6 *Admiralty Handbook of Wireless Telegraphy*, Volume 2, section 46.

■

Getting the Most from Half-Wave Sloper Arrays

So you want to put up a really big 160-meter directional array? Here are some tips.

By Rudy Severns, N6LF

For those who have a single tall support, $\lambda/4$ or higher, the half-wave sloper family of antennas described by K1WA,[1] K8UR,[2] K3LR[3,4] and others can be a relatively simple way to make an antenna with modest gain, good F/B and an electrically steerable pattern on 80 or 160 meters. Previous articles have provided much information on this family of antennas. But having just come through a cycle of building several variations, I found that a lot more needed to be said. Fig 1A shows two different half-wave sloper array element shapes that I will refer to as the "K1WA" and the "K8UR," with the understanding that many other arrays use these shapes. Fig 1B shows some other possible element shapes.

This article presents a 160-meter variation of this family of antennas and, more importantly, a discussion of the details of how to get such a beast working really well. You could simply put up four precut dipoles, with $^3/_8$-λ phasing lines *à la* K1WA and the array will work with reasonable F/B. However, with extensive modeling I discovered that fanatical attention to detail and tuning and adding a first-class ground system will greatly enhance performance.

In the summer of 2000, I put up a pair of 150-foot wooden poles placed east-west along a ridge. I called George, W2VJN, who had been using half-wave slopers for years and asked for his advice. That began a long series of conversations and experiments. George supplied many key insights while I was doing field testing and modeling work. I very quickly learned how difficult it is to actually obtain the performance predicted from modeling in a real 160-meter antenna. The configuration reported here is a bit different from earlier versions but is simpler to build and, more importantly, easier to get up and running at full performance.

Initial Experiments

You can use several slopers spaced uniformly around the support to produce a steerable pattern. These can be simple $\lambda/2$ slopers (K1WA) or diamond-shaped (K8UR), with the lower ends brought back to the base of the support. The slopers may be driven as a phased array (K8UR) or as a parasitic array (K1WA and K3LR). It is very common to drive and/or load the center of each element.

There is another possibility, however. You can voltage feed at the lower ends of the elements. This approach, while certainly known, has not gotten much press. It has some advantages when K8UR-shaped elements are used.

My experiments began with a 2-element version of the K3LR antenna, where the length of the feed line from the center of an element to a switch box is adjusted to tune one element as a reflector while the other element is driven. The non-driven element is open-circuited at the switch box. This allowed me to switch the direction of the main lobe from east to west with a SPDT relay, selecting one or the other feed line. I quickly discovered how much the

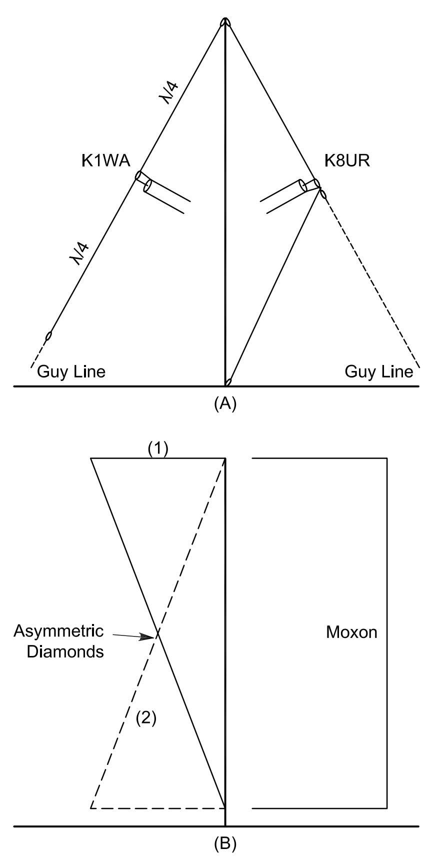

Fig 1—Half-wave sloper array element shapes.

[1]Notes to appear on page 80.

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shape of the actual array differs from the nice straight-line wires we use in modeling.

First, there is the sag in the 130-foot wire spans on each side of the feed point where the guy lines used to spread the array are attached (see Fig 1A). I found I had to move the guy-line anchor point much farther away from the support to get enough tension to control sag. This was made worse by the weight of the roughly 114 feet of RG-8X feed line going back to the support.

Testing showed that the F/B was not very high—a few decibels at most—and the feed-point impedances were substantially different from predicted, making for a poor match. Further checking with a clip-on RF ammeter showed lots of current on the feed lines.

I then modeled the antenna with the actual sags in the elements and feed line I use Nittany Scientific's *GNEC*, which implements the *NEC*-4 catenary wire (CW) and insulating sheath (IS) cards. Guess what? Lots of RF on the feed lines, lousy F/B and mediocre gain were indicated. I put common-mode chokes baluns at the feed points (more weight, more sag, more loss) and that helped, but only a bit. The extra weight also increased the tension in the wire and guy line to the point where wire stretch and subsequent detuning became a problem.

I also found that if I wanted to actually tune the elements so that they behaved as a parasitic array, I had to make measurements at ground level—at the end of about 200 feet of coax. I had to calibrate the coax and then transform the measurements by the transmission-line equation to get the actual feed-point impedance.

Then I had to lower the array and trim each end of the elements. This was doable but what a pain! It was clear from the modeling and measurements that the actual shape of the elements and whether or not insulated wire was used had a significant effect on the behavior of the array.

I groused about all this to George as we drove up to the Northwest DX convention last June (2001) and he said "Why not use voltage feed instead?" The light went on. I wanted a parasitic array in the K8UR configuration, but voltage-fed at the bottom of the driven element. The other elements would be open, acting as reflectors. This would have some advantages:

1. All the coax, baluns, relay boxes, etc, hanging up in the air are eliminated. That removes a lot of stress on the array and lowers the expense and the loss in the cable, even on 160 meters. The extent of cable loss was pointed out in the K3LR articles.
2. There is no longer any need for the elements to assume a symmetrical shape (equal lengths at top and bottom) to minimize coupling to the feed line hanging from the center point. They can have considerable deviation from symmetry, as shown in Fig 1B.
3. All the measurements, pruning, tuning and switching can be done at the base of the support, right at ground level. Very convenient!
4. With much less weight and windage the array is less susceptible to damage. This is particularly important if you live in an area where icing is a problem.
5. The loading on the support is much less. Not a big deal with a guyed tower, but important when using a tall wooden pole or other light support.

Of course, there are some disadvantages too:

1. The switching relay(s) must now be capable of handling high voltage (>5 kV), mandating vacuum relay(s).
2. A tuning unit is required at the base of the antenna.

The Array At N6LF

Fig 2 shows a side view of the array presently installed at N6LF. This has performed very well this winter (2001-2002). Note that the shape of the individual elements is not symmetrical—the triangle apex is well above the midpoint. In my installation I have another 150-foot pole 300-feet east and an anchor point at 100 feet in a tree 400 feet west of the main support. These allowed me to raise the apex (corner) of the element farther above ground, reducing ground losses somewhat.

The elevation pattern for this array is shown in Fig 3 at several points across the band. I maximized gain at 1.830 MHz, where the F/B is about 7 dB. Below that both the gain and F/B drop off but the gain is still quite useable. As you go up the band the gain falls slowly but the F/B improves. At 1.890 MHz the low-angle F/B is very good (about 24 decibel) but if you look at the full rear quadrant the F/R is only 12 dB, pretty much in line with the expectations from free-space modeling.

Because I almost always use a Beverage antenna for receiving I elected to go for maximum gain at 1.830 MHz. You could just as well go for high F/B and sacrifice a half dB or so of gain. I adjusted the tuner for minimum SWR at 1.830 MHz. This gave an SWR of 1.1:1 at 1.800 MHz and 2:1 at 1.970 MHz. It would have been quite possible to adjust for an SWR < 2:1 over the whole band but the gain starts dropping off above 1.900 MHz.

This antenna has been up since August 2001 and was used in the ARRL, Stew Perry and CQ CW 160-meter contests. It has performed very well indeed, despite the truly terrible conditions on 160 meters at this part of the sunspot cycle and due to my less-than-ideal location. During the

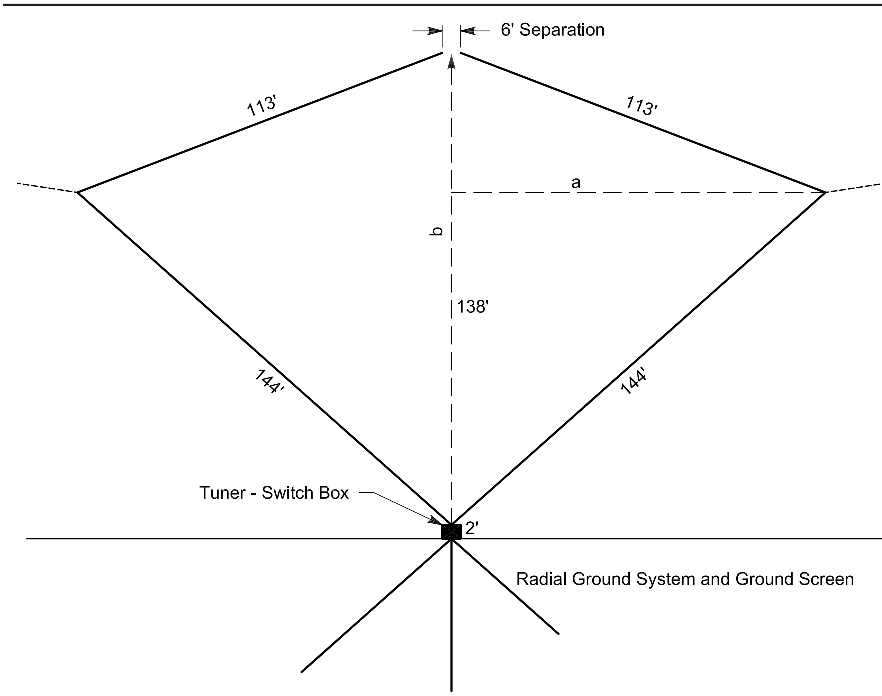

Fig 2—Present array at N6LF.

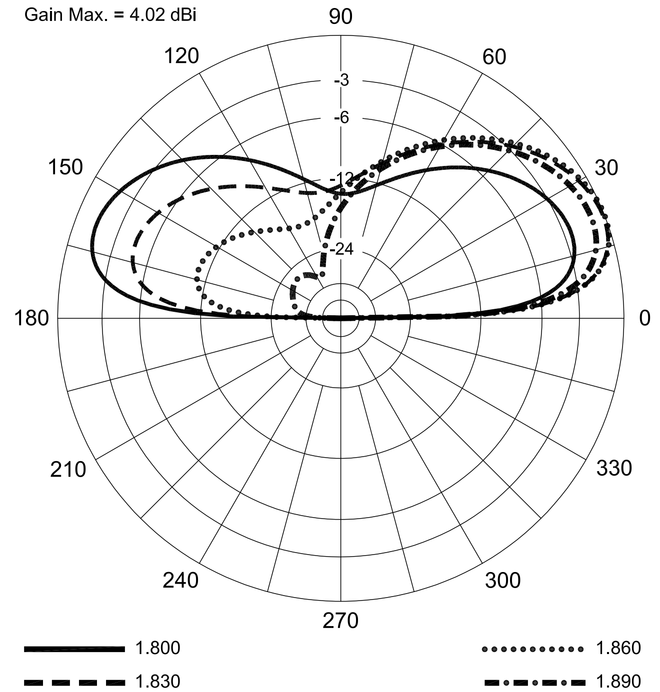

Fig 3—Elevation radiation patterns at 1.800, 1.830, 1.860 and 1.890 MHz for the N6LF two-element array.

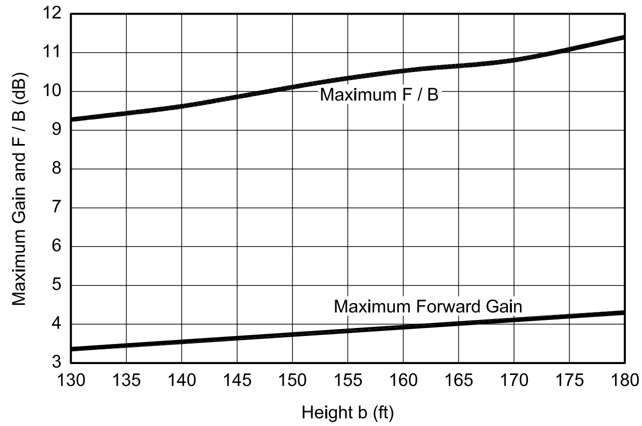

Fig 4—Typical maximum free-space gain and F/B for two-element diamond-shaped array.

ARRL 160-meter contest George was operating on the east coast from W3BGN's shack and he compared signals from the west-coast stations.

Of course, K6SE (who was using a balloon vertical over the Salton Sea salt flats) beat us all hands down. Compared to the other big stations, however, my signal was right in there, so the antenna is clearly starting to work well. And there is even more I can do to improve it. The following details how I achieved my level of performance and what could be done to improve it.

Comparison To Other Antennas

Even with the best design and construction, this antenna will not beat out an equally well-designed and installed four square. It will also be outperformed by the Spitfire antennas[7] that are similar to this antenna. The Spitfire uses the supporting tower as the driven element and the parasitic elements as reflectors and directors to form a 3-element, rather than a 2-element, vertical Yagi with a steerable pattern. However, when done well, the full-wave sloper family of antennas is not hopelessly outclassed—and they are far easier and less expensive to build compared to a four-square system if a suitable support is already in place.

In all of the modeling to follow, ground is assumed to have σ = 0.005 S/m (conductivity) and ε = 13 (relative dielectric constant). For the radiation patterns, the main axis of the array is in the (y, –y) direction (90° to 270°).

Element Length

In most two-element Yagi designs the length of the driven element is adjusted so that the feed-point impedance is resistive. The parasitic element length is adjusted to perform either as a director or a reflector. In low-frequency arrays, the size is usually much too large to allow the array to be physically rotated, and the driven and parasitic elements must be interchanged to switch the pattern. This can be accomplished in several ways. The most common is to add or subtract some length or loading and then interchange the element you wish to feed as the driven element.

There is another possibility that has not received much attention. If you take two equal-length parallel conductors, spaced on the order of 0.1 to 0.3 λ, one of which is driven and the other parasitic and do some free-space modeling, you will find that as you increase the length of the elements (keeping both the same length) that the parasitic element will first act as a director and then as a reflector as *both* are made longer. The advantage of this is that both elements are identical and there is no change in length or loading when you change direction. You simply change which element is driven.

This is particularly helpful when multiple elements are used and only one is driven at a time. For an end-fed element this makes it very easy to change directions. You simply use a system of relays to select which element is to be driven, leaving the other element open to act a reflector. A single tuning network at the base of the antenna is required, and it sees an impedance that does not change as the pattern direction is changed.

Of course the driven element in this case will not be resonant and will exhibit some reactance. With a simple parallel L-C tuner at the base, that is not a problem. Typically, the reactance will be equivalent to 5 to 10 pF, which can easily be accommodated by adjusting the tuning capacitor in the tuner.

Modeling two elements in free space gives a general idea of how this works for K8UR-shaped elements. The gain and F/B will depend on the overall height of the diamond (dimension "b" in Fig 2) and the width (dimension "a" in Fig 2). Fig 4 graphs typical free-space gain and F/B for elements varying in height from 130 to 180 feet at 1.830 MHz, using #12 bare copper wire. Notice these are the maximum values found by fixing the height and adjusting the width in the model.

As in any Yagi, maximum gain and maximum F/B do not occur for the same dimensions. In general, at the maximum F/B point the gain will be down by about 0.5 dB. There are no surprises here—the taller the array, the more gain and F/B you can obtain. However, even at $\lambda/4$ ($\approx$ 130 feet), there is usable gain and F/B, even though this is half the length ($\lambda/2$) of normal Yagi elements.

Notice also from Fig 2 that I set the separation distance between the top ends at 6 feet. This is not a magic number, but the distance between the ends of the elements does affect the behavior. Spacings of order of 1 to 3 feet each side away from the support structure seem to work fine, although others should work also. Chose a spacing during the design phase and be careful to stick with it when erecting the array.

Element Shape

Besides the obvious mechanical difference between the K1WA and K8UR elements, there are important radiation-pattern differences too. If you start with a single half-wave sloper, with the top at 150 feet, the radiation pattern in Fig 5 will have a combination of vertical and horizontal radiation. That's no real surprise, since you have a slanting dipole.

When you combine this into a two-element array, however, some funny things start to happen, as shown in Fig 6A for in-phase and Fig 6B for 180° out-of-phase excitation. The pattern doesn't look anything like the broadside-endfire you expect in a 2-element vertical array. The problem is that the vertical and horizontal fields add up differently and the array does not behave quite as you might expect. While 160-meter operators generally favor vertical polarization for transmitting, for receiving the combination of vertical and horizontal polarizations may help. I hasten to say that this is speculation on my part.

For a four-element half-wave sloper array, where three of the elements are reflectors, the radiation pattern is shown in Fig 7. The total pattern is quite reasonable but is made up of vertical and horizontal components that individually have very different patterns. Again, it is not clear if there are any advantages or disadvantages to this mixed polarization.

The K8UR-element shape has a very different pattern. Fig 8 shows the

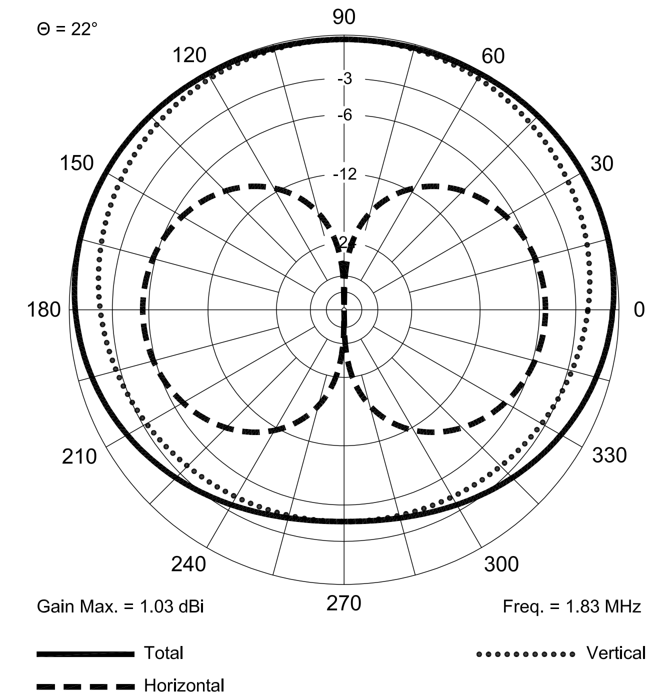

Fig 5—Vertical, horizontal and total pattern at 22° elevation for a single half-wave sloper with the upper end at 150 feet.

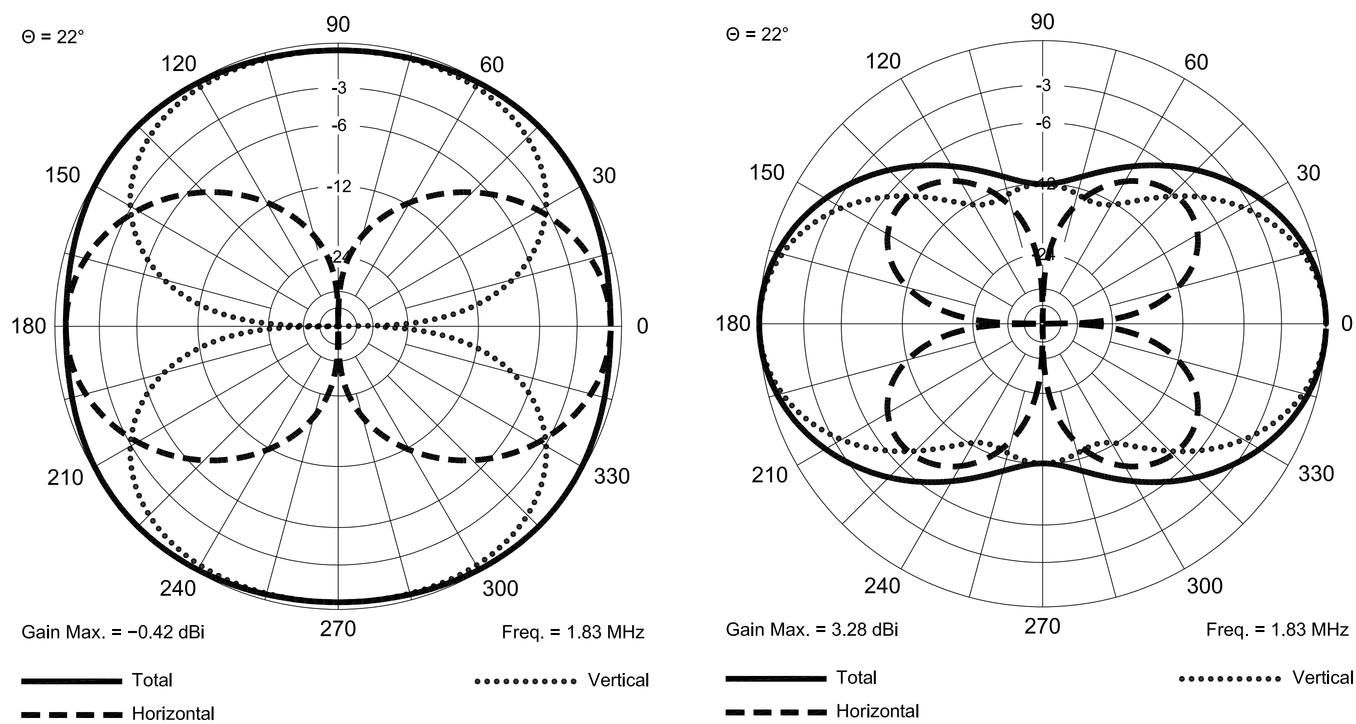

Fig 6—Vertical, horizontal and total pattern at 22° elevation for two half-wave slopers. At A, driven in-phase and at B, 180° out of phase, with the upper ends at 150 feet.

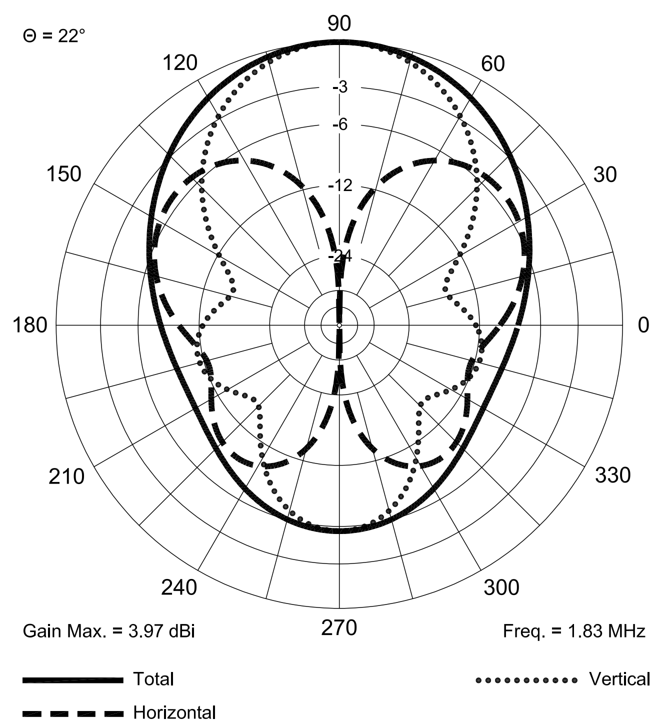

Fig 7—Vertical, horizontal and total pattern at 22° elevation for a four-element K1WA array, one driven element and three reflectors.

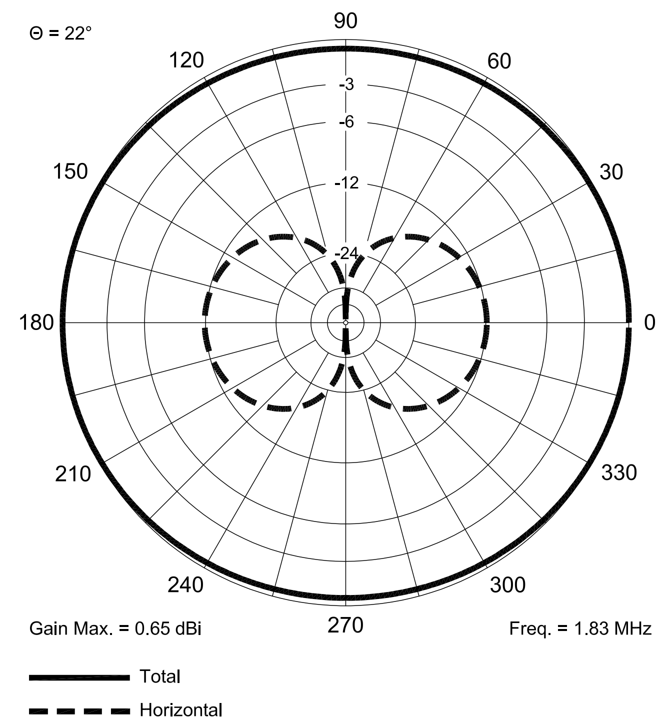

Fig 8—Vertical, horizontal and total pattern at 22° for a single K8UR-shaped element.

patterns for a single K8UR element. The horizontal component is much lower, –12 dB or more, and contributes little to the total pattern. This is one of the reasons that this shape is usually preferred if you are building a vertically polarized array.

In the K1WA and K8UR antenna models, I fed the elements at the center and I made every effort to keep things symmetrical to minimize coupling to the feed line. However, when fed from the end there is no necessity to make the element shape symmetrical. Fig 1B shows two asymmetric shapes (1 and 2). The advantage of shape 2 is that the anchor point for the guy line is much closer to the support. The overall space required for the antenna is greatly reduced. The downside of shape 2 is that it places a high E-field close to ground for a considerable distance. This increases ground losses *if an extensive ground system is not used under the antenna*.

Lifting the apex up, as shown in shape 1, reduces the ground loss significantly but requires a high anchor point for the guy lines. In the installation at N6LF these two points were available and the initial design did not use an extensive ground system. Later I realized just how much could be gained by adding a ground system. With a good ground system the additional loss due to shape 2 can be almost eliminated and the guy-line anchor points moved in much closer to the main support.

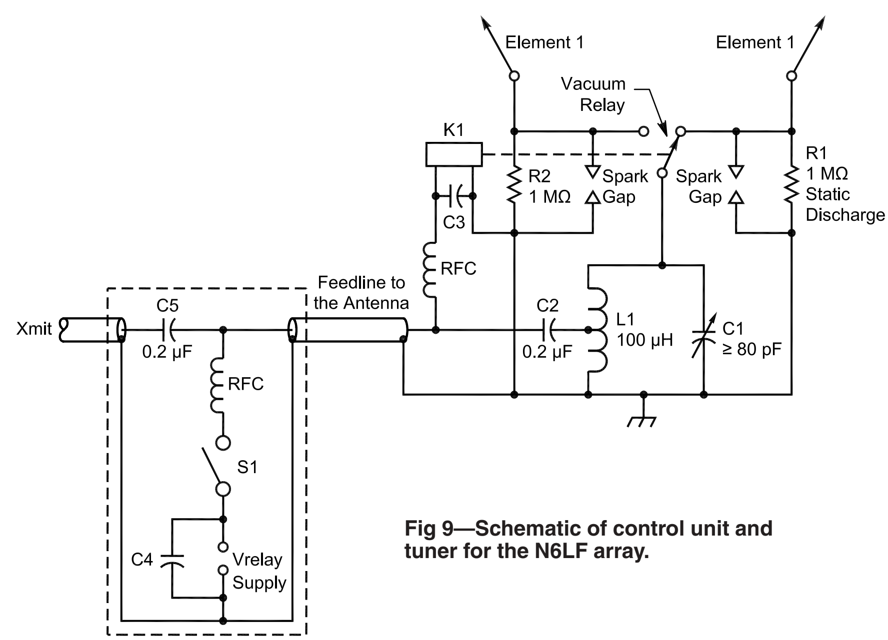

Fig 9—Schematic of control unit and tuner for the N6LF array.

If two high supports are available, then you can use the Moxon rectangle (shape 3 in Fig 1B). This yields somewhat better gain and F/B but does require two supports.

Tuner Design

Fig 9 is a schematic of the tuner. The heart of the tuner is a simple parallel-resonant L-C circuit, with a tap on the inductor for matching to the feed line. *GNEC* predicted a feed-point impedance of 5318 – *j* 1776 Ω and the actual array impedance was within 5% of this. Notice that this is the series-equivalent impedance shown in the sidebar, "Design of the Tuner").

To design the matching network,

the series-equivalent is transformed to the parallel-equivalent circuit. The parallel equivalent impedance is about 6 kΩ in parallel with 5 pF. The next step is to chose a loaded Q. Typically this would be in the range of 5 to 10, so I chose Q = 5 to minimize the size of the tuning capacitor (C1), which must be rated for > 5 kV peak at 1.5 kW operation. A lower loaded Q also reduces the circulating current and increases the match bandwidth somewhat.

The downside of a low loaded Q is that the inductor is larger, as are its losses. However, as shown in the sidebar, for unloaded coil Q > 200 the loss is less than 0.1 dB. The coil I used was 6 inches in diameter by 5 inches long, with 29 turns of #12 wire. It had a measured unloaded Q higher than 400 on an HP 4342A Q-meter. The coil loss is thus quite small.

I use 7/8-inch CATV cable for the long runs to the shack. To match the 75-Ω feed line, the tap was 4.5 turns from the bottom of the coil. George reminded me that this would be a good place to use a shielded loop made of coax with a series-tuning capacitor. This would give better harmonic suppression and provide dc decoupling and some improvement in lightning protection. It also would provide more isolation from BC station pickup, which can be a real problem in an antenna this large. In my case I went with the simpler direct tap and it has worked well but when I improve (translation: rebuild because I can't stop fooling with it) the antenna next summer I will probably incorporate a shielded coupling loop.

C_1 is a vacuum variable, but it could just as well be an air variable with widely spaced plates. The capacitance required is only of the order of 80 pF and not all of that needs to be variable. You could for example use a fixed 50-pF capacitor in parallel with a 30-pF air variable, which would be relatively small physically. Keep in mind that these network values are for a particular design. Other designs may have somewhat different impedances and the component values must be selected accordingly.

Relay K1 switches between the east and west elements in the array to switch the pattern. I used a surplus RB1H Jennings SPDT vacuum relay rated for 12 kV. The relay coil called for 26.5 V but I found that it would start to pull in at 16 V and worked just fine with 20 V or more to activate it. For the dc power source I used a wall transformer power supply rated for 18 V, but which actually puts out 22 V. The relay is activated through the feed line using dc-blocking capacitors (C2 and C5) and RF chokes. The control unit is located in the shack and I simply flip switch S1 to change directions.

If you want to use three or four elements, then more relays will be needed. Fig 10 shows an arrangement of two relays for three elements. It is possible to use ac combined with dc and some diodes to control as many as three relays from the shack through the coax, as is done in the Ameritron RCS-4 remote coaxial switches. Of course, a separate control cable can be used also. In my case the distance from the shack to the array is > 700 feet, so I opted for feed-line control.

Capacitors C2 and C5 are for dc blocking. They must carry the full RF current, about 5.5 A at 1.5 kW when the load is matched to 50 Ω. I chose to use multiple NPO disk ceramic capacitors in parallel because they were readily available and inexpensive. NPO capacitors are larger for a given capacitance than other ceramic capacitors, but they have lower losses. You may be tempted to use 0.1 µF capacitors instead of a number of 0.01 or 0.02 µF capacitors in parallel, but be careful. The self-resonant frequencies for the larger disk ceramics can approach 1 MHz and you don't want the capacitor to be operated at or above its self-resonant frequency. In addition, a number of smaller capacitors in parallel will have much more surface area and cool much better, enhancing the current-carrying capability, which is primarily limited by temperature rise. Arrange the parallel capacitors with space between them so each one can cool itself.

There are a few other parts in the box that deserve some attention. The 1-MΩ resistors connected from the end of each element to ground are there for static discharge. The long wires in the array can develop high static potentials under some conditions. That potential on the free-floating reflector element can cause the relay to arc when transmitting. I happened to have on hand a bunch of 2-W, 100-kΩ carbon-composition resistors, so I simply built up R1 and R2 using 10 of these in series.

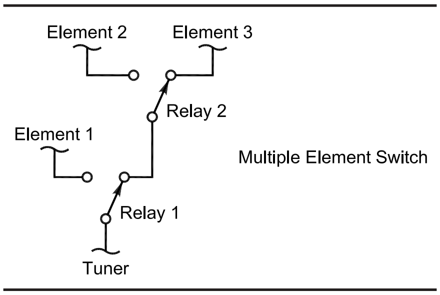

Fig 10—Relay connections for a three-element array.

The 20-W overall power rating was not really necessary, but using several resistors in series increased the voltage rating. Thus I did not have to worry about arcing the resistors while transmitting, when there is a high potential at the ends of both the driven and parasitic elements. The loss introduced by these resistors is small. I also placed a spark-gap to ground across the drain resistors for lightning protection. A lightning strike anywhere within a quarter mile of this large antenna will induce very high voltages and full-up lightning protection is absolutely necessary.

The layout of the tuner is shown in Fig 11. I chose a plastic container for the enclosure because they are readily available in a wide variety of sizes and are economical. The use of a plastic enclosure also keeps the coil's unloaded Q high by keeping conducting surfaces away from it. A large metal box would also work and might have some advantages. One disadvantage of the plastic box is that ultraviolet from the sun will degrade it. In Oregon that is not a big problem but I do keep it covered with a shade cloth.

For ground within the box I used a 2-inch copper strap, which is brought out the bottom of the box to real ground. It is very important to have a good RF *and* lightning ground at this point. I use a 24-inch diameter by 8-foot culvert pipe surrounding the base of the support pole acting as a socket so that the pole can be removed with a crane for repair and alterations. This provides an excellent ground. If

Fig 11—Photograph of the tuner.

you use a tower, there should be a series of ground rods at the base for lightning grounding in any case and these can be used as a starting point for your RF ground system.

Tuning and Adjustment

One of the advantages of parasitic arrays is that the phasing of the element currents is automatically taken care of by tuning the element lengths properly. You can thus avoid the multiple matching networks and feed lines used in a phased array, where every element current and amplitude must be adjusted.

Unfortunately, the tuning in a parasitic array is strongly effected by the size and shape of the elements, which vary with tension and wire size. Sometime I wonder whether the phase of the moon manages to get into the act!

When you use insulated wire for the elements, the insulation material itself has a considerable effect. For a typical 20-meter array made with aluminum tubing, dimensions derived from modeling are usually very close and any adjustments needed are merely for matching. For a large 160-meter wire array with an arbitrary element shape that is not the case. The elements *must* be carefully tuned in the field for full performance.

What I elected to do was to design the array in *GNEC* with the element shapes as close to reality as possible, including insulation, sag, etc. When I optimized the array, I modeled one element alone and determined its self-resonant frequency. In the field I then erected one element at a time and adjusted it to be resonant at the same frequency as the model. I used solid #12 THHN insulated wire because it was much more economical than bare #12 (for some strange reason) and available in 2500-foot reels at a retail outlet near me.

Besides cost, I prefer to avoid the surface oxidation normal in bare wire. As I showed in my *QEX* article,[5] insulation in reasonable condition introduces very little loss, while an oxidized surface introduces significant loss, at least in low-impedance arrays. However, the insulation significantly changes the resonant frequency of an element and it increases the weight, requiring more tension to maintain the shape.

For the first pass I erected an element with the shape shown in Fig 2. The upper dimension for this first try was 113 feet and the lower section 153 feet. With bare wire, the resonant frequency was 1.838 MHz and with insulated wire the resonant frequency dropped to 1.789 MHz. That's a shift of almost 3%—no big deal in a dipole but bad news for a Yagi element. I experimented with other wires and insulations that had even larger frequency shifts.

So I went back to *GNEC* and modeled the resonant frequency with bare wire and with two different types of insulation. The insulation on THHN wire is listed as having a dielectric constant in the range of 3 to 4, so I used a value of 4. Back in the field I erected elements using bare wire and the two different insulations. The correlation between the *GNEC* insulating sheath (IS card) calculation and the actual measurements in the field was very good. It was better than 0.1%, so long as I kept sufficient tension on the element. I did repeated measurements as a check.

For tensioning I used a filled 2.5-gallon water jug, approximately 25 lbs, on the halyard for hoisting the upper end of the element. Higher tension had very little effect on element resonant frequency. However, reducing the tension below about 15 lbs allowed the sag to visibly increase and the resonant frequency dropped by nearly 60 kHz. These two effects combined were more than sufficient to seriously mistune the element.

By trimming the length of the lower section to resonate the individual elements (one at a time, with the other element not present) and maintaining a constant tension, I was able to get the array to work very well. Testing of F/B in the ARRL, Stew Perry and CQ 160-meter contests when numerous stations were available showed that the antenna had a F/B of 8 to 10 dB. This was just about where it should have been and the performance was all I could ask for.

One problem I encountered was how to measure the resonant frequency of an end-fed element. For a single element, the feed-point impedance is approximately 6 kΩ at resonance. This is out of the range of most amateur impedance bridges. You could use a more professional bridge, such as a General Radio 916 or 1606A, but again the impedance is outside of the normal range and some range-extending tricks have to be used.

I tried using a dip meter, with very poor results. The frequency calibration is very poor in most dip meters and there is considerable frequency pulling at resonance. Even using a frequency counter to track the dip meter was not totally satisfactory because of the effect of the meter itself and the fact that hand capacitance altered the resonant frequency. The resonant frequency of the elements is very sensitive to small amounts (a few pF) of capacitive loading at the ends—right where you are trying to make the measurement.

Another problem with the dip meter and with other ham test gear can come from broadcast-band (BC) signals. In my case there is a 1-kW BC station a few miles away. At the station frequency I get induced voltages of a volt or more at the open end of an element under test, and almost 100 mV on the transmission line back in the shack. I used an MFJ-249 SWR analyzer and the AEA complex-impedance analyzer to check the match at the tap point. Both instruments go bonkers in the presence of a large BC signal.

I could make the measurement with these instruments if I placed a BC high-pass filter between the instrument and the tap, but that doesn't help with the resonant frequency measurement. I found the use of a Bird directional wattmeter to be more satisfactory for SWR adjustment and used a Boonton 250A RX meter for the resonance check. It may be possible to adapt a noise bridge with a tuned detector to make direct measurements on the antenna but I did not try that.

Indeed, I am very fortunate to have my old Boonton 250A RX meter. This is a vacuum-tube instrument that seems to shrug off the BC signal. The RX meter measures parallel impedance up to 100 kΩ and proved ideal for these measurements. I picked up mine for $35 at a corporate surplus sale many years ago and recently bought another for $46 on eBay. For low-band antenna enthusiasts this is a very nice instrument to have. Keep an eye out at flea markets and on eBay.

The frequency calibration is not adequate in the 250A but I fixed that with an inexpensive external frequency counter to monitor the internal generator frequency. I also calibrated the RX meter using 1% film resistors to further improve the accuracy. These are inexpensive and readily available.

There are other impedance-measuring instruments on the used market that appear regularly on eBay and at flea markets. A more modern instrument that is fairly common is the HP 4815A vector impedance meter. These also go up to 100 kΩ but suffer from much greater sensitivity to BC interference than the Boonton 250A. While the HP 4815A is relatively inexpensive on the used market, you have to be very careful to get one with a functioning probe. The probes are easily damaged and prohibitively expensive to have repaired.

In making the actual measurements, I was very careful to keep the layout as close as possible to the final one. I positioned the Boonton 250A in the same location the tuning unit would occupy. I then brought a 2-inch ground strap up to the same point it would be in the tuner and connected the strap to the "low" terminal of the 250A. I brought the end of the element down to where the tuner would be with a 12-inch pigtail from the insulator at the lower end of the element.

I zeroed the meter and then connected the pigtail to the "high" terminal of the 250A. Yes, all of this fussing around is necessary to get accurate measurements! An important check is to see if placing your hands on the test gear has any effect on the readings. There should be none. If there is, then you have to work on your layout, most likely the grounding. You should also try to keep away from the bottom of the element. Holding a hand near the element will shift the resonant frequency.

In the end the array has worked very well, but at the low end of the band the F/B appears to be higher than

Design of the Tuner

The tuner is a simple parallel-tuned L-C network, with a tap on the inductor to match to the feed line, as shown in Fig 6 in the main article. The task at hand it to determine the values for L1 and C1.

An equivalent circuit for the tuner and the antenna is given in **Fig A1A**. The antenna is represented by R_a and X_a in series and the tuner by the parallel combination of L1, C1 and R_1, where R_1 represents the loss in the L-C network, almost entirely due to the finite unloaded Q of L1.

The values for R_a and X_a are determined using modeling and confirmed by measurements on the completed array:

R_a = 5318 Ω and X_a = –1776 W (capacitive reactance)

The next step is to convert the series-equivalent circuit for the antenna to a parallel equivalent, as shown in **Figure A1B** using the following expressions:

$$Q_a = \frac{|X_a|}{R_a} = \frac{|-1776|}{5318} = 0.334$$

$$Q_a^2 = 0.334^2 = 0.112$$

Parallel equivalents of R_a and X_a are:

$$R_p = R_a \times \left[1 + Q_a^2\right] = 5318 \times \left[1 + 0.112\right] = 5914\,\Omega$$

$$X_P = X_a \times \left[1 + \frac{1}{Q_a^2}\right] = -1776 \times \left[1 + \frac{1}{0.112}\right] = 17633\,\Omega$$

X_P is the impedance at 1.830 MHz for a shunt capacitance of 4.9 pF.

Selection of L1 and C1

The next step is to choose a loaded Q_L for the tuned circuit when the antenna is connected. A Q_L in the range of 2 to 10 would be typical. Smaller Q_L means a smaller capacitor and a larger inductor, along with somewhat wider matching bandwidth. The problem is that a larger inductor will have greater loss. I chose Q_L = 5, which works out very well, with minimal coil loss. For:

$$Q_L = 5$$

$$X_L = 2\pi f L_1$$

$$Q_L = \frac{R_p}{X_L} \text{ meaning } L_1 = \frac{R_p}{2\pi f Q_L} = \frac{5914}{2 \times 3.14 \times 1.83 \times 5} = 103\,\mu\text{H}$$

$$C_0 = \frac{1}{4\pi^2 f^2 L_1} = \frac{1}{4 \times 3.14^2 \times 1.83^2 \times 103 \times 10^{-6}} = 73.4\,\text{pF}$$

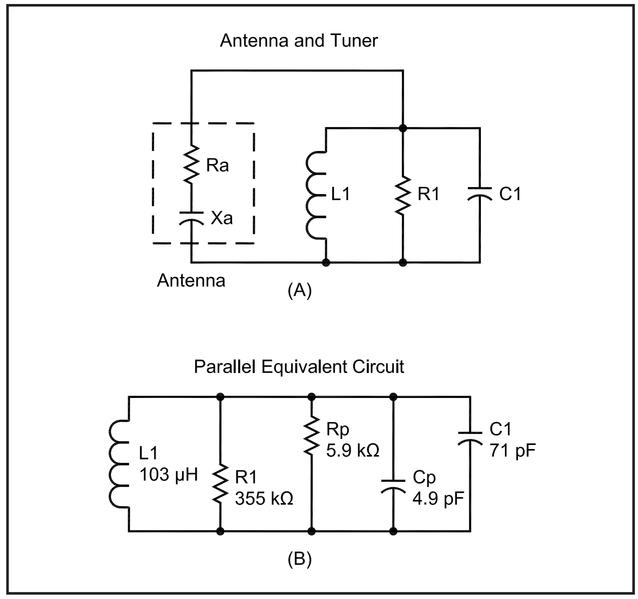

Fig A1—Equivalent circuits for the antenna and tuner.

C_0 is the capacitance needed to resonate at 1.830 MHz with L_1.

$$C1 = C_0 - C_p = 73.4 - 4.9 = 68.5\,\text{pF}$$

Loss Due to L1

R1 represents the loss in L1 and depends on the unloaded Q_1 of L1:

$$Q_1 = \frac{R1}{X_L} \text{ which means that } R1 = X_L\, Q_1$$

$$X_L = \frac{R_p}{Q_L} = \frac{5914}{5} = 1183\,\Omega$$

For $Q_1 = 300$

$$R1 = 1183 \times 300 = 355\,\text{k}\Omega$$

$$\text{Loss Ratio} = \frac{R_p}{R1} = \frac{5914}{355000} = 0.0167$$

$$\text{Loss Ratio} = 10 \times \log\left(1 - 0.0167\right) = -0.07\,\text{dB}$$

This is pretty small, and you can ignore the coil loss as long as Q_1 > 200. Coil Qs of 400 or more are not very difficult to obtain with a little care in construction.

predicted. I suspect that the final array is tuned a bit low in frequency due to stray capacitance loading at the bottom of the array, probably caused by the tuner and the final layout. Of course in an 80-meter array this effect could be exploited by switching in a small amount of capacitance to shift down from 3.790 to 3.510 MHz.

Ground System

One of the underlying assumptions for this family of antennas has been that since they use full-size half-wave dipoles, fed at the center, no ground system is required. It is true that the antennas will work reasonably well without the extensive ground system typical of a $\lambda/4$ vertical. However, the lower ends of the elements have a very high potential to ground. Using *GNEC* to plot the near-field electric (E) and magnetic (H) field intensities shows that the E field intensities are >800 V/m for 1.5 kW at ground level beneath the ends of the elements. This translates into high ground losses in the near field.

The K3LR articles mention the use of four elevated radials to improve performance somewhat, but that is about all that has been said on the subject. I began by modeling the fields under the array to get a feeling for ground losses and then modeled the array with 60 buried radials of progressively longer length out to 0.3 λ. The result was a steady increase in peak gain due to lower ground loss. The gain increase amounted to 0.6 to 1.5 dB, depending to some extent on the modeling approach. Even at the low end of this range, this is a very worthwhile improvement.

In the present N6LF array there is a ground screen made from 2-inch mesh chicken wire with a radius of 50 feet. From there, I go out another 150 feet with #12 insulated radials lying on the surface of the ground. Because I use only two elements at present, the field intensities are not uniform in all directions around the array, being higher under the elements and lower off to the sides. I therefore have placed more copper and ground screen in the high-field regions. With three or four elements the field intensities are much more uniform as you go around the array and standard symmetrical radial systems would be more appropriate. The ground system is not yet complete but already it appears to make a difference. Certainly the modeling says it should.

Wire Issues

Conductor loss, using #12 solid copper wire, is about 0.5 dB, which is reasonable but it could be reduced. Using a larger-diameter copper wire would help but also increases the weight of the element. Aluminum wire, although it has a lower conductivity than copper, can provide less resistance for the same weight. For example, a #7 aluminum wire will weigh about the same as a #12 copper wire, but will have a loss about 40% lower (taking into account skin effect, where resistance varies with the square root of conductivity). Of course, it will have more windage and the loss improvement is only a fraction of a decibel, so going to large aluminum wire may be a bit too picky.

Whether you decide to use copper or aluminum wire, stretching of the wire is a concern because it detunes the array. I tried a simple experiment: I took a 100-foot piece of copper wire, anchored one end and yanked really hard on the other end. It stretched a bit, about 6 inches ($\approx$ ½%). Conventional wisdom says that stretching the wire this way will increase its resistance by work-hardening the copper and also by reducing the diameter. I measured the wire resistance very carefully before and after stretching, using a Kelvin bridge good to a fraction of a milliohm. The dc resistance increase was right in line with the increase in length, $\approx$ ½%.

Work hardening and diameter reduction effects were too small to detect. For this reason I pre-stretched my elements and then trimmed them to length because it was more important to have the element correctly tuned than worry about a very small loss effect. This winter I had a lot of strong winds push the array around but no icing, which is very rare in any case. So far the pre-stretched elements have been stable. If you live in an area where icing is a problem, then you probably need to use either Copperweld or Alumoweld wire, both of which are much stronger but real pains to work with.

Another problem that caught me by surprise was the simple act of accurately measuring the length of a long piece of wire. I began by pulling the wire off the reel simultaneously with a long tape measure, both held in my hand. Every time I tried it I got a different final length—by a foot or more. The problem is that the wire slips with reference to the tape. So next I tried anchoring the end of the tape and stretching it out on the ground beforehand and then pulling the wire out and tensioning both the wire and the tape.

This was much more accurate and repeatable, but it also was a lot of trouble and requires a clear space of nearly 300 feet. George showed me his solution: a wire-length meter[6] like you see in hardware stores. It measures wire length to an inch in 300 feet without having to go out in the cold and wet. (It has been known to rain occasionally in Oregon.) It does the job quickly and easily and I was particularly glad I bought my own when I started to cut the numerous radials for the ground system. You have to build a simple 2×4 frame to hold the meter and a reel of wire but that's not difficult. If you want to do it right you can also buy an adjustable reel for the wire you cut off. That makes handling the long lengths much easier, especially when cutting numerous long radials.

Safety Issue

While end feeding the elements has many advantages, it presents a safety hazard because the fed ends are so close to ground level, where someone might be able to touch them. The voltages on the lower ends of the wires are very high while transmitting at high power. Some form of guard fence or safety screen is advisable if there is any likelihood of people or animals coming in contact with the wires.

Modeling Comments

Throughout this discussion I have emphasized the need for careful modeling. In my case I have no tower in the middle of things but most installations are likely to have one with HF Yagis attached. I began modeling a tower by obtaining an antenna file from Al Christman, K3LC (ex-KB8I), for a Rohn 55 tower, which models essentially every strut in the tower. The normal thing to do is to calculate the self-resonant frequency of the tower and then model it using a single wire with a diameter that results in the same resonant frequency. You can then use the simpler model in the overall antenna model. I found that I could find such an equivalent wire but the variation in feed-point impedance around resonance was not the same as for the tower. I got a better match in impedance characteristics by adjusting both the diameter *and* the height.

Using this equivalent model I then modeled George's antenna system. I found that his tower did not interact very much with his array. However, that represents a sample of one. It is perfectly possible that another tower, with a different collection of HF Yagis on it, might interact strongly and greatly modify the behavior. This has to be dealt with on a case-by-case basis for each installation.

The W2VJN Antenna

W2VJN has built a number of K1WA arrays over the years. When George was living in New Jersey, he had used a K1WA configuration on 80 meters with very good results. He began with a one-element sloper, then added another element and finally went to four elements. After moving out to the wilderness in Oregon he erected a 150-foot Rohn 55 tower with an array of HF Yagis on it. About 2½ years ago he put two elements on 160 meters and then a year later added two more elements. The array drives one element at a time, with the remaining three acting as reflectors.

The original K1WA array used λ/2 elements, with the length of the coax cable going to the switch box tuned to make the element a reflector when not driven. The result is that the match at the drive element is not all that good for a 50-Ω line—SWR is typically on the order of 1.8:1 or so. George has a variation that improves the match. The elements are cut for 1.850 MHz (by calculation, $L=492/f_{MHz}$). With three elements open and acting as reflectors, the apparent resonant frequency measured at the switch box is 1.770 MHz. This means that at 1.830 MHz there is an inductive reactance at the feed point in the switch box. This is tuned out with a 1200 pF capacitor, resulting in a much better match, close to 1:1 at 1.830 MHz.

George's array can be switched in four different directions and he uses it for receiving as well as transmitting. He has found that selecting the right direction can make a considerable difference in some cases. He has found his antenna to be very effective on receive despite (or perhaps because of) the mix of vertical and horizontal polarization.

Future Improvements at N6LF

While the present array works very well, there is more that I can do. One idea is to add directors. At N6LF I have three tall poles in a row that would allow me to hang director elements for increased gain. I have already done this accidentally. After finishing with the 160-meter array I put up an 80-meter dipole on the east side of the 160-meter array, suspended between the poles that support the 160-meter array, as shown in Fig 12.

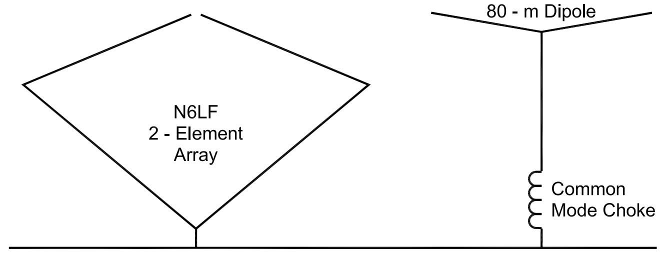

Fig 12—Combination of the N6LF array and an 80-meter dipole for modeling.

I checked to see if the 80-meter dipole had any effect on the 160-meter array by modeling the combination. It certainly did have an effect! With the 80-meter feed line grounded, the 80-meter dipole acted like a reflector and killed my gain to the east. Adding a coax common-mode choke balun turned the dipole into a director and this increased the gain to the east. The dipole is not a very reliable director, however, because as the wind blew it moved up and down, changing its characteristics. One minute it might be a director but a reflector at another. For now I drop the 80-meter dipole for contests or if I think there is the possibility of an opening to Europe. Next summer I plan to make other arrangements for the 80-meter antenna so that it does not interact with the 160-meter array.

And of course a three-element Yagi would have more gain and better F/B than a two-element Yagi. Next summer I will expand the array to three elements. I had originally planned to suspend the directors between the other available poles but after looking at the Spitfire antenna[7] I changed my mind. Since I already have an extensive ground system in place, it makes more sense for me to simply hoist a wire up along the supporting pole and use it as a driven element, and then use the other two elements as director/reflectors. I could even suspend a second director between the supports and go to a four-element (or even five-element) Yagi on 160-meters. Of course, the beamwidth will narrow and I would have to go to at least four-direction switching for the pattern to have reasonable coverage.

Although I use only two elements that allow me to switch the pattern from east to west, the present array has been very useful. Going to three elements (one driven and two reflectors) would be worthwhile. The gain is changed very little by having two reflectors but there is some improvement in F/B. The real improvement would be the ability to slew the pattern in three different directions rather than two. It is possible to have two driven elements and have one as a reflector. This in combination with one driven and two reflectors would give six headings for the pattern. I am not convinced that this would be worth the trouble, however.

Acknowledgements

Much of the progress I made in this project was the result of endless discussions with George Cutsogeorge, W2VJN. He was a tremendous help. I would also like to thank Dean Straw, N6BV, for putting me onto the antenna file for the Rohn tower and to Al Christman, K3LC, for sending it to me. This was very helpful in evaluating the effect of the tower on George's array and sloper arrays in general.

I have referenced only a few of the multitude of articles on sloper antennas. George told me to go to Google. com and enter "sloper arrays". I got over 500 references, about two thirds of which were for sloper antennas! No doubt everything said here has already been said many times but perhaps this forum will reach a wider audience.

John Devoldere's *Low-Band DXing*[8] is another great reference source. At the last moment in preparing this article John, EI7BA, made me aware of a particularly good article by Tony Preedy, G3LNP.[9] It deals with bent verticals less than λ/2wavelength long and arrays made from them, which are very much like the arrays in this article. This is also a "must read" article.

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5 R. Severns, N6LF, "Conductors for HF Antennas," *QEX*, Nov/Dec 2000, pp 20-29.

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7 Kaufmann, W1FV, and Hopengarten, K1VR, "The FVR Spitfire Array," Dayton Hamfest presentation, 1998. Copies of this can be found on the Yankee Clipper Contest Club Web site; **www.4CCC.org/Articles/Spitfire/spitfire.html**.

8 J. Devoldere, *Low-Band DXing*, 3rd Ed. (Newington, ARRL: 1999).

9 T. Preedy, G3LNP, "Single Support Directional Wires," *RADCOM*, Aug and Sep 1997.

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Experimental Determination of Ground System Performance for HF Verticals Part I Test Setup and Instrumentation

This description of the test setup used by the author for a series of experiments sets the stage for a series of articles describing his results.

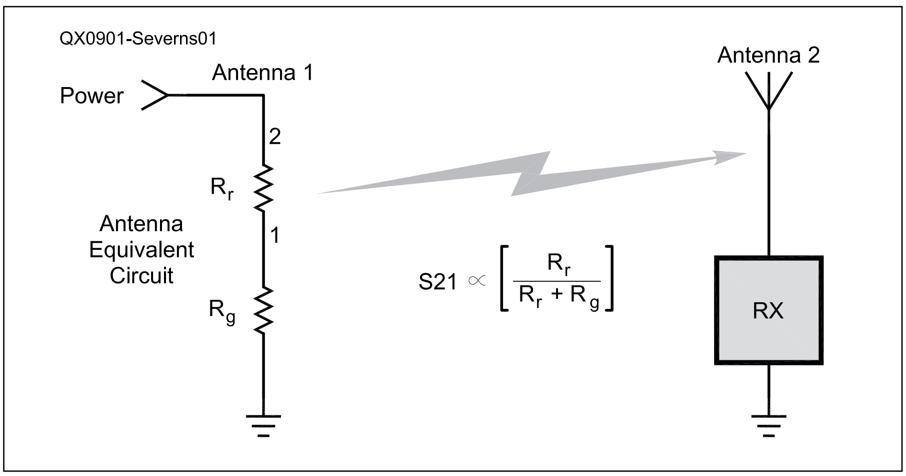

Figure 1—This drawing illustrates the traditional measurement scheme.

HF verticals located on or near ground are a perennial topic among amateurs. Over the past several years this discussion has been illuminated (and in some cases obscured!) by the advent of really good modeling software based on NEC (numerical electromagnetic code). This has resulted in a vast literature on antennas using the results of modeling. However, these results are not without some controversy. In particular the relative merits of a large number of buried radials versus a few elevated radials has been especially contentious. What has been missing from the discussion are careful field measurements done with good instrumentation and technique to see if the NEC predictions actually hold up in the real world. To address this problem I performed a series of field experiments, over a period of a year, to examine how different ground system arrangements affected the behavior of a vertical antenna and to see if field measurements on a real antenna would correlate with NEC modeling.

The results of these experiments will be presented in a series of *QEX* articles. There is no pretence that these experiments will answer all questions or even definitively settle some of the arguments, but at least they should give us something to think about.

In Part 1, I will discuss the test range, test instrumentation and test procedures used for all the experiments. Part 2, which is included in this issue of *QEX*, discusses an earlier and apparently overlooked prediction from NEC, that in sparse (<10 radials) radial systems lying close to ground, there can be a substantial increase in ground loss when the radials are made much longer than ⅛ wavelength. This is a case of more copper = more loss, which is not at all intuitive! Part 3 will compare verticals with a large number of ground surface radials to verticals with four elevated radials. This part will directly address the elevated radial controversy. Part 3 will also have comparisons between several different elevated radial configurations. Part 4 will look at the effect of radial numbers on the characteristics of ¼ wavelength and several shorter loaded antennas. Part 5 will take a look at the problems of ground systems for multiband verticals, where a range of 7 to 30 MHz must be accommodated. Finally in Part 6, I will report on some experiments with a full size ¼ wavelength vertical on 160 m. In addition, because this series will take many months to be published, there will be lots of time for feedback. I plan to include some of this in Part 6.

Test Setup

The physical layout of the test range, the instrumentation employed and the test procedures were all key elements in obtaining reliable results. The following discussion provides descriptions of these elements which remained essentially constant for the experiments. The majority of measurements were done at 7.2 MHz although there was some work at 160, 30, 20, 17, 15 and 10 meters. The information given here is intended to provide information common to all the experiments.

Test Concept

The traditional test procedure for these kinds of measurements is well known. As shown in Figure 1, a test antenna is excited with a known power, and the resulting signal is measured at a remote point. A change is then made in the test antenna and the measurement is repeated. The difference between the two measurements is a measure of the effect of the change in the antenna and/or ground system on performance. The signal transmission to antenna 2 from the excitation of antenna 1 (S21) will be proportional to the radiation efficiency of the antenna. In other words, S21 ~ input power × Rr / (Rr + Rg) where Rr is the radiation resistance and Rg is the ground loss. For our purposes we can assume that losses due to conductors are small. Both Rr and Rg will vary as we change the ground system but the final goal is to see the effect on the transmitted signal.[1]

The standard way to make these measurements is to use a transmitter combined with forward and reflected power meters to excite the test antenna (antenna 1) with a known power. A calibrated receiver is connected to a remote receiving antenna (antenna 2) to measure the resulting signal. In my initial tests I used both an HP3586C and an HP3585A spectrum analyzer for the receiver. I wished to measure the performance differences between configurations to within 0.1 dB if possible, and these instruments were capable of that. However, the limiting factor turned out to be my ability to measure the excitation power; 0.1 dB corresponds to about 2%. To make repeatable measurements to 0.1 dB you would need to measure power to better than 1%.

To get around that problem I decided to use the instrumentation scheme illustrated in Figure 2. I chose to make the measurements with a vector network analyzer (VNA) in the transmission mode (S21 is the response at port 2 due to the excitation at port 1). The transmission path was from the VNA output port, out to the test antenna via a transmission line, from there to the receive antenna and back to the VNA input port via another transmission line.

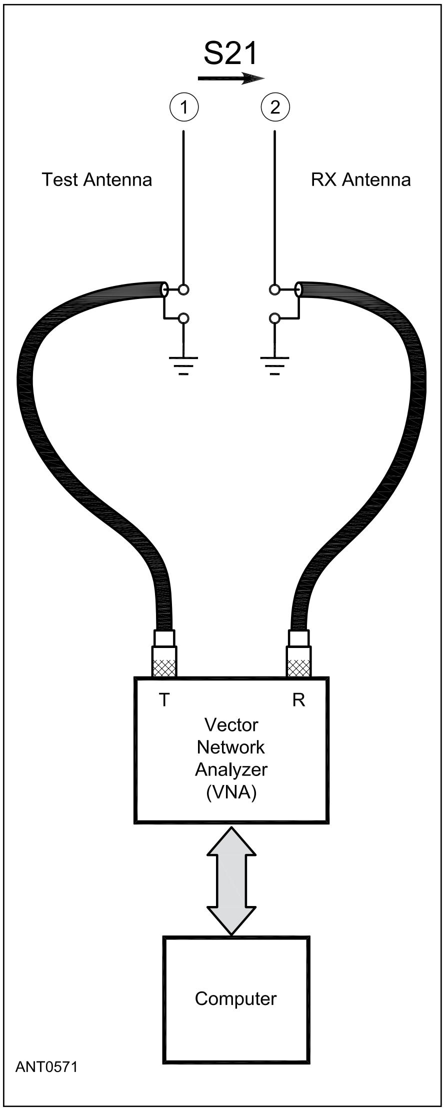

Figure 2 — This diagram shows the vector network analyzer approach for measuring antenna performance.

Amplitude measurements with a professional VNA are typically displayed to 0.001 dB, but of course nothing else in the system is stable to that level. In practice I found that measurements made over a short period of time (2-3 hours) were repeatable to within 0.05 dB. That is more than adequate for these experiments. A weakness of this measurement method is that as the separation between the test antenna and the receiving antenna is increased, the attenuation around the transmission loop becomes quite large, –40 to –60 dB. For instrumentation and a physical setup with a noise floor and stray coupling below –110 dBm, this is acceptable but it did limit the separation distance on 40 m to about 2.25 wavelengths for the particular receiving antenna employed. This is in the far field but not by much. Another limitation was that ± 0.05 dB repeatability was possible only when the antenna under test and the receive antennas were actually stable to that level. This usually meant that measurements had to be made in early morning when the test range was in the shade or late in the day when things had reached thermal equilibrium. It was very easy to detect a cloud passing over by the small changes due to temperature changes in the antennas. I could readily detect the effect of the wind on the vertical, causing it to move slightly. In the end the A-B comparison measurements were probably within a few tenths of a dB but only when I carefully attended to all the details.

This brings us to an important point. The purpose of the experiments was to determine the effect of different ground system arrangements from their effect on S21. All the measurements were relative A-B comparisons. In other words, they were comparisons between two different configurations. There was no

Figure 3— A view of the test antenna area and test equipment shelter. The receiving antenna is at the far end of the pasture.

[1]Notes appear on page 25.

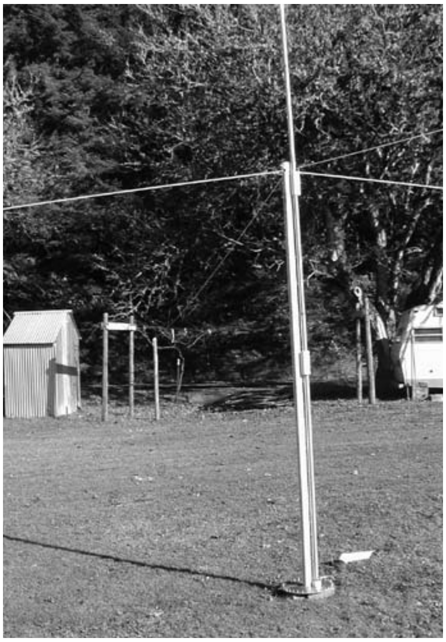

Figure 4 — This photo shows a typical test antenna and center post support.

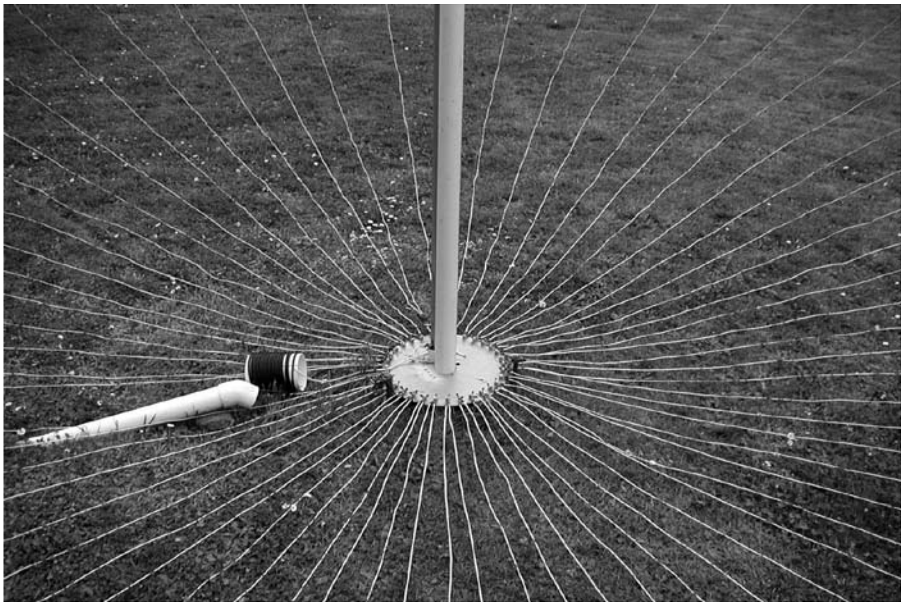

Figure 5 — Here is the test antenna base at ground level, with 64 radials.

Figure 6 — The base plate is in position for elevated radials.

attempt to measure absolute signal strengths or radiation patterns. The separation distance between the test antenna and the receiving antenna was sufficient to place the receiving antenna outside the reactive near field but the groundwave was still significant. This was not a problem for the type of measurements being made. The presence of a metal pump house and a travel trailer, both of which are small in terms of a wavelength might have had an impact on pattern measurements but should not have affected the type of A-B measurements being made in this series of experiments.

Physical Arrangement

The test range was set up in a field as shown in Figure 3, with an area for the test antennas (including ground systems), a remote receiving antenna (in the far distance) and a small travel trailer to provide shelter for the instrumentation.

The eight poles, in an 80 foot diameter circle around the test antenna, were used to support elevated radials as needed. When more than eight elevated radials were needed, a ½ inch Dacron line was stretched around the posts at the desired height and tightened with a turn-buckle. Each post has a backstay to a buried deadman anchor so the radials could be well tensioned. Radial heights on each post were located using a laser level to keep the radial fan flat around the circle.

In the center of the circle there is a support post (PVC pipe) as shown in Figure 4, with Dacron support lines attached to the top. This post is intended to hold the antenna under test and allow it to move up and down to vary the height for elevated radial tests. An example of the base plate at ground level with 64 radials attached is shown in Figure 5.

The base plate is isolated from ground but there are three ground stakes (4 foot copper-clad steel rods) close to the plate for those tests where grounding is desired. The ground stakes have short pig-tail leads to connect to the base plate when desired.

Figure 6 shows an example of the base plate positioned for elevated radial tests. The base plate, the radials and the entire test antenna are elevated by sliding them along the support pipe. This arrangement made it very easy to change the height of the radials in small increments up to 4½ feet above ground. The radials lying on the ground in Figure 6 were *not* present during elevated radial tests!

As shown in Figures 5 and 6, a coaxial common mode choke (balun) was used to isolate the transmission line from the test antenna. This was done for all measurements whether or not ground stakes were engaged. The choke has an impedance of >3 kΩ at 7.2 MHz. For those tests in which the SteppIR vertical was employed, the balun that comes with that antenna was used in lieu of the choke shown in the photos.

The receiving antenna was a 3-turn diamond loop with a diagonal dimension of

24 inches, as shown in Figure 7. The loop was resonant at 8.2 MHz. This loop was installed at the top of a 40 foot mast, as shown in Figure 8.

The distance from the base of the test antenna to the receiving loop is a little over 300 feet, about 2¼ wavelengths at 7.2 MHz. The elevation angle from the base of the test vertical is about 8°.

The coax from the VNA output port to the base of the test antenna was ½ inch Andrews Heliax with N connectors. The coax from the receiving antenna back to the VNA was LMR400. Low loss coax was used because it provided better shield attenuation to reduce coupling and in the case of the heliax running out to the test antenna, the very low loss removed the need for an additional correction factor for the change in cable loss with variations in SWR.

Test Instrumentation

Feed point impedance, transmission gain (S21) and radial current measurements were all made using a VNA. Two analyzers were available: an HP3577A with an HP35677A S-parameter test box and an N2PK analyzer with dual fast detectors. Figures 9 and 10 are photos of these instruments.

Note the organic automatic heating unit on top! Critical for maximum accuracy! The common mode choke in the photo is undergoing characterization for transmission loss and series impedance at 7.2 MHz. It turned out however, that the impedance of the choke was much greater than the 50 Ω reference impedance of the VNA. Above about 2 kΩ even an HP VNA becomes inaccurate for a direct measurement. For choke measurements, I used an HP4815 analyzer, which is well suited for high-impedance measurements.

After careful comparisons between the HP and N2PK VNAs, the N2PK was selected for

Figure 7 — This photo shows the loop receiving antenna.

Figure 8 — Here is the receiving antenna atop a 40 foot mast. N7MQ assisting!

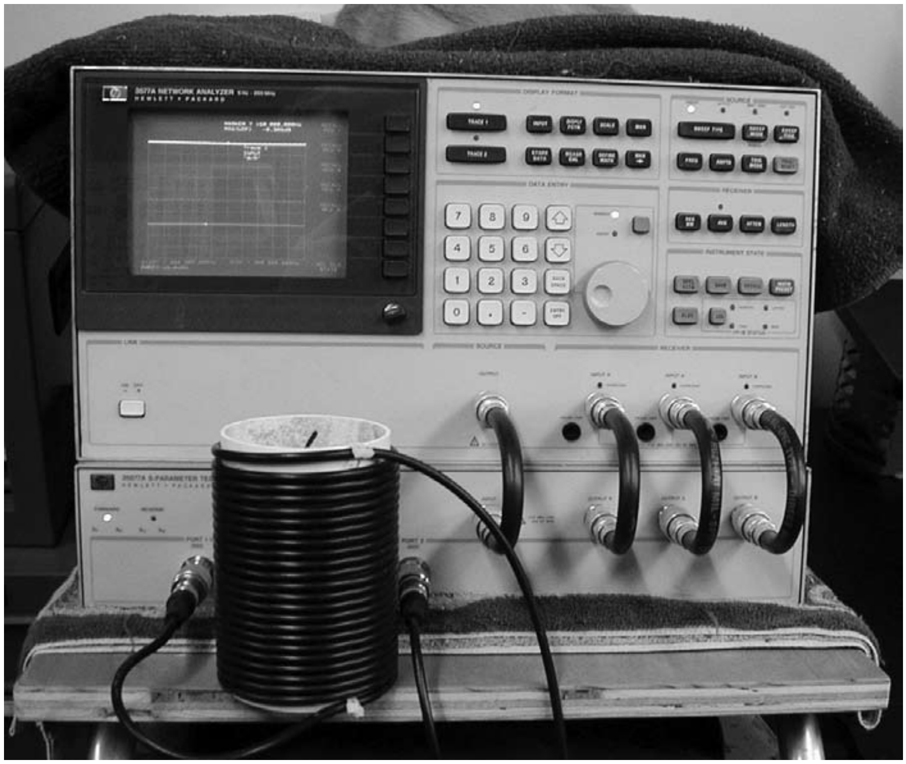

Figure 9 — HP3577A with an HP35677A S-parameter test box.

Figure 10 — Here is my test bench, showing the N2PK VNA with the associated laptop computer and HP calibration loads.

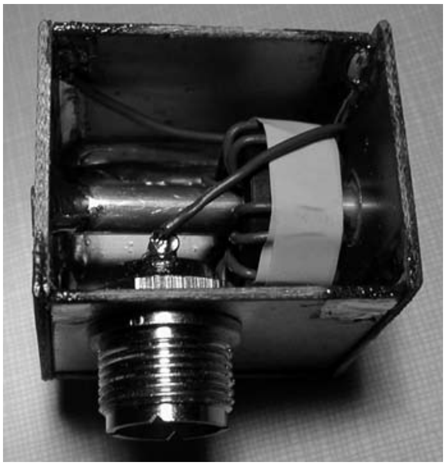

Figure 11 — This photo shows a typical shielded current transformer.

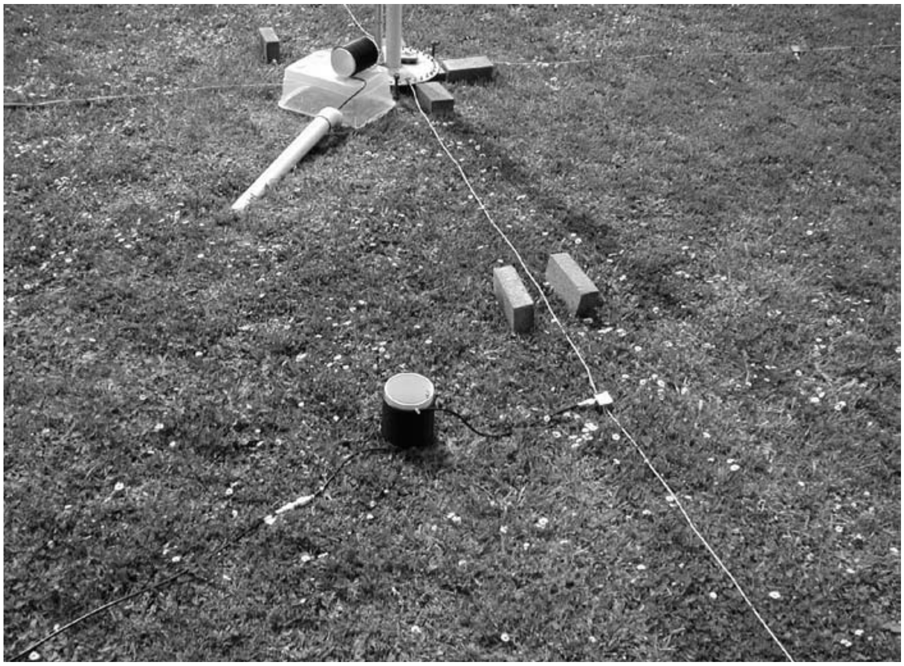

Figure 12 — Here is the test setup for a typical radial current measurement.

most of the measurements because its performance was very close to the HP and had the advantage of direct readout to a computer, which made data reduction much easier. The N2PK VNA was also much lighter than the HP (70+ pounds!) and much more suitable for field measurements.

On several occasions it was necessary to measure the current division ratios between the radials and in some cases, the relative current distribution along a radial. To make these measurements a set of shielded current transformers, like the one shown in Figure 11 were used.

To make a current measurement, a radial wire was passed through the current transformer, as shown in Figure 12. Current transformers were placed in the same location simultaneously on all the radials during a measurement. The transformer being used to sense current was terminated in 50 Ω by the instrumentation, so all of the dormant current transformers were also terminated in 50 Ω. This was done to compensate for any interaction introduced by the current transformer. At the very least, the effect of the current transformer would be the same on all radials. The active current transformer was isolated with a choke as shown in Figure 12.

Even with this degree of care, the current measurements were still a bit tricky because of the residual interaction between the cable from the current transformer and nearby radials. In some cases I actually used four identical cables in a symmetrical layout to try to minimize imbalance due to this interaction. I believe the resulting measurements were reasonable and useful but not especially precise!

The relative value of the current was determined by using the VNA in the transmission mode, measuring S21 for the loop from the VNA output port to the base of the antenna, out along the radial to the current transformer and back to the VNA input port. This was a convenient way to measure the current division between radials and the relative current distribution along a radial.

Comments on test procedures

A good physical setup and professional instrumentation are a very good start, but to obtain reliable data great care must be exercised in using and calibrating this equipment. For feed point impedance measurements, at the beginning and end of every test run an OSL (open, short, reference load) calibration was performed with the calibration plane at the test antenna feed point. At the beginning and end of each test run a transmission calibration was also performed.

In addition, before beginning a series of measurements a measurement of stray coupling and possible interference was performed. The procedure was to disconnect the feed line from the base of the test antenna, terminate the feed line with a 50 Ω load and then measure the transmission gain of the entire system in this state. Throughout the series of experiments, this transmission level was never higher than –110 dBm and usually –115 dBm or lower, at 7.2 MHz. As a further check on results, most experiments were run several times to verify consistency and repeatability. All of this was very time consuming but absolutely necessary to assure the best possible measurements. I did not delude myself, however, into thinking the measurements were perfect and cannot be improved on. I do believe the results make sense, fit well with NEC modeling predictions, give useful insights into vertical antenna/ground system behavior, and potentially can be of practical help in optimizing a given antenna installation.

Acknowledgement

This experimental work was inspired by the earlier work of Jerry Sevick and Arch Doty.[2, 3] Some of my experiments were a repeat of their earlier work with more advanced instrumentation. I would also like to thank Mark Perrin, N7MQ and Paul Thompson, W8IEB for the many hours of help they provided during the experiments. Without their help, I would still be out in the field taking measurements!

Notes

[1]R. Severns, N6LF, "Radiation Resistance Variation With Radial System Design," Sep 2008. Available at: **antennasbyn6lf.com**.

[2]J. Sevick, W2FMI, *The Short Vertical Antenna and Ground Radial*, CQ Communications, Inc, 2003. Jerry's work also has appeared in a number of *QST* articles.

[3]A. Doty, K8CFU, "Improving Vertical Antenna Efficiency, A Study of Radial Wire Ground Systems," *CQ Magazine*, April 1984, pp 24-31. This article also has a very nice list of earlier references related to ground systems for verticals.

Rudy Severns, N6LF, was first licensed as WN7WAG in 1954 and has held an Extra class license since 1959. He is a consultant in the design of power electronics, magnetic components and power-conversion equipment. Rudy holds a BSE degree from the University of California at Los Angeles. He is the author of two books and over 80 technical papers. Rudy is an ARRL Life Member, and also an IEEE Fellow.

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Experimental Determination of Ground System Performance for HF Verticals Part 2 Excessive Loss in Sparse Radial Screens

These experimental results may surprise you, and might turn "conventional wisdom" upside down.

In 1998, Jack Belrose, VE2CV, used *NEC* modeling to show the effect of resonant and non-resonant radials placed very close to the ground surface on the behavior of a ¼ wavelength vertical.[1] One of the observations in that article was that the use of a small number of ¼ wavelength (free space) radials, lying on the ground surface, could lead to much higher losses than expected, and that shortening the radials could actually reduce ground loss. This seems counter to more classical analyses which show that making radials too long may be a waste of wire but does no harm. The classic analysis, however, does not take into account the possibility of resonances in the radial screen that might amplify the radial current, increasing ground loss.

The purpose of this experiment was to see if a real antenna would actually demonstrate the predicted behavior, and validate the *NEC* predictions experimentally.

Description of the Experiment

The experiment was done in six parts spread over a three week period:

1) The antenna for part 1 was a telescoping aluminum-tubing vertical, averaging 1 inch in diameter, with a fixed height of 34 feet. The test frequency was 7.2 MHz. I used four no. 18 insulated wire radials lying on the ground surface. All four radials were of equal length, which was varied from 33 feet down to 18 feet. The impedance at the feed point, the transmission gain (S21) and the current division ratios between the radials were measured and recorded. The antenna and radials were isolated from ground and the feed line with a common mode choke.

2) For part 2, part 1 was repeated, first isolated from ground and then with one or more ground stakes connected, to evaluate the effect of using ground stakes at the base of the antenna. Tests were also done without any radials, and with just 1, 2 or 3 ground stakes connected to the base plate.

3) Part 3 of the experiment was the same as part 1 except with 8 radials (no ground stakes).

4) For part 4, the antenna was changed from the fixed tubing vertical to a remotely adjustable SteppIR vertical. In parts 1, 2 and 3, the antenna height was kept constant at 34 feet, but in this part of the experiment the height was changed to re-resonate the antenna as the radial number and radial lengths were changed. The test frequency was 7.2 MHz.

5) After completing the first four parts of the experiment it was clear that shortening the radials from the standard free space ¼ wavelength value did indeed improve the signal, at least in the case of 4 and 8 radials, so I wanted to see what the effect was for 16 and 32 radials. Trimming that many radials to gradually shorten them, however, was a bit more work and wasted wire than I was prepared for. Instead, I ran this part of the experiment first with 4, 8, 16 and 32, thirty-three foot radials, which I had on hand, and then with 4 ,8, 16 and 32, twenty-one foot radials, which were also on hand. This gave me two data points for each number of radials. Again, the test frequency was 7.2 MHz, with measurements of S21 and feed-point impedance.

6) Part 6 of the experiment was a check to see if the same kind of improvement would be seen at 30, 20 and 15 m by shortening the radials from ¼ wavelength (free space). This part of the experiment was not nearly as thorough as the first five parts but did confirm that the same basic behavior was present at the higher frequencies as that seen on 40 m. The test frequencies were 10.120 MHz, 14.200 MHz and 21.200 MHz.

[1]Notes appear on page 52.

Experimental Results

Part 1

Figure 1 shows the variation in |S21| (magnitude of the transmission gain) as a function of radial length. The amplitude scale is normalized to 0 dB for a radial length of 33 feet, which is approximately a ¼ wavelength in free space at 7.2 MHz. The Y-axis shows the improvement in dB as the radials are shortened.

The improvement is quite large, about 2.8 dB, which would have a noticeable effect on signal strength. In Belrose's paper the improvement was about 3.5 dB but that was for average soil. My average ground characteristics are *approximately* $\sigma = 0.015$ S/m and $\varepsilon_r = 30$, which is quite a bit better than average ground. These values were derived from ground probe measurements.[2] One would expect more improvement for poorer soil.

An earlier experiment in which the current distribution on a 33 foot radial, at 7.2 MHz, was measured, gave the results shown in Figure 2.

A quick check was made during the present experiment, and the current distribution appeared to be essentially the same. From the current distribution we can see that the radial in Figure 2 is resonant well below 7.2 MHz. To move the current maxima back to the base of the vertical we would have to reduce the radial length by about 10 feet. Looking back at Figure 1, we see that we are very close to the maximum |S21| when the length has been reduced by 10 feet to 23 feet. What appears to be happening is that we are tuning the radials to resonance (or at least close to it) at 7.2 MHz to compensate for the loading effect of the soil in close proximity to the radial wire.

The division of current between the radials was measured for 18 foot and 33 foot lengths. Table 1 shows the results. The current division was quite uniform and the differences too small to have significant effect on the observed gain changes.

Table 1
Current Division Between Radials Normalized to 1 A of Total Base Current.

Radial Number	*I_n, 33-Foot Radials (A)*	*I_n, 18-Foot Radials (A)*
1	0.24	0.26
2	0.24	0.25
3	0.25	0.25
4	0.27	0.24

The variation of feed-point impedance as the radial lengths were shortened (with the vertical height constant at 34 feet) is shown in Table 2.

Table 2
Measured Feed Point Impedances

Radial length (ft)	*Feed Point Impedance (Ω)*
33	135 + *j* 28
30	108 + *j* 55
27	83 + *j* 51
24	67 + *j* 37
21	60 + *j* 22
18	57 + *j* 8

Parts 2 and 3

Part 1 was done during a week of heavy rain. Parts 2 and 3 were performed 8 days after part 1, when the soil had drained and dried out significantly so the ground characteristics may have changed somewhat.

The next step in the experiment was to expand the radial count from 4 to 8 radials and also to investigate the effect of using grounding stakes (4 foot copper clad steel rods) connected at the base of the antenna. Measurements with 4 and 8 radials were repeated in each run. This run was with a fixed height for the vertical (34 feet). The results are shown in Figure 3.

At all lengths, 8 radials are an improvement over 4. With 8 radials, the amount of improvement with radial shortening is smaller but still useful. We can also see that adding a ground stake in the case of 4 radials also makes a substantial improvement but we should keep in mind that my soil would be classified as "very good" so we would expect ground stakes to be more effective than they would be in poorer soil.

The results for the case of no radials and 1, 2 or 3 ground stakes, normalized to the cases of four 33 foot radials and four 21 foot radials, with no ground stakes, are given in Table 3. Vertical height was constant at 34 feet.

Part 4

In part 4 I changed to the SteppIR vertical and adjusted the height to re-resonate the vertical for each radial length. The results are shown in Figure 4, which are very similar to the results for constant height given in Figure 3. No ground stakes were employed.

Part 5

From the earlier test results, I could see that the improvement due to radial shortening decreased as the number of radials increased.

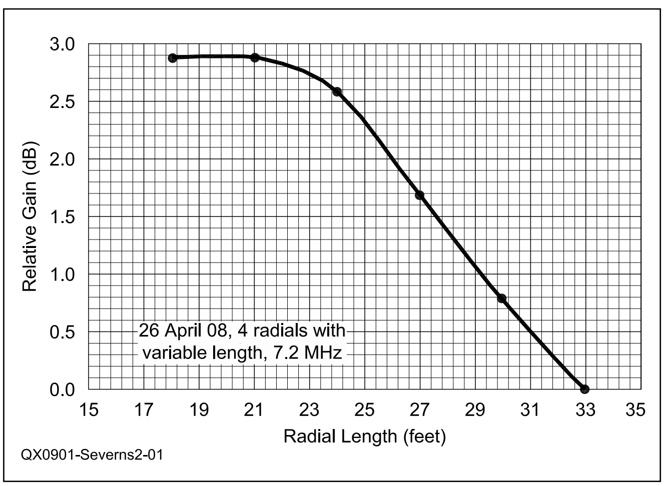

Figure 1 — This graph shows the improvement in |S21| as the radials were shortened. There were four radials lying on the ground surface.

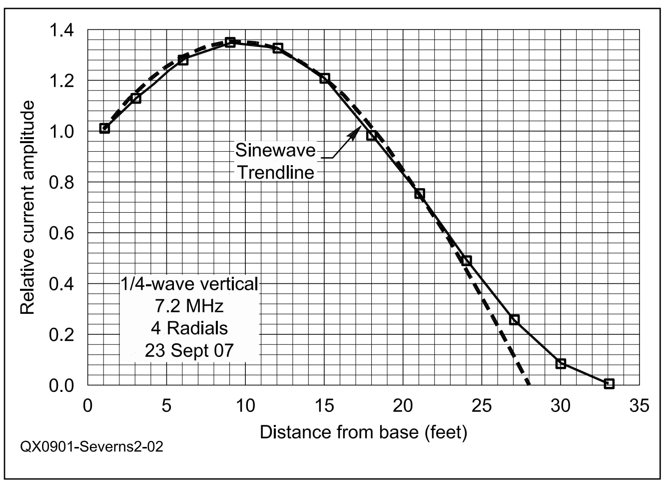

Figure 2 — This graph shows the relative current amplitude along a radial.

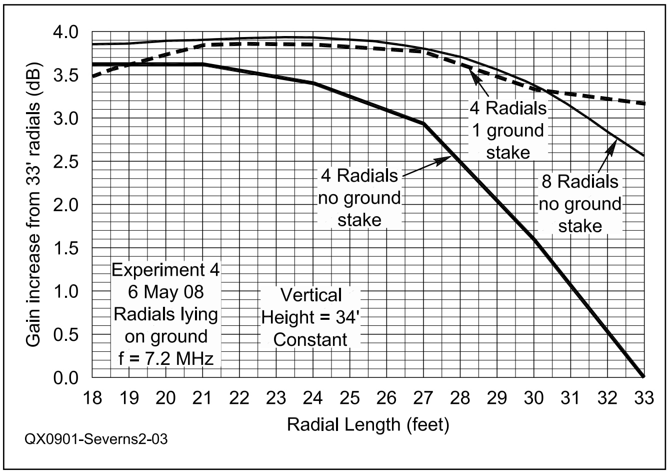

Figure 3 — This graph shows the change in |S21| with radial length. The vertical antenna height was a constant 34 feet.

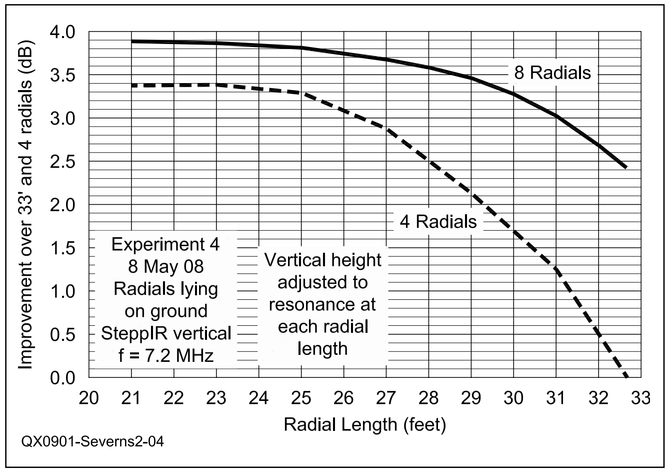

Figure 4 — This graph shows the change in |S21| with radial length. I adjusted the SteppIR antenna height to resonance for each radial length.

Table 3
Test Results for no Radials and 1, 2 or 3 Stakes, Compared to 4 Radials with no Ground Stakes.

Number of Stakes	*Feed Point Z (Ω)*	*Compared to Four 33-Foot Radials, No Ground Stakes (dB)*	*Compared to Four 21-Foot Radials No Ground Stakes (dB)*
1	77 + *j* 40	2.67	–0.95
2	69 + *j* 30	3.09	–0.53
3	66 + *j* 26	3.25	–0.37

Table 4
Results for 4, 8, 16 and 32 Radials, with Lengths of 33 Feet and 21 feet.

Number	*33-Foot Radials Feed Point Impedance (Ω)*	*21-Foot Radials Feed Point Impedance (Ω)*	*33-Foot Radials |S21| Relative to Four 33-Foot Radials (dB)*	*21-Foot Radials |S21| Relative to Four 33-Foot Radials (dB)*	*Delta Gain Change (dB)*
4	89.8	52.5	0	3.08	+3.08
8	51.8	45.6	2.26	3.68	+1.42
16	40.5	42.8	3.76	3.95	+0.19
32	37.7	41.6	4.16	4.04	–0.12

In this part of the experiment the number of radials was extended to include 16 and 32 radials to quantify that difference. The test was conducted with sets of 4, 8, 16 and 32 thirty-three foot radials, and then repeated with the same numbers of 21-foot radials. The StepIR antenna was used, and its height was adjusted to re-resonate as the radials were altered. The results are tabulated in Table 4. These measurements were made several days after those used in Figure 4, so there are some differences because of small changes in the ground characteristics, radial layout, and other conditions. These day-to-day variations are a major reason for repeating some parts of earlier experiments multiple times and trying to do a complete experiment in a short period of time (a couple of hours).

It should be noted that a ground system consisting of only four radials is really flaky. Measurements vary significantly with small variations in radial layout, changes in soil moisture, placement of the feed line relative to the radials, and so on. Shortening the radials does seem to reduce this sensitivity, but even so, a four radial system should only be an emergency measure.

As expected, as the number of radials is increased the change due to radial shortening gets much smaller. Over the very good ground on which these measurements were made, shortening the radials gave only a modest advantage when more than 8 radials were used. Over poorer soils, however, radial shortening with 16 radials might be worth doing. The lower value for feed point impedance (Z_i) with 33-foot radials is at least in part due to the shorter height needed to resonate. For 21-foot radials the height had to be increased to re-resonate the antenna.

It is interesting to note that with 32 radials, the 33-foot radials were actually slightly better (0.12 dB) than 21-foot radials. Quite probably there was some optimum length in-between that may have been slightly higher than either, but that is not likely to be very large and I decided it wasn't worth the trouble to cut up a set of 32 radials to find out. The important point is that the changes in gain, input impedance and height variation to re-resonate all get much smaller when more radials are used. I would think that with 32 or more radials you wouldn't worry about resonances in the radial screen. The problem is only important when fewer than 16 radials are deployed.

Table 5 shows the antenna height (h) in inches. This is the reading from the control box. The actual height is about 12 inches longer due to the height above ground of the

Table 5
Indicated Height of the Vertical.

Number of Radials	*33-Foot Radials h (inches)*	*21-Foot Radials h (inches)*
4	357	381
8	366	382
16	374	382
32	377	382

Table 6
30 m, ¼ Wavelength Free Space = 24.3 Feet.

Radial Length (ft)	Z_i (Ω)	*|S21| (dB)*	*h (in)*
21	44.4	–62.31	260
20	41.6	–61.12	261
18	41.0	–61.84	264
16	42.6	–61.78	267

reel and the lengths of connecting wires, plus the length of radials from the reel box to ground surface. The columns for h do, however, give an idea of the change in height. In the case of 33-foot radials the change is quite large (20 inches) between 4 and 32 radials. On the other hand with 21-foot radials the change in h with radial number is very small, factions of an inch. The values in the Table are rounded off to the nearest inch.

Part 6

In the final part of this experiment the effect of radial shortening on 30, 20 and 15 m was examined. This was really just a quick look using radials left over from the earlier parts of the experiment, cut down from them rather than making up a new set of ¼ wavelength (free space) radials for each band. *In all three cases 8 radials were used*. The test frequencies were: 10.120 MHz, 14.200 MHz and 21.200 MHz. The corresponding free space ¼ wavelengths would have been, 24.3 feet, 17. 3 feet and 11.6 feet respectively. The results are shown in Tables 6, 7 and 8. The value for |S21| is the actual measurement.

One oddity in this data was that the best radial length on both 30 and 20 m was the same, about 15 feet. There is some dispersion (variation with frequency) in the soil characteristics but I don't think that's a full explanation. In all cases the optimum length was well short of the free space ¼ wavelength. I think this part of the experiment needs to be rerun cutting down from full length radials. This will be done at some future time.

NEC Modeling

At this point it was clear that Belrose's original work was basically confirmed experimentally, but I was curious to see how closely this data could be replicated using *NEC4-D* modeling software (*EZNEC Pro + MultiNEC*). The first trial model employed 4 radials with lengths from 6.4 m (21 feet) to 10 m (33 feet). The wire table for this model is given in Table 9. The radials were placed 5 mm above 0.01/14 soil. The test frequency was 7.2 MHz and the vertical height was adjusted to maintain resonance as the radial number was changed.

We can compare the maximum gain data against the experimental data for 4 radials (from Figure 4) as shown in Figure 5.

The match in gain data is very good, as was the current distribution on the radials. The impedance data was also close. We can also see what *NEC* predicts about the current distribution on a radial as we change the length. Figure 6 shows the current distribution on a 33-foot radial for *NEC* model 1.

Figure 6 looks very similar to the experimental measurement shown in Figure 2. When we shorten the radials to 21 feet, we get the current distribution shown in Figure 7. This is very close to resonance.

The match in gain and current distribution, however, is really too good to be believed. First of all, this is not an exact model of the real antenna. The vertical uses a strip of beryllium-copper, not a no. 12 wire, and I believe my ground characteristic is better than the 0.01/14 used in the model. Models with wires very close to the ground sur-

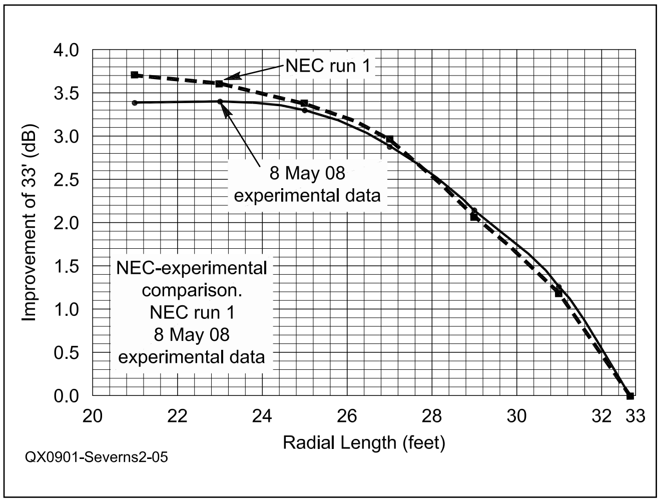

Figure 5 — Here is a comparison between *NEC* modeling run 1 and the experimental data using 4 radials taken on May 8, 2008.

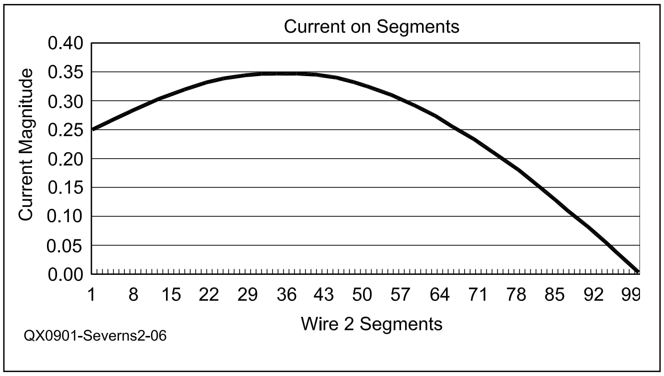

Figure 6 — This graph shows the current distribution on a 33 foot radial (*NEC* model).

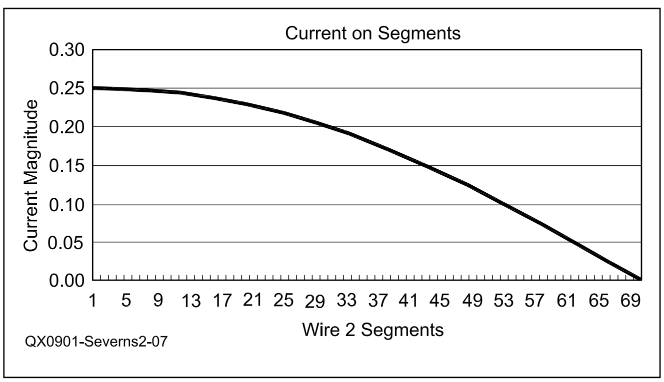

Figure 7 — Here is the current distribution on a 21 foot radial (*NEC* model).

Table 7
20 m, ¼ Wavelength Free Space = 17.3 Feet.

Radial Length [ft]	*Z_i (Ω)*	*|S21| (dB)*	*h (in)*
16	37.8	–62.03	178
15	36.0	–61.84	179
14	35.0	–61.91	181

Table 8
15 m, ¼ Wavelength Free Space = 11.6 Feet.

Radial Length [ft]	*Z_i (Ω)*	*|S21| (dB)*	*h (in)*
9	27.3	–60.34	60
8	30.0	–60.29	60
7	34.3	–60.11	60
6	41.0	–60.46	60

Table 9
Model Wire Table

End 1			***End 2***			***Diameter***	***Segs***	*Show Lengths in ● m O wl*		
X (m)	*Y (m)*	*Z (m)*	*X (m)*	*Y (m)*	*Z (m)*	*(mm or #)*	*(359)*	*Wire*	*Length*	*Seg Len*
40 m gp 4 rad A										
0.000	0.000	0.005	0.000	0.000	10.306	#12	103	W1	10.301	0.100
0.000	0.000	0.005	6.400	0.000	0.005	#12	64	W2	6.400	0.100
0.000	0.000	0.005	0.000	6.400	0.005	#12	64	W3	6.400	0.100
0.000	0.000	0.005	-6.400	0.000	0.005	#12	64			
0.000	0.000	0.005	0.000	–6.400	0.005	#12	64			

Table 10
Z_i and Peak Gain

Freq (MHz)	*L*	*M*	*R at Src1*	*X at Src1*	*SWR(50 Ω)*	*Max Gain*
7.200	9.056	10	83.15	0.03	1.663	–4.41
7.200	9.275	9.45	65.72	0.01	1.314	–3.22
7.200	9.535	8.84	54.59	0.00	1.092	–2.12
7.200	9.757	8.23	49.83	–0.01	1.003	–1.45
7.200	9.955	7.62	48.23	–0.02	1.037	–1.04
7.200	10.136	7.01	48.48	0.01	1.031	–0.81
7.200	10.306	6.4	49.91	–0.02	1.002	–0.70

Where L is the height of the vertical in meters and M is the length of the radials in meters.

face are very sensitive to small changes in the model and wire segmentation. A change in height as small as 1 mm when the wires are at 5 mm above ground, makes a very substantial change in the results. By diddling the model, I can get the kind of match shown in Figure 5, but when I go the other way and attempt to use the model to predict the behavior of the real antenna, the results could be way off. When it comes to wires very close to ground — distances comparable to the wire diameter — *NEC* replicates the general behavior but you do not know enough of the details of the real antenna and it's immediate environment to expect exact quantitative results from the model.

In addition, the characteristics of real soil vary widely even at a fixed location: vertically, horizontally and over time. The soil will very likely have grass (weeds?) over it, which varies in length and water content during the year. We will seldom have more than a general idea what our ground characteristics are even with ground probe measurements. We will also not really know the height above ground to a fraction of mm! The radials will be buried somewhere in the grass, so who knows what the effective height really is.

Final comments

The effect that showed up initially in Belrose's article and in later *NEC* modeling appears to be real. I think it is clear that in a sparse radial system lying directly on the ground surface, it is possible to incur substantial additional ground losses over what we might expect. The prediction from *NEC* modeling of this effect appears to be confirmed, at least qualitatively. I have been able to reproduce it experimentally multiple times, on multiple bands, with different antennas.

While *NEC* predicts the effect, you can't rely on *NEC* modeling for exact predictions. You will have to do final adjustment in the field. This is not a general indictment of *NEC*. When the antenna has not been right down next to the ground surface, I have found *NEC* predictions to be very good when I went out and built the actual antenna.

We have a couple of ways to attack the problem of radial resonance and excess ground loss: first, cut the radials to be near resonance while lying on the ground. That works if you have the instrumentation, but is hardly a practical approach in general. The second and much more practical approach is to use at least 16, or better yet, 32 radials. As I pointed out earlier, ground systems using only a few radials are a poor idea for many reasons.

Notes

[1]J. Belrose, VE2CV, "Elevated Radial Wire Systems For Vertically Polarized Ground-Plane Type Antennas, part 1 — Monopoles," *Communications Quarterly*, Winter 1998, pp 29-40.

[2]R. Severns, N6LF, "Measurement of Soil Electrical Parameters at HF," ARRL, *QEX*, Nov/Dec 2006, pp 3-9.

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Experimental Determination of Ground System Performance for HF Verticals Part 7 Ground Systems With Missing Sectors

Here is the author's research on radial systems that do not make a full circle around the vertical antenna.

A very common problem with vertical ground systems is the impracticality — in many situations — of laying down a symmetric circle of radials. Some object, frequently a structure or a property line, may make it impossible to place radials in certain areas around or near the base of the antenna. I have received many questions on this subject so I decided to do some experiments where I compared the signal strength (S_{21}) of a ¼ λ vertical antenna that has a full 360° radial fan to one with a substantial portion of the radial fan missing in one sector.

The first part of the experiment was done at four frequencies: 7.2, 14.2, 21.2 and 28.5 MHz. The second part the experiment was done at 7.2 MHz only.

Radial Fan Configurations

For this series of tests I chose to use a symmetric 360° radial fan with thirty two 33 foot radials (¼ λ on 40 m) as the reference configuration (C1). As shown earlier in this series, a radial system with thirty two ¼ λ radials is usually pretty good. You can add more radials, but the gain is relatively small, so a 32-radial system is a good compromise, and probably more typical of amateur installations. The radials were close to ¼ λ on 40 m. Figure 1 shows a plan view of the initial radial fan geometries.

The four 180° sectors were arranged in relation to the receiving antenna as follows:

1) Radials toward (C2),
2) Radials away (C3),
3) Radials to the left (C4), and
4) Radials to the right (C5).

Both right and left configurations, which ideally should be identical, were run as a check on the consistency of the measurements.

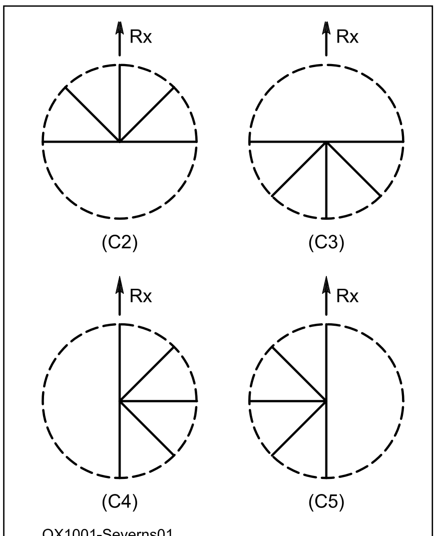

Figure 1 — Missing sector radial layouts.

After running tests using configurations C1 through C5, I realized that some additional radial configurations might be interesting. In particular I wanted to see how much adding some short radials in the missing sector would improve things.

I added the configurations shown in Figure 2 to the experiment:

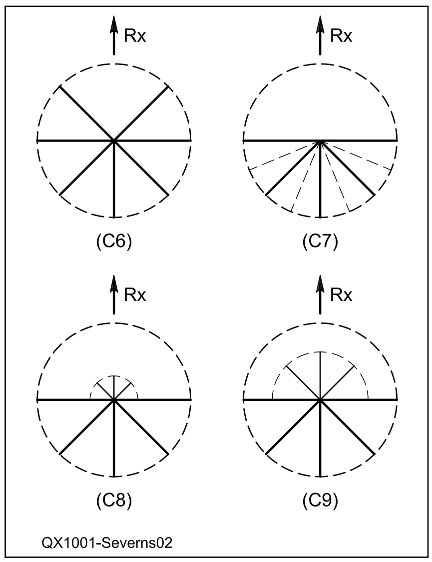

Figure 2 —Additional asymmetric ground systems.

Table 1

Effect of a 180° Sector Ground System on Signal Strength (S_{21}) in a Given Direction Relative to the Receive Antenna

Frequency (MHz)	*C2 Toward RX (dB)*	*C3 Away from RX (dB)*	*C4 Left (dB)*	*C5 Right (dB)*
7.2	–0.42	–1.91	–0.82	–0.94
14.2	–0.57	–2.42	–1.20	–1.24
21.2	–0.69	–3.00	–1.24	–1.33
28.5	–0.55	–3.23	–1.26	–1.58

Table 2
S_{21} Test Results for the Added Radial Configurations

Radial Configurations	*$\|S_{21}\|$ Referenced to C1 (0.0 dB)*
C6	–0.44
C3	–1.91
C7	–1.39
C8	–1.52
C9	–0.34

5) A 90° missing sector (7 radials removed, 25 radials remaining) (C6). The axis of the missing sector was pointed at the receiving antenna.

6) To C3, which has 17 radials facing away, I added an additional sixteen 33 foot radials between the seventeen already there (33 radials total) (C7). The missing 180° sector was facing the receiver.

7) To C3 I added fifteen 8.5 foot radials in a fan towards the receiving antenna. These are 1⁄16 λ radials on 40 m (C8). C9) To C3 I added fifteen 17 foot radials in a fan towards the receiving antenna. These are 1⁄8 λ radials on 40 m.

Test Results

Modeling ground systems with missing sectors using *NEC* indicates that compared to a full 360° system we should see both a reduction in the peak signal and a distortion in the pattern; in other words, a front-to-back ratio not equal to 0 dB.

Experimental results are given in Tables 1 and 2. Note that Tables 1 and 2 show the *difference* in dB from the 360° radial fan (C1), which is the reference.

Clearly sector radial systems have an impact on the radiated signal. In the direction of the remaining radials the signal loss is on the order of 0.5 dB, but in the direction of the missing sector the loss is from 1.9 to over 3 dB. If you have a 3 dB loss, that means you have lost half your power. Not good!

The test results qualitatively agree with *NEC*, the peak amplitude is reduced and the pattern is distorted when only a partial radial fan is employed. The radial system used for the tests reported in Table 1 has 33 foot radials, which of course are long for frequencies above 7.2 MHz. As we saw in the discussion for multi-ground systems (Part 6), the system with all 40 m radials gives the best performance, even better than if we used thirty two ¼ λ radials tailored for each band.

The test results for radial configurations C6 through C9 are given in Table 2. All of these tests were done at 7.2 MHz.

The first thing we see is that omitting the seven radials in a 90° sector (C6) does not seem to do too much harm, only –0.44 dB. Eliminating all the radials in a 180° sector (C3) is not good, however (–1.91 dB). The loss jumps by almost 1.5 dB over the 90° case!

Taking the radials removed from C1 (to form C3) and adding them between the remaining radials in C3 (C7) helps a little bit, reducing the loss by 0.5 dB. If, instead, we add fifteen 1/16 λ radials (C8) in the missing sector we get a similar improvement, about 0.4 dB. Despite some improvement, the signal loss for both C7 and C8 is still substantial. What really seems to help is to put fifteen 1/8 λ radials (C9) in the missing sector. Unfortunately, that may not always be possible.

Some Closing Comments

Overall, it's pretty clear both from modeling and experiment that sector ground systems can reduce your signal substantially in some directions and produce a distorted pattern.

What can we do about this? The first thing is to remember that the field intensity around the vertical increases rapidly as we get near the base of the antenna.[1] If we move the base of the antenna away from the obstacle as little as 1/16 λ or better yet 1/8 λ, so that we can have at least some radials in the sector towards the obstacle, the losses will be reduced. As shown above, 1/8 λ spacing can be quite effective. In the process of moving the base away from the obstacle you may have to shorten some of the other radials on the side away from the structure but that may be acceptable. Another possibility would be to move the base from the side of the building to a corner which might allow the radial fan to be increased from 180 to 270°. As the test data shows, this can be very helpful.

These experiments were done in an ideal situation. There was no actual structure next to the antenna. In addition to the losses we see in this idealized situation, it is very likely that the structure blocking the radial fan will increase the loss. It is difficult to estimate how much the loss will increase, but it's not likely that the building will improve your signal! Another factor to consider is the soil characteristics. My soil, over which these tests were conducted, would be rated as good or even very good, depending on the time of year. Poorer soils would result in even larger negative effects due to the use of a sector ground system than those shown in Tables 1 and 2.

What I have shown here represents only a few of many possibilities. It's not possible to experimentally examine all possible situations, but *NEC* modeling should give you a good qualitative feeling for your particular situation. One common situation that I did not have time to examine experimentally is the case where the base is alongside the house but not too far from a corner. The conventional wisdom is that you should run the radials along the side of the house to the corner and then fan them out from there. I don't think that can hurt but keep in mind that the farther you are from the corner, the less effective this scheme is likely to be.

[1]Rudy Severns, N6LF, "Verticals, Ground Systems and Some History," *QST*, Jul 2000, pp 38-44.

Rudy Severns, N6LF, was first licensed as WN7WAG in 1954 and has held an Extra class license since 1959. He is a consultant in the design of power electronics, magnetic components and power-conversion equipment. Rudy holds a BSE degree from the University of California at Los Angeles. He is the author of three books and over 90 technical papers. Rudy is an ARRL Life Member, and also an IEEE Fellow.

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Experimental Determination of Ground System Performance for HF Verticals Part 5 160 Meter Vertical Ground System

How much will the signal strength and feed point impedance change as radials are added?

This experiment was actually the first of the series of experiments on ground systems that have been the subject of this series of articles. The experiment involved measuring the change in signal strength as radials are added to the ground system of a vertical antenna, beginning with four radials and going up to 64 radials. The intent was to determine the additional gain in signal for each doubling of radial number, and to determine the point of vanishing returns. In addition, the changes in feed point impedance due to changing radial number were of interest.

While the results of this initial experiment were quite interesting, a more important result was an appreciation of the difficulties of making these measurements accurately. This experience led to a modification in the test procedure and a shift to 40 m verticals, which have been described earlier.

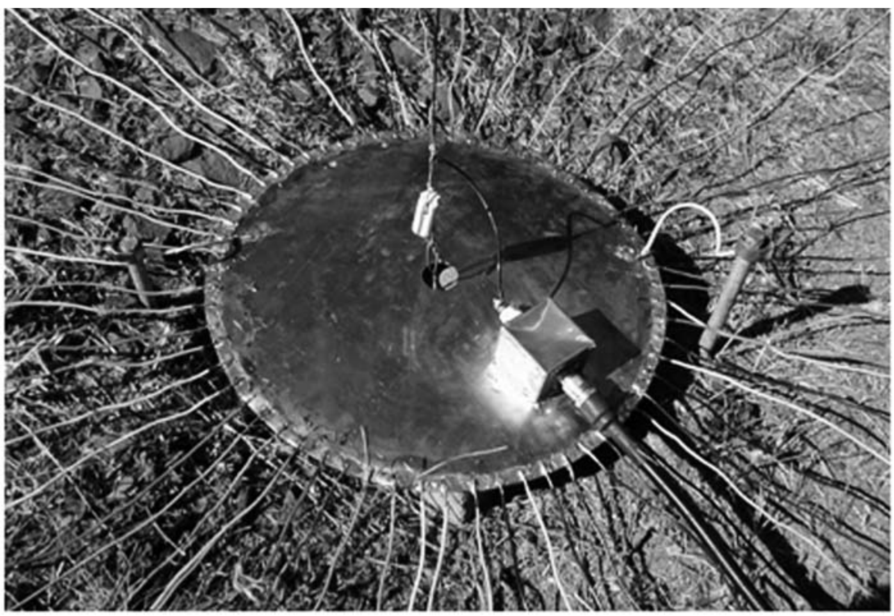

Figure 1 — This photo shows the antenna base with radials attached.

Test Antenna Description

The test frequency for this experiment was 1.800 to 2.000 MHz. The vertical was 125 feet of no. 12 AWG insulated copper wire suspended from a Dacron line hung between two 150 foot poles.

At the base of the antenna there was an 18 inch diameter copper disk, as shown in Figure 1. The inner ends of the radials and the shield of the coax feed line were attached to the disk. There were also two galvanized ⅝ inch × 4 foot ground stakes connected to the disk. The radials were 130 foot lengths of no. 12 insulated (THHW) wire lying on the ground surface. Radials were put down in the sequence of 4, 8, 16, 32 and 64.

The terrain around the antenna was not flat, but rather on a narrow ridge about 40 to 50 feet wide. The result is that many of the radials were in part bent down at about a 45° angle as they ran down the steep slope on either side. Along the ridge, however, the radials are more or less level.

The test antenna was erected 700 feet to the east of my house with a 50 foot deep gully in between. The ridge is in a Douglas fir forest with 100 plus foot trees within 50 feet of the test antenna at some points. The radial system ran along the ridge and also down the sides of the ridge into the forest.

To excite the test antenna, between the house and the antenna there was a 700 foot length of 1⅝ inch coax, with an additional 75 feet of ½ inch coax. Both were Andrews heliax.

Measurement Equipment

The signal source was a Yaesu FT1000MP transceiver with two Bird Model 43 wattmeters on the output (forward and reflected power). The wattmeters were used to set the forward power to a constant 50 W and also to measure reflected power to calculate SWR. The SWR measurement is needed to correct for the power reflected from the antenna and not radiated. This correction was applied to the received signal amplitude.

The receiving antenna was a 10 foot vertical wire driven against a 4 foot ground stake, next to my house. The receiver was an HP3585A spectrum analyzer. The amplitude resolution was about ± 0.1 dB.

Base impedance measurements were made at the antenna using an N2PK vector network analyzer (VNA). The impedance measurements were accurate to better than 1%.

The test procedure was very straightforward. For each number of radials, the FT1000MP output was adjusted to 50 W and received signal strength on the spectrum analyzer recorded along with the SWR for that measurement and the input impedance at the base of the antenna.

Test Results

Three complete runs were made to verify repeatability of the measurements. Each run included a complete stepping through the number of radials in the sequence, 4, 8, 16, 32 and 64. Typical received (and corrected for SWR) signal strengths versus radial number are given in Table 1. This data is graphed in Figure 2.

The data in Figure 2 has one obvious oddity. You would expect that the incremental difference as the radial numbers are doubled would be monotonically decreasing as the radial number rises. The step between 16 and 32 radials does not do this and it appears that the value for 16 radials is too small. This anomaly was noted during the experiment, however, and checked carefully as the radial count was redone three times. The anomaly was there in all three cases. I have no explanation for this other than the irregularity of

Table 1
Typical Test Data for Received Signal Strength with P_o = 50 W.

Number of Radials	*Corrected Signal Strength*	*Relative Signal Strength*
4	–30.1 dBm	0.0 dBm
8	–29.3 dBm	0.8 dBm
16	–28.9 dBm	1.2 dBm
32	–28.0 dBm	2.1 dBm
64	–27.7 dBm	2.4 dBm

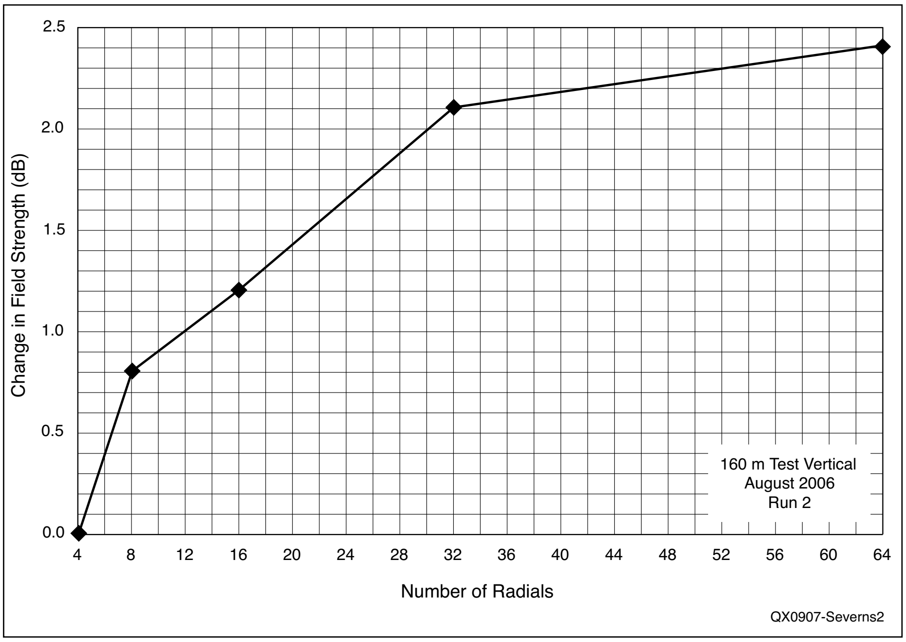

Figure 2 — Here is a graph of the typical signal strength change with radial number.

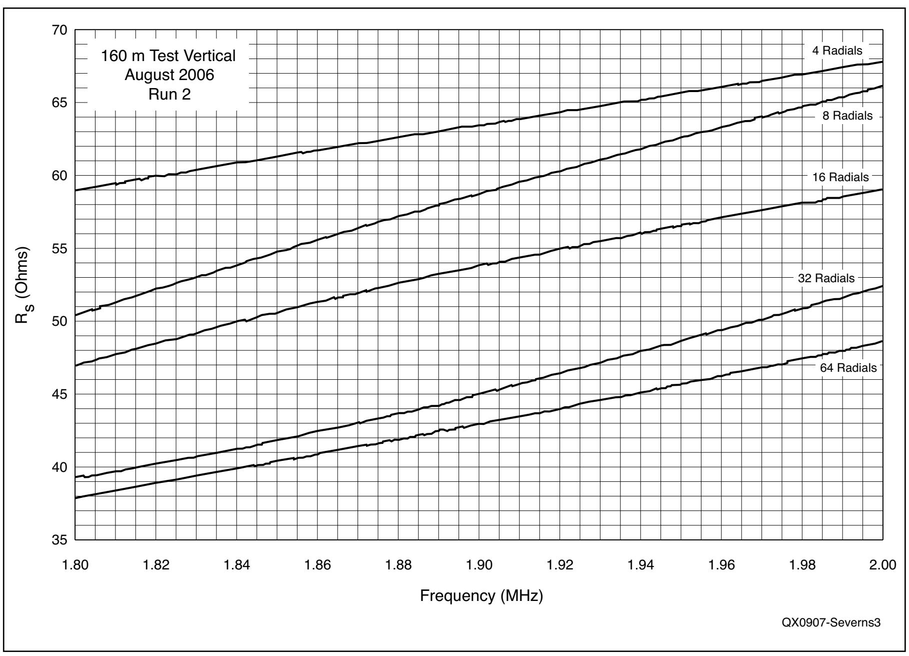

Figure 3— This graph gives the resistive part of the base impedance over the 160 m band for different radial numbers.

the site, which forced the radial layout to be far from flat or level. Later experiments with more regular radial systems on other antennas all showed the expected monotonic decrease in improvement with increasing radial number.

In any case, it's pretty clear that 32 radials do a good job and by 64 radials you are well into the region of vanishing returns. I certainly could not justify doubling the radial count to 128!

The results of feed-point impedance measurements are given in Figures 3, 4 and 5.

As discussed in Part 2 of this series, we would expect the resonant frequency to vary with the number of radials, due to the shift in radial resonance because of soil loading. The 40 m experimental work was done over an essentially flat pasture and the resonant frequency change was regular and monotonic. The gross irregularity of the ground surface in this earlier experiment, however, resulted in the erratic frequency changes shown in Figure 5. This problem was a primary reason for moving the experimental site from the narrow ridge to a pasture. Unfortunately, the 150 foot support poles were not available in the pasture so it was necessary to change the experimental frequency to 40 m to make the vertical height manageable.

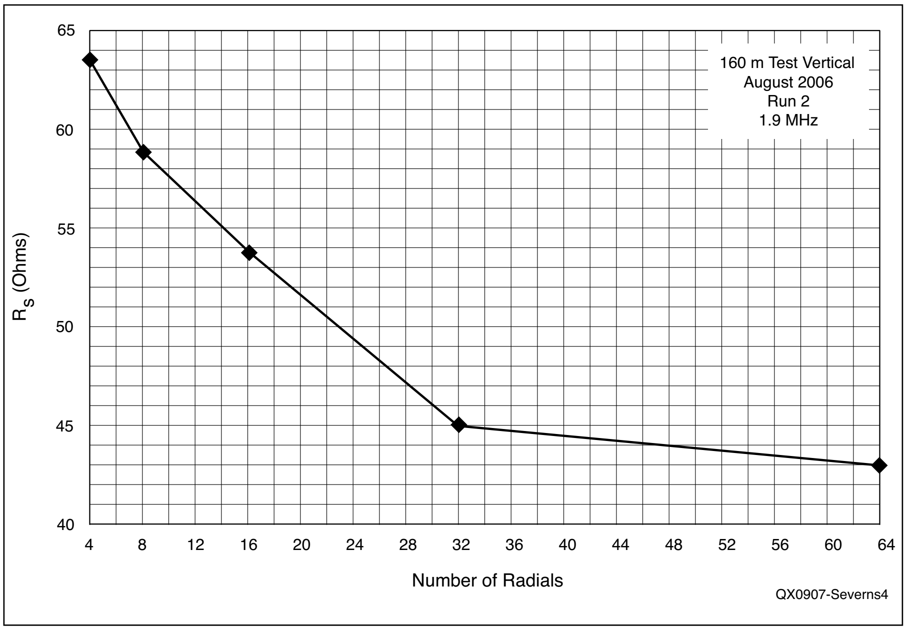

Figure 4 —This graph shows the base resistive component versus radial number at 1.9 MHz.

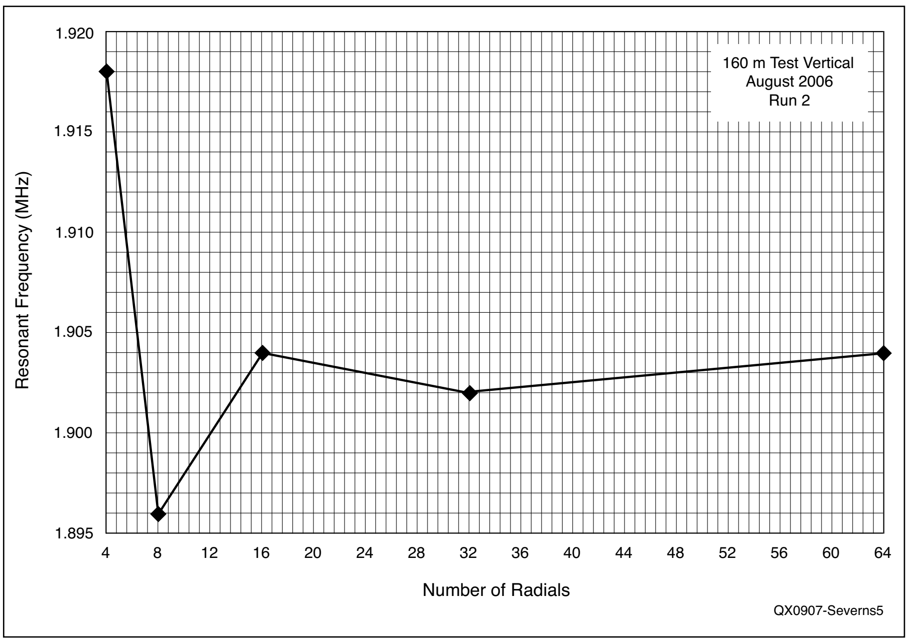

Figure 5 —This graph shows the antenna resonant frequency for different numbers of radials.

Summary

This initial experiment helped me to understand the problems inherent in making accurate comparisons between different ground systems. I had to change the site, the test frequency, the test instrumentation and the test methodology to get to the point where I could have confidence in the test results and draw conclusions from them.

This experiment was by no means a failure, however. We can see that the change in signal strength is very much in line with what we saw in the 40 m work. It also supports the conclusion that we should use at least 16 radials, but when we use more than 32 radials we are definitely reaching the point of vanishing returns. For most amateur installations the Standard Broadcast ground system of one hundred twenty 0.4-wavelength radials could not be justified by any useful increase in signal strength.

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QEX

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Radiation and Ground Loss Resistances In LF, MF and HF Verticals: Part 1

With the impending FCC announcement about the release of a new LF and a new MF band, hams will be interested in practical antennas and learning how to calculate EIRP to legally operate on those bands.

Unlike the higher bands, where the maximum transmitting power limit is stated in terms of transmitter output power, on the (soon to be released) 630 m (472 to 479 kHz) and 2200 m (135.7 to 137.8 kHz) bands, the maximum allowable power is stated in terms of the effective isotropic radiated power (EIRP) from the antenna. On 630 m the maximum EIRP allowed is 5 W, which for the short verticals likely to be used at 475 kHz, translates to a radiated power (P_r) of 1.7 W. (For more information on EIRP, see the sidebar.)

This raises the question, "How do we determine P_r?" As shown in the sidebar, the standard professional approach has been to measure the field strength at a point some distance from the antenna and then calculate EIRP. That's fine for the pros, but for most amateurs, that method won't be practical. There are other ways we might go about it, however. For example, if we can measure the current at the feed point (I_o) and if we know the radiation resistance (R_r) referenced to the feed point, we can find the radiated power from Equation 1.

$$P_r = I_o^2 \times R_r \quad \text{[Eq 1]}$$

An alternative would be to measure the feed point resistance (R_i) and the input power (P_i) and then calculate P_r using Equation 2.

$$P_r = (R_r / R_i) \times P_i \quad \text{[Eq 2]}$$

We can measure quantities like I_o, P_i, and R_i, but there is no way to measure R_r directly.

Feed Point Equivalent Circuit Model

Figure 1 shows the traditional equivalent circuit used to represent the resistive part of an antenna's feed point impedance (R_i) when describing what happens to the input power, P_i. The radiation resistance, R_r, represents the radiated power.

$$P_r = I_o^2 \times R_r \quad \text{[Eq 3]}$$

where:

I_o is the current at the feed point in rms amperes.

The power lost in the soil close to the antenna is represented as R_g. The sum of other ohmic losses such as conductor loss, insulator leakage, and so on is represented as R_L. The input resistance at the feed point is assumed to be the sum of these resistances.

$$R_i = R_r + R_g + R_L \quad \text{[Eq 4]}$$

Determining P_L is reasonably straightforward, but P_g is trickier. In the following discussion I will be ignoring R_L. In other words, we will assume lossless conductors. This is not because these losses are unimportant but the interest here is in R_r and R_g, and how they vary with frequency, ground system design and soil characteristics. P_L is certainly a worthy subject, but we will save that for another day.

The traditional assumption has been that R_r for a vertical over real ground is the same as it would be for the same antenna over perfect ground. The value we measure for R_i is assumed to be the sum of the R_r for perfect ground and additional loss terms that result from ground and other loss elements. I've certainly gone along with the conventional thinking, but over the years I've become skeptical after seeing experimental and modeling results and calculations that didn't fit. I've come to the conclusion that at HF at least, R_r for a given vertical over real soil, is not the same value for the same antenna over perfect ground.

The following discussion focuses on the concept illustrated in Figure 1, with R_L = 0. The discussion will show that at HF (1.8 MHz and higher frequencies), R_r

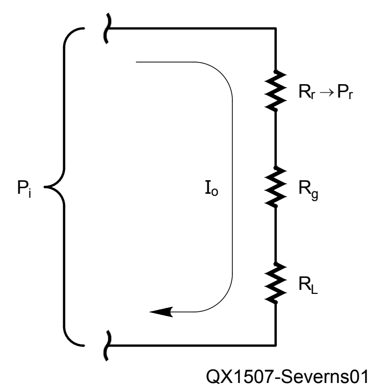

Figure 1 — This is a typical equivalent circuit for an antenna feed point resistance.

differs significantly from the value over ideal ground. At LF (137 kHz) and MF (472 to 479 kHz), however, the variation of R_r from the ideal value is much smaller, which is very helpful for determining P_r.

To make this article easier to read I've placed almost all the mathematics and the many supporting technical details in an extensive set of Appendices.

Appendix A — Shows how to calculate R_r using the Poynting vector.

Appendix B — Gives a review of soil characteristics.

Appendix C — Describes the E and H fields and power integration.

Appendix D — Covers other miscellaneous bits.

Pushing material into appendices makes life much easier for the casual reader, but provides the gory details for those who want them. These appendices are available on my web site: **www.antennasbyn6lf.com** and are also available for download from the ARRL *QEX* files web page. Go to **www.arrl.org/qexfiles** and look for the file **7x15_Severns.zip**.[1]

R_r For A Lossless Antenna

We need to be careful with our use of the term "radiation resistance." A definition of R_r associated with a lossless antenna in free space, can be found in almost any antenna book. A typical example is given in *Radio Engineers' Handbook* by Frederick Terman:[2]

"*The radiation resistance referred to a certain point in an antenna system is the resistance which, inserted at that point with the assumed current* I_o *flowing, would dissipate the same energy as is actually radiated from the antenna system. Thus:*

$$\text{Radiation resistance} = \frac{\text{radiated power}}{I_o^2}$$

Although this radiation resistance is a purely fictitious quantity, the antenna acts as though such a resistance were present, because the loss of energy by radiation is equivalent to a like amount of energy dissipated in a resistance. It is necessary in defining radiation resistance to refer it to some particular point in the antenna system, since the resistance must be such that the square of the current times radiation resistance will equal the radiated power, and the current will be different at different points in the antenna. This point of reference is ordinarily taken as a current loop, although in the case of a vertical antenna with the lower end grounded, the grounded end is often used as a reference point."

Discussions of R_r for the lossless case are common but I've not seen a discussion of R_r where the effect of near-field losses are considered. In his book, *Antennas*, Kraus does tease us with a comment:[3]

"*The radiation resistance* R_r *is not associated with any resistance in the antenna proper but is a resistance coupled from the antenna* ***and its environment*** *to the antenna terminals.*"

The bold type is mine! The implication that the environment around the antenna plays a role is important but unfortunately Kraus does not seem to have expanded on this observation.

Calculation of R_r and R_g

As pointed out earlier if you know I_o and P_r, you can calculate R_r. A standard way to calculate the total radiated power is to sum

EIRP and Radiated Power, P_r, From Verticals

On 630 m the maximum allowable power is stated in terms of effective isotropic radiated power (EIRP), which is not the same as the radiated power ($P_r = R_r \times I_o^2$, where I_o is the rms current). It is important to understand the difference. As shown in Figure SB1, an isotropic radiator is one that radiates uniformly in all directions. The power density, P_{di}, is the same in all directions at a given radius. If you place a short monopole over a perfect ground plane, for the same P_r, the power density at the same radius will be greater by a factor of 3 (+4.77 dB). The factor of 3 occurs because the power density is doubled (+3 dB) by going from free space to the perfect ground plane, and there is a further increase of 1.5 × (+1.77 dB) because of the directivity of the short monopole.

To achieve the same P_d at the same radius, if we excite the isotropic antenna with P_r = 5 W, we can only excite the monopole with P_r = 1.7 W.

To determine the power density (P_d) in the wave front, we can make a field strength ($|E_z|$) measurement at some distance r from the antenna.

$$P_d = \frac{|E_z|^2}{377} \approx \frac{|E_z|^2}{120\,\pi}\left[\frac{W}{m^2}\right] \quad \text{[Eq SB1]}$$

Note, E_z is in V/m and 377 Ω represents the impedance of free space. Implicit in Equation SB1 is the assumption that the measurement of E_z has been taken far enough from the antenna to be in the far field, where $|E_z| / |H_y| \approx 377\ \Omega$. At 630 m, you need to be at least 5 λ away, or about 3 km, and 5 km would be better.

Assuming P_d is constant over a sphere with radius r (in meters) you can multiply P_d by the area of the sphere to obtain EIRP.

$$EIRP = \frac{r^2 |E_z|^2}{60}[W] \quad \text{[Eq SB2]}$$

The point is that while we are allowed an EIRP = 5 W, the allowed P_r is about 1.7 W!

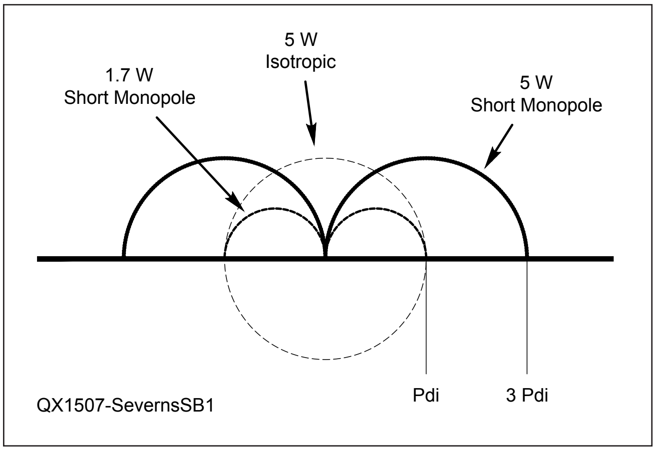

Figure SB1 — Radiation power density at the same radius from an isotropic radiator in free space and a short monopole over perfect ground.

[1]Notes appear on page 34.

(integrate) the power density (in W/m^2) over a hypothetical closed surface surrounding the antenna. For lossless free space calculations the enclosing surface can be anywhere from right at the surface of the antenna to a sphere with a very large radius (large in terms of wavelengths). For P_r calculations, a large radius has the advantage of reducing the field equations to their far-field form, which greatly simplifies the math. This is fine for lossless free space or over perfect ground, where near-field or far-field values give the same answer. When we add a lossy ground surface in close proximity to the antenna, however, things get more complicated. Note that the terms near-field, Fresnel, and far-field are carefully defined in Appendix C.

Take for example a vertical ½ λ dipole with the bottom a short distance above lossy soil. You could create a closed surface that surrounds the antenna but does not intersect ground, and then calculate the net power flow through that surface. When you do this you find the R_i provided by *EZNEC* (my primary modeling software) will be the same as the R_r calculated from the power passing through the surface. Technically, this is R_r by the free space definition, since the antenna is lossless, as is the space within the enclosing surface, but that's not how we usually think of the relationship between R_i and R_r. The conventional point of view is that the near-field of the antenna induces losses in the soil, which we assign to R_g, separate from R_r, as indicated in Figure 1. The power absorbed in the soil near the antenna is not considered to be "radiated" power although clearly it is being supplied from the antenna. When we run a model on *NEC* or make a direct measurement of the feed point impedance of an actual antenna, we get a value for R_i from Equation 5.

$$R_i = R_r + R_g \qquad \text{[Eq 5]}$$

Can we separate R_r from R_g, and if so, how? Assuming we're going to use *NEC* modeling, we could simply use the average gain calculation (G_a). The problem with G_a is that it includes all the ground losses, near and far-field, ground wave, reflections, and so on. For verticals, G_a gives a realistic, if depressing estimate of the power radiated for sky wave communications, but the far-field loss is not usually included in R_g. Typically, R_g represents only the losses due to the reactive near-field interaction with the soil. In the case of a ¼ λ ground based vertical for example, that would be the ground losses out to ≈ ½ λ (see Appendix C). Instead of using G_a we can have *NEC* give us the amplitudes and phases of the *E* and *H* fields on the surface of a cylinder, which intersects the ground surface as indicated in Figure 2.

The power density is integrated over the

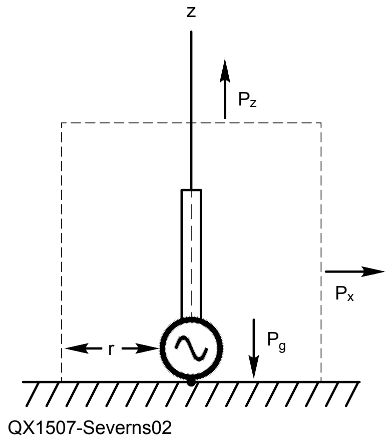

Figure 2 — We can use *NEC* modeling to calculate the *E* and *H* fields on a cylindrical surface enclosing a ground mounted vertical.

surface of the cylinder (P_x) and over the surface of the disc (P_z) that forms the top of the cylinder, giving us P_r directly. Instead of integrating the power over the surface of the cylinder we could sum the power passing through the soil interface at the bottom of the cylinder, which gives P_g directly. From either P_r or P_g we can calculate R_r using Equation 6.

$$R_r = \frac{P_r}{I_o^2} = \frac{\left(P_i - P_g\right)}{I_o^2} \qquad \text{[Eq 6]}$$

Of course this is more complicated than simply using G_a! It turns out, however, that if you're moderately clever in your choice of surface and field components, it can be quite practical to calculate the values using a spreadsheet like Microsoft *EXCEL*. The mathematical details are in Appendix A. Because the fields near a vertical are sums of decaying exponentials (1/r, 1/r^2, 1/r^3) the boundaries between the field regions are not sharply defined, the choice for the cylinder or disc radius (r) is somewhat arbitrary. The rather messy details of the choice of integration surface radius are discussed in Appendix C.

R_r and R_g for a ½ λ Vertical Dipole

For simplicity, I began this study using a resonant vertical ½ λ dipole like that shown in Figure 3, with the bottom of the antenna placed 1 m above ground. The analysis was done at several frequencies, two of which are reported here — 475 kHz and 7.2 MHz. Note the frequencies are a factor of ≈ 16× apart. In a later section, I give an example at 1.8 MHz. The antennas heights (*h*) were adjusted for resonance over perfect ground and that height was retained for modeling

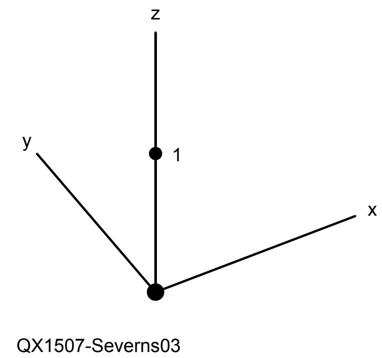

Figure 3 — This model shows a ½ λ vertical dipole, with the bottom of the antenna 1 m above ground.

over real soil.

Figures 4 and 5 show the variation in R_i at 7.2 MHz and 475 kHz for a wide range of soil conductivity (σ) and permittivity (ε_r, relative dielectric constant). The notation "J =" on the Figures indicates the height of the bottom of the antenna above ground.

As we would expect, in free space $R_r \approx 72\ \Omega$ and over perfect ground Rr ≈ 95 – 100 Ω for these antennas. Over real ground R_i varies dramatically with both soil characteristics and frequency. One point is obvious:

R_i is not a combination of R_r over perfect ground and some R_g!

On 40 m, values for R_i over real soils are all lower than the perfect ground case, but the values on 630 m vary from well below the perfect ground case to slightly above. In both cases, as ground conductivity increases, R_i converges on the perfect ground case as one would expect. For very low conductivities, we can see that ε_r has a profound influence on R_i, but its effect is greatly reduced for high conductivities. Note that at 475 kHz for σ ≧ 0.0001 S/m, R_i rapidly converges on the perfect ground value, and the effect of ε_r is minimal. On the other hand, at 40 m the jump in R_i doesn't occur until σ ≧ 0.003 S/m, that's more than an order of magnitude higher than 475 kHz. It would appear that at 475 kHz the value for ε_r doesn't matter much over most common soils, but at 7.2 MHz it has a major influence for some typical values of σ. What's going on here?

Soil Characteristics

It is important to understand that the characteristics of a given soil will vary with frequency. The following is a brief overview. You can find a much more detailed discussion in Appendix B. Figures 6 and 7 are examples of σ and ε_r for a typical soil over a frequency range from 100 Hz to 100 MHz. These graphs

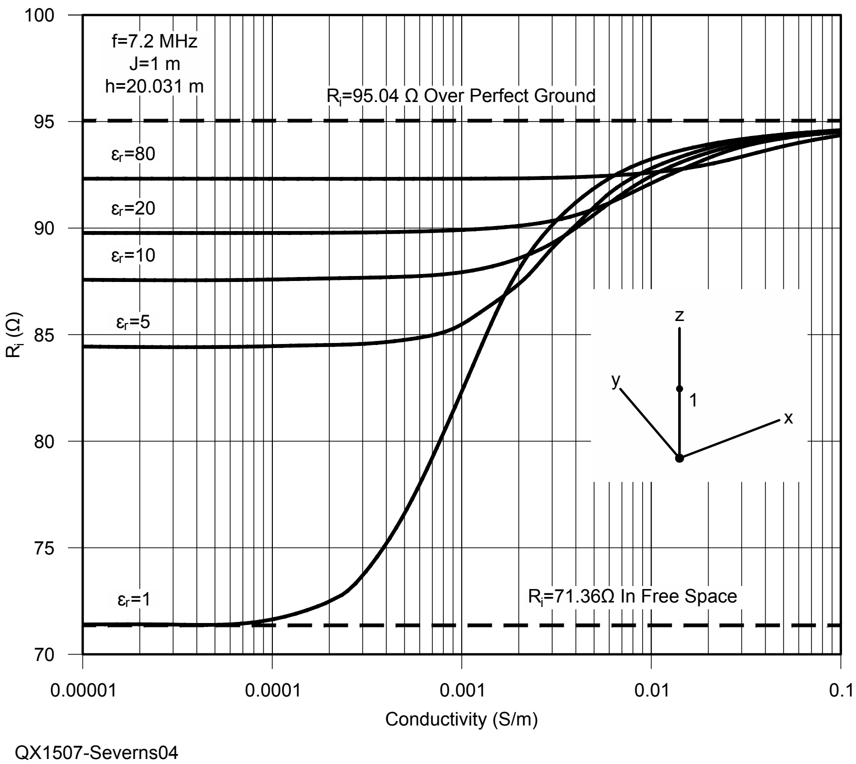

Figure 4 — Here is a graph of R_i versus ground conductivity for a ½ λ vertical dipole at 7.2 MHz.

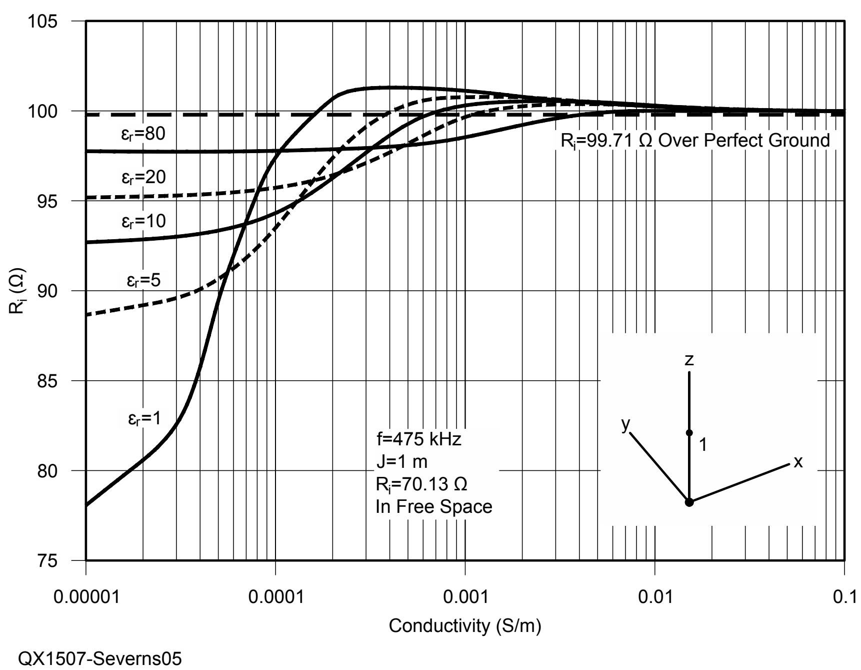

Figure 5 — This graph shows R_i versus ground conductivity for a ½ λ vertical dipole at 475 kHz.

were generated using data excerpted from *Antennas in Matter* by King and Smith.[5] In this example, at 100 Hz $\sigma \approx 0.09$ S/m and that value is relatively constant up to 1 MHz, beyond which σ increases rapidly. The behavior of the relative dielectric constant (ε_r) is just the opposite, decreasing with frequency until about 10 MHz and then leveling out. We can combine σ and ε_r by using the loss tangent (D).

$$D = \tan\delta = \frac{\sigma_e}{2\pi f \varepsilon_e} \qquad \text{[Eq 7]}$$

where:

$\varepsilon_e = \varepsilon_o \, \varepsilon_{er}$ = effective permittivity or dielectric constant (in farads/m)

ε_o = permittivity of a vacuum = 8.854×10^{-12} farads/m.

For a good insulator, D<<1 and for a good conductor, D>>1. For most soils at HF 0.1<D<10, but it is often close to 1.

We can combine the data in Figures 6 and 7 into a graph for D, as shown in Figure 8.

Figure 8 shows that something interesting happens when we go from HF down to MF. At HF, D is usually not far from 1, but at MF, D is usually much higher. This implies that the soil characteristics are dominated by conductivity. Figures 4 and 5 show that at MF, conductivity becomes the dominant influence at much lower conductivities than at HF. This explains some of the features of Figures 4 and 5.

Relationships Between *D*, *R*$_r$ and *R*$_g$

The role of the loss tangent, D, is worth exploring a bit further. Figure 4 showed the variation in R_i as ε_r and conductivity were varied. In a similar way we can examine the variation in R_r and R_g over the same range of variables as shown in Figure 9, which is a graph of R_i, R_r, and R_g with $\varepsilon_r = 10$ for the 40 m ½ λ vertical. On the chart there is a vertical dashed line corresponding to values of σ where $D = 1$ for $\varepsilon_r = 10$ ($\sigma \approx 0.004$ S/m in this example). Something interesting happens in the region around the point where the loss tangent equals one.

A very prominent feature of Figure 9 is that R_r and R_g are *not* constant as we vary σ. The value for R_g (which represents ground loss) peaks near $D = 1$, which is what dielectric theory predicts for the maximum dissipation point. We can take one further step with the data in Figure 9, and graph the ratio R_r / R_i (which is the radiation efficiency) as shown in Figure 10. The minimum efficiency (≈0.66) occurs at $\sigma \approx 0.0025$ S/m.

This graph emphasizes the effect of the loss tangent on ground loss.

Acknowledgements

I want to express my appreciation to Steve

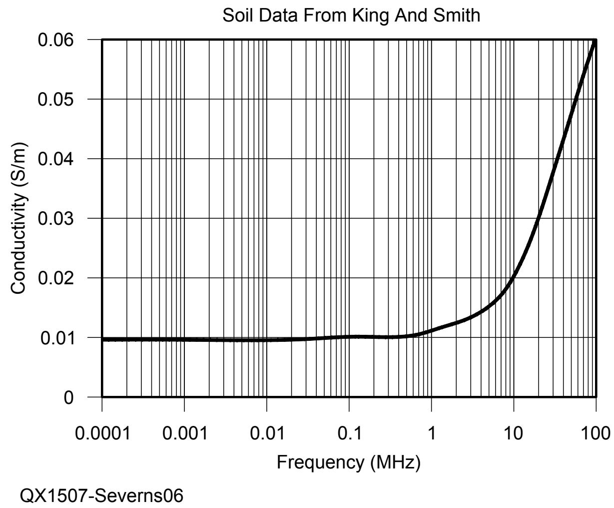

Figure 6 — This graph gives an example of how soil conductivity varies with frequency.

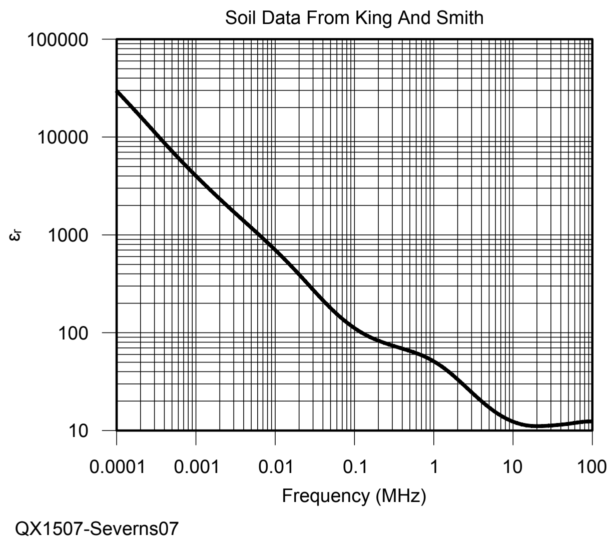

Figure 7 — This graph shows soil permittivity variation with frequency.

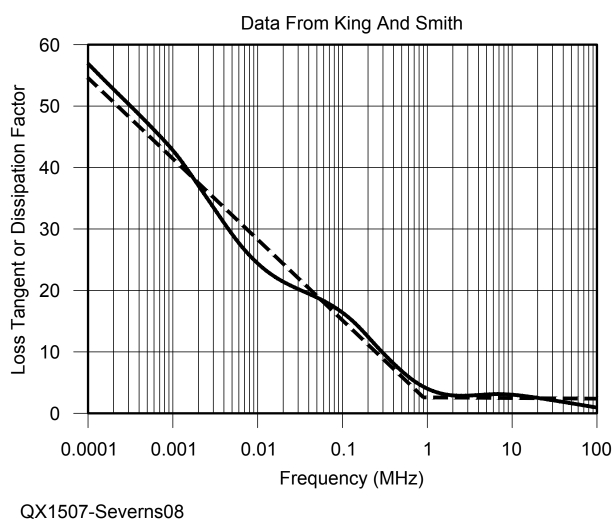

Figure 8 — Here is a graph of the loss tangent associated with the soil in Figures 6 and 7.

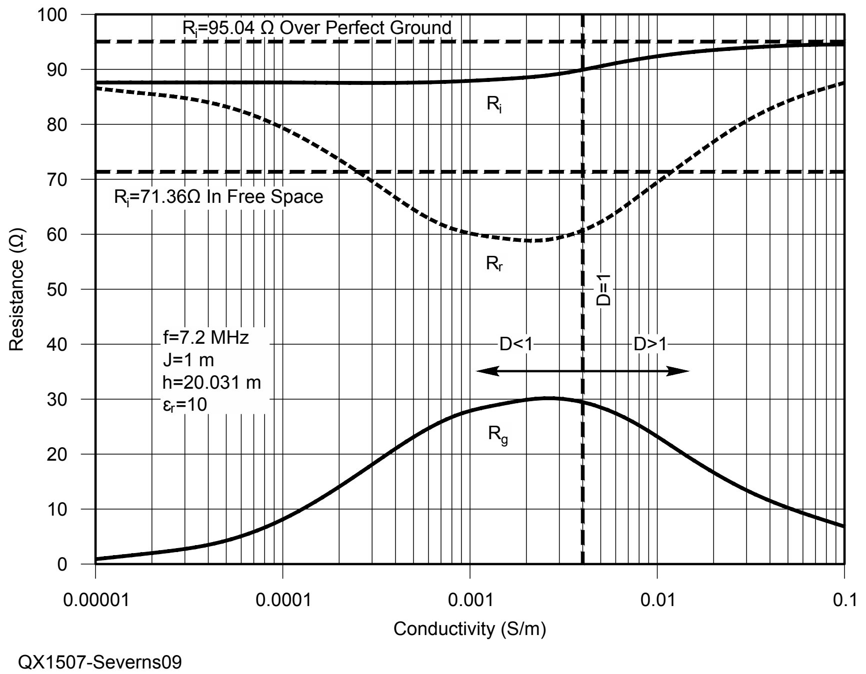

Figure 9 — Variations in R_i, R_r, and R_g with ε_r = 10.

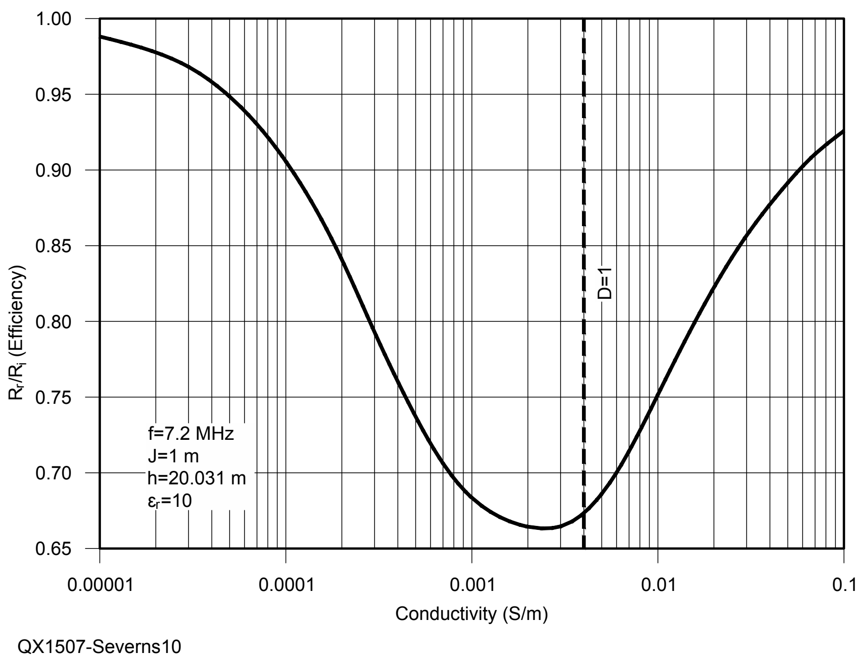

Figure 10 — Here we see the variation of radiation efficiency with ε_r = 10.

Stearns, K6OIK, for his very helpful review of this article. He put in a lot of effort and I've incorporated many of his suggestions in the main article and in the Appendices. I also appreciate the comments from Dean Straw, N6BV, and Al Christman, K3LC. All of the modeling employed a prototype version of Roy Lewallen's (W7EL) *EZNEC Pro/4* modeling software (see Note 4) that implements *NEC* 4.2, and Dan MaGuire's (AC6LA) *AutoEZ*, which is an *EXCEL* spreadsheet that interacts with *EZNEC* to greatly expand the modeling options. Without these wonderful tools this study would not have been practical and I strongly recommend both programs.

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Notes

[1]The Appendices and other files associated with this article are available for downloading from the ARRL *QEX* files web page. Go to **www.arrl.org/qexfiles** and look for the file **7x15_Severns.zip**.

[2]Frederick Terman, *Radio Engineers' Handbook*, McGraw-Hill, 1943.

[3]John Kraus, *Antennas*, McGraw-Hill, 1988, second edition.

[4]Roy Lewallen, W7EL, *EZNEC pro/4*, **www.eznec.com**.

[5]King and Smith, *Antennas in Matter*, MIT Press, 1981, p 399, Section 6.8.

[6]J. Wait, R. Collin and F. Zucker, *Antenna Theory*, Chap 23, Inter-University Electronics Series (New York: McGraw-Hill, 1969), Vol 7, pp 414 – 424.

[7]Dan McGuire, AC6LA, *AutoEZ*, **www.ac6la.com/autoez.html**.

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The Case of Declining Beverage-on-Ground Performance

Detailed modeling and measurements that validate the use of NEC help explain why over the course of two winter seasons the performance of the Beverage on the Ground (BOG) antenna dropped off dramatically as the antenna slowly sank into the ground.

In midsummer of 2013 I placed a 450 foot length of insulated wire in my pasture configured as a Beverage-on-the-Ground (BOG) receiving antenna. At the same time I erected a terminated loop receiving antenna — a triangle, 70 feet high by 30 feet on the base. I already had a 30 foot vertical working as a non-directional *E*-probe with an amplifier. Over the last 18 months I've been decoding WSPR transmissions — which provide S/N estimates — and comparing reports between the antennas in an attempt to quantify their relative performances.

Initially the BOG and the loop were clearly superior to the vertical, and throughout the 18 months the loop performance was very consistent. The BOG worked well at first. However, over time and especially during the two intervening winter wet seasons, I noticed the BOG signal amplitudes dropping off significantly (-15 dB) and the S/N improvement dropped to no better than the vertical. With the coming of the last summer's dry season the BOG improved somewhat but never really came back. This winter the BOG was not very useful. I checked the connections, feed lines and all associated hardware carefully but found no problems, so this rather radical decline in performance was a mystery!

Figure 1 — Test antenna #1.

Recently, I received an email from Al Christman, K3LC, relaying a question he received from Carl Luetzelschwab, K9LA, regarding the reliability of NEC modeling for wires close to, or on the surface, or buried in the soil. There has been some skepticism regarding the validity of NEC modeling in these situations. Over the years I've often compared my modeling predictions with finished antennas and generally found very good correlation. However, while modeling *E*-and *H*-fields for verticals close to the soil-air interface I saw some anomalies in the *H*-field calculations when using NEC4.1, which uses the GN2 ground code.

These problems have long been recognized but recently Jerry Burke modified the NEC code to NEC4.2 upgrading to GN3, improving modeling of the ground interaction. I've had a chance to try GN3 (incorporated into NEC4.2) and it did not generate the anomalies I'd seen with GN2. This prompted me to ask, "does NEC4.2 model antennas with wires close to and/or buried in soil well enough to explain why the performance of my BOG was declining so badly?" To answer that question I felt I had to validate NEC4.2 modeling to my satisfaction before I could confidently move on to my BOG problem.

I decided to perform a series of field experiments to see how well NEC predictions would correlate with actual antennas having wires parallel to the soil at low heights or buried in the soil. I also wanted to investigate an antenna that employed a ground rod. Since my interest is in antennas for 80 m and 160 m, I used test frequencies ranging from 1 to 4 MHz. By no means do my examples cover all possibilities but they are representative. Here is what I found.

Modeling Software and Instrumentation

NEC solves for the currents on the wires. From these currents both the feed-point impedance and the radiation pattern are calculated. If the impedances from the NEC model agree with the values measured on the actual antenna over a wide range of frequencies you can be reasonably sure the modeling is reliable. In the case of my BOG it would also be helpful to see if NEC4.2 would predict the current distribution along the wire at a given frequency, for example 1.83 MHz.

For the modeling part of this experiment I used *EZNEC Pro4 v6*, courtesy of Roy Lewallen, W7EL.[1] That version of *EZNEC* uses NEC 4.2. I also used the latest version of *AutoEZ* from Dan Maguire, AC6LA.[2] *AutoEZ* is an Excel® spread sheet with macros that automate a wide range of modeling tasks using *EZNEC* as the engine. For impedance measurements I used a vector network analyzer (VNA), either the VNA2180 from W5BIG or a homebrew N2PK VNA. I've made it a point to display the raw measurements without any "corrections" to the data points. That is why you can see noise present on the graphs of VNA measurements at frequencies associated with my local broadcast stations and, in one case, coupling to nearby verticals. The soil electrical characteristics were calculated at the same frequencies as the impedance measurements. This ground data was then inserted into the model. *AutoEZ* makes it easy to blend this kind of data into a model.

The following discussion addresses only NEC4.2, since NEC2 does not allow buried wires and does not do a very good job when the wires are close to ground. It is very possible that GN3 was not required for all the comparisons. NEC4.1 might very well have returned very similar results. I didn't repeat the modeling with NEC4.1 (GN2).

Soil Surface

First let's clarify the nature of the ground surface. When modeling, we assume the air-ground interface is a distinct line with the properties of air above it and the soil below it. NEC in its present form cannot model a "transition" zone. It's important to recognize that with real antennas the soil-air interface is not smooth nor sharply defined. Unless carefully reworked, the soil surface will be lumpy with varying characteristics both vertically and horizontally. As we'll see later, the characteristics of an antenna close to, or buried in, the soil are very sensitive to soil electrical characteristics so this "lumpiness" in the surface makes it difficult to get good correlation when modeling wires that are between one inch above and one inch below the surface. In effect there is no distinct soil-surface interface. What we do have in reality is a transition zone from air to soil, which we can model only approximately.

For example, in a pasture as you get closer to ground, first there is grass, then there is the body of grass plant, then there is the root system, and finally you reach actual soil. Even then you're still not home free. The moisture in the top few inches of soil varies quickly with rain and subsequent drying. If the antenna is installed in a forest, initially a surface wire will be lying on top of leaves or needles in various stages of decay, and other woody debris. In summer time this surface may be quite dry, so in effect the antenna is at a height of a few inches.

My experience, and that of others, as well as the modeling, show that this can provide a very good receiving antenna. However, with

Figure 2 — Center connector, common mode choke and feed point support.

the arrival of fall, leaves and needles will drop down on the wire, burying it to some degree. Also it's likely that the forest floor will be quite wet or even frozen.

I had an interesting exchange with Don Johnson, N4DJ, about his work with BOG antennas in a forest. His results were very good, and he did not notice the severe degradation in performance that I had experienced. It appears that the degradation over time is highly variable and specific to a particular installation, so we want to be careful about drawing general conclusions. If you live in the desert you may be able to place a wire directly on the soil surface and have that remain relatively unchanged for an extended period of time.

I think it is important to reiterate that modeling a wire lying on the ground surface is a special problem. My test antennas #1, #2, and #3 were modeled with the assumption that the air-soil interface was distinct, not fuzzy, and that seems to have worked well. In my case, the BOG wire (test antenna #4) was placed on the surface of a pasture in the summer time when the grass had been mowed and was very dry. The soil also was very dry, so the wire was effectively 1 to 3 inches above the soil. But over the period of 18 months the wire was swallowed up by the weeds, and by this winter it was buried in wet sod and tall grass. There really is no way to model this transition layer between air and the actual soil. What I've done is to compare a BOG antenna one inch above the soil to a BOG antenna one inch below the soil. There was good agreement between modeling and experiment.

Test antenna #1

The first test antenna was a center-fed dipole. I chose a length of 300 feet because that included both series (odd half-wave multiples) and parallel (even half-wave multiples) resonances within the test frequency range. This presented a wide range of impedance values at the feed point, from a few tens of ohms to several thousand ohms. I varied the height above ground from 48 inches down to 1 inch in the sequence 48, 24, 12, 6, 3 and 1 inch. A common mode choke was used for isolation. The feed-point impedance was measured with a VNA. The VNA calibration plane was directly at the antenna terminals. Soil electrical characteristics were measured concurrently. The details of the soil measurements are given in articles on soil electrical characterization. [3]

Figure 1 shows a view along the length of test antenna #1. The #17 AWG aluminum electric fence wire was supported on 5-foot fiberglass wands with plastic wire clips. The clips were moved up and down to adjust wire height. The wands were spaced 10 to 20 feet apart and the wire was anchored at the ends to steel fence posts that were more than 6 feet away from the ends of the wire. Multiple support points and significant wire tension kept the droop to less than a quarter of an inch. I used high quality insulators and non-conducting Dacron line at the wire ends, and a Budwig center connecter. Figure 2 shows the Budwig connector and common-mode choke at the feed point.

Another view of the center connector is shown in Figure 3, which also shows a measurement of the shunt capacitance (C_p) across the feed point introduced by the Budwig and the cable shield. The center wire of the cable connecting the fitting to the choke was open-circuited so only the capacitance of the fitting and the outside of

Figure 3 — Shunt capacitance measurement of the center fitting.

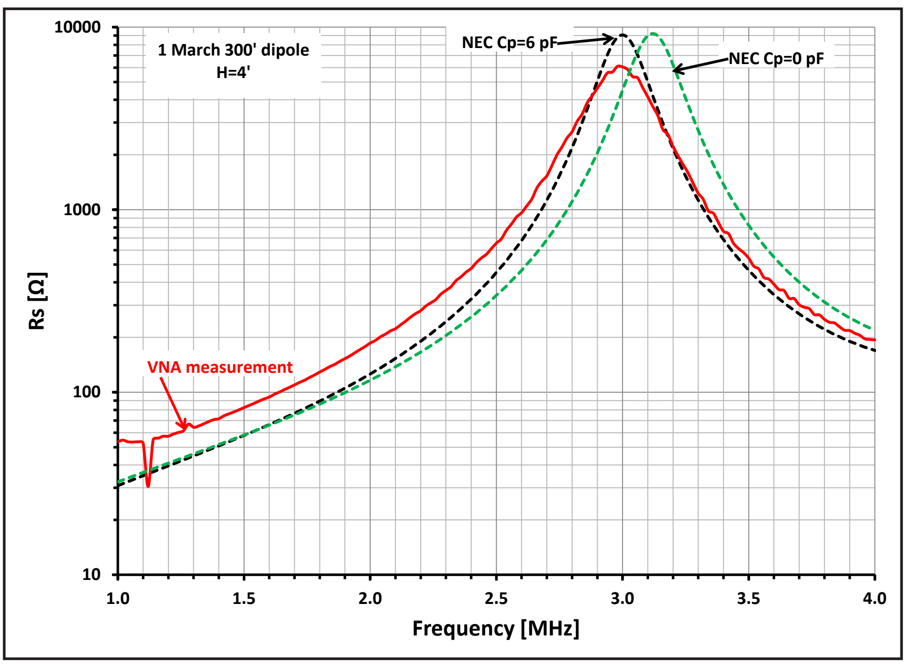

Figure 4 — Modeling with and without C_p.

the cable was included. Shunt capacitance C_p turned out to be about 6 pF, which was added to the model as a capacitive load in parallel with the source. In the 1 to 4 MHz range a shunt capacitance of 6 pF would not seem to matter but, as seen in Figure 4, when added to the model, significantly improved the correlation around the high impedance point.

Figure 5 shows the measured impedance of the common mode choke. While the choke impedance is more than 2 kΩ, at some frequencies the feed-point impedance was even higher. For this reason the graphs show some reduction in measured compared to predicted impedance at the high impedance points.

The measured and computed comparisons of test antenna #1 resistance and reactance are shown in Figures 6 through 17 for heights of 48, 24, 12, 6, 3 and 1 inch above the soil. Note that there are glitches in the VNA measured data around 1.2 to 1.6 MHz on many of the figures. These correspond to local radio station transmissions. These spurious signals are obvious and can be ignored.

NEC4.2 based calculations appear to do a very good job of matching measurements down to 1 inch above ground. I didn't go lower because the soil surface had variations of more than a half inch, and despite weed-whacking closely, there were still grass lumps under the antenna. The zero reactance measurements of Figure 18 show how the resonant frequencies, both series (odd half wave multiple) and parallel (even half wave multiple), vary with height.

Figure 18 illustrates the important point that the resonant frequency goes down in frequency as the antenna comes closer to ground, and that the change is relatively slow until you get to very low heights (less than 3 inches) at which point the change is rapid.

Test antenna #2

The second test antenna was a 40 foot dipole using #26 AWG insulated wire buried 1 inch below ground surface. I wanted to have both series and parallel resonances like I had with the 300 foot dipole but that wasn't possible over the 1 to 4 MHz range so I settled for a 40 foot length that was resonant at about 2.5 MHz. The length of test antenna #2 is 1/9 the length test antenna #1 but we still have a series resonance frequency comparable to the 300 foot above-ground dipole. This observation reinforces the message in Figure 18, that placing the antenna close to or in the soil drastically and rapidly decreases the resonant frequency. As shown in Figure 19, I cut a slot in the soil with a lawn edger. I then inserted the antenna and backfilled the slot with compacted dirt.

After inserting the wire into the slot but before backfilling it, I measured the

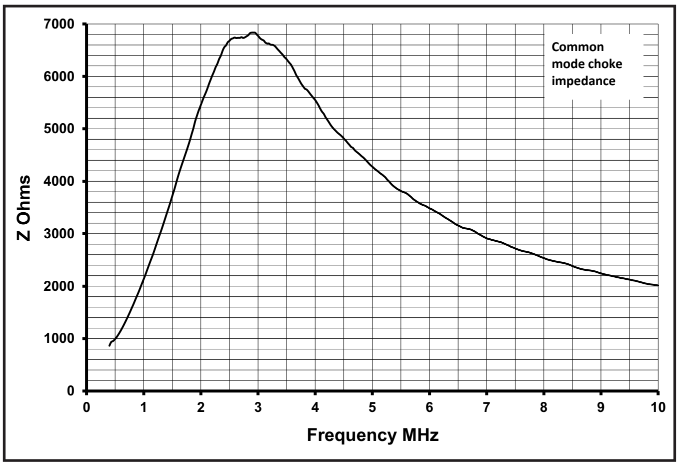

Figure 5 — Measured impedance of the common mode choke.

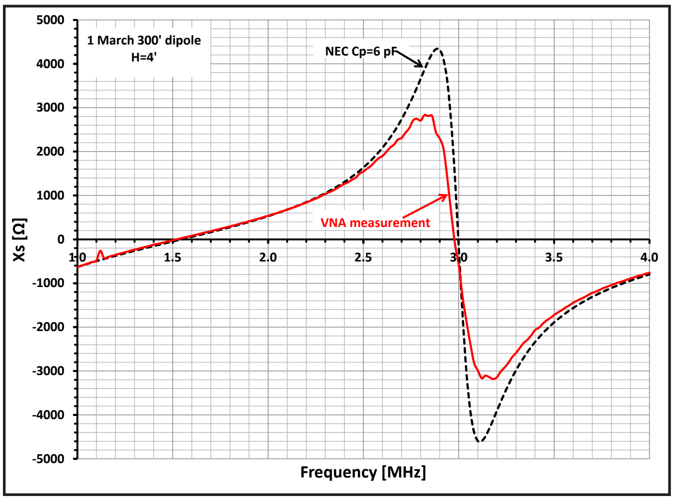

Figure 7 — Reactance measurement at antenna height of 48 inches.

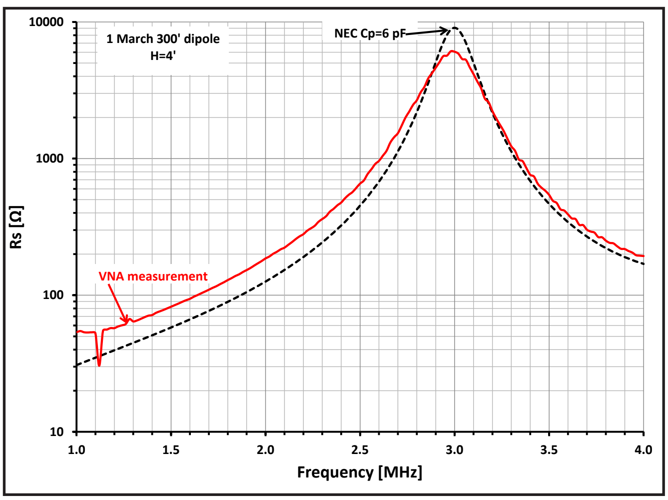

Figure 6 — Resistance measurement at antenna height of 48 inches.

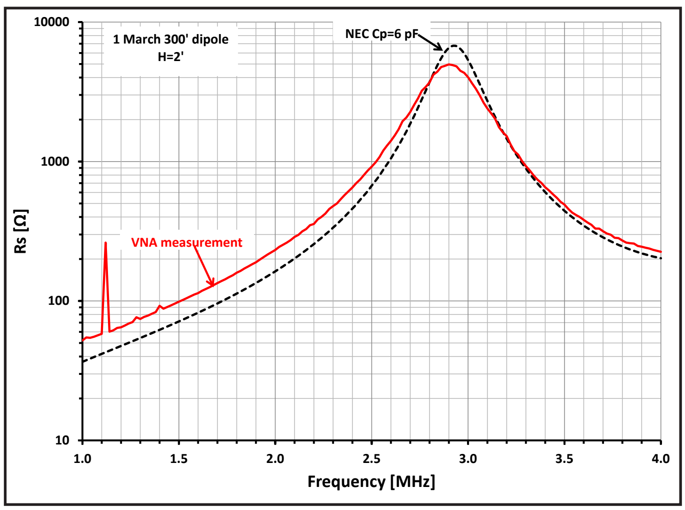

Figure 8 — Resistance measurement at antenna height of 24 inches.

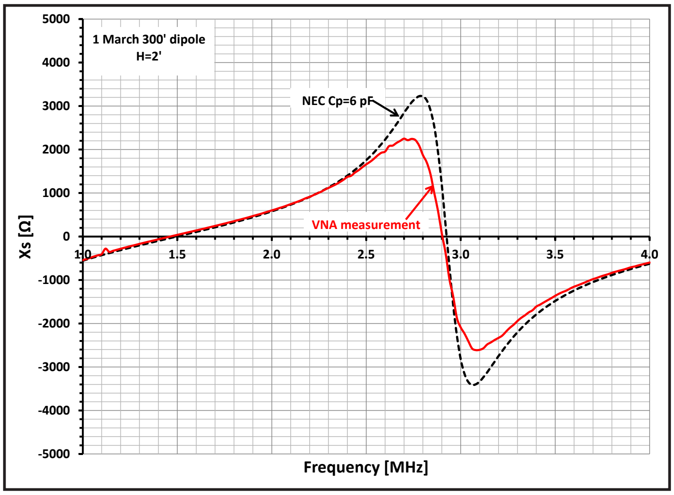

Figure 9 — Reactance measurement at antenna height of 24 inches.

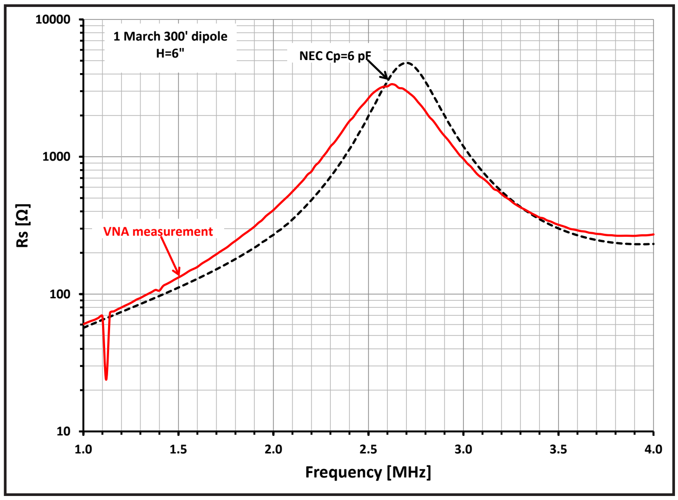

Figure 12 — Resistance measurement at antenna height of 6 inches.

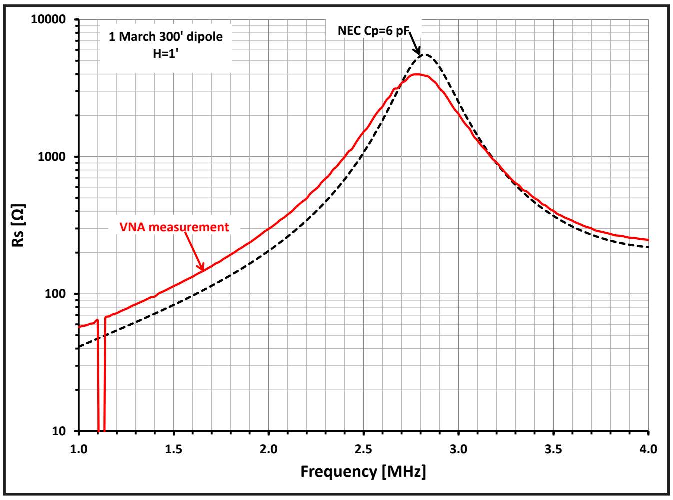

Figure 10 — Resistance measurement at antenna height of 12 inches.

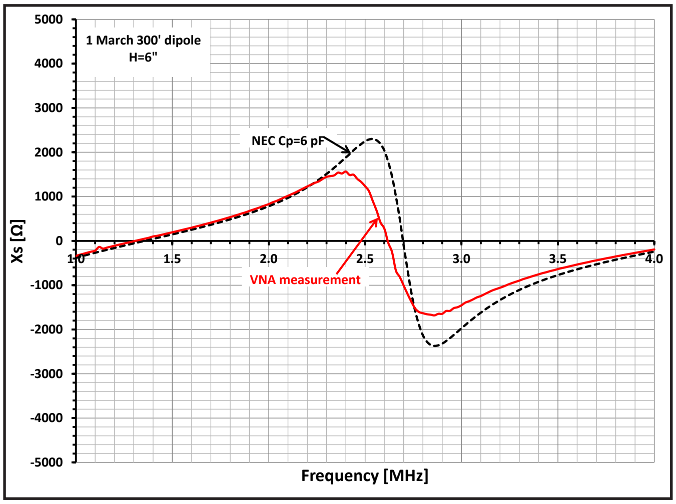

Figure 13 — Reactance measurement at antenna height of 6 inches.

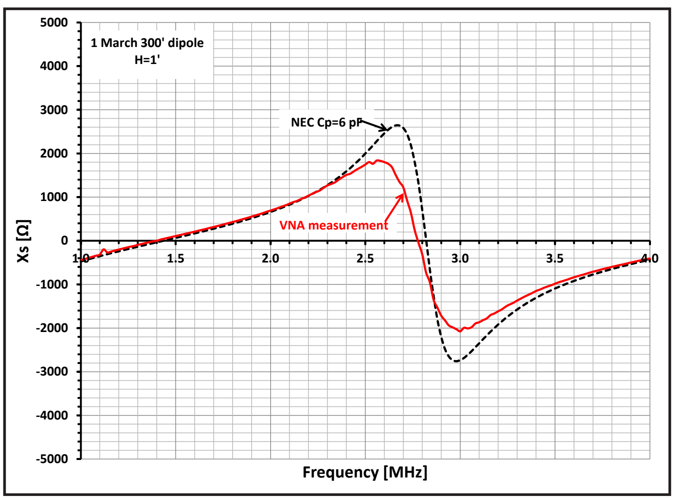

Figure 11 — Reactance measurement at antenna height of 12 inches.

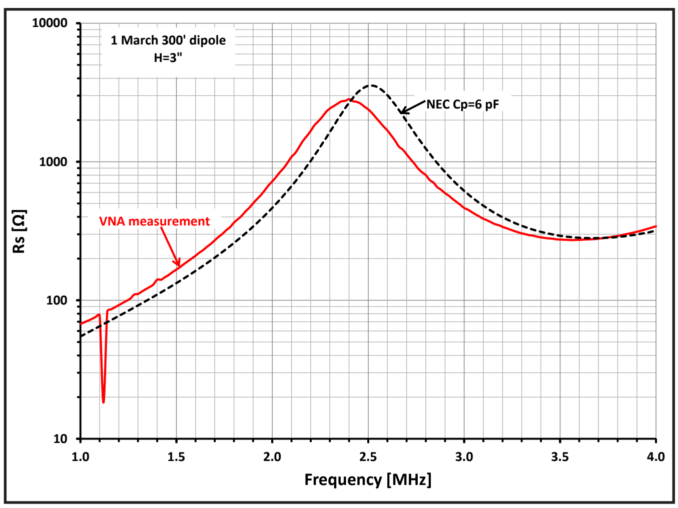

Figure 14 — Resistance measurement at antenna height of 3 inches.

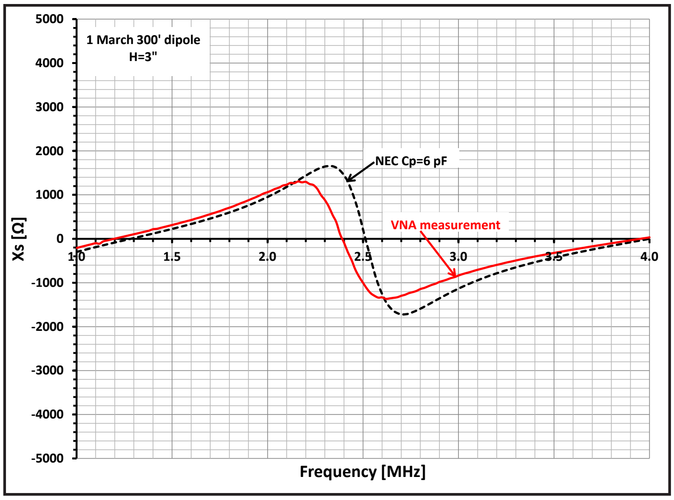

Figure 15 — Reactance measurement at antenna height of 3 inches.

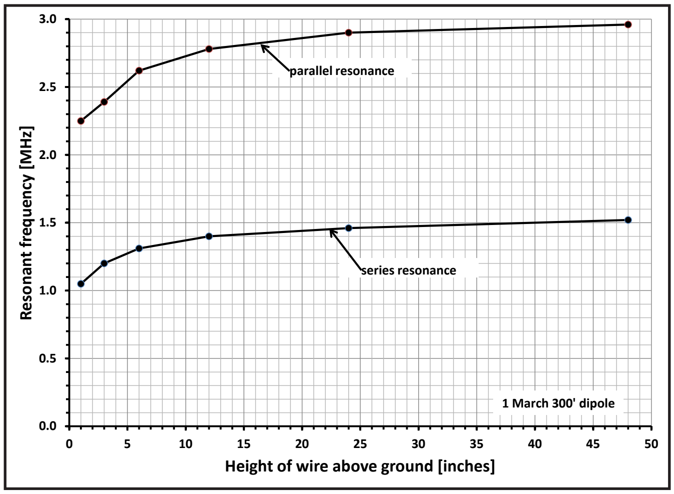

Figure 18 — Resonance variation with height for the 300 foot dipole.

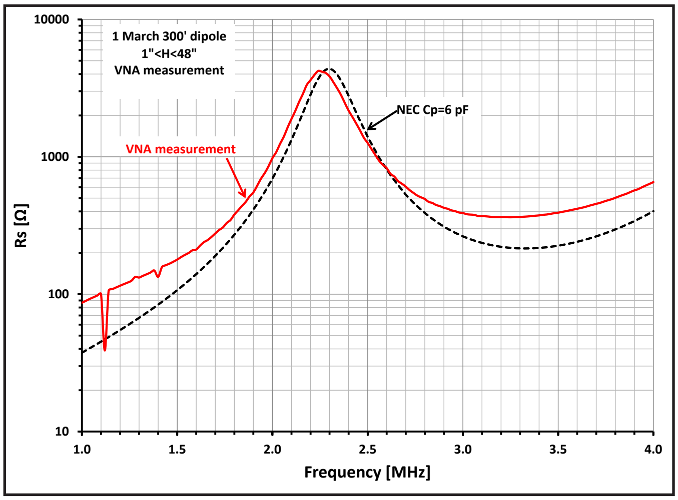

Figure 16 — Resistance measurement at antenna height of 1 inches.

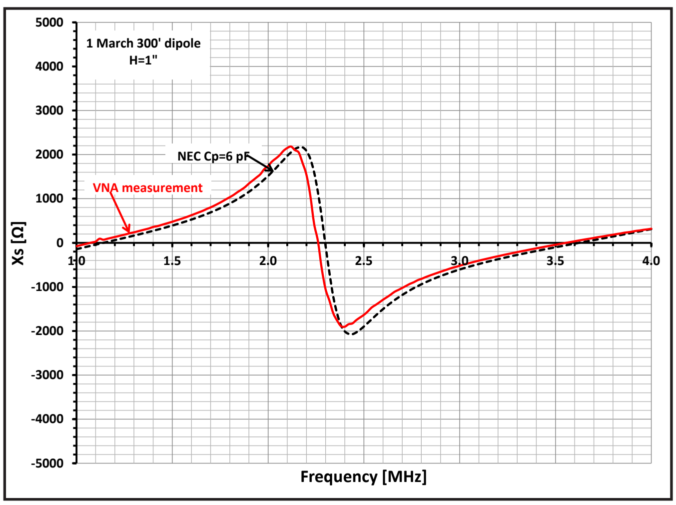

Figure 17 — Reactance measurement at antenna height of 1 inches.

Figure 19 — Cutting a slot in the soil for the 40 foot buried dipole.

impedance. The result was very different from the NEC-based calculation for a buried antenna, and instead behaved as though the antenna were lying on the surface. However, as soon as I backfilled the soil slot and re-measured the impedance, I obtained the results shown in Figures 20 and 21. The good agreement in Figures 20 and 21 between measurements and calculations indicates the NEC model provides reasonable predictions.

I tried both a 19-inch monopole probe and a 12-inch open wire line probe (OWL) to measure the soil characteristics.[3,4,5] The monopole probe gives a good estimate of the average soil characteristics from the surface down to three feet or so. The OWL probe, on the other hand, measures a cylinder of soil just 12 inches from the surface. Figures 22 and 23 illustrate the differences in measurements between the two probes in the same soil.

I felt the OWL data was more appropriate for a wire buried only 1 inch deep. OWL measured values yielded better correlation with modeled values.

Because soil measurements are not perfect, I wondered just how sensitive the model was to variations in the soil characteristics. I reran the VNA measurement of the buried dipole nine days later after it had rained. A comparison between the two measurements is shown in Figures 24 and 25. After the rain, soil moisture was higher, which increased significantly in both conductivity and permittivity, and lowered the resonant frequency from 2.4 to 2.2 MHz.

We can get a feeling for the sensitivity of the modeling to variations in soil electrical characteristics by taking a soil measurement and varying the values ±10% as shown in Figure 26. This example illustrates why good soil measurements are needed to get reasonable correlation, at least for antennas with wires close to or buried in soil.

The sensitivity of modeled resistance calculations is shown in Figure 27 for variations of the insulation relative dielectric constant, and in Figure 28 for insulation thickness. The choices for insulation thicknesses in Figure 28 were not random. The wire used for the antenna had an insulation thickness of 0.008 inches marked on the reel label, however my actual measurements, using a micrometer, of the total outer diameter minus the wire diameter revealed that the actual thickness was 0.009 inches. Using the measured value in the model improved the correlation as shown in Figure 28. Figures 24 though 28 illustrate the sensitivity of resistance and reactance of buried wires to different variables, such as the effect of rain, ground constants, insulation permittivity and insulation thickness.

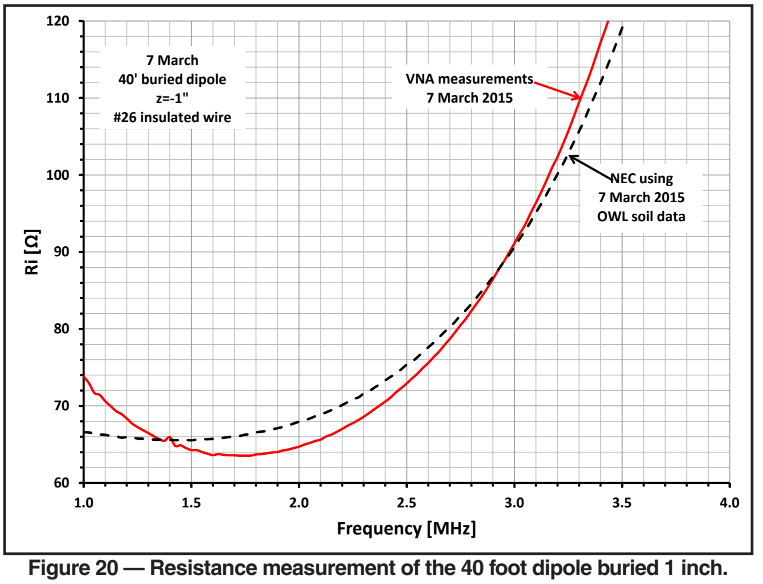

Figure 20 — Resistance measurement of the 40 foot dipole buried 1 inch.

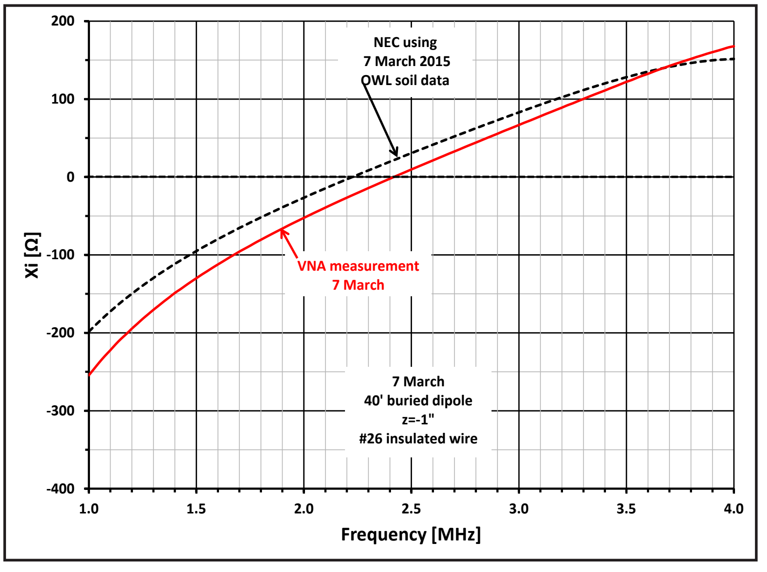

Figure 21 — Reactance measurement of the 40 foot dipole buried 1 inch.

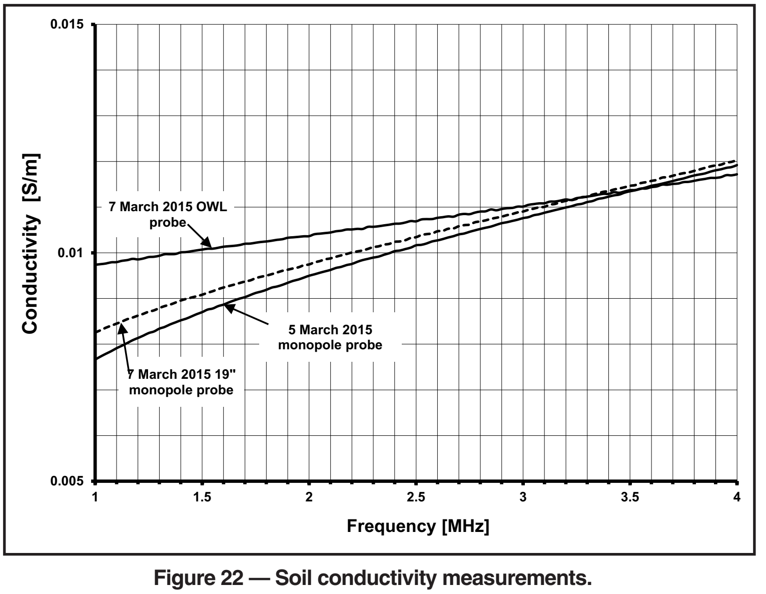

Figure 22 — Soil conductivity measurements.

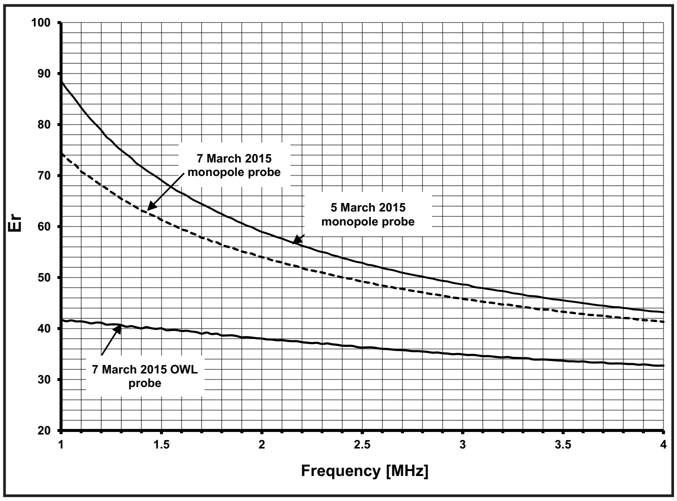

Figure 23 — Soil relative permittivity measurements.

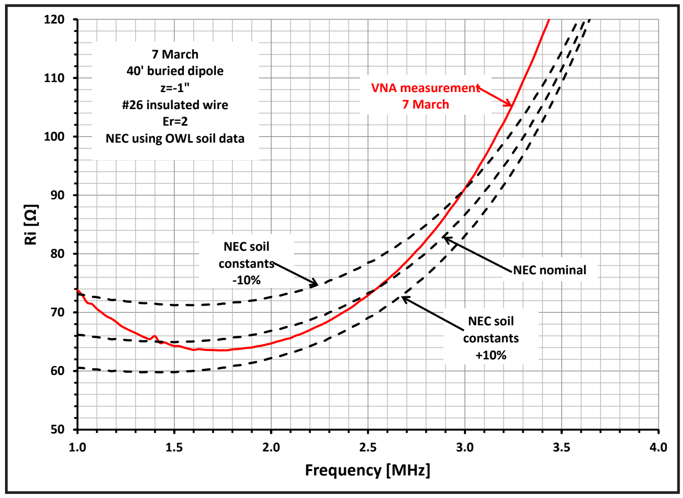

Figure 26 — Variations in modeled resistance for different ground constants.

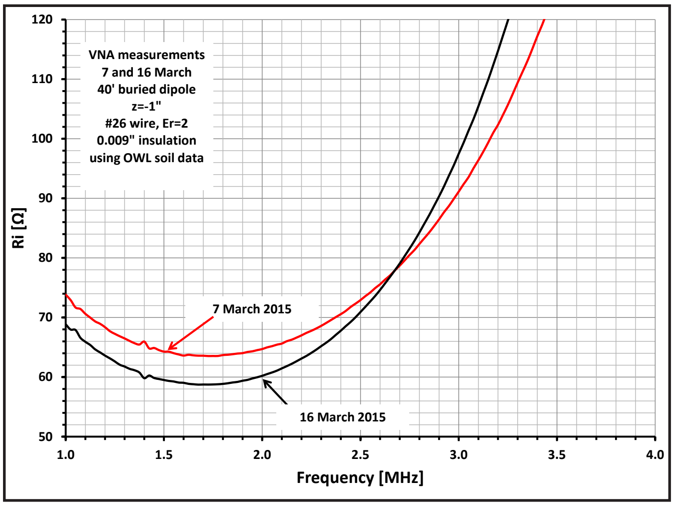

Figure 24 — Resistance measurement of the buried 40 foot dipole on March 7, and on March 16 following rain.

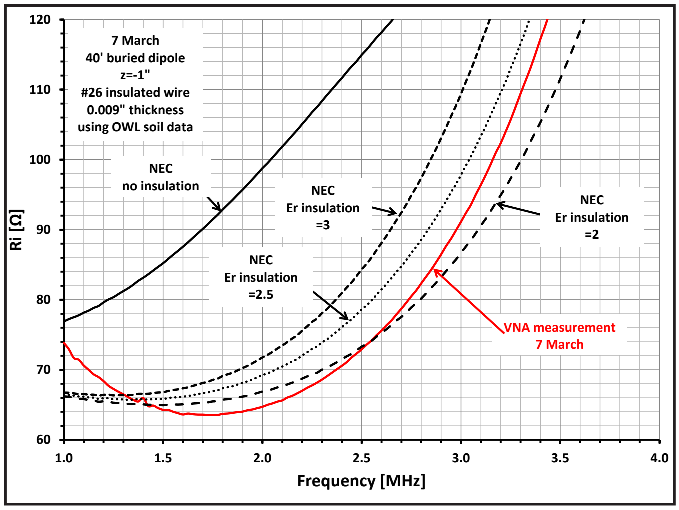

Figure 27 — Effect of wire insulation relative dielectric constant.

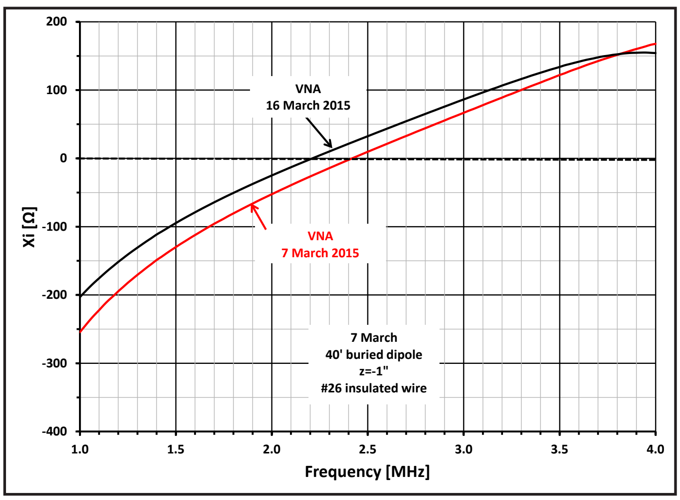

Figure 25 — Reactance measurement of the buried 40 foot dipole on March 7, and on March 16 following rain.

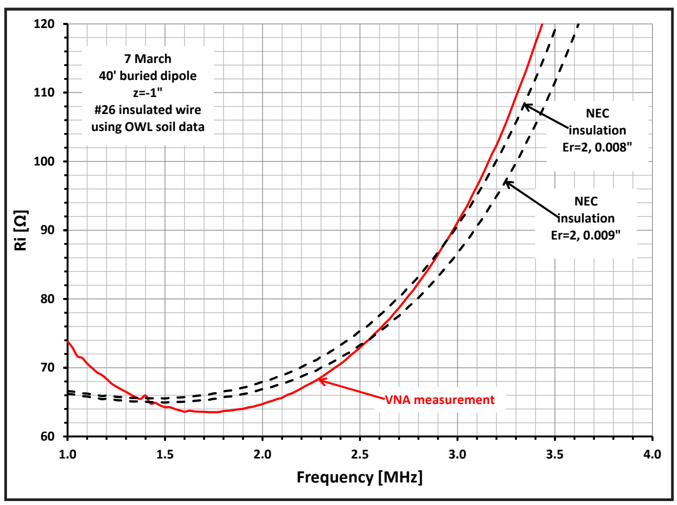

Figure 28 — Effect of insulation thickness.

Test antenna #3

I wanted to test an antenna that incorporated a ground rod, and one that would have a radiation resistance comparable to the loss resistance associated with a rod to get a feeling of how well ground rods are modeled. I have a pair of tall support poles so I simply suspended a 77 foot length of #26 AWG insulated wire from the midpoint of a Dacron line stretched between the poles directly over the ground stake shown in Figure 29. One of the rules for NEC modeling is that a source cannot be on a segment directly adjacent to a wire-size discontinuity. In this case that would be the ground stake to the #26 AWG wire connection. In the model, the source must be in the center of three consecutive segments of the same length and wire diameter. To meet those requirements I used 3-inch segments in the model and placed the source at the center of the second segment (at 4.5 inches), which matched the actual feed point configuration of the test antenna. Using concurrent soil measurements, I got the results shown in Figures 30 for the resistance, and Figure 31 for the reactance.

The overall agreement between measurements and calculations is good, and the resonant frequency is particularly close. The noise introduced into the VNA from local AM broadcast stations picked up by the tall vertical is also obvious. There were other antennas and a metal building within 150 feet of the test vertical, which also introduced some spurious resonances. Unfortunately there's not much I can do about the local AM signals. Their bandwidths are all narrow so I fit a 3rd order polynomial trend line (R^2=0.987) into the VNA data, which pretty well filtered out the noise. The NEC calculation is a good fit to the trend line.

Test antenna #4

This entire exercise had been prompted by a mystery concerning the declining performance of a BOG, and by questions regarding the validity of NEC modeling of BOGs so, appropriately, my final test antenna was a BOG.

Using the 450' BOG already in place I measured the feed point impedance from 400 kHz to 4.4 MHz. I also measured the current amplitude and phase along the wire at 1.83 MHz. I added the current measurements as a further confirmation of the NEC modeling predictions, that is, the rapid exponential decrease in current with distance along the wire. Figures 32 shows the BOG in relation to a measuring tape alongside the wire to locate the sampling points. Figure 33 shows the instrumentation position. Figure 34 shows the probe for

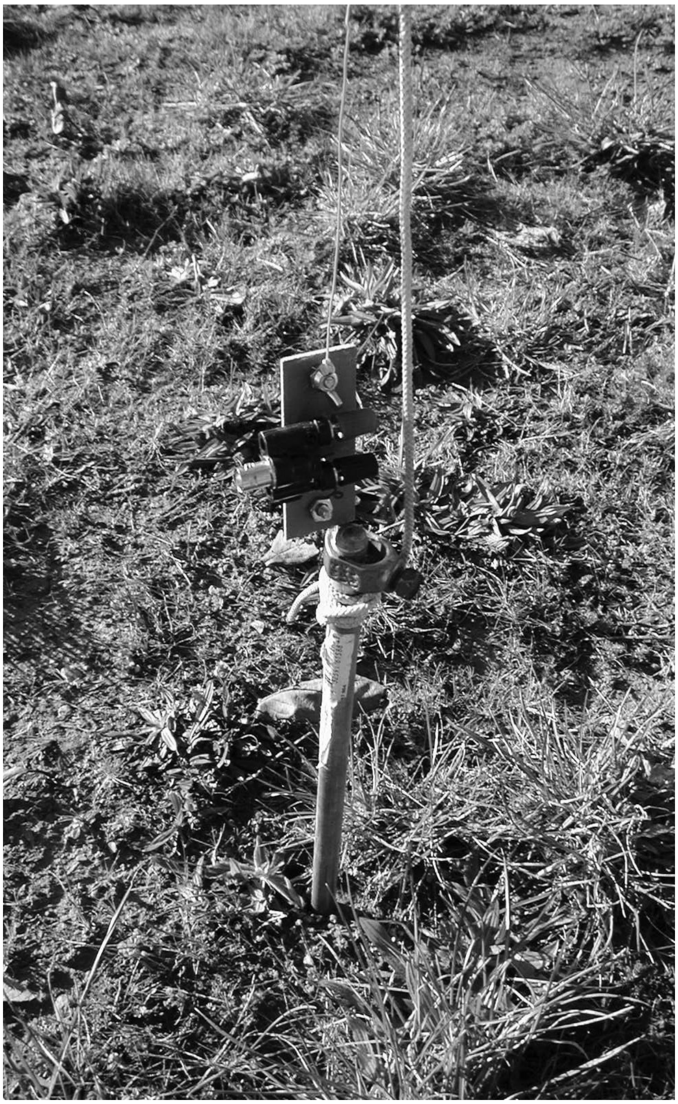

Figure 29 — Feed point and ground rod of test antenna #3.

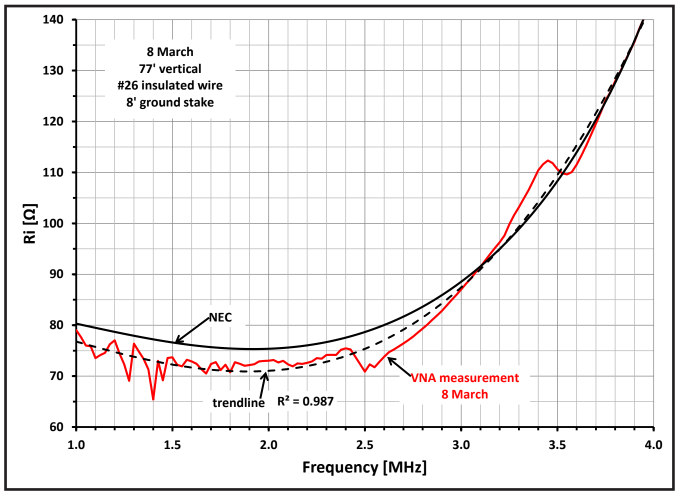

Figure 30 — Measured and computed resistance of the 77 foot vertical with a single ground stake.

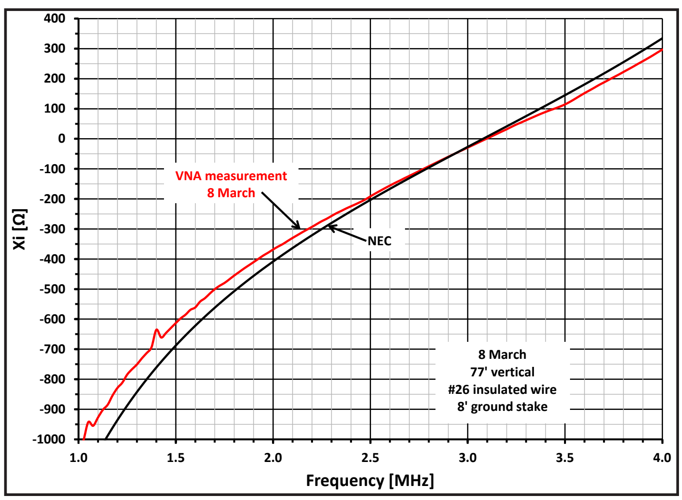

Figure 31 — Measured and computed reactance of the 77 foot vertical with a single ground stake.

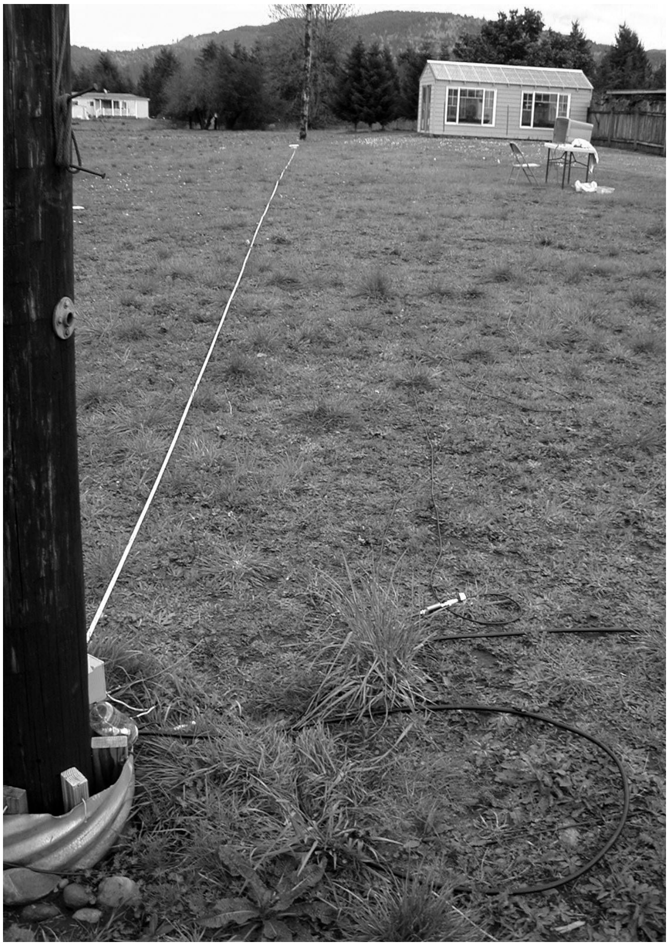

Figure 32 — View of the BOG with measuring tape.

Figure 33 — Instrumentation position.

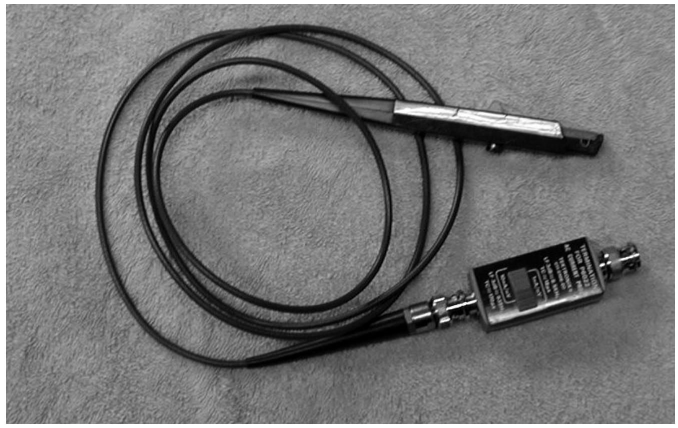

Figure 34 — Scope probe used for current pickup.

Figure 35 — Base excitation and current sampling example.

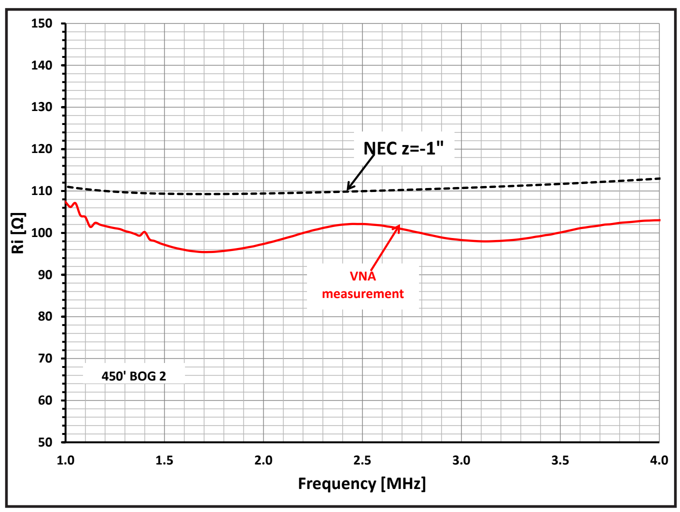

Figure 36 — Measured and computed BOG resistance.

picking up the antenna currents. Figure 35 shows the excitation point at the base, and a current sampling example. For the current measurements, the VNA was in the transmission mode where the antenna was excited at the feed point and the transmission gain (S21) was sampled at several points along the wire using the oscilloscope current probe shown in Figure 34. S21 is a surrogate for the current.

The antenna was modeled one inch below the soil. Modeling results and comparisons to the VNA measurements are shown in Figure 36 (resistance), Figure 37 (reactance) and Figure 38 (current amplitude). The impedance and current distribution graphs show good correlation between NEC and the real antenna despite the uncertainties in the ground surface transition zone.

The rapid exponential decay of the antenna current was a surprise, but the field measurements confirmed it. This goes a long way towards explaining why the antenna performance was so poor. Functionally it behaves more like a short radial than an antenna! Disconnecting the ground rod at the far end had no effect on either the current distribution or feed point impedance, which was no surprise since there was very little current at the far end of the antenna.

Next, I modeled the BOG with the antenna wire one inch above and one inch below the soil to approximately represent the changes from the time it was first installed to the present. The radiation patterns are compared in Figure 39.

I think antenna patterns of Figure 39 solves the initial mystery! The larger pattern with receive directivity factor (RDF) of 12 dB and peak gain *Gp* of -21.47 dB represents the initial condition of the antenna. The smaller pattern with an RDF of 6 dB and *Gp* of -37.4 dB is the present condition of the BOG. These patterns make it clear just how severely the performance was declining as the BOG gradually sank into the sod and soil through two winters. At the time of the measurements spring had arrived and the grass was growing rapidly. The pattern differences shown in Figure 39 agree well with S/N comparisons made over the past 18 months.

Insulated wire

One of the small mysteries was the observation that placing the dipole loosely in the ground slot — which was quite narrow — without packing it with soil had much less affect on the antenna impedances than when the soil was packed around it. One way to explore this is to model a buried dipole as if it were inside a hollow pipe. We can do this with NEC by setting the insulation parameters σ=0 and ε_r=1, that is, air insulation. We can then vary the radius of the insulation from 0.001 to 3 inches as shown in Figure 40.

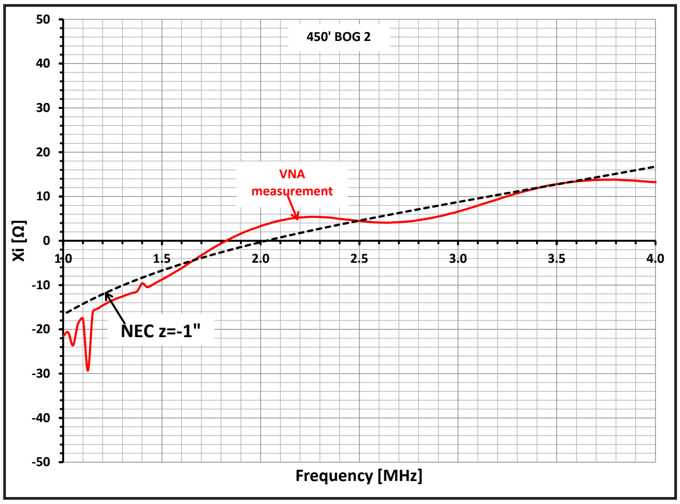

Figure 37 — Measured and computed BOG reactance.

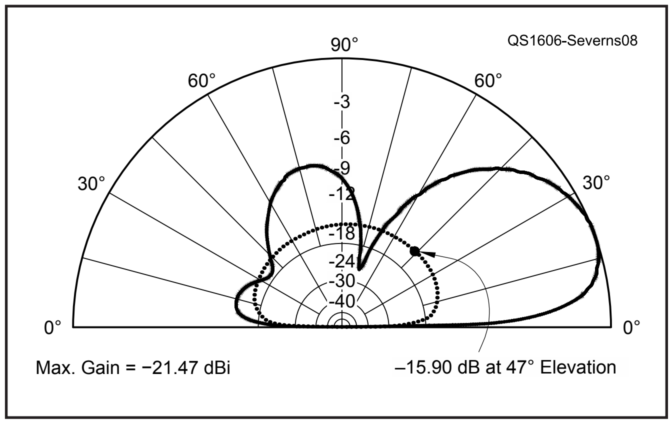

Figure 39 — Computed elevation antenna patterns for the BOG one inch above and one inch below ground.

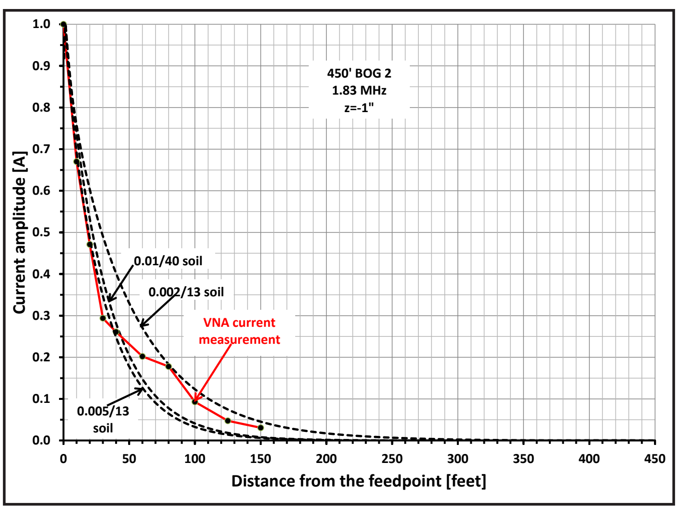

Figure 38 — Measured and computed BOG current amplitude.

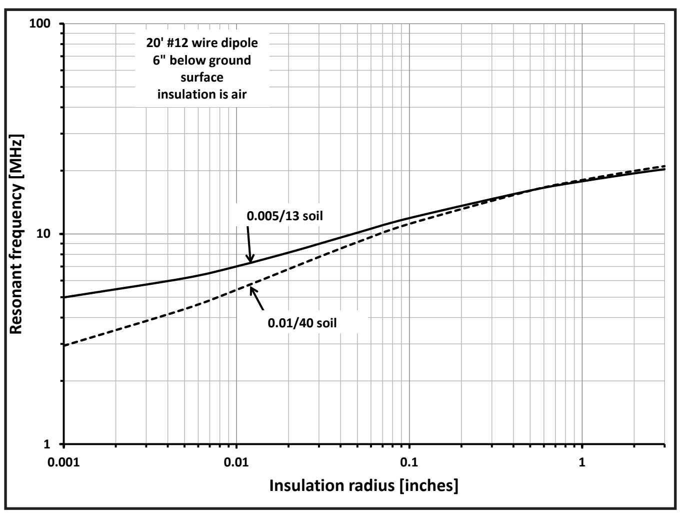

Figure 40 — Resonance frequency in two different soils for different air insulation thickness.

What we see is that even a very thin layer of air around the wire will rapidly increase the resonant frequency. In effect, laying test antenna #2 directly into the soil slot resulted in a layer of air around the wire except at a few points where it was resting on the soil. This also affects test antenna #4, the BOG. The vegetation had grown up gradually around the wire so that it was embedded in the weeds and sod with very little air gap. The same wire BOG centered within a small diameter plastic pipe would behave quite differently. Buried Beverages in plastic pipes?

Conclusions

In the four examples, correlation between measurement and modeling was excellent. These do not by any means represent all the possibilities but the antennas chosen cover a range of practical examples using very low or buried wires.

Based on this work I believe that if we use NEC4.2, and follow the NEC modeling guidelines closely, make sure the model is dimensionally as close as possible to the actual antenna, and make careful soil measurements, then NEC modeling will give reliable results. The practical limitations of NEC4.2 modeling are not due to computational shortcomings in the NEC code. What limits us is our knowledge of the details of the actual antennas and the associated soil characteristics and our ability to replicate these in a model.

As a practical matter we can never be perfect, but modeling should get us close. I think we can use NEC to compare elevated radials and buried radials, both insulated and non-insulated, with reliable results.

There are many other questions we can ask, like what happens when interlaced elevated radials are used in vertical arrays. I think that NEC should give reliable results. The results for Beverage antennas, both elevated and buried with resistor and ground rod terminations should also be reliable.

In the case of the BOG the news is bit ambiguous. NEC modeling demonstrates that the BOG antenna can work very well, and from my experience I agree. However, your results may vary. High conductivity soil, for example, may result in very low signal levels. If the BOG is slowly being covered by whatever grows around it or falls from the sky, you may experience significant degradation in performance over time. As always, buyer beware!

Acknowledgements

I express my sincere appreciation to both Roy, W7EL, and Dan, AC6LA, for the use of their latest software. Without these tools the study would have been impractical. My thanks to Don, N4DJ; Greg, W8WWV, and Carl, K9LA, for reading and helpful comments on draft versions of this article.

Photos courtesy of the author.

Rudy Severns, N6LF, was first licensed as WN7AWG in 1954. He is a retired electrical engineer, an IEEE Fellow and ARRL Life Member.

Notes

[1]Several versions of *EZNEC* antenna modeling software are available from developer Roy Lewallen, W7EL, at **www.eznec.com**.

[2]*AutoEZ* automates use of *EZNEC*, see **www.ac6la.com**.

[3]Rudy Severns, N6LF, "Experimental Determination of Ground System Performance for HF Verticals", *QEX,* in seven parts, Jan/Feb 2009 pp 21-25 and pp 48-52, Mar/Apr 2009 pp 29-32, May/Jun 2009 pp 38-42, Jul/Aug 2009 pp 1-3, Nov/Dec 2009 pp 19-24, Jan/Feb 2101 pp 18-19.

[4]Rudy Severns, N6LF, "An Experimental Look at Ground Systems for HF Verticals", *QST* Mar 2010 pp 30-33.

[5]Rudy Severns, N6LF, "A Closer Look at Vertical Antennas With Elevated Ground Systems", *QEX*, Part 1 Mar/Apr 2012 pp 32-44, Part 2 May/Jun 2012 pp 24-33.

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Conductivity of Trees at HF

N6LF publishes his measurements of tree dielectric parameters.

The effect of trees on HF antennas has been a very long running discussion in the amateur community with little resolution or hard data. During the 1960s and 1970s much work was done for the military on propagation through jungle forests, but much of this work was for frequencies above 50 MHz, so it didn't really answer the questions. In the February 2018 edition[1] of *QST* Kai Siwiak, KE4PT, and Richard Quick, W4RQ, took a serious look at this using NEC modeling [as well as infinite cylinder analytical modeling — *Ed.*] to quantify the impact of trees on vertical radiators, which it turns out can be significant. The article is a real step forward.

Electrical Parameters of Trees

A critical part of the analysis is a determination of the electrical characteristics of trees, that is, their conductivity (S/m) and relative permittivity ε_r. After reading their article I realized that I had already performed measurements on both coniferous (Douglas fir) and deciduous (big leaf western maple) trees, which might help. In 2007 I had a 3-element vertical array[2] on 160 m located in a dense fir forest where the trees were conveniently approximately $\lambda/4$ high and close to the antenna, within 50 ft, well within the near-field. While the array seemed to work okay I wondered just how much I was losing to the forest so I made some measurements on actual trees.

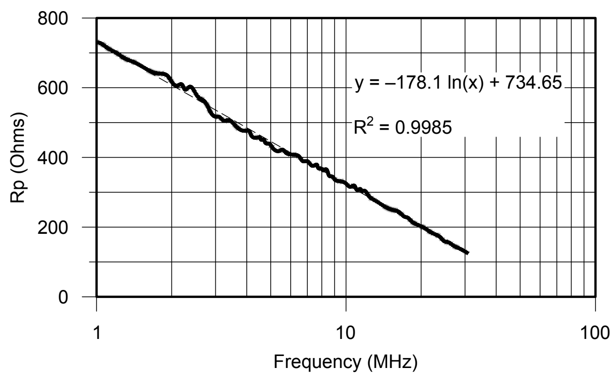
I assumed that the primary loss would be from the longitudinal E-field, that is, the vertical polarization, and that a tree could be viewed as a cylindrical vertical impedance which could be measured experimentally. For the experiments I drove a series of nails approximately 2 inches long, connected with a wire to form two rings about one foot apart as shown in Figure 1. The impedance between the two rings was measured using a vector network analyzer (an N2PK VNA). Measurements were made on Douglas fir — diameter at the inner bark of 10 inches — and big leaf maple — 8 inch diameter — trees in late March when the sap was up.

One problem when using a VNA is the need to properly calibrate out the effect of the cable and leads to the two rings, to isolate the impedance of the tree between the two rings. For the open-circuit, short-circuit, load calibration procedure I used a plastic trash can as shown in Figure 2.

The trash can diameter was about the same as the trees being measured. The interconnected nails in each ring were inserted into holes. The open-circuit calibration is shown in Figure 2, for the short-circuit calibration I used 6 parallel wires distributed symmetrically around the trash can each end

Figure 1 — Impedance measurement on a Douglas fir tree.

Figure 2 — Calibration test fixture shows the AIM4170 as the VNA, but the same fixture was used to calibrate the N2PK vector network analyzer that was used for the actual measurements.



QX1805-Severns03

Figure 3 — Fir tree equivalent parallel resistance, *Rp*, second run, 25 Mar. 2007.

connected to the ring. For the load calibration I inserted resistors in series with these wires with a total parallel resistance of 50 ohms.

Test Results

In the first test I connected a dc ohmmeter between the rings. What I noticed immediately was the resistance changing slowly over time much like what you see when checking an electrolytic capacitor for leakage current. The sap of the tree is an electrolyte so that behavior was not a surprise. For the impedance measurements I assumed a parallel *Rp Cp* equivalent circuit. Samples of typical measurements are given in Figures 3, 4, and 5. The general behavior was much the same for both the fir and the maple trees.

The conductivity and permittivity, as a function of frequency, appear to behave very much like soil[3]; conductivity (σ) goes up with increasing frequency — *Rp* goes down — and ε_r goes down with increasing frequency to a point where it flattens out (*Cp* is a function of ε_r).

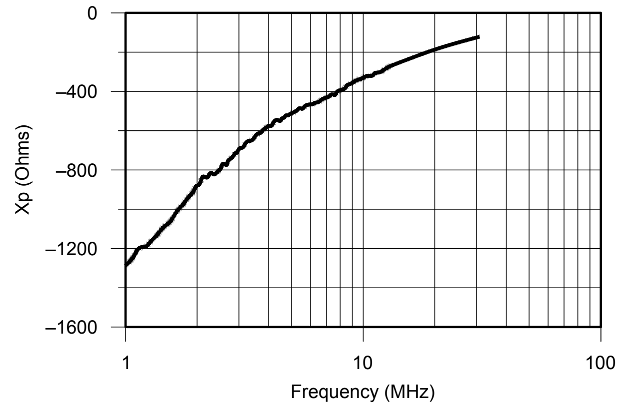
I made an estimate of σ from the equation for a resistor:

$$\sigma = \frac{L}{(Rp)(A)}$$

where *L* is the 12 inch (0.3048 m) distance between rings; *A* is the effective cross sectional area in square meters.

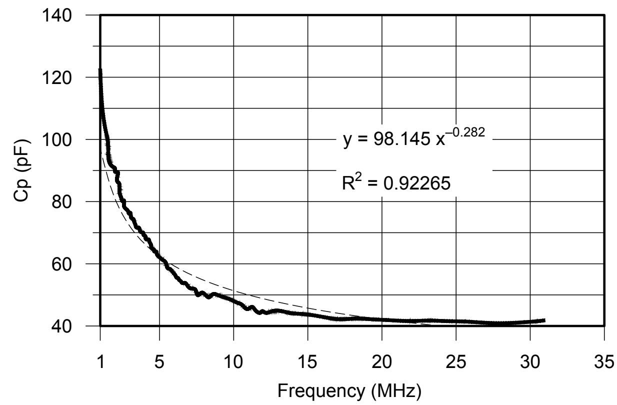
Determining the cross sectional area, *A*, is a bit tricky. If you assume the conduction is limited to the cambium, a thickness of about 0.125 inches (0.003175 m), and the diameter is 10 inches (0.254 m), then $A = 0.00253$ m^2. From Figure 3, *Rp* is about 325 Ω at 10 MHz. This gives $\sigma = 0.37$ S/m, which seemed pretty high! However, that number is based on a 1/8” conduction layer. Kai, KE4PT, sent me an extract from a book on wood characterization by Bucur[4], which indicates that the characteristics across the entire diameter do not vary greatly, at least for the case of young trees with little or no heartwood. If the wood across the diameter also conducts, then the calculated conductivity is lower. For example, for a diameter *d* of 10 inches (0.254 m), $A = 0.016$ m^2, $\sigma = 0.059$ S/m. This gives a range of conductivity at 10 MHz of about 0.06 to 0.4 S/m. The actual average conductivity is likely somewhere in between.

At this point in my 2007 experiments I found it hard to believe such high values for tree conductivity. Because I did not have any backup from other sources for my measurements I have been reluctant to publish this work. However, in the February 2018 *QST* article the authors assume[5] $\sigma = 0.17$ S/m, which lies within the range of my measurements. Their value was derived from extensive earlier



QX1805-Severns04

Figure 4 — Equivalent parallel impedance, *Xp*, second run, 25 Mar. 2007.



QX1805-Severns05

Figure 5 — Equivalent parallel capacitance, *Cp*, second run, 25 Mar. 2007.

work in the professional literature so I now have some faith in my measurements. The only additional comment I would add is that the values of conductivity used in the NEC model should include the variation with frequency (dispersion) so clearly shown in my measurements.

I think at this point we can use NEC modeling with some confidence to estimate the effect of trees on HF antennas. Unfortunately that effect appears to be substantial and not a good thing!

Rudy Severns, N6LF, was first licensed as WN7AWG in 1954. He is a retired electrical engineer, an IEEE Fellow and ARRL Life Member.

Notes

[1] K. Siwiak, KE4PT, and R. Quick, W4RQ, “Live Trees Affect Antenna Performance”, *QST* Feb. 2018, pp. 33-37.
[2] Rudy Severns, N6LF, “A 3-Element 160 Meter Vertical Array”, NCJ May/June 2009, pp. 12-13.
[3] Rudy Severns, N6LF, “Measurement of Soil Electrical Characteristics at HF”, *QEX* Nov/Dec 2006, pp. 3-9.
[4] Voichita Bucur, **Nondestructive Characterization and Imaging of Wood**, Springer Series in Wood Science, Chapter 7.
[5] Tree data; D. Tomasanis, “Effective Dielectric Constants of Foliage Media,” RADC-TR-90-157, Interim Report AD-A226 269, Jul. 1990.

Another Way to Look at Vertical Antennas

What's the difference between a dipole and a vertical? Maybe not as much as you think. Come along and try another point of view.

By Rudy Severns, N6LF

The grounded vertical is one of the earliest radio antennas, well known to Marconi and widely used today by amateurs, particularly for 80 and 160 meters. VHF verticals with "ground planes" are also popular. Traditionally, ground has been viewed as an integral part of the antenna—in effect supplying the "missing" part of the antenna, since, at low frequencies at least, the vertical portion of the antenna is usually less than λ/2. Even when the antenna is not grounded, but raised above ground, we still use the terms "elevated ground system," "counterpoise ground," "ground plane" and so on. In this view, we retain the concept that ground is an integral part of the antenna and an ungrounded vertical must have some structure that replaces the "real" ground. While this conceptual framework has served us well for over 100 years, it tends to limit our thinking to more traditional solutions. A change in viewpoint exposes useful variations, better suited for particular applications.

The traditional view, stemming largely from the work of Brown, Lewis and Epstein[1] in the 1930s, is that a λ/4 vertical, with a ground system of 100 or more long radials, is the ideal—anything else is an inferior compromise.

Recent work,[2,3] using primarily *NEC* modeling, has indicated that elevated ground systems with only 4 to 8 λ/4 radials can be very competitive with the more-traditional 120-buried-radial antenna, although that is the subject of some controversy, due to the difficulties experienced with experimental verification. There is even the heresy that radials as short as λ/8 may be only marginally less effective than full λ/4 radials and have significant practical advantages. Elevated-radial systems have their own drawbacks, such as (1) nonuniform radial currents,[4] which lead to asymmetrical patterns and perhaps increased loss, and (2) the need for an isolation choke at the feed point. A network of wires, arranged in a circle λ/2 in diameter and suspended above ground, may be more trouble than simply burying the wires. There has been considerable

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[1]Notes appear on page 32.

discussion—regarding traditional λ/4 radials used in elevated ground systems—as to whether these are a poor choice or not and whether other arrangements may be superior.[4, 5, 6, 7, 8]

Because most amateurs are severely limited by available space and the cost of towers and extensive ground systems, the traditional buried-radial or even the elevated λ/4-radial systems are frequently infeasible. What is needed is a wide range of other choices for the antenna structure from which to choose the best compromise for a given situation. Obviously, the final design should sacrifice as little performance as possible.

An alternate way to look at verticals has been suggested by Moxon (see Note 5) and others:

1. The antenna is a shortened (less than λ/2) vertical dipole with loading. The loading may be symmetrical or asymmetrical, lumped or distributed, inductive or capacitive, or a combination of all of these. Usually, the loading contributes little to the radiation, although some loading structures may radiate.

2. Ground is *not* part of the antenna. However, the interaction between ground and the antenna—and the loss in the ground—must certainly be taken into account. This includes both near and far fields.

This view can the maintained even when a portion (or all!) of the antenna is buried.

At first glance, this seems a trivial conceptual change. Nonetheless, looking at a vertical as a short, loaded dipole in proximity to ground—rather than as a grounded monopole—opens possibilities not usually considered with the more traditional point of view. For example, with a full λ/4 vertical, one would not normally consider adding a top hat for loading. However, in so doing, the diameter of an elevated ground system at the base of the antenna can be drastically reduced, seemingly out of proportion to the size of the top loading hat. This can be a very real advantage by reducing the footprint of the antenna. A shortened, horizontal dipole antenna with a hat at each end is very well known; it draws little comment. Nevertheless, vertically orienting the antenna and manipulating the end-loading devices to suit the application is not so common—although the antennas are conceptually identical!

Loaded Dipoles in Free Space

One of the simplest ways to resonate a shortened dipole (less than λ/2) is to add capacitive elements or “hats” at the ends, as shown in Fig 1. As indicated, the feed point may be anywhere along the radiating portion of the antenna. Fig 1 shows symmetrical end loading. Fig 2 shows extreme asymmetrical loading, where only one capacitive loading structure is used. This is, of course, the familiar groundplane antenna being viewed as an asymmetrical dipole. Actual antennas can vary between these two extremes, since they incorporate various sizes and geometries of loading hats to suit particular applications.

When the vertical portion of the antenna, *h*, is less than λ/4, top loading is commonly employed. However, top loading is usually not considered when $h \geq \lambda/4$. This may be due to our past view that we need an extensive set of buried radials, or equivalently, an elevated system of λ/4 radials. For a λ/4 vertical, the diameter of the radial system will be ≈λ/2, changing only slowly as the number of radials is varied. On the other hand, if we lengthen the vertical section beyond λ/4, add some top loading or even some inductive loading, the diameter of the bottom radial structure drops rapidly.

A simple example illustrating this point is given in Figs 3 and 4. Fig 3 shows an asymmetrical λ/4 dipole with two radials (L1 and L2) at each end. L2 is varied from zero to 22.3 feet, and L1 is readjusted, as needed, to resonate the antenna at 3.790 MHz.

Clearly, adding even a small amount of top loading (L2) greatly reduces the length of the bottom radials (L1), and consequently the land area required

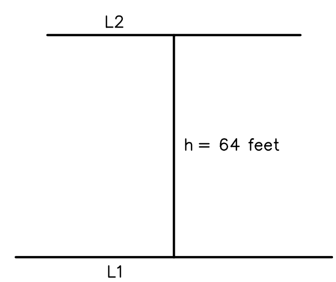

Fig 3—1Asymmetric two-radial dipole. F_R = 3.790 MHz.

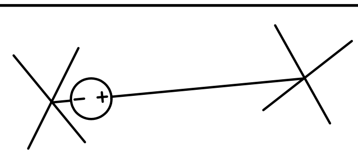

Fig 1—Short loaded dipole.

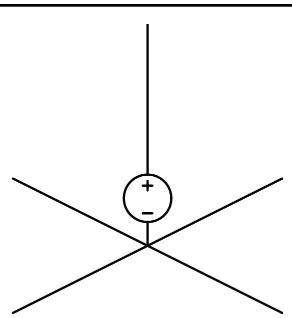

Fig 2—Asymmetrical dipole.

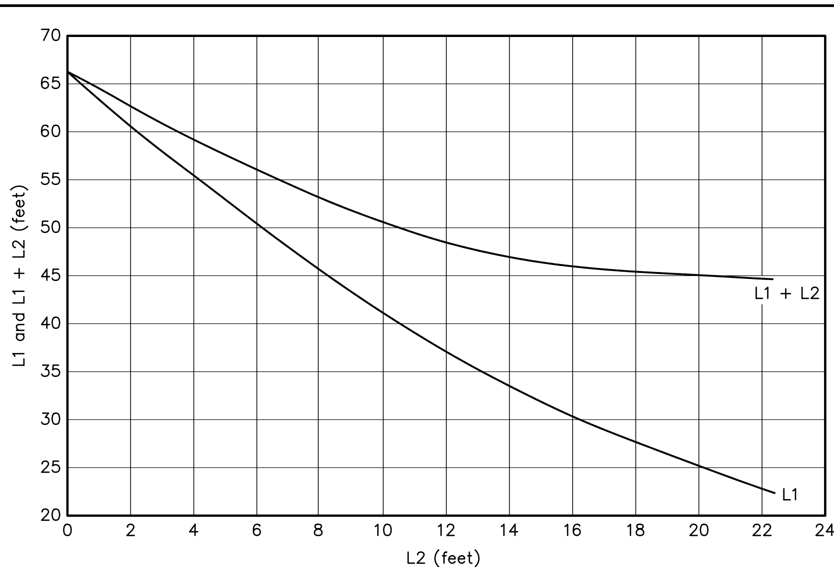

Fig 4—Effect of top loading on radial length.

for installation. This is a matter of considerable practical importance to those with restricted space in which to erect an antenna. With somewhat more complex loading elements, the footprint can be reduced even further.

In addition to greatly reducing the length of the radials, a number of other things happen during the above exercise:

1. With only two radials and no top loading, the radiation pattern varies with azimuth by about 0.7 dB, making the pattern slightly oval. This pattern asymmetry essentially disappears as the radials (L1) are shortened with top loading (L2).

2. When placed over ground, the currents in individual λ/4 radials are rarely equal. This can lead to asymmetric patterns and increased loss. The current asymmetry rapidly decreases as the radials are shortened.

3. The peak gain, and the angle at which it occurs, changes relatively little as top loading is added and bottom radials shortened while keeping the vertical section the same length.

4. Small amounts of inductive loading could also be used to supplement or even replace the top loading. As long as the vertical section is close to λ/4, the radial lengths can be reduced to λ/8 without seriously increasing losses.

Modeling Issues

The realization that everything—from the length of the radiator to the type and distribution of loading—is a potential variable that may be adjusted to achieve specific ends, is a very liberating idea. Unfortunately, it brings its own set of problems. Which variations are best for a given application? A multitude of questions arise when judging any particular variation.

The large number of possibilities and questions cannot be dealt with analytically, at least beyond an elementary level. The only practical way to deal with the many variables is to systematically explore the possibilities with *NEC*, *MININEC* or other CAD modeling software. Yet, even that is not a simple matter. Each modeling program has particular strengths and weaknesses that affect its use for this problem. The bottom portion of a vertical for 80 or 160 meters is usually very close to ground (less than 0.05 λ). For these applications, the modeling software should implement the Norton-Sommerfeld ground and properly model the current distribution in the lower part of the antenna as modified by induced ground currents. Only *NEC 2* and *4* do this. Of course, if the lower part of the antenna is buried in the ground, only *NEC 4* is suitable.

Loading structures may consist of a web of wires with multiple wires at each junction, perhaps of different diameters, and with small angles (less than 90°) between adjacent wires attached to the same node. *MININEC*-based software can model multiple acute angles if segment tapering is used, but if many wires are used in the structure, the number of segments becomes quite large. *MININEC* Broadcast Professional, using a different segment-current distribution, does an even better job without the need for tapering. However, both of these programs do not model the interaction properly for very low antennas over real ground. *NEC 2* can model the ground effects correctly, but may not handle the multiple small angles properly, especially if different diameter conductors are connected. *NEC 4* is much better in this respect, but is not widely used by amateurs because of its expense.

Real grounds are frequently stratified beginning only a few feet down. On 160 meters, the skin depth is of the order of 15 to 20 feet, and it is common to have several layers with different electrical properties over that distance. Even in homogeneous ground, the effect of rain and subsequent drying creates a varying conductivity profile. None of the presently available software addresses this problem. The validity of *NEC 2* or *4* modeling for ground has been questioned because of differences between experimental measurements and predictions made by modeling. This is a critical issue. If NEC is fundamentally deficient with regard to ground modeling, then the comparisons to date between buried-radial and elevated-radial systems are invalid. That includes the work reported in this article! On the other hand, NEC modeling may be fine, but the problem lies with the highly variable nature of real ground. This is particularly so down to depths of 15 to 20 feet, which cannot be simulated with NEC, but that could greatly modify experimental results. Some support for this view comes from experimental work at higher frequencies. There the skin depth is much less, and modeling predictions are in much better agreement with experiments.

The presently available software, while a remarkable achievement, is not totally satisfactory to fully exploit the possibilities. The suggested point of view brings this out. A great deal of care must be used when modeling a vertical with a complex loading system near ground.

A Design Example

The advantages of employing a different conceptual approach can be illustrated using the 160-meter vertical used at N6LF, where an effective antenna was built on a very difficult site at low cost.

The site is on a narrow ridge—approximately 60 feet wide at the top—in a forest. There is no possibility of installing an extensive buried radial system because of the dense forest, heavy underbrush, steep slopes and very large old-growth stumps. Even an elevated system of normal size, about 260 feet in diameter, is not practical.

A support for the antenna was constructed from three Douglas fir trees, fastened together to form an **A** frame (see the sidebar "A Large **A**-Frame Mast, Inexpensively" for details). This resulted in a support 135 feet high. Allowing eight feet from the bottom of the antenna to ground and a few feet of slack at the top for sway in high winds, the final vertical length is 120 feet—very close to λ/4. Because the antenna is located over 700 feet from the shack, 75 Ω Hardline coax (a freebee from the local CATV company) is used for the transmission line. The antenna was designed to have a 75-Ω feed-point impedance to match the transmission line. The feed-point impedance at the junction of the lower hat and the vertical wire was manipulated by adjusting the relative sizes of the bottom-hat and top-loading wires. Alternately, I could have used a larger hat on the bottom and moved the feed point up into the vertical part of the antenna, but this was not done because of the limited space available for the bottom hat. I also tried some inductive loading at the base and at the junction of the top-loading wires. Relatively small amounts of inductive loading—with very little additional loss—would further reduce the size of either or both of the capacitive loading elements. I did not keep any inductive loading because sufficient space was available for the arrangement shown.

The final antenna is shown in Fig 5. There are four radials at the bottom, connected by a skirt wire at the ends. The diameter of this bottom-loading structure is only 40 feet, compared with 260 feet for normal λ/4 radials. Two sloping wires are used for loading at the top. A sloped top hat may not be optimal when compared to horizontal wires: The radiation resistance is somewhat lower. Nevertheless, this arrangement is very simple and allows the antenna to be tuned by changing the angle of the wires with the vertical portion of the antenna. This can be done from ground level by shifting the attachment points for the guy lines supporting the sloping wires.

Christman's comparison (see Note 2) between a 120-buried-radial vertical and an elevated four-radial vertical (both with $h = \lambda/4$) indicates that the gain and radiation-pattern differences between the antennas are quite small: 0.35 dB for peak gain, 1° for peak gain angle. Because the difference is so small, I have chosen to use the four-radial elevated antenna as the reference antenna, since it is much easier to model than a complete 120-buried-radial antenna.

Using *NEC4D* for modeling, radiation-patterns for a four-radial ground-plane antenna and this antenna were compared. The result is presented in Fig 6. The model assumes ground of average electrical characteristics under the antenna ($\sigma = 0.005$ S/m; $\varepsilon = 13$). The wire used was #13 copper, and its loss was included in the modeling. The price paid for drastically reducing the diameter of the bottom loading structure is a peak-gain reduction of 0.5 dB. This is a fair trade for dramatically easing the installation of the lower loading element because 0.5 dB will probably not be detectable in actual operation. In the real world, where full-size ($\lambda/4$) radials very likely have varying currents (see Notes 4 and 8), the smaller antenna may not, in fact, be inferior at all. In this particular example, full-size radials would need to zigzag down a steep hillside at various angles. It is very doubtful they would have been any better than the small hat that was adopted.

Any antenna with an elevated radial system needs an isolation choke (common-mode choke, or *balun*, if you prefer) on the transmission line near the feed point. One effect of moving the loading from the bottom to the top of the antenna is to increase the potential between the feed point and ground. This requires more inductance in the isolation choke to properly decouple the transmission line. For this application, I happened to have a roll of ½-inch Hardline. The roll was about two feet in diameter, so I simply expanded it into a coil three feet long and two feet in diameter with a simple wood framework to hold it in place. Fig 7 is a photo of this king-sized decoupling choke.

The result was a choke with 350 μH of inductance (4 kΩ at 1.840 MHz). When this value of inductance was placed in the model with a buried transmission line, there was still some interaction; resonance was displaced downward. This was also found true on the actual antenna. This illustrates one of the drawbacks of very small bottom-loading structures: A choke with enough inductance to avoid interaction may not be practical, at least on 160 meters. Since the current in the choke is relatively small, additional losses due to ground currents will not be very large. The Q of the choke, however, must be high to limit losses in the choke itself.

The monster balun shown here is extreme and not required. A much smaller choke could be used. The large structure was used because it was actually very convenient with the materials on hand.

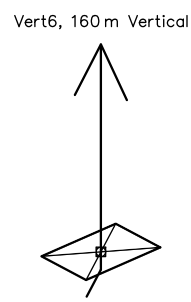

Fig 5—Antenna configuration.

Fig 7—Rudy and the "small" decoupling choke.

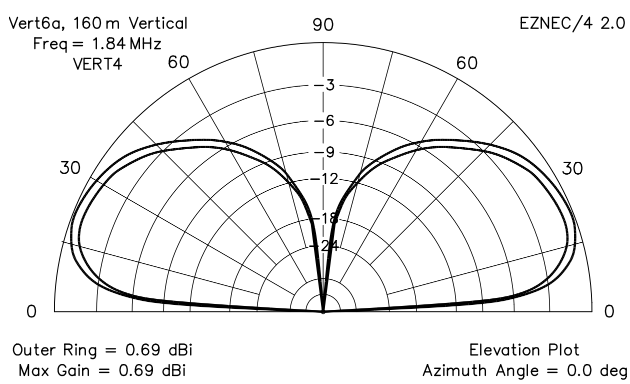

Fig 6—Comparative radiation pattern.

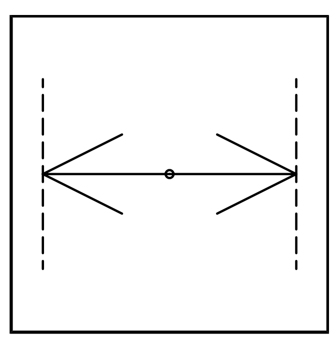

Fig 8—Flat versus drooping loading wires.

A Large A-Frame Mast, Inexpensively

A λ/4 vertical is about 70 feet tall on 80 meters, and 130 feet on 160 meters. Getting this height with a tower can be expensive. I needed a less-expensive alternative. In the Pacific Northwest, fir trees with heights greater than 100 feet are common, and can usually be purchased locally and inexpensively if they are not already growing on your property. In the southeastern US, there are extensive pine forests which, while not typically as tall as the firs, can be used in the same way. I have many tall Douglas Fir trees on my property, so I selected three of them, two with 12-inch diameter bases and one of about 8 inches. I trimmed the top off the two larger trees at a point where they were about five inches thick. This gave me two poles approximately 80 feet long. Since I was only going to support a wire vertical, I topped the smaller tree at a point where it was roughly two inches thick. This gave me a pole 60 feet long. I was trying to have the cross-sectional area at the top of each large pole roughly equal to the area at the base of the smaller pole when they were overlapped.

The next step was to drag the poles to the antenna site and assemble the A-frame shown in Fig 9:

1. I bought a large, used railroad tie and cut it in half at the middle of its length. I then buried each half vertically with about 18 inches above the ground to form a pivot post. I placed the posts about 10 feet apart.

2. I placed the two large poles, side-by-side, midway between the two posts.

3. I placed the smaller pole on top of the two large poles—overlapping by about five feet—and lashed the three poles together using #9 galvanized smooth iron fence wire as indicated in Fig 9C. To begin the lashing, I stapled the end of the wire; as I applied each turn, I tightened it with a claw hammer. After 15 turns or so, I stapled the free end.

4. I then spread the butt ends of the large poles out to the pivot posts. [Did you use a team of mules, or just your burly "pecs"?—*Ed.*] This spreading tightened the lashings very nicely (!) so that the three poles were solidly connected.

5. I wanted to raise and lower the A frame at will and keep the pole ends away from soil contact (rot!). Therefore, I created a pivot at each post by drilling a 2-inch-diameter hole through the post and pole butt. I then inserted a length of 1.5-inch galvanized iron water pipe as the shaft for the pivot. To keep the pipe from slipping out, I put a pipe cap on each end as a retainer.

6. The next step was to attach *two* halyards (one spare, just in case!) to the top of the mast. I used two small pulley blocks—the kind typically used on sailboats—and then rove a length of black, sun-resistant, $^{3}/_{8}$-inch Dacron line through each block. The lines were long enough to form a continuous loop reaching the ground, so I could hoist or recover the antenna at will.

7. Finally I erected the **A** frame. In my case, I used a nearby tree as a gin pole (suitably guyed!!) along with three steel blocks and a long length of wire rope. Hoisting power was supplied by a small tractor. I took great care because of the forces involved. The initial lift required a pull of over 1000 pounds and the **A** frame weighs over a ton. (Green trees are heavy!) If I were more patient, I could have allowed the trees to dry out (months!), which would have greatly reduced the weight.

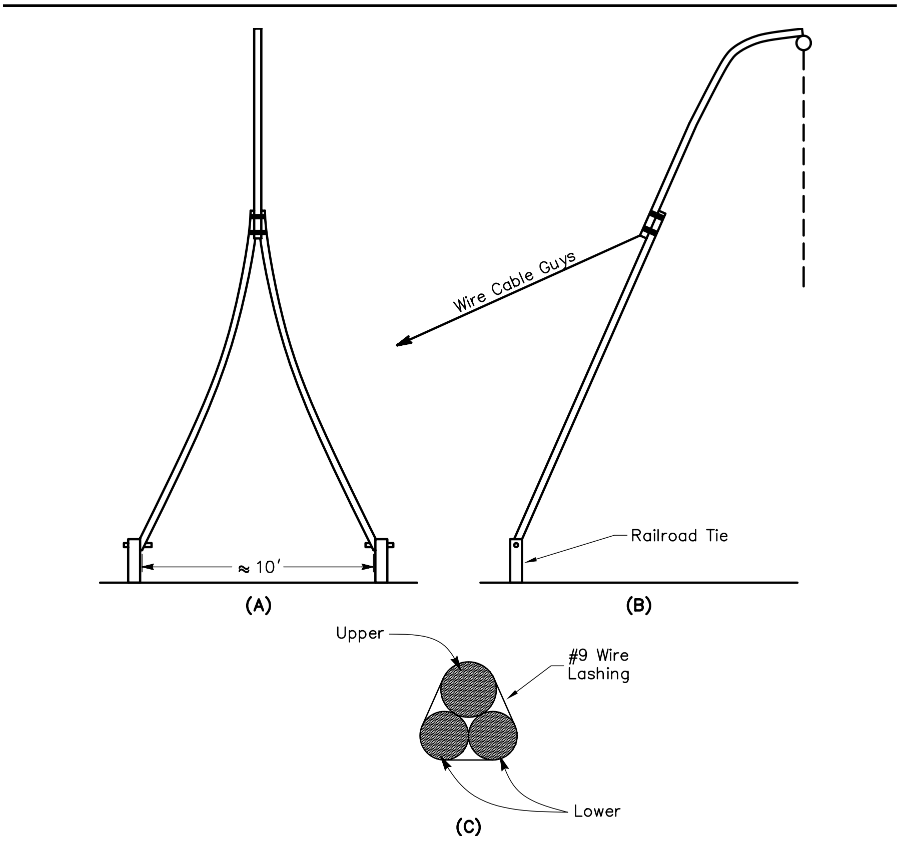

Fig 9

I choose not to raise the mast to a vertical position because I wanted the antenna and the loading structures to stand clear of the mast and any guys. As shown in Fig 9B, I left frame tilted about 15° from vertical and bent the top over like a fishing pole, so it is even farther out from the base. The green pole bent relatively easily, and the bend became permanent when the wood dried out.

I used two wire-cable back-guys, anchored at the junction of the poles, to hold the mast in place. Although the weight of the mast makes it unlikely it would blow over towards the guys, I use the spare halyard as a guy from the top of the mast in the opposite direction to the wire guys. This arrangement minimizes conductors in the near field of the antenna.

The cost of the entire exercise was less than $75, and I expect to get many years of use from the mast. Of course, I had the trees, the tractor and the hoisting tackle, which kept the cost very low.

Table 1—Antenna Comparison at 3.510 MHz

Antenna	*h (ft)*	*L1 = L2 (ft)*	Z_{middle} Ω	Z_{end} Ω	*Peak gain (dBi)*	*Peak angle °*	*Wire loss (dB)*	*2:1 BW (kHz)*
λ/2	137	0	91	>5000	+0.30	16	0.08	270
Lazy-H	120	4.4	96	1096	+0.28	17	0.07	280
Lazy-H	100	10.4	94	384	+0.12	19	0.07	280
Lazy-H	80	17.4	81.3	180	-0.06	20	0.08	260
Lazy-H	69.8	21.6	71.2	127	-0.07	21	0.09	240
Lazy-H	60	26.3	59.7	90.9	-0.15	22	0.10	200
Lazy-H	40	38.3	33.7	40.8	-0.38	24	0.16	140
Lazy-H	30	45.6	21.5	23.8	-0.59	25	0.23	100
λ/4 (2 radials)	69.8	—	—	38.8	+0.11 by-0.39	22	0.15	200
λ/4 (4 radials)	69.8	—	—	35.7	+0.21	22	0.13	175

More Modeling

In the process of developing this antenna, a great deal of additional modeling was performed to explore the effect on performance of different loading arrangements. One of the more interesting variations was a symmetrically loaded, two-radial antenna called a Lazy-H vertical (see Note 6). This antenna is intended to be supported between two trees. The antenna is identical to that shown in Fig 3, except that L1 = L2. Table 1 gives a comparison between a full λ/2 vertical, a λ/4 ground-plane with two and four radials and the Lazy-H with different values of *h* (height of the vertical portion) varying from 120 down to 30 feet. Note that the λ/4 Lazy-H is within 0.3 dB of the four-radial λ/4 vertical and has greater bandwidth. If two supports are available, the Lazy-H is much easier to fabricate than the four-radial version, and has significant size in only two dimensions instead of three. I assumed #13 copper wire and average ground for the models. Z_{end} is the impedance at the junction of the vertical section's lower end and the lower radials. The bottom of all the antennas is assumed 10 feet above ground.

In the 160-meter example given earlier, the top loading structure was simply a pair of drooping wires led to anchor points near ground. The question arises as to the comparison between flat configurations, like that shown for the Lazy-H and the drooping-wire alternative. This question can be quickly answered by modeling an end-loaded dipole in free space with two different configurations as shown in Fig 8. The modeling shows that the drooping wires must be lengthened to achieve resonance, the radiation resistance is significantly lower with drooping wires and the far-field pattern is essentially the same. From a practical point of view, the use of drooping wires greatly simplifies the structure, and has very little effect on the far-field pattern. It may reduce the efficiency of the antenna if the radiation resistance is lowered too much, however. This is the kind of trade-off information critical to a new design.

In general, modeling this class of antennas shows that peak gain and peak-gain angle primarily determined by ground characteristics and the height of the vertical radiator, *h*. The loading means has only a second-order effect on the radiation pattern. A variety of loading arrangements can satisfy a particular situation with little loss of performance—as long as we keep the radiation resistance high enough to control losses.

Conclusions

This article has advocated a different conceptual view of vertical antennas: They can be viewed as loaded dipoles close to ground. Changing the point of view makes it easier to recognize the wide range of options available for configuring a high-performance vertical to meet the needs of a particular site and set of limitations. To assess the many options, we need the help of software. Unfortunately, no available software package provides the desired computational capabilities. Users of any antenna modeling software should be very careful when setting up the model and interpreting results.

Acknowledgement

In addition to the referenced papers, other workers in this field have pointed out the advantages of the point of view presented here. This idea is certainly not the author's creation, although I wholeheartedly endorse it. Moxon's work deserves careful reading. I am indebted to Dr. L. B. Cebik, W4RNL; Dick Weber, K5IU, and Grant Bingeman, KM5KG, for their comments and support.

Notes

[1]Brown, Lewis and Epstein, "Ground Systems as a Factor in Antenna Efficiency," *IRE Proceedings*, June 1937, Vol 25, No. 6, pp 753-787.

[2]A. Christman, "Elevated Vertical Antennas for the Low Bands: Varying the Height and Number of Radials," *ARRL Antenna Compendium*, Vol 5, 1996, pp 11-18.

[3]A. Christman, "Elevated Vertical Antenna Systems," *QST*, Aug 1988, pp 35-42.

[4]R. Weber, "Optimal Elevated Radial Vertical Antennas," *Communications Quarterly*, Spring 1997, pp 9-27.

[5]L. Moxon, "Ground Planes, Radial Systems and Asymmetric Dipoles," *ARRL Antenna Compendium*, Vol 3, 1992, pp 19-27.

[6]R. Severns, "The Lazy-H Vertical Antenna," *Communications Quarterly*, Spring 1997, pp 31-40.

[7]J. Belrose, "Elevated Radial Wire Systems For Vertically-Polarized Ground-Plane Type Antennas," Part 1, *Communications Quarterly*, Winter 1998, pp 29-40; Part 2, Spring 1998, pp 45-61.

[8]D. Weber, "Technical Conversations," *Communications Quarterly*, Spring 1998, pp 5-7 and 98-100.

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Experimental Determination of Ground System Performance for HF Verticals Part 3 Comparisons Between Ground Surface and Elevated Radials

Experimental results from another of the author's antenna experiments.

Over the years there has been a great deal of discussion regarding the relative merits of a vertical antenna with a few elevated radials versus one with a large number of radials either lying on the ground or buried just below the surface. *NEC* modeling predicts that as few as four radials, a few feet above ground, will provide as efficient a ground system as a large number of on-ground radials. Whether this prediction is valid is a matter of some dispute. Resolving this issue is important for amateurs using HF vertical antennas.

The first segment of the experiment was a comparison of the performance of a ¼-wavelength vertical antenna with a large number of ground surface radials (64) to one with only four elevated radials. From the results in segment one it appeared that elevated radial systems for HF verticals have some merit. But there are a number of different ways to implement an elevated radial system. The purpose of the second segment of the experiment was to evaluate the relative performance of several different elevated radial schemes.

Segment One

All measurements were made at 7.2 MHz using a 33.5 foot tubular aluminum vertical antenna. The experiment began with sixty four, 33 foot no. 18 AWG insulated wire radials lying on the ground surface. The antenna was insulated from ground and used a common mode choke (balun) in the feed line. With a height of 33.5 feet and 64 radials, the vertical was close to resonance at 7.2 MHz.

During the experiment, |S21| (magnitude of the transmission gain, see Part I of this series)[1] and the input impedance at the feed point (Z_i) were measured and recorded as the radial system was changed. The experiment began with 64 radials lying on the ground surface. Without changing the height of the vertical, |S21| and Z_i were measured as the radial number was reduced in the following sequence: 64, 32, 16, 8, 4. The next step was to make a series of measurements, beginning with the four radials on the ground and then elevating the radials and the base of the vertical to 6 inches, 12 inches and finally 48 inches. At the 48 inch height, a measurement of the current division between the radials was made.

This entire sequence was repeated three times on different days. The results did not change significantly between test runs.

[1]Notes appear on page 32.

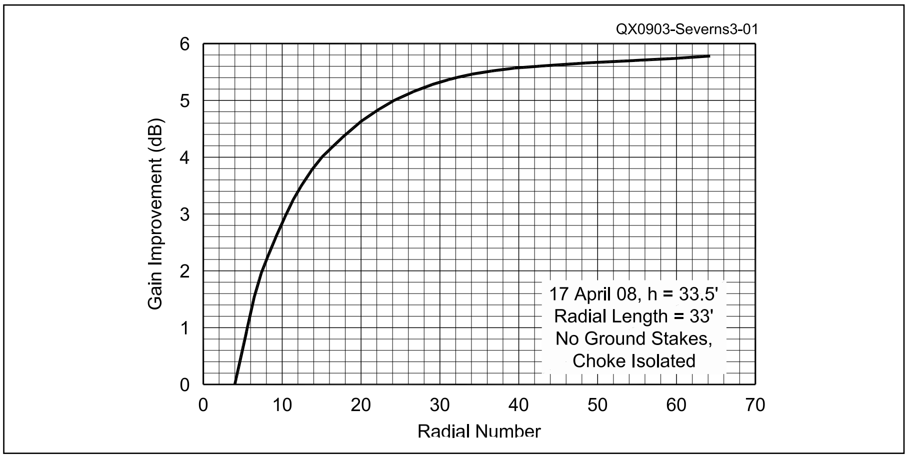

Figure 1 — |S21| as a function of radial number. All radials are lying on the ground surface.

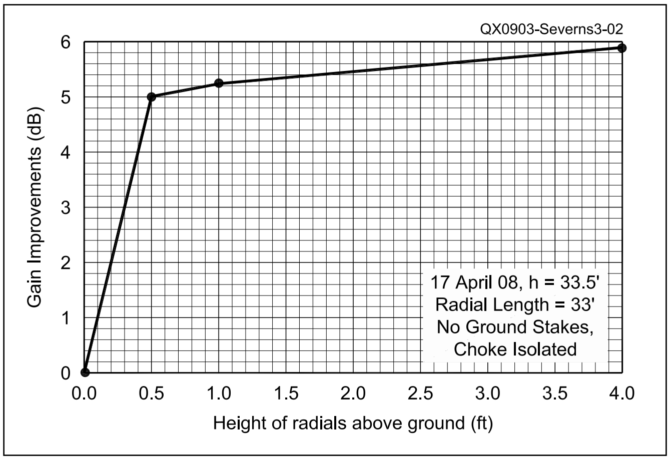

Figure 2 — |S21| with 4 radials and the antenna base at different heights.

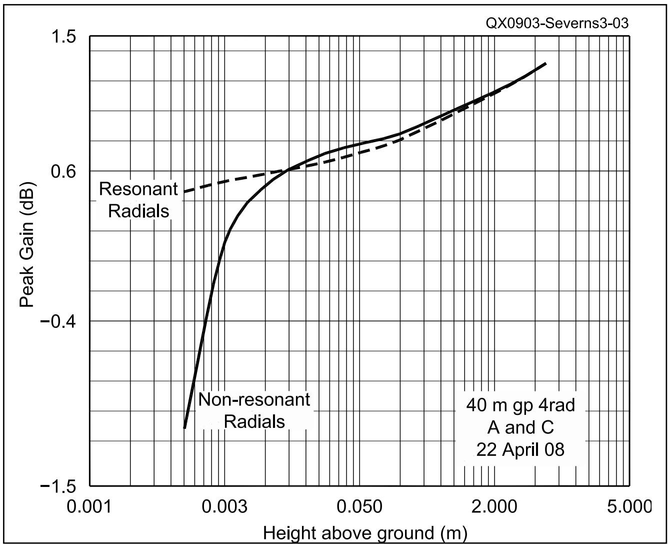

Figure 3— *NEC* prediction of peak gain versus radial height for 4 radials.

Experimental Results

The observed variations in |S21| as radial number and height were changed are shown in Figures 1 and 2. In the graphs, |S21| has been normalized (0 dB) to the value for 4 radials lying on the ground surface, so that the graphs show the *improvement* in dB as either the radial height or number were increased.

From Figure 1, we see that with 64 radials lying on the ground surface |S21| = +5.8 dB. From Figure 2, for four radials and the base of the antenna elevated 48 inches above ground, we see that |S21| = +5.9 dB. The difference is only 0.1 dB. For any practical purpose, the two ground systems are equivalent, which is in accord with *NEC* predictions.

The large change in |S21| with radial number in Figure 1, which is predicted by *NEC*, is mostly the result of additional loss caused by resonances present in sparse radial screens. This effect was discussed in Part 2 of this series.[2]

The very large change between 0 inches and 6 inches in elevation shown in Figure 2 was also predicted by *NEC*. A typical prediction from *NEC* of peak gain versus radial height is shown in Figure 3.

The data line labeled "nonresonant radials" corresponds to constant length (33 feet) radials, which are not shortened to compensate for the effect of the soil characteristics on the radial resonant frequency. The other data line shows the effect of adjusting the length of the radials to re-resonate the antenna as the height above ground is altered.

Typical measured values for Z_i during the experiment are given in Table 1.

The measured current division between the radials, normalized to 1 A of total base current, is given in Table 2.

The radial current asymmetry was small

Table 1
Experimental Values for Feed Point Impedance.

Number of Radials	*Radial Height (Inches)*	*Z_i (Ω)*
64	0	39.7 – *j* 1.2
32	0	42.9 + *j* 2.1
16	0	56.1+ *j* 6.2
8	0	85.5 + *j* 8.0
4	0	137 + *j* 14.9
4	6	43 + *j* 6.4
4	12	40.6 + *j* 0.08
4	48	34.8 – *j* 9.7

Table 2
Current Distribution in the Radials When Elevated to 48 Inches.

Radial Number	*Relative Current (A)*
1	0.235
2	0.271
3	0.247
4	0.247

Table 3
Gain Comparisons With One and Four Radials.

Radial Number (dB)	*Azimuth (Degrees)*	*Peak Gain (dBi)*	*Elevation (Degrees)*	*Delta from 4 Radial Case (dB)*	*Delta from 4 Radial Case*
4	0	+1.15	21.4	0	X
4	0	–1.12	8	X	0
1	0	+0.38	22.8	–0.77	X
1	0	–2.04	8	X	–0.92
1	90	–0.36	22.8	–1.51	X
1	90	–2.79	8	X	–1.67
1	180	–2.19	19.8	–3.34	X
1	180	–4.59	8	X	–3.47

enough to not have any meaningful effect on |S21|. Earlier measurements on radial systems with 64 radials, lying on the ground surface, also showed little asymmetry in the current division.

Effect Of Radial Current Division Asymmetry

As shown by Weber, it is very common for the current division between the radials in an elevated radial system to be unequal, especially if there are only a few radials.[3] This asymmetry can affect the radiation pattern, and may possibly explain some of the variation in earlier comparisons. For this reason, I was very careful to minimize that asymmetry.

To get worst case estimates of the effect of current asymmetry on the pattern, I did some *NEC* modeling. Two models, the first with four radials and the second with one radial, are shown in Figures 4 and 5.

Comparisons between the peak gain and the gain at 8° elevation are given in Table 3. I have shown the peak gain and its associated angle, and also the gain at 8°, which corresponds to the angle to the test range receive antenna. As Table 3 shows, that makes little difference in the magnitude of the pattern distortion.

The worst case signal reduction from the four-radial case is at the 180° azimuth, with one radial. If all the current were in the radial pointing away from the receive antenna, the signal strength would be a bit over –3 dB from the case where all four radials had the same current. I examined models with 1, 2, 3 and 4 radials, but the worst case is for a single radial. That is hardly surprising.

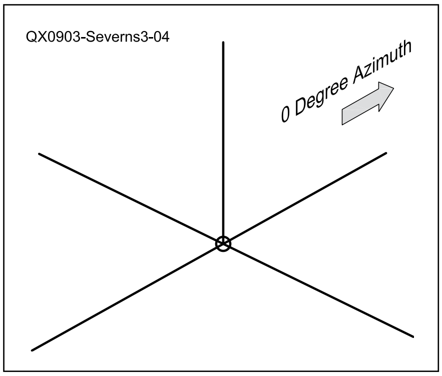

Figure 4 — Four elevated radials, 48 inches above 0.015/30 soil.

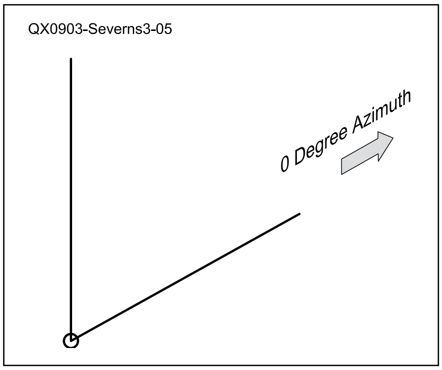

Figure 5 — One elevated radial, 48 inches above 0.015/30 soil.

Segment 2

The "standard" elevated radial scheme has four or more radials elevated above ground by 4 feet to 10 feet, with the base of the vertical antenna also elevated so that the radial fan is essentially flat. For a variety of practical reasons, however, somewhat different radial configurations are often used and it is of some interest to see what effect these variations have on the performance of the antenna.

Description of the Experiment

All the experimental runs were done with four 35 foot radials (except as noted), the length of the vertical set to 34 feet and a test frequency of 7.2 MHz. The antenna, including radials, was isolated from ground with a common mode choke (balun) in the feed line. Measurements of |S21| and Z_i were made for each test configuration.

The following configurations were tested:

1) Radials and antenna base elevated at 48 inches above ground.

2) The far end of the radials at 48 inches sloping down to the base at ground level.

3) A "gullwing" configuration as suggested by Dean Straw, N6BV, and later extensively modeled by Al Christman, K3LC.[4] The base was at ground level with the radials rising from the base at a 45° angle until they reached 48 inches above ground. The rest of the radials beyond this point were kept at 48 inches above ground from this point out to the far ends.

4) Radial lengths cut to 17.5 feet (≈ ⅛-wavelength). Radial and base height set to 48 inches. Antenna resonated with a 2.2 µH inductor.

5) For reference purposes, a run was made with the radials lying on the ground surface and the antenna base at ground level. This was done as a check because segment one of this experiment had been done earlier and ground conditions at the site had changed. Also a slightly different radial length was used (35 feet versus 33.5 feet).

Experimental results

The experimental results are summarized in Table 4. The values for |S21| were normalized by setting the value for configuration 1 to 0 dB and the rest to the difference between them and configuration 1. A line of data from an earlier experiment has been added for comparison. (See Note 2.)

As a check, for configuration 1, the current division between the radials was measured. Those results are summarized in Table 5.

Comments on Segment Two

The most important observation is that radically changing the radial geometry does not seem to have a major impact on performance (|S21|).

Table 5
Measured current division between radials, normalized to 1A total base current.

Radial number	*Normalized Current (A)*
1	0.249
2	0.269
3	0.260
4	0.221

Table 4
Experimental Results

Configuration Number	*|S21| Normalized (dB)*	*Z_i (Ω)*	*Test Configuration*
1	0	39 + *j* 6.3	Base and 4 radials elevated at 48 inches
2	–0.47	36 + *j* 6.2	Base at ground level, radials ends at 48 inches
3	–0.65	29 – *j* 11	Gullwing, base at ground level radial ends at 48 inches
4	–0.36	39 + *j* 0.9	Base and radials at 48 inches radial length = 17.5 feet 2.2 µH inductor to resonate
5	–5.19	132 + *j* 22	Base and radials on ground surface, four 35 foot radials
Earlier Experiment (See Part 2)	–1.79	51 + *j* 1	Base and radials on ground surface, Four 21 Foot Radials

Cutting the radial lengths in half (configuration 4) and adding a small loading inductor reduced the gain by only –0.4 dB. The use of shorter radials has been suggested by Weber (see Note 3) and Moxon to either make the radial screen footprint smaller and/or reduce asymmetry in the current division between radials.[5]

I was surprised to see that the gain reduction for the gullwing configuration (configuration 3) was slightly worse than simply running the radials straight up to the far end (configuration 2). It may have something to do with the higher feed point impedance in configuration 2. In the case of the gullwing, the radials rise close to the vertical element, resulting in some cancellation between the vertical element and radial currents depressing the feed-point resistance. We see a similar effect in top-loaded antennas with sloping wires. From the standpoint of keeping the radials above head height for safety reasons, the gullwing is more attractive than just sloping up the radials.

It would seem that anything done to get the radial wires away from ground makes a great improvement as you can see from configuration 5, where the radials are lying directly on the ground surface. Even using shorter, resonant radials on the ground surface is not as effective as simply elevating the radials. Modeling and experimental work shows that you don't have to get very high to make a substantial improvement but greater heights are used for safety reasons to keep the radials above head height.

One thing missing from this experiment was the use of more than four radials. An earlier experiment which compared four elevated radials to eight in configuration 1, showed very little difference in $|S21|$ (about + 0.2 dB). The advantage of more radials is not so much improved efficiency but rather reduced chances for radial current asymmetry and a lower Q, which can improve the SWR match bandwidth.

Summary

The experiments seem to show that a few elevated radials can work well as a replacement for a large number of ground radials. The experiments also show that alternate elevated radial geometries can work nearly as well as the "standard" and may have practical advantages.

Certainly this set of experiments does *not* completely resolve the debate regarding a large number of ground radials versus a few elevated radials, but it does lend some credence to the *NEC* modeling. To finally resolve these questions we need other experimenters to repeat these and/or similar experiments. We should also recognize that these experiments were done at a particular site, which has good to very-good soil. Repeating the tests over other soils, particularly poor ones, would be of considerable interest. It is at least possible that larger differences between the ground surface and elevated radials might be seen.

Even if these tests and *NEC* modeling are in fact correct and a few elevated radials can, in principle, provide equivalent performance to a large number of ground radials, this does not mean we should dash out and convert all our ground systems to four elevated radials. Because of their much higher Q, elevated radial systems are subject to a number of ills. They are very sensitive to details of layout, soil characteristics, nearby conductors, coupling to feed lines, and other factors. Like ground radials, elevated radial systems work much better if the screen is not too sparse: in other words, try to use 12 or more radials. You will be much happier.

Notes

[1] Rudy Severns, N6LF, "Experimental Determination of Ground System Performance for HF Verticals, Part 1," *QEX,* Jan/Feb 09, pp 21 - 25.

[2] Rudy Severns, N6LF, "Experimental Determination of Ground System Performance for HF Verticals, Part 2," *QEX* Jan/Feb 09, pp 48 - 52.

[3] Dick Weber, K5IU, "Optimum Elevated Radial Vertical Antennas," *Communication Quarterly*, Spring 1997, pp 9 - 27.

[4] R. Dean Straw, N6BV, "Antennas Here Are Some Verticals On The Beach," *ARRL Antenna Compendium, Vol 6*, pp 216 - 225.

[5] L. Moxon, G6XN, "Ground Planes, Radial Systems and Asymmetric Dipoles," *ARRL Antenna Compendium Vol 3*, pp 19- 27.

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A Closer Look at Vertical Antennas With Elevated Ground Systems

N6LF shares his results from more vertical antenna experiments.

[This article is being published in two parts. — Ed.]

Among amateurs, there has been a long running discussion regarding the effectiveness of a vertical antenna with an elevated ground system compared to one using a large number of radials either buried or lying on the ground surface. NEC modeling has indicated that an antenna with four elevated λ/4 radials would be as efficient as one with 60 or more λ/4 ground based radials. Over the years there have been a number of attempts to confirm or refute the NEC prediction experimentally, with mixed results. These conflicting results prompted me to conduct a series of experiments directly comparing verticals with the two types of ground systems. The results of my experiments were reported in a series of *QEX*[1-7] and *QST*[8] articles (Adobe *Acrobat* .pdf files of these articles are posted at **www.antennasbyn6lf.com**). From these experiments I concluded that at least under *ideal* conditions four elevated λ/4 radials could be equivalent to a large number of radials on the ground.

Confirmation of the NEC predictions was very satisfying but that work *must not be taken uncritically*! My articles on that work failed to emphasize how prone to asymmetric radial currents and degraded performance the 4-radial elevated system is. You cannot just throw up any four radials and get the expected results! I'm by no means the first to point out that the performance of a vertical with only a few radials is sensitive to even modest asymmetries in the radial fan.[9, 10, 11] It is also sensitive to the presence of nearby conductors or even variations in the soil under the fan.[12] These can cause significant changes in the resonant frequency, the feed point impedance, the radiation pattern and the radiation efficiency. While these problems have been pointed out before, as far as I can tell no detailed follow-up has been published. Besides the practical problem of construction asymmetries, at many locations it's simply not possible to build an ideal elevated system even if you wanted to. There may not be enough space or there may be obstacles preventing the placement of radials in some areas or other limitations. I think it's very possible that some of the conflicting results from earlier experiments may well have been due to pattern distortion and increased ground loss that the simple 4-wire elevated system is susceptible to.

As the sensitivity of the 4-radial system and its consequences sank into my consciousness I began to strongly recommend that people use at least 10 to 12 or more radials in elevated systems. Although I have heard anecdotal accounts of significant improvements in antenna performance when the radial numbers were increased to 12 or more, I have not seen any detailed justification for that. What follows is my justification for my current advice.

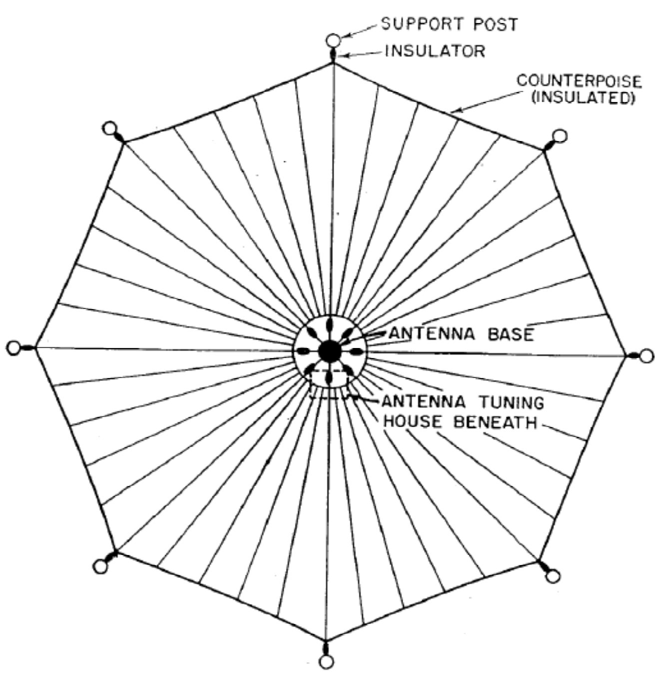

Figure 1 — A typical counterpoise ground system. Figure adapted from from Laport.[14]

[1]Notes appear on page 41

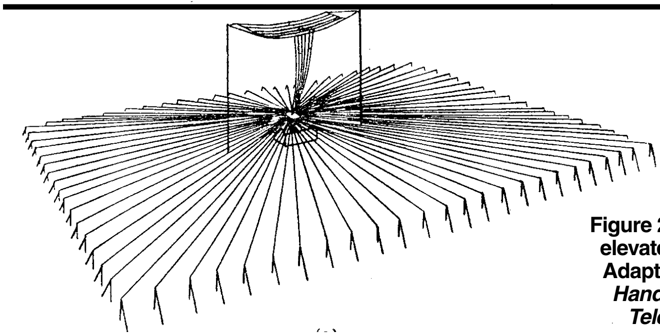

Figure 2 — A very large LF elevated ground system. Adapted from *Admiralty Handbook of Wireless Telegraphy*, 1932.[34]

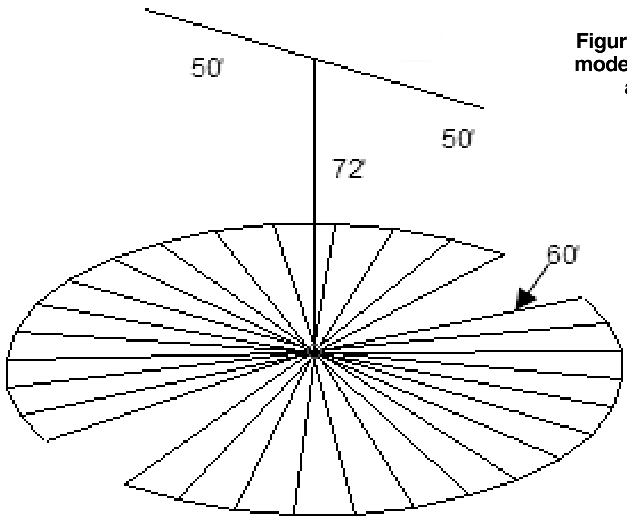

Figure 3 — *EZNEC* model of the 1BCG antenna.

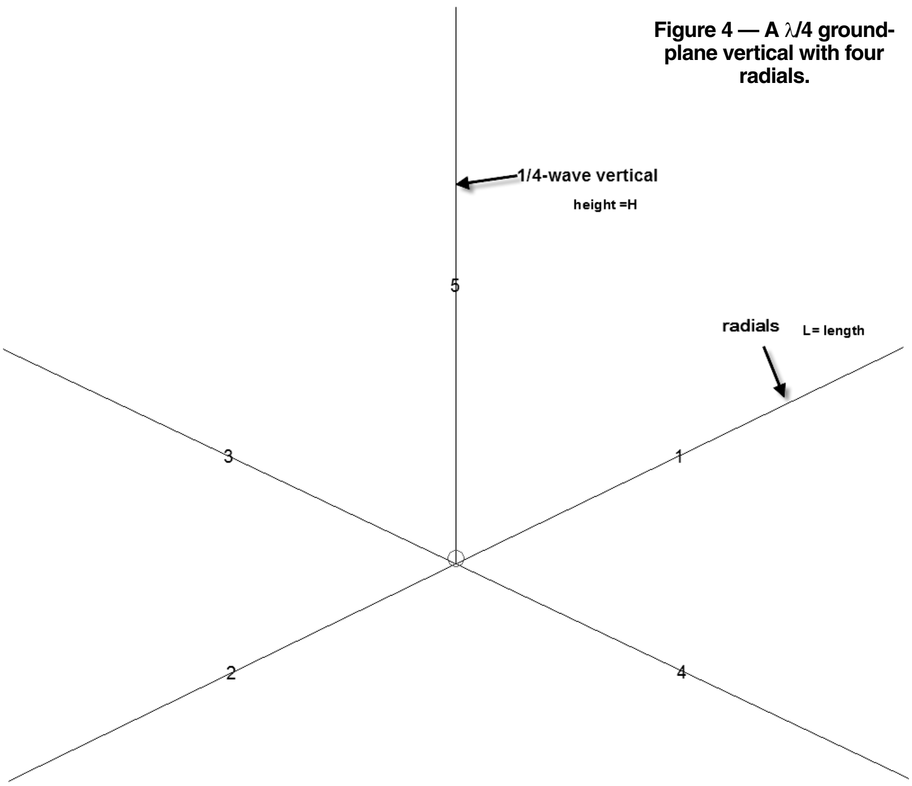

Figure 4 — A $\lambda/4$ ground-plane vertical with four radials.

My original intention for this article was to illustrate the problems introduced by radial fan asymmetries and to discuss some possible remedies. In the process, however, I came to realize that before going into the effects and cures for asymmetries it was necessary to first understand the behavior of ideal systems. Ideal systems can show us when and why they are sensitive and point the way towards possible cures or at least ways minimize problems. The discussion of ideal antennas (over real ground however!) also illustrates a number of subtleties in the design and possibly useful variations that differ somewhat from current conventions.

For these reasons, after some historical examples of elevated wire ground systems, I'll spend a lot of time analyzing ideal systems and then move on to the original purpose of this article: asymmetric radial currents and how to avoid them. At the end of this article I summarize my advice for verticals using elevated ground systems. While much of what follows is derived from NEC modeling, I have incorporated as much experimental data as I could find and compared it to the NEC predictions to see if NEC corresponds to reality.

Prior Work on Elevated Ground Systems

There is a lot of prior information on elevated ground systems: Moxon,[10, 11] Shanney,[13] Laport,[14] Doty, Frey and Mills,[12] Weber,[9] Burke and Miller,[15, 16] Christman,[18 to 33] Belrose[39, 42] and many others. There is also my own work, some published but most not.

Some History

In the early days of radio, operating wavelengths were in the hundreds or thousands of meters. Ground systems with $\lambda_0/4$ radials were rarely practical but very early it was recognized that an elevated system called a "counterpoise" or "capacitive ground," with dimensions significantly smaller than $\lambda_0/4$, could be quite efficient. Note, λ_0 is the free space wavelength at the frequency of interest. Figure 1 shows a typical example of a counterpoise.

Here is an interesting quotation from *Radio Antenna Engineering* by Edmund Laport[14] regarding counterpoises:

"*From the earliest days of radio the merits of the counterpoise as a low-loss ground system have been recognized because of the way in that the current densities in the ground are more or less uniformly distributed over the area of the counterpoise. It is inconvenient structurally to use very extensive counterpoise systems, and this is the principle reason that has limited their application. The size of the counterpoise depends upon the frequency. It should have sufficient capacitance to have* *a relatively low reactance at the working frequency so as to minimize the counterpoise potentials with respect to ground. The potential existing on the counterpoise may be a physical hazard that may also be objectionable.*"

Laport was referring to counterpoises that were smaller than $\lambda_0/4$ in radius. In situations where $\lambda_0/4$ elevated radials are not possible amateurs may be able to use counterpoises instead. Unfortunately, beyond the brief remarks made here, I have to defer further discussion of counterpoises to a subsequent article.

Rectangular counterpoises, some with a coarse rectangular mesh, were also common. A rather grand radial-wire counterpoise is illustrated in Figure 2.

Amateurs also used counterpoises. Figure 3 is a sketch of the antenna used for the initial transatlantic tests by amateurs (1BCG) in 1921-22.[35, 36] The operating frequency for the tests was about 1.3 MHz (230 m). At 1.3 MHz, $\lambda_0/4$ = 189 feet, so the 60 foot radius of the counterpoise corresponds to $\approx 0.08\ \lambda_0$.

Note that in all these examples, a large number of radials are used. The use of only a few radials, initially with VHF antennas elevated well above ground, seems to have started with the work of Ponte[37] and Brown.[38]

Behavior With Ideal Radial Fans

In this section we'll look at verticals with a length (H) $\approx \lambda_0/4$ (λ_0 is the free space wavelength) and symmetric elevated radial systems where the height above ground (J) and the number (N) and length (L) of the radials is varied. We'll also look at the effect of soils with different characteristics from poor to very good. Even though we will be looking at verticals with H $\approx \lambda_0/4$, keep in mind that elevated ground systems can also be used with verticals of other lengths, with or without loading, inverted Ls, and other antenna types. Elevated radials can also be used with multi-band antennas.

NEC Modeling

Figure 4 shows a typical model of a vertical with a radial system. Except as noted, the following discussion will focus on operation on 3.5 to 3.8 or 7.0 to 7.3 MHz as the operating band and 3.65 or 7.2 MHz as a spot frequency near mid-band. The conductors (both the vertical and the radials) are lossless no. 12 wire. Most of the modeling was done over real grounds. The modeling used *EZNEC* Pro4 v.5.0.45, using the *NEC4D* engine. The use of *NEC4D* over real soils gives the correct interaction between ground and the antenna. Excellent free programs based on *NEC2* are available, but these do not properly model the ground-antenna interaction, so that results obtained from them must be used with some caution.[41] For HF verticals close to ground this is an important limitation.

The Effect of Element Dimensions on Performance

The simplest idea of a ground-plane antenna is that you take a quarter-wave vertical and add four quarter-wave radials at the base. It is well known that the elements of a dipole will be a few percent shorter than λ_0 so it is usually assumed that in a ground-plane antenna the vertical and the radial lengths will also be a few percent less than λ_0. Typically

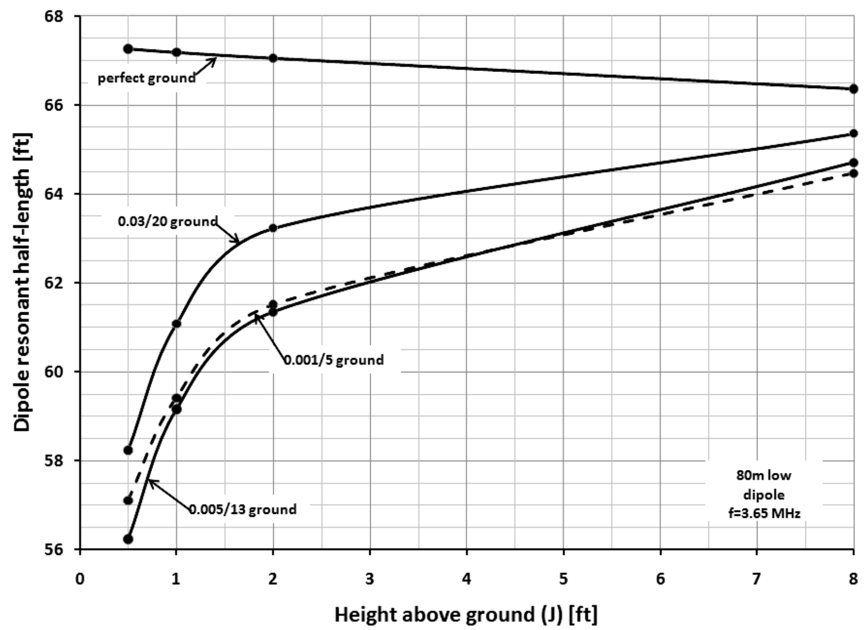

Figure 5 — Dipole half-length for resonance for different values of J and different soils.

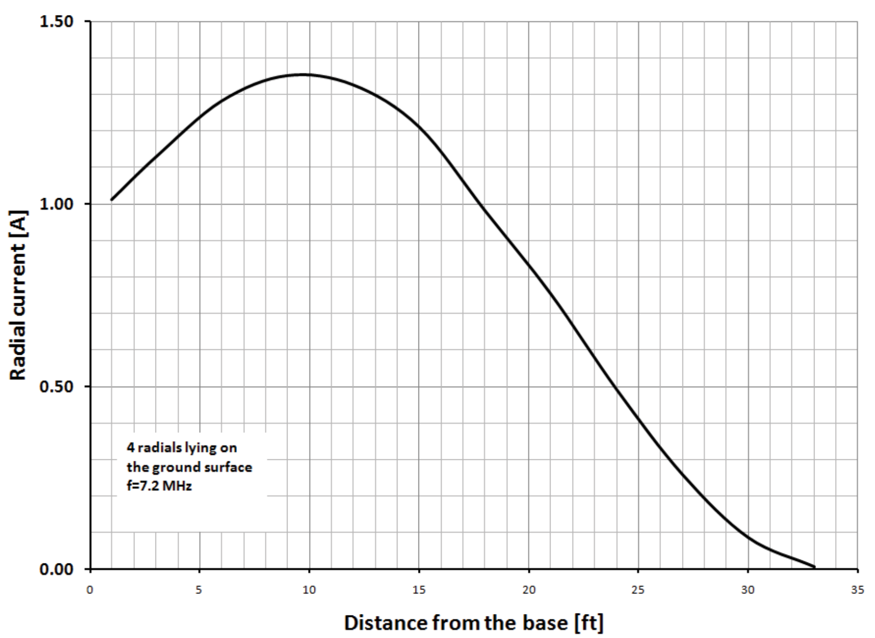

Figure 6 — Measured current on a 33 foot radial at 7.2 MHz. This antenna uses four radials lying on the ground surface.

it is assumed that the vertical and the radials will be *individually* resonant at the operating frequency. Unfortunately it's not that simple, because the vertical is coupled to the radials and both interact strongly with ground because, at least at lower HF (<20 m), the base of the vertical and radial fan will usually be only a fraction of λ_0 above ground. What you have in reality is a coupled multi-tuned system with complicated interactions. It turns out that there are a wide range of pairs of values for H and L that result in resonance, or $X_{in} = 0$ at the feed point (where $Z_{in} = R_{in} + j\,X_{in}$ and Z_{in} is the feed point impedance). Some of these combinations where neither the vertical nor the radials are individually resonant may be useful.

Antenna Resonance and Element Dimensions

The free space wavelength (λ_0) at a given frequency in MHz (f_{MHz}) is given as:

$$\lambda_0 = \frac{299.792}{f_{MHz}}[m] = \frac{983.570}{f_{MHz}}[feet] \quad \text{[Eq 1]}$$

At 3.65 MHz, $\lambda_0/4$ = 67.368 feet. If we model a resonant $\lambda/4$ vertical over perfect ground using no. 12 wire, we find that at 3.65 MHz, $\lambda/4$ = H = 65.663 feet, which is about 3.5% shorter than $\lambda_0/4$.

To take into account the effect of ground on radial resonance for a given value of J and soil characteristic, it has been suggested that we can erect a low dipole at the desired radial height (J) and trim its length to resonance. An example of this is given in Figure 5.

For J = 8 feet, depending on the soil, L varies from 64.5 feet to 66.4 feet. As we reduce J we find that L gets smaller. The shift in resonance for radials close to ground has also been demonstrated experimentally. (See Note 2.) Figure 6 shows the measured radial current at 7.2 MHz on 33 foot radials (sum of four radials). Clearly this radial is $\lambda/4$ resonant at a lower frequency than 7.2 MHz! As Figures 5 and 6 show, the effect gets much larger for small values of J.

What do we mean by "resonant" values for H and L "independently"? It's not just that the reactances cancel at the feed point. When I say "the resonant length for H or L" I'm talking about the case where the current distribution on the vertical and the radials independently corresponds to resonance: in other words, the current just reaches a maximum at either the base of the vertical or at the inner ends of the radials. If either H or L is made longer than resonance, the current maximum will move out onto the radials or up the vertical. Figure 7 shows the current distribution on a vertical and the radials for three combinations of H and L, each of which yield $X_{in} = 0$ at the feed point.

To better understand what's happening we can expand Figure 7 around the 1 A feed point (indicated by the arrow) as shown in Figure 8.

For H = 64 feet and L = 80.85 feet, the current on the vertical has not peaked so the vertical is too short for resonance. The radial current peak is well out on the radials, however, so clearly the radials are too long for resonance. The reactance of the vertical and the radials cancels at the feed point so the antenna is "resonant" but not the vertical and radials individually. Similarly, for H = 69 feet and L =58.8 feet, the current in the vertical peaks and begins to fall (moving from the top to the bottom of the vertical) before the feed point is reached. Again, we have a resonant antenna but the vertical and the radials are not

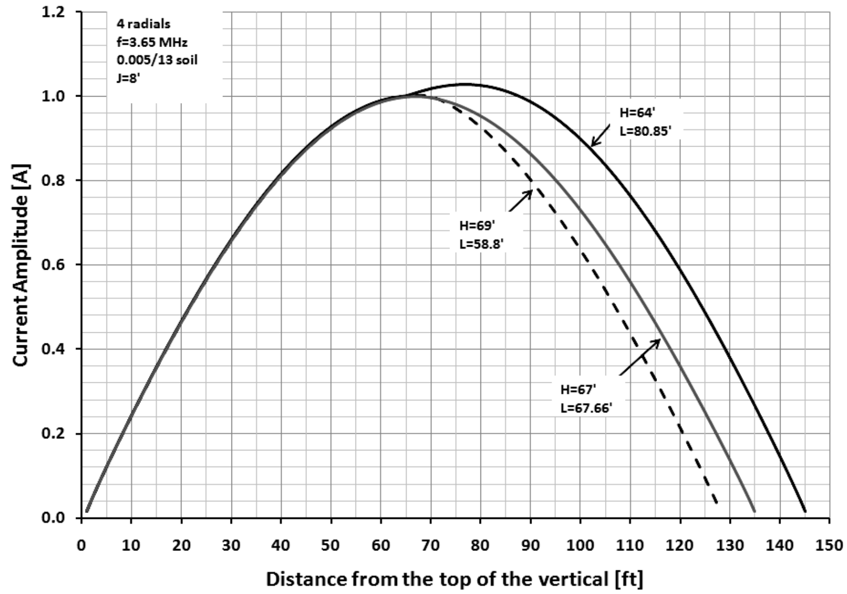

Figure 7 — Current distribution on the vertical and the radials. The current starts at the top of the vertical, runs to the base and then out along the radials. The radial current is the sum of the currents in the four radials. The currents are for 1 A_{rms} at the feed point.

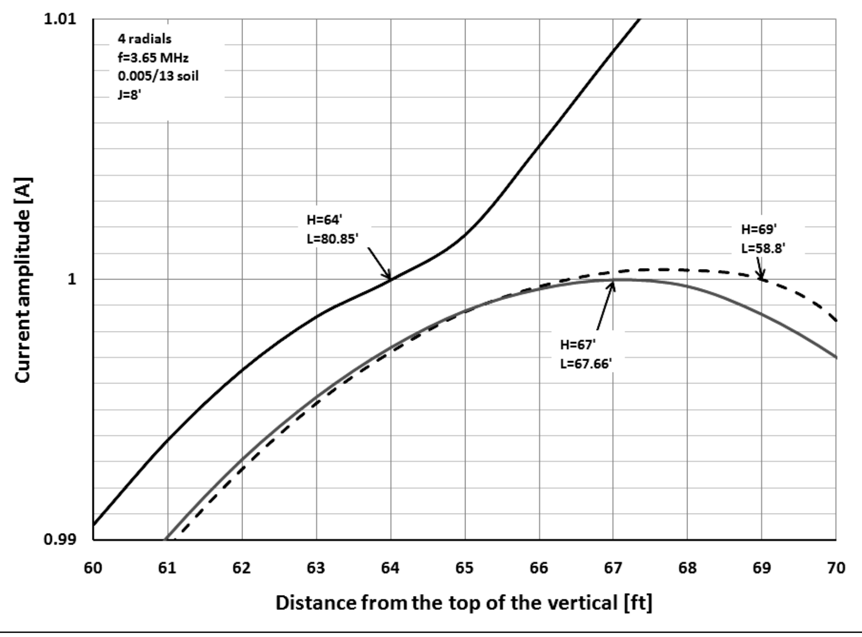

Figure 8 — Current distribution on the vertical and the radials expanded around the feed point. The arrows point to the junctions between the vertical and the radials.

individually resonant. If we set H = 67 feet and L = 67.66 feet, however, both the vertical and the radials are λ/4 resonant individually.

The "resonant length" (by the definition given above!) of the vertical is 67 feet and the "resonant" length for the radials is 67.7 feet, both of these lengths are substantially different than the value we got earlier for λ/4 resonance for a vertical over an infinite perfect ground-plane (65.7 feet). The "resonant" radial length of 67.7 feet is quite different from the dipole 8 feet above average ground (64.7 feet). H and L are actually closest to λ_0 (67.4 feet). What we have just seen is only one particular example. If we change J and/or the soil characteristics and/or the number of radials, these lengths will change!

Setting up the antenna so that both the vertical and the radials are individually resonant turns out to not be so simple and we might ask, "Is it really necessary to have both the vertical and the radials resonant individually?" It turns out that there are other considerations besides the current distribution with regard to the choice of L for a given H. It is possible to use values of L where $X_{in} \neq 0$ and compensate for that with a tuning impedance at the feed point for example, or perhaps use some top-loading. In addition, in some situations it may not be possible to have radials long enough to make $X_{in} = 0$ while keeping the radial fan symmetric. Further, Weber has suggested that radials with L <λ/4 or >λ/4 are a possible cure for radial current division inequality. (See Note 9.) So we have reasons to investigate the effect of variations in vertical height and radial length on antenna behavior.

For each value of H, number of radials (N), height above ground (J), ground characteristic (σ = conductivity and ε_r = permittivity) and choice of operating frequency, there will be some radial length (L_r) that makes the antenna resonant. That's a lot of variables! So we will look at only a few examples to get a general idea of what happens.

Figure 9 gives an example of the variation in the value for L (L_r) that results in resonance at the feed point ($X_{in} = 0$) as a function of N and several values of H, with fixed values of f, J and soil.

Notice how widely L_r varies with N for most values of H although there is one value for H (66.71 feet) that seems to have only a small variation in L_r as N is changed. Note also how much shorter L_r becomes when H is increased by a few feet. This could be very useful in situations where space for the radial fan is limited. On the other hand note how quickly L_r grows when H is shortened. For N = 16 we see that when H = 64 feet, L_r = 106 feet but for H = 69 feet, L_r is only 39 feet! That's a difference in L_r of almost 3:1. If you cannot make H long enough, all is not lost! A bit of top loading has an effect much like increasing H.

Another way to explore the interaction between L and N is to set L equal to L_r for some value of N (say 16 radials) and while watching the resonant frequency (f_r), vary the number of radials as shown in Figure 10. Note that the most stable f_r is where H = L = 66.71 feet. That is relatively close to the values we got earlier for independently resonant vertical and radials. (Be careful, this is particular to this example; things will vary with different J, ground type, and other variables). Note also that for H a bit tall, f_r decreases as radials are added, but if H is a bit short f_r increases as radials are added. This kind of behavior can be confusing if you are trimming the radials to resonate at a particular frequency, especially if you add some radials. It is possible you could add some radials and then have to make all the original radials longer!

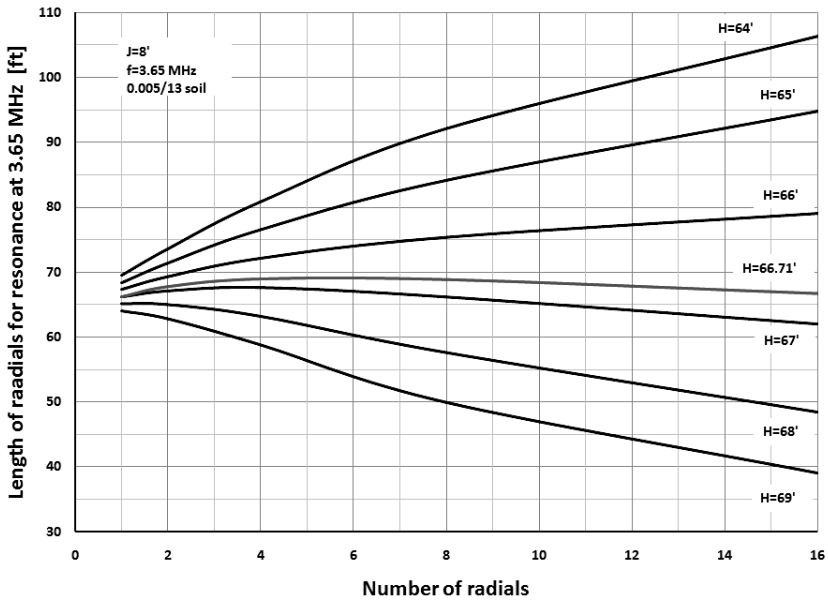

Figure 9 — Examples of the effect of radial number on the radial length for resonance at 3.650 MHz (L_r) for several different values of H.

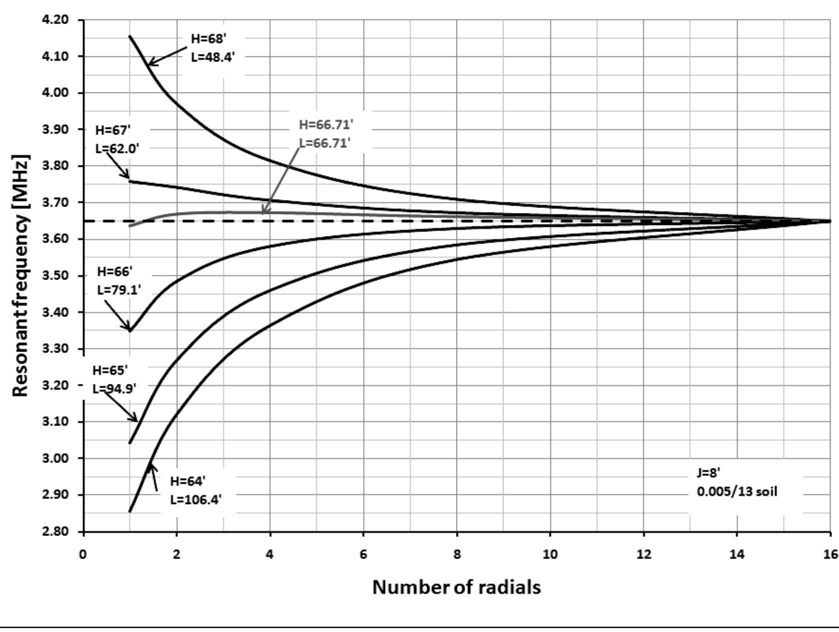

Figure 10 — Resonant frequency of the antenna as a function of radial number for several combinations of H and L that are resonant at 3.650 MHz with N = 16.

This raises the question, "Do real antennas actually behave this way?" During the ground system experiments, I saw exactly this kind of behavior. For the 160 m vertical, f_r went down as I added radials but for the 40 m verticals, f_r went up with radial number. Figure 11 shows graphs of experimental measurements, one for 160 m and the other for 40 m. Real antennas can behave as the modeling predicts.

At this point it's pretty clear that there is considerable interaction between the variables (H, L, J, and so on) but it's not obvious yet if there are optimum combinations (some better than others).

The effect of radial length on efficiency

It turns out that the values for both N and L can have a significant effect on the efficiency of the antenna. Burke and Miller published a very interesting paper in 1989 with the results of NEC modeling of both elevated and buried radial systems for a wide range of N, L, J and soil characteristics.[15] I read this paper many years ago but I have to admit that it did not dawn on me just how much important information was there. Recently the light dawned as I re-read the paper and some additional graphs that Jerry Burke kindly sent me, so I have been redoing some of their modeling. Some of the Burke-Miller graphs were plots of average gain (G_a) versus radial length with radial number as a parameter. G_a is a useful proxy for radiation efficiency in that it gives the proportion of the input power to the antenna that is actually radiated into space. G_a is the ratio of the radiated power (P_r) to the input power (P_{in}) in dB ($G_a = 10 \text{ Log } [P_r/P_{in}]$). All of the power dissipated in the earth, including the near-field losses and reflections in the far-field, are subtracted from the input power. What is actually done is to integrate the power flow across a hemisphere with a very large radius centered on the antenna. The total power flowing through the surface of the hemisphere is P_r. I should emphasize that this is the power radiated towards the ionosphere, power in the ground-wave is considered a loss. For Amateurs, where sky-wave propagation is the norm at HF, this makes sense.

The Burke-Miller graphs used a constant value for H. I will begin with similar graphs but for Amateurs it is more likely that as L is increased H will be decreased to maintain resonance at a given frequency, so I will also show that variation.

Figure 12 is an example of the effect of radial length and radial number on G_a of the antenna when H is kept constant (68 feet in this example).

Figure 12 has some interesting features:

1) Beginning with short values for L, G_a increases slowly up to a maximum. Below maximum, using radials somewhat shorter

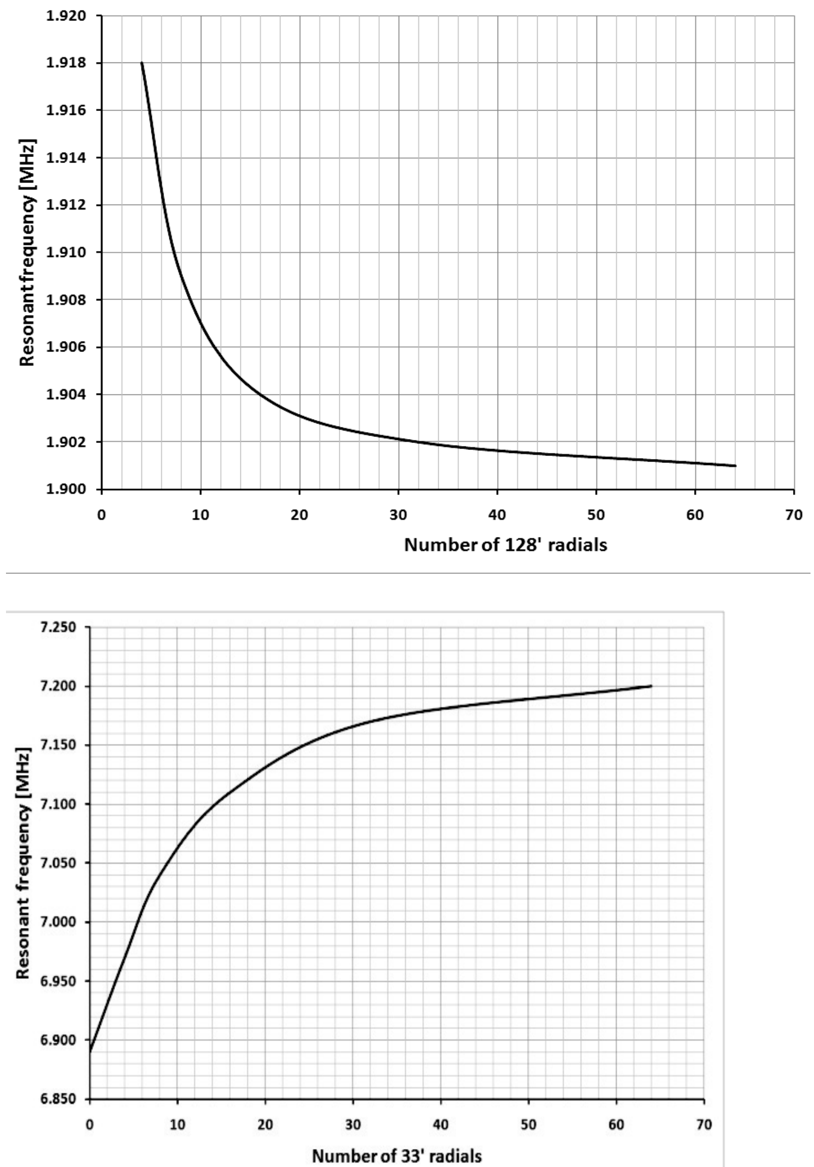

Figure 11 — Experimental measurements of the effect of radial number on resonant frequency.

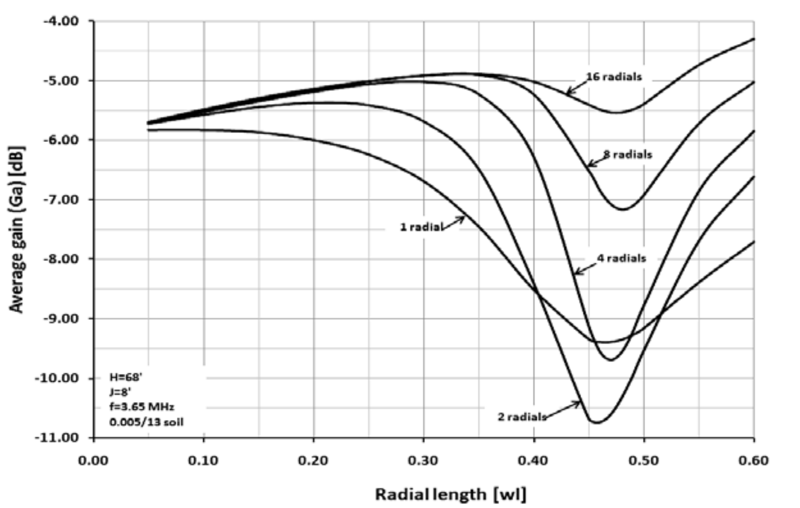

Figure 12 — Average gain as a function of radial length (in wavelengths, λ_0) and number of radials. H = 68 feet, J = 8 feet, f = 3.650 MHz and 0.005/13 soil.

than λ/4 does not seriously reduce the efficiency.

2) Above the maximum, however, there is a large dip! The bottom of the dip can be as much as –7 dB before G_a rises again for longer lengths.

3) Up to the length where G_a starts to fall, increasing N doesn't make much difference in G_a as long as you have four or more radials, but increasing N does push the dip towards longer radial lengths and reduces the depth of the dip.

Figure 12 is for the case where J = 8 feet. If we reduce J, the G_a graphs will change, as illustrated in Figure 13.

As the antenna is moved closer to ground, the efficiency starts to fall, the maximum is lower and the dip gets deeper and occurs at shorter values of L. In fact, if you push J down to 1 inch or less (the case for radials lying on the ground surface) the notch gets even deeper and begins to fall off at lengths well below $\lambda_0/4$. Note, however, that the effect is substantially reduced when larger numbers of radials are used.

One of the suggestions for improving current division between radials was to make them substantially longer than $\lambda_0/4$, in other words, L = 3 $\lambda_0/8$. (See Note 9.) As Figures 12 and 13 show, that's probably not a good idea unless you're using 16 or more radials, but with that many radials current division will already be much improved, as we'll see shortly. Before getting carried away with conclusions we have to ask, "Do real antennas actually behave this way and do we have any experimental verification?" As part of the ground system experiments reported in *QEX* and *QST* (see Notes 1 to 8), I measured the signal strength as N and L were varied with H constant. Figure 14 is a typical result.

I have to admit that during the experiments I did not make the connection between my measurements and the work of Burke and Miller (see Note 15) so I only extended the radial lengths out to slightly less than $\lambda_0/4$. But we can still see the predicted behavior:

1) For short L, the gain rises slowly to a point where it starts to fall.

2) When L is large the dip in gain is large.

3) Increasing N reduces the dip and moves it to larger values for L.

Besides the data shown in Figure 14, I ran spot checks on the gain with sixteen and thirty two 33 foot radials. These were also in agreement with the NEC predictions. I think it's pretty clear that NEC is telling us the truth and we need to pay attention! Radial length is an important consideration.

Figures 12 and 13 are for σ = 0.005 S/m and ε_r = 13, Figure 15 shows the effect of different soil characteristics on G_a for given H, J and N.

As we saw in Figure 6, close proximity

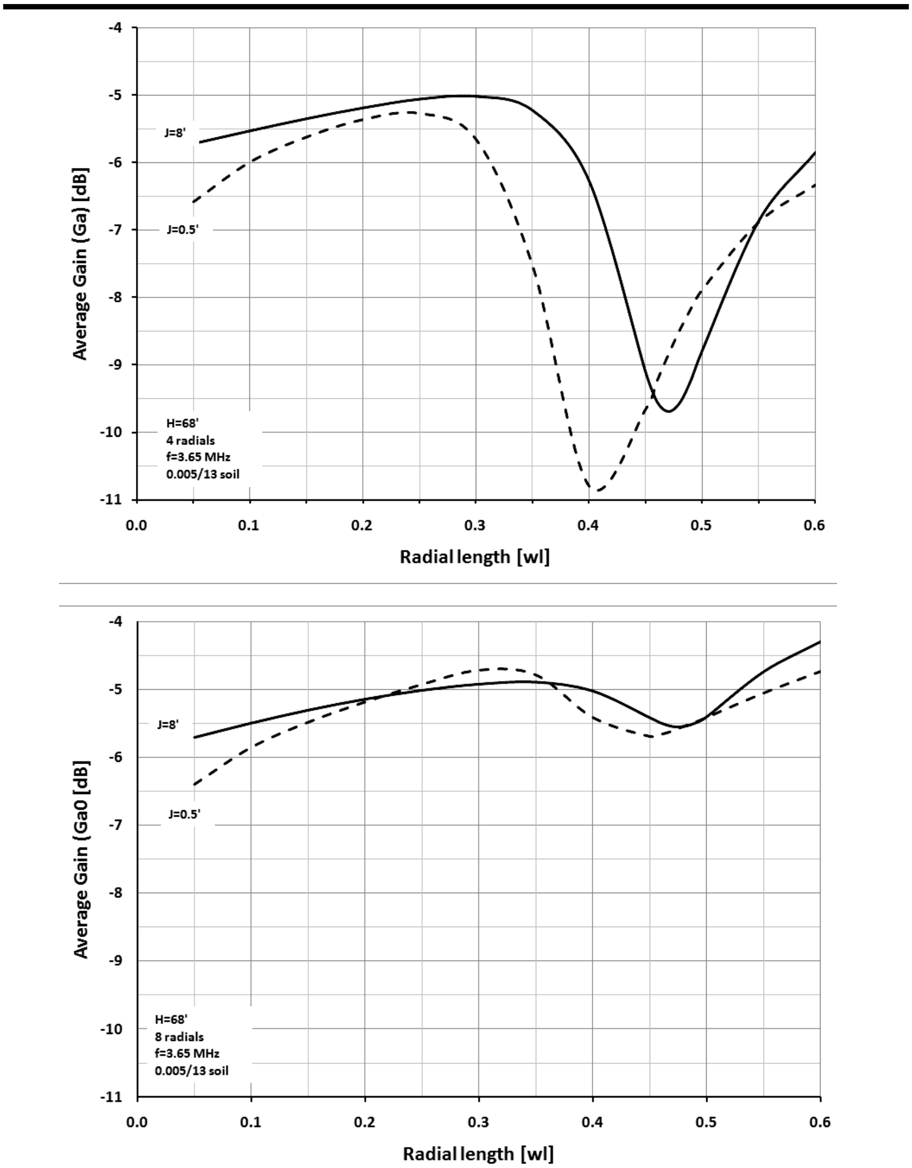

Figure 13 — Comparison of G_a for J = 8 feet and 0.5 feet. N = 4 and 8, and L is in λ_0 = wl.

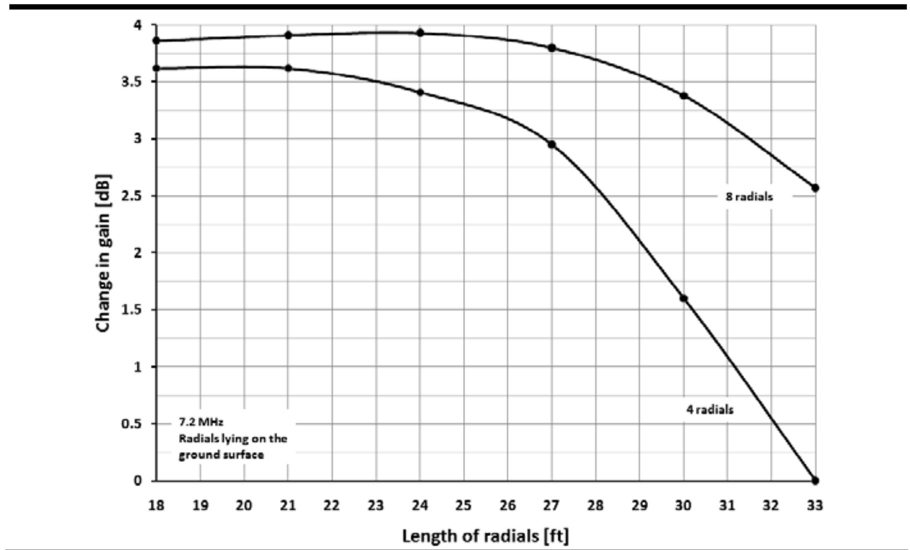

Figure 14 — Far-field change in signal strength as L and N are varied. Radials are lying on the ground surface. f = 7.2 MHz.

to ground has great effect on the radial resonant frequency. John Belrose, VE2CV, has modeled G_a for radials lying close to ground and the effect of different numbers of radials as shown in Figure 16.[42] Note that the data points in the graph were taken from Belrose's article and re-graphed.

The dashed line in Figure 16 represents the case where the lengths of the four radials are adjusted so that the radials are resonant. The predictions in Figure 16 agree with the experimental work shown in Figure 14 showing the effect of shortening the length of radials close to ground. Figure 16 also predicts that even a very small increase in height above ground for the radials will make a large difference in loss, especially if N is small. This large change in G_a with small elevations has been verified experimentally (see Note 3) as shown in Figure 17.

In some cases it may be necessary to use a vertical with H other than $\lambda/4$. Figure 18 shows G_a as a function of L for H = 100 feet ($\approx 3\ \lambda_0/8$), H = 68 feet ($\approx \lambda_0/4$) and H = 34 feet ($\approx \lambda_0/8$) with and without top-loading. Compared to H = 68 feet, the notch for H = 34 feet begins a lower value of L and is much deeper. Putting a short base loaded vertical over an elevated ground-plane may not be a good idea. (Note: this is something that needs to be explored further!) If we add two horizontal top-loading wires that restore the resonance of the 34 foot wire to that of the 68 foot wire, G_a is greatly improved. With the top-loaded vertical, the peak value for G_a is a few tenths of a dB lower than for the full height vertical but that may be acceptable because the vertical is only half as tall. That's something to think about for 160 m verticals. It is also interesting to note that the taller vertical ($H \approx 3\lambda/8$) while more tolerant of longer radials is somewhat less efficient ($\approx$ –0.5 dB). The lesson to draw here is that using elevated ground systems with short verticals can be problematic but really tall verticals may not be all that great either. You have to model the specific situation carefully to make sure you understand what's going on.

The graphs in Figure 12 assume that H is constant. We could also have varied H so that $X_{in} = 0$ for every value of L. This may give us some insight into optimum combinations (with regard to G_a!) of H and L. Figure 19 shows what happens when we do this compared to the case where H was constant for N = 4 and 16. The curves for a fixed H (solid lines) and variable H (dashed lines) are very similar, except that for the four radial case, the dip sets in a bit earlier and is somewhat deeper. The maximum G_a point is about 0.28 λ_0 with four radials and about 0.35 λ_0 with sixteen radials, but in both cases the maximum is very broad. As long as you stay

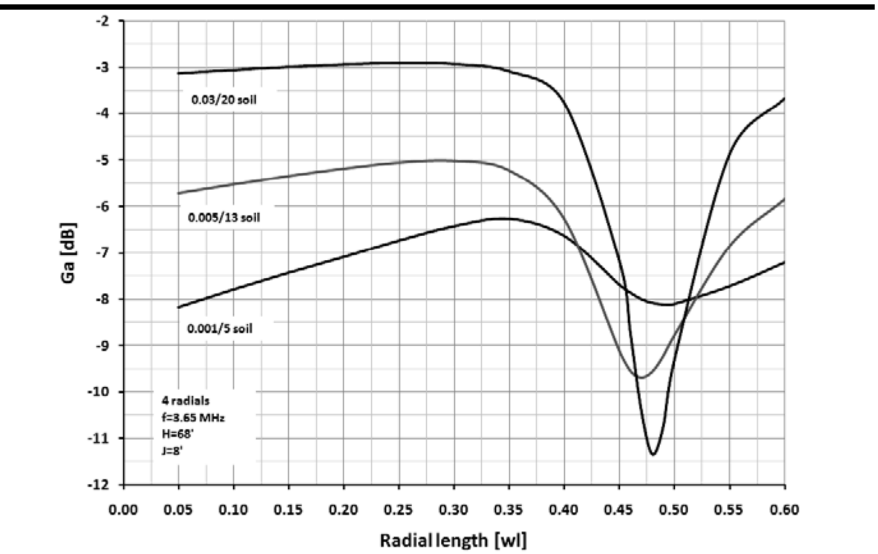

Figure 15 — Effect on G_a of different soils for H = 68 feet, J = 8 feet and N = 4.

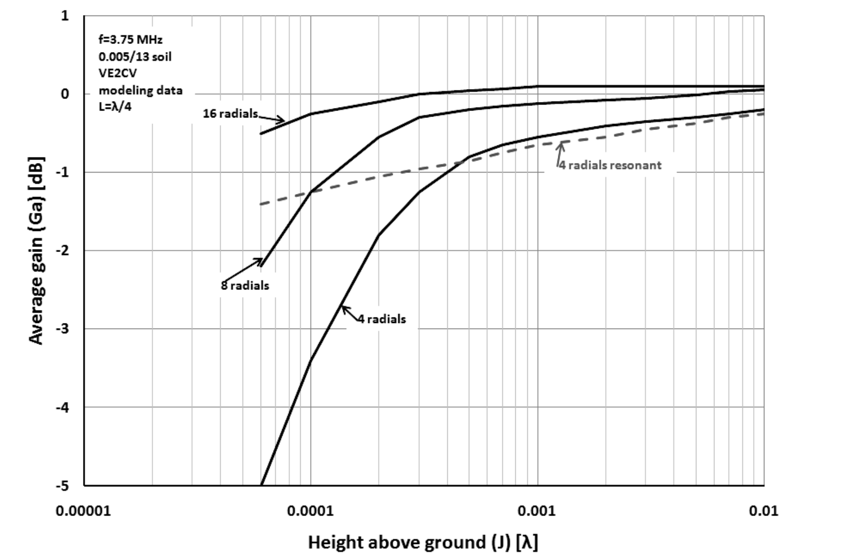

Figure 16 — Average gain when radials are placed close to ground.

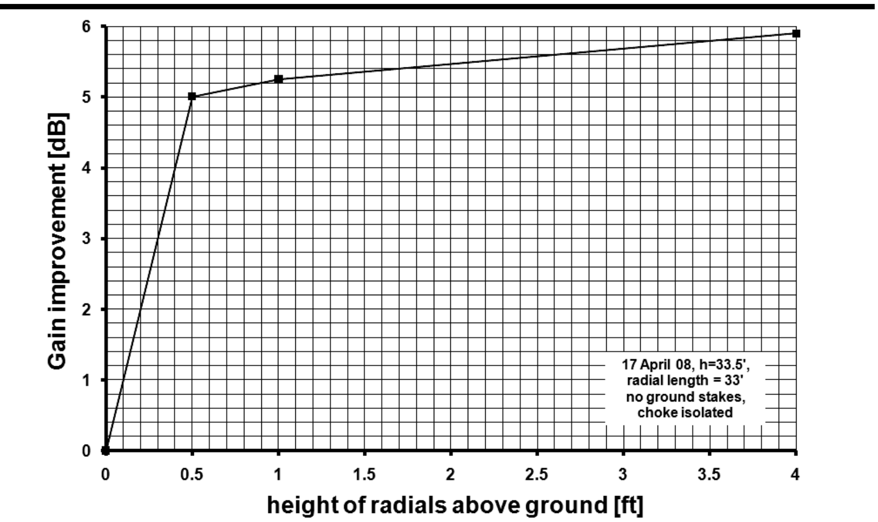

Figure 17 — Measured change in gain as four radials are elevated above ground.

below the point where G_a starts to fall, the value of L is not critical.

Figure 20 shows the values for H that result in resonance at 3.650 MHz for each radial length in Figure 19.

Again we see that the sensitivity to radial length is smaller when more radials are used. We can also look at the effect on R_{in} at resonance as we vary the H + L combination. An example is given in Figure 21.

When four radials are used there is also an important effect on the radiation pattern when the radials are too long.

Figure 22 compares the radiation patterns for two different combinations: L = 0.29 λ_0 and L = 0.46 λ_0. The first is close to the peak G_a value and the second is at the minimum of G_a. In the case of the long radials, not only is G_a much smaller but the peak of the radiation pattern has moved from about 22° to 45°! Clearly if you are using only a few radials, long radials are bad idea.

An Explanation for the Dips in G_a

Why do we see these large dips in G_a for some values of L? We can investigate this by looking at the current distributions on the radials and the associated E and H-field intensities close to ground under the radials. Figure 23 shows examples of the current distribution on the radials as a function of distance from the base (feed point) for several different radial lengths; 64, 70, 80, 100 and 121 feet. The graphs are for N = 4 except for the dashed line, where N = 16 and L = 121 feet.

For the same current at the feed point, with longer radials the currents are much higher as we go out from the base. We would expect these higher currents to increase both E and H-field intensities at ground level under the radials. Using the near-field plotting capability of NEC we can visualize the field intensities as shown in Figure 24.

Figure 24 shows the drastic increase in field intensities with longer radials. In this case I've chosen the longer radial length (121 feet) to correspond to the dip in G_a in Figure 12. Since the power dissipation in the soil will vary with the square of the field intensity, it's pretty clear why the efficiency takes such a large dip when the radials are too long. Figure 25 illustrates what happens to the fields under the radial fan when more radials are employed.

The earlier quotation from Laport stated that the use of more radials would make the fields under the radial fan more uniform. Figure 25 certainly supports that but we can go one step further to show how much the fields are smoothed with more numerous radials. Figure 26 makes that point.

Figure 26 is the E-field intensity just above ground level at points lying on a 90° arc with a radius of 40 feet (centered on the base) for two radial lengths (L = 64 feet and 121 feet) and N = 4 and 16. We can see that with only 4 radials, the E-field peaks sharply directly under the radials but with 16 radials the field is much more uniform.

In Part 2

In the second part of this series, we will examine radial systems for multiband verticals. We also take a look at the effect of various asymmetries in the radial fan.

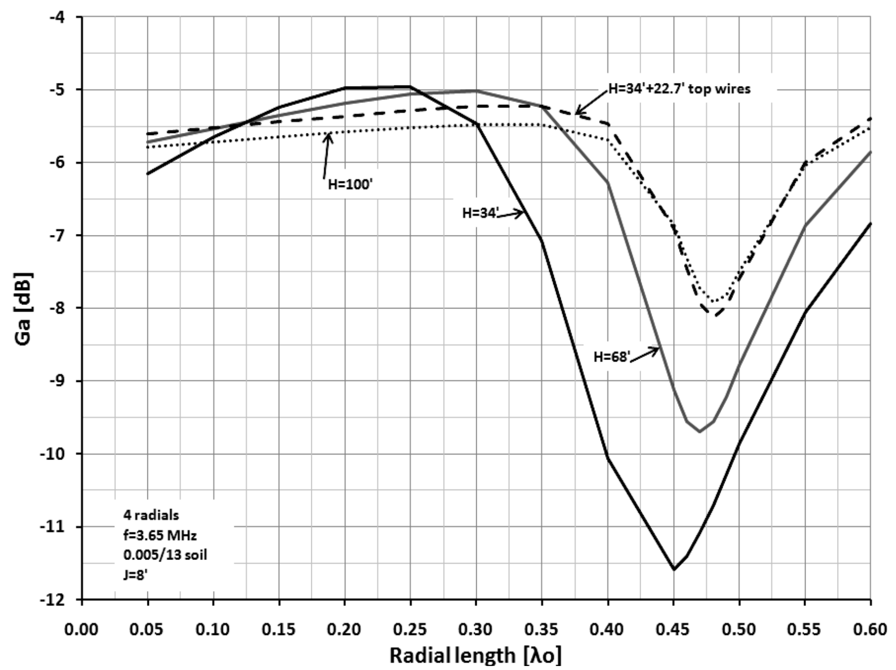

Figure 18 — Effect on G_a of short verticals. H = 100 feet, 68 feet, 34 feet and 34 feet with top-loading.

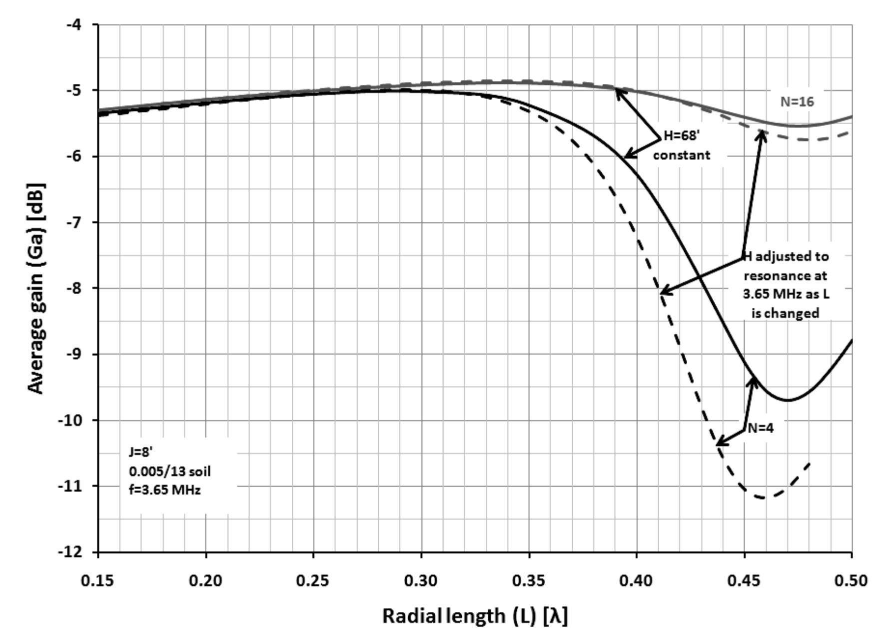

Figure 19 — Effect on G_a of radial length when H is varied to keep X_{in} = 0 at 3.650 MHz compared to the case where H is constant at 68 feet (from Figure 12).

Rudy Severns, N6LF, is a retired electrical engineer (UCLA '66). He holds an Amateur Extra class license and was first licensed in 1954 as WN7WAG. He is a life fellow of the IEEE and a live member of the ARRL. His current Amateur Radio interests are antennas, particularly HF vertical arrays and interactions between towers and arrays. He also enjoys 600 m operation as part of the group under the WD2XSH experimental license. Some of his publications about antennas are posted on his website at ***www.antennasbyn6lf.com****.*

Notes

[1]Severns, Rudy, N6LF, "Experimental Determination of Ground System Performance for HF Verticals, Part 1, Test Setup and Instrumentation," *QEX*, January/February 2009, pp 21-25.

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[4]Severns, Rudy, N6LF, "Experimental Determination of Ground System Performance for HF Verticals, Part 4, How Many Radials Does My Vertical Really Need?" *QEX*, May/June 2009, pp 38-42.

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[9]Dick Weber, K5IU, "Optimum Elevated Radial Vertical Antennas," *Communication Quarterly*, Spring 1997, pp 9-27.

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[14]Laport, Edmund A., *Radio Antenna Engineering*, McGraw-Hill, 1952. Note:

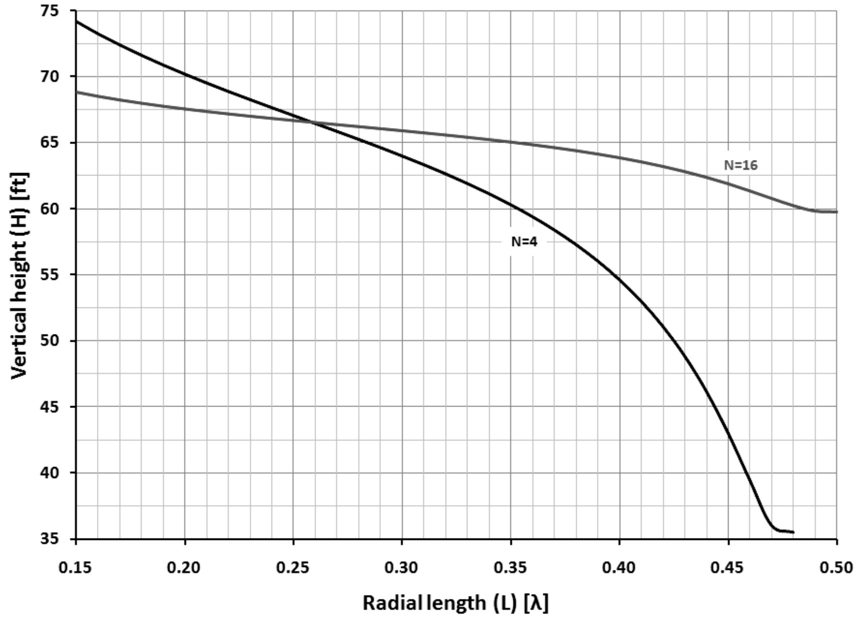

Figure 20 — Values for H that make $X_{in} = 0$ as L is varied.

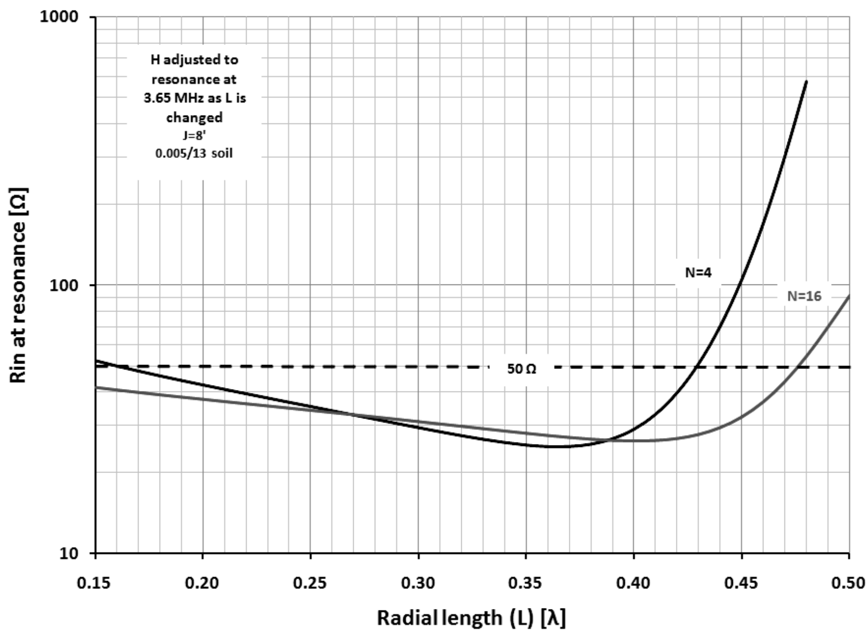

Figure 21 — R_{in} at resonance as a function of L.

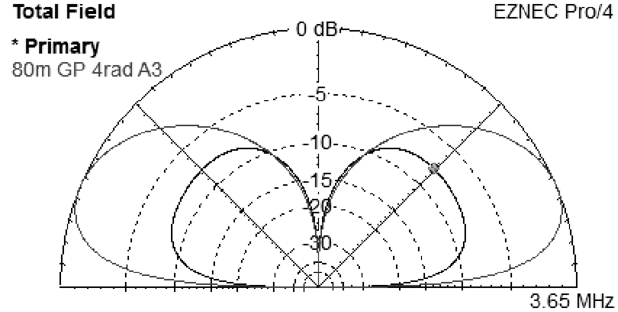

Figure 22 — Radiation pattern for H = 64.64 feet – L = 78.15 feet and H = 39.49 feet – L = 123.96 feet. N = 4 in both cases.

this book is available on-line in .pdf format at **http://snulbug.mtview.ca.us/books/RadioAntennaEngineering/**.

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[20]Christman, Al, K3LC, "Elevated Vertical-Antenna Systems", *QST*, August 1988, pp 35-42. Additional comments related to this article, *QST* Technical Correspondence, May 1989, pg.50.

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[25]Christman and Radcliff, "AM Broadcast Antennas with Elevated Radials: Varying the Number of Radials and Their Height Above Ground," *IEEE Transactions on Broadcasting*, 1996.

[26]Christman, Al, K3LC, "Elevated Vertical Antennas for the Low Bands: Varying the Height and Number of Radials," *ARRL Antenna Compendium*, Vol 5, 1996, pp 11-18.

[27]Christman, Al, K3LC, "Dual-Mode Elevated Verticals," *ARRL Antenna Compendium*, Vol 6, 1999, pp 30-33.

[28]Christman, Al, K3LC, "Elevated Vertical Antennas Over Sloping Ground," ARRL Antenna Compendium, Vol 6, 1999, pp 189-201.

[29]Christman, Al, K3LC, "'Gull-Wing' Vertical Antennas," *National Contest Journal (NCJ)*, Nov 2000, pg 14.

[30]Christman, Al, K3LC, "Using Elevated Radials With Grounded Towers," *IEEE Transactions on Broadcasting*, Vol 47, No. 3, September 2001, pp 292-296.

[31]Christman, Al, K3LC, "A Study of Elevated Radial Ground Systems for Vertical Antennas," Part 1, *National Contest Journal (NCJ)*, January/February 2005, pp 19-22. Part 2, *NCJ*, March/April 2005, pp 17-20.

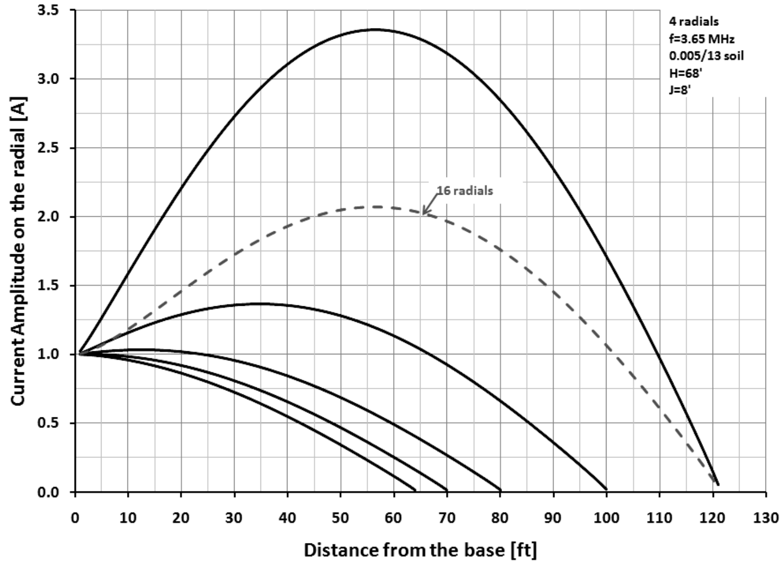

Figure 23 — Radial current distribution as a function of distance from the base. N = 4, H = 68 feet, f = 3.65 MHz, J = 8 feet and average soil.

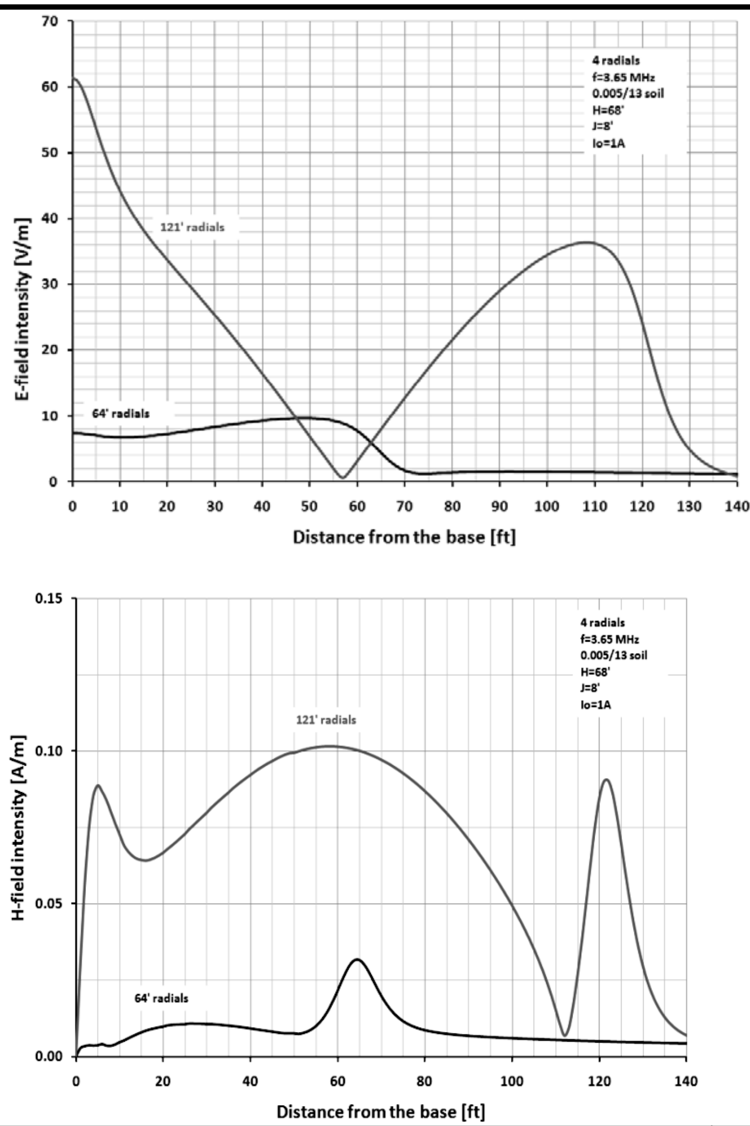

Figure 24 — E and H field intensities close to the ground surface directly below the radials with N = 4.

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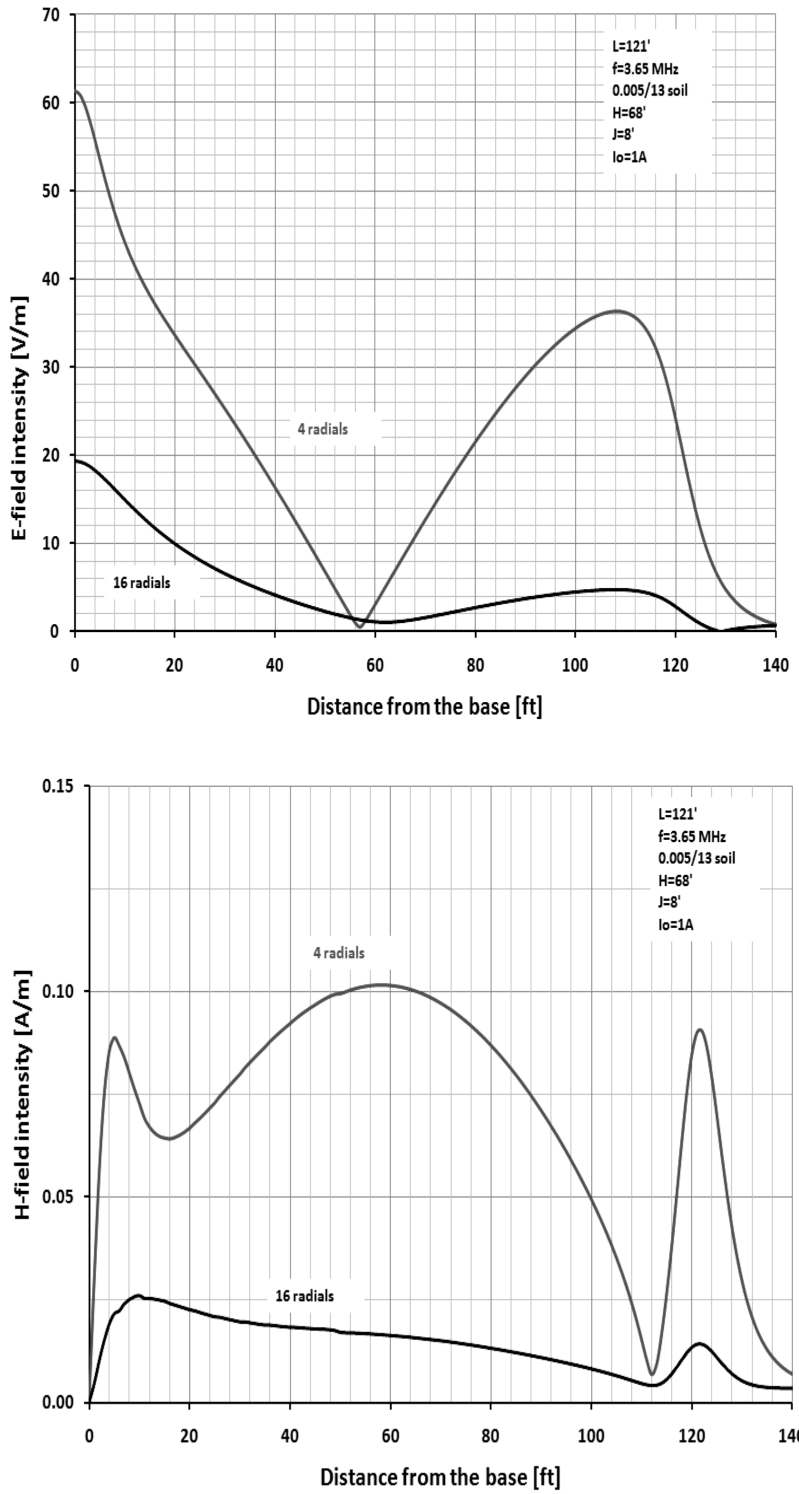
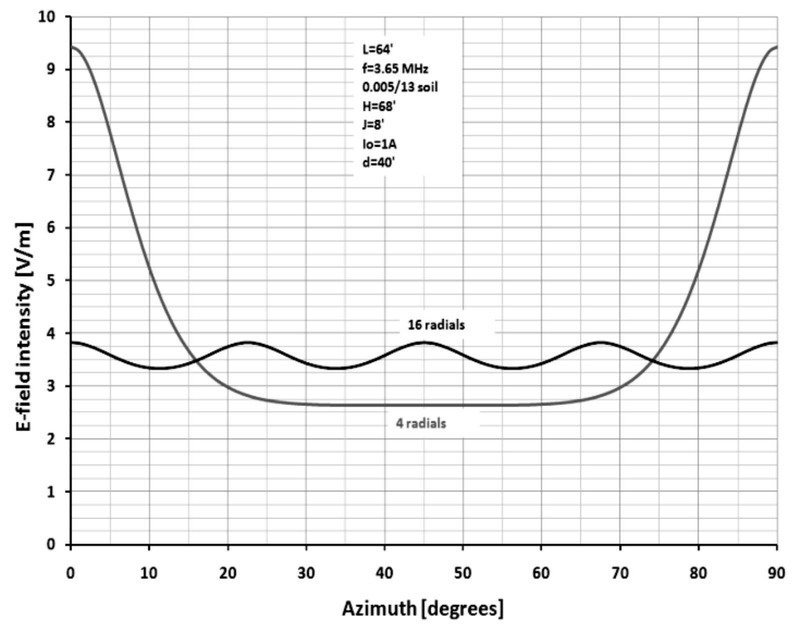


Figure 25 — E and H field intensities close to the ground surface directly below the radials. N = 4 and16.



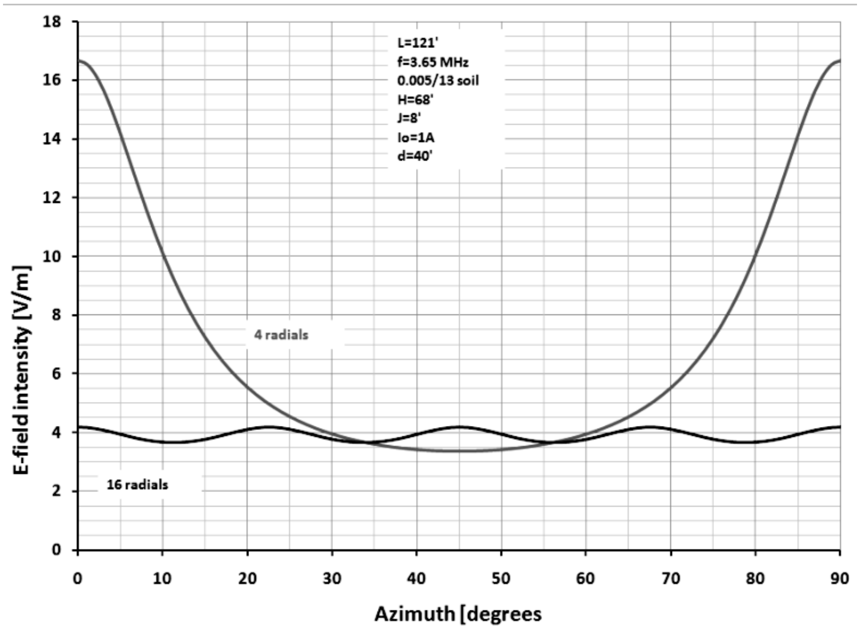


Figure 26 — E-field intensity just above ground on a 90° arc 40 feet from the base.

Tech Notes

[Tech Notes seems the ideal forum for Rudy to present some supporting documentation regarding his earlier stance on using foil conductors in antennas. Are the benefits obtained by using a thin foil element outweighed by increased edge-current losses? Read on and then decide for yourself. We are always in need of short, interesting technical articles for future Tech Notes columns. If you have something that may be of interest, please contact us.— Peter Bertini, K1ZJH, *QEX* Contributing Editor, **k1zjh@arrl.org]**

Resistance of Foil Conductors For Antennas

By Rudy Severns, N6LF

In the Nov/Dec 2000[1] issue of *QEX,* I presented an overview of conductor resistance for antennas. One of the suggestions offered was to use foil conductors to reduce resistance for a given cross-section of copper. Obviously, just using round copper wire of larger and larger diameters would be a heavy and expensive way to reduce conductor loss in low-impedance antennas. The premise was that the resistance of a round wire (which is more than a few skin depths in diameter) will be reduced by rolling it out into a foil. Several readers challenged this, stating that "The current in a foil is concentrated at the edges, and so, in effect, you don't gain anything." This illustrates a very common misconception, which I will address.

Relative Loss in Wider Foils

While it's certainly true that current densities at the edges of a foil can be much higher than in other parts, this does not mean that you cannot substantially reduce losses for the same area of copper by going from a round to a foil conductor.

I ran a very simple model using Finite Element Modeling (FEM) software,[2] which allows the loss in a given conductor of arbitrary shape to be determined at high frequencies, while accounting for eddy current effects. I chose a foil thickness of 8 mils and a frequency of 14 MHz, with a constant current of 1 A rms. The skin depth in copper at 14 MHz is about 0.7 mils, so this represents a relatively thick conductor. I then varied the width from 125 mils (1/8 inch) to 1000 mils (1 inch) and computed the losses. If it were true that all of the current would be concentrated in the edges, then making the foil wider should have little effect on the losses. However, if this view is indeed incorrect, you would expect to see the loss decrease as the foil is made wider.

The results are shown in Fig 1. The loss is normalized to 1 for a strip width of 125 mils. As we increase the width, loss decreases, but not as quickly as it would if it strictly followed the area ratio or dc resistance. It is pretty clear that the current is probably not entirely—or even largely—flowing in the edges, but there is something going on that is probably related to edge effects. Time to take a closer look!

Current Distribution in a Foil

One of the nice things about FEM CAD software is that you can graph the current density in the conductor. Fig 2 is a plot of the current density in the foil; the lines represent constant current densities. The greatest current density is indeed at the outer edge, and in fact, at the outer corners as indicated. Nonetheless, it is also clear that there is current flowing elsewhere. Because the foil is about 11 skin-depths thick, we see that there is essentially no current inside the conductor. This is due to skin effect and comes as no surprise.

Now let's look more closely at the

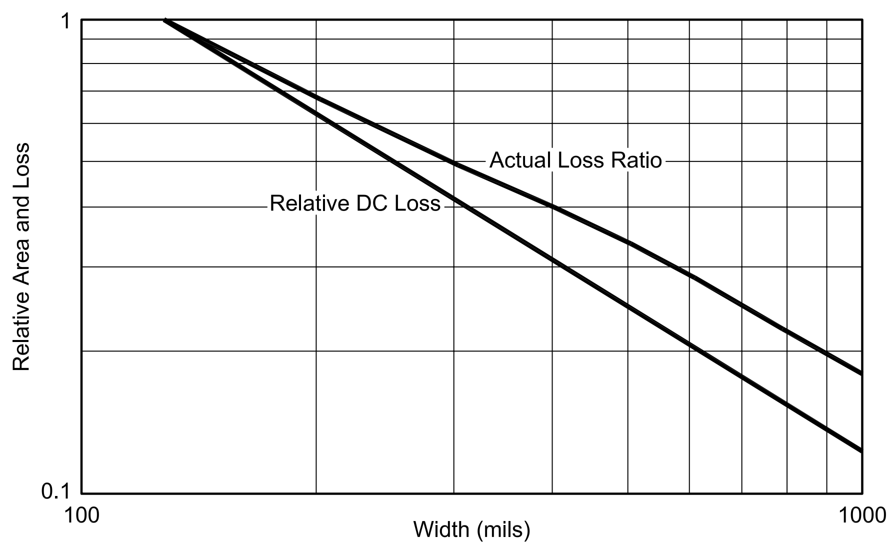


Fig 1—Comparison of dc and actual ac loss based on an increase of the 0.008-inch foil width.

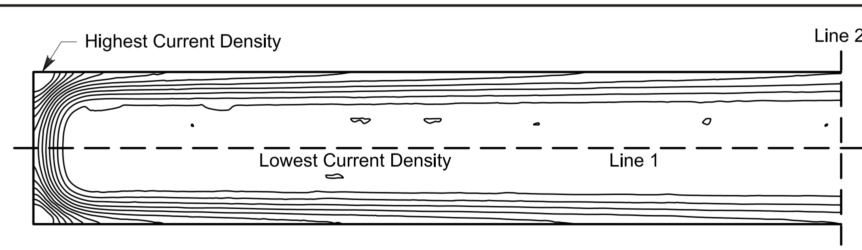


Fig 2—Current density distribution on the left half of the foil. By symmetry, the other half is a mirror image.

[1]Notes appear on page 64.

current density at the outer edge. Fig 3 is a graph of the current density along Line 1 defined in Fig 2. Sure enough, the current density at the ends is quite high, but the area of that region is relatively small so it represents only a portion of the total current in the entire conductor. There is significant current in other areas.

Fig 4 is plot of the current density along Line 2 in Fig 2, which is roughly at the middle of the foil. In line with what we know about skin effect, the current density is highest on the surface of the foil and decreases as we go inside. Yet, there is still significant current flowing on the surface of the foil away from the end edges. For this example, I chose a thick foil (11+ skin depths). Using a thinner foil would have shown that the edge effect was less pronounced, and in fact, thinner foils have less loss contributed by the edges.

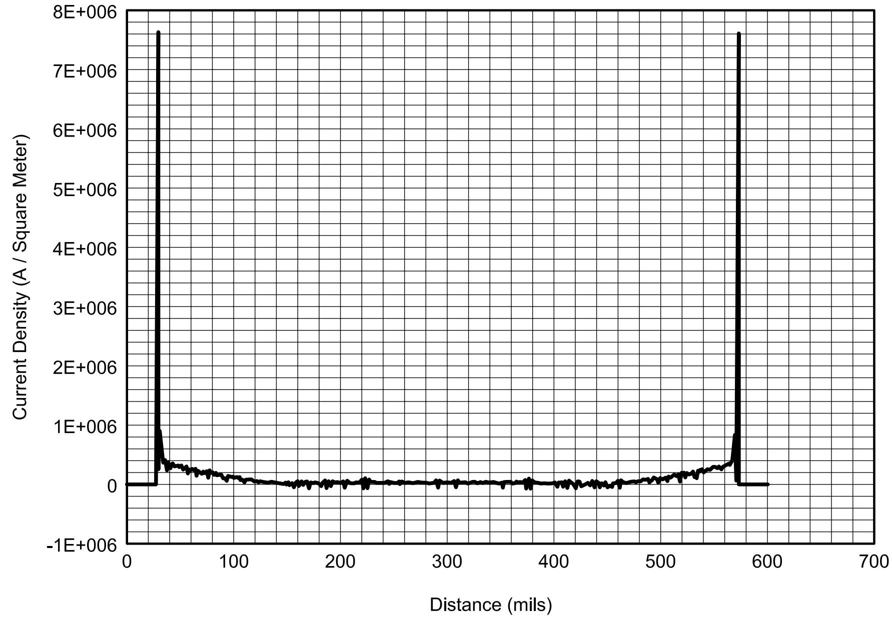


Fig 3—Current density in amperes per meter along line 1 (see Fig 2).

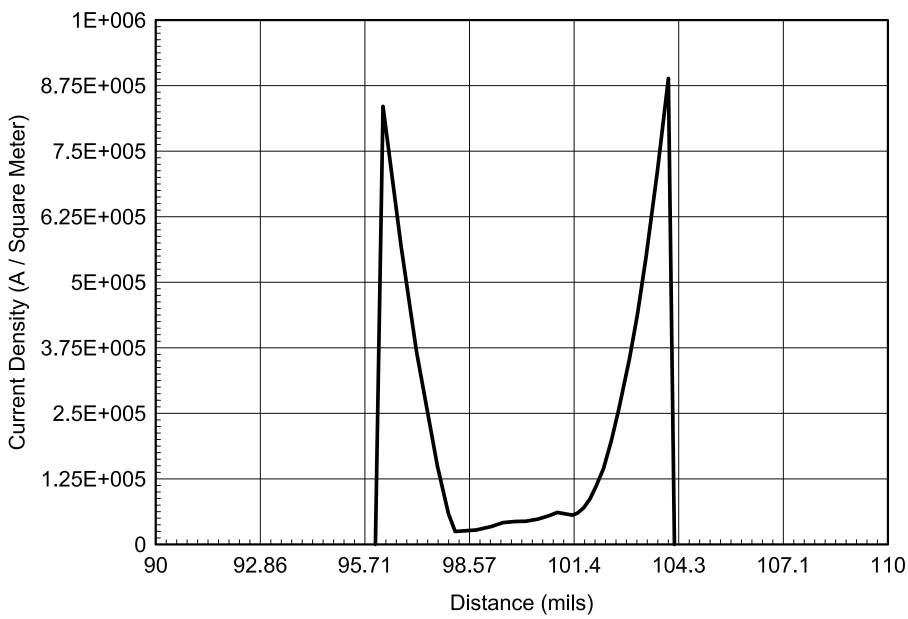


Fig 4—Current density in amperes per meter along line 2 (see Fig 2).

Summary

If a round wire is run through a roller so it flattens while keeping a constant cross-sectional area, we will discover that the HF resistance initially increases when the wire is formed into a square. It then begins to decrease as it is flattened further. As the conductor is made thinner, the resistance decreases, and when the thickness is about one skin depth, the difference between the ac and dc resistances will be small. There will also be little loss from the edges. All of this has been long known and experimentally verified in the early 20th century. Unfortunately, the idea that all the current flows in the edges is still part of our lore.

Another fact has been long known[3] but often forgotten: For a given external diameter (which is large compared to a skin depth), you can reduce the ac resistance by removing copper from the inside—that is, use a "thin-wall" tube. For a given diameter, the minimum ac resistance is reached when the wall thickness is roughly two skin depths.

Foil conductors do have disadvantages: They flutter in the wind, and very thin foils have little mechanical strength if the foil is unsupported. Several years ago, while building an antenna for my sailboat, I needed a low-loss and lightweight design to mount at the masthead. I bought some thin copper tape and applied it to a fiberglass fly-rod blank. It worked great and survived many thousands of miles of sailing across the Pacific. In effect, it was a "thin-wall" tube. Alternately, the foil could have been *inside* the fiberglass tube. Another time, losses in a stainless steel backstay antenna were reduced by bending a thin foil strip, with PVC tape on both edges, around the backstay in a U shape. This worked great. The copper tape was the conductor, and the stainless-steel backstay kept it and the mast supported.

I've noticed that the new motorized dipole being sold by Fluid Motion uses a copper-foil element inside a fiberglass tube. Reeling the foil in or out sets the length of the element. I think this shows that there is a practical use for foil conductors in some antenna installations.

Notes

[1]R. Severns, N6LF, "Conductors For HF Antennas," *QEX*, Nov/Dec 2000, pp 20-29.

[2]*Maxwell* software by Ansoft Corporation; **www.ansoft.com/products/em/max3d/index.cfm**.

[3]F. Terman, *Radio Engineers Handbook* (New York: McGraw-Hill, 1943), Fig 2, p 33. □□

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Experimental Determination of Ground System Performance for HF Verticals Part 4 How Many Radials Does My Vertical Really Need?

Experimental results to answer an often-asked question.

A frequently asked question is "How much of a ground system do I really need to make my vertical antenna work?" Usually, what's wanted is an answer in the form of "This much ground system will improve your signal by X dB." Another common question is "Does it matter if I lay the radials on the ground surface instead of burying them?" This is a practical consideration because it's often much easier to lay out the radials on the surface and let them vanish into the grass.

These questions can be addressed analytically and with modeling, but for most of us that's not very convincing. It's much more satisfying to see actual field measurements on real antennas. In the past there has been professional work at MF broadcast frequencies and also the excellent work by Jerry Sevick, W2FMI, at HF.[1, 2] The problem with an experimental approach is the practical limit on the number of test examples: you can't do all the possible variations! What's needed are reliable field measurements that can be compared to calculations and/or modeling to see if there is reasonable correlation. If there is, we can use calculations or modeling for the wide variety of antennas and soil characteristics we which we couldn't test.

Some of the material that follows represents a redo of Sevick's work with better instrumentation, but the material in this section, along with the other five parts of the series, goes well beyond Sevick's work. The details of the test equipment and experimental setup were given in Part 1 of this series.[3]

Efficiency Limitations

The purpose of the ground system is to improve antenna efficiency so that less power is lost in the soil and more is radiated. Efficiency is the ratio of the power radiated to the total input power at the feed point. Of course what we want is to radiate all the input power (100% efficiency) and maximize our signal, but there are practical limits. We can represent the resistive part of the feed point impedance (Rs) by three series resistors as shown in Figure 1.

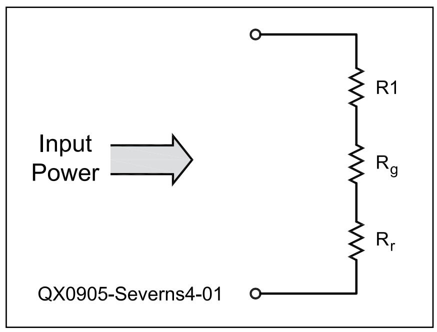

Figure 1 — An antenna input equivalent circuit. Rl represents the ohmic loss due to conductors, loading inductor series resistance, and so on. Rg represents the power dissipated in the soil by the near-field of the antenna. Rr is the radiation resistance, which accounts for the radiated power.

The input resistance at the feed point is Rs = Rr + Rg + Rl. We have to be a bit careful what we mean by "radiation resistance." Rr is usually defined as the value of the resistance at a current maximum attributable to radiation. In a vertical antenna with a height of ¼ λ or less over perfect ground, this point is at the base of the antenna, which is the usual feed point. In real antennas with various numbers of ground surface radials, however, the height of the antenna may have to be modified to maintain resonance and the current maximum may actually be out on the radials or possibly even back up into the vertical. What this means in practice is that the fraction of the feed point impedance we attribute to Rr may not be converging to the ideal value from theory as we add radials or change radial lengths. For example,

[1]Notes appear on page 42.

a resonant, very thin ¼ λ vertical over perfect ground will have Rr = 36.2 Ω but a real antenna may converge to a somewhat different value as we add radials and reduce ground loss.

With a ¼ λ vertical it is often assumed that if Rl is small, then Rg is simply Rs – 36.2 Ω. This is not the case and should not be assumed. The radiation resistance varies as the ground system changes, and does not approach 36 Ω until the ground system is relatively large. In a broadcast antenna with 120 radials 0.4 λ long, this approximation is very good, but in the limited ground system typical of amateur antennas at HF, it is not. A detailed discussion of this point can be found in an article available on my Web site, "Radiation Resistance Variation with Radial System Design."[4] (This may become a *QEX* article in the future.)

Because we are interested in the effect of efficiency on signal strength, it is handy to express efficiency (η) in terms of dB:

$$\eta = 10\mathrm{Log}\left(\frac{1}{1+\dfrac{Rg}{Rr}+\dfrac{Rl}{Rr}}\right) \qquad \text{[Eq 1]}$$

For 100% efficiency, Rl = Rg = 0 and η = 0 dB. If we increase Rl and/or Rg, η will decrease. For example 80% efficiency would be about –1 dB.

Experimental Tests

All of the measurements were made on 40 m, 7.2 MHz in most cases. I chose 40 m verticals for their manageable size. Even at that size, the ground system that had to be laid down and taken up numerous times, required over 2000 feet of wire.

I used five different antennas:

- A ¼ λ, 1 inch aluminum tubing vertical, adjusted to resonate at 7.2 MHz.
- An 1/8 λ, 1 inch aluminum tubing vertical with three top loading wires sloping at roughly 45°, again, resonated at 7.2 MHz.
- An 1/8 λ, 1 inch aluminum tubing vertical with no top loading, but resonated to 7.2 MHz with a base inductor.
- A 40 m Hamstick mobile whip (about 7.5 feet high), the top section adjusted for resonance at 7.2 MHz.
- A Cushcraft R7000 vertical.

The minimum conceivable ground system for a vertical would be a single ground stake with a coaxial feed line back to the shack. In this case, the feed line acts as a single random length radial. For these measurements I adopted this as the "zero radial" system, where the stake was a 4 foot copper-clad steel rod with ½ inch Andrews Heliax, buried 6 inches below the ground surface, back to the shack. The ground system was improved progressively by adding 33 foot (no. 18 AWG) radials in the progression: 0, 4, 8, 16, 32 and 64. This was repeated for each antenna. A ¼ λ in free space is close to 33 feet at 7.2 MHz. As was shown in Part 2, however, the *electrical* length of the radials changes when the radials are placed close to the soil.[5]

The soil characteristics under the radial system were measured using the technique given in *QEX*.[6] The average soil constants in the test field were: conductivity, $\sigma = 0.02$ S/m and relative dielectric constant, $\varepsilon_r = 30$. I will refer to this as "N6LF soil."

For each number of radials and each antenna, two measurements were made: the input impedance and the relative signal strength at a point 1.8 wavelengths away from the test antenna, at an elevation angle of about 8 degrees. Because the number of radials affected the resonant frequency, each antenna was re-resonated by adjusting its height as the number of radials was changed.

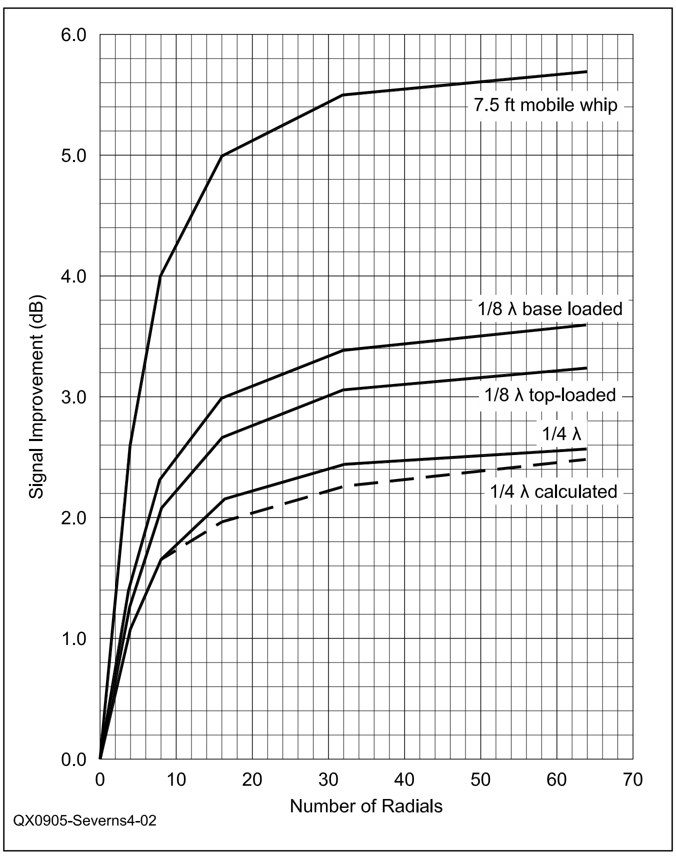

Figure 2 — Typical improvement in signal as ¼ λ radials are added to the basic ground system (a single ground stake).

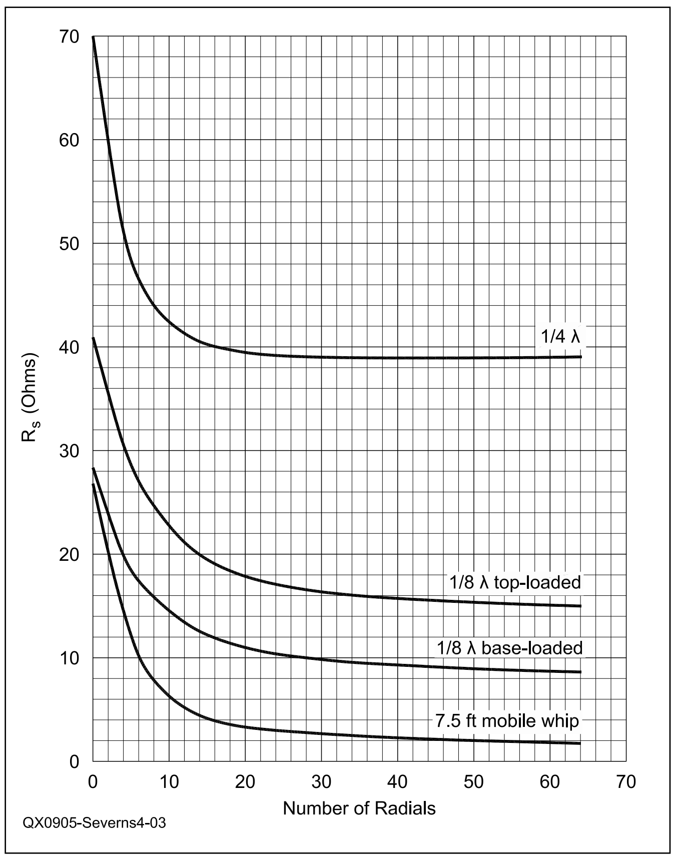

Figure 3— Measured input resistance (Rs) at resonance as a function of the number of radials.

Experimental Results

When we compare the results for different numbers of radials on a given antenna, the change in relative signal strength directly answers the question of how much signal improvement we get by adding radials. Typical test results are shown in Figure 2.

Note that the graph is in terms of the *improvement* in signal over the single ground stake with no radials for *each* antenna. The graph does *not* compare the relative worth between each antenna. Obviously a short, lossy mobile whip will yield much less signal (–10 dB or worse!) than the ¼ λ vertical.

The effect of radial number on input resistance (Rs) is shown in Figure 3.

In the case of the Hamstick mobile whip, I have subtracted Rl from the measured input resistance because it has a fixed value independent of radial number. Rl is determined by the loading coil Q. We can see that as we add larger numbers of radials the values for Rs begin to level out and approximate, but not equal, values for ideal lossless antennas.

Interpreting the Data

One of the interesting things about Figure 2 is that it shows that *the shorter and more heavily loaded the antenna, the more you have to "gain" from an aggressive ground system*. For example, the improvement for the ¼ λ vertical, going from 0 to 64 radials, is about 2.6 dB, but for the 1/8 λ base loaded vertical it's more like 3.4 dB, and for the mobile whip, nearly 6 dB.

What's going on here? As I pointed out in my July 2000 *QST* article on ground systems, when we shorten an antenna but keep the input power the same, both the magnetic and electric field intensities in the immediate vicinity of the antenna increase dramatically.[7] This translates to much higher ground losses. What we see in Figure 2 is that adding the radial system reduces these losses, but since the losses are higher to start with for the shorter antennas, the improvement is greater. No mystery!

From Figure 2 we can see that for all the test antennas, most of the improvement comes with the first 16 radials. As we add more radials beyond 16, there is still improvement but it is proportionately smaller. You gain perhaps another fraction of a dB going to 32 radials but by the time you reach 64 radials there isn't much change. The broadcast standard of 120 radials 0.4 λ long is hard to justify for amateur use, particularly given the present price of copper wire!

Figure 2 also has a dashed line very close to the curve for the ¼ λ vertical. This is a prediction using Abbott's calculation method.[8] I could have also added calculated lines for the other antennas and would have seen the same reasonable correlation, but that would have really cluttered the graph so I left them off.

We do have to be a little careful in using these graphs as general guides. They represent experimental results over my particular soil, at one frequency. Can we really draw any general conclusions? In lieu of running tests on all possible soils, we can get a feeling for this by calculating the signal improvement for different soils using Abbott's calculation method. (See Note 8.) Typical calculated results for different soils, at 7.2 MHz, are shown in Figure 4. This graph starts at 8 radials and goes to 64 radials. Smaller numbers of radials are omitted because the underlying calculation becomes inaccurate as the angle between the radials increases beyond 45°, the 8 radial case. From a practical point of view this is not a serious limitation. As I pointed out in Part 2 in the Jan/Feb 2009 issue of *QEX* (see Note 5), and as the data in Figure 2 shows, a four-radial ground system has very minimal performance; 8, or better yet 16 radials, should be the minimum, except perhaps in an emergency.

For the soil over which these tests were done (N6LF), the calculated 8 to 64 radial change is about 0.8 dB. Going back to Figure 2 we see that the measured change for the ¼ λ vertical is 0.9 dB (8 to 64 radials). The calculation agrees quite well with the measurements. Figure 4 tells us that when the soil is better, a given number of radials gives somewhat less improvement and with poorer soils there is more improvement. Again, no surprise. If you have better soil, you have lower losses to start with, so the improvement will be less. But even with very good soil it's still worthwhile to use at least 16 radials.

What about frequencies other than 40 m? There are a couple of complications to extending the 40 m work to another band. First, the graph in Figure 4 does not scale directly with frequency because the field intensity at a given distance (feet or meters), for a given base current, does *not* scale linearly with frequency. Second, at a given site the ground characteristics will vary with frequency. (See Note 6) The result is that the ground loss is not the same for the scaled antennas at other frequencies, even though the input power may be similar.

As we go down in frequency, soil conductivity typically decreases, which tends to increase ground loss but the relative dielectric constant goes up, which tends to decrease ground loss. For N6LF soil at 7.2 MHz, σ = 0.020 S/m and ε_r = 30, but at 1.8 MHz, σ = 0.013 S/m and ε_r = 68. The net effect on signal improvement (8 to 64 radials) is shown in Figure 5.

If you examine Figures 2 and 3 closely and compare the curves for the ¼ λ vertical, you may see something funny going on. In Figure 2, even when we go from 32 to 64 radials, there is still some improvement in signal. But if you look at Figure 3, there appears to be no change in Rs, so how can the antenna be more efficient? This same paradox shows up in the Brown, Lewis and Epstein data (see Note 1) taken 70 years ago, and has been the subject of comment ever since. What's going on? Several things are going on simultaneously. First, the number of radials is increasing, which reduces Rg. Second, we are steadily increasing the height

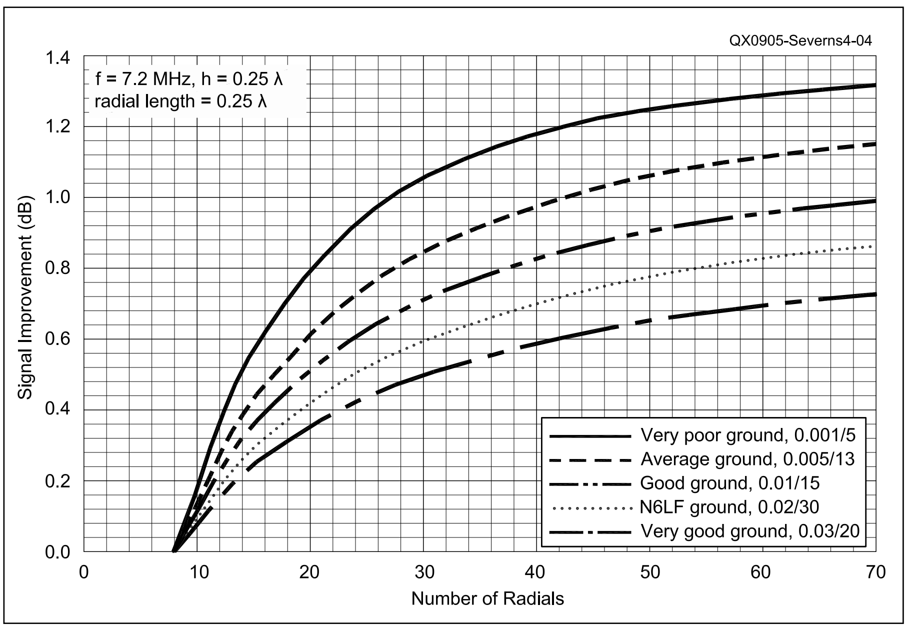

Figure 4 — Calculated signal improvement as we vary the number of radials over different soils with a ¼ λ vertical with ¼ λ radials at 7.2 MHz. Note: 0 dB is for the 8 radial case.

of the antenna to re-resonate it due to the effect of the radials on the ground, which we will look at shortly. This tends to raise Rs. In the case of the measurements for the ¼ λ antenna, the two effects cancel to some extent. Notice that for the other antennas, Rs is still trending down as signal strength goes up with number of radials. Altering the height as we add radials is not the full story, however, Rr is also affected by the radial system. (See Note 4.)

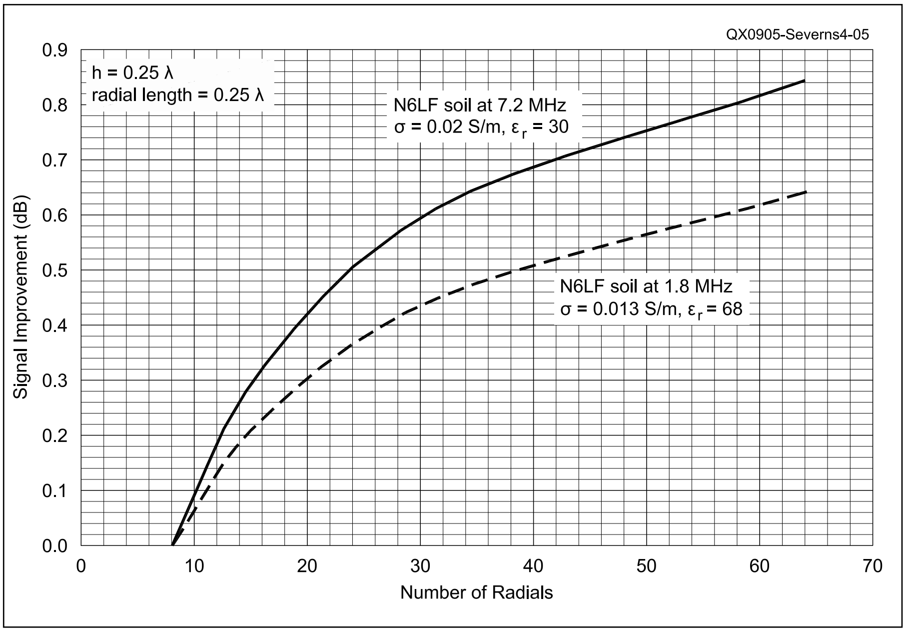

Figure 5 — Difference in signal improvement between 1.8 and 7.2 MHz over N6LF soil using the same vertical height and radial length in wavelengths (scaled with frequency). 0 dB is for the 8 radial case.

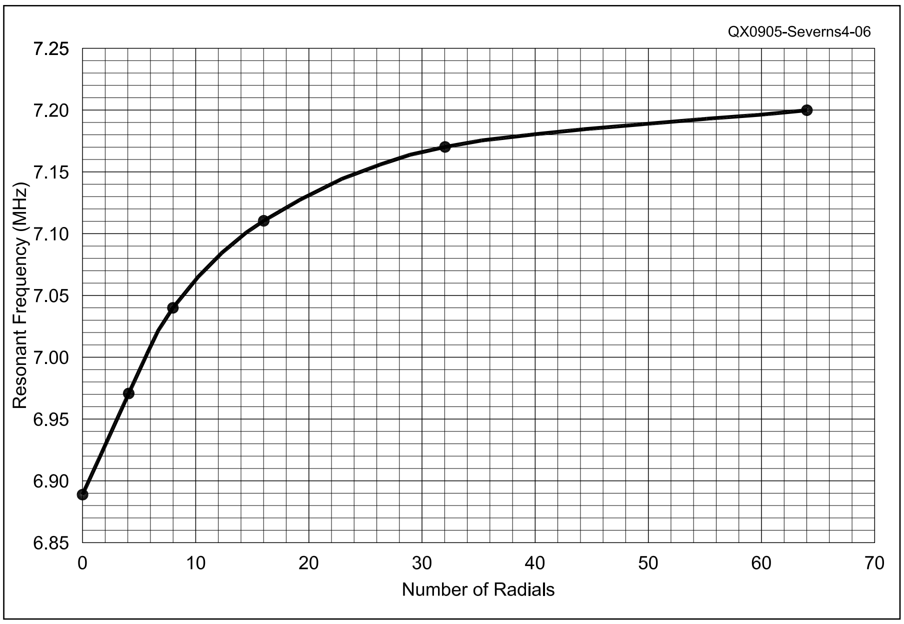

Figure 6 — Resonant frequency of a vertical antenna resonated at 7.2 MHz with sixty four 33 foot radials, as a function of the number of radials.

Additional Tests

In addition to the tests where antenna height and number of ¼ λ radials were the variables, I ran a few others. In one, I compared the performance of the ⅛ λ top-loaded vertical with 64 radials, with and without, an ⅛ λ circular ground screen (diameter = 36 feet) added over the radial fan. The addition of the ground screen made no detectable difference, which is in line with previous work. See Note 1. Obviously, if you have only a few radials, then a ground screen would help.

Modeling of gain versus radial number and radial length indicates that a larger number of shorter radials may be just as good or better than fewer longer radials, assuming both radial systems use the same amount of wire.[9] To check this out I ran a test using the top-loaded ⅛ λ vertical, comparing sixteen ¼ λ (33 ft) radials versus thirty two ⅛ λ (17 ft) radials. In line with the modeling and also calculations, the signal strengths were almost the same. The feed point impedances were substantially different however. I had to lengthen the vertical to re-resonate it with the 32 short radials. This is a good example of the interaction between the feed point impedance and the radial system. If space is restricted, then more short radials in place of fewer long radials may work just fine, but to properly evaluate that option it would be best to do the modeling or calculation for a particular vertical and soil characteristics.

I made measurements on the R7000, with and without an external ground system, which showed that adding a 64 radial ground system had almost no effect on signal strength (+0.1 dB). This surprised me until I had an e-mail conversation with Joe Reisert, W1JR, the original designer. The antenna was designed to work without a ground system and although the antenna is physically less than ¼ λ on 40 m (25 ft), the loading is arranged so that it behaves more like a ⅜ λ. There are a set of 48 inch radials at the base, which are isolated from ground. The current maximum is well up into the antenna and the base is a high impedance point. The conventional wisdom, to which I have been a subscriber, is that even with a ½ λ vertical, adding an extensive ground system will improve performance. I did not see that here. This is a subject for more experiments, perhaps.

Measured Resonant Frequency

During the experiments, I found that changing the number of radials changed the resonant frequencies of all the antennas except the R7000. For example, using the ¼ λ vertical, I laid down 64 radials and adjusted the height of the vertical so that it was resonant at 7.2 MHz. I then started removing radials (but not changing the height), measuring the resonant frequency as I went down to zero radials. The results are shown in Figure 6.

Obviously the resonant frequency is affected by the radials. You can of course re-resonate the antenna by changing its height or loading. During the experiments for signal strength and input impedance, I adjusted the height to restore resonance at 7.2 MHz. With 64 radials resonance at 7.2 MHz was obtained with h = 33 feet 7 inches. With no radials, the 7.2 MHz resonant height was 32 feet 11 inches, 8 inches shorter.

What's going on? When there are no radials, only the ground stake and the random length of feed line, the resonant frequency is low primarily because the upper portion of the stake effectively adds to the antenna height. Even though the stake is driven into the soil, the top layer of soil, at least in summer when these measurements were made, is quite dry. The effective ground surface is actually somewhat below the physical surface. There was also some inductance in the lead connecting to the ground stake. As we add radials this effect is reduced but only slowly because, as shown in Part 2 (see Note 5), the radials are heavily loaded by their close proximity to the soil. They are resonant below 7.2 MHz so they are inductive at 7.2 MHz. This shunt inductance is across the base of the antenna. As we add more radials we are adding more inductors in parallel, which reduces the effective reactance and increases the resonant frequency.

Conclusions

The answer to our original question, "*Does laying the radials on the surface matter?*" is a little clearer now. *For the same number of radials of the same length*, the efficiency will be pretty much the same whether buried or on the surface, but the effect on feed point impedance may be somewhat different. This can become a practical problem if the antenna tuning varies with the season (wet or dry or frozen ground). Radials lying on the ground surface really behave more like elevated radials even though they may be lying right in the dirt.

We can summarize all this with the following advice:

- Try to use at least sixteen ¼ λ radials.
- If you don't have the space for ¼ λ radials, lay down a larger number of shorter ones.
- More than 16 radials will help but give only a fraction of a dB over average or better soils.
- The shorter your antenna, the more you need a good ground system.
- The poorer your soil the more you need a good ground system.
- A surface-radial ground system will affect the resonant frequency and you may have to adjust the vertical height for that.
- Work hard at making the antenna itself more efficient. In other words,. use high-Q loading coils, use top loading to minimize the size of loading coils, minimize conductor loss, and so on.
- Modeling and calculations seem to be in reasonable agreement with measurements and, with some caution, can usefully be used to estimate the magnitude of improvement when adding to a ground system.

Acknowledgments

This work was inspired by the classical articles by Jerry Sevick, W2FMI, which have served us so well.[2, 10, 11, 12, 13] In many ways my experiments are just an update and reconfirmation of Sevick's work.

I want to thank Mark Perrin, N7MQ, for his help in making many of the measurements. Especially helping to drag the monstrously unwieldy chicken wire ground screen into position and out again.

In addition to the references already cited in this article, I have included several more related references, which the reader may find useful. See Notes 14 through 21.

Notes

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QEX

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A Closer Look at Vertical Antennas With Elevated Ground Systems—Part 2

N6LF shares his results from further HF vertical antenna experiments.

[Part 2 concludes this article, which began in the Mar/Apr 2012 issue of *QEX*. — Ed.]

Multiband Verticals

For a single band antenna we can avoid the problems of long radials by simply using radials that are short enough or by increasing the number of radials, but what about the case of multiband antennas where you typically have four λ/4 radials for each band? For example, if you have 40 m λ/4 radials, these will be λ/2 on 20 m, ¾ λ on 15 m, and so on. In light of the information we found for G_a as a function of L in Part 1, is that a problem? I don't have the space here to explore it in detail with modeling, but I have looked at multiband elevated verticals experimentally. The information was in Part 5 of my *QEX* series, "Experimental Determination of Ground System Performance for HF Verticals." Part 5 was in the July/August 2009 issue of *QEX*, pp 15-17. That series of articles is available for viewing on my website: (**www.antennasbyn6lf.com**). The experimental work indicated that as long as there are a large number of radials (whether they are the same length or of different lengths) you don't have a problem but if you try to use only a few long radials you will have problems. Read the article for the details.

Potentials on the Radials

As Laport stated, elevated ground systems can have very high voltages between the wires and ground. Figure 27 shows examples of the voltage from a radial wire to ground for ideal 4, 12 and 32 $\lambda_0/4$ radial systems.

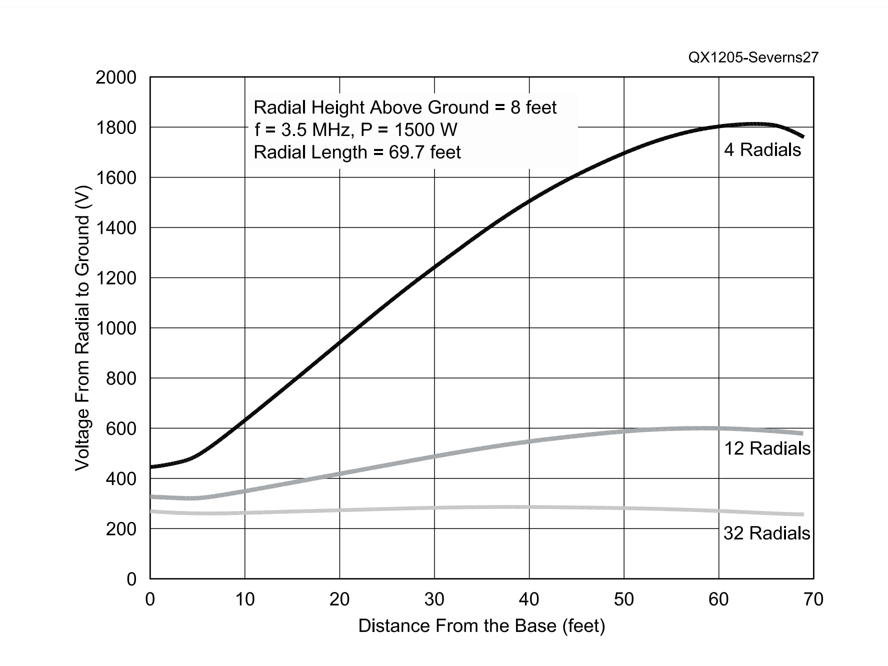

Figure 27 — Examples of the voltage from a radial wire to ground with different numbers of radials. The input power to the vertical is 1500 W, the operating frequency is 3.5 MHz and the radial system is elevated 8 feet above ground.

I think this Figure makes it clear why you want to keep the radials out of reach! Note that as more radials are added the potential difference between the radials and ground drops significantly and becomes more uniform as we go away from the base of the antenna. This is a reflection of the reduction in E-field amplitude with more numerous radials, as was shown in Figures 24, 25 and 26 in Part 1 of this article (Mar/Apr 2012 *QEX*). Even with a large number of radials that voltage is still high. This voltage will vary with the square root of the power level so that going down from 1500 W to 100 W, a change of 15:1 (0.067), the voltage only drops by 0.26! Be careful!

Feed Point Impedances

The behavior of the feed point impedance over the band (3.5 to 3.8 MHz for these examples) as we vary H, L, J, N and soil characteristics is an important factor. The point I want to make in this section is how widely the input impedance of ground-plane antenna can vary as we change one or more of the variables. There is no one number for Z_{in}! We will also look at variations in SWR bandwidth.

A graph of the feed point impedance ($Z_{in} = R_{in} + j\ X_{in}$) from 3.5 to 3.8 MHz for different numbers of radials is shown in Figure 28. Note that in Figures 28 to 31, H = L and is adjusted so that the model is resonant at 3.65 MHz for each variation of parameters. As the parameters N, J and soil characteristics are changed, the values for H and L vary somewhat. From Figure 28 we can see that N has a strong effect on the feed point impedance (Z_{in}) although that effect diminishes as N increases. As shown in Figure 29, we can convert the information in Figure 28 to SWR. In this case the Z_0 impedance for the SWR calculation is taken to be R_{in} at resonance (3.65 MHz) for *each* value of N.

Figure 29 shows that the 2:1 SWR bandwidth increases somewhat as N is increased but by N = 16 we are approaching the point of vanishing returns for bandwidth.

Figure 30 shows the effect of height above ground of the radial fan (J) on Z_{in} for N = 4. It's pretty clear that the value for J has a strong effect on Z_{in}. The effect of different soil characteristics for a given value of J (8 feet in this example) is shown in Figure 31.

The information in Figures 28 to 31 represents only a few possible combinations, but the graphs make the point that the feed point impedance of an elevated radial vertical is a strong function of all the variables, so that each installation is unique.

We can also see the behavior of Z_{in} over the band for different combinations of H and L that are resonant at 3.65 MHz. Some examples are given in Figure 32 and the associated graphs for SWR, are given in Figures 33 and 34. N = 4 and the H&L combinations are shown on the graphs.

The combination H = 73.25 feet and L = 43.11 feet has the very nice property that Z_{in} = 50 Ω at 3.65 MHz. As shown in Figure 33, this results in a relatively wide 2:1 SWR bandwidth compared to the other combinations.

The greater match bandwidth is not just because Z_{in} = 50 Ω at resonance. The combination also has intrinsically more bandwidth as shown in Figure 33, where the Z_0 at resonance is set to R_{in} at resonance for *each* combination of H and L separately.

The idea of increasing the feed point impedance at resonance to 50 Ω by making the vertical taller and the radial fan radius smaller has actually been around for many years: R_{in} at resonance can be increased by sloping the radials downwards from the base. In effect you are making the vertical taller and reducing the radial fan radius, which is what we did in the above example.

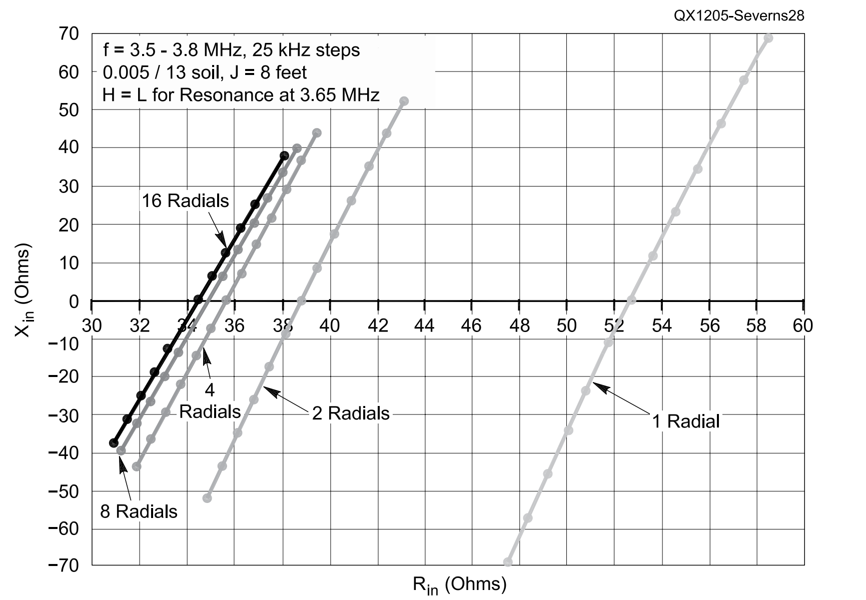

Figure 28 — X_{in} versus R_{in} ($Z_{in} = R_{in} + j\ X_{in}$) where frequency varies from 3.5 MHz (lower left ends of the curves) to 3.8 MHz (upper right ends of the curves) for different values of N. Frequency is stepped in 25 kHz intervals.

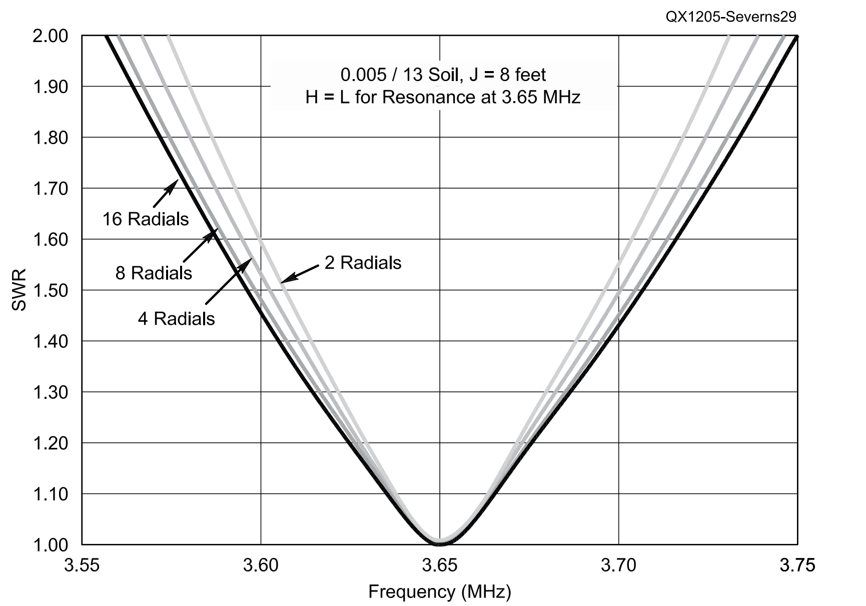

Figure 29 — Feed point SWR as a function of N.

Figure 9 (in Part 1) showed how L_r varied for different values of N and H. For H = 69 feet, L_r decreased rapidly as more radials were added. We can play this game to find designs where Z_{in} = 50 Ω at resonance. Figure 35 is a graph where L is varied from 15 feet to 100 feet for two values of H (72 feet and 77.6 feet). Note that H in the range of 72 feet <=> 77.6 feet represents the limit that allows R_{in} = 50 Ω. Longer or shorter values for H do not have a point where R_{in} = 50 Ω for L = 15 feet <=> 100 feet. The combination of H = 72 feet, L = 25 feet, N = 16 and J = 8 feet over average ground will give us $Z_{in} = 50\ \Omega$ at 3.65 MHz. Figure 36 shows the comparison for SWR between two combinations where N = 4 and N =16. This illustrates one of the advantages of using more radials.

For H = 72 feet and N = 16, L is only 25 feet that represents a drastic reduction in the radius of the radial fan. In exchange for an increase in height on the order of 6 feet, we have a good match over a wide portion of the band and a small diameter radial fan. Instead of increasing the height we could have just added a couple of short top-loading wires. This is very nice but it's not entirely for free. When compared to the normal four radial system (H = 67 feet, L = 67.7 feet), G_a for the H = 72 feet, L = 25 feet combination is lower by about 0.25 dB. You sacrifice a small amount of gain. Whether that is acceptable for the improvement in matching is an individual decision.

In a private communication with Dick Weber, K5IU, he made a suggestion that overcomes the reduction in gain associated with small radial length: use longer radials. This will result in $X_{in} \neq 0$ but you can tune out the reactance with a series impedance. He has also pointed out that if X_{in} is inductive (+) then you can tune out the reactance with a series capacitor at the feed point. Looking back at Figure 35, we see that this trick will work for H > 72 feet. (That is for this particular case, where N = 16, J = 8 feet over average ground!). If we chose H = 75 feet, L = 70 feet, N = 16 and adjust the series capacitor at the feed point as we move across the band, we get the result shown in Table 1. Note that X_{in} is given in the Table, but C_s (the added series capacitor) tunes it out.

What we see is a vertical that can have a very low SWR across the entire 75/80 m band. It isn't necessary that C_s be adjusted at every point. Three or four values of C_s switched with relays would probably still provide acceptable SWR over the entire band. For the case where H = 72 feet, L = 25 feet and N = 16, G_a = –5.52 dB. When we change to H = 75 feet, L = 70 feet and N = 16, G_a = –5.03 dB. That's an improvement of +0.5 dB in signal strength.

There is another option to make $Z_{in} = 50 + j0\ \Omega$ at resonance. Instead of making the antenna taller (or top-loading it) and the radials shorter, you can simply shift the feed point up into the vertical to a point where $R_{in} = 50\ \Omega$. This is just a matter of moving the base insulator up into the antenna. You won't get quite as much match bandwidth as with the taller vertical but it will be close and you can use longer radials that give a better G_a. Whether this trick is mechanically feasible depends on the particular implementation.

All the examples to this point have assumed that the excitation at the base of the

Table 1

Z_{in} and SWR from 3.5 to 4.0 MHz for H = 75 Feet, L= 70 Feet and n = 16

Frequency (MHz)	R_{in} (Ω)	X_{in} (Ω)	C_s (pF)	*SWR*
3.50	43.7	69.6	654	1.14
3.65	49.4	113.7	384	1.01
3.80	56.0	159.4	263	1.12
4.0	66.6	223.6	178	1.33

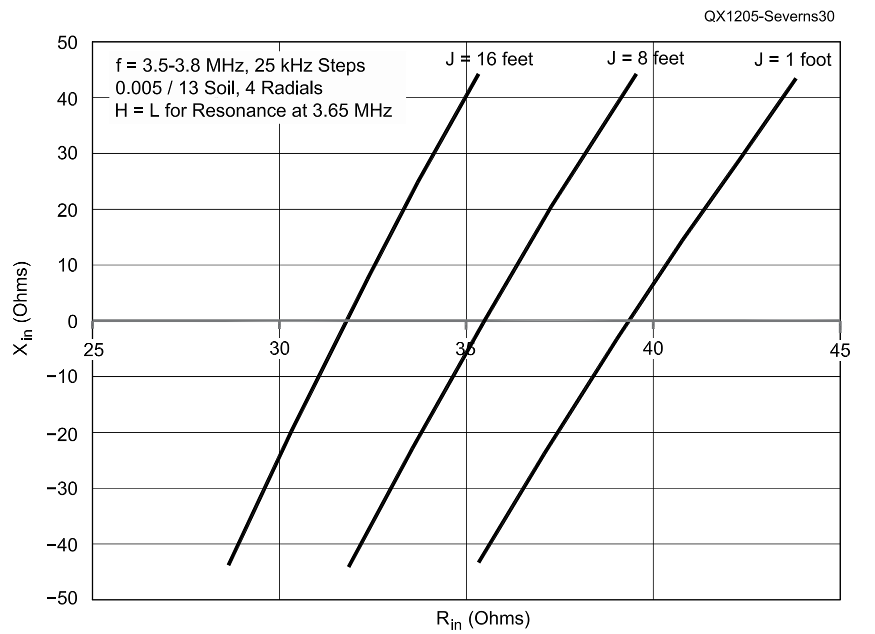

Figure 30 — The effect of height above ground on Z_{in}.

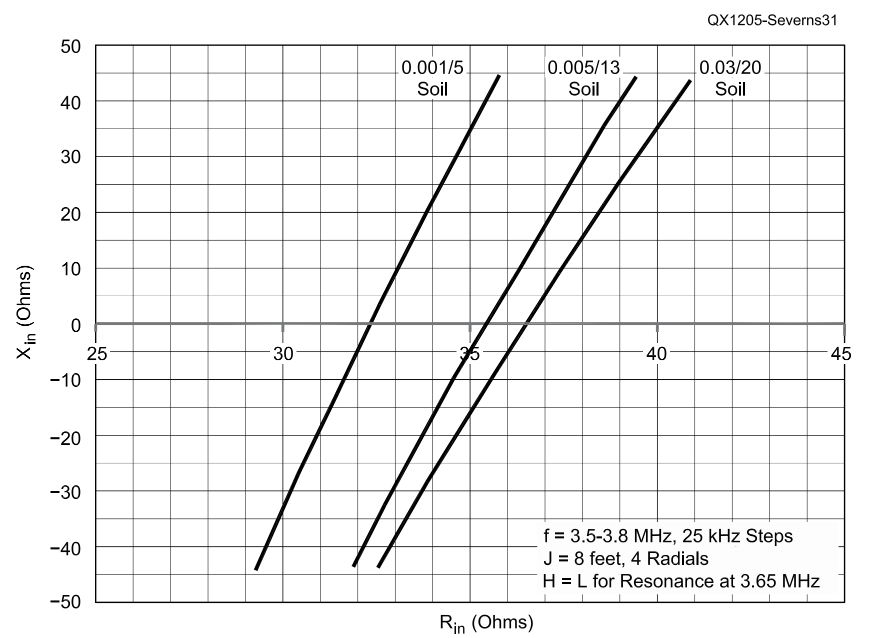

Figure 31 — The effect of different soil characteristics on Z_{in}.

vertical was isolated from ground: a choke (balun) was used in series with the feed line. If a choke is not used and the coaxial feed line is simply connected to the antenna and run down to ground, usually with the shield connected to the radials and the center conductor to the vertical, there will be additional ground currents that increase loss. In a 4-radial elevated system, G_a typically falls -0.25 to –0.5 dB or even more for lossy soils if a choke is not used. If 12 to 16 radials are used, the increased loss is much smaller, usually only a few tenths of a dB. You might argue that when N is large a choke is not needed but I think it is better to be cautious and use a choke even in that case.

Earlier we saw how the radial length (L) affected the efficiency (G_a) of the antenna. We also saw that the effect was reduced when more radials were used. It is useful to look at Z_{in} as both N and L are varied, especially around values of L near $\lambda_0/4$. Figure 37 shows the effect of varying L on X_{in}.

Figure 37 is particularly interesting in that it shows how sensitive the X_{in} component of Z_{in} is to radial length when only a few radials are used. The R_{in} component is not nearly as sensitive. This becomes important when we look at current asymmetries in the radials. Adding more radials reduces the sensitivity of Z_{in} to radial length and also the susceptibility to radial current asymmetry. Dick Weber, K5IU (see Note 43) generated a graph very similar to Figure 37 by assuming the radials were open circuit transmission lines and plotting the impedance at the feed point as more radials were added in parallel. I have more on radials as transmission lines in the next section.

Effect of Asymmetries in the Radial Fan

Is there significant current division asymmetry among the radials of typical installations and, if there is, do we need to be concerned about it? To answer the first part of this question, Dick Weber, K5IU, made a series of measurements on representative 80 m and 160 m verticals with two and four elevated radials. Dick's work was published in "Optimum Elevated Radial Vertical Antennas," in the Spring 1997 edition of *Communication Quarterly*. (See Note 9 in Part 1 of this article.) I have summarized some of his data in Table 2 but I strongly recommend reading his complete article.

Data tables are helpful but sometimes a graph of the data has more impact. Figure 38 compares the radial current division for Weber's 80 m vertical with four radials. Figure 38 shows two things: the radial current division between the radials is far from equal and the division ratios change as we move across the band. Unfortunately, this is typical of elevated ground systems with only a few radials, as shown in Table 2.

Weber explains this behavior by pointing out that a horizontal radial above ground is actually a section of single wire transmission line open-circuited at the far end so that in the region where $L \approx \lambda_0/4$ it acts like a series resonant circuit. Figure 39 shows an equivalent circuit.

Individually the radials may have differ-

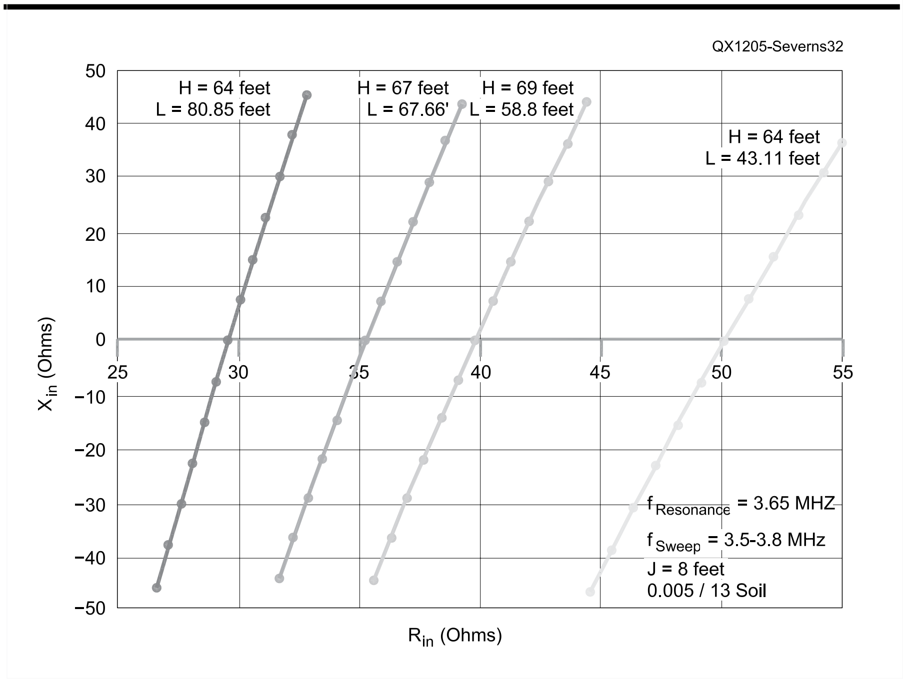

Figure 32 — Z_{in} variation for different combinations of H and L that are resonant at 3.65 MHz.

Table 2
Radial Current Comparisons from K5IU Measurements
(See Note 9 in Part 1 of this article for a reference to the source of this data.)

Antenna #	*Station ID*	*Frequency (MHz)*	*Relative Current Radial 1*	*Relative Current Radial 2*	*Relative Current Radial 3*	*Relative Current Radial 4*
1	K5IU	3.528	1.00	0.52	0.27	0.27
1	K5IU	3.816	0.96	1.00	0.51	0.51
1	WXØB	1.805	1.00	0.01	-----	-----
1	WXØB	1.885	1.00	0.05	-----	-----
2	WXØB	1.805	1.00	0.80	-----	-----
2	WXØB	1.885	1.00	0.10	-----	-----
1, 0.125 λ radials, w/inductor	WXØB	1.805	1.00	0.83	-----	-----
1, 0.125 λ radials, w/inductor	WXØB	1.885	1.00	0.76	-----	-----
1	W7XU	1.805	1.00	0.00	0.00	0.00
1st trim	W7XU	1.805	0.03	1.00	0.10	0.07
Last trim	W7XU	1.805	1.00	0.00	0.00	0.00
Last trim	W7XU	1.900	0.03	1	0.10	0.07

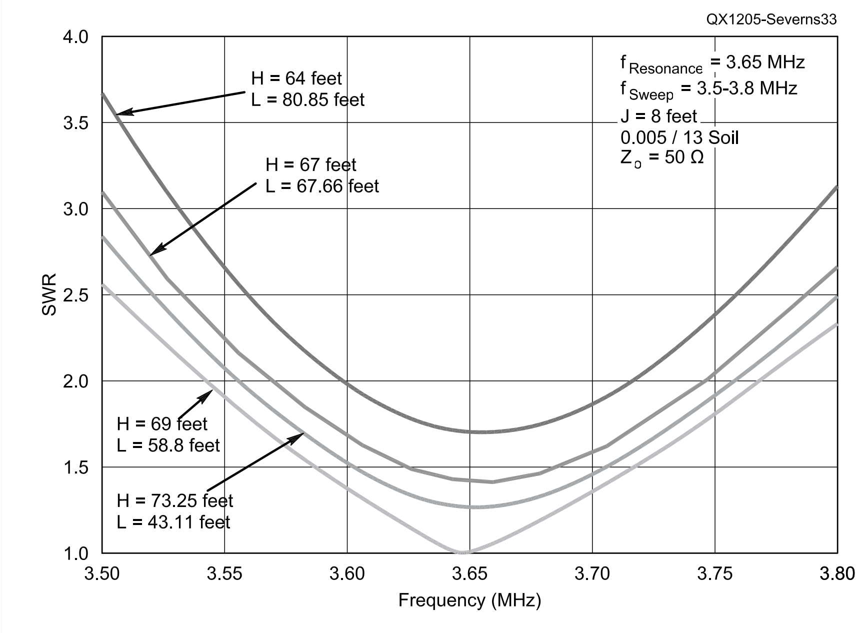

Figure 33 — SWR for various combinations of resonant H and L. Z_0 = 50 Ω for all curves.

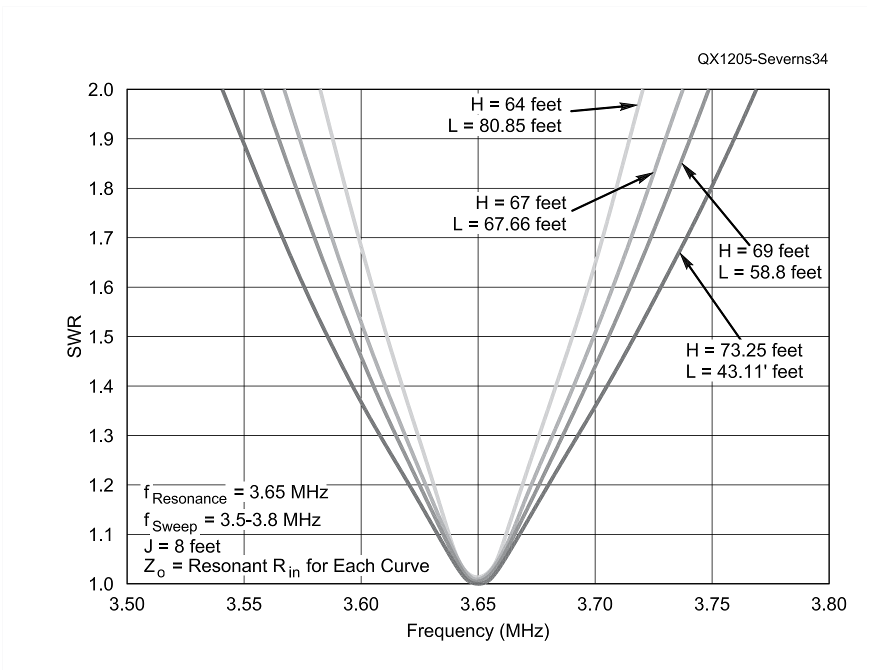

Figure 34 — SWR with Z_0 equal to R_{in} at resonance for the particular combination of H and L.

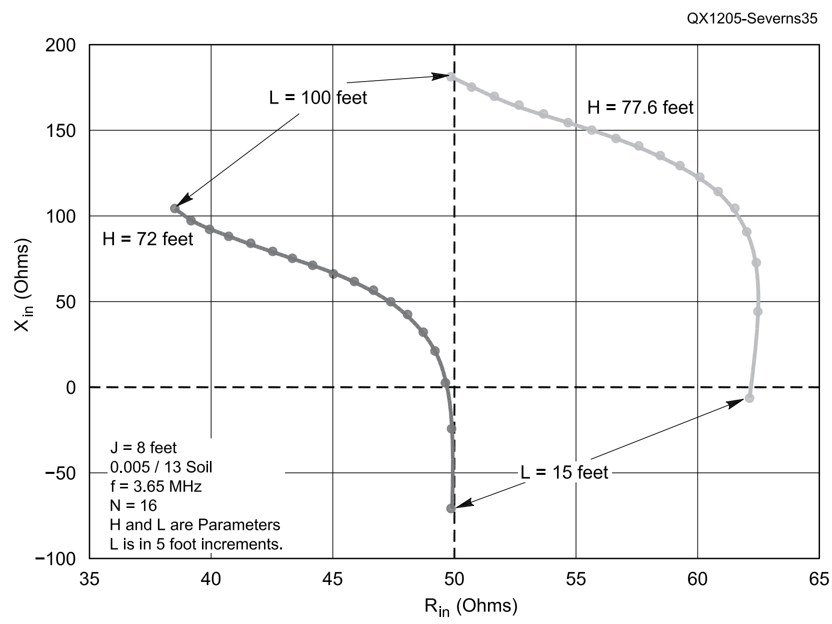

Figure 35 — Z_{in} as a function of radial length (L) for H = 72 feet and 77.6 feet with N = 16.

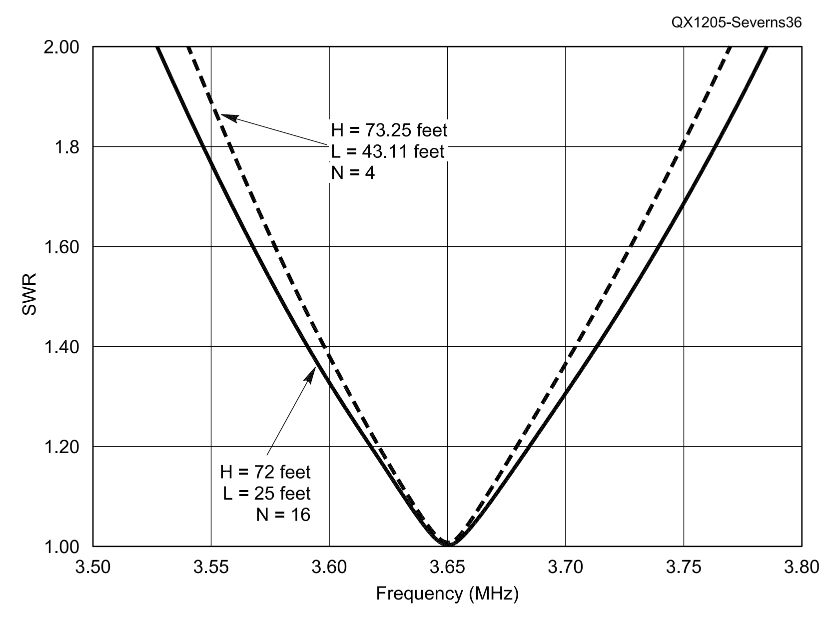

Figure 36 — SWR over 3.5 to 3.8 MHz for two different combinations of H and L.

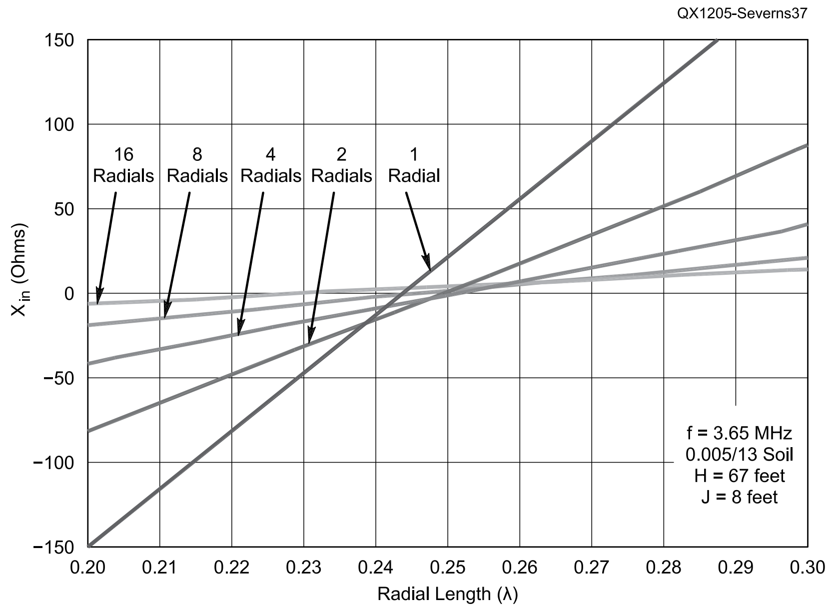

Figure 37 — Effect of changing L in the neighborhood of λ/4 as a function of radial number.

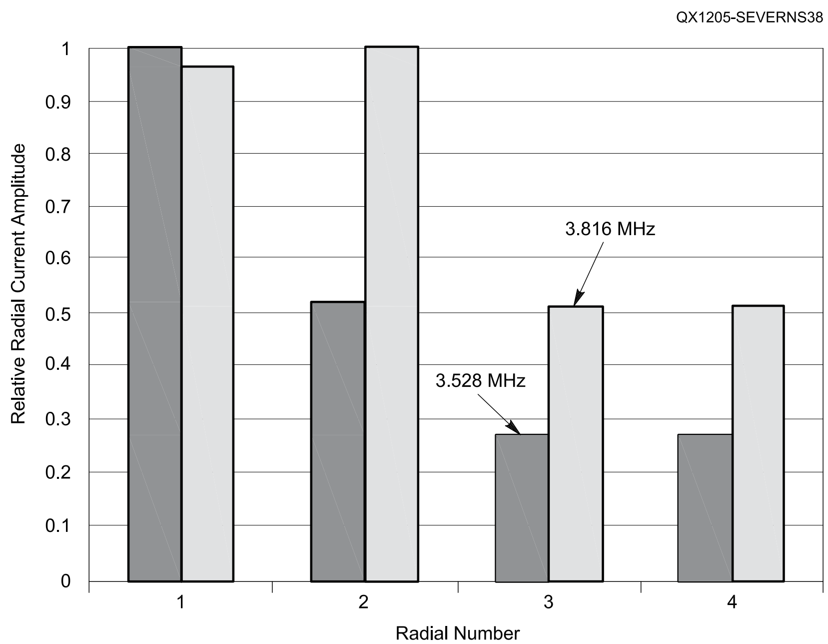

Figure 38 — Current division between the four radials at 3.528 and 3.816 MHz for the 80 m vertical at K5IU.

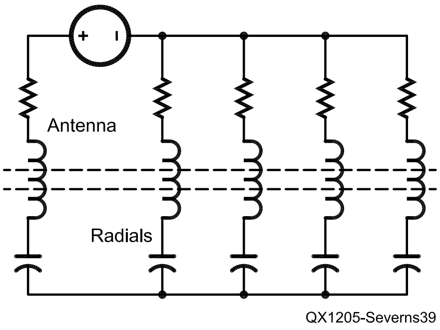

Figure 39 — Equivalent circuit for a vertical with elevated radials.

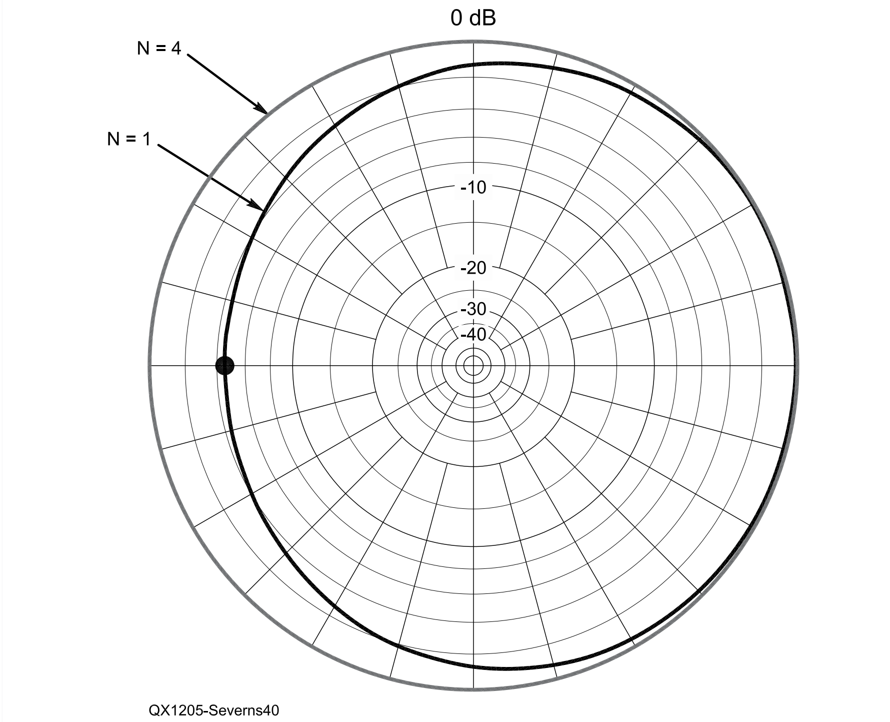

Figure 40 — Azimuth radiation pattern comparison between one and four elevated radials. J = 8 feet, f = 7.2 MHz over average ground. The elevation angle for these plots is 22°.

ing resonant frequencies due to length variations, varying ground characteristics under a particular radial, nearby conductors, and other factors. (See Note 12 in Part 1, Doty, Frey and Mills, "Efficient Ground Systems for Vertical Antennas," *QST*, Feb 1983, p 20.) At a given frequency, a particular radial may be close to series resonance, which means it has a low input impedance and may therefore take the majority of the current. This is a reasonable idea but the basic model in Figure 39 doesn't take into account the coupling between the individual radials or between the radials and the vertical. It would be more correct to add mutual coupling between all the inductive elements of Figure 39 as shown by the dashed lines. In the case of four radials, the radials are at right angles to each other and to the vertical so that the mutual coupling is small (but not zero). When you go to eight radials, for example, the angle between the radials goes from 90° to 45°. That greatly increases coupling between the radials.

All this is very interesting but so what? Does current-division asymmetry in the radials cause any problems we should worry about? One way to look into this is to model a system with only one radial, which might be a worst case. Several of the examples in Table 2 show almost all the radial current to be in one radial. Figure 40 shows a comparison in the azimuth radiation patterns between one and four radials with J = 8 feet and f = 7.2 MHz, at an elevation angle of 22°. Note that I have changed from 80 m to 40 m for the following examples simply because this work was already on hand. With four radials, the pattern is symmetric within 0.1 dB but with only one radial the pattern is distorted with a F/B ratio of 4.6 dB. In addition, the average gain for one radial is about 0.5 dB lower than G_a with four radials. There is substantial signal reduction (almost 5 dB!) in the direction away from the single radial. Over poor soil, G_a is even lower and the F/B can be 6 dB or more.

Does having all the current in one radial actually represent the worst case or can we have even more pattern distortion and/or lower G_a in some other cases? NEC modeling can be used to investigate this question. We'll start with a 40 m $\lambda/4$ vertical with four radials (see Figure 4 in Part 1). Radials 1 and 2 form an opposing pair with a length = L. Radials 3 and 4 are a second opposing pair with length = M. First we'll model the antenna with all the radials the same length (L = M) and then with radials that differ in length ($L \neq M$).

The feed point impedances for three different radial length configurations are compared in Figure 41 as the frequency is varied from 7.0 to 7.3 MHz. The plot on the left is for the case where all the radials are identical (L = M = 34.1 feet). The looping plot on the right is for the case where L = 35.6 feet and M = 33.1 feet. This represents a length error of ± 2.9%. The middle plot is for L = 34.6 feet and M = 33.6 feet. That is a length error of ± 1.4%. Clearly even modest radial length asymmetry can have a dramatic effect on the feed point impedance and resonant frequency. The resonant frequency is the point at which $X_{in} = 0$.

Feed point impedance is not the only problem associated with asymmetric radial lengths. Figure 42 compares radiation patterns between symmetric and asymmetric systems at 7.25 MHz. The amount of pattern distortion varies across the band from a fraction of a dB at 7.0 MHz to 3 dB at 7.25 MHz. Besides the distortion, the gain in all directions is smaller for the asymmetric case. Computing the average gains for the symmetric and asymmetric cases, there is about a 1.6 dB difference. What this tells us is that asymmetric radials can lead to significantly higher ground losses!

Pattern distortion and increased ground loss with asymmetric radials occurs because the radial currents with asymmetric radial lengths are very different from the symmetric case. An example is given in Figure 43.

The graph bars represent the current amplitudes at the base of the vertical and each of the radials immediately adjacent to the base of the vertical. The grey bars are for symmetric radial lengths (L = M = 34.1 feet) and the black bars are for asymmetric radials (L = 35.1 feet and M = 33.1 feet). In the symmetric case, each of the radials has a current of 0.25 A, which sums to 1 A, the excitation current at the base of the vertical. The radial currents are also in phase with the base current.

With asymmetric radials the picture is very different: the current amplitudes are different between radial pair 1 and 2 and pair 3 and 4, and the sum of the current amplitudes is *not* 1 A (the base current amplitude), it is much larger! This would seem to violate Kirchhoff's current law that requires the sum of the currents at a node to be zero. In this

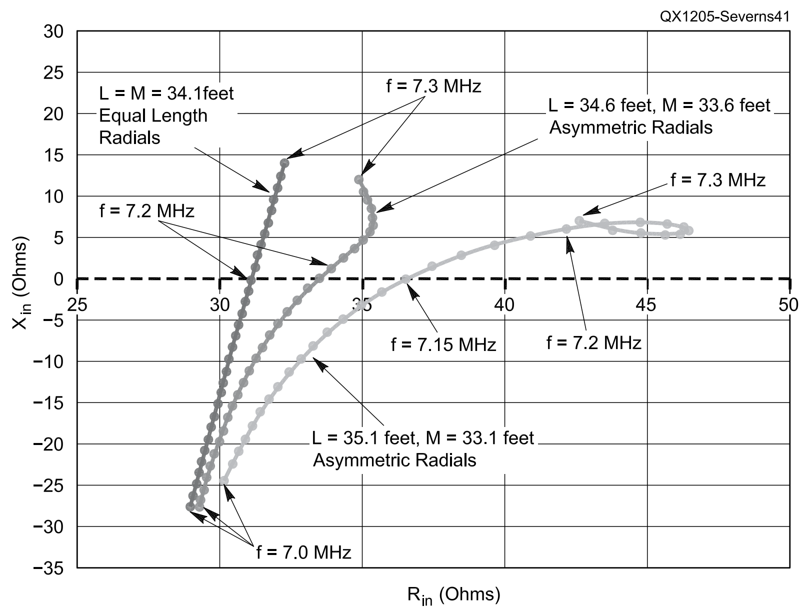

Figure 41 — A comparison of the input impedances ($Z_{in} = R_{in} + j\,X_{in}$) from 7.0 to 7.3 MHz at the feed point of the vertical, for symmetric and asymmetric radial lengths. The frequency is stepped in 10 kHz increments.

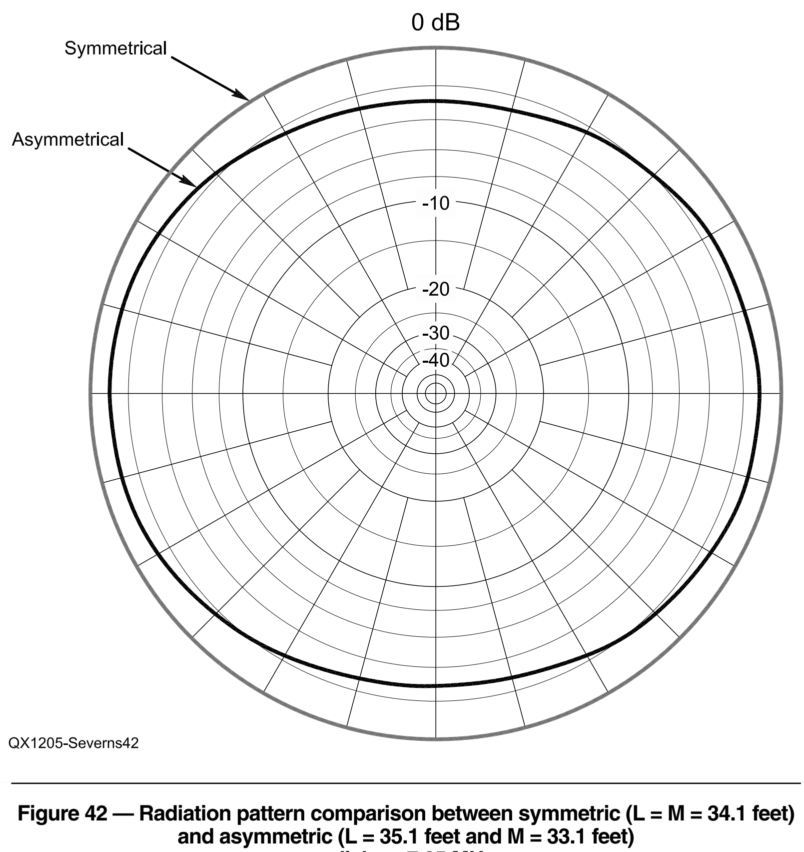

Figure 42 — Radiation pattern comparison between symmetric (L = M = 34.1 feet) and asymmetric (L = 35.1 feet and M = 33.1 feet) radials at 7.25 MHz.

case the radial currents in the two pairs of radials are not in phase with each other or the vertical base current. The current in radials 1 and 2 is shifted by –62° from the base current and the current in radials 3 and 4 is shifted by +89°. The base and radial currents sum *vectorially* to 0 however. That satisfies Kirchhoff's law! These large asymmetric currents go a long way towards explaining the increased ground loss and pattern distortion. Note that the current asymmetry shown in Figure 43 is for f = 7.25 MHz. As the frequency is changed the pattern for the asymmetric currents in Figure 43 will change in a way similar to Weber's data shown in Figure 38.

If we take the example of L = 35.6 feet and M = 33.1 feet and add a wire from the junction of the radials to a ground stake, the G_a drops *another* –0.5 dB and the radial current asymmetry increases.

These examples represent only two particular cases. Obviously there are an infinite variety of radial fan distortions including radial lengths, azimuthal asymmetry, droop of the radials, and on and on. As we increase the number of radials what we see is a rapid decrease in the sensitivity to asymmetric radial lengths. A primary effect of additional elevated radials (>4) is to reduce the sensitivity to radial asymmetry, nearby conductors, variations in ground conductivity or objects under the radial fan, and, as shown in Figure 27, more numerous radials reduce the potentials on the radials.

How can we tell if there is a problem in an existing radial fan? One way is to measure the current amplitudes in the individual radials close to the base of the vertical. (See Part 1 of my series, "Experimental Determination of Ground System Performance on HF Verticals; Test Setup and Instrumentation," in the Jan/Feb 2011 issue of *QEX*.) If the current amplitudes are significantly different between the radials and/or if the sum of the current amplitudes in the radials is greater than the base current, then you have a problem. Current amplitude measurements can be made with an RF ammeter. More accurate measurements that also show the phase can be made using current transformers and an oscilloscope or a vector network analyzer.

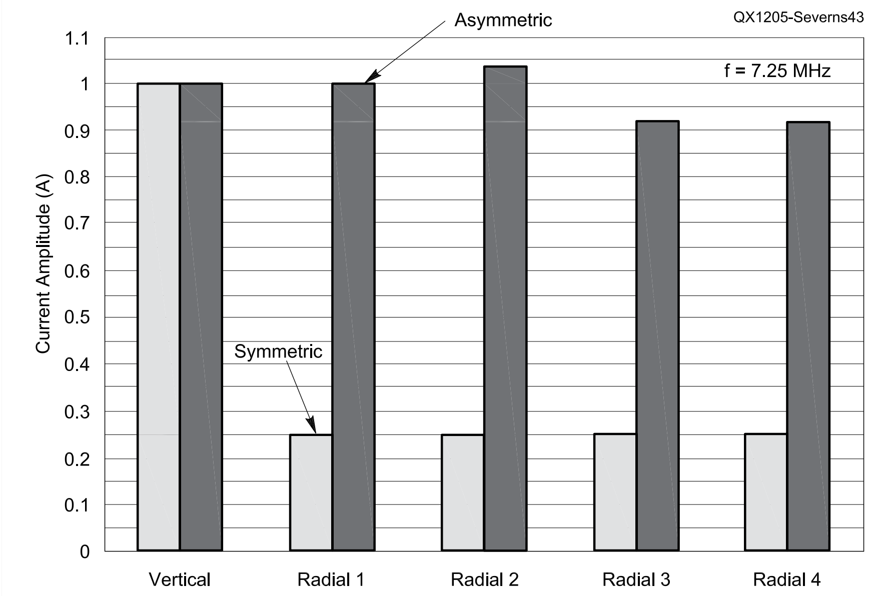

Figure 43 — Comparison of currents between symmetric and asymmetric radials.

Final Comments

This discussion has shown that a vertical with an elevated ground system has many subtleties and many potentially useful variations, but it has also shown that you cannot simply throw up a vertical with a few radials dangling in various directions and expect it to work properly. You have to take some care. Are there a few simple rules that will keep us out of trouble?

Here's my advice:

1) Use at least 10 to 12 radials.

2) Make an effort to have the radial system as symmetric as possible.

3) Keep the radial system as far as possible from other conductive objects.

4) While it is certainly possible to use almost any height for the vertical, I suggest you start with H = $\lambda_0/4$ and trim the radials for resonance. This makes H a little tall, but it shortens your radials (especially if you're using 10 to 12) and raises the feed point impedance a bit.

5) Use a balun or common mode choke on the feed line at the base of the vertical. To be effective, the balun should have a shunt impedance of >2 kΩ.

6) If you have a special problem situation by all means model some trial solutions first. That will save you a lot of time over cut-and-try in the field. If you can't afford NEC4 software, the free NEC2 software will still be very helpful. (See **www.4nec2.com**.)

This article has covered a lot of ground looking in detail at the behavior of verticals with elevated ground systems. Despite the length of this article, it really just scratches the surface of the subject. There are many other topics that deserve attention. For example: a more detailed look at counterpoises, or, in an array, the interaction between the radial systems associated with the individual verticals, the effect of non-level terrain, and so on. I particularly recommend the articles by Al Christman, K3LC, that address many of these issues. (See Notes 18 through 33 in Part 1.) While I hope the work reported here is helpful, there's still lots more to be done before we can claim to really understand this class of antennas.

Acknowledgement

Much of the modeling data for the graphs was derived using *MultiNEC*, an *Excel* program developed by Dan Maguire, AC6LA, that interfaces with NEC modeling programs allowing multiple runs for parameter studies. This is truly a wonderful program. Unfortunately it is currently being revised and not available as of this writing. I also want to thank my reviewers Al Christman, K3LC, George Cutsogeorge, W2VJN, Mark Perrin, N7MQ, Ward Silver, NØAX, Dean Straw, N6BV, and Dick Weber, K5IU. They gave freely of their time going through my drafts, making many important suggestions and corrections.

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Some Ideas for Short 160 Meter Verticals

Few amateurs have room for full sized vertical antennas on 160 meters. Shorter verticals are possible, but you have to be creative.

While it's desirable for a vertical to be a full ¼-wavelength high, on 160 meters that's ≈130 feet and many times that's not possible. For a variety of reasons we may be restricted to much shorter verticals. The late Jerry Sevick, W2FMI, showed us how to build efficient short verticals for 20 and 40 meters using a flat circular top-hat, which is very effective for capacitive loading and practical at 40 meters.[1, 2, 3] But a flat top becomes mechanically difficult on 160 meters, at least for really short verticals where a large diameter is needed. However, capacitive top-loading is still the key to maximizing efficiency in short verticals. This drives us to consider other forms of top-loading. One traditional approach has been the "umbrella" vertical shown in Figure 1. The attraction of this approach is its simplicity: just hook some wires to the top and pull them out at an angle.

Umbrella verticals aren't new, they've been around since the early days of radio and some really excellent experimental work has been done at MF.[4] Large antennas are difficult to work with so there hasn't been a lot of experimental optimization although Belrose, VE2CV, has written about his work with VHF models and at MF.[5, 12] The advent of *NEC* modeling software has made it much easier to explore antenna optimization and this article is mostly a *NEC* modeling study. While *NEC* can be very informative, it's my policy to compare my *NEC* modeling to reliable experimental data whenever possible and I do so near the end of this article.

[1]Notes appear on page 45.

What's a "Short" Antenna?

What's meant by a "short" vertical? In professional literature the definition is usually a vertical shorter than one radian (1 radian = 57.3° = $\lambda/2\pi$ = $0.16\lambda_0$) where λ_0= free space wavelength. Sometimes "short" is defined as a vertical with a physical height $H<\lambda_0/8$ or 45°. At 1.83 MHz $\lambda_0/8 \approx 67$ feet. The focus of this article will be antennas with $H<0.125\lambda_0$.

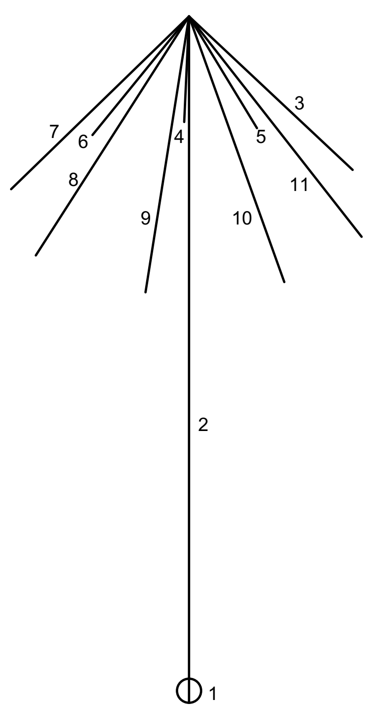

Figure 1 — Example of an umbrella vertical.

Is There a Problem?

Before starting a discussion on capacitive top-loading we need to ask if there is a problem with short verticals that justifies the added complexity of a top hat. After all, we could put up a simple vertical and load it with an inductor as is done for mobile antennas. There is certainly lots of information on optimizing mobile verticals. For a lossless antenna the radiation pattern of a very short vertical is almost the same as a $\lambda/4$ vertical. The differences between short and tall verticals show up when losses are taken into account. We also know that as H is reduced Q rises rapidly and the match bandwidth narrows.

Real antennas have several sources of loss:

- Loading coil resistance — R_L
- Equivalent ground loss resistance — R_g
- Conductor resistance — R_c
- Loss due to leakage across insulators (at the base and at wire ends) — R_i
- Corona loss at wire ends — R_{cor}
- Matching network losses — R_n

In general R_L and R_g are the major losses but in short antennas conductor currents and the potentials across insulators can be much higher than in taller verticals. In fact the shorter the antenna the greater the losses from all causes and a major part of the design effort is directed towards minimizing losses.

The impedance at the feed point is $Z_{in} = R_a - jX_c$, where $R_a = R_r + R_L + R_g + R_c + R_i + R_{cor}$, and X_c is the capacitive reactance. R_r is the radiation resistance which represents the desired power "loss." Note that when modeling lossless examples, $R_a = R_r$.

Figure 2 shows a graph of Z_{in} for an ideal vertical (R_a=R_r) over a range of heights: $0.01\lambda_0 < H < 0.125\lambda_0$. Note how rapidly R_a falls ($\propto H^2$) and X_a rises ($\propto 1/H$).

In most of the following graphs and discussion H is given as a fraction of λ_0. The physical height in feet (H') at 1.83 MHz is given by:

λ_0 = 537.471 feet → H' = 537.471 × H

For example H = 0.05 λ_0 → H' = 26.9 feet and H = 0.125 λ_0 →67.2 feet

In Figure 2 $Q_a = X_c/R_a$. Because R_a falls rapidly as H is reduced and simultaneously X_c increases rapidly, Q_a becomes very large for small values of H. Q_a varies as $1/H^3$!

For H ≤ 0.125, the capacitive reactance dominates Z_{in} which implies that short antennas are basically just small capacitors in series with small resistances, with the equivalent circuit shown in Figure 3.

To tune out the capacitive reactance at the feed point we can add a series inductor as shown in Figure 4 where $X_L = X_c$ and R_L is the loss resistance associated with X_L ($R_L = X_L/Q_L$).

The efficiency (η) for the circuit in Figure 4 can be expressed by:

$$\eta = \frac{power\ radiated}{input\ power} = \frac{R_r}{R_a + R_L} \quad \text{[Eq 1]}$$

Where $R_a = R_r + R_g + R_c + R_i + R_{cor}$. Ignoring for the moment $R_g + R_c + R_i + R_{cor}$, we can graph Equation 1 to show how the efficiency of a short vertical depends on Q_L and H as shown in Figure 5. A Q_L of 200 represents a pretty mediocre inductor. Q_L values of 400 to 600 are practical with a little care. A Q_L = 1000 is possible, but not easy. The efficiencies in Figure 5 are expressed in dB of signal lost due to power absorbed in the inductor. Table 1 shows the correlation between efficiency in percent and dB where η in dB = 10 Log(η(in %)/100).

Table 1
Relationship Between Efficiency in % and dB

Efficiency in %	*Efficiency in dB*
50%	–3 dB
10%	–10 dB
1%	–20 dB
0.1%	–30 dB

For small values of H, the efficiency is pretty depressing. What's even more depressing is that Figure 5 only shows the effect of R_L. When we include other losses the efficiency will be even lower.

Given the practical limitations on Q_L it's clear that short base-loaded verticals can be very inefficient. Mobile antenna work has shown that we can improve the efficiency by moving the inductor from the base up into the vertical itself. While this can help, we can do much better by adding capacitive top loading, which is practical for fixed installations.

Besides efficiency there are other problems. The match bandwidth will be proportional to $1/Q_a$, becoming very narrow as the vertical is shortened. Of course, higher losses provide damping, which increases the bandwidth somewhat, but that's not the direction we want to go. For a given input power, short antennas can have much higher conductor currents and very high voltages at the feed point. For example, if we set H = 0.05 λ_0, $R_r \approx 1\ \Omega$ and $X_c \approx 1500\ \Omega$. If the base inductor Q_L = 400, then X_L = 3.75 Ω. $R_r + R_L$ = 4.75 Ω. For P_{in} = 1500 W the current into the base will be ≈ 18 A_{rms} and the voltage at the feed point (and across the inductor) will be ≈ 27 kV_{rms}! In addition, the inductor will be dissipating ≈1200 W. Clearly, base loaded short verticals have problems. Capacitive top-loading is the way out of this box.

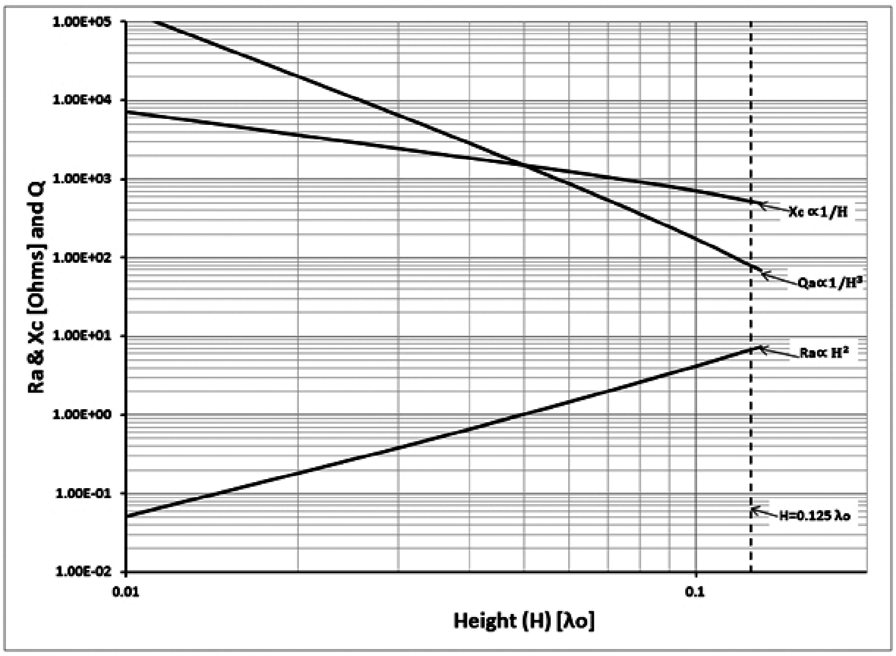

Figure 2 — Feed point impedance at the base of an ideal vertical.

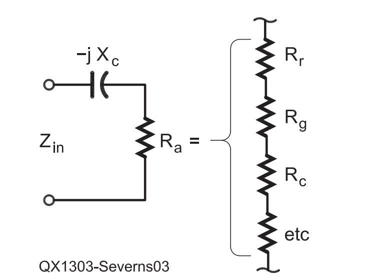

Figure 3 — Equivalent circuit for Z_{in}.

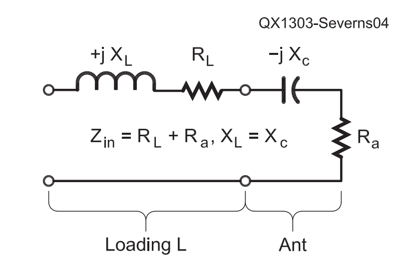

Figure 4 — Equivalent circuit for the input impedance with a series inductor.

Design Variables

There are many variables, all of which can affect performance:

- The height (H)
- The number of umbrella wires (N)
- The length of the umbrella wires (L)
- Whether or not there is a skirt tying the ends of the umbrella wires together
- The apex angle (A) between the top of the vertical and the umbrella wires
- Whether or not a loading coil is used
- The location of the loading coil if one is used
- Q_L of the loading coil
- Conductor sizing and losses in conductors

• Insulator losses
• Matching network design and losses
• Possible corona losses
• Currents and potentials on the antenna
• The characteristics of the ground system and surrounding soil.

There are many variables and we cannot work with all of them at once. What I've elected to do is deal with one or a few at a time, adding loss elements as a better understanding of the antenna develops. The initial models are very idealized, but in the end we'll be including a real ground system, inductor and conductor losses, etc. I've chosen the 8-wire umbrella with a skirt for this discussion because it's relatively simple and it works well, but we should keep in mind that this is only one of many possibilities.[6] An example is shown in Figure 6. The apex angle (A) will be varied from 30° to 90°. The modeling was done at 1.83 MHz. For the moment the ground is assumed perfect and there are no conductor losses.

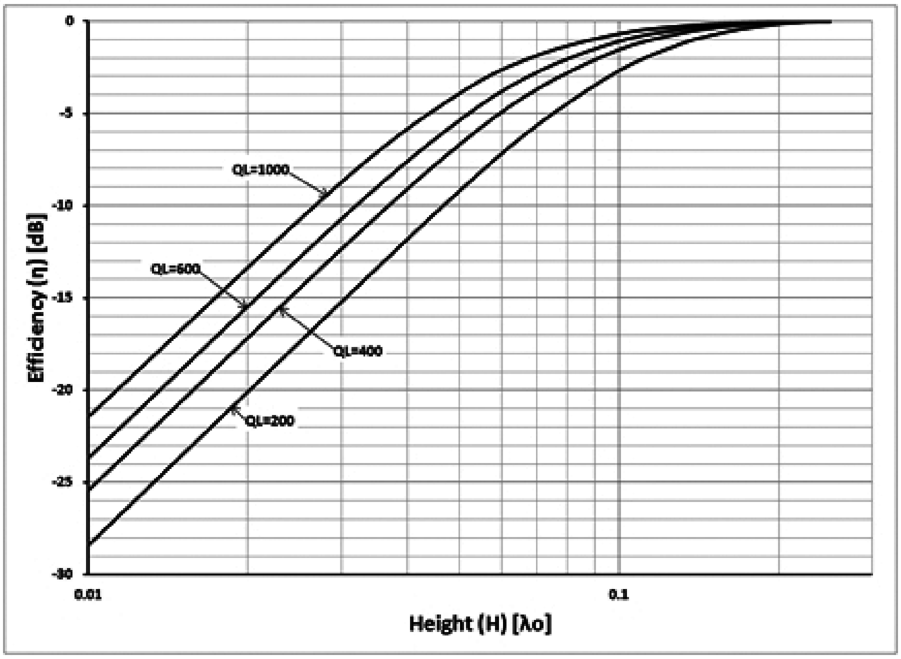

Figure 5 — Variation of efficiency in dB as a function H and Q_L.

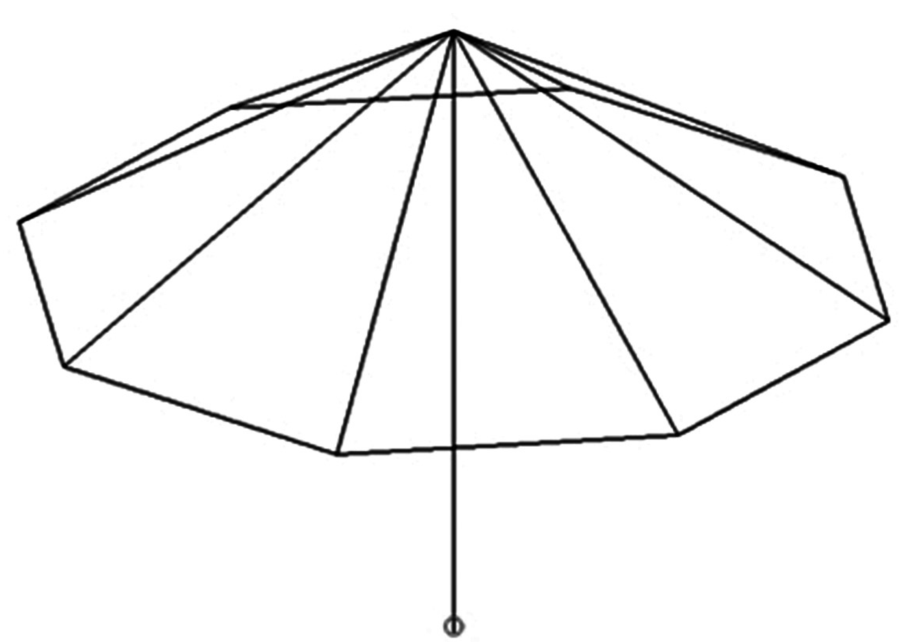

Figure 6 — *NEC* model.

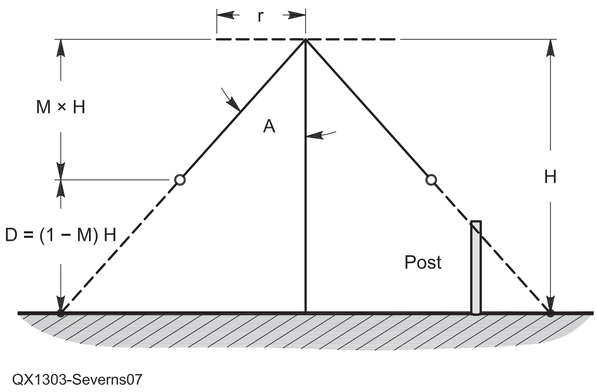

Figure 7 — Model dimensions.

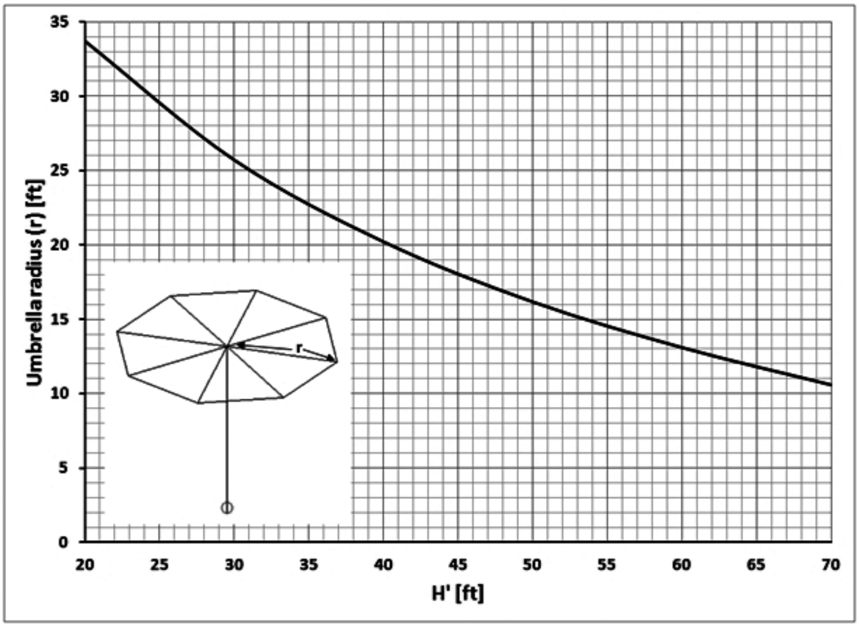

Figure 8 — Radius of the horizontal umbrella needed to resonate the vertical as a function of H'.

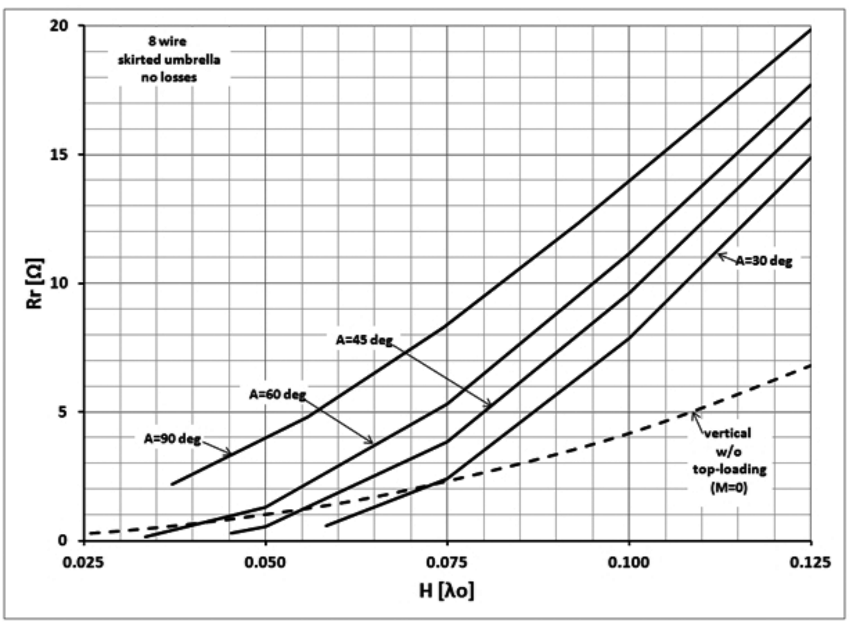

Figure 9 — R_r at resonance as a function of A and H compared to an unloaded vertical.

Figure 7 is a sketch of a top loaded vertical identifying the dimensions. The height of the vertical is H and the vertical dimension of the umbrella is M × H (from the top of the vertical to the bottom of the skirt wires). M is a fraction of H (0<M<1). As we increase M, the bottom of the umbrella moves closer to ground. The distance from the bottom of the umbrella to the ground is D = H(1 – M). Another dimension we may use is the radius (r) from the vertical to the outside of the umbrella skirt. All these dimensions are in λ_0 except M which is a dimensionless ratio. The angle between the umbrella and the vertical at the top is A (in degrees). Initially all the conductors are #12 perfect conductors.

Idealized Top-Loaded Verticals

There are many possible combinations of top and inductor loading we could use, but given the losses associated with loading coils, our first instinct might be to resonate the antenna *without a base inductor*, using only top-loading. This is possible for a wide range of H. We don't want to fool ourselves, however. Even without the need for a resonating inductor, we will very likely need a matching network with an inductor. Top-loading for resonance is not the only option. One widely held idea is that the top-loading should be adjusted to maximize R_r and then an inductor or capacitor should be used to resonate. It's also possible that some other combination may yield the best efficiency. We'll look at these possibilities after we've added a ground system to the model to introduce R_g into the efficiency calculation.

Horizontal Umbrellas

Jerry Sevick used flat or horizontal umbrellas (A=90°) for top loading on 40-meter verticals. This form of top-loading is very effective, but it may not be practical on 160 meters. Figure 8 shows how large the umbrella radius must be to resonate the vertical at 1.83 MHz for 20 feet ≤H'≤70 feet. To give a better feeling for the mechanical dimensions I've shown H and r in feet (H' and r').

For H' = 40 feet, resonance requires an umbrella with r' = 20 feet. An umbrella with r = 10 feet is pretty easy, but going to r = 20 feet or more becomes a mechanical challenge, at least if the umbrella is a free standing "wagon wheel." Mechanically, it's much simpler to just attach the umbrella wires to the top of the vertical and slope them towards ground. But there's a price to pay as shown in Figure 9. For most values of H and A, R_r is higher than its value without top-loading, but for sloping umbrellas R_r is substantially lower than for A = 90°. If it's possible to use a horizontal umbrella by all means do so, but for the rest of this article, we will assume we can't do that and we'll be considering umbrellas with sloping wires.

Umbrellas with Sloping Wires

Figure 9 makes the importance of A clear. For a given M and H, the larger we make A the larger r will be and the greater the top-loading capacitance. This allows us to reach resonance with smaller values of M. However, larger values of A require the umbrella wire anchor points to be farther from the base of the vertical, increasing the ground footprint. One way to reduce the footprint would be to place the umbrella wire anchor points on posts above ground as indicated in Figure 7. In a given installation the value for A is likely to be limited by the available space.

Resonating the vertical using only capacitive loading helps a great deal by eliminating R_L, but we still have the problem of low R_r for small values of H as shown in Figure 9. The dashed line represents R_r for a bare vertical, without top-loading. Over much (but not all!) of the graph we see that top-loading not only resonates the antenna but also increases R_r. That's great but for really short antennas, R_r with capacitive loading can be little better or even *lower* than the simple vertical.

Figure 10 shows the relationship between H and M *for resonance* for skirted umbrellas with 4 and 8 wires, for three apex angles (A).

Whether we can reach resonance depends

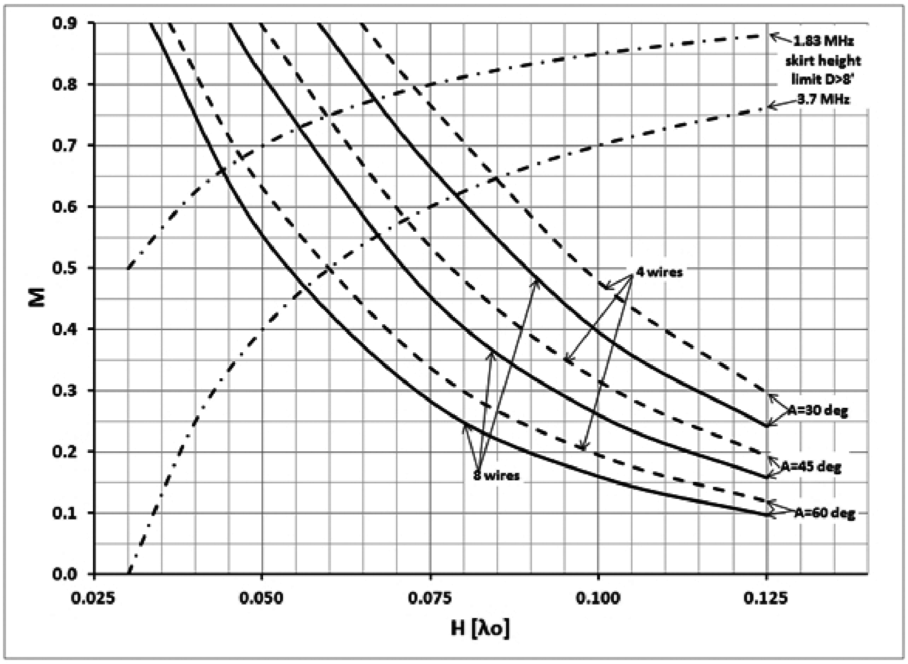

Figure 10 — Values of M for resonance when using 4 or 8 umbrella wires and a skirt.

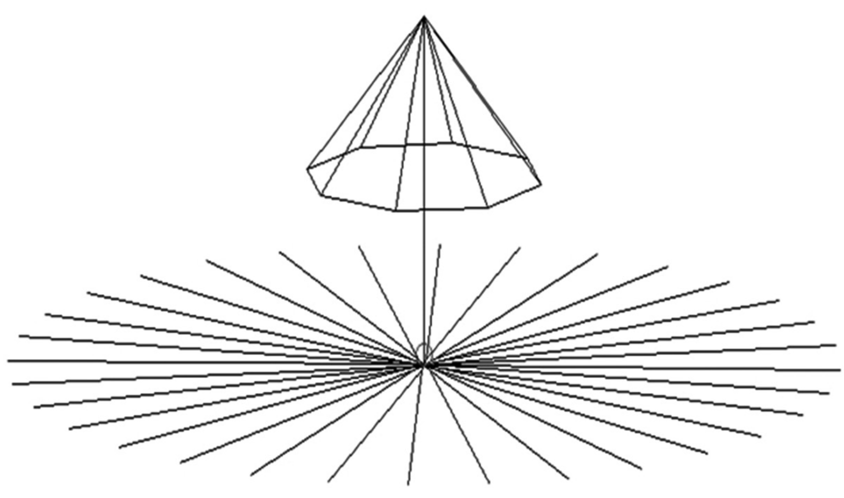

Figure 11 — *NEC* model for a top-loaded vertical with a ground system.

on H, A and the number of umbrella wires, but as Figure 10 shows we can do pretty well for antennas down to H ≈ $0.04\lambda_0$ or a bit shorter on 160 meters if we use a large value for A and more wires in the umbrella. At 1.83 MHz, $0.04\lambda_0$ = 21.5 feet, which is definitely a "short" vertical. Figure 10 shows that increasing the number of wires in the hat increases its effectiveness, but the point of vanishing returns sets in quickly. The improvement gained by doubling the eight wires to 16 wires would be relatively small. The number of umbrella wires becomes a judgment call: is it worth the cost and increased vulnerability to ice loading? The major drawback to wire umbrellas is their vulnerability to ice loading. If you live in an area where ice storms are common you'll have to carefully think through your mechanical design.

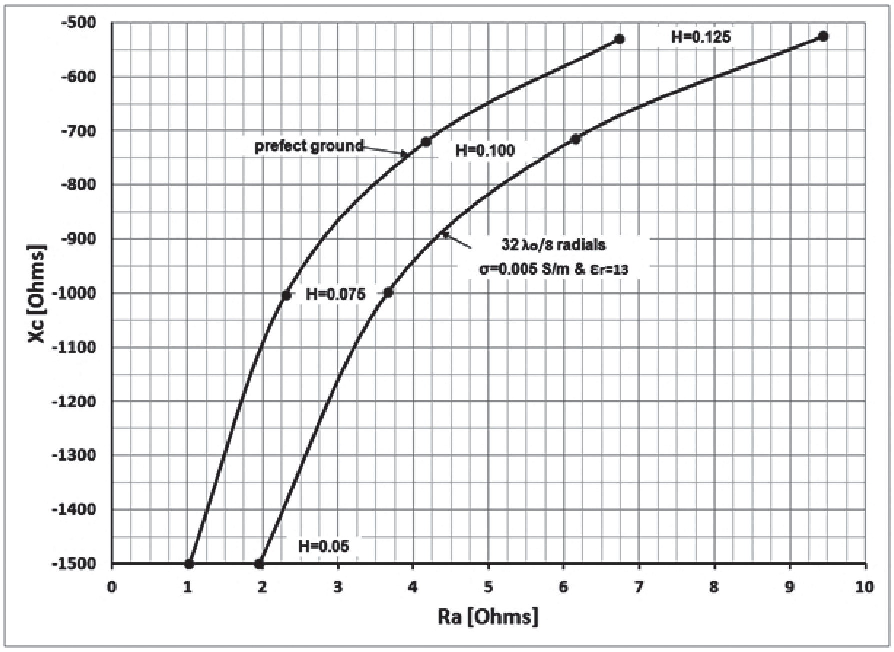

Figure 12 — R_a versus X_c as a function of H with no top-loading, with perfect and real ground systems.

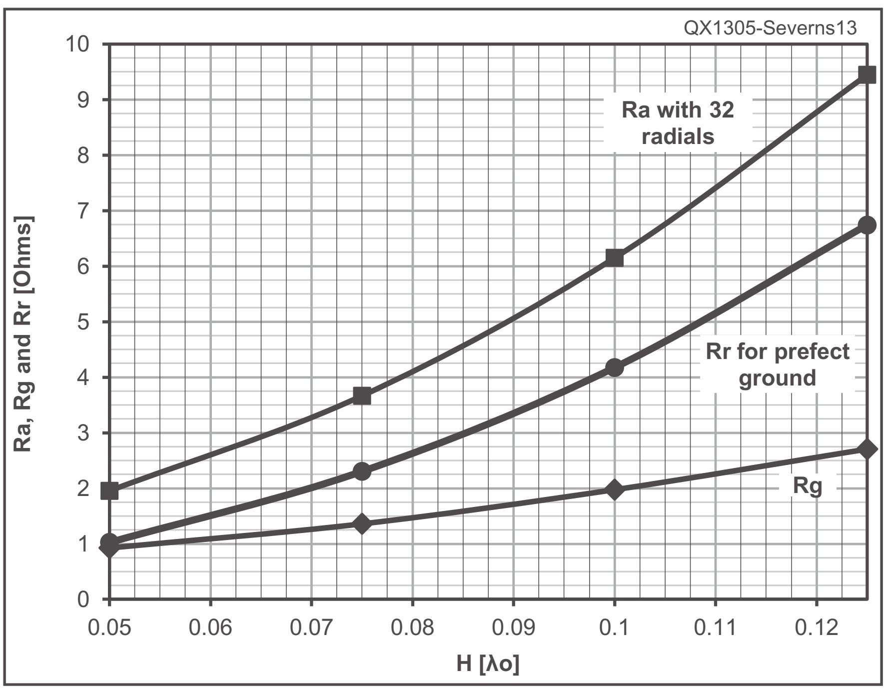

Figure 13 — R_r, R_a and R_g as a function of H *without top-loading*.

There is an important limitation on M, especially for small values of H: *the distance above ground of the lower edge of the umbrella*. Because there can be very high potentials on the skirt you must *keep the skirt out of reach*, at least 8 feet above ground so you can't touch it. This limitation is indicated in Figure 10 by the dash-dot lines. There is one set of limits for 1.83 MHz and a second for 3.7 MHz. You are limited to values of M below these boundary lines.

Non-Ideal Verticals

Now it's time to include losses in addition to R_L.

Affect of Ground System Losses

A model that includes a ground system is shown in Figure 11.

I've chosen to use 32 $\lambda_0/8$ radials (Lr ≈ 65 feet) buried 6 inches in average soil (σ = 0.005 S/m and ε_r = 13). This represents a compromise system; real systems may be larger or smaller depending on the limitations of a given installation. A = 45° is a common apex angle where the radius of the umbrella wire anchor points is about the same as H. To keep the number of graphs in bounds I've set A = 45° for many of the examples.

We need to keep our goal in mind. For a given set of limitations on H, the footprint area of the ground system and the distance to umbrella anchor points on the ground, etc, we want to achieve the maximum possible efficiency. For the moment we'll work with the major losses: R_g and R_L. In this part of the discussion we are *not* going to assume the umbrella loading alone is enough to resonate the antenna. We may use some X_L.

We can start by looking at the effect of real ground on R_a as shown in Figure 12 which compares R_a versus X_c between models with and without the ground system for four values of H. The dots correspond to the values for H at that point.

We can see that R_a increases substantially when a real ground system is used but we also see that X_c is not greatly affected. This indicates that using R_r for the perfect ground as the R_r value with a real ground is a reasonable approximation. This lets us calculate R_g from the model values for R_r and R_a:

$$R_g = R_a - R_r \qquad \text{[Eq 2]}$$

Figure 13 is a graph using Equation 2 to calculate R_g with the ground system shown in Figure 11 but without top loading.

Even though we've kept the ground system and soil characteristics constant as we varied H, R_g is *not* constant. There is a common misconception that at a given frequency, with a given ground system design and soil characteristics, that R_g is some fixed number without regard to the details of the vertical. This is not the case! R_g is not something you measure with an ohmmeter. It is how we account for the ground losses (P_g) associated with a given antenna for a given base current (I_o).

$$P_g = R_g I_o^2 \quad \text{[Eq 3]}$$

P_g is created by E and H-fields which in turn are a function of both the base current and the details of the antenna. As we change the antenna, for a given I_o and ground system, P_g will change and that means R_g will change.

Z_{in} with a Ground System

Figure 14 shows the feed-point impedance ($Z_{in} = R_a + jX_c$) as a function of H and M: where H = 0.05, 0.75, 0.100 and 0.125 and M is varied from 0 (no umbrella, just a bare vertical) to a limit imposed by the minimum allowed ground clearance (8 feet) for the umbrella skirt. The dashed line represents Z_{in} for a bare vertical as H is varied. We can see that the addition of an umbrella drastically changes Z_{in} and Z_{in} is a strong function of both H and M. There are some square markers in Figure 14, which correspond to points of maximum efficiency. We'll discuss these shortly.

Efficiency

In terms of R_r, R_g and R_L, the efficiency will be:

$$\eta = \frac{R_r}{R_r + R_g + R_L} \quad \text{[Eq 4]}$$

We know that $R_L = X_c / Q_L$ and we'll set $Q_L = 400$ which is a reasonable value. The *NEC* model gives us R_r from the ideal antenna and R_a from the antenna with the ground system.

Figures 15, 16 and 17 show how R_r and the loss resistances R_g and R_L vary as a function of M. In Figures 15 and 16 there are markers (the diamonds) for the values of M which correspond to resonance. Note that for H = 0.050 resonance is not reached with the maximum value of M so there is no diamond marker. In Figures 15 and 18 the circles mark the values of M corresponding to maximum R_r. In all these graphs M = 0 corresponds to no umbrella.

In Figure 15 as we enlarge the umbrella (increase M) R_r rises initially but there is a maximum point which depends on H. Increasing M further reduces R_r. This is not surprising given that the currents on the umbrella have a component ≈180° out of phase with the current on the vertical. This results in some cancellation, which increases as M increases. For H = 0.125 and 0.100, R_r maximum and resonance are fairly close

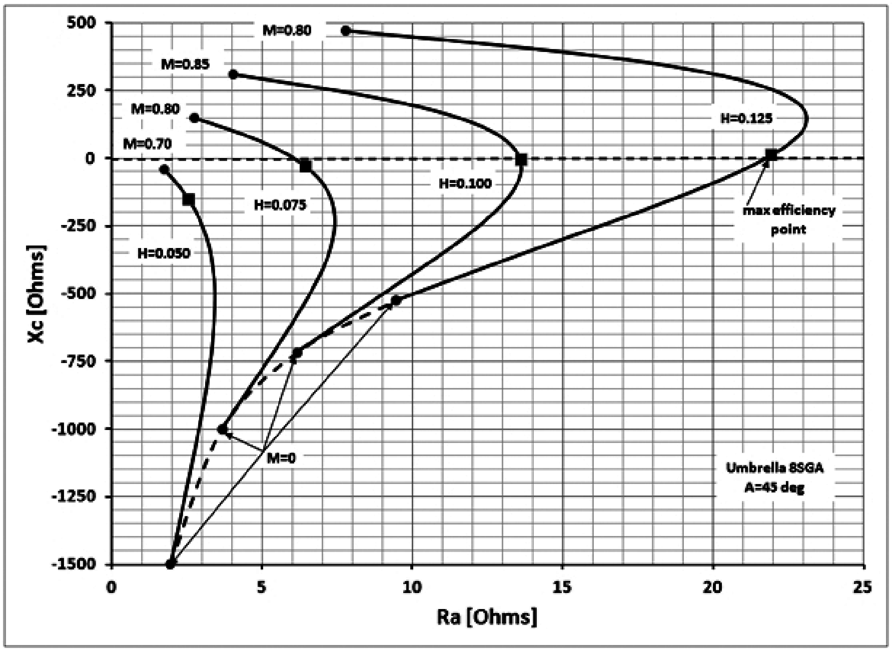

Figure 14 — Feed point impedance as M is increased for H = 0.05, 0.075, 0.1 and 0.125 and A = 45°.

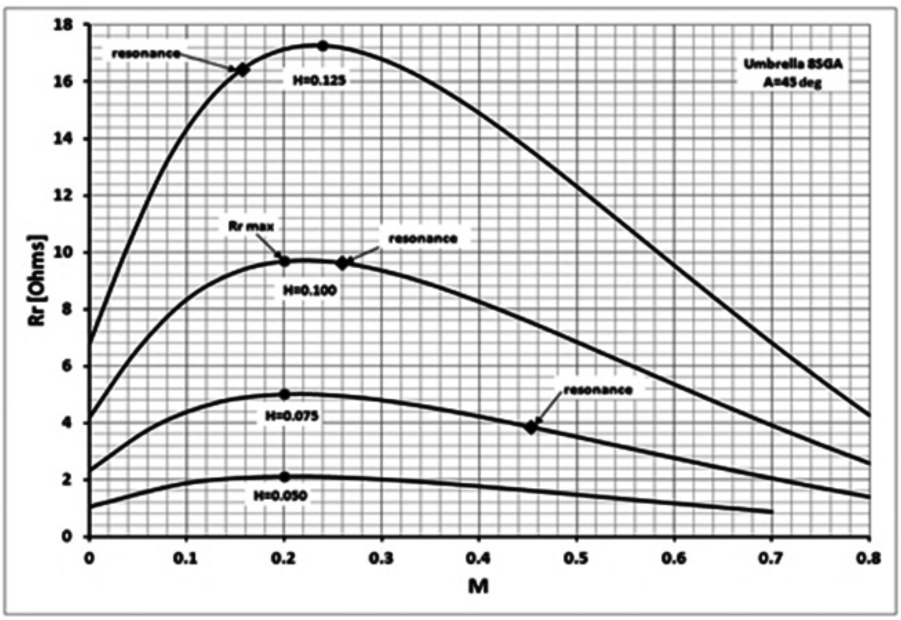

Figure 15 — R_r as a function of M with H as the parameter.

Table 2
L-Network Values and 2:1 SWR Bandwidths

H (λ_o)	R_a (Ω)	X_a (Ω)	X_s (Ω)	R_s (Ω)	X_p (Ω)	*2:1 Bandwidth*
0.050	2.56	–152.5	163.5	0.41	–12.56	15 kHz
0.075	6.46	–30.67	47.44	0.12	–19.26	33 kHz
0.100	13.60	–5.92	28.17	0.07	–30.56	56 kHz
0.125	21.94	11.42	13.39	0.03	–44.21	75 kHz

together, but for shorter antennas the two points are widely separated.

As shown in Figure 16, R_g behaves very much like R_r for smaller values of M; R_g rises but then reaches a peak and begins to fall as M is increased further.

Figure 17 shows R_L decreasing as M is increased and at some point resonance is reached ($X_c = 0$, except for H = 0.050). Above this point we no longer need X_L to resonate (X_c>0) so in Figure 17, $R_L = 0$ above resonance.

All three loss resistances vary with M so it's hard to see simply by inspection where the minimum loss or highest efficiency point is. Better to plug in values for R_r R_g and R_L into Equation 3 and see where the maximum efficiency occurs as shown in Figures 18 and 19.

Figure 18 shows the efficiency in dB where 100% efficiency would be 0 dB. Besides circles for maximum R_r and diamonds for resonance, there are squares to indicate values of M corresponding to maximum efficiency. One important point to notice is that while there are distinct points of maximum efficiency these maximums are very broad. For H = 0.125, resonance and maximum efficiency coincide and for H = 0.100 and 0.075 they're also nearly coincident. The choice for M is not critical but in general the shorter the vertical the larger the optimum value for M. It's also interesting to note that the points of maximum R_r don't coincide with either resonance or maximum efficiency. This brings into question the common assumption that designing for maximum R_r will result in maximum efficiency. That's actually a shame because if maximum R_r is our goal then *NEC2* modeling could easily be used to determine the value. Unfortunately, we need *NEC4*, which is often not available, to determine R_g as it varies with the design of the vertical. However, it is possible to use E and H near-field values from *NEC2* and a spreadsheet to calculate R_g as shown in the *ARRL Antenna Book* (the equations are given in the *Excel* files on the associated CD).[8]

As shown in Figure 19, the apex angle of the umbrella (A) has an effect on the value for M at the maximum efficiency point. The larger A the lower the losses and the smaller (in terms of M) becomes the umbrella. Note that for larger values of A the efficiency peaks are higher but narrower. Making A as large as practical is very helpful for shorter antennas.

Figures 18 and 19 indicate that it's possible to build very short verticals with efficiencies better than 50%. Figures 18 and 19 also bring out another important point. For the examples shown, with the exception of H = 0.125 in Figure 18, resonance occurs for values of M larger than those for maximum efficiency. This implies that it might be better to not load to resonance and use a small loading inductor. However, the differences in efficiency between the maximum and the values at resonance are small in most cases, at least for H> 0.050. From a practical point of view it's simpler top-load to resonance. That value for M can easily be obtained using *NEC2* and some field tuning adjustments. For really short verticals it may pay to do some *NEC4* modeling to see where the maximum efficiency occurs. You could also make field strength measurements with a given input power or use a VNA.[9]

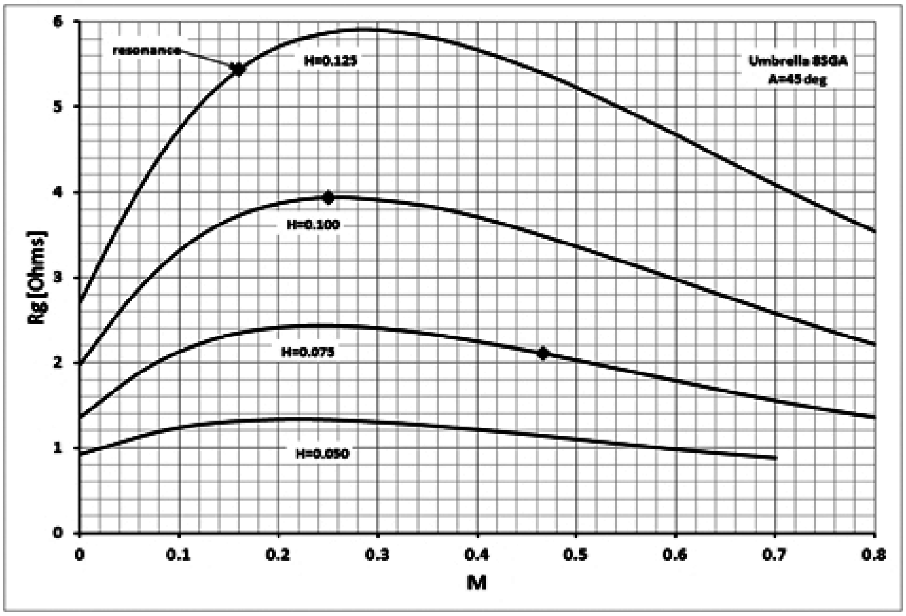

Figure 16 — R_g as a function of M with H as the parameter.

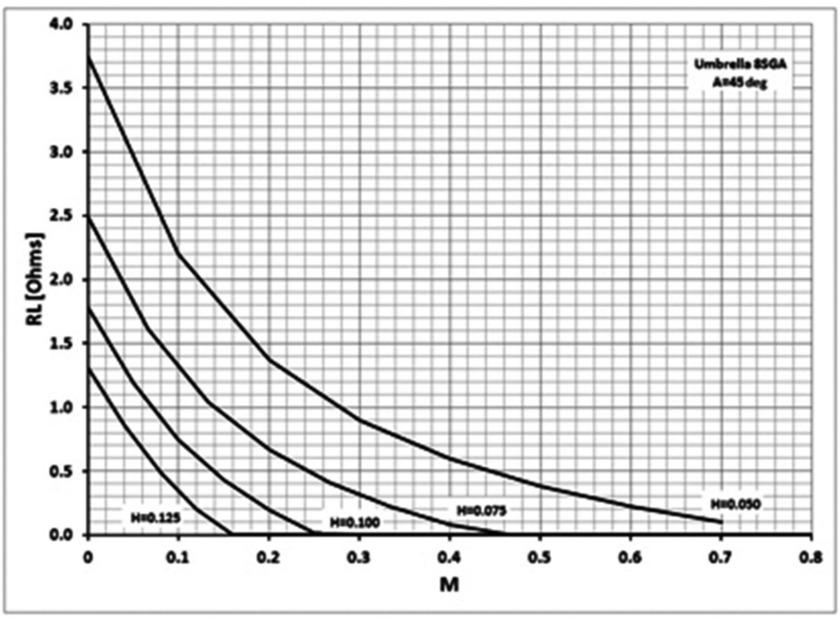

Figure 17 — R_L as a function of M with H as the parameter.

Conductor Losses

It's time to consider conductor losses (R_c). Figure 20 gives examples of how the current at the feed point (I_o), for a given input power (1.5 kW in this example), can vary with H and M. A is fixed at 45° and the squares mark points of maximum efficiency. Figure 20 shows how rapidly I_o increases as H is reduced. Conductor loss varies as I_o^2 so the conductor losses grow rapidly as H reduced. It isn't only that I_o is larger but the current along the entire vertical that increases with more capacitive loading as illustrated in

Figure 21, which shows examples of the current distributions on an H = 0.075 vertical. Note that these current distributions are for I_o = 1 A. As shown in Figure 20, for a given P_{in}, the value for the base current (I_o) will depend on R_a, where

$$I_0 = \sqrt{P_{in} / R_a}$$

As we vary the power level I_o will vary but the ratio I_{top}/I_o, where I_{top} is the current at the top of the vertical, will remain the same as shown.

The current distribution for M = 0.50 has I_{top}/I_o = 0.99, in other words the current is almost constant along the vertical part of the antenna. I_{top}/I_o ratios greater than 0.9 are typical for short antennas top-loaded to near resonance. As shown in Figure 21, the current without top-loading (M=0) falls almost linearly to zero (or close to it) at the top. In the case of mobile antennas the current distribution can be significantly improved by moving the loading inductor up into the vertical, which raises the question if that idea is also useful when heavy top-loading is used. It turns out that when the current distribution is nearly constant the loading coil position has limited effect on the current distribution. From a practical point of view, moving the inductor up into the vertical is a nuisance, but in some cases you may be able to gain some improvement by relocating the inductor if the top-loading is not great enough to be close to resonating the vertical. This may be the case when H < 0.05.

We can get a good measure of conductor loss by turning on the conductor loss option and then calculating the average gain (G_a) with only the conductor losses. Figure 22 illustrates conductor losses for two different conductor sizes for the vertical part of the antenna with 0.05 < H < 0.125. In each case shown the antenna is resonant with only top-loading.

The initial model had #12 wires for the vertical and four umbrella wires with a skirt. As can be seen, the conductor losses at H = 0.05 are very high, ≈–4.5 dB. Most of the loss is in the vertical conductor so increasing its diameter from 0.08 to 0.5 inch cuts the loss almost in half. An even larger diameter conductor along with eight umbrella wires would reduce the conductor loss to less than 1 dB. For example, at 1.83 MHz, 0.05 λ_0 ≈ 27 feet, a 30 foot length of 4-inch aluminum irrigation tubing along with a skirted 8-wire top-hat could have low conductor losses.

The message here is to be very aggressive in conductor sizing. If we are, we can keep conductor losses low even in very short antennas!

Voltage at the Feed Point

Not only is I_o large in short verticals but the voltage at the feed point can also be very high due to the high reactances below resonance (see Figures 12 and 14 for X_c). Figure 23 shows typical values for the feed point voltages for P_{in} = 1.5 kW as M is varied for several values of H.

Note that the vertical scale is in kV_{rms}! Fortunately, for H≤0.075 the highest efficiency point is close to resonance so the feed point voltages are relatively low. However, with H ≤0.05, you can't reach resonance, at least with A = 45° and 8 wires, and the feed point voltage is much higher. One way to improve both efficiency and reduce the feed point voltage would be to increase A to 60°. At 1.83 MHz, 0.050λ_0 ≈ 27 feet so it may be practical to increase A in shorter antennas.

If the power is reduced from 1500 W to 100 W we're still not out of the woods because the voltage varies as the square root of P_{in}. Going from 1500 W down to 100 W reduces the feed point voltage by a factor of

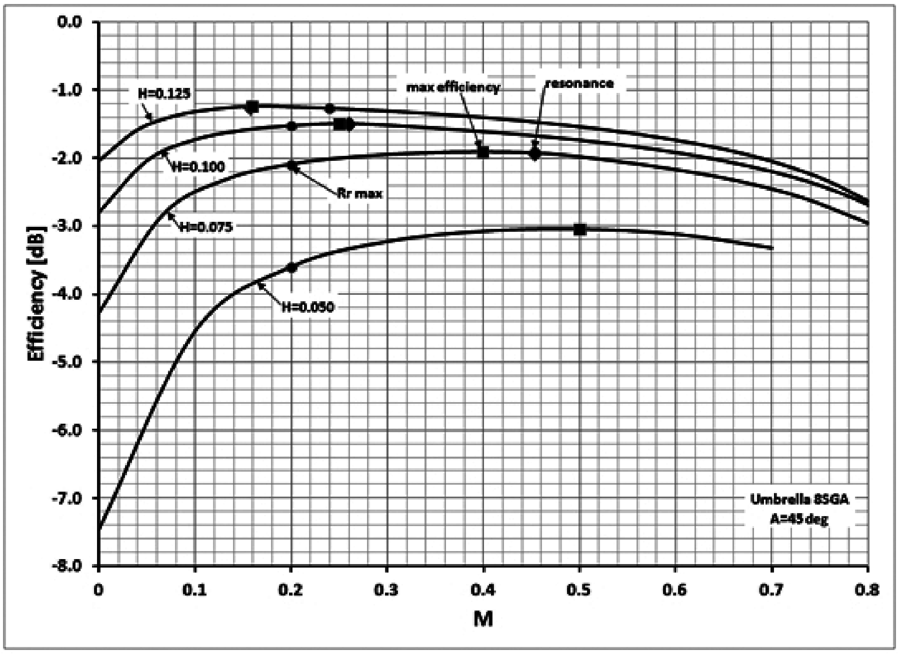

Figure 18 — Efficiency in dB as a function of M with H as the parameter and A = 45°.

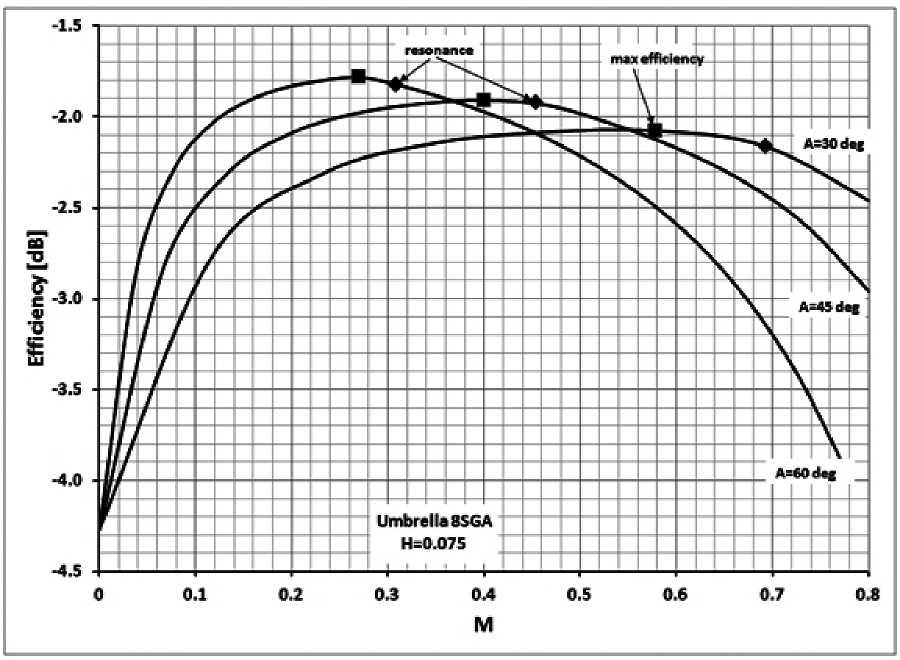

Figure 19 — Efficiency in dB as a function of M with A as the parameter and H = 0.075.

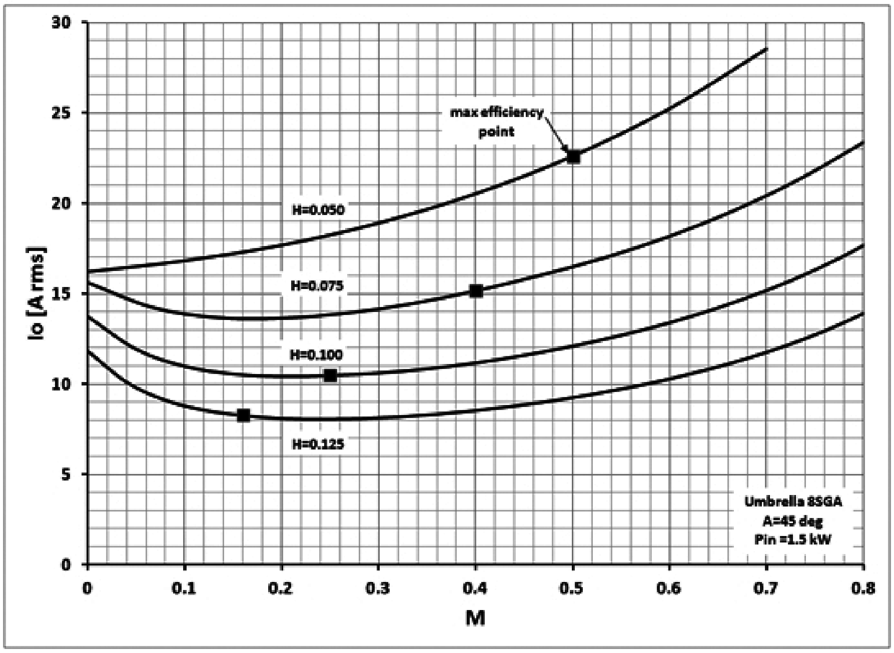

Figure 20 — I_o as a function of M with H as the parameter.

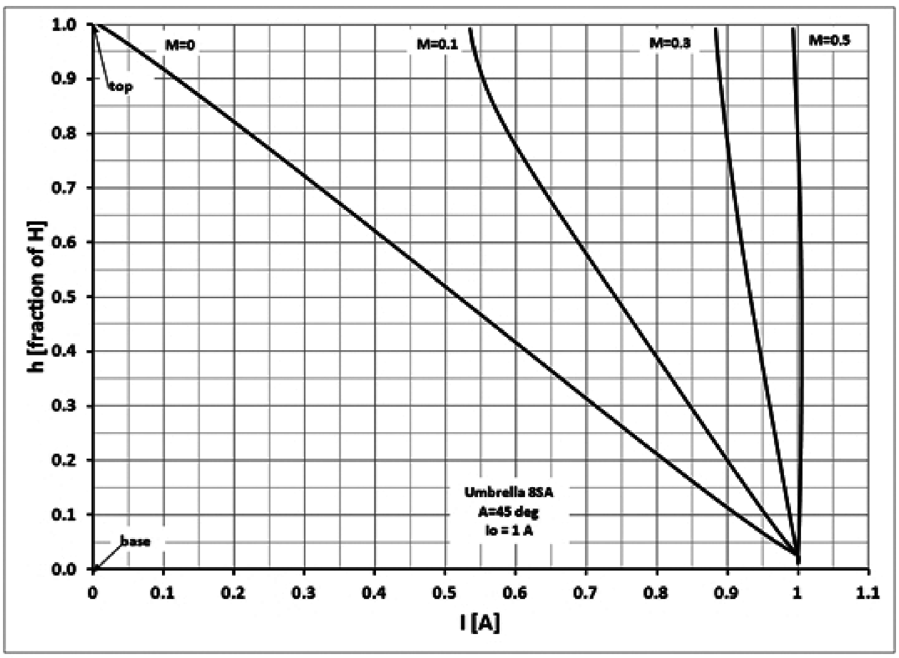

Figure 21 — Examples of the current distribution on a top loaded vertical.

1/3.9 not 1/15! Even at low power levels the voltages can be dangerous. These voltage levels at RF frequencies can introduce significant loss associated with leakage across the base insulator. A plastic bottle base insulator doesn't cut it! Keeping the insulator surface clean and dry is also important. Some form of plastic shield can help to keep achieve this. The use of equipotential rings can also help.

Besides the base insulator these voltages will appear across the base loading inductor if one is present and/or the output of the matching network. There is also the problem of dealing with the power dissipation in the loading inductor. In addition there will be very high potentials on the lower part of the umbrella. These potentials are lower with skirted umbrellas and such umbrellas are usually further above ground, but you still have to consider corona losses. Any sharp points where the umbrella and skirt wires are joined or where insulators are connected can result in substantial losses due to corona, especially if you live at higher altitudes such as Denver, Colorado. You should use high grade insulators on the support lines spreading the umbrella even if they are non-conducting.

SWR Bandwidth

The final step is to match the feed point impedance to 50 Ω. This can be done in many ways but for this discussion I assume the use of a simple L-network matching the feed-point impedance at the highest efficiency point.[10] Assuming $A = 45°$ and $f = 1.83$ MHz, Table 2 summarizes the L-network components and the 2:1 SWR bandwidth for each antenna. X_s is the series matching reactance, R_s is the loss resistance associated with X_s and X_p is the shunt reactance. In this example all the X_s are inductors with $Q_L = 400$ and the X_p are capacitors. The ground system in Figure 11 is included. Note that R_s (due to the loss in the matching inductor) has only a small effect on efficiency except for smaller values of H.

Table 2 illustrates the sharp reduction in match bandwidth associated with shorter verticals. For a given H, one way to improve bandwidth without reducing efficiency is to make A larger. Making the diameter of the vertical conductor larger will also help especially if you can go to a wire cage several feet in diameter! There's a big bag of tricks along those lines that deserve discussion but this article is already too long.[11, 12, 13]

Experimental Verification

As mentioned in the introduction, *NEC* modeling is a powerful tool, but it's not perfect. Whenever possible I like to compare my results with high quality experimental work. Fortunately, such work is available for this discussion. In October 1947 Smith and Johnson published an IRE paper on the "Performance of Short Antennas" which presented their experimental work at MF on a 300 foot tower with eight sloping umbrella wires and a loading inductor at the base. (See Note 4.) This paper is a beautiful example of first class experimental work. Measurements were made at several frequencies from 120 to 350 kHz with the umbrella wire lengths varied in steps from 100 feet to 450 feet. Figure 24 is a sketch of the tower and umbrella arrangements. The angle between the tower and the umbrella wires was ≈ 48°. H = 300 feet represents $0.037\lambda_0$ at 120 kHz and $0.107\lambda_0$ at 350 kHz so despite the large physical size, this is still a "short" vertical.

The ground system had five hundred 75-foot radials and 250 400-foot radials. The 400-foot radial wires extend a short distance past the outer edge of the umbrella when its wires are at maximum length. At 120 kHz, 75 feet = $0.009\lambda_0$ and 400 feet = $0.03\lambda_0$. At 350 kHz, 75 feet = $0.027\lambda_0$ and 400 feet =

$0.14\lambda_0$. Compared to standard broadcast practice ($0.4\lambda_0$ radials) this is a very abbreviated ground system. A small ground system is just what we might expect with a short amateur vertical. The 500 75-foot radials are in effect a ground screen close to the base of the vertical where the E-fields can be very intense.

Part of the experiment was a measurement of field strength at one mile with 1 kW of excitation. This was done at several frequencies with a range of umbrella wire lengths and loading coil Qs. An example of the results is given in Figure 25 for a loading coil $Q_L = 200$.

Changing frequency with a fixed H is equivalent to changing H at a fixed frequency. Figure 25 sends a clear message: the taller the better! H is a dominate factor in achievable efficiency. There are two sets of data on the graph: the first is the solid line for the case of no skirt wire around the outer perimeter of the umbrella and the second (the dashed line) is for the case where a skirt wire connects the outer ends of the umbrella wires. The point of maximum signal can be viewed as the optimum length for the umbrella wires. The relative field intensity can be used as a surrogate for efficiency. The higher the field intensity, at a given distance, for a given input power, the higher the efficiency.

Note the correspondence between the experimental work in Figure 25 and the *NEC* results in Figure 18. Both figures tell the same story!

Using a skirt provides more capacitive loading for a given length of umbrella so we see the peak move to the left, toward shorter umbrella wires. In both cases the peak is quite broad especially for the un-skirted umbrella.

It is also interesting how the peak field point moves towards longer umbrella wires at lower frequencies (corresponding to smaller H in λ_o) and the peak field also declines indicating lower efficiency. No surprise really, the antenna is electrically smaller at the lower frequencies and less efficient. The shift of the peak towards longer umbrella wires is a reflection of increased loss (lower efficiency). Again, this agrees well with the *NEC* modeling.

I strongly recommend reading the Smith and Johnson paper as well as Belrose and Sevick. See the detailed reference information in the Notes.

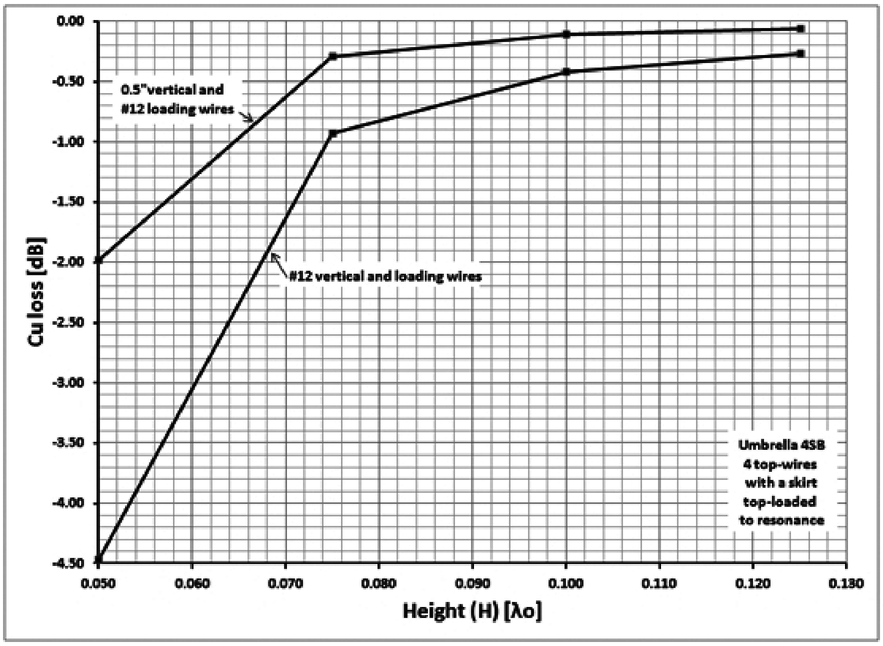

Figure 22 — Examples of conductor loss in short antennas.

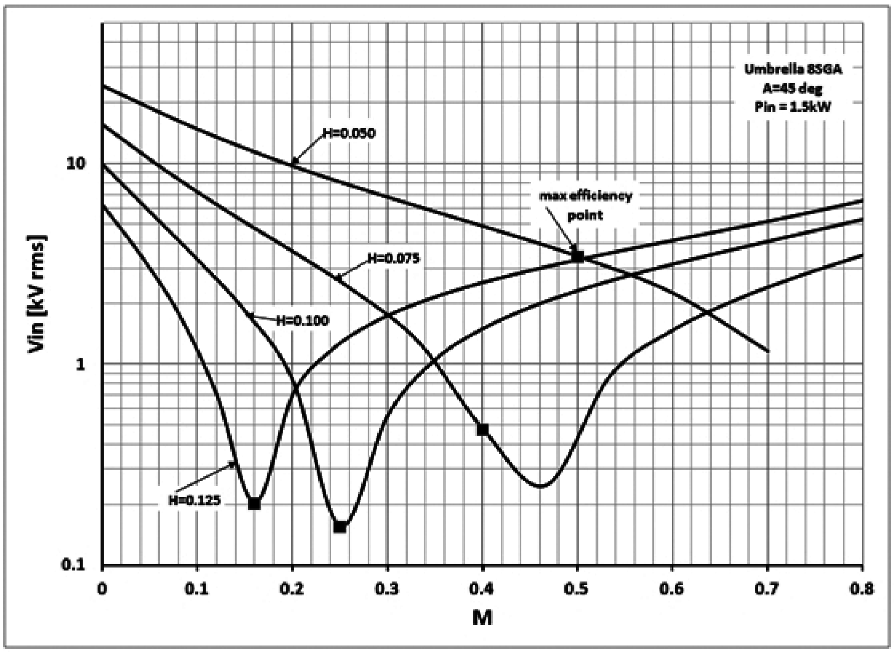

Figure 23 — Feed point voltage as a function of M with H = 0.050, 0.075, 0.100 and 0.125.

Summary

From both modeling and experimental work we can draw some general conclusions:

1. Make the vertical a tall as possible.

2. Make the ground system as large and dense as practical.

3. Make the apex angle (A) as large as practical.

4. Use at least eight wires and a skirt in the umbrella.

5. Be very aggressive in conductor sizing especially for the center conductor.

6. Use high-Q inductors for loading/matching networks.

7. Use high quality insulators both at the base and for the umbrella.

If you do these things then it is possible to have reasonable efficiencies even in very short antennas. Despite the length of this discussion there's far more that could be said and many more ideas for improving short antennas are out there.

Acknowledgements

The work in this paper was prompted by some questions on a short 160-meter antenna from Paul Kisiak, N2PK. Because I hadn't dug deeply into this subject I couldn't help him much beyond some general comments on ground systems. I want also to thank my reviewer Mark Perrin, N7MQ.

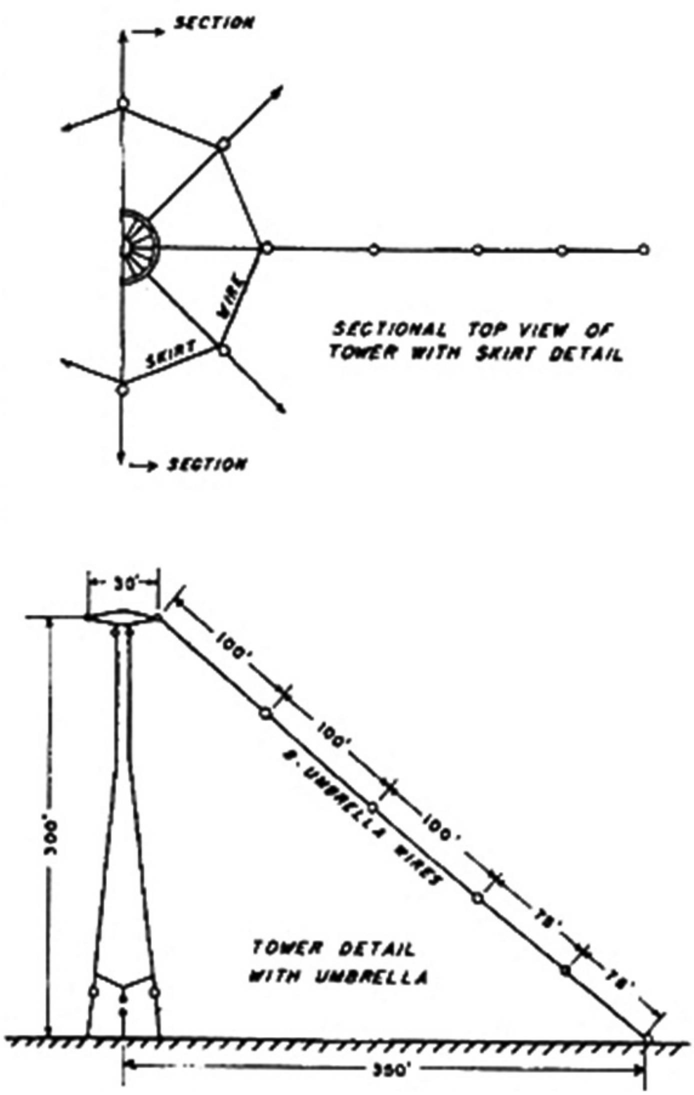

Figure 24 — Sketch of the experimental antenna from Smith and Johnson.[4]

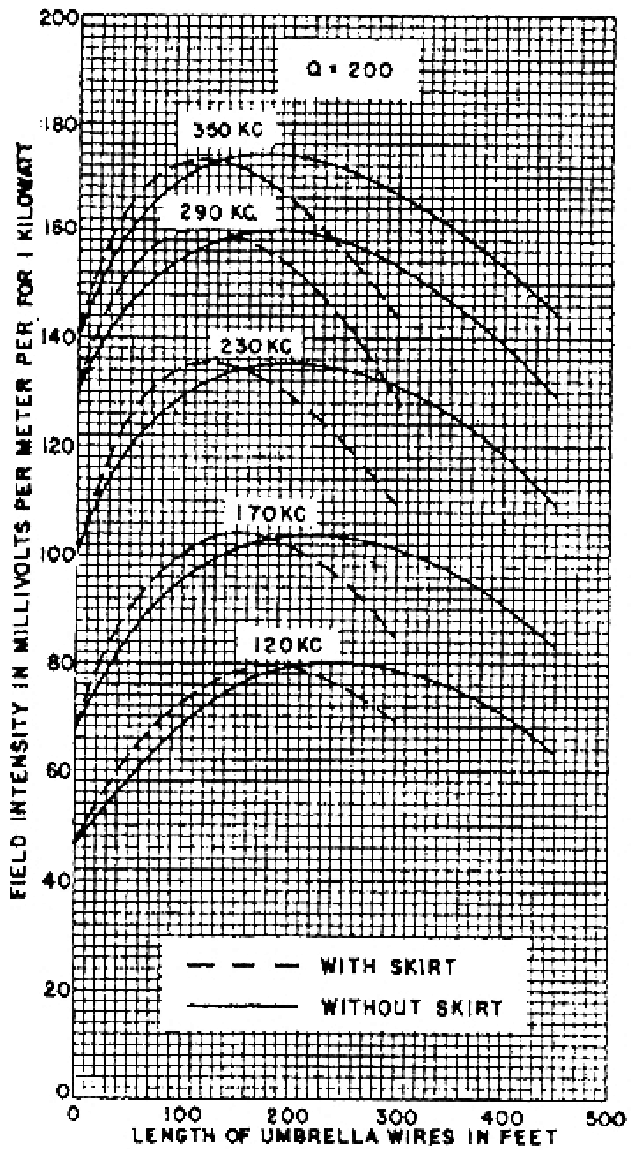

Figure 25 — An example from Smith and Johnson.[4]

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Notes

[1]Jerry Sevick, W2FMI, "The W2FMI Ground-Mounted Short Vertical," *QST*, March 1973, pp. 13-18 and 41.

[2]Jerry Sevick, W2FMI, "Short Ground-Radial Systems for Short Verticals," *QST*, April 1978, pp. 30-33.

[3]Jerry Sevick, W2FMI, "The Short Vertical Antenna and Ground Radial," *CQ Communications*, 2003.

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[5]John Belrose, VE2CV, "Folded Umbrella Top Loaded Vertical Antenna," *Ham Radio Magazine*, September 1982, pp. 12-17.

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[7]Howard Shepherd, W6US, "A High-Efficiency Top-Loaded Vertical," *Ham Radio Magazine*, October 1984, pp. 65-68.

[8]*The ARRL Antenna Book*, 22nd edition, 2011. See the discussion in Chapter 3 and the Chapter 3 material on the accompanying CD including the *Excel* graphs with the equations already loaded.

[9]Severns, Rudy, N6LF, "Experimental Determination of Ground System Performance for HF Verticals, Part 1, Test Setup and Instrumentation", *QEX*, January/February 2009.

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QEX

Conductors for HF Antennas

Putting up an antenna for the low bands? What kind of wire will you use? This analysis may change your plans.

By Rudy Severns, N6LF

Most of us give little thought to the wire from which we fabricate antennas. Most of the time that's okay, but some antennas are quite sensitive to conductor loss. Then we need to think carefully about our choice of wire or other conductor. Recently, I have been building 160-meter wire arrays using hundreds of feet of wire in each. Some of the spans are over 600 feet, and they are attached to poles and trees that move in the wind. For this reason, I initially used #12 stranded Copperweld with PVC insulation. One of the antennas is a two-element, end-fire array—essentially a vertically polarized W8JK. It is a problem with any end-fire array that to obtain gain, the radiation resistance must be lowered by closely spacing the elements. In the case of a W8JK array, the impedance is in the range of 8 to 20Ω. As Krause pointed out in Reference 1, this makes the obtainable gain very sensitive to conductor resistance. The problem is particularly severe on 160 meters because the wire used is very long (over 700 feet in my array) and tubing is impractical.

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The performance of the W8JK array was good, but I had a feeling that I could get much more from the antenna. This led me on a hunt to identify possible losses: to measure wire resistance, to analyze expected conductor losses, to finite-element model solid-copper and Copperweld (copper-clad steel) wire and to model the effects of wire losses on antenna performance. The results are interesting and give insight into appropriate conductor selection. It turns out my intuition was right, the conductor loss was high. The wire resistance was double the expected value, but the reason for that was a surprise.

Conductors

Many types of wire, conductive strips and tubes can be and are used for antennas. The reference against which other wires are judged is solid #12 AWG, soft-drawn, bare copper. Other common choices are:

- seven-strand, hard-drawn copper
- solid #12 AWG Copperweld
- 19-strand Copperweld (#12 AWG)
- aluminum electric-fence wire, in various sizes

- Alumoweld (aluminum-clad steel, see Reference 2)
- #8 AWG aluminum clothesline
- aluminum tubing
- thin copper or aluminum strips
- stainless steel tubing
- towers and galvanized steel guy wires

Occasionally galvanized steel fence wire, stainless steel or copper plated steel electric fence wire is suggested for antennas. These are *very poor* choices, as I will show shortly. Table 1 lists the resistivity and conductivity for some common conductors. The values for steel are only approximate because they vary greatly with the exact composition and processing history.

Sometimes silver plating is suggested for conductors. The conductivity of silver is only 6% better than copper, but when the surface oxidizes, silver oxide is a much better conductor than copper oxide. We will not be considering silver conductors for the rest of this article, however.

Skin Effect

The resistance of wire at a given frequency depends on three things: size, electrical properties of the material (including surface corrosion!) and the

Table 1—Conductivity and Resistivity of Conductors

Material	*Conductivity (σ)* siemens/meter	*Resistivity (ρ)* ohm-cm
Silver	6.2×10^{7}	1.62×10^{-7}
Copper (annealed)	5.8×10^{7}	1.7241×10^{-6}
Aluminum (99.9%)	3.81×10^{7}	2.62×10^{-6}
Iron	1.03×10^{7}	9.71×10^{-6}
Low-carbon steel (AISI 1040)	0.5×10^{7}	20×10^{-6}
Stainless steel (AISI 304)	0.11×10^{7}	90×10^{-6}

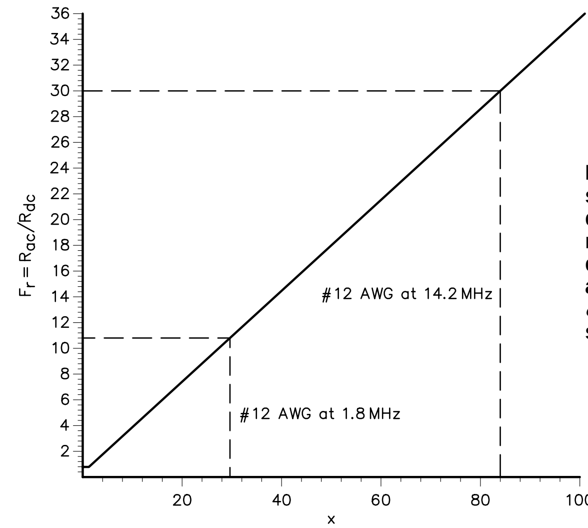
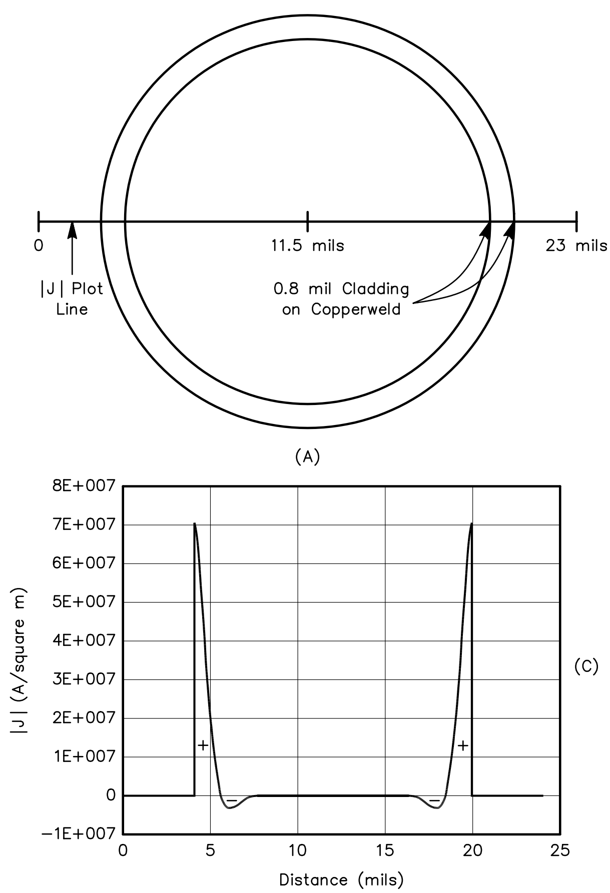


Fig 1—R_{ac}/R_{dc} ratio for solid round wire. Wire diameter (X) is normalized to the skin depth, , where d is the actual wire diameter and δ is the skin depth in the same units.



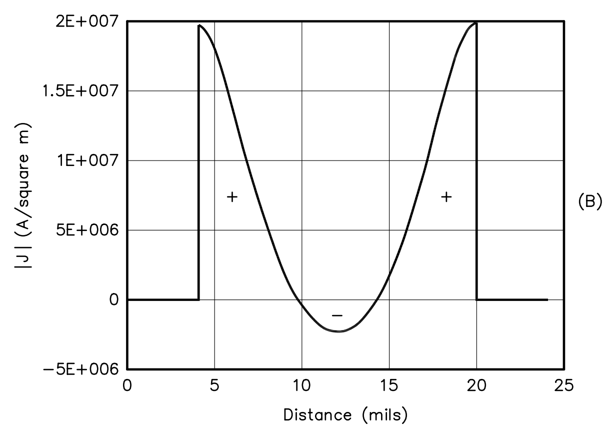


Fig 2—Current density (J) in a solid copper #26 AWG wire (cross section A) at 1 (B) and 16 (C) MHz.

resistance increase due to skin effect. Skin effect is the tendency for current to crowd to the outer perimeter of a conductor as frequency is increased. It is characterized by the depth at which the current density (J) has fallen to about 0.37 ($1/e$, where e=2.718). For good conductors, the skin depth (d) is expressed by:

$$\delta = \sqrt{\frac{1}{\pi\sigma\mu f}} \text{ meter} \qquad \text{(Eq 1)}$$

where:

δ = skin depth (meters)

μ = permeability = $\mu_r\ \mu_o$; μ_o= 4×10^{-7} H/m; μ_r = relative permeability

σ = conductivity in siemens/m (mho/m)

f = frequency (hertz)

For copper at room temperature:

$$\delta = \frac{2.602}{\sqrt{f_{MHz}}} \text{ mils} \qquad \text{(Eq 2)}$$

For f = 1.8 MHz, δ =1.94 mils. For f = 14.2 MHz, δ = 0.69 mils. The Appendix contains a graph of the relation between skin depth and frequency for copper at 20° and 100°C.

For round wire, the variation of R_{ac}/R_{dc} (F_r, or resistance factor) with normalized wire diameter $X = d/\delta\sqrt{2}$ is shown in Fig 1. The variable d is the wire diameter, in the same units as δ. The equation from which the graph is derived is given in the Appendix. For #12 AWG copper wire at 1.8 MHz, X = 29.5 and F_r = 10.8. For the same wire at 14.2 MHz, X = 83 and F_r = 30. This thirty-fold resistance increase at 20 meters is due to skin effect! It cannot be ignored on any amateur band.

I am fortunate to have access to finite-element modeling (FEM) CAD software that can directly calculate and graph current distribution and power loss in conductors such as solid copper wire or Copperweld, which is made of two different materials. The graphs in Figs 2 through 5 were generated using FEM software (see Reference 2).

Figs 2B and 2C give plots of the current density (J in A/m^2) along the line shown in Fig 2A, for solid #26 AWG copper wire (δ = 15.9 mils) at 1 and 16 MHz. The crowding of current to the outside perimeter of the wire and how crowding worsens as frequency increases is clearly shown. This is why the apparent resistance of the wire increases so much. At some points within the wire, the instantaneous current is actually flowing backwards (minus signs) due to the self-induced eddy currents that are the underlying phenomena responsible for skin effect. These currents must be balanced by more forward current (+) to keep the average current unchanged. That is, the same number of carriers must come out one end of the wire that you put in the other end. The net result is increased power dissipation for a given RMS current.

In Copperweld wire, the copper cladding on the outside of the wire is typically about 10% of the wire radius. For #26 AWG wire, the cladding thickness would be about 0.8 mils (0.0008 inches). Fig 3 graphs J for #26 AWG Copperweld. It is clear that the current is flowing only in the copper cladding; there is almost no current in the steel core. This is predicable from the skin-depth equation; δ is inversely proportional to the square root of the permeability. For steel, μ_r is highly variable, affected by the composition of the steel, the processing and even

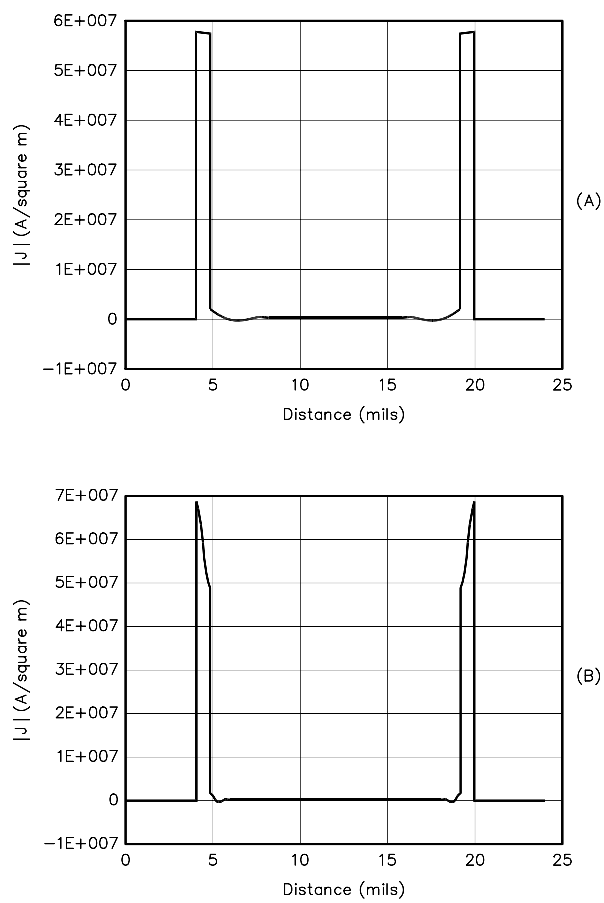


Fig 3—Current density in a #26 AWG Copperweld wire with 0.8-mil cladding at 1 (A) and 16 (B) MHz.

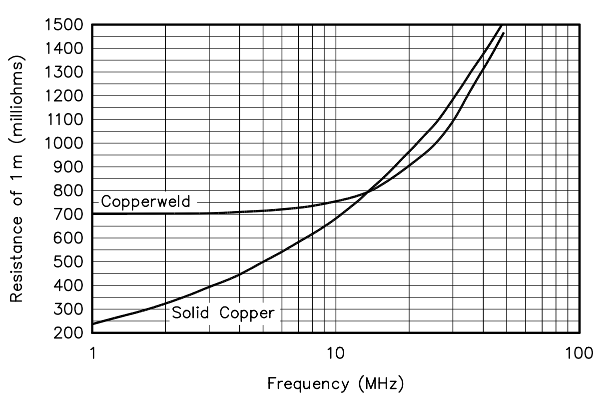


Fig 4—Resistance comparison of 1-meter lengths of #26 AWG solid copper and #26 AWG Copperweld with 0.8-mil cladding from 1 to 30 MHz. Derived from FEM modeling.

the current level. Losses can actually increase as the current increases because μ_r increases with flux density (B), reducing the skin depth and increasing R_{ac}. Thus, μ_r can be from 1000 to 10,000 or more, which means that the skin depth at 1 MHz and above is very small. Copperweld behaves very much like a tubular conductor. This can allow the conductor loss to actually be less or greater than a solid conductor of the same outside diameter, depending on the wall thickness and frequency.

A graph of R_{ac} for 1-meter lengths of #26 AWG solid copper and Copperweld (0.8-mil cladding) wires is given in Fig 4. Below about 14 MHz, the solid copper wire has less resistance. In fact at 2 MHz (160 meters), the Copperweld has more than twice the resistance of solid copper wire. This is simply because current in the Copperweld is crowded into a thin layer. The tube is too thin! Above 14 MHz, however, the tube has less resistance and the Copperweld is superior. Notice also that at low frequencies, the resistance of the Copperweld is nearly constant.

This can be explained from Fig 3, which shows that at low frequencies the current density is basically uniform and changing frequency doesn't change J much. As you reach the middle range of frequencies, current distribution in the tube is better than that in the solid wire and the loss is less. At some high frequency, current distribution in the tube will equal that in the solid wire (the core no longer matters) and its resistance will be the same. In Fig 4, the resistances begin to converge above 50 MHz. The resistances shown in Fig 5 for #12 AWG wires clearly illustrate the convergence at high frequencies. Thus, there is a region, depending on the cladding thickness, where Copperweld is superior to solid wire, but below this region, it is inferior!

Fig 5 is a graph of R_{ac} for 1-meter lengths of four different #12 AWG wires: solid copper, Copperweld with 4-mil and 2-mil cladding and an approximation for 19 strands of #26 Copperweld with 0.8-mil cladding. Again, we see the excess resistance for the 0.8-mil Copperweld at 160 meters, but now the crossover frequency with solid copper is just above 7 MHz. For 30 through 10 m, the stranded Copperweld is somewhat better (5-10%) than solid copper. Copperweld with 4-mil cladding (which is standard for solid #12 AWG Copperweld) is slightly better (≈5%) than solid wire on 160 meters and equal at higher frequencies. While the electrical properties are good and the wire is very strong and durable, the stiffness of Copperweld and its strong desire to remain coiled make it the devil's own invention to work with. Wear gloves and eye protection when working with it!

For 40 meters and up, stranded Copperweld is a good choice: It has low resistance, good strength and is reasonable to work with. For 80 and 160 meters however, the resistance is quite a bit higher and may be a problem for some antennas. Solid copper or Copperweld would be a better choice. In the case of iron fence wire, stainless steel wire or copper-plated steel electric-fence wire, the skin depth will be very small and the ac resistance very large. The copper plating on electric-fence wire is simply too thin to be of any help at HF.

We must also consider that the current distribution on all but the shortest antennas is not constant but nearly sinusoidal or a portion of a sinusoid. Because the losses are proportional to I^2R_{ac}, the loss will be different in different parts of the antenna. This can be accounted for by placing an equivalent resistance (R_{eq}) at the current loop, such that R_{eq} dissipates the same total power as the wire. The efficiency (η) of an antenna, taking into account only the radiation resistance (r_r) and the equivalent wire resistance, will be $\eta = r_r/(r_r+R_{eq})$. For $\lambda/2$ or $\lambda/4$ conductors with sinusoidal current distributions, $R_{eq} = R_{ac}/2$, where R_{ac} is the ac resistance for the entire wire length. A derivation of this result is given in the Appendix. For constant current distribution along the conductor, $R_{eq} = R_{ac}$.

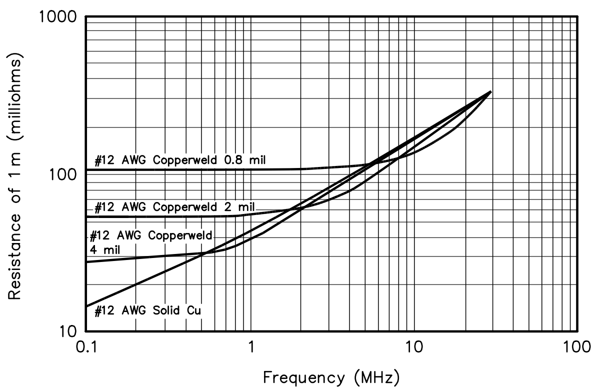


Fig 5—Resistance comparison of 1-meter lengths of #12 AWG solid copper, #12 AWG Copperweld with 4-mil cladding and an equivalent wire for 19-strand #26 AWG Copperweld with 0.8-mil cladding from 1 to 30 MHz. Derived from FEM modeling.

Table 2—Wire loss comparison for #12 wires.

	14.2 MHz Dipole		*1.85 MHz Dipole*		*1.85 MHz Ground-Plane*		*1.85 MHz W8JK*	
Conductor	*Gain (dBi)*	*Loss (dB)*	*Gain (dBi)*	*Loss (dB)*	*Gain (dBi)*	*Loss (dB)*	*Gain (dBi)*	*Loss (dB)*
Perfect	2.14	0	2.14	0	5.27	0	5.93	0
Copper	2.09	–0.05	2.01	–0.13	5.10	–0.17	4.92	–1.01
19–strand Copperweld	2.09	–0.05	1.88	–0.26	4.81	–0.35	3.93	–2.0
Aluminum	2.07	–0.07	1.94	–0.20	5.01	–0.26	4.47	–1.46
Iron	–1.88	–4.02	–4.99	–7.13	–2.58	–7.85	–9.58	–15.5

Table 3—Wire loss comparison for #18 wires

	14.2 MHz Dipole		*1.85 MHz Dipole*		*1.85 MHz Ground–Plane*		*1.85 MHz W8JK*	
Conductor	*Gain (dBi)*	*Loss (dB)*	*Gain (dBi)*	*Loss (dB)*	*Gain (dBi)*	*Loss (dB)*	*Gain (dBi)*	*Loss (dB)*
Perfect	2.13	0	2.13	0	5.27	0	5.94	0
Copper	2.08	–0.05	1.99	–0.14	4.93	–0.34	4.06	–1.88
Aluminum	2.06	–0.07	1.92	–0.21	4.75	–0.52	3.29	–2.65

Effects of Wire Loss on Gain

Okay, so as frequency increases, the resistance of the wire increases and different conductors have more or less loss. *So what! Does it really matter?*

One way to get a handle on this question is to model some typical antennas and determine the effect of different wire sizes and materials on gain. You can also calculate R_{eq} and then calculate the efficiency of the antenna. This is done in the Appendix. Tables 2 and 3 show the results of modeling three different antennas using perfect, copper (Cu), aluminum (Al) and iron (Fe) conductors of two different sizes. I assumed a resistivity of 10^{-7} Ω-m and a relative permeability of 1000 for the iron wire. Steel wire could actually be worse (lower conductivity and higher permeability). The dipoles and the W8JK array are modeled in free space. The W8JK array has two λ/2 dipoles, spaced λ/8 apart and fed 180° out of phase. The ground-plane antenna has four radials, 10 feet above perfect ground.

The tables show several things of interest. First, for the same wire size, as frequency decreases the wire loss increases. This is because even though the wire resistance per-unit-length is decreasing ($1/\sqrt{f}$) the wire length is increasing (1/f). The net wire resistance increases as frequency decreases if the antenna length is scaled. This increase in wire loss can become important in low-band antennas. Second, except for the iron wire, the effect of wire loss and wire size is very small in dipole antennas. You can use copper or aluminum wire in fairly small sizes without compromising performance much. It is also clear that using iron fence wire is bad news.

The ground-plane antenna is more sensitive to wire characteristics than are the dipoles because of its lower impedance, but again the changes are small as long as copper or aluminum wire is used. The use of more radials will reduce wire loss.

The W8JK array, however, is very sensitive to wire size and material. With perfect conductors, the gain over a dipole is 3.8 dB. Using #18 AWG aluminum wire gives away most of that gain (–2.65 dB). Even with #12 AWG copper wire, there is still a loss of over 1 dB. In the W8JK, changing to a #6 AWG wire or two parallel, spaced #12 AWG wires reduces the wire loss to –0.53 dB.

Any low-impedance antennas, such as Yagis, end-fire arrays or short loaded verticals will be sensitive to wire size and conductivity. On 80 and 160 meters, many verticals are short and heavily loaded!

Flat-Strip Conductors

Up to this point, we have been considering round conductors. An alternative would be to use thin, flat conductors of either copper or aluminum. Fig 6A is a graph of R_{ac}/R_{dc} for thin, flat-strip conductors (see the Appendix for generating equation). For #12 AWG round copper wire at 1.8 MHz, $F_r \approx 11$. If we take the same wire and roll it out into a strip approximately 0.010×0.625 inches, the thickness of the strip in skin depths will be about 5. Looking at Fig 6 we see that for $X = 5$, $F_r = 2.4$, which is a factor of 4.6 lower than for the equivalent round wire. By dividing F_r by the corresponding values of X, we can create Fig 6B, which is a graph of resistance normalized to 1 Ω for a thickness of 1 skin depth. Notice that for $X < 1.5$, the $R_{ac} = R_{dc}$, but as foil thickness increases the resistance goes through a minimum at $X = \pi$ and then back up about 9% to level out at a constant R_{ac} regardless of the thickness. For $X < 1.5$, the current distribution in the conductor is almost uniform, so $R_{ac} = R_{dc}$. Above this point, the distribution in increasingly on the outer surfaces of the strip.

At high frequencies, all the current is on the outer perimeter of the strip so the thickness of the inside doesn't matter. Only the length of the perimeter counts. This is the same as for a round wire. The important difference between round and strip conductors is that for a round wire, you have to increase the diameter to reduce R_{ac}. This means you have a lot of unused copper (inside the wire) to buy. Strip or foil conductors can be kept thin and simply made wider to reduce R_{ac}. You put the extra copper to good use and in the end buy less.

Of course, there is the issue of increased wind area with a foil conductor. Foil also tends to "sail" and/or flutter in the wind, distorting the antenna shape and stressing the array. That is a downside! Putting a spiral twist in a foil conductor helps to keep it from flying around in the wind. I have found that 0.010×0.5- to 1-inch strip works pretty well and doesn't fly around or flutter too much. Unfortunately, copper and alu-

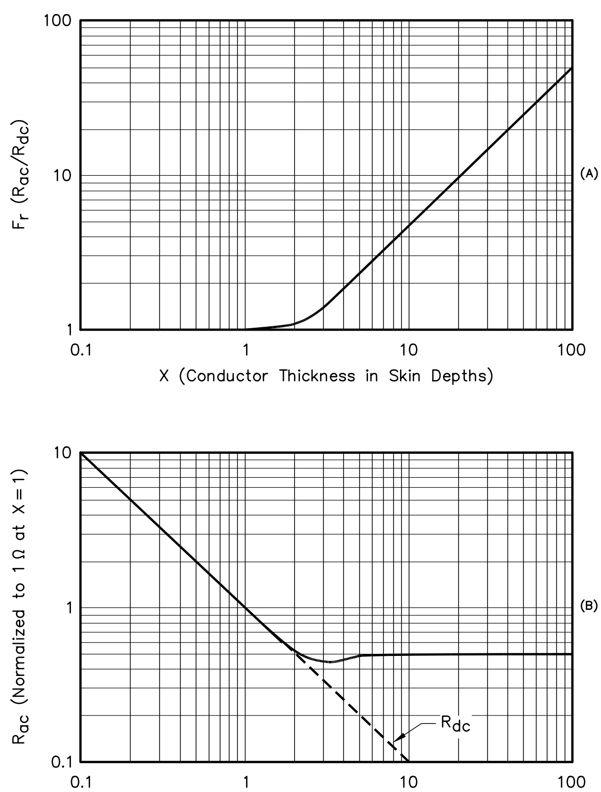


Fig 6—Resistance factor ($F = R_{ac}/R_{dc}$) for flat-strip conductors as a function of thickness in skin depths.

minum foils of appropriate sizes are not so readily available as round wire.

Composite Antenna Assemblies

In some cases, straight copper wire simply does not have the strength required, but the alternatives may have too much resistance. It is possible to compromise by using different conductors at different places in the antenna. Keep in mind that the losses are I^2R in nature. This means that the bulk of the losses occur in the high-current regions of the antenna. Fig 7 shows a 160-meter, two-element end-fire array mentioned earlier. The vertical portions have high current levels; they are made from 0.010×0.625-inch copper strip. The horizontal portions have much less current. The antenna is supported from the top between two poles 300 feet apart, so the upper wires have considerable stress. These are stranded Copperweld. The lower horizontal wires have very little stress; they are copper. The result is an antenna with minimum loss but strength where it is needed.

The 50-pF capacitors tune out the inductive reactance at the feedpoint. These must be high-voltage, high-current capacitors, which usually come in only a few standard sizes. The position of the capacitors and the lengths of the upper horizontal wires can be adjusted to give 450 Ω resistive at the feedpoint. That allows the use of 450-Ω ladder line as the feedline to ground level, where a 9:1 balun transforms to 50 Ω for the run back to the shack. Stub matching could be used instead.

Measurement of Wire Resistance

Theory and modeling are nice, but I wanted to make some actual measurements of wire resistance to confirm the modeling and calculations. Unless you have access to an impedance analyzer such as an HP4192 ($50,000 please!), this is not an easy measurement to make directly. After several false starts, I found it best to wind the wire into a large coil of well spaced turns and measure the Q on a Boonton 260A Q-meter. This gave reasonable results that are shown in Tables 4, 5 and 6. The values for resistance are probably not very precise, but the relative differences between different wires are clearly shown.

The coil is 17 turns (except for the #8 AWG aluminum wire which used 16 turns) spaced 1.5 wire diameters (with ⅛-inch Dacron rope) on a 4.2-inch ID PVC-pipe form. 4.5 inches long. The coil requires 19.5 feet of wire. The copper wire was #12 AWG, the antenna wire was 19 strands of #26 AWG Copperweld. (The Copperweld is nominally #13, but when I put a micrometer on the two wires, the sizes were not very different: 0.077 inches for the antenna wire, versus 0.082 inches for the solid copper wire. This results in only a 6% resistance difference.)

I took great care to make the two coils identical. They were both wound on the same form. Measurements were done with the same lead lengths and coil position relative to the Q-meter. A frequency counter was used to set the Q-meter frequency. The Q-meter zeros were carefully adjusted, and so on. The close values for the resonating capacitance show that the coils were very nearly identical.

Three things jump out at you from Tables 4 and 5:

- The Q for the coil made with stranded Copperweld is substantially lower than that of the solid copper wire coil.
- The variation in Q over frequency is different for each coil.
- The coil Qs begin to converge as the frequency is increased.

Remember that $Q = X_L/R_S$, where $X_L = 2\pi fL$ is the impedance, and R_s is the total series loss resistance.

In an antenna, we are interested in the resistance due to skin effect (R_{ac}), so we must separate the components of coil loss to get an estimate of the skin-effect loss. R_s has several components:

- skin effect in the conductor
- turn-to-turn and geometric proximity effects
- losses in the coil form
- loss in wire insulation
- radiation from the coil
- losses due to eddy currents in nearby conductors

Skin effect can be calculated quite accurately using the equation of Fig 1

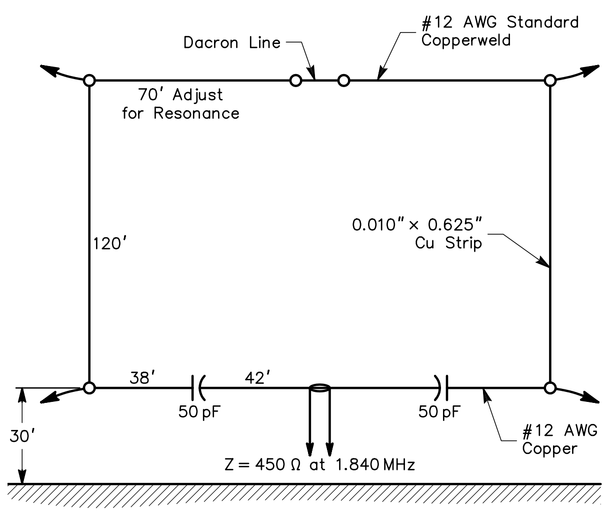


Fig 7—N6LF W8JK two-element vertical array using composite-wire construction.

Table 4 —#12 AWG bare solid copper wire test results using new wire

Frequency	*Resonating Capacitance*	*Measured Q*	X_L	R_S
1.8 MHz	359 pF	410	246.3 Ω	0.601 Ω
3.9 MHz	69 pF	360	591 Ω	1.64 Ω

Table 5—19-strand #26 AWG Copperweld test results using new wire

Frequency	*Resonating Capacitance*	*Measured Q*	X_L	R_S
1.587 MHz	460 pF	250	218 Ω	0.87 Ω
1.8 MHz	358 pF	270	247 Ω	0.92 Ω
3.9 MHz	68 pF	323	600 Ω	1.85 Ω

for a solid, round conductor. For 19.5 feet of #12 AWG copper wire at room temperature and 1.8 MHz:

R_{dc} = 0.031 Ω, from wire table (I measured the coil as 0.030 Ω on a bridge),

$R_{ac}/R_{dc} = F_r = 10.79$, from Fig 1,

$R_{ac} = R_{dc} \times F_r = 0.334$ Ω

At 1.8 MHz, the total R_s in Table 4 is 0.601 Ω, which indicates an additional loss resistance of 0.267 Ω beyond the skin effect.

Given the close similarity between the two coils, we can estimate the stranded Copperweld coil resistance component due to skin effect to be:

$R_{skin} \approx R_s - 0.267 = 0.915 - 0.267 = 0.648$ Ω

This is 1.9 times the resistance of solid copper wire! This agrees rather well with the comparison in Fig 5 between 0.8-mil clad Copperweld and solid copper wires. In a dipole, I don't think this would matter but in a W8JK array, it's bad news.

Looking again at Fig 5, we would expect the skin-effect loss for the two types of wire to converge as we go higher in frequency, reflected in more similar Qs. This is what we see in Tables 4 and 5. We would also expect the Q of the Copperweld coil to decrease with frequency because X_L is decreasing, but R_s is not. For the solid wire coil, both X_L and R_s are decreasing, so Q is more stable.

Emboldened by these results, I wound coils using several other wires I had on hand or was able to scrounge from friends. The test results are given in Table 6. I threw in the iron fence wire just for kicks!

The differences in the 14 different wires tested are quite easy to see:

- The #12 AWG wire is better than #14 AWG
- New insulation has very little effect (but weathered insulation may not be so benign!)
- Oxidation of bare wire definitely reduces the Q. Both samples were only mildly oxidized. Longer exposure would have further reduced the Q.
- Stranded wire is inferior to solid
- Very fine stranding (168-strand sample) reduces the Q significantly
- For the same size wire, solid Copperweld is just as good as solid copper
- At least at low frequencies, stranded Copperweld is inferior to solid Copperweld and other copper wires, solid or stranded
- Iron fence wire is bad news!

I also wanted to verify the advantage of Copperweld wire implied by Fig 4. Using #14 AWG solid copper and solid Copperweld, I wound free-standing three-turn coils and then two coils on a ceramic coil form. The results are shown in Table 7.

In both cases, the Copperweld produced a coil with somewhat higher Q, as predicted by Figs 4 and 5. Remember that only part of R_s results from skin effect, so the difference between the two wires is diluted by other losses. The tests were run a number of times to be sure the differences were real and repeatable.

Aluminum Wire Connections

Aluminum wire has the advantages of very low cost and a better strength-to-weight ratio (≈3×) than copper. The reduced conductivity (σ) of aluminum can be accommodated by using a larger wire size. For an equal resistance, it will still weigh less than copper. Keep in mind we are talking about equal R_{ac} not R_{dc}! The difference arises because of skin effect, which is proportional to $1/\sqrt{\sigma}$. The skin depth will be greater in aluminum than in copper (at the same frequency) because of the lower conductivity. The lower weight and higher strength is helpful in long spans and may put off the need to use Copperweld conductors.

However, aluminum has one major disadvantage. Making a low resistance connection that will remain low during extended exposure to the elements is not a trivial exercise. It is very possible for a poor connection to introduce significant loss, especially if it is at a high-current point. There are also corrosion problems with connecting copper conductors to aluminum conductors.

Alumoweld Wire

In addition to Copperweld, aluminum-clad steel wire is available under the name Alumoweld. It is available in a variety of sizes, although the smallest size available is #12 AWG. It is also available as stranded wire and stranded guy wire equivalent to the galvanized wire used for guys. While it is very stiff—handling very much the same as Copperweld or steel wire—it has some advantages. In most atmospheres, it is much more resistant to corrosion than galvanized steel. It is electrolytically compatible with the aluminum tubing frequently used in antennas, so it can be used for support wires in aluminum antenna structures to avoid dissimilar-metal corrosion.

Towers and Supports

It is quite clear that iron fence wire is a very poor choice for antennas, but what about steel towers and the use of galvanized or stainless steel guy wires as antenna elements? In towers, the surface area is much larger than that

Table 6—Comparison of Q for coils made with various wires at 1.8 MHz

Wire description	*Q*
New #12 bare soft-drawn solid copper	410
New insulated solid #12	410
New insulated stranded #12 THWN	350
New insulated 19 strand #26 Copperweld	270
New #14 bare soft-drawn solid copper	353
New #14 bare solid Copperweld	360
New #14 bare stranded Copperweld	194
Oxidized #14 bare stranded Copperweld	162
New #14 bare 7/22 stranded hard-drawn copper	338
Oxidized #14 bare 7/22 stranded hard-drawn copper	300
New #14 168 strand superflex	225
#14 aluminum electric fence wire	260
#8 aluminum clothesline	360
#13 iron fence wire	25

Table 7—Coil Qs measured at 25 MHz

Coil form	*Wire*	*Measured Q*	X_L	R_S
Air	Copper	285	145 Ω	0.51 Ω
Air	Copperweld	310	138 Ω	0.45 Ω
Ceramic	Copper	266	186 Ω	0.70 Ω
Ceramic	Copperweld	282	193 Ω	0.68 Ω

Appendix

A. Skin depth in copper

Fig A is a graph of skin depth in copper as a function of frequency for two temperatures.

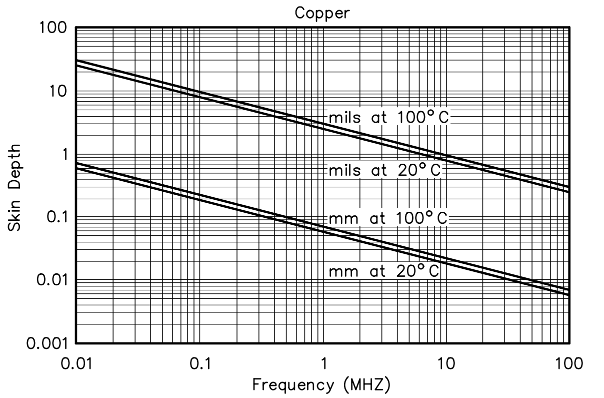


Fig A—Skin depth in copper at 20°C and 100°C; dimensions are in mils and millimeters.

B. R_{eq} Derivation

The current distribution in an antenna is usually a sinusoid or a portion thereof as indicated in Fig B. With the center as the origin:

$$I = I_{o(RMS)} \cos\left(\frac{\pi}{2}\right)\left(\frac{x}{l}\right) \qquad \text{(Eq A)}$$

Note that I_o, the current at $x = 0$, is RMS! The wire loss is:

$$\Delta P = \Delta R I^2 dx \qquad \text{(Eq B)}$$

Where ΔR is the resistance per unit length. The total power loss is then:

$$P = \int_a^b \Delta R I^2 dx = \Delta R I_o^2 \int_a^b \cos^2\left(\frac{\pi x}{2l}\right) dx$$

$$P = \left(\Delta R I_o^2\right)\left[\frac{l}{\pi}\sin\left(\frac{\pi x}{2l}\right)\cos\left(\frac{\pi x}{2l}\right) + \left(\frac{x}{2}\right)\right]_a^b \qquad \text{(Eq C)}$$

$$R_{eq} = \frac{P}{I_o^2} = \Delta R\left[\frac{l}{\pi}\sin\left(\frac{\pi x}{2l}\right)\cos\left(\frac{\pi x}{2l}\right) + \left(\frac{x}{2}\right)\right]_a^b$$

For $a = 0$ and $b = l$:

$$P = I_o^2\left(\frac{\Delta R l}{2}\right)$$

$$R_{eq} = \left(\frac{\Delta R l}{2}\right) \qquad \text{(Eq D)}$$

Where $\Delta R l$ is the total R_{ac} for the length of wire.

R_{eq} can be used directly to calculate the gain decrease due to conductor loss. The loss is simply the log of the efficiency:

$$loss = 10\log\left(\frac{r_r}{r_r + R_{eq}}\right) \text{ dB} \qquad \text{(Eq E)}$$

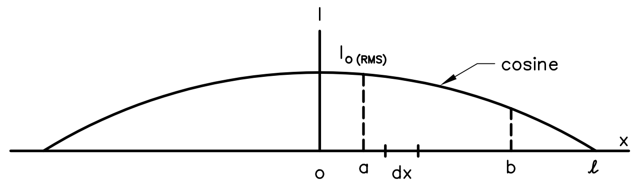


Fig B—Current distribution on an antenna wire and definition of equation quantities.

where r_r = radiation resistance. For a dipole in free space where r_r = 73 Ω, see Table 8.

The loss in gain by this calculation agrees with the gain loss in Table 2 that was derived using MOM (method of moments) in an antenna-modeling program.

C: Resistance Factor for Round Wire

$$F_r = \frac{R_{ac}}{R_{dc}} = \frac{\gamma}{2}\left[\frac{ber\gamma\, bei'\gamma - bei\gamma\, ber'\gamma}{ber'^2\gamma + bei'^2\gamma}\right]$$

$$\gamma = \frac{d}{\delta\sqrt{2}} \qquad \text{(Eq F)}$$

$$\delta = \text{skin depth},\ d = \text{ wire diameter}$$

where *ber* and *bei* are the real and complex parts of Bessel functions with complex arguments. They are often called Kelvin or Thompson functions. Most spreadsheet programs do not have these functions. Math programs like *Maple* or *Mathmatica* do have them. Fig 1 was done with Maple. It is possible to use series summation approximations that can be found in advanced math tables (see Reference 6).

D. Resistance Factor for Thin, Flat Foil where X = Thickness in Skin Depths

$$F_r = \frac{R_{ac}}{R_{dc}} = \frac{X}{2}\left[\frac{\sinh X + \sin X}{\cosh X - \cos X}\right] \qquad \text{(Eq G)}$$

This equation may readily be evaluated with a spreadsheet. Most spreadsheets have both circular and hyperbolic functions.

Table 8—Loss due to conductor resistance for dipoles using #12 AWG solid copper wire

Frequency (MHz)	δ (mils)	*X*	F_r	*L* (feet)	R_{eq} (Ω)	*Loss* (dB)
1.84	1.92	29.8	10.8	267	2.29	–0.13
3.75	1.35	42.5	15.3	131	1.59	–0.09
7.15	0.98	58.7	21.0	68.8	1.15	–0.07
14.2	0.69	82.7	29.5	34.6	0.81	–0.05

of a wire. Although the skin depth will be very small, the large surface area should help greatly. I would be more concerned with the joints between tower sections, particularly in high-current regions. This problem has been addressed by attaching copper wire jumpers across tower joints. The problem will be much worse in crank-up towers, where the sections have sliding joints between them.

I would be more concerned about loss in a steel tower if it were being used as part of an array with low impedances, especially if the tower is electrically short and heavily loaded. In that case, I would consider installing a collar at the top of the antenna and attaching several parallel copper wires in a cage around the tower from top to bottom. This way the copper is the conductor not the tower. This allows the tower to be grounded directly but still have the feed point open. If the collar were made significantly larger than the tower, then it would not only reduce loss but also increase bandwidth and reduce the loading necessary because of the larger effective diameter of the antenna.

Sometimes the guy wires on a tower or the rigging on a sailboat are used as antennas. Depending on the antenna, these can be very lossy and should be

Why is there Skin Effect?

When a time-varying current flows in a conductor, a time-varying magnetic field will be created around the conductor. A simple example is shown in Fig C. A current flowing in the wire creates a magnetic field around the wire as indicated. The direction of the magnetic field in relation to the current obeys the "right-hand rule"—that is, if the thumb of your right hand extends in the direction of positive current flow as shown, the magnetic field will curl around the wire in the same direction as your fingers.

Just as a current creates a magnetic field, a time-varying field, from some external or internal source, will induce a time-varying current in a conductor. This is called an "eddy" current and higher frequencies yield greater-amplitude eddy currents in a given conductor. The direction of the eddy current is such that its magnetic field opposes the inducing field.

We can see how these currents and fields create skin effect by examining Fig D. This is a section of a round wire carrying a current from one end to the other. This current is labeled "A." It is simply the net current flowing through the wire. This current creates a magnetic field both inside and outside the wire as indicated by the dashed lines "B." This field, in turn, creates an eddy current ("C") as shown.

Notice that near the center of the wire, the eddy current opposes the desired current, but on the outer part of the wire, the eddy current aids the desired current. If we look at a cross-section of the wire, we see that the current density near the center is reduced, but near the outside, the current density is increased. As frequency increases, less current flows on the inside of the wire and more flows near the outside surface. Of course, the *net* current stays the same, but it is crowded into a smaller and smaller portion of the wire's cross-sectional area.

The result is that the apparent resistance of the wire increases because we are using only a small portion the available copper area to carry current. This means that the loss for a given current will be higher. In copper at HF, the current is crowded into a layer of 2 mils, or less, in thickness. The rest of wire only provides mechanical support for the thin outer layer that conducts!

There is another way to look at skin effect. If you have a large sheet of conductor and you irradiate it with a electromagnetic wave perpendicular to the surface, the wave will penetrate the surface for some small distance. The amplitude of the wave decreases exponentially and the depth at which the amplitude has decreased to $1/e \approx 37\%$ ($e \approx 2.718$, the base of natural logarithms) is referred to as the penetration or skin depth (δ). Increasing frequency decreases δ.—*N6LF*.

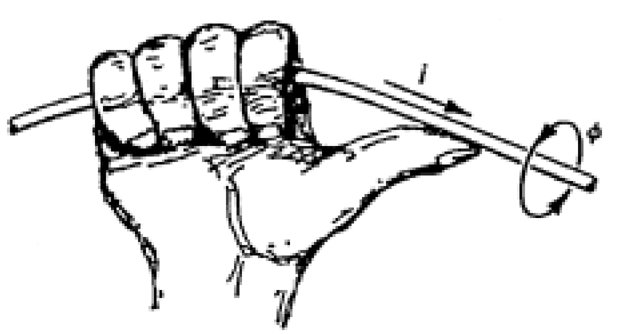


Fig C—The "right-hand rule" relates the direction of current flow to the magnetic field it produces.

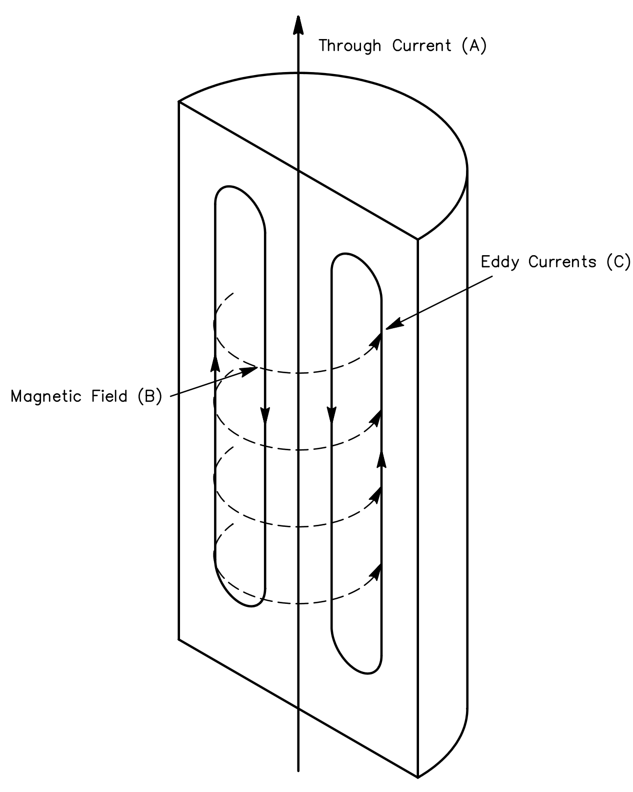


Fig D—Eddy currents in wire produce the skin effect. The through current (A) produces a magnetic field (B) that induces eddy currents (C). The eddy currents offset through current near the wire center and add to through current near the wire surface.

used with some caution. Note from Table 1 that the standard marine stainless steel (304) has a resistivity greater than 50 times that of copper. A number of years ago, I used an insulated backstay on my sailboat as a half-sloper, fed at the top and driven against the aluminum mast. To minimize the loss in the stainless steel backstay, I used a strip of copper (encased it in plastic tape to control corrosion) bent over the backstay in the form of a U for the distance between the two insulators. This proved very satisfactory during several years of cruising in temperate and tropical waters.

Stainless Steel and Mobile Antennas

Most mobile antennas are electrically short and heavily loaded, especially at and below 7 MHz. The result is very low radiation resistances. Because of its very high resistivity, stainless steel may not be a very good choice for these antennas despite the obvious mechanical and corrosion-resistance advantages. For example, consider an 8-foot center-loaded whip with a 0.5-inch diameter base section and a 0.125-inch diameter top section. The loss due to conductor resistance using stainless steel is 0.6 dB at 7.150 MHz, 1.3 dB at 3.8 MHz and 3.1 dB at 1.84 MHz. The use of stainless steel wire would result in losses very similar to steel fence wire.

Conclusions

For antennas with current-loop impedances above 35 Ω or so, any copper, Copperweld or aluminum wire in a variety of sizes will work just fine; however, for lower-impedance antennas, copper or Copperweld wire size #12 AWG or larger should be used. Copper or aluminum tubing is very effective for low-impedance antennas. For 80- and 160-meter antennas, the resistance of stranded Copperweld may be unacceptably high.

New insulation does not seem to affect loss, at least at 1.8 MHz, but surface oxidation does. Thin insulation should have only a very small effect on tuning but will suppress oxidation. This is a consideration for low-impedance antennas only.

By careful choice of conductor or combinations of conductors, considering both electrical and mechanical properties, it should be possible to keep the conductor loss low in almost any kind of antenna, with the possible exception of very small antennas.

Loose Ends

Despite the extensive discussion in this article, several subjects need more attention. I think the losses in steel towers need to be analyzed more closely. I also have not addressed losses from currents induced in guy and support wires. Usually these currents are small if the wire is short compared to $\lambda/2$, but steel wire can be quite lossy even with small currents. This subject needs some scrutiny. In searching through the literature, I found very little in the way of measurements or even discussion of antenna conductors. Books in the reference list contain some very useful tables, but if you know of any important articles I have missed please tell me.

Acknowledgment

I would like to express my thanks for the review and helpful comments provided by George Cutsgeorge, W2VJN. Mark Perrin, N7MQ, and Joe Brown, N7EZG, were very helpful in digging out great examples of "cruddy" wire for me to test. Tom Schiller, N6BT, told me about Alumoweld wire and related a number of hilarious anecdotes concerning antenna conductors.

Notes

1. J. Kraus, "Antenna Arrays with Closely Spaced Elements," *Proceedings of the IRE*, February 1940, pp-76-84, see Fig 5.
2. "Maxwell" by Ansoft Corporation, Four Station Square Ste 200, Pittsburgh, PA 15219-1119; tel 412-261-3200, fax 412-471-9427; e-mail **info@ansoft.com**; **http://www.ansoft.com**.
3. United States Alumoweld Company, Inc, 115 USAC Dr, Duncan, SC 29334; tel 800-342-8722, 864-848-1901, fax 864-848-1909; e-mail **sales@alumoweld.com**; **www.alumoweld.com**.
4. F. Terman, *Radio Engineers Handbook* (McGraw-Hill, 1943), pp 32-33.
5. D. Fink and W. Beaty, *Standard Handbook for Electrical Engineers*, 11th Ed., ISBN: 0070220050 (McGraw-Hill, 1999) Chapter 4.
6. *Radiotron Designer's Handbook*, 4th Ed. (RCA, 1952), pp 1408-1417.
7. M. Abramowitz and I. Stegun, *Handbook of Mathematical Functions*, National Bureau of Standards republished by Dover Publications (New York, 1965), pp 379-385.
8. R. Straw, N6BV, Ed., *The ARRL Antenna Book*, 18th Ed. (Newington: ARRL, 1997), Chapter 20.
9. J. Devoldere, ON4UN, *Low-Band DXing*, 3rd Ed. (ARRL, 1999), pp 8-5 through 8-7.

Measurement of Soil Electrical Parameters at HF

The author describes the technique he uses to measure soil conductivity and relative dielectric constant over a range of frequencies on the HF bands.

Rudy Severns, N6LF

Introduction

Modeling of antennas over real ground requires at least a reasonable guess of the values for the soil conductivity (σ) and relative permittivity (ε_r), also referred to as "relative dielectric constant." Unfortunately these numbers are usually not readily available. From broadcast (BC) work we have charts of ground conductivity covering large areas, but these numbers give only σ, not ε_r. In part, the absence of ε_r data is because, for sites where you would want to build a BC station, the soil characteristics are usually dominated by σ, and ε_r has only a second-order effect. This is often true at BC frequencies but is usually not the case in all but the most conductive soils at HF. In addition, the values for σ will be different between BC frequencies and HF. Another problem is that the BC ground conductivity charts cover much too large an area to take into account the details of local ground variation, which can deviate greatly from local averages.

It would appear that the best approach is to simply measure your local soil characteristics at the frequencies of interest. Unfortunately, this is much easier said than done. None of the known methods is anywhere near perfect, and many are difficult to implement. In fact there is a school of thought that the problem is impossible and we should not waste our time worrying about it. I don't share that view as a general proposition, but it is not without some justification, given the difficulties involved.

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There are common situations where the values for the soil constants (which are anything but constant!) are really not very important for modeling purposes. For example, for horizontal polarization with antenna heights above $^1/_4$-λ, the numbers are not very critical for determining feed-point impedances, near-field losses or the formation of the far-field radiation pattern. Another case would be for vertical antennas where one has the space, money and patience to lay down a large number of long radials. With this brute force approach, the near-field ground loss can be made arbitrarily small regardless of the soil, and you really don't care what the ground constants are, at least from a local loss point of view. The far-field pattern, however, is still just a guess without real data.

If your space and/or financial resources are more limited, then a modestly accurate estimate of your ground characteristics will allow you to design a ground system that minimizes the loss within the constraints of the space and resources you do have. There is also the situation that arises fairly often on 80 m. On that band a $^1/_2$-λ is about 130 feet, which is not an exceptionally tall tower for amateur use. Horizontal gain antennas are certainly practical but they're not easy, being large and relatively expensive. The option is to go to a vertical array, which may be easier. Accurately predicting the performance of a possible vertical array in comparison to a competing horizontal array requires at least a reasonable guess for the ground characteristics on that band. The decision as to which way to go may depend, at least in part, on having a reasonable estimate of the ground characteristics.

Another problem with present ground modeling practice is the assumption that soil parameters, whatever they may be in a given location, are constant over frequency. For example, for a given soil, the assumption is that ground constant values at 160 m are the same as the values at 20 m. That's not the case. As pointed out by Bob Haviland, W4MB, and in many professional papers, ground parameters at HF vary substantially with frequency.[1,2]

There is a need for a practical method to estimate soil parameters at HF for amateurs. By practical, I mean a mechanically simple test apparatus and measurement equipment

[1]Notes appear on page 8.

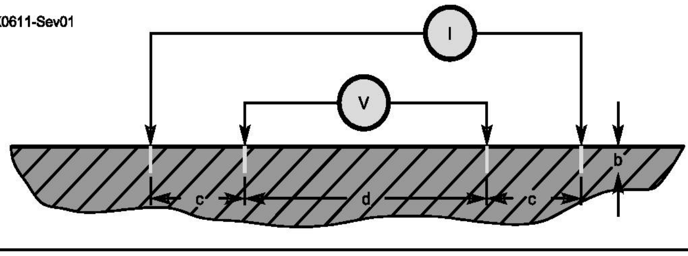

Figure 1 —This diagram shows the Wenner array, the traditional method used by amateurs to measure ground parameters.

no more advanced than an AEA or MFJ impedance analyzer. Fortunately, great accuracy is not required and it makes little difference in the modeling if the values are off by 25%. This article will discuss one particular approach, the use of ground probes, which at least approximate this ideal.

Soil Parameter Measurement

There are many ways to measure ground parameters. Each has advantages and limitations. None is the perfect answer. The traditional method used by amateurs is the Wenner array (or similar variations), which uses four probes in line as shown in Figure 1, excited with line-frequency ac.[3,4,5] This approach can give a good estimate of ground conductivity at 50 to 60 Hz, and by varying the spacing of pairs of probes, can be used to define subsoil layering. It gives no information on ε_r, however. A measurement of this type provides only a lower bound on soil conductivity, which will be higher at HF.

Another technique, frequently used in BC work, is to measure the rate of decrease of the E-field intensity as you go away from the antenna on a radial line. It is possible, by some judicious curve fitting to the measured data, to infer the average ground conductivity along the measured line. This is a reasonable approach at BC frequencies, where the soil characteristics are dominated by the conductivity. At HF, however, the soil is both resistive and capacitive. Typically, when trying this technique at HF, more than one pair of parameters (σ and ε_r) may generate curves that fit the data. This ambiguity is a problem. In addition, the measurements need to be made at some distance from the antenna where there are significant amplitude differences between measuring points and so do not give a very good idea of the ground characteristics within a $^1/_2$-λ of the base. Information on ground parameters close to the antenna is needed for ground system design, especially in the initial design stages for a new vertical.

A technique that would seem to fit our requirements is to insert a probe into the soil and measure its impedance.[6] In the simplest case the probe is basically just a capacitor, and the ground parameters are inferred from the change in impedance of this capacitor from when the probe is in air and to when it's in soil. This approach can yield a detailed characterization of the soil in the immediate area of the antenna and at distance also. A basic limitation of this procedure is that it is usually not possible to use a probe that reaches very far down into the soil. The result is characterization mainly of the top few feet of soil, which is usually substantially less than the skin depth. By making measurements at many spots over the area of interest the probe method can give a very good picture of the lateral variation of soil parameters. We know that the properties will also vary vertically (variations in moisture content, stratification, and so on) and we would like to know the variations down to a skin depth. It is possible to take a surface measurement, then dig down three feet or so in the same spot and reinsert the probe in the undisturbed soil at that level and make another measurement. This can be repeated until sufficient depth is achieved. That, however, defeats our goal of keeping the process simple, and is not practical for large-area surveys.

Is a fairly accurate characterization of only the top layer of soil of any real use? Certainly that's debatable but I think it is worth doing. There will be cases where the soil characteristics change slowly and the probe measurements are pretty close. It is also possible to have an entirely different strata a few feet down, with completely different characteristics. It probably is a good idea to dig one test hole as suggested above, to get a feeling for the local stratification and then do a survey with surface probes in the near area. In any case, I think probe measurements are a vast improvement over nothing but we should not be fooled into thinking the results are exact. Like everything in modeling, the information has to be used cautiously.

Monoprobe Technique

This method uses an impedance analyzer to measure the impedance of a single ground probe with a ground screen, as shown in Figure 2. The ground screen can be either square or circular, with a radius greater than the length of the longest probe. Rupar used a copper sheet for the ground screen.[7] I initially used a sheet of $^1/_8$-inch-thick aluminum, but a large metal sheet is awkward to work with and I found that a piece of ½ inch galvanized hardware cloth (as shown in Figure 3) worked just as well. Note the weights on the screen. The hardware cloth is flexible, and the weights are used to keep it in contact with the soil. This is an advantage if the ground is a little uneven in that the flexible screen may fit it better, minimizing any air gap between the screen and soil. More on this later. Anything will do for weights; bricks or rocks are fine. The flexibility of the hardware cloth means you can roll up the wire to make an easier package for

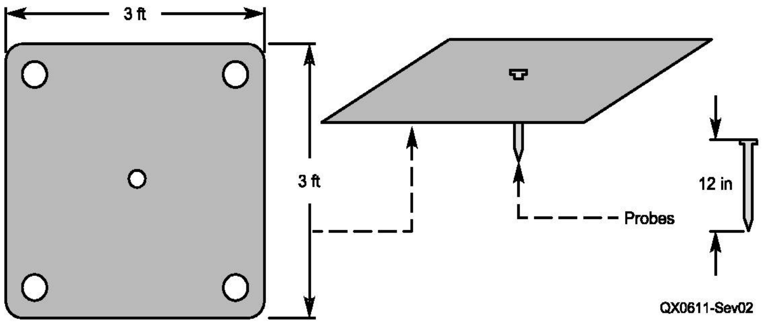

Figure 2 — This illustration shows the construction of the monoprobe. A three-foot square of half-inch hardware cloth forms the base. A hole cut in the center provides space for the probe to go through the screening without contacting the wire.

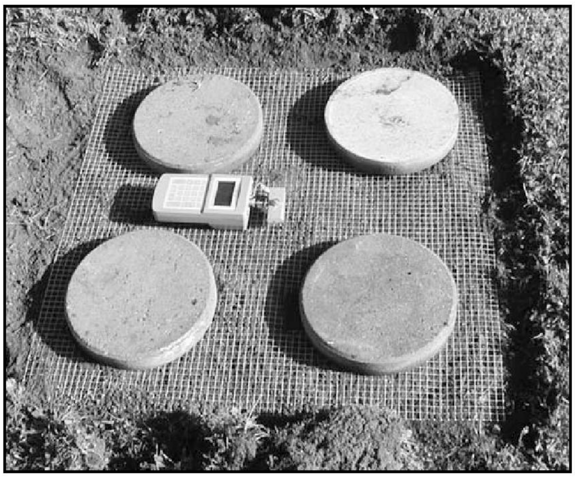

Figure 3 — This photo shows the monoprobe in use. Note the circular blocks used to hold the hardware cloth in contact with the ground, and the AEA impedance analyzer connected to the probe in the center.

carrying. The example in Figure 3 shows an AEA complex impedance analyzer being used. An MFJ-259B or other impedance analyzer will work just as well.

Figure 4 shows examples of typical probes (12 inches and 18 inches long). The crossbars in this example are phenolic, but any reasonable insulating material will work fine, even wood. The crossbar is there to help push the rod into the ground and pull it back out. The rod is inserted through a small hole (1 inch or so) in the screen and pushed down until the crossbar is pressed firmly onto the screen. The rods shown are brass but that's not essential. I just happened to have some $^3/_8$ inch brass rod stock on hand. For later experiments I found some inexpensive $^7/_{16}$ inch aluminum rod at a scrap yard. Anything from $^1/_4$ inch to $^1/_2$ inch should work fine. In fact, you can even use square rods if you wish. You can find suitable rod stock at most hardware stores. The larger diameters make for more sturdy probes, with a little more capacitance, but they may be harder to push into the ground. The presence of the thin insulating layer of oxide on aluminum rods has essentially no effect on the measurements.

Initially I threaded the top of the rod and the crossbar, then screwed them together, adding a nut on the top for a connection. You don't need to be so fancy. Later on I just drilled a tight fitting hole in the crossbar, drove the rod into it and added a cross 6-32 machine screw to hold it and to provide an electrical contact.

The impedance is measured between the top of the rod and the ground screen as shown in Figure 5. Note that I have used a lead from the top of the rod to the analyzer and a ground clip on the analyzer to connect to the screen. You could also mount a coaxial connector on the screen with a lead going to the top of the rod. The choice of which way to go affects the stray inductance and capacitance and is discussed in the appendix in the context of probe calibration.

OWL probes

The OWL (Open Wire Line) probes are simply two parallel rods and a crossbar without a ground screen, as shown in Figure 6.[6] The impedance is measured between the tops of the two rods. For a battery powered impedance analyzer like an MFJ, the measurement is floating (once you take your hands off the instrument!) and no balun is needed. If you want to use a more advanced analyzer, such as the Ten-Tec TAPR or N2PK vector network analyzers, with a cable, then a balun would be a good idea. I made up a test balun, which is included in the Figure 6 photo. I used a Fair-Rite FT240-43 ferrite core. This is standard core available from Amidon. The winding is a 3 foot length of RG-58 with BNC connectors at the ends. This length results in about 12 turns, and should give adequate isolation down to 1 MHz.

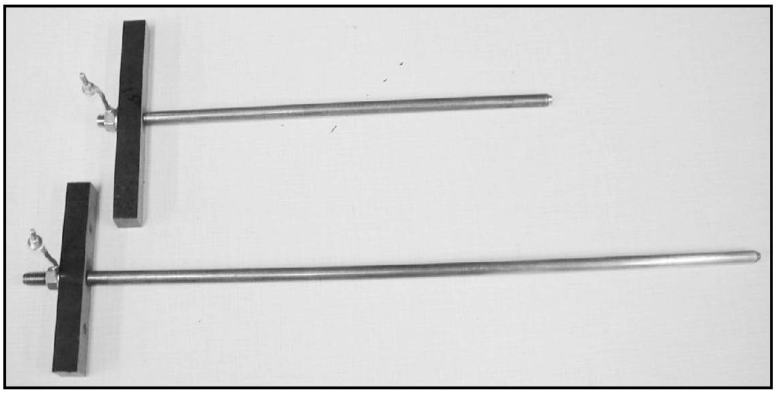

Figure 4 — Examples of 12-inch-long and 18-inch-long probes for use with the monoprobe technique.

Figure 5 — This photo shows the details of how the impedance analyzer is connected to the probe and ground screen.

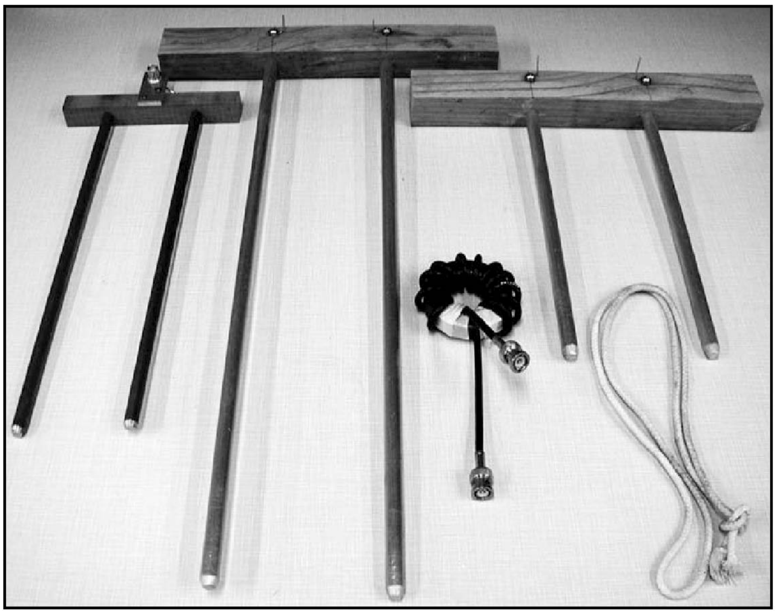

Figure 6 — Here are three open wire line (OWL) probes, along with a balun and a loop of rope used to pull the probes out of the ground after the measurements have been made.

The probes with 4 inch spacing use clip leads like those shown in Figure 5 but the 3 inch spacing probe (the probe on the left in Figure 6) has a BNC connector on the cross-bar, to which the balun is connected. There is nothing magical about either arrangement.

Figure 6 also shows a vital piece of equipment: the cord! Before pushing a probe into the soil it's a really good idea to loop the cord around the crossbar. You will use it to pull the probe out of the ground. In hard soils, with the bar pressed against the ground, getting the probe out without the cord can be a chore. It also helps to put a handle on the cord.

I'd like to emphasize that the diameter, spacing and length of the probe rods is not critical. The only thing you must do is to measure or calculate the probe capacitance (as shown in the appendix) for your particular probe. Larger diameter rods and closer spacing result in lower measured impedances. With the AEA and MFJ analyzers, measurements of impedances above 200 Ω or so become less reliable and it is better to work with lower impedances. In soils with poor conductivity the measured impedance, with the same probe, will be higher than in more conductive soils, and it may be better to use closer rod spacing.

Choosing Between a Monoprobe or an OWL Probe

Both types of probes will work just fine, but each has advantages and disadvantages. The single probe is much easier to insert than a double probe. There is also the issue of keeping the rods parallel with the OWL. If the rod spacing varies between air and in the soil then the calibration of C_0 will be off. Given the modest accuracy required for ham applications this is usually not a problem. The OWL is much more compact to carry around because you don't need the large ground screen and weights to hold it down. That's a very practical advantage!

The monoprobe is influenced by a much larger volume of soil and provides an average over that volume whereas the OWL pretty much characterizes a small cylinder of soil. The monoprobe measurement is intrinsically unbalanced whereas the OWL may require a balun or other isolation for measurement with non-isolated instruments.

In the end, either will work. You just have to decide what suits you.

Taking and Reducing the Impedance Data

The procedure is very straightforward. You simply lay the screen on the ground if using a monoprobe, insert the probe into the soil and record the impedance reading on the analyzer at each frequency of interest. In the case of an OWL probe, you simply insert the probe into the ground up to the bar and make an impedance measurement. This should only take a few minutes and then you move on to the next points, recording the impedance measurements as you go. You can cover a lot of ground in an hour or so.

The next step is to convert the impedance readings to σ and ε_r using the equations given below. Putting these equations into a Microsoft *Excel* spreadsheet makes the whole process very painless.

The impedances can be in either of two formats: R + *j*X or magnitude (|Z|) and phase angle (θ). The equations for converting the measured impedances using R and X are:

$$\sigma = \frac{8.84}{C_0}\left[\frac{R}{R^2+X^2}\right] \qquad \text{(Eq 1)}$$

$$\varepsilon_r = \frac{10^6}{2\pi f_{\text{MHz}} C_0}\left[\frac{X}{R^2+X^2}\right] \qquad \text{(Eq 2)}$$

If you prefer to work with |Z| and θ, the equations take the form:

$$\sigma = \frac{8.84}{C_0}\left[\frac{1}{|Z|\sqrt{1+\tan^2\theta}}\right] \qquad \text{(Eq 3)}$$

$$\varepsilon_r = \frac{10^6}{2\pi f_{\text{MHz}} C_0}\left[\frac{\tan\theta}{|Z|\sqrt{1+\tan^2\theta}}\right] \qquad \text{(Eq 4)}$$

where C_0 = capacitance in pF of the probe in air. This can be either measured or calculated quite closely, as shown in the appendix. Frequency is in MHz and impedances are in ohms.

These equations assume the probe is simply a capacitor. If you make the probe longer at a given frequency, or push the measurement frequency up for a given probe length, there comes a point where the probe is no longer simply a capacitor, it becomes an antenna, buried in the soil. It can still be used but data reduction is more complex.

Some Actual Measurements

Now it's time to look at actual measurements taken on my property. Tables 1 and 2 show typical impedance measurements taken at two different locations, and their reduction to σ and ε_r. The data in Tables 1 and 2 is graphed in Figures 7 and 8.

Because of the relatively poor accuracy of the AEA analyzer, the graphs are a bit "lumpy."

Table 1

18 Inch Monoprobe, C_0 = 7.41 pF. On my antenna hill with an AEA-CIA analyzer.

Frequency (MHz)	*Resistance (Ω)*	*Reactance (Ω)*	*Conductivity, σ (S/m)*	*Relative Permittivity, ε_r*
1	129	–134	0.0044	83
2	83.3	–95.8	0.0062	64
3	66.3	–76.2	0.0078	53
4	56.7	–65.2	0.0091	47
5	51.5	–57.2	0.010	41
6	46.2	–47.8	0.012	39
7	40.2	–46.2	0.013	38
8	35.1	–40.4	0.015	38

Table 2

Four Inch × Nine Inch OWL, C_0 = 2.71 pF. In my backyard with AEA-CIA, no balun.

Frequency (MHz)	*Resistance (Ω)*	*Reactance (Ω)*	*Conductivity, σ (S/m)*	*Relative Permittivity, ε_r*
1	176	–137	0.0042	59
2	123	–119	0.0050	44
3	95.2	–98.5	0.0061	38
4	83.4	–86.3	0.0069	32
5	77.7	–75.0	0.0079	28
6	73.0	–63.4	0.0093	24
7	60.9	–60.9	0.0098	25
8	54.7	–52.8	0.011	25

In Figure 7 I have smoothed things out by inserting a linear trend line, which fits quite well. The lumpiness is typical using this class of instrument for measurement, but the lumps are still small enough not to matter. The data in Tables 1 and 2 is a bit sparse, but taking a large number of closely spaced data points and then smoothing with a trend line works even better.

You may notice that in Figure 3, the grass has been dug away so that the screen is in direct contact with the soil. I made measurements with and without the grass to see what the effect the grass would have. The results are shown in Figures 9 and 10.

As you can see from the graphs, the presence of the grass doesn't have much effect on the conductivity measurements, but does substantially affect the ε_r measurements. What the grass does is to insert a layer of air under the screen, which reduces the effective capacitance. That, in turn, reduces the value for ε_r. This is not a big issue but you should at least take a string trimmer and cut the grass as low as possible. If you are using an OWL probe then the effect of the grass is very small if the probe is pushed firmly down against the ground.

Another concern is the effect of using different probe lengths. The moisture in the very uppermost layer of soil responds rather quickly to weather conditions. When it rains, σ and ε_r go up and, when things dry out, σ and ε_r fall. This rate of variation with time and depth depends on the soil itself but for the most part the soil characteristics respond much more slowly at depths beyond 12 inches or so. This effect on measurements is illustrated in Figure 11. The soil at W6XX is quite sandy and the top layer dries out fairly rapidly. We can see this in the graph. σ is substantially lower in the upper layer which is being measured by the short probe. The longer probes reach down into soil that dries much more slowly, and as you can see the two longer probes give essentially the same data. A close look at the 24 inch probe data line illustrates a limitation mentioned earlier on probe length. As the probe is made longer the current distribution along the probe is no longer essentially constant. Instead of behaving like a simple capacitor (which Equations 1 and 2 assume) it is starting to act like an antenna. Notice how the 24 inch probe curve starts to bend over at the higher end. This can be corrected by using more complex equations for the data reduction, but for most users that may be more trouble than it's worth. The usable range is still above 40 m. Very high conductivity soils may require shorter probes.

Comments on Ground Data

The conductivity graph (Figure 7) has an important feature: the ground "constants" are not constant at all with frequency. It is very typical in the HF region for σ to increase with

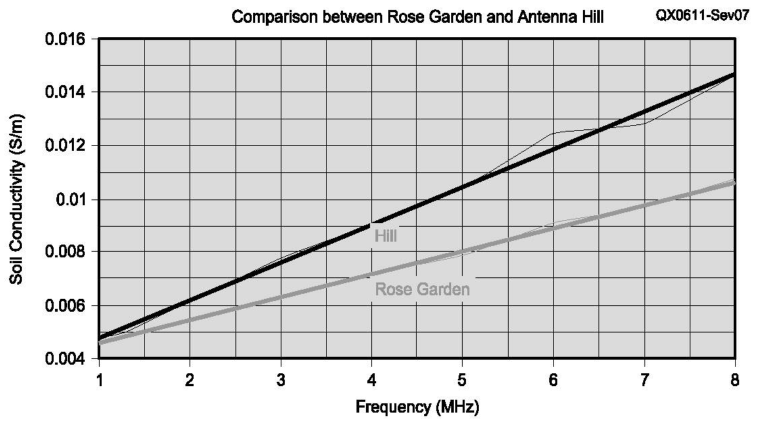

Figure 7 — This graph compares the conductivity I measured in two areas of my property.

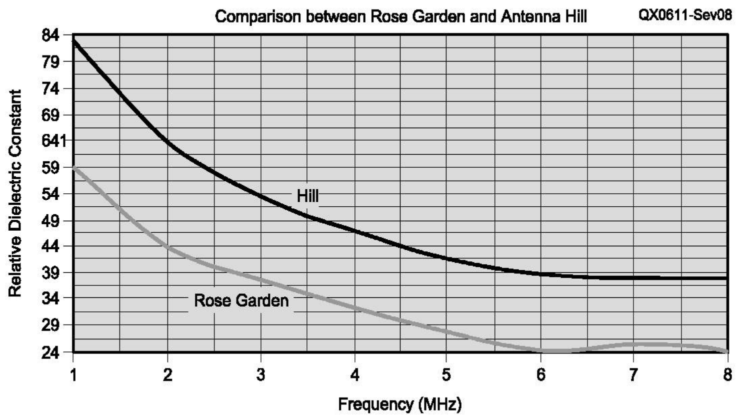

Figure 8 — The relative dielectric constant values on the antenna hill and in the rose garden, based on the measurements taken from 1 to 8 MHz.

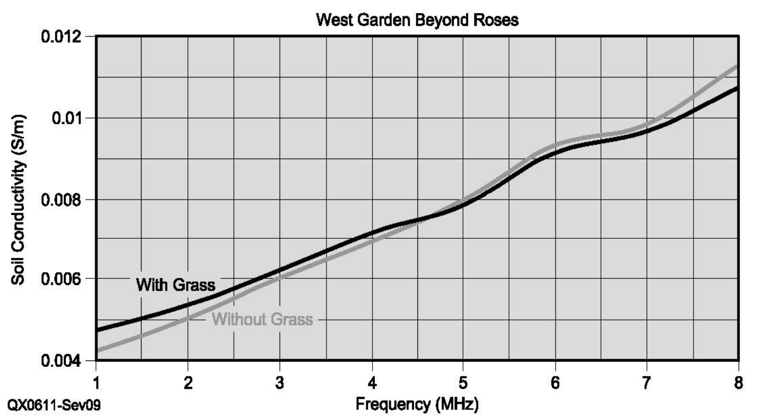

Figure 9 — I measured soil conductivity in an area of grass in my west garden, beyond the roses. Then I dug up a patch of grass and repeated the measurements to estimate the effects of grass.

frequency. In addition, as shown in Figure 8, ε_r is not constant either and tends to decrease as you go up in frequency to about 5 MHz, but then stabilize above that. The general shape and trends displayed in Figures 7 to 11 agree very well with those seen in the large body of professional work on ground parameter values.

At both sites in Figure 7, σ corresponds to what is generally called "average ground." Average ground is usually defined as σ = 0.005 S/m and ε_r = 13. In Figure 8, however, ε_r is much larger than 13, especially below 5 MHz. This is particularly characteristic of soils with a lot of clay particles. For many years there was a great deal of controversy over the large values of ε_r measured at low frequencies. The consensus is that it is very real. The following quote is from the King and Smith book, *Antennas In Matter*, which is considered a definitive work:[8]

"For some time, the high values of permittivity and the dispersion at these lower frequencies were thought to be artifacts of the measuring procedure; that is, it was thought that they were caused by electrochemical effects at the interface between the metallic electrodes and the sample of rock or soil. Measurements made using several different materials for the electrodes, however, indicate that there is a high permittivity associated with the geological material apart from any electrode effects."

Summary

How should we use the numbers we get? First, I try to take my readings at the end of the driest part of the year. Because both σ and ε_r are strong functions of soil moisture content, measuring near the end of the dry season will give you a conservative estimate. One exception I make is for my 80 and 160 m antennas, which I normally only use during the winter, which is definitely the wet season in Oregon. I use the winter ground parameters for these bands. Second, I average the readings found at different places over the site.

These are the values I use when designing a new antenna. Am I kidding myself? Well, perhaps, but I find it hard to believe that I am worse off than if I simply guessed and took the traditional value for mountains of σ = 0.001 S/m and ε_r = 5, which would appear to apply in my location even though these values are much lower than what I measure at my QTH.

I think that ground probe measurements are worth doing and I use them, but with care. Of course we would like even better methods, and in fact a number of workers are trying to find better ways.

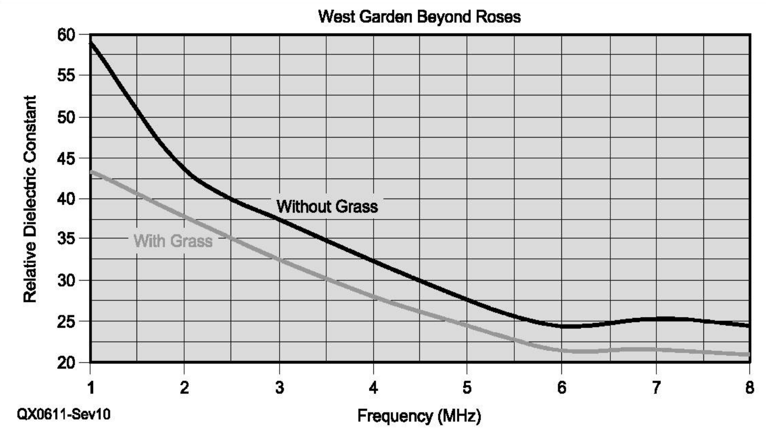

Figure 10 — This graph compares the relative dielectric constant as measured with and without grass.

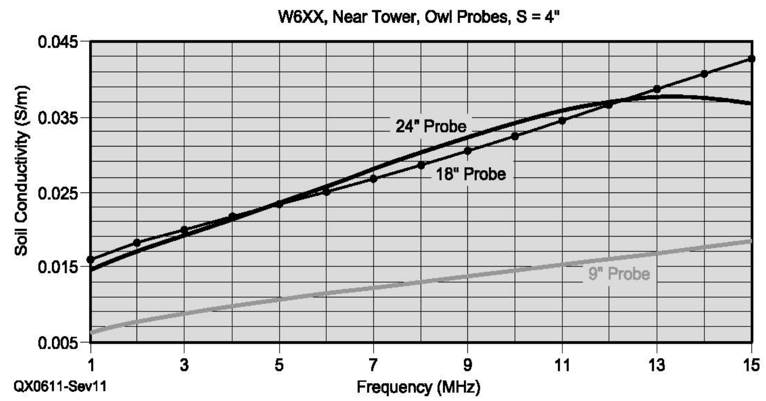

Figure 11 — This graph shows the data collected by Pete Gaddie, W6XX, near his tower, using three different-length OWL probes.

Acknowledgement

Pete Gaddie, W6XX, and I have been building and testing a variety of ground probes and much of what is in this article comes from our results. I would like to acknowledge the great assistance Pete has given me in supplying technical papers, extended discussion on technical details, measurement advice and a great deal of parameter measurement on his own. Thanks, Pete.

Notes

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Appendix — Determining C_0

The value for C_0, which appears in both of the equations for σ and ε_r, is the capacitance of the probe in air. It has to be determined before the impedance data can be reduced to σ and ε_r. There are two ways to go about finding C_0: direct measurement and calculation. For the OWL probes C_0 can be determined very closely from the following equation taken from Terman:[13]

$$C_0 = \frac{3.677}{\log_{10}\left\{\frac{D}{d}\left(1+\sqrt{1-\frac{1}{\left(D/d\right)^2}}\right)\right\}} \frac{\text{pF}}{\text{ft}} \qquad \text{(Eq A1)}$$

where:

D = center-to-center distance between rods
d = diameter of the rods.

Dimension units for D and d must be the same but can be anything.

For an 18.5-inch OWL probe with D = 4 inches and d = 0.44 inches, this gives C_0 = 4.51 pF. Of course there will also be a small additional capacitance due to end effect. A later measurement gave C_0 = 4.83 pF, which indicates that the end effect adds about 10% to the calculated capacitance. Unfortunately, there isn't a similar simple expression for the monoprobe.

Measuring C_0 poses a problem because it is so small, typically less than 10 pF. I use an inexpensive L/C meter made by Almost All Digital Electronics, model L/C meter IIB, shown in Figure A1.

This meter operates at about 1 MHz. By being very careful to zero the instrument just before a measurement and taking great care not to change the layout between zeroing and measuring, I have found that this instrument does measure small values of capacitance very well. In the case of the OWL probes I always measured a value which was just a little bit higher than calculated, which is what you would expect taking end effect into account.

A direct measurement of a probe will give a capacitance that is the sum of C_0 and the part of the probe that sticks out of the ground and is connected to the impedance analyzer. This is a parasitic capacitance (C_p), which has to be subtracted from the total measurement. I determined C_p by building a dummy probe that is identical in all respects to the actual probe except that the portion of the rod or rods that would normally be in the soil is cut off. The mechanical layout for the part sticking out of the ground is carefully replicated and a direct measurement of C_p made. This is then subtracted from the total capacitance measurement for the probe. In principle C_p is in parallel with the impedance you want to measure to determine σ and ε_r and causes a small error. In practice, C_p will be roughly the same magnitude as C_0.

Figure A1 — I used this Almost All Digital Electronics L/C meter to measure the capacitance of the probes in air. By measuring the total capacitance of the probe and leads, and then measuring the parasitic capacitance, C_p, of the part of the probe above ground along with the connecting leads, I am able to calculate the capacitance of the probe alone, C_0.

When the probe is inserted into soil, however, C_0 is multiplied by ε_r and the effective capacitance is much larger than C_p. You can modify the equations to take C_p into account but except for soils with very low ε_r I don't think it matters much.

Again, it is important to realize how small the measured capacitances are. You have to keep your body and any other conductors well away from the probe and the L/C meter. I place the probe and meter on top of a large plastic garbage can, well away from benches and other objects. I zero the meter by holding it with one stick and pushing the zero button with another, so the effect of my body is minimized.

Even with this simple and inexpensive instrument I believe I get quite accurate values for C_0. I confirmed the measurements using an HP 3577A vector network analyzer. Table A1 shows the parameters for my measurements.

C_p is in shunt with the measured impedance and might cause some error. You can, of course, modify the equations to remove this effect when C_p is known but I found that for most soils the values for the measured impedances were much lower than the shunt impedance presented by C_p and adding a correction factor was unnecessary.

Table A1

Measurement Parameters

Probe Type	*Rod Diameter (Inches)*	*Spacing (Inches)*	*Length (Inches)*	*Ground Screen (feet)*	C_o
Monoprobe	0.375	—	18	3 × 3	7.41 pF
OWL	0.44	4	18.5	—	4.83 pF
OWL	0.44	4	9.5	—	2.7 pF
OWL	0.44	3	12	—	3.4 pF

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Experimental Determination of Ground System Performance for HF Verticals Part 6 Ground Systems for Multiband Verticals

How much will the signal strength and feed point impedance change as radials are added?

The first five parts of this series have focused on ground systems for a single-band vertical (mostly on 40 m).[1, 2, 3, 4, 5] This part of the series will address multiband radial systems, and give us an opportunity to see if the performance equivalence shown earlier between a large number of radials lying on the ground and a few elevated radials will hold with a multiband radial system.

The experiments were performed in two phases. The first was for radials lying on the ground and the second was for elevated radials. These represent two typical scenarios for amateurs: in other words. "Do I put the antenna in the back yard or up on the roof?" These are quite different arrangements, so the discussion is divided into two parts, beginning with the radials lying on the ground surface and then moving on to elevated radials.

The Test Antenna

For this series of tests I used a SteppIR III vertical antenna. The SteppIR has the advantage that its height can be adjusted to be resonant anywhere between 40 m and 6 m. The height adjustment is motorized and controlled remotely, so it is very convenient for tests on multiple bands.

[1]Notes appear on page 24.

Test Frequencies

Most of the measurements were taken at spot frequencies of 7.2, 14.2, 21.2 or 28.5 MHz. I did make a limited number of measurements across each band, however, and some of those results will be discussed.

Radial System Configurations

For these experiments I made up four sets of thirty two ¼ λ radials, one set for each band (40, 20, 15 and 10 m). The radial lengths are given in Table 1 along with the corresponding free space ¼ λ. As is the usual practice, the radials are a few percent shorter than the free space ¼ λ. The radials were fabricated from AWG no. 18 stranded, insulated wire.

Table 1
Description of Radial Lengths

Frequency (MHz)	*Free-Space ¼ λ (Feet) / (Inches)*	*Radial Length (Feet)*
7.2	34.2 / 410	33
14.2	17.3 / 208	16.8
21.2	11.6 / 139	11.3
28.5	8.63 / 104	8.4

Table 2
Total Length of Wire in Each Configuration.

Configuration	*C1*	*C2 and C8*	*C3*	*C4*	*C5*	*C6*	*C7*
Total Wire (ft)	2240	280	560	1056	528	264	132

Table 3
Transmission Gain (S21) in dB for Each Configuration Relative to C2 (0 dB).

Frequency (MHz)	*C1*	*C2*	*C3*	*C4*	*C5*	*C6*	*C7*
7.2	+0.9	0.0	+0.2	+0.9	+0.4	+0.1	–3.2
14.2	+0.8	0.0	+0.3	+1.0	+0.5	–0.6	–1.8
21.2	+0.3	0.0	+0.3	+0.8	+0.2	–1.1	–2.6
28.5	–0.6	0.0	0.0	+0.4	–0.5	–1.3	–3.8

During the experiments I used several different configurations:

C1) Sets of 32 single-band radials, one set at a time. In this way I had an optimized ¼ λ vertical over a ground system of thirty two ¼ λ radials on each band. These antennas were then measured individually on the appropriate single band.

C2) Four ¼ λ radials on each band (16 total radials), connected *all at the same time*.

C3) A repeat of C2 except using eight radials for each band (32 total radials) .

C4) Thirty two 33 foot radials.

C5) Sixteen 33 foot radials.

C6) Eight 33 foot radials.

C7) Four 33 foot radials.

C8) For some elevated radial tests, I used four ¼ λ radials on each band, *one set of radials at a time*. The set of four was chosen for the particular band.

C1 and C8 were used for comparison purposes in that they represent a monoband antenna on each band. Obviously with a multiband antenna you would not run out to the antenna and change the radials whenever you changed bands! But this can give us feeling for any compromise in going from monoband to multiband verticals.

C2 represents the most common multiband ground system in general use both for elevated and ground surface radial systems, and so it was an obvious choice. I could have chosen many other possible combinations but those I did choose are at least reasonable. In particular I wanted to show that a few long radials (C6 and C7) don't work very well whether on the ground or elevated. Table 2 shows the total length of wire in each configuration.

Radials Lying on the Ground

The experimental results for radials lying on the ground are shown in Tables 3, 4 and 5. In Table 3 the values for S21 are in dB *relative* to the measured S21 value for C2 (0 dB). This was done to make it easier to compare each configuration to the de facto standard (C2).

The results for C7 show the same problem when used with a multiband vertical as shown earlier for a single band vertical — the ground loss is very high. Increasing the radial number from 4 to 32 (from C7 to C4) shows improvement.

C2 is our "standard" ground system (at least in practice) and we can see that its performance in comparison to the other configurations is quite good. It is true that individual sets of 32 radials on each band (C1) are somewhat better (except on 10 m, for which I have no explanation!) but the compromise is less than 1 dB. Even though C2 has only four radials cut for 40 m, the other twelve shorter radials seem to take up most of the slack, and we do not see the very poor performance that four radials by themselves displayed. By doubling the number of radials in C2 to eight for each band (C3), we see some improvement over C2, although it's only a fraction of a dB.

The best performer is C4, which is 0.4 to 1 dB better than C2, depending upon the band. C4, however, requires almost four times as much wire. If we cut the amount of wire in half (C5) we still have some improvement over C2 (with the exception of 10 m). C3 and C5, which use approximately the same amount of wire, behave very similarly.

In the final analysis it appears that the standard ground system (C2) works just

Figure 1 — Here is a view of the vertical with elevated radials.

Table 4
Physical Height of the Vertical for Each Frequency and Ground System Configuration.

Configuration Frequency (MHz) (Inches)	*Free Space ¼ λ (Inches)*	*C1 (Inches)*	*C2 (Inches)*	*C3 (Inches)*	*C4 (Inches)*	*C5 (Inches)*	*C6 (Inches)*	*C7*
7.2	410	391	406	394	391	386	371	369
14.2	208	201	202	201	198	199	200	201
21.2	139	137	137	137	137	137	137	138
28.5	104	103	102	102	102	102	103	104

Table 5
Measured Feed Point Impedances With the Vertical Height Adjusted for Resonance at the Test Frequency.

Configuration Frequency (MHz)	*C1 (Ohms)*	*C2 (Ohms)*	*C3 (Ohms)*	*C4 (Ohms)*	*C5 (Ohms)*	*C6 (Ohms)*	*C7 (Ohms)*
7.2	40.0	54.4	51.7	40.0	43.5	56.3	92.4
14.2	35.1	50.0	44.5	42.7	51.2	62.4	85.8
21.2	36.0	40.5	38.4	42.0	48.9	66.3	102.9
28.5	34.4	48.2	39.3	43.8	51.6	67.8	105.6

fine. You can add more wire and get some improvement but whether that improvement is worthwhile depends on the user.

As shown in Tables 4 and 5, there is some interaction between the tuning or resonant height of the vertical and the individual ground system configurations. We've seen this effect in earlier experiments. The heights shown are a bit of an approximation. The control unit display for the SteppIR gives the length of the tape (the vertical conductor) above a certain point but between that point and the actual ground radial plate there is approximately another 12 inches of wire. The wire is bent within the base housing so you can't assign an accurate additional length. I have used 12 inches as a reasonable approximation.

The measured feed point impedances are given in Table 5.

Elevated Radials

Having four sets of 32 radials (one set on each band) on hand from the ground surface tests I decided to use these same radials for the elevated radial tests. With the exception of C1and C3, I used the same configurations (C2, C4-C7) for the elevated tests. In the elevated radial testing, I used C8 in place of C1. Like C1, C8 is not practical, being a series of monoband verticals, but it serves as a reference against which to judge the compromise from using a multiband radial system. For comparisons between elevated and ground surface radials I have added a column (C1) to Tables 7 and 9 for the *on-the-ground* data associated with C1. We will use these when we discuss elevated versus ground radials.

A photograph of the experimental arrangement for the elevated radial tests is shown in Figure 1.

Because of the need for easy access to the radial base plate to make the many changes in radial configuration, I had to place the base of the antenna only 6 feet above ground.

Six feet high for the base is a bit low if we want to improve the feed point match by sloping the radials downward. In free space the input impedance of a 4-radial ground-plane antenna is about 22 Ω. As we bring the antenna closer to the ground, the impedance will vary around this number but in general is well below 50 Ω. Often the SWR will be high. One common means to improve the match is to slope the radials downward from the base, which raises the feed point impedance and lowers the SWR. Because of the limited height at the center, I could only lower the outer ends of the radials a small amount. Keep this in mind when we look at the measured impedances and SWR plots.

Experimental results are given in Tables 6, 7 and 8. A few of the columns have blanks. These are cases where that configuration, on that band, performed so poorly as to be unac-

Table 6
Transmission Gain (S21) in dB for Each Configuration Relative to C2 (0 dB).

Frequency (MHz)	*C2 (dB)*	*C4 (dB)*	*C5 (dB)*	*C6 (dB)*	*C7 (dB)*	*C8 (dB)*
7.2	0.0	–-0.1	–0.2	–0.2	0.0	0.0
14.2	0.0	+0.2	–0.8	–4.0	—	+0.2
21.2	0.0	+0.4	+0.2	+0.2	—	+0.4
28.5	0.0	+1.1	+1.8	+0.7	—	+0.2

Table 7
Physical Height of the Vertical for Each Frequency and Configuration.

Configuration Frequency (MHz)	*Free Space ¼ λ (Inches)*	*C1 (Inches)*	*C2 (Inches)*	*C4 (Inches)*	*C5 (Inches)*	*C6 (Inches)*	*C7 (Inches)*	*C8 (Inches)*
7.2	410	391	403	397	397	400	403	403
14.2	208	201	208	190	180	150	—	208
21.2	139	137	143	142	143	145	—	142
28.5	104	103	104	100	97	88	—	104

Table 8
Measured Feed Point Impedances with the Vertical Height Adjusted for Resonance.

Configuration Frequency (MHz)	*C2 (Ohms)*	*C4 (Ohms)*	*C5 (Ohms)*	*C6 (Ohms)*	*C7 (Ohms)*	*C8 (Ohms)*
7.2	43.4	42	41.0	42.1	43.0	43.0
14.2	34.2	38.9	41.1	83.9	—	33.9
21.2	36.8	52.3	49.5	48.4	—	31.4
28.5	23.9	34.8	38.3	73.2	—	24.5

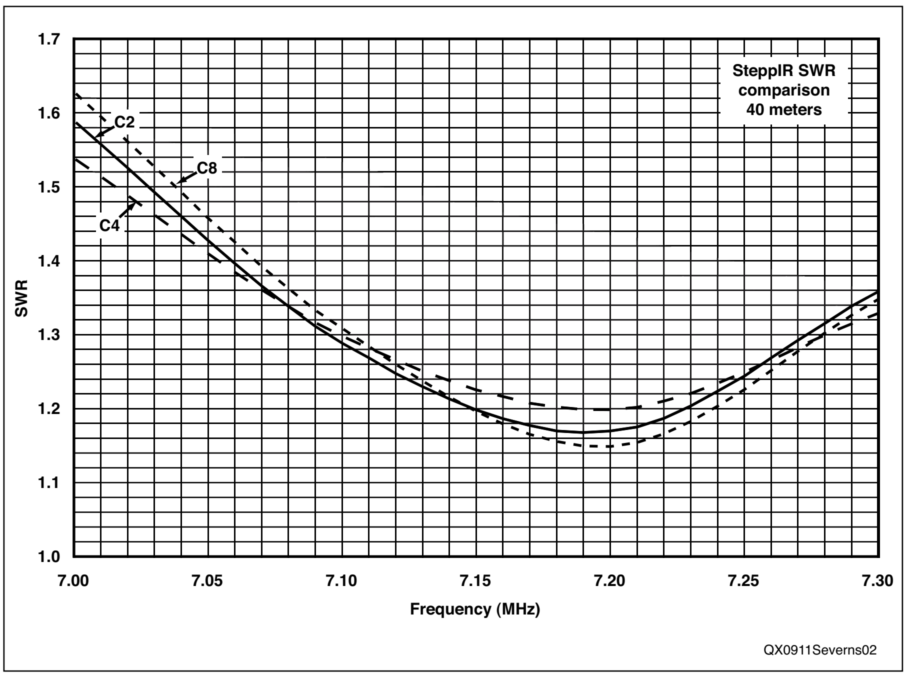

Figure 2 — Feed point SWR comparison on 40 m.

ceptable and I didn't see any point in recording that information.

From the data in Table 6, the standard multiband radial system (C2) appears to work very nearly as well as C4 and it takes only a quarter as much wire! The only band on which C4 appears to have a significant advantage is 10 m. C2 is also very close to C8 so there is very little compromise from the monoband case. As we move to fewer long radials (C5-C7) we see there is an immediate problem on 20 m, where the gain starts to fall quickly. From Table 7 we see that on 20 m the resonant height of the vertical starts to change radically as we go to fewer long radials, so clearly there is some funny business going on. This is related to the fact that ¼ λ radials on 40 m are close to ½ λ long on 20 m. Except on 20 m, C5 and C6 seem to be okay on 15 and 10 m, but by the time we get to C7 (four 33 foot radials) the performance was so poor I haven't even entered the data. The four long radials don't even work well on 15 m, where they are close to ¾ λ long.

From a loss point of view there appears to be little advantage to using anything other than the standard four radials cut for each band (C2). There is, however, the question if there is any matching (SWR) improvement from using more wire — for example C4 instead of C2. Figures 2 through 5 show a comparison of the feed point SWR between C2, C4 and C8 on the four bands.

On 40 m the differences are insignificant. On the higher bands we see little difference between C2 and C8. C2 is behaving pretty much as we would expect. However, C4 does seem to offer some improvement above 40 m. It is especially noticeable on 15 m, where the 32 radials are all near ¾ λ resonance. From some of my earlier work I was not surprised that increasing the number of radials beyond four did not give much improvement in S21, but I was expecting to see much flatter SWR curves. This just doesn't seem to happen on 40 m but does appear on 15 m with ¾ λ radials.

We should keep in mind that the feed point impedances and associated SWR will be affected by the height above ground, which in this case is very low. For well

Modifying the Ground Radial Connections on the SteppIR

Before conducting the experiments, I modified the ground radial connection on the standard SteppIR and also made up a special feed line choke that would have an impedance greater than 1000 Ω on all bands.

As the SteppIR comes from the manufacturer, it has a single no. 12 brass machine screw to which the ground radials can be attached. I felt this was not adequate and certainly not very convenient for the many radial changes necessary during the experiments. I changed the single brass no. 12 screw to a pair of ¼-20 machine screws spaced about 6 inches apart, as shown in Figure 1A.

I then fabricated an aluminum disk with fifteen ¼-20 bolts with wing-nuts around its perimeter. The disk was attached to the base of the SteppIR housing as shown in Figure 2A.

For all the measurements in the experiments, but particularly for the elevated radial measurements, I wanted to have a common mode choke (balun) in the feed line and the cabling at the base of the antenna. The choke I used is shown in Figure 3A. The choke has 6 turns of RG8X coaxial cable wound on two stacked type 43 cores (Fair-Rite #2643803802, available from Mouser Electronics). Also shown in the picture is the probe from the HP4815A vector impedance meter used for impedance measurements. The measured shunt impedance was between 2 and 3 kΩ from 7 through 30 MHz.

Figure 4A shows both the chokes installed at the base of the antenna.

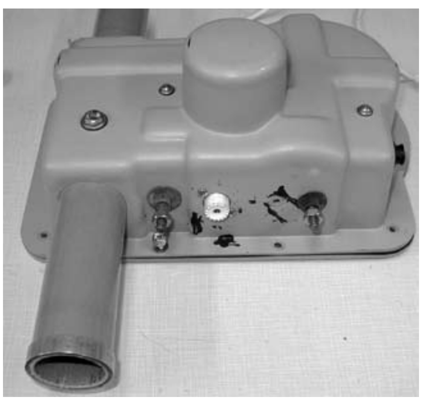

Figure 1A — Modified radial attachment scheme for the SteppIR.

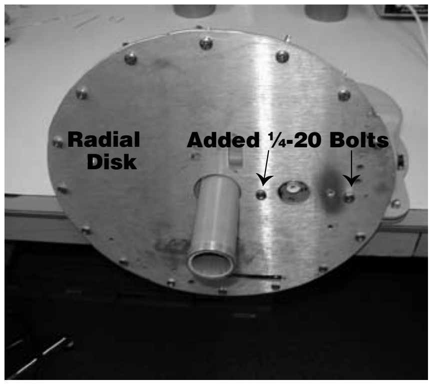

Figure 2A — SteppIR with radial disk attached.

Figure 3A — Common mode choke for the feed line.

Figure 4A — Chokes installed in the feed line and control cables at the base of the antenna.

elevated radials, where the slope can be adjusted to provide a better match, the results may be much better than shown here.

Elevated Versus Ground Radials

Another key question is "How do the elevated radial systems compare to a large number of radials on the ground on each band?" Table 9 makes that comparison using the results from this series of experiments. C1 is used as the reference (0.0 dB).

C1 uses radials lying on the ground surface and C2, C4 and C8 are elevated. When we compare the signals for C1 to those for C8, which is a direct comparison between four elevated radials against 32 ground surface radials, one band at a time, we see only small differences: four elevated radials seem to perform much the same as large numbers of ground surface radials. This is in keeping with what we saw in Part 3, only now extended to bands from 40 through 10 meters.

C4 (which is thirty two elevated 33 foot radials) is also only marginally different from C1 and C2 except on 10 m, where the difference is 1.4 dB. Considering it has four times the wire, I doubt it's justified.

Some Final Comments

In summary, I don't see any compelling reason to use more than four radials on each band for a multiband vertical. The "standard" system (C2) does in fact seem to work well. If you want to lay out or hang up more wire, you can get some small improvement but generally the maximum improvement seems to be on the order of 1 dB or less, although the improvement might be somewhat higher over poorer soil than mine. In a way, this was a bit of a disappointment. It would have been nice to discover some magic new ground system for multiband verticals, but that was not to be. All I've really accomplished is to show that the old standard works just fine, and it appears that a few elevated radials can work as well as a large number of on-the-ground radials! Be careful, however! As I pointed out earlier in the series, elevated monoband radial systems with only a few radials are very susceptible to local effects that can cause unequal radial currents, which can degrade performance.

Keep in mind when comparing the data in this part with some of the data reported in earlier parts of this series, that this set of measurements were made in mid-summer when the temperature had been 85° and 108° F over the preceding month. The soil will have dried out considerably compared to that for most of the earlier experiments. This can cause the impedance and S21 measurements to vary substantially between seemingly identical experiments. This is why I emphasized in Part 1 the need to do all com-

Table 9
Transmission Gain (S21) in dB for Each Configuration Relative to C1 (0 dB).

Frequency (MHz)	*C1 (dB)*	*C2 (dB)*	*C4 (dB)*	*C8 (dB)*
7.2	0.0	+0.2	+0.1	+0.1
14.2	0.0	+0.1	+0.3	+0.3
21.2	0.0	–0.5	+0.4	–0.1
28.5	0.0	–0.3	+1.1	–0.1

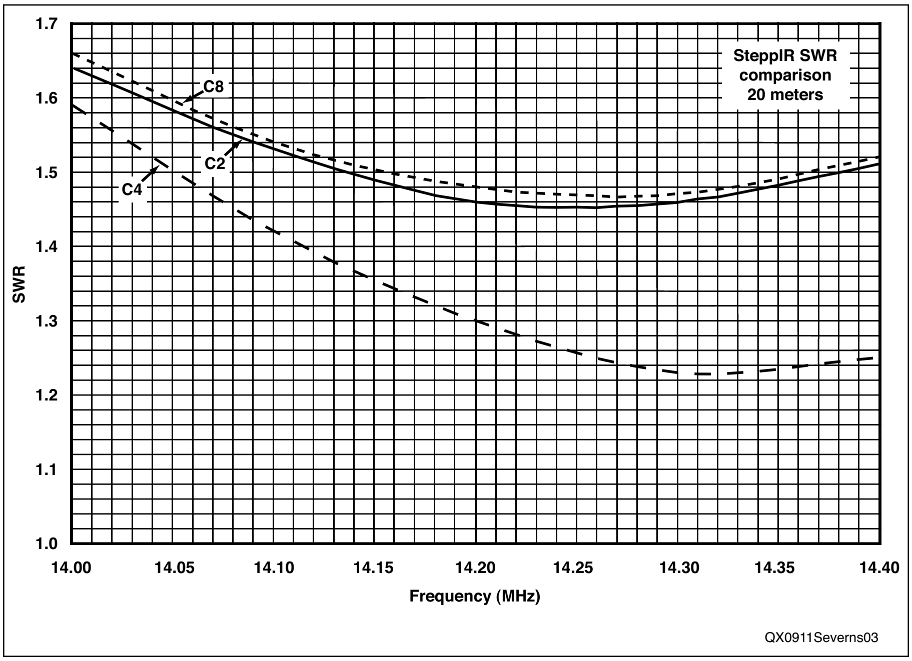

Figure 3 — Feed point SWR comparison on 20 m.

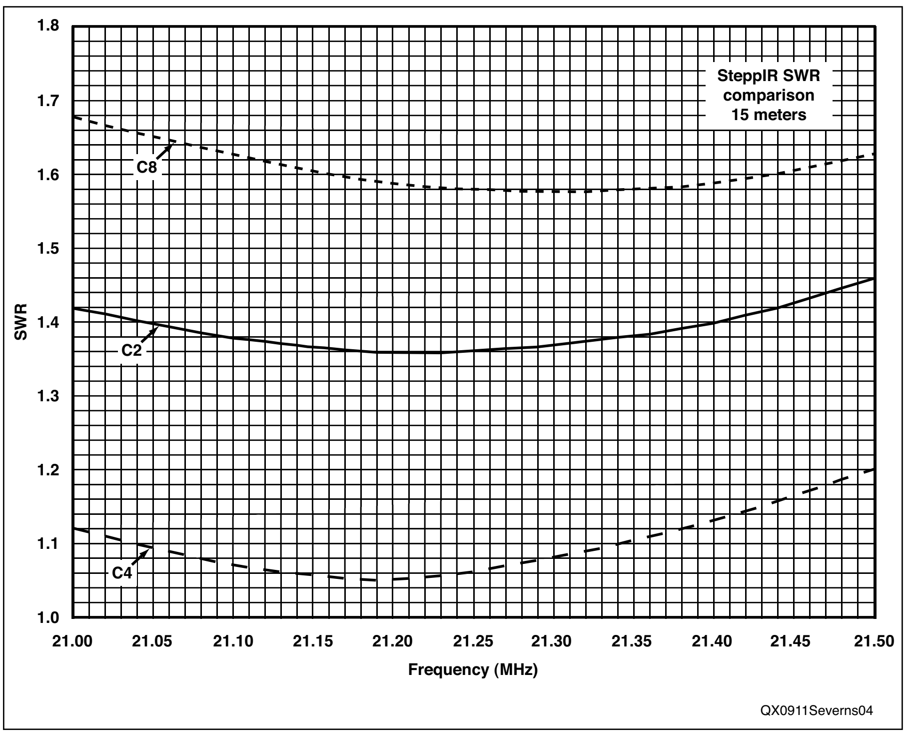

Figure 4 — Feed point SWR comparison on 15 m.

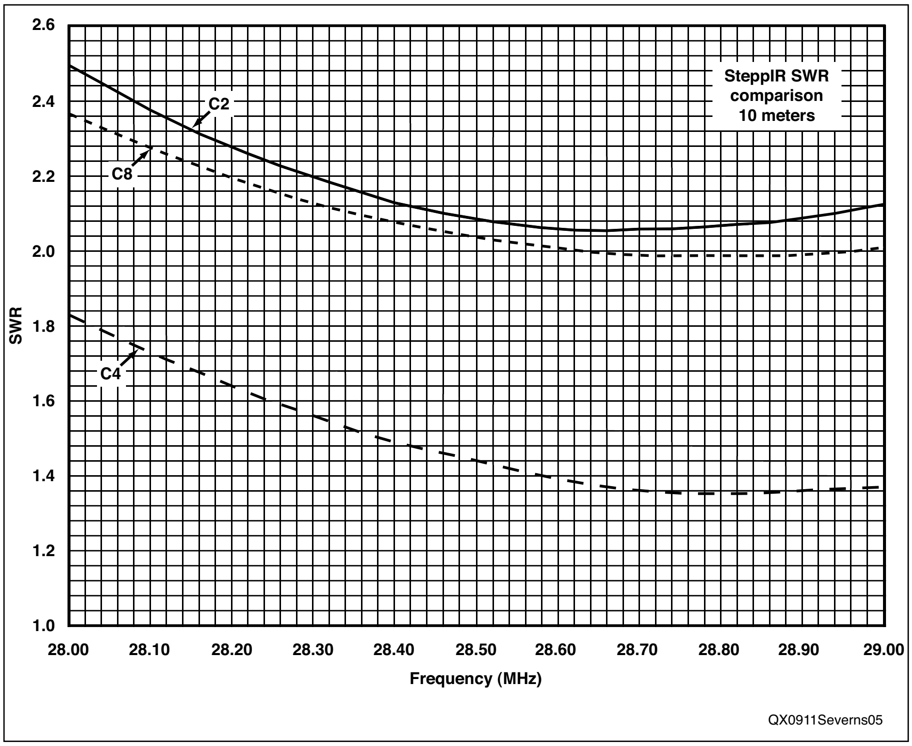

Figure 5 — Feed point SWR comparison on 10 m.

parison experiments in as short a time interval as possible. This sensitivity to changes in ground characteristics is also the reason I have emphasized that the specific numbers derived from these experiments must *not* be taken as absolutes. They are intended only to show the trends in performance between different ground systems. In addition, the frequency range in this series of tests goes much higher than those for the earlier experiments. The soil characteristics at a given location and time will vary with frequency.[6] In other words, your mileage may vary!

Despite the extensive experimental work reported in this series there will still be many unanswered questions regarding ground systems for verticals. Answers will have to be deferred to future experiments and computer modeling. Hopefully, others will be inclined to join in this effort making their own contributions. Of course not all questions have to be answered experimentally. As some of this work has indicated, *NEC* modeling can shed a lot of light on many questions, although in the end it's always more convincing if there is at least some experimental confirmation.

Acknowledgement

I would like to express my appreciation to Mike Mertel, K7IR, for the loan of the SteppIR vertical antenna used in these experiments. That antenna made the experiments much easier. I would also like to thank Mark Perrin, N7MQ, for his help at some key points in the experiments, when another hand was really helpful.

Rudy Severns, N6LF, was first licensed as WN7WAG in 1954 and has held an Extra class license since 1959. He is a consultant in the design of power electronics, magnetic components and power-conversion equipment. Rudy holds a BSE degree from the University of California at Los Angeles. He is the author of two books and over 80 technical papers. Rudy is an ARRL Member, and also an IEEE Fellow.

Notes

[1] Rudy Severns, N6LF, "Experimental Determination of Ground System Performance - Part 1," *QEX*, Jan/Feb 2009, pp 21-25.

[2] Rudy Severns, N6LF, "Experimental Determination of Ground System Performance - Part 2," *QEX*, Jan/Feb 2009, pp 48-52.

[3] Rudy Severns, N6LF, "Experimental Determination of Ground System Performance - Part 3," *QEX*, Mar/Apr 2009, pp 29-32.

[4] Rudy Severns, N6LF, "Experimental Determination of Ground System Performance - Part 4," *QEX*, May/June 2009, pp 38-42.

[5] Rudy Severns, N6LF, "Experimental Determination of Ground System Performance - Part 5," *QEX*, Jul/Aug 2009, pp 15-17.

[6] Rudy Severns, N6LF, "Measurement of Soil electrical Parameters at HF," *QEX*, Nov/Dec 2006, pp 3-9.

QEX

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Determination of Soil Electrical Characteristics Using a Low Dipole

N6LF shows how to create a universal chart showing antenna impedance values for a wide range of soils that map to the average values of σ and Er for the soil over which the antenna is installed.

Rick Karlquist, N6RK, asked on the top-band reflector about placing a dipole on the ground surface to derive soil electrical characteristics — conductivity (σ) and relative dielectric constant (*Er*) — from impedance measurements of the dipole. A short discussion of this technique has appeared in the last few editions of *The ARRL Antenna Book*.[1] For some years I've used the ground probe approach[2] to measure soil characteristics so I hadn't paid much attention, but in some situations this method may have advantages over the soil probes and is worth considering. The probe approach gives the values for a small volume of soil around the probe, down to a depth of 3 ft or so. If you want to map the properties of a large area you need to make multiple measurements at different locations. The low-dipole approach on the other hand intrinsically averages the properties of a much larger area below the antenna and for a couple of skin depths down into the soil. The *ARRL Antenna Book* discussion was pretty limited so I decided to expand on it using antenna modeling software combined with a spreadsheet.

If you have a program that accurately models the soil-antenna interaction (such as NEC4) then you can use the antenna of your choice at whatever frequency you are interested in, see Example 2. Most amateurs don't have this software but the technique can still be used. With some prompting from Rick, N6RK, I realized that if the antenna dimensions — length, height, wire size, etc. — and measurement frequency are predefined then it is possible to create a universal chart with contours showing values of *Ri* and *Xi* for a wide range of soils. If the antenna is fabricated as specified, and impedance is measured at the specified frequency, the measured impedance can be plotted directly on the graph yielding a good estimate of the average values of σ and *Er* for the soil over which the antenna is installed. As a practical matter the reference antenna needs to be something easy and inexpensive to build. For that purpose a low dipole works well, and details of a suggested design are given in Example 1. From a practical point of view it is necessary to have a predefined antenna for each band. In this article I've chosen 80 m for demonstration purposes.

What frequency, lengths and heights?

The height above ground *z* and test antenna length *L* will depend on the frequency of interest. At what frequency within the band should we make the measurement or do we need to measure across the band? Figures 1 and 2 show examples of actual measured values for σ and *Er* at my home site using soil probes.

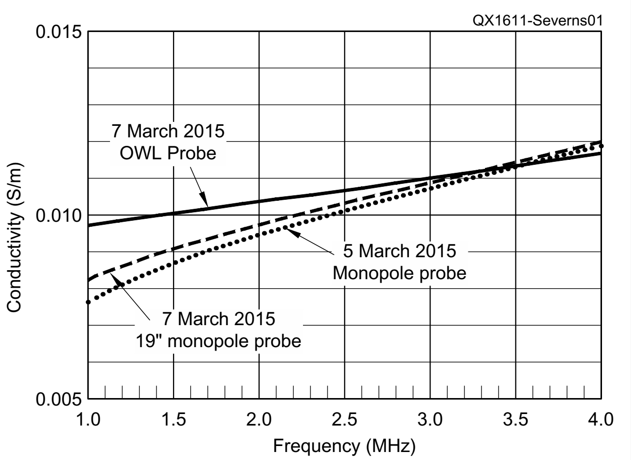

Figure 1 — Soil conductivity σ at N6LF.

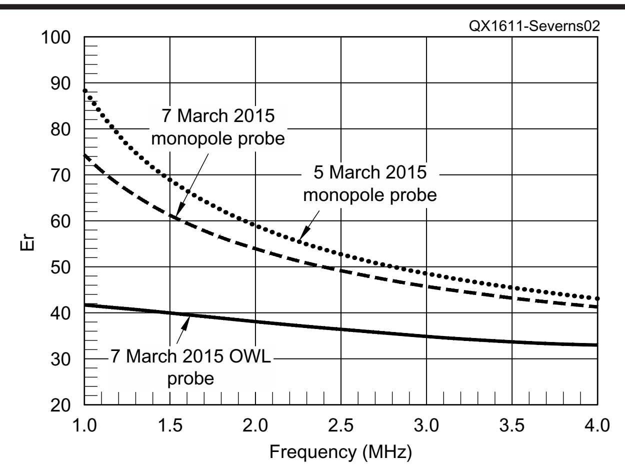

Figure 2 — Soil relative permittivity *Er* at N6LF.

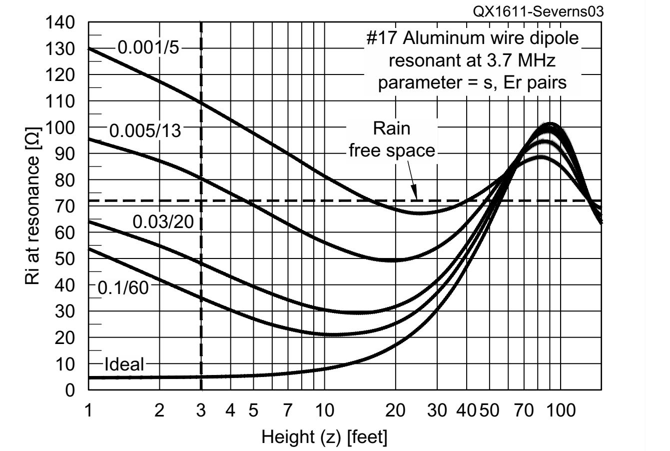

Figure 3 — *Ri* at resonance versus *z* for typical σ, *Er* pairs.

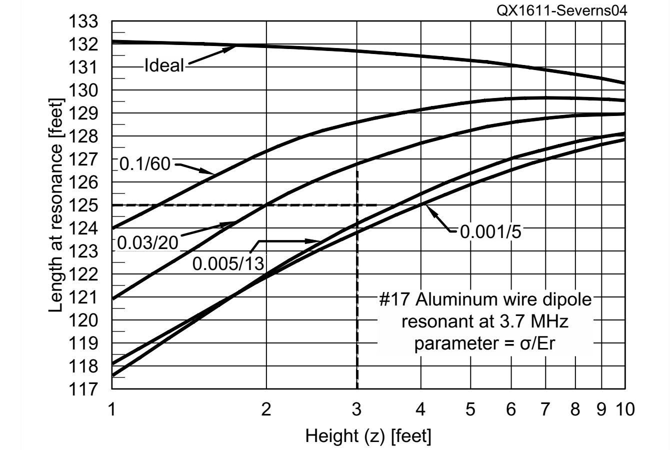

Figure 4 — Effect of height and ground constants on resonant length at 3.7 MHz.

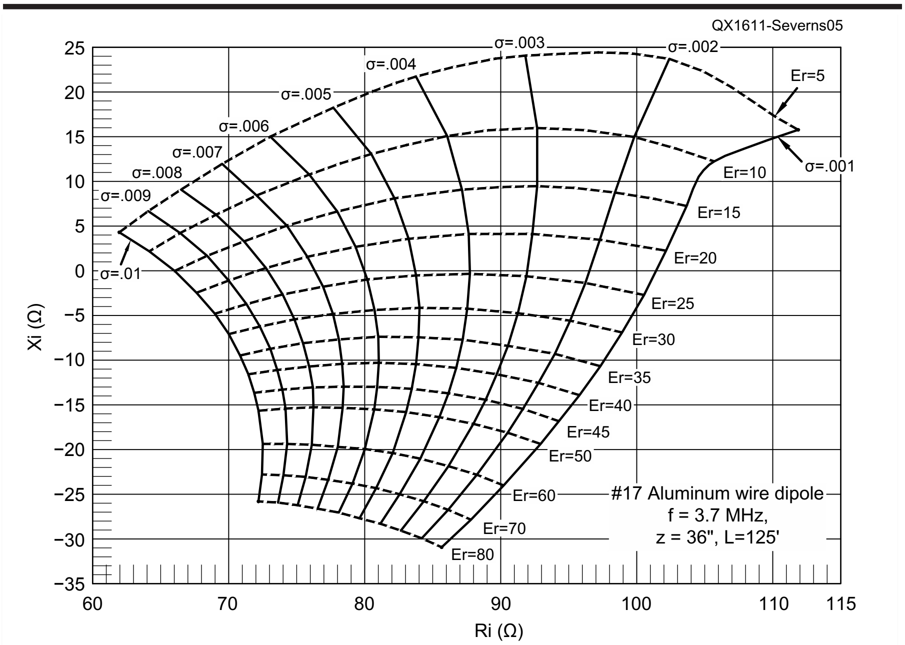

Figure 5 — *Xi* versus *Ri* for 0.001<σ<0.01 and 5<*Er*<80.

Over the 80 m band (3.5-4.0 MHz), conductivity is 0.011<σ<0.0.012 S/m and relative permittivity is 41<*Er*<43. This is a pretty small range and a measurement near mid-band, say 3.7 MHz, should be more than accurate enough. Remember, we are not trying for 1% accuracy, ±20% will do just fine. The modest change of values shown over the 80 m band is typical of most soils. Other bands are much narrower in percentage of center frequency so the changes are even smaller. A single frequency measurement is adequate for each band.

Strictly speaking, the test antenna does not have to be resonant but there are practical measurement advantages to not being too far from resonance. As you move away from resonance the values for *Ri* and *Xi* will begin to change fairly rapidly. Many of the instruments used to measure impedance don't handle very well impedances less than 10 Ω or greater than a few hundred ohms. The impedance values are smaller close to series resonance.

The next question is "how high"? Figure 3 shows the effect of various soils (typical σ and *Er* pairs) at a range of heights when the antenna is tuned to resonance at each point. For heights between 1 and 10 ft the contours are well separated, promising reasonable resolution for variations in σ and *Er*. However, at greater heights the contours begin to tighten up making resolution a problem. It looks like any height *z* between 1 and 10 ft should work. I chose 36 in because it's a very convenient working height. Since standard electric fence hardware is well suited for this kind of field measurement, 36 in corresponds to a standard insulated electric fence post — a practical detail passed to me by N6RK.

For a given height and resonant frequency, the resonant length will depend on the values for the ground constants as shown in Figure 4. For calculations at 3.7 MHz with *z*=36 in, *L*=125 ft is a reasonable compromise.

A universal graph for 80 m

If we have a physical description of the antenna in terms of height above ground *z*, length *L*, wire size, etc., we can model the antenna at a single frequency *f* using a wide range of values for σ and *Er*. This will give us values for the feed-point impedance *Zi*=*Ri*+*jXi* at a given frequency for each pair of σ and *Er* values. Using a spreadsheet we can then graph *Ri* versus *Xi* — which are the quantities we can actually measure on a test antenna — as functions of σ and *Er*, with *Ri* on the *x*-axis and *Xi* on the *y*-axis, where σ and *Er* are parameters defining the contours. After measuring the feed-point impedance at *f* we can plot the measured *Ri* and *Xi* pair as a point on the graph. I used *EZNEC pro*[3] with NEC4.2 and an *Excel*® spreadsheet software, *AutoEZ*[4], to automate the calculations and graph them. From earlier work I did on verifying the accuracy of NEC4 for wires close to ground I found that the fitting at the feed point has a shunt capacitance of about 6 pF. This has been added to the model.

With *L*=125 ft, *z*=36 in and *f*=3.7 MHz we graph *Xi* versus *Ri* as functions of σ and *Er* (Figures 5 and 6). The dashed contours represent 5<*Er*<80 and the solid contours represent 0.001<σ<0.01 S/m (Figure 5), and 0.01<σ<0.03 S/m (Figure 6). This range of values should cover most common soils that amateurs are likely to encounter. If this doesn't work for your site then you can use the

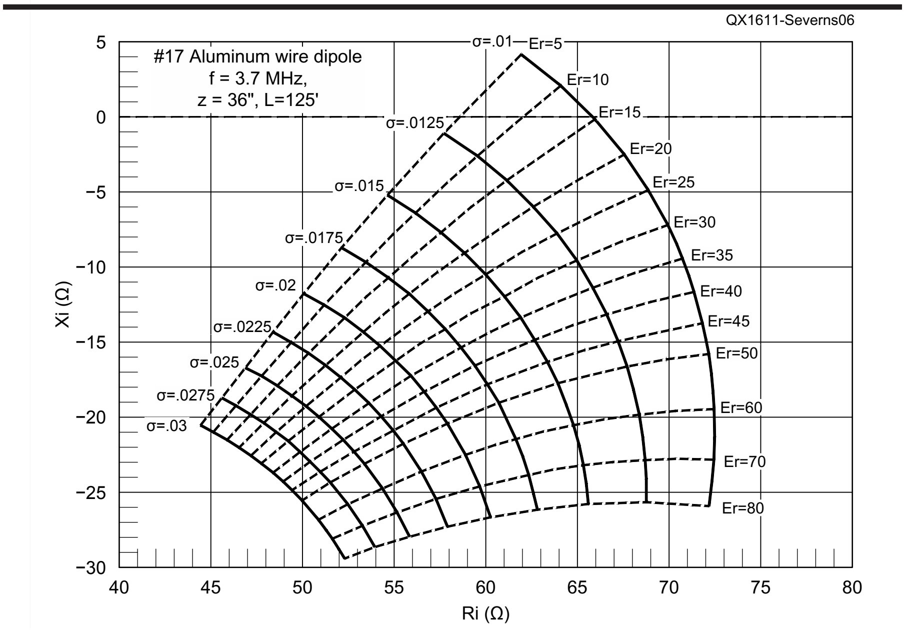

Figure 6 — *Xi* versus *Ri* for 0.01<σ<0.03 and 5<*Er*<80.

Figure 7 — Center connector and feed-point support. [Rudy Severns, N6LF, photo.]

procedure described in Example 2 to generate your own graph using NEC4 software.

Note that I've cut Figure 6 off for σ greater than 0.03 S/m. As the conductivity increases the scale compresses rapidly. In fact if we push σ all the way to infinity (perfectly conducting soil) *Zi* converges to a single point at *Zi*=4.2-*j*76.5 Ω. Most amateurs are not blessed with soil of this high conductivity so this limitation is not that serious. For higher conductivity soils ground probe measurements are probably a better method.

Example 1

Figures 7 and 8 are photos of the mechanical arrangements for typical test antenna using standard #17 AWG aluminum electric fence wire and hardware widely available in hardware and farm stores. The electric fence wire is suspended at 36 in on fiberglass (F/G) wands, with yellow plastic wire clips that slide up/down the wands for height adjustment. The wands were spaced 10 to 20 ft apart and the wire is anchored at the ends to steel fence posts 6 to 10 ft away from the ends of the wire. Multiple support points and significant wire tension kept the droop to less than 0.25 in. High quality insulators and non-conducting Dacron line were used at the wire ends. Figure 7 shows the Budwig center connecter and the common mode choke (balun) at the feed-point. The center connector and choke introduce approximately 6 pF of shunt capacitance across the feed point, which must be added to the model. The steel fence post at the midpoint shown in Figure 8 was replaced with the F/G wand shown in Figure 7.

The measured impedance of the common mode choke is shown in Figure 9. The choke comprises two Fair-Rite 2631665702 type 31 cores taped together to form a binocular core. The winding is six turns of RG174/U 50 Ω mini-coax.

Example 2

If NEC4 based software is available then you can create your own charts using your choice of antenna, as follows. We assume a horizontal center-fed dipole made with #17 AWG aluminum wire at a height *z* of 36 in. After tuning to resonance at 3.5 MHz the length *L* is 131.11 ft. The measured feed-point impedance *Zi* at 3.5 MHz is 80.26+*j*0 Ω. From this we can determine the values for σ and *Er* at 3.5 MHz. First create the NEC4 model using #17 AWG aluminum wire 131.11 ft long and 36 in above ground. Since we do not know the values for σ or *Er*, we'll run the model repeatedly with a range of possible values for σ and *Er*. If we're too far off in our choice of values the process should point the way to go. In this case the trial values will be 0.001<σ<.0.01 S/m and 1<*Er*<50. Running the model repeatedly, we can determine *Zi* for a matrix of σ and *Er* values. A spreadsheet, sample included in the *QEXfiles*, is a good way to keep track of results.[5]

Using the spreadsheet we can graph a more restricted set shown in Figure 10. The measured value of *Zi* for the antenna at 3.5 MHz is 80.26+*j*0 Ω. A dot with a label has been placed at that value on the graph. We see our matrix of values has bracketed this value nicely. The σ=0.005 S/m line passes right through *Zi*. Also, *Zi* lies between the *Er*=10 and *Er*=15 lines, right around *Er*=13. We could repeat the process for multiple values of *Er* around 13 to refine the answer further, but from a practical point of view we're already close enough. With σ=0.005 S/m and *Er*=13, we have average soil.

Figure 8 — Test antenna supported with F/G wands. [Rudy Severns, N6LF, photo.]

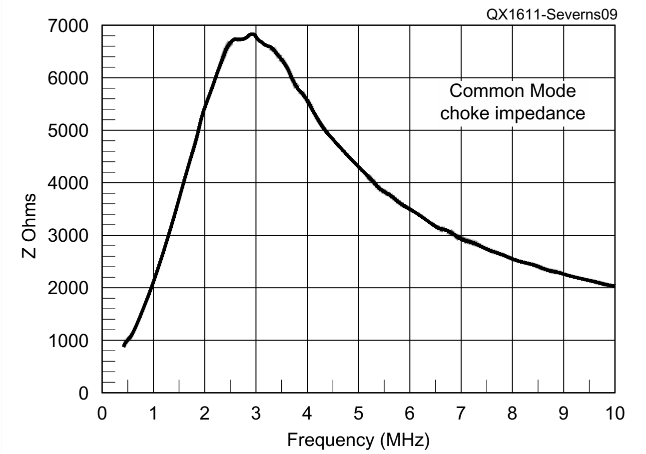

Figure 9 — Common mode choke impedance.

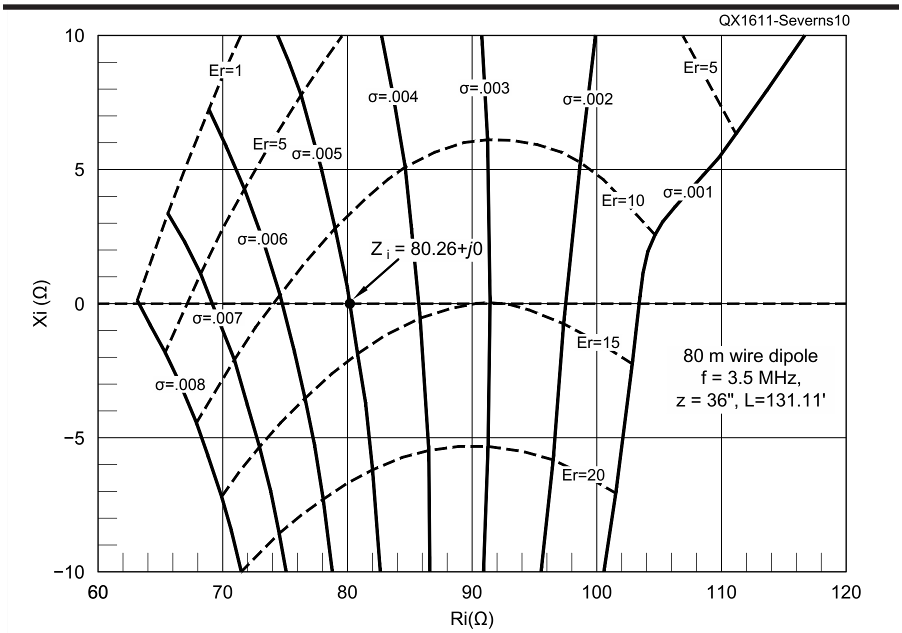

Figure 10 — Graph of *Ri* versus *Xi* for a range of σ and *Er* values at 3.5 MHz.

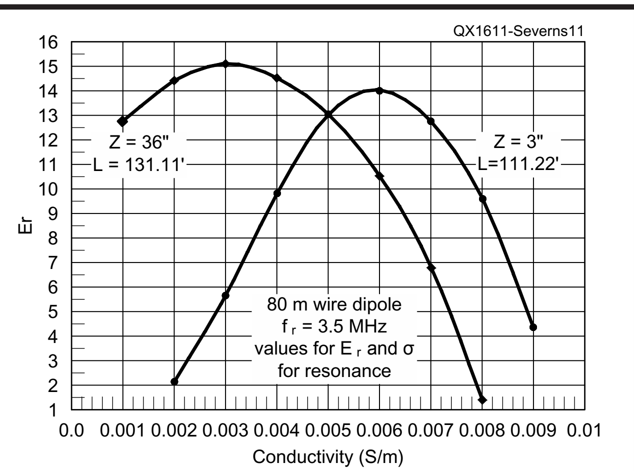

Figure 11 — Values for σ and *Er* that result in the resonant lengths shown at 3.5 MHz.

Example 3

If you have the requisite modeling software but not the impedance measuring equipment it is possible to determine σ and *Er* by resonating the antenna at a given frequency at two different heights and then, modeling these two configurations — trying different *L* and *z* — and graphing the values for σ and *Er* that correspond to the same resonant frequency. Figure 11 shows the procedure. Here *f*=3.5 MHz, and at *z*=3 in length *L*=111.11 ft, while at *z*=36 in *L*=131.11 ft. The two curves intersect at σ=0.005 S/m and *Er*=13.

Summary

There are several ways to use a low dipole to determine soil electrical characteristics. However, you will need either NEC4 software or a good impedance measuring instrument or both to do this. The ground probe method does not rely on modeling but it does require a reasonably good impedance measuring instrument capable of showing *R* and *X* as well as the sign of *X*. Low dipole measurements have the advantage of giving a realistic average of the soil characteristics over a substantial area and down a few skin depths into the soil. Ground probe measurements generally give the characteristics over a small volume of soil, and multiple measurements are required to cover a large area. Each has advantages and limitations but both will work.

Acknowledgements

I would like to acknowledge the help that Rick Karlquist, N6RK, provided by reading my early drafts and asking many questions that needed some thought to answer. My thanks to George Cutsogeorge, W2VJN, for relaying Ricks original top-band query.

Rudy Severns, N6LF, was first licensed as WN7AWG in 1954. He is a retired electrical engineer, an IEEE Fellow and ARRL Life Member.

Notes

[1]Pages 3-31 to 3-33 in, *The ARRL Antenna Book*, 22nd edition, 2011. Available from your ARRL dealer or the ARRL Bookstore, ARRL item no. 6948. Telephone 860-594-0355, or toll-free in the US 888-277-5289; **www.arrl.org/shop**; **pubsales@arrl.org**.

[2]R. Severns, N6LF, "Measurement of Soil Electrical Parameters at HF", *QEX* Nov/Dec 2006, pp 3-9. Available at **www.antennasbyn6lf.com**.

[3]Several versions of *EZNEC* antenna modeling software are available from developer Roy Lewallen, W7EL, at **www.eznec.com**.

[4]*AutoEZ* for *EZNEC*, see **www.ac6la.com**.

[5]See **www.arrl.org/qexfiles**.

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Radiation and Ground Loss Resistances In LF, MF and HF Verticals; Part 2

With the impending FCC announcement about the release of a new LF and a new MF band, hams will be interested in practical antennas and learning how to calculate EIRP to legally operate on those bands.

Soil-Antenna Interaction

As illustrated in Figure 11, one way to analyze a vertical antenna over ground is to use a hypothetical image. If the ground is perfect then the image antenna will be a duplicate of the actual antenna with the same current amplitude and phase. For a dipole a short distance above ground, the image is another dipole the same distance below ground. We now have a system of two coupled dipoles and it's no surprise that R_i of the upper dipole is no longer ≈ 72 Ω, but in these examples R_i ≈ 94 – 100 Ω. What's happening is that the upper vertical (the real one) has a self resistance of ≈ 72 Ω, but added to that is a mutual resistance (R_m) coupled from the image antenna.

If the ground is not perfect, however, then the image antenna will not be an exact replica of the real antenna. The current amplitude and phase on the image will be different, so we should not be surprised if R_i does not have the same value as either the free space or perfect ground cases. Viewing R_i as a combination of the free space value and some mutual ± R_m because of the soil is perfectly valid, and this was Wait's approach in *Antenna Theory*.[8] He calculated the ±Δ R_i as the soil and/or radial fan is changed. This ±Δ R_i was a combination of changes in R_r and R_g, however, and not R_g alone.

[8]Notes appear on page 28

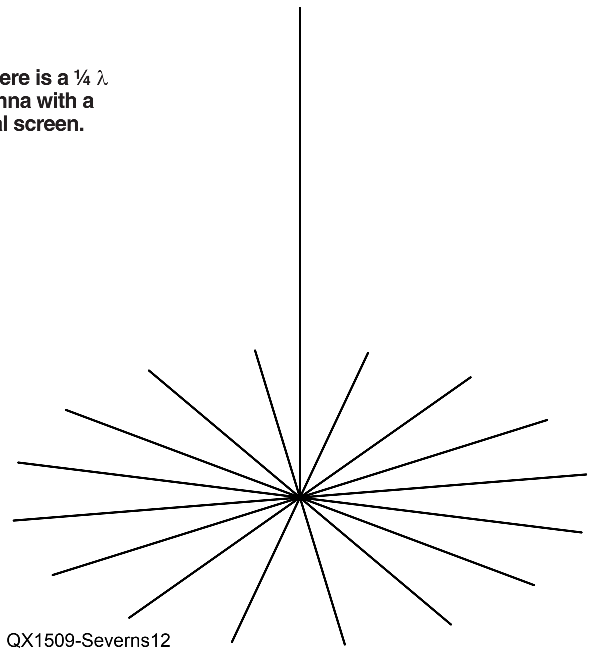

Figure 12 — Here is a ¼ λ vertical antenna with a buried radial screen.

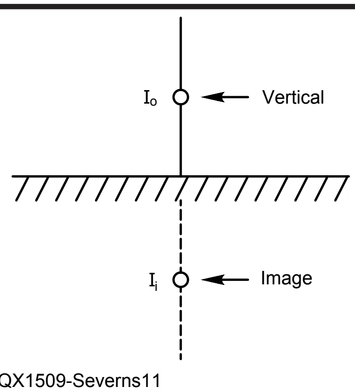

Figure 11 — This is an example of an antenna and its image.

R_r and R_g for a ¼ λ Vertical Antenna at 7.2 MHz

The ¼ λ vertical antenna with a buried radial screen shown in Figure 12 is more representative of typical amateur antennas for 40 m than a full-height ½ λ vertical dipole. Amateurs are not likely to use a full

¼ λ vertical antenna on 630 m, however. Such an antenna would be ≈ 500 feet high! We'll look at a more typical 630 m antenna in a later section.

I calculated data points for 16, 32, and 64 radials, with lengths of 2, 5, 10, and 16 m over poor (0.001/5), average (0.005/13) and very good (0.03/20) soils. Figure 13 is a graph showing the behavior of R_i, R_r, and R_g as a function of radial length when 64 radials are employed over average ground at 7.2 MHz.

On the graph there is a dashed line labeled "36 Ω" corresponding to the value of R_r for a resonant ¼ λ vertical antenna over infinite perfect ground.

The fact that R_i does not decrease or even flatten out for radial lengths > ¼ λ but instead starts to increase has been predicted analytically (for example in Wait — see note 8), my earlier *NEC* modeling (see Appendix D) and as seen in practice. (Note: Appendices A, B, C, and D are available for download from the ARRL *QEX* files website.[9]) What's interesting is that $R_r \neq 36\ \Omega$! R_r starts out well below the value for an infinite perfect ground plane, but as the radial length is extended it approaches 36 Ω. Increasing the radial number and/or extending radial length also moves R_r closer to 36 Ω. Figure 13 represents only one case: 64 radials over average ground.

Figure 14 gives a broader view of the behavior of R_r for different soils and radial

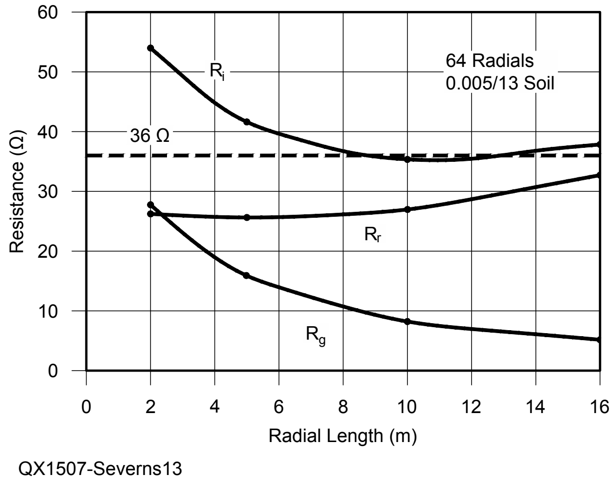

Figure 13 — This graph shows R_i, R_r, and R_g as a function of radial length for a 40 m ¼ λ vertical antenna.

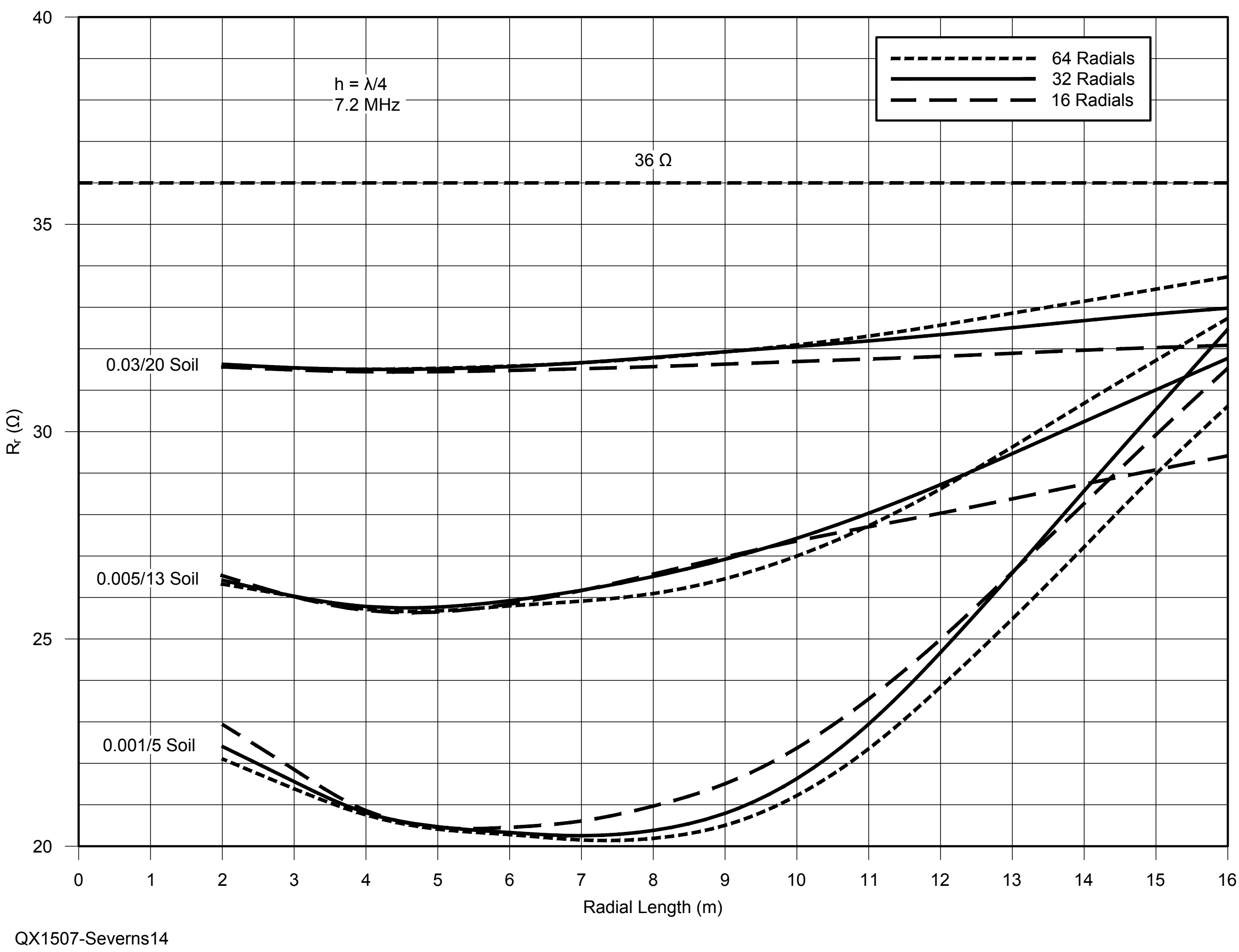

Figure 14 — Here is a graph that plots R_r as a function of soil, radial number, and radial length.

numbers as radial length is varied.

It's abundantly clear that $R_r \neq 36\ \Omega$ but as we improve the soil conductivity and/or increase the number and/or length of the radials R_r converges on 36 Ω. We can also graph the values for R_g as shown in Figure 15, which nicely illustrates how more numerous and longer radials reduce ground losses.

For a given model, *NEC* will give us R_i, G_a, and the field data from which we can determine R_r using the Poynting vector and a spreadsheet. With this information we can have some fun! R_r / R_i is the radiation efficiency, including only the ground losses within the radius of integration, which in this case is ≈ ½ λ. G_a is the radiation efficiency including all the losses, near and far field. The ratio G_a / (R_r / R_i) gives us the loss in the far field, separate from the near field losses. Figure 16 graphs all three, G_a, R_r / R_i, and G_a / (R_r / R_i) with various numbers of radials over average ground. Note that the far field loss is almost independent of the radial

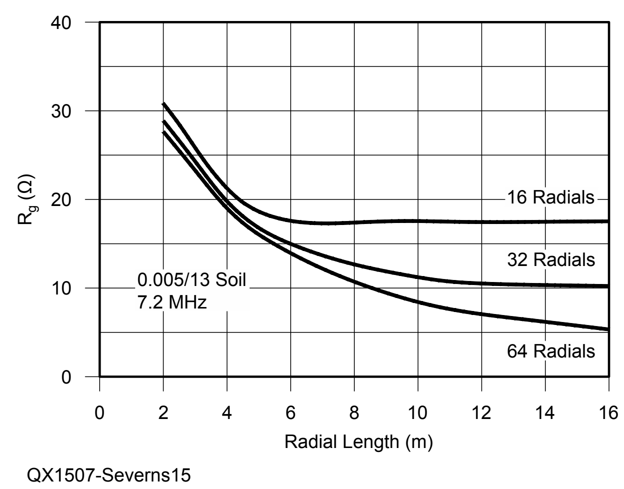

Figure 15 — This graph plots R_g as a function of radial length and number.

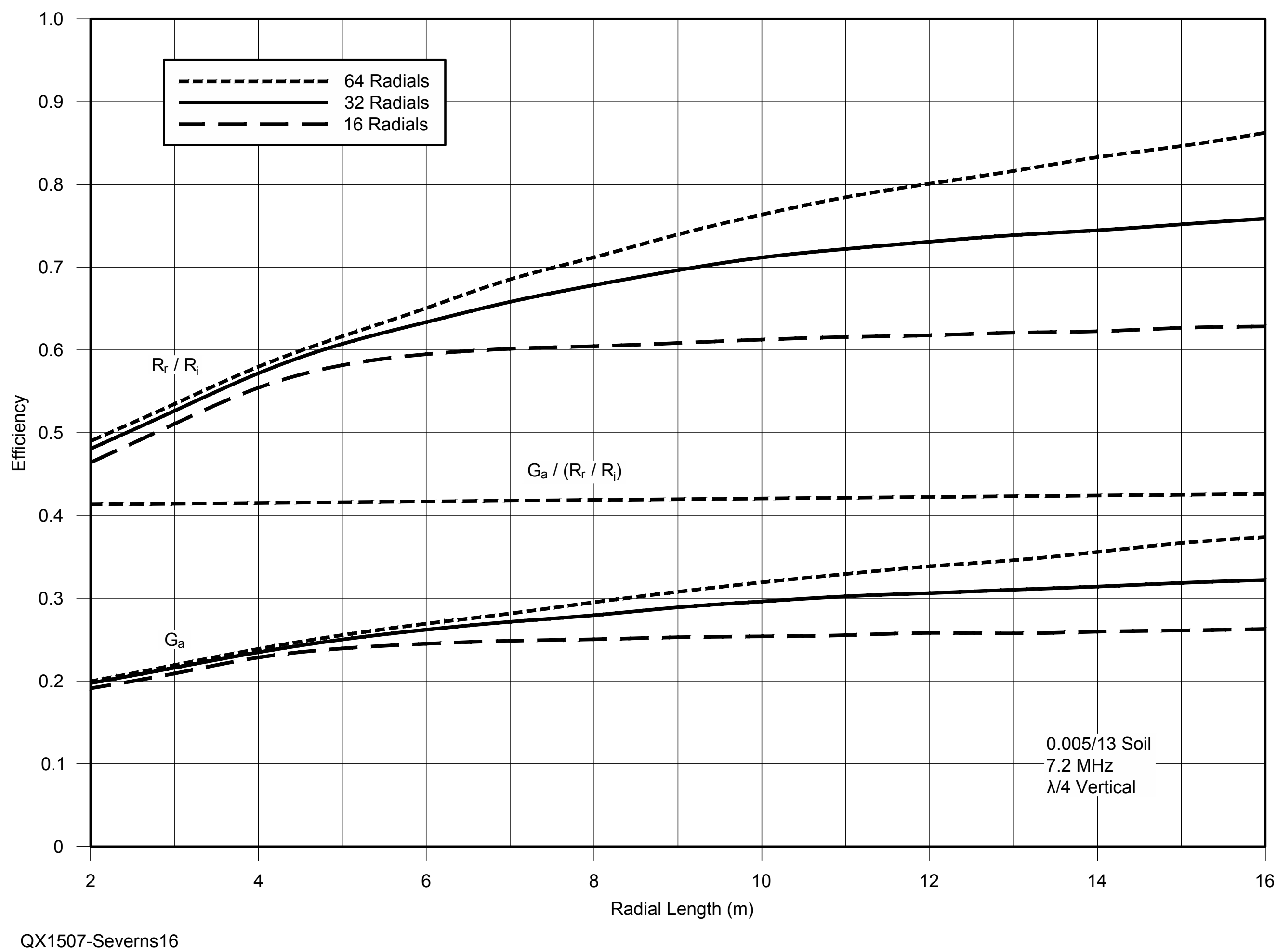

Figure 16 — Here is a graph of antenna efficiency as a function of radial length and number.

number or radial lengths, which is what you would expect because we haven't changed anything in the far field as we modified the radials. In fact any bumps or anomalies in that graph would indicate a screw-up in the calculations! It serves as a much needed cross check on the calculations.

After seeing Figure 16, Steve Stearns, K6OIK, suggested adding a graph of (R_i / R_g) – G_a, which is the ground wave radiation efficiency. This is shown in Figure 17.

By repeating the calculations for a ¼ λ vertical at 1.8 MHz, we can compare the results to expose the effect of frequency on R_i, R_r, and R_g for the same type of antenna. An example is given in Figure 18. The solid lines are for 1.8 MHz and the dashed lines 7.2 MHz.

What we see is that even though both antennas are ¼ λ, with the same length radials (in λ) and the same soil characteristic, the values for R_i, R_r, and R_g are substantially different. At 1.8 MHz, R_r is much closer to 36 Ω. Using ¼ λ radials at 7.2 MHz and integrating the radiated power, $R_g \approx 8$ Ω. If you subtracted the R_i value given by *NEC* from 36 Ω, however, you would think R_g was essentially zero! At 1.8 MHz, $R_g = 36 - R_i \approx 2$ Ω, which seems reasonable. The power integration for the 1.8 MHz vertical gives $R_g \approx 6$ Ω, however, which means the efficiency is lower than we thought. As the soil conductivity (σ) increases, the values for R_r move closer to 36 Ω. If we lower the frequency to the lower AM broadcast band (say 600 kHz) using a ¼ λ vertical with 120 0.4 λ radials, R_r will be very close to 36 Ω. This is a frequency range where a great deal of profession work has been done, which might explain why the discrepancy between estimated and actual R_g and R_r went unnoticed. The difference would be very small, easily within the range of experimental error!

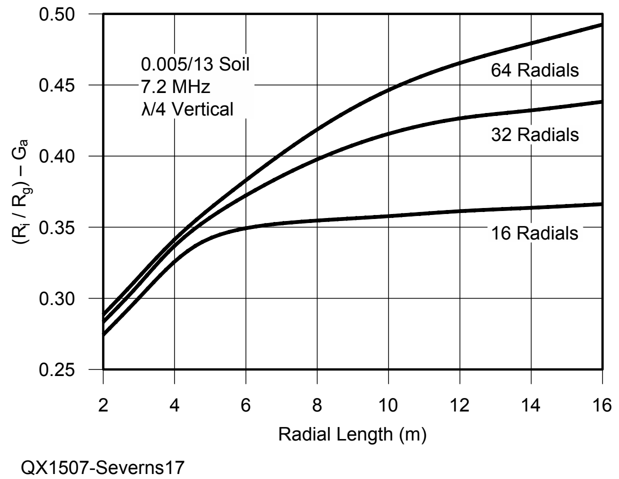

Figure 17 — This graph plots ground wave radiation efficiency; (R_i / R_g) – G_a.

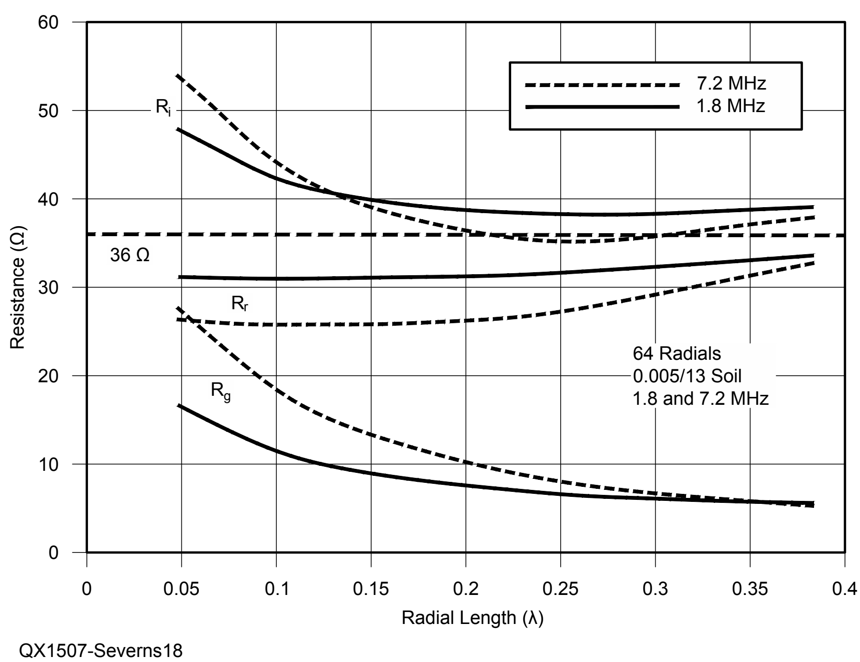

Figure 18 — Here is a graph showing R_i, R_r, and R_g for a ¼ λ vertical antenna with 64 radials at 1.8 and 7.2 MHz.

A Small 630 Meter Vertical Antenna

On 630 m (472 to 479 kHz), where 1 λ ≈ 2000 feet, any practical antenna is very likely to be small in terms of wavelength. Figure 19 shows an example of a short top-loaded vertical for 630 m. The vertical is 15.24 m high (50 feet, 0.024 λ) with 7.62 m (25 feet, 0.012 λ) radial arms in the hat. The usual practice for very short verticals is to have a dense ground system that extends some distance beyond the edge of the top-hat and/or a bit longer than the height of the vertical. Two cases were modeled: 64 and 128 radials, all 18 m long.

The results calculated from the *NEC* field data are given in Table 1. Over perfect ground $R_r = 0.7$ Ω

For this antenna with real soils, R_r is somewhat higher than the perfect ground case and converges on the perfect ground case as the soil conductivity improves. In this example using the perfect ground value for R_r yields an efficiency somewhat lower than real soil, as shown in the 0.7 / R_i column, but the difference is not very large. We should also keep in mind, as shown in Appendix C, that the computed values for R_r depend on the integration radius, which is somewhat arbitrary. If I had used a slightly larger radius, the R_r values would have been a bit lower, or closer to the ideal ground value.

Summary

For a lossless antenna in a lossless environment, the calculation of radiation resistance is very straight forward: integrate the power density over a hypothetical surface enclosing the antenna. The net power outflow divided by the square of the rms current at the feed point gives R_r. We can extend this technique to antennas in a lossy environment by using the field values obtained from *NEC* modeling and a spreadsheet.

At HF, values for R_r over real soils appear to be significantly lower than the values for the same antennas over perfect ground, at least in the case of ¼ λ and ½ λ vertical antennas! For short verticals at LF and MF, however, the real-ground R_r appears to be close to the ideal-ground value depending on the details of the soil and the ground system. It is my opinion that calculating P_r and efficiency using the perfect ground value for R_r is a reasonable approximation for the vertical antennas likely to be used by Amateur Radio operators at 630 m and 2200 m.

Measuring E-field intensities accurately many km from the antenna at low power levels and also figuring out the ground wave attenuation factors from soil measurements isn't practical for most hams. At LF and MF, forget the E-field measurements; just do some simple modeling to determine R_r over perfect ground and measure your base current: $P_r = I_o^2 R_r$!

Acknowledgements

I want to express my appreciation to Steve Stearns, K6OIK, for his very helpful review of this article. He put in a lot of effort and I've incorporated many of his suggestions in the main article and in the Appendices. I also appreciate the comments from Dean Straw, N6BV, and Al Christman, K3LC. All of the modeling employed a prototype version of Roy Lewallen's *EZNEC Pro/4* that implements *NEC 4.2* and Dan MaGuire's (AC6LA) *AutoEZ* which is an *EXCEL* spreadsheet that interacts with *EZNEC* to greatly expand the modeling options.[10, 11] Without these wonderful tools this study would not have been practical and I strongly recommend both programs.\

Rudy Severns, N6LF, was first licensed as WN7WAG in 1954 and has held an Amateur Extra class license since 1959. He is a consultant in the design of power electronics, magnetic components and power conversion equipment. Rudy holds a BSE degree from the University of California at Los Angeles. He is the author of three books, more than 90 technical papers and a past editor of QEX. *Rudy is an ARRL Life Member and an IEEE Life Fellow.*\

Notes

[8] J. Wait, R. Collin and F. Zucker, *Antenna Theory*, Chap 23, Inter-University Electronics Series (New York: McGraw-Hill, 1969), Vol 7, pp 414 – 424.

[9] The Appendices and other files associated with this article are available for download-ing from the ARRL *QEX* files web page. Go to **www.arrl.org/qexfiles** and look for the file **7x15_Severns.zip**.

[10] Roy Lewallen, W7EL, *EZNEC pro/4*, **www.eznec.com**.

[11] Dan McGuire, AC6LA, *AutoEZ*, **www.ac6la.com/autoez.html**.

Table 1A
630 m Vertical 64 Radials, Integration Radius = 100 m.

Soil	R_i [Ω]	R_r [Ω]	R_g [Ω]	R_r / R_i	$0.7 / R_i$	G_a
0.001/5	5.50	1.01	4.49	0.18	0.13	0.060
0.005/13	2.01	0.844	1.17	0.42	0.34	0.232
0.03/20	1.09	0.76	0.32	0.70	0.63	0.533
Perfect	0.69	0.69	0	1.00	1.00	1

Table 1B
630 m Vertical 128 Radials, Integration Radius = 100 m.

Soil	*Ri [W]*	*Rr [W]*	*Rg [W]*	*Rr / Ri*	*0.7 / Ri*	*Ga*
0.001/5	4.90	1.009	3.895	0.21	0.14	0.067
0.005/13	1.883	0.843	1.04	0.45	0.37	0.247
0.03/20	1.033	0.78	0.253	0.76	0.67	0.561
Perfect	0.69	0.69	0	1.00	1.00	1.00

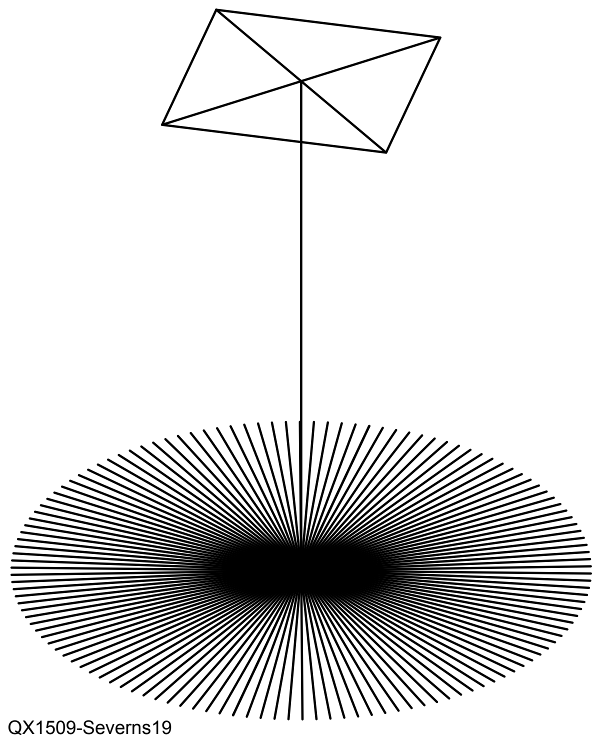

Figure 19 — This model shows a 630 m antenna example.

There was an error in the way Equation SB2 was printed in Part 1 of this article. That equation, in the **EIRP and Radiated Power, P_r, From Verticals** sidebar on page 29 of the July/August 2015 issue of *QEX*, was printed without the subscripts, superscripts and equals sign. There was also an error in the denominator of the equation. The correct equation is reproduced here. We apologize for this error, and offer our thanks to Andy Talbot, G4JNT, for being the first to point out the problem with this equation. — *Ed.*

$$EIRP = \frac{r^2 \left|E_z\right|^2}{30}\left[W\right] \qquad \text{[Eq SB2]}$$

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A Receiving Array for 160 m Through 2200 m

N6LF presents study of an antenna with low back lobes and the ability to switch the pattern direction and shape from the shack in a simple structure with no phasing networks.

For the past ten years I've participated in the ARRL 600 m experimental license group, WD2XSH, and tried a variety of receiving antennas from phased verticals (*E*-probes) to BOG's (Beverage on the ground) to terminated loops. I've also used regular Beverages on 160 m but at 475 kHz a 1.5 λ Beverage would be ≈ 3000 ft long and at 137 kHz over 10,000 ft, not very practical for most of us.

With the imminent authorization of the 2200 m and 630 m bands I needed an LF-MF receiving antenna with good performance from 100 kHz through 2 MHz. What I wanted was an antenna with low side lobes off the back (azimuths 90° through 270°) and the ability to switch the pattern direction and shape from the shack. All this of course is in a simple structure with no phasing networks.

Comments on Terminated Loops

Resistively terminated loops have many names: flags, pennants, EWEs, and so on. These antennas are usually electrically small — loop perimeters smaller than 0.1 λ — where λ is a wavelength at the operating frequency. Given the long wavelengths this will be the case for any practical antenna at 630 m or 2200 m. Because of the small size the current amplitude will be almost the same along the wire. The small variation in current magnitude translates into an insensitivity to the shape of the loop. Round, square or triangular makes little difference. This encourages us to use shapes that fit the available space and supports. Changing the size (area) of the loops has little effect on the pattern, it mostly affects the amplitude of the received signal. The greater the area of the loop, the greater the signal voltage *V* amplitude at a given frequency. It's just Faraday's law,

$$V = n\frac{d\phi}{dt}$$

where ϕ is the total flux and n is the number of turns. As we go down in frequency, for the same physical size, the signal decreases.

An essential feature of terminated loops is the use of a resistive termination somewhere in the loop. The value of the terminating resistor is typically in the range of 200 – 1200 Ω , which is much greater than the self-impedance of a small loop without the termination. The result is a feed-point impedance dominated by the fixed termination resistance. The feed-point impedance changes little as the frequency and/or loop size are changed. Another effect

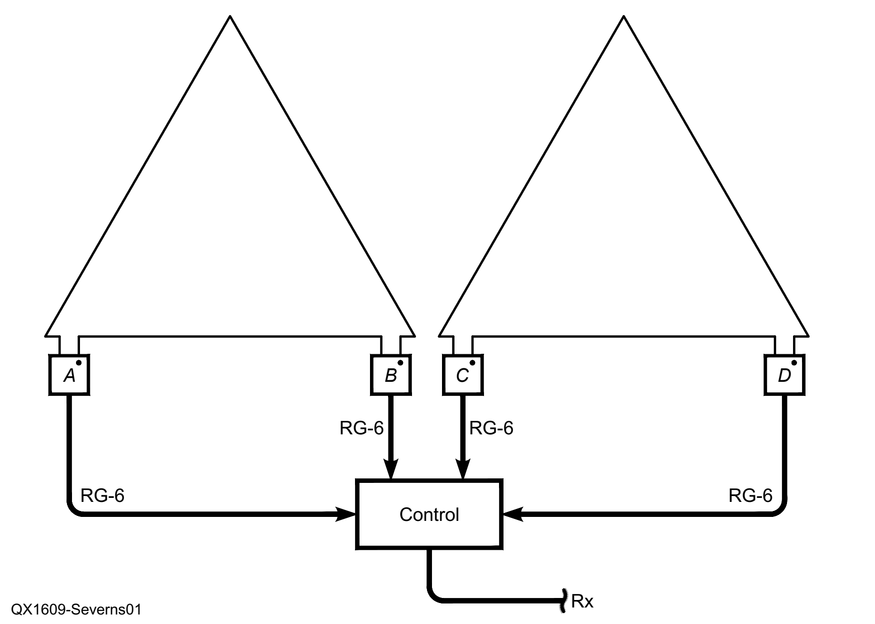

Figure 1 — *EZNEC* model for the receiving antenna.

of using a termination is to swamp out the mutual impedance due to coupling between loops. Changing the phase differences or the spacing between the loops has little effect on the feed-point impedances, which simplifies feed network design. This reduction in mutual coupling is exactly the same effect seen in phased arrays using short vertical elements (*E*-probes).

The properties of terminated loops lead me to think about combining them in an array. About that time the March 2015 issue of *QST* arrived with an article by Chris Kunze, DK6ED, on a his version of a double loop antenna.[1] This antenna is basically two triangular terminated loops in a line, fed 180° out of phase. What attracted my attention was the good pattern off the back of the antenna, sharp broadside nulls and the simplicity of the phasing scheme, which might allow the antenna work from 100 kHz to 2 MHz if it could be made large enough to have sufficient received signal on 2200 m but still be small enough to behave like a "small" loop on 160 m.

A bit of modeling with *EZNEC* was very encouraging so I built and tested an antenna.[2] This note describes that antenna in some detail. However, the reader should keep in mind this is just one example that happens to fit my particular location.

These antennas can be scaled up or down in size to suit a particular situation. The primary effect of scaling is to change the received signal strength. The directive patterns change very little.

The Antenna

The antenna is shown in Figure 1. I have two ≈80 ft poles, spaced 150 ft in my pasture from which I could suspend the antenna. Each loop is an equilateral triangle 73 ft on a side. The bottom wires are 8 ft above ground and the corners at the mid-point are 2 ft apart. At each end of each of the bottom wires (points *A*, *B*, *C* and *D*) there is a 1 kΩ to 75 Ω impedance transformer with a common-mode choke for isolation (Figure 2). Each choke is connected to a length of 75 Ω RG-6 leading back to the control box in the shack. The control box determines how the feed points are driven — which are terminated, which are driven and what the phase relationship will be between the two loops. The cables back to the control box can be of any length but all four cables must be the same electrical length! It's best if all four cables are cut to the same physical length from the same roll of cable.

The 100 kΩ resistor in Figure 2 is for

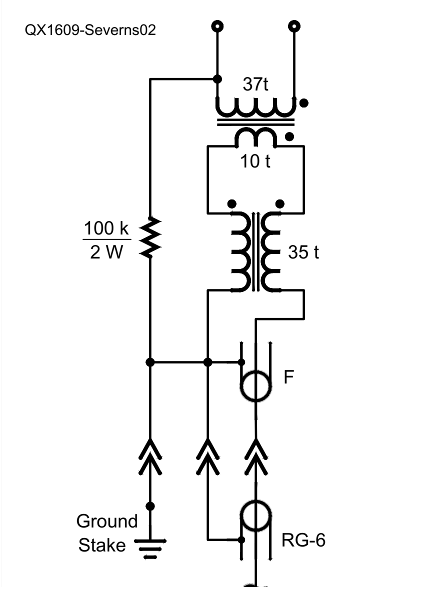

Figure 2 — Impedance transformer and common mode choke. RG-6 with F-connectors runs to the control box.

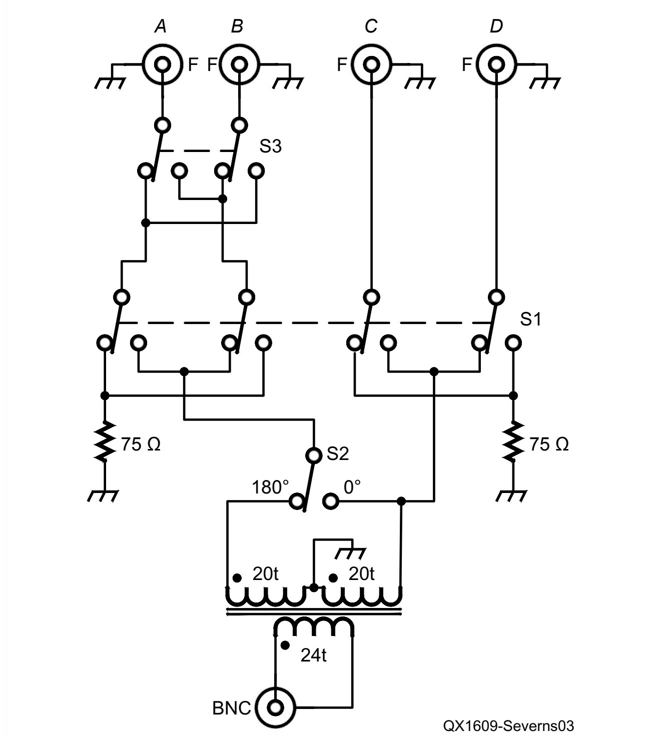

Figure 3 — Control unit schematic. F-connectors are used at *A*, *B*, *C* and *D* in the 75 Ω portion of the system, and a BNC connector is used at the 50 Ω connector to the receiver.

Table 1
Source and termination locations.

Configuration	*Left source*	*Right source*	*Left termination*	*Right termination*	*Relative phasing*
1	*B*	*D*	*A*	*C*	0
2	*A*	*C*	*B*	*D*	0
3	*B*	*D*	*A*	*C*	180°
4	*A*	*C*	*B*	*D*	180°
5	*A*	*D*	*B*	*C*	0
6	*B*	*C*	*A*	*D*	0
7	*A*	*D*	*B*	*C*	180°
8	*B*	*C*	*A*	*D*	180°

static discharge, these are large wire antennas that could accumulate a charge under some weather conditions. Construction details for the transformer-chokes and the control box are in the last section of this article. The control box contains only three switches and a phase-inversion transformer as shown in Figure 3.

The terminations are 75 Ω resistors placed in the control box. The 75 Ω is transformed to 1 kΩ at the antenna with the transformers at *A*, *B*, *C* and *D*. Whether a cable is acting as a source or as a termination is determined in the control box. If *A* and *C* are terminated and *B* and *D* are sources, the radiation maximum is to the right, from the terminations towards the sources. The transformer provides 180° phase inversion and, with the turns ratios shown, also transforms the 75 Ω impedances to 50 Ω at the receiver output.

There are eight different combinations of sources, terminations and relative phasing (0° or 180°). These combinations are summarized in Table 1.

Each combination has a specific pattern although configurations 5 and 6 have the same pattern as do 7 and 8. The result is four different patterns, two of which are reversible, that can be selected from the control box in the shack.

Figures 4 through 7 are for 475 kHz but the patterns at 1.83 MHz and 137 kHz are very similar except for differences in peak gain. This is illustrated in Figures 8 through 11, which compare the directivity patterns for 160 m and 630 m. The outer (higher gain) patterns are configuration 1, the loops are driven in-phase. The inner patterns are for configuration 3, loops driven 180° out of phase.

At 160 m, Figures 8 and 9 illustrate significantly improved directivity going from the loops in-phase to 180° out of phase, it also shows the significant reduction in peak gain (≈ -5 dBi). Figures 10 and 11 are for 630 m and again we see a significant improvement in directivity with 180° phasing, but an even larger reduction in peak gain (≈ -16 dBi). The patterns for 2200 m are very similar to 630 m except that there is another 20 dB of

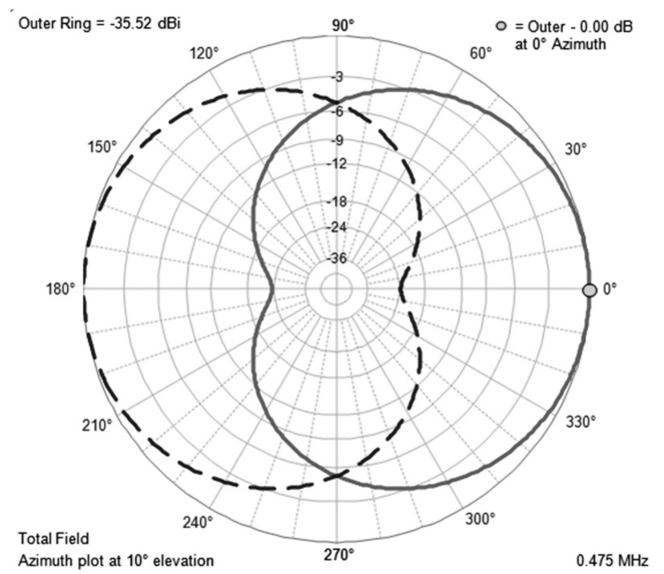

Figure 4 — Pattern for configurations 1 and 2.

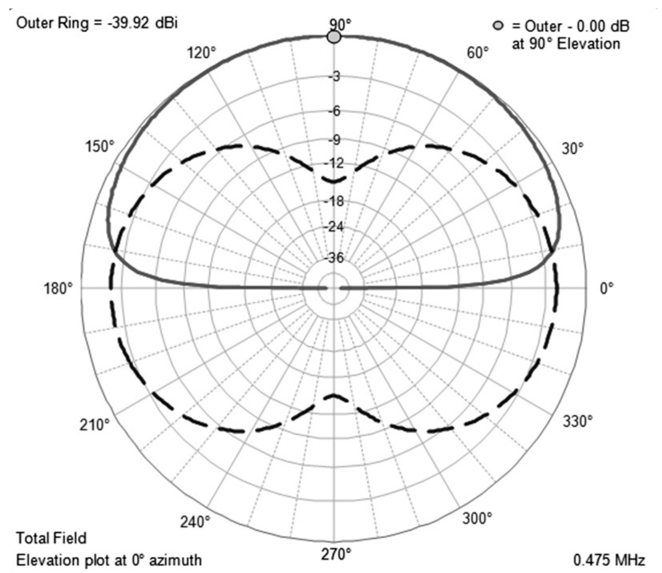

Figure 6 — Pattern for configurations 5 and 6.

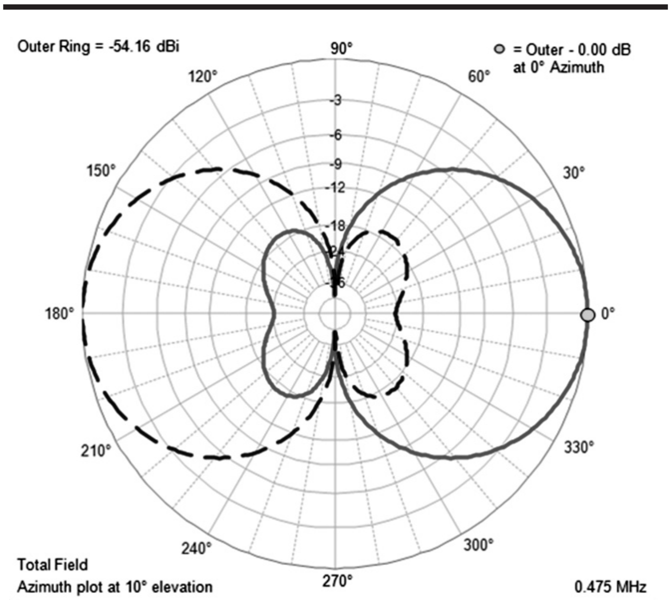

Figure 5 — Pattern for configurations 3 and 4.

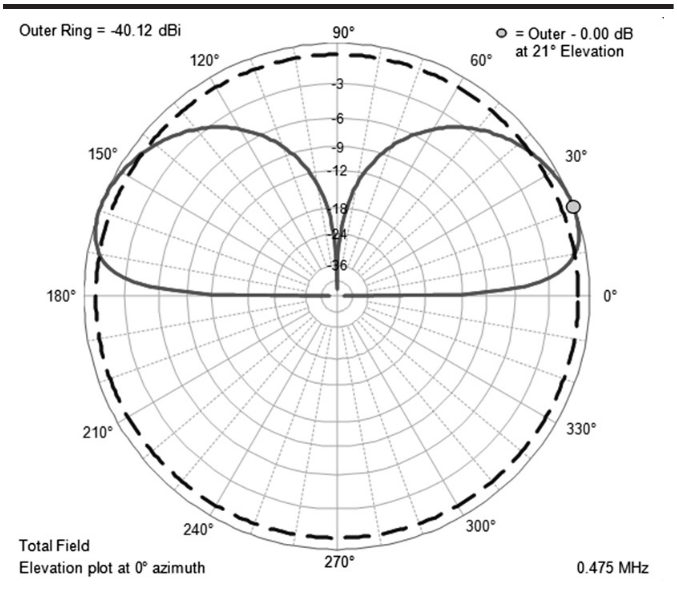

Figure 7 — Pattern for configurations 7 and 8.

gain reduction. The signal levels on 160 m and 630 m are not alarming low and on-the-air testing has shown that an amplifier is not needed. However, on 2200 m a preamp would be helpful — between 20 to 40 dB would be adequate — although I have been using my antenna successfully on 137 kHz for WSPR signals without additional receiver gain.

The predicted performance on 160 m, 630 m and 2200 m for different configurations is summarized in Table 2.

Near-field Patterns

All of the directivity patterns shown to this point have been for the far-field — many wavelengths from the antenna. At 475 kHz λ is ≈ 2,000 ft and at 137 kHz λ is ≈ 7,200 ft. The directivity pattern for any noise source — like a utility line or neighbors TV — within that distance will be the near-field pattern, which can be very different from the far-field pattern. Figures 12 and 13 show a comparison between near and far-field patterns with the noise source at a distance of 400 ft at 475 kHz for the near-field pattern.

The solid lines represent the far-field patterns and the dashed lines the near-field patterns. Note the scale is in m*V*/m not dB. When the loops are both driven in phase (configuration 1) there is some degradation in the near-field pattern compared to the far-field but it's not too severe. However, the difference between the near and far-field patterns with 180° phase difference (configurations 3 and 4) is very great. This is a very important observation for locations in congested urban environments. Although the far-field pattern with 180° phase difference is much more directive, the local noise rejection is grossly inferior. Configurations with 180° phase difference may not be usable in these situations.

Sensitivity to Shape

The configurations listed in Table 2 assume two symmetric triangles. To illustrate how insensitive to loop shape the antenna is, I modeled the variation shown in Figure 14, and show a performance comparison in Table 3. The first entry is Figure 1 and the second Figure 14.

The differences are very small. This implies that the primary driver for loop shape will be the available supports.

An Extended Version

I happen to have another 80 ft pole in line with the first two, again spaced 150 ft. I've considered duplicating the present antenna and extending it to four loops as shown in Figure 15. Figures 16 – 18 show patterns associated with Figure 15. Receive directional factor (RDF) is 13.6 dBi at 475 kHz with an antenna that is only 300 ft long! A comparable Beverage would be almost a mile long. However, the Beverage would have a lot more signal coming out of it.

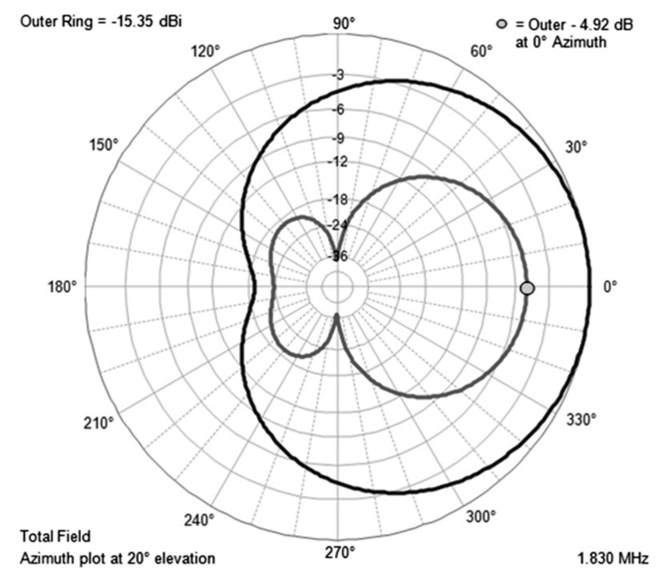

Figure 8 — 1.83 MHz azimuth plot at 20°.

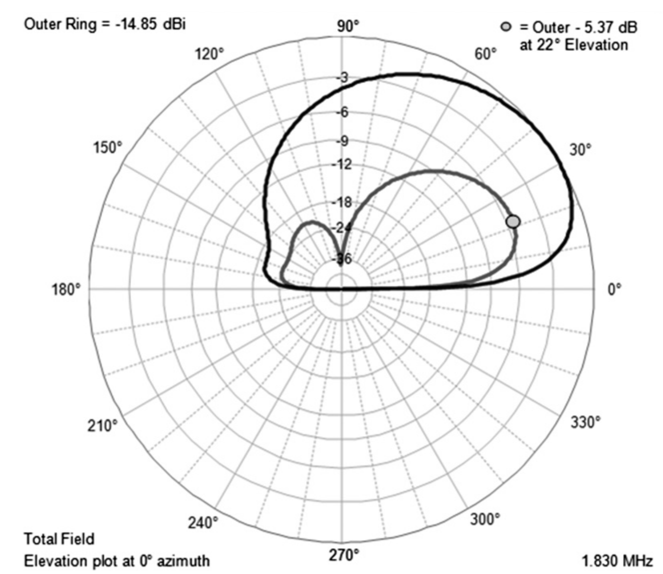

Figure 9 — 1.83 MHz elevation plot.

Table 2
Performance summary.

Band	*Configuration*	*F/B [dB],10° elev.*	*F/R [dB],10° elev.*	*RDF*	*Max gain [dBi]*	*at Az°*	*at El°*
160 m	1 & 2	18.39	3.91	7.13	-12.48	0	38
160 m	3 & 4	18.07	15.07	11.22	-20.12	0	22
160 m	5 & 6	0.00	0.00	6.33	-15.81	0	90
160 m	7 & 8	0.00	0.00	5.01	-17.40	0	26
630 m	1 & 2	23.49	5.22	7.71	-34.44	0	26
630 m	3 & 4	24.43	16.73	11.52	-53.55	0	18
630 m	5 & 6	0.00	0.00	5.47	-39.92	0	90
630 m	7 & 8	0.00	0.00	4.77	-40.12	0	20
2200 m	1 & 2	23.63	5.33	7.71	-55.46	0	20
2200 m	3 & 4	14.63	14.63	11.08	-85.18	0	14
2200 m	5 & 6	0.00	0.00	5.22	-61.38	0	90
2200 m	7 & 8	0.00	0.00	4.71	-61.08	0	16

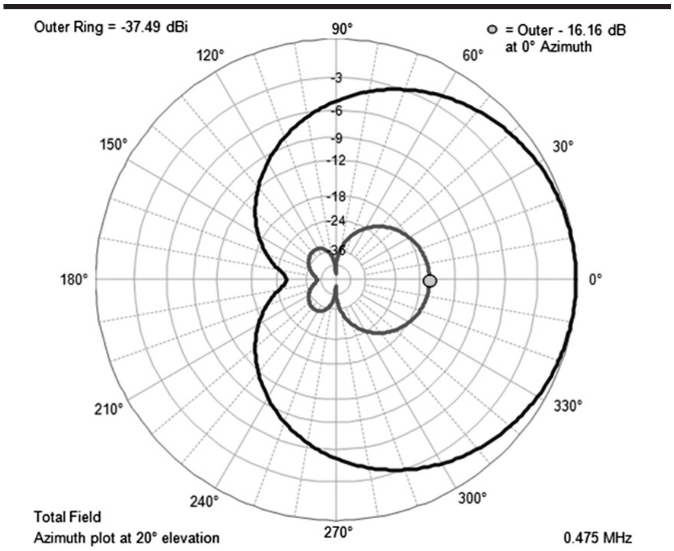

Figure 10 — 475 kHz azimuth plot.

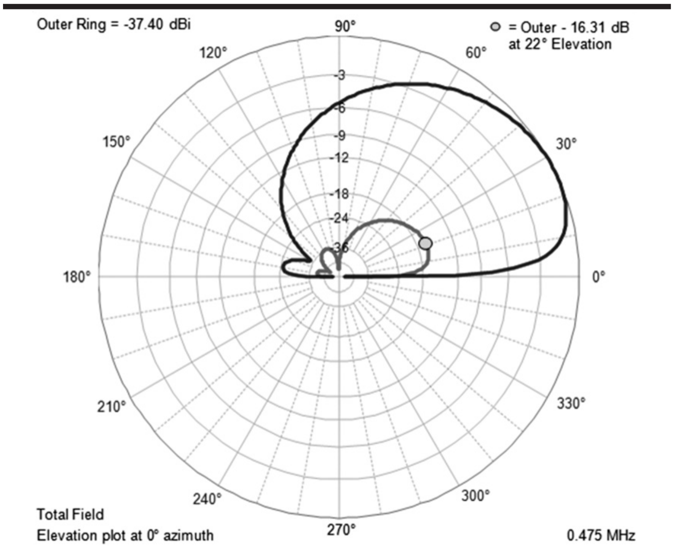

Figure 11 — 475 kHz elevation plot.

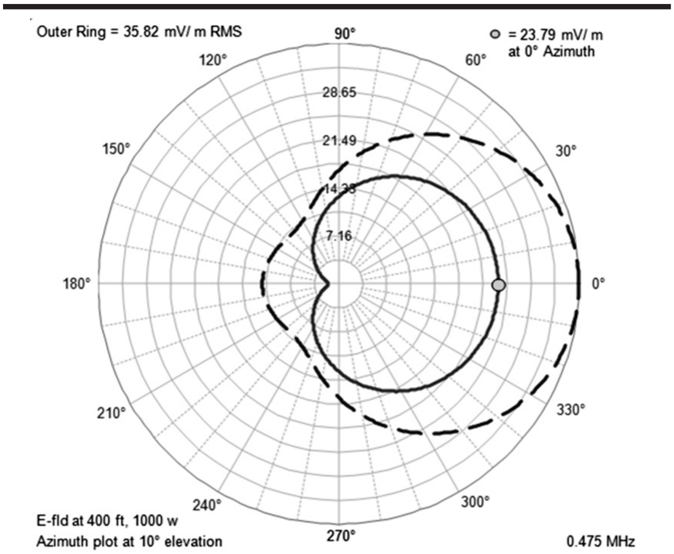

Figure 12 — Comparison between near and far-field patterns for zero phase difference.

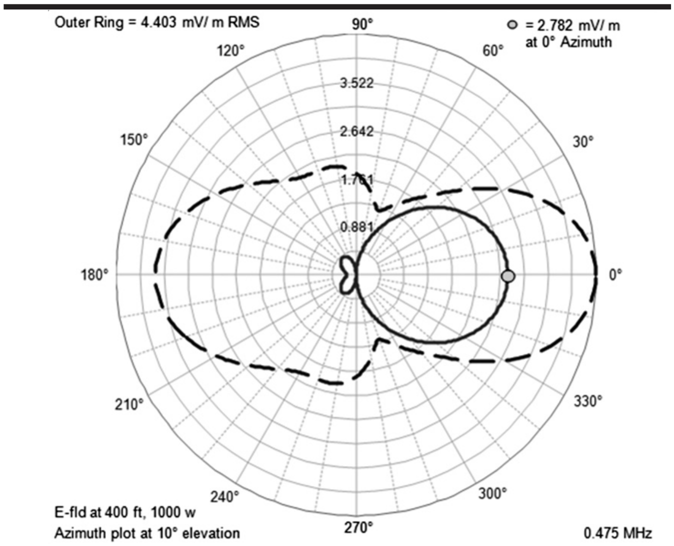

Figure 13 — Comparison between near and far-field patterns for 180° phase difference.

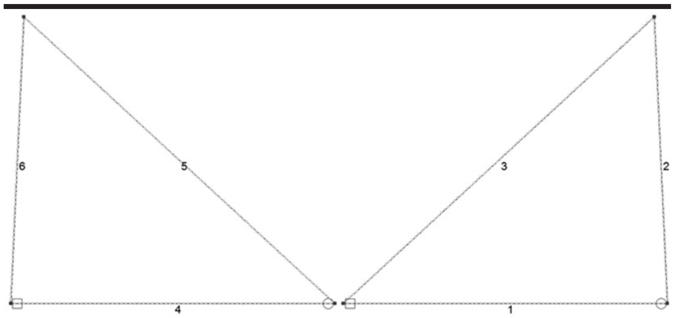

Figure 14 — An alternate loop shape.

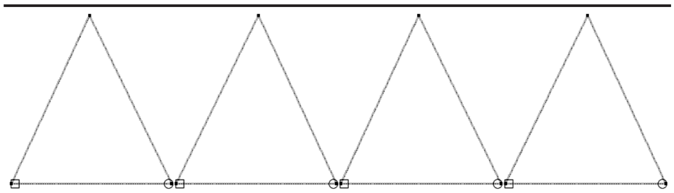

Figure 15 — Four loop version.

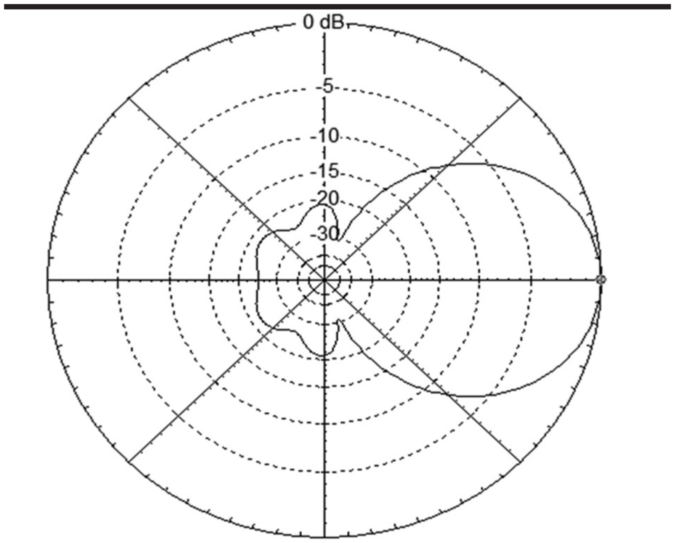

Figure 16 — Four loop azimuth pattern.

Table 3
Performance comparison.

Band	*Configuration*	*F/B [dB]at 10° elev.*	*F/R [dB] at10° elev.*	*RDF*	*Max gain [dBi]*	*at Az°*	*at El°*
630 m	1 & 2	23.49	5.22	7.71	-34.44	0	26
630 m	1 & 2	21.92	5.10	7.68	-34.36	0	26

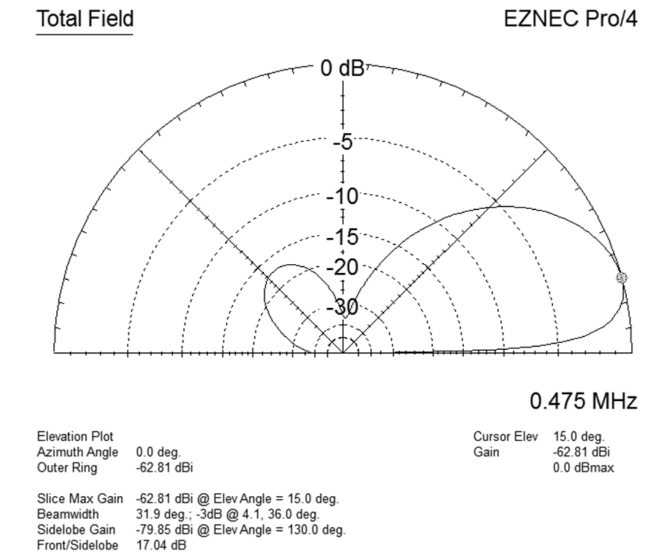

Figure 17 — Four loop elevation pattern.

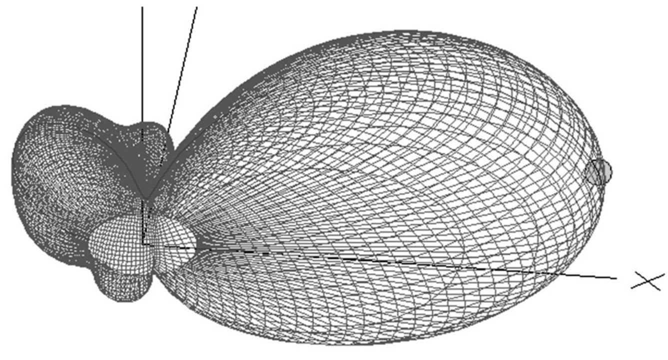

Figure 18 — 3-D pattern for four loops.

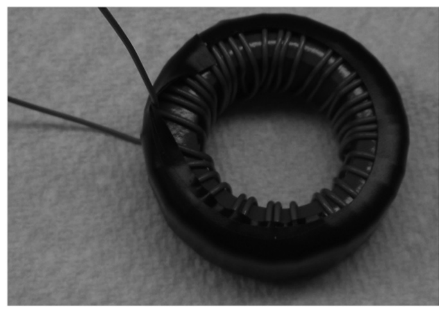

Figure 19 — Secondary winding on the impedance transformer.

Verification

Modeling is a great tool, providing reliable predictions, but in the end it's necessary to verify the predictions and that the antenna is correctly assembled. Does this contraption actually work? After a careful visual check that all the electrical connections are correct, and that all of the transformer/chokes are correctly connected to provide proper phasing. Figures 1 and 2 have prominent phasing dots to indicate the proper connections. Even with careful assembly it is possible to switch one or more of the connections. There are a couple of ways to quickly check the polarity of the transformers. First, set the control to 0° phasing (configuration 1), then switch the direction (configuration 2). There should be no significant change in signal level for the background noise. If there is a large change then at least one of the transformers is reversed. Next change the phasing to 180° (configuration 3). There should be a substantial drop in signal level but the new level should not change much when the pattern is reversed (configuration 4). Finally, select a strong signal with a known direction, more or less in line with the main lobe, then reverse the pattern. This should show the F/B of the array and confirm the directions are correct. If all these are as expected then you probably have the phasing correct.

You can also make some impedance measurements. The feed system is designed for 75 Ω up to the control box, and the impedances within the feed system should be close to this over the entire frequency range. Using a VNA2180 vector network analyzer I measured the impedances at several points from 100 kHz to 2 MHz as I switched the control box through the various configurations. The first point was the output port to the receiver. The impedance was close to 50 Ω as designed. Tthe phase inversion transformer converts the 75 Ω impedance of the feed system to 50 Ω for the receiver. I next measured the impedances at the control box end of the feed cables one at a time while switching between configurations. Each of these measurements was a sweep over the frequency range. All of the graph plots were very similar with an SWR < 1.5:1, indicating there were no major errors. The antenna impedances agreed with predictions.

That was the easy part! The next step was to verify that the antenna had the predicted directivity patterns associated with each configuration. The ideal procedure would be to place a signal source well beyond the Fresnel zone, that is, more than 10 λ distant at various azimuths and measure signal strengths as the pattern was switched. At 137 kHz or even 475 kHz the distances to the sources would have to be many miles although at 1.8 MHz the distances are not so great. My location is in a small valley surrounded in most directions by hills so this approach did not seem practical except perhaps for checking the depth of a null in a particular direction on 160 m. I needed to be a bit more crafty! Because the patterns are basically the same from 100 kHz to 2 MHz, I realized I could use signals anywhere in that range. There are a large number of well defined signals in this range, most prominently AM broadcast stations. There are also aeronautical and coastal navigation beacons and the WSPR transmissions by Amateur Radio experimental stations. From long experience with Yagis and other arrays we know that the null depth and location is much more sensitive than the details of the main lobe. In general if the nulls are where they should be and the null depth anywhere near what it should be, then we can have confidence that the pattern is close to its predicted form. Locating and measuring pattern nulls can take us a long way towards verifying the actual pattern.

To identify and measure signals I have an old HP3585A spectrum analyzer. This allowed me to see the station signals and measure their amplitudes. The instrument displays the amplitude to 0.01 dB but that's deceiving. Even strong local BC signals have several dB of variation (noise) even with very narrow scans, which makes resolution of the main lobe impractical but it's still possible to get a good estimate of null depths and locations by observing the signal while switching the pattern direction. Switching the pattern doesn't help however, with the nulls to the side (±90°, see Figure 5). I was able to find BC stations lying along the axis of the array which showed the predicted F/B ratios reasonably well. The preliminary measurements with BC and 630 m WSPR stations indicate the patterns are close to the NEC predictions, at least the nulls.

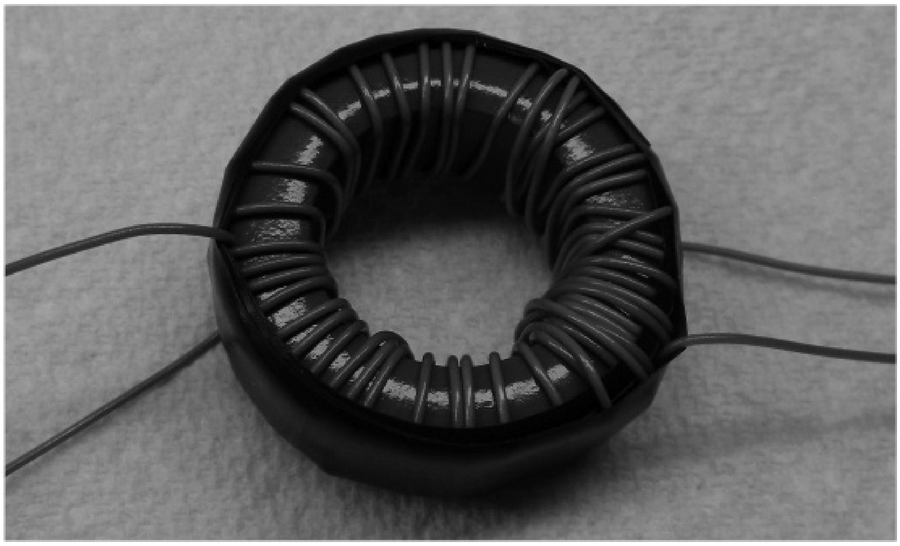

Figure 20 — Primary winding added to the impedance transformer.

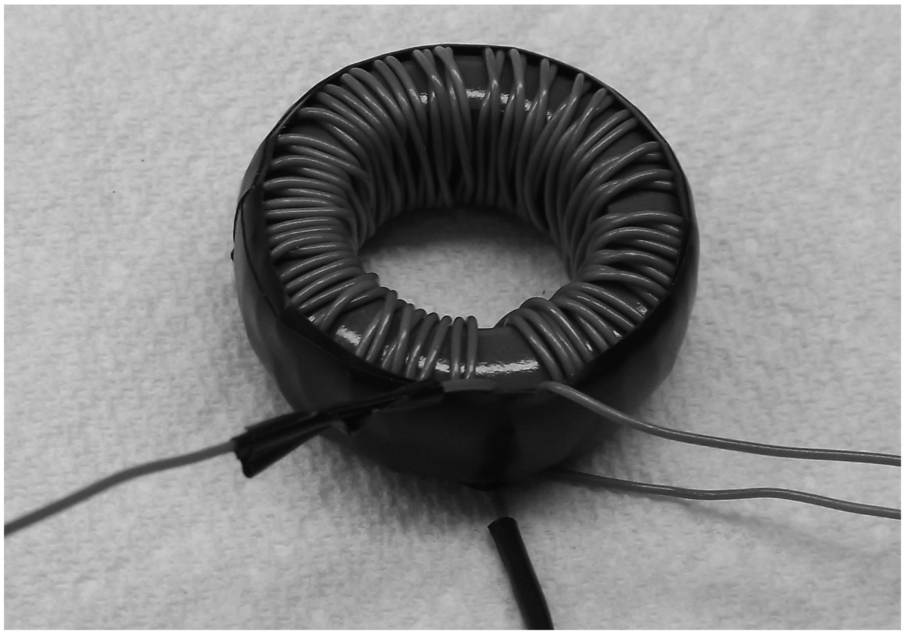

Figure 21 — Common mode choke.

Transformers and Control Unit Details

As indicated in Figure 1, the loops are fed or terminated at the lower corners. At each point (*A*, *B*, *C* and *D*) there is an isolated impedance transformer, 1000 Ω to 75 Ω like the one shown in Figure 2. To further isolate the transmission lines from the antenna, on the primary of the impedance transformer there is a common mode choke. Note the use of winding polarity dots in the transformer-choke schematic of Figure 2. Keeping track of the phasing is critical! When toroidal cores are used, two windings are in phase — the same dot — when both wires come out of the core in the same direction.

The impedance transformers, the common mode chokes, and the phase inversion transformer are all wound on the same toroidal ferrite core, Fair-Rite #5977002721. Nine cores are needed for this project. I obtained them from Mouser Electronics for about $3.75 US each.[3] These cores are type 77 ferrite, recommended for use in low flux applications below 3 MHz. All of the windings used #26 AWG insulated wire. Neither the wire size nor the insulation type is critical. I simply used what I had on hand. You have to use wire small enough for the windings to fit on the cores. The magnetic components must to work from 137 kHz through 1.9 MHz. The feed-point transformers are used to isolate the antenna from the feed system and to transform the to 75 Ω resistance on the primary to1000 Ω on the secondary to properly terminate the loops. The transformer shunt impedance has be significantly greater than1000 Ω to maintain proper termination. This has to be the case over the entire range of 137 kHz to 2 MHz. At the low frequency the issue is enough inductance with a reasonable number of turns. The type 77 ferrite has high permeability, about 2000, up to 1 MHz, above which it starts to decrease but is still adequate for this application at 2 MHz. We also have to maintain a sufficiently high self resonant frequency, f_r, so that there is sufficient shunt impedance, Z_s, at 2 MHz. Like the transformer, the choke also needs to have sufficient Z_s over the entire range. This becomes a bit of a balancing act, more turns give more low frequency impedance but lower f_r with reduced impedance at 2 MHz. 35 turns gave f_r=700 kHz, with Z_s=2.8 kΩ at 137 kHz, 20 kΩ at 475 kHz and 6.1 kΩ at 1.8 MHz. These values, while not ideal, are an acceptable compromise. Figures 19 – 21 show some of the winding details.

The common mode choke has 35 turns wound bifilar (two wires twisted together). Note the careful marking of one pair of wires, these allow us to indentify each of the windings. As shown in Figure 2, for correct phasing the center conductor of the feed line must be connected to the dotted end of the primary winding. As shown in Figure 21, I placed a small piece of tape on one winding. On the bottom of the choke I connected the taped winding to the center conductor of the input F connector. I then connected the other end of the taped winding to the dotted end of the transformer.

Note also that the ends of two windings come out on the same side of the toroid, the windings from the same side have the same polarity — they share the same "dot". This convention applies also to the impedance transformer.

The transformer-chokes were installed in insulated junction boxes (Figure 22) available at most hardware stores. The left box is for point *A* in Figure 1. Points *B* and *C* are combined in a common box (middle) and point *D* is in the box on the right. The cores are secured with some silicone caulk/adhesive. The terminals to which the antenna wire is attached were simple SS machine screws in holes through the sides of the boxes. The holes were tight and caulked with silicone.

The installation at point *B* – *C* at the center of the antenna is shown in Figure 23. Notice the careful markings on the box and the cables to keep track of proper phasing and cable connections. For the antenna to work as expected it is vital that all the connections are correct. To this end every cable was marked at both ends, *A*, *B*, *C*, etc. Every RF connector on the feed point boxes and the control unit was also carefully marked to avoid confusion during assembly. The antenna was made from #17 AWG aluminum electric fence wire.

Summary

The final version of my antenna is basically the same as DK6ED's, just scaled up and with some added switching to give additional patterns. There are four modes of operation, two of which are reversible. On several occasions while using the antenna I've found the pattern associated with 180° phase shift to be too narrow for general listening. The deep side nulls cut out stations north and south of me. In fact most of the time I leave the loops in-phase, switching to 180° phasing only when it seems to help. I have been using the antenna on 160 m, 630 m and 2200 m without an amplifier. This has worked very well, however, if the antenna were scaled down in size, an amplifier might be needed especially on 2200 m.

I spent a great deal of time trying to optimize this antenna, varying the shape, relative phasing, termination resistances and even exploring reactive terminations. I found all this made very little difference. The antenna seemed to work about the same no matter what I did to it. Even changing the soil characteristics under the antenna has only modest effect. The received signal amplitude is a function of the size of the loops. Bigger loop mean more signal, but that's about all that changes as the loop size is varied.

Rudy Severns, N6LF, was first licensed as WN7AWG in 1954. He is a retired electrical engineer, an IEEE Fellow and ARRL Life Member.

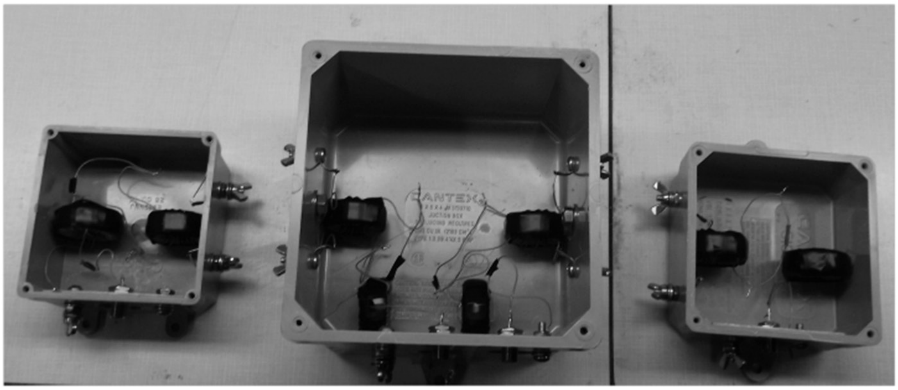

Figure 22 — Feed point boxes with transformer-chokes installed.

Figure 23 — Transformer box at the center of the array.

Notes

[1]Chris Kunze, DK6ED, "The DK6ED Double Loop", *QST* Mar 2015, pp. 34-37.

[2]Several versions of *EZNEC* antenna modeling software are available from developer Roy Lewallen, W7EL, at **www.eznec.com**.

[3]**www.mouser.com**.

A Wideband 80-Meter Dipole

This worthy antenna is so simple and inexpensive you'll want more than one!

By Rudy Severns, N6LF
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Cottage Grove, OR 97424
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The 500-kHz width of the 80-meter band makes it by far the widest HF amateur band on a percentage basis—13% of the center frequency. Over the years, a legion of articles have described antennas that purported to provide an SWR of less than 2:1 over the whole band. Some did, some didn't. With my two transceivers—a Drake TR-7 and Yzaesu FT-757GX—even a 2:1 SWR isn't low enough because the rigs automatically begin to reduce output power before a 2:1 SWR is reached. I suspect this is not an uncommon occurrence with other rigs (not equipped with built-in automatic antenna tuners) as well.

What's really needed is an antenna that provides an SWR below 1.6 or 1.7:1 over the entire band. It'd be really convenient to jump from one end of the band to the other without having to think about retuning the antenna tuner or rig, or buying an automatic antenna tuner. Such a requirement makes antenna design tough!

The following is a description of an antenna that meets the need. This one has been built—and it works great with no noticeable SWR degradation caused by rain, snow, wind or other weather elements. Surprisingly, it's a simple wire antenna that's only as long as a standard dipole.

Earlier Antennas

My idea has its roots in two well-known antennas: the open-sleeve dipole[1,2,3] and the folded dipole.[4] With an open-sleeve dipole, additional conductors are added in close proximity to—but not connected to—a common single-wire dipole, as shown in Figure 1. In addition to the fundamental resonance of the simple dipole, the added conductors create new resonances. This effect can be used to multiband or broadband an antenna—and it's an idea that's been around since WW II.

A folded dipole's bandwidth is greater than a single-wire dipole made of the same wire size. Although the bandwidth attainable with a folded dipole is better, by itself, it's still not good enough for our needs. Figure 2 shows the typical SWR plot for a folded dipole, using 12-inch element spacing, #14 wire and centered on 3.750 MHz. This antenna's 2:1 SWR bandwidth is approximately 375 kHz. You can improve things a bit by using greater element spacing, but then the weight and length of the spacers gets to be a hassle and you still won't have sufficient bandwidth.

Antenna Height

One common problem with any 80-meter antenna is installing it high enough to do some good. Because the current maximum is at the center of the dipole, it's important to keep that part of the antenna as high as possible. For most installations, 70 feet is pretty high, but at 80 meters, 70 feet is only one quarter of a wavelength.

A dipole's radiation angle is largely determined by the height of its center. If the antenna is strung between two supports, there's bound to be some sag, height is lost and the radiation angle raised. Weight of any sort—baluns, long lengths of coax, matching networks, etc, particularly near the antenna's center—contribute to sag. The resultant high-angle radiation is great for local QSOs, but bad news for DXing.

If I can't provide support for the antenna center, I prefer to use lightweight transmitting twinlead (weighing in at 2.4 lb/100 ft versus the 9.4 lb/100 ft of RG-8 coax), with the balun at ground level. The 450-Ω ladder line is quite efficient, relatively light and costs about 16 cents a foot, much less than coax.

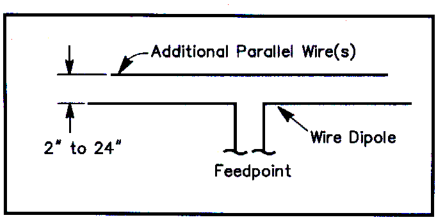

Figure 1—An open-sleeve dipole example.

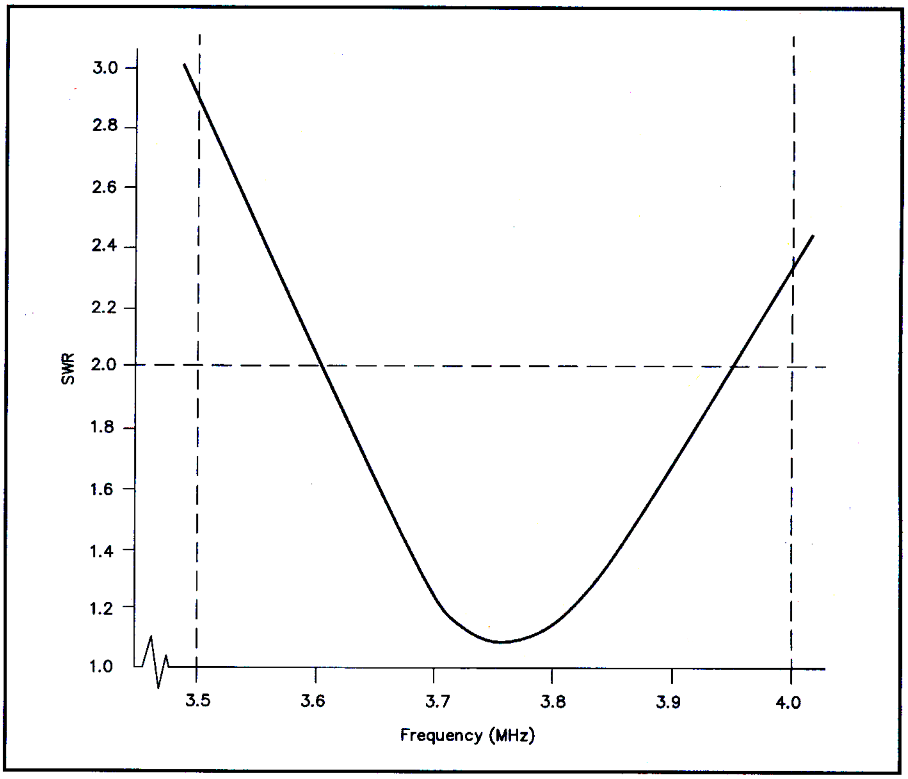

Figure 2—Typical SWR for a folded dipole resonant at 3.75 MHz.

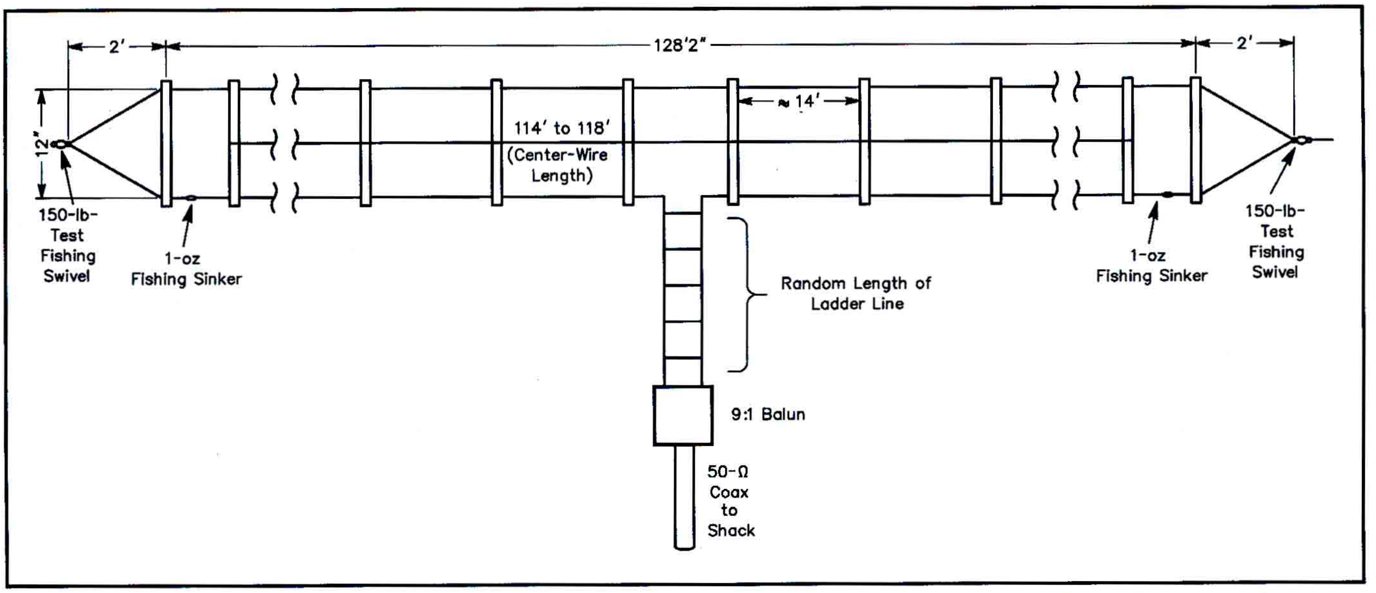

Figure 3—The open-sleeve folded dipole: simple and inexpensive. The antenna is fed with a random length of 450-Ω open-wire transmission line through a 9:1 balun.

The feed-point impedance of a folded dipole is about 300 Ω. Although 300-Ω ladder line is available, making the transition from 300 Ω to 50 Ω requires a 6:1 balun. Such baluns can be bought or made, but 4:1 or 9:1 baluns are much more common.

A Broadband 80-Meter Antenna

The antenna I came up with is shown in Figure 3. It's simply an open-sleeve version of the folded dipole. The resonator wire added midway between the two folded dipole elements is supported by the spacers already used for the folded dipole. That's all there is to it: a single wire down the middle of a folded dipole! One interesting result of adding the wire is that not only is the antenna very broadband, but by juggling the spacing and wire lengths a bit, Z_r is very close to 450 Ω, which fits in nicely with available transmission lines and a 9:1 balun. The transmission line operates with a very low SWR and can be of virtually any length.

A graph of the measured SWR for two lengths of the center wire (L_c) is shown in Figure 4. The measurements were made with considerable care, using Bird wattmeters. For L_c = 118 feet, the highest SWR is 1.5: 1, and is less than that over most of the band. For L_c = 114 feet, the worst-case SWR is 1.8:1, but the overall 2:1 bandwidth is extended to 800 kHz. This would be advantageous to MARS operators operating just above the upper band edge. Experimenting further, I shortened L_c to 112 feet, which pushed the 2:1 bandwidth up to nearly 1 MHz (3.3 to 4.25 MHz). For most hams, that may not be of great importance, but it's something to keep in mind.

Figure 3 shows the number and separation of the wire spacers. It's important to keep the spacers as light (and inexpensive!) as possible. The two spacers on each end have to be fairly stiff, so I used sections cut from solid fiberglass electric-fence wands.[5] The rest of the spreaders are made from half sections of ½-inch CPVC plastic pipe. They're about half the weight of the fiberglass wand spreaders. I could have used full sections of the CPVC pipe for the end spreaders but, for the same weight, they would have had more wind loading.

Summary

Modeling this antenna, which is essentially a transmission line, doesn't work very well on *MININEC*-based programs.[6] *NEC* programs such as *NecWires*[7] are needed, and even then, you have to use 50 to 100 segments per λ/2 for the final design. Using *NecWires*, the total computed loss was only 0.07 dB (1.6%) for #14 wire and 0.09 dB (2%) for #16. Combined with the very low loss of the open-wire transmission line, if a good three-core, 9:1 Guanella balun[8] is used, the overall efficiency will be quite good.

At best, these antennas will be close to the ground in terms of wavelengths. The

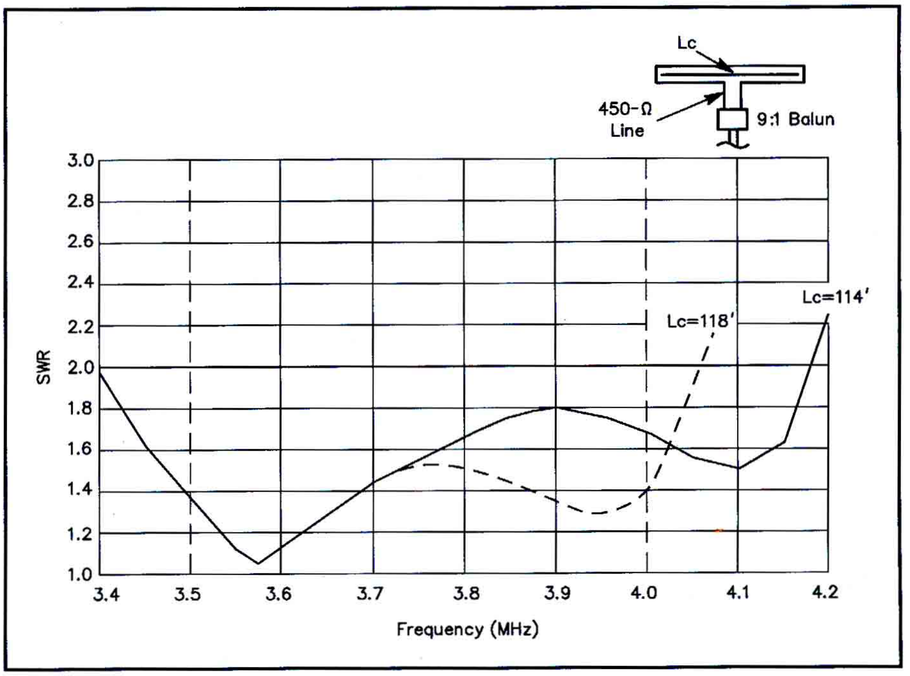

Figure 4—SWR curve of the open-sleeve dipole of Figure 3, showing curves for different lengths of the center wire (L_C).

ground effects are important and will affect the impedances and final dimensions. This antenna was modeled at a height of 70 feet over poor ground ($\varepsilon_r = 13$, $\sigma = 2$ mS), which corresponds (more or less) to my location and support height. I only had to adjust the center wire a bit to get the predicted performance. At another location or antenna height, the final performance and dimensions may be different.

A folded dipole *loves* to rotate when being hoisted and it twists when the wind blows, which really upsets the SWR if the parallel wires short together. In Figure 3, I've included a couple of details that help reduce this problem. The ends of the dipole are not symmetrical. To aid in avoiding antenna twist, 1-oz fishing sinkers are added to the bottom wire on each end. I also use two heavy-duty (150-pound-capacity) fishing-line swivels at the antenna support points.

Performance isn't the only criterion for a good antenna. For most hams, cost is always a consideration. This design uses 380 feet of wire (a total cost of $34 at 9 cents per foot[9]) and about a buck's worth of 1/2-inch CPVC pipe. The CPVC can also be used for the center and end insulators. The 450-Ω open-wire transmission line costs 14 to 16 cents per foot (see Note 9), so add another $15 to the total. So, for $50, you've got everything but the balun and the lead-in coax. You've also got a darn good antenna.

Notes

[1]Roger Cox, WBØDGF, "The Open-Sleeve Antenna: Development of the Open-Sleeve Dipole and Open-Sleeve Monopole for H.F. and V.H.F. Amateur Applications," *CQ,* Aug 1983, pp 13-19.

[2]Gary Breed, K9AY, "Multi-Frequency Antenna Technique Uses Closely Coupled Resonators," *RF Design,* Nov 1994, pp 78-85.

[3]Bill Orr, W6SAI, "Radio FUNdamentals," The Open Sleeve Dipole, *CQ,* Feb 1995, pp 94-98.

[4]R. Dean Straw, N6BV, Ed, *The ARRL Antenna Book* (Newington: ARRL, 17th ed, 1994), p 2-32.

[5]These wands, which measure 3/8-inch in diameter and are 4 feet long, are available from farming supply stores and Sears.

[6]*ELNEC* is available from Roy Lewallen, W7EL, PO Box 6658, Beaverton, OR 97007. *AO 6.1* is available from Brian Beezley, K6STI, 3532 Linda Vista Dr, San Marcos, CA 92069, tel 619-599-4962.

[7]*NEC-Wires* is available from Brian Beezley, K6STI; see Note 6.

[8]Jerry Sevick, W2FMI, *Transmission Line Transformers* (Newington: ARRL, 2nd ed, 1990), p 9-28.

[9]Several *QST* advertisers offer 450-Ω open-wire transmission line. See the ads at the back of this issue.

Rudy Severns was first licensed as WN7WAG in 1953, became W7WAG the following year and got his Amateur Extra class license in 1959. An electrical engineer who received his BS from UCLA in 1966, Rudy is a senior member of the IEEE, an ARRL life member and presently works as an independent consultant for power electronics and power conversion design.

By Rudy Severns, N6LF

Verticals, Ground Systems and Some History

What makes a vertical antenna cook? Here you can gain some insight as to what this popular antenna likes and dislikes.

Over the past 100 years, beginning with Marconi and continuing to this day, vertical antennas and their associated ground systems have received considerable attention. Many fine articles and technical papers have explained the finer points of vertical antenna operation. Sometimes we forget the information's origins—and sometimes the wisdom gets a little distorted. Occasionally it's worthwhile to revisit the earlier work and recognize how the old relates to present-day applications.

Research

A few years ago, I decided to get on 160 meters and wanted an effective antenna. I decided on a vertical of one form or another, but soon realized that I really didn't have a good understanding of how to get the best performance from a vertical. That led me to research the amateur and professional literature and discover a treasure trove of information.

Examining these early papers, I was struck by the depth of understanding and the quality of the work, both analytical and experimental. These papers represent a tremendous amount of effort—especially when you realize that up until a few years ago, all the computations were done manually with nothing more advanced than a pencil, a slide rule or a mechanical adding machine! Today, personal digital computers, equipped with a variety of software quickly manipulate the most complex expressions. With the software, it's easy for us to examine and manipulate mathematical expressions derived in earlier work and mine them for new understanding and insights. We now have antenna-modeling programs that are nothing short of magical, although their magic must be used with some caution. It's important to not only have a fundamentally solid understanding of antennas, but the modeling programs as well.[1]

[1]Notes appear on page 49.

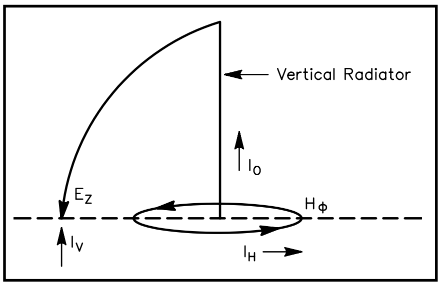

Figure 1—Fields and ground currents near the base of a vertical antenna.

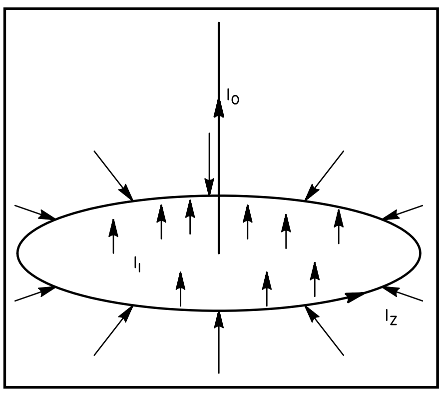

Figure 2—Definition of the current zone near the base of a vertical antenna. I_z represents the total current flowing through a zone at a given radius (r_1) by assuming the current is u niform to a depth of one skin depth (δ) as shown in Figure 13.

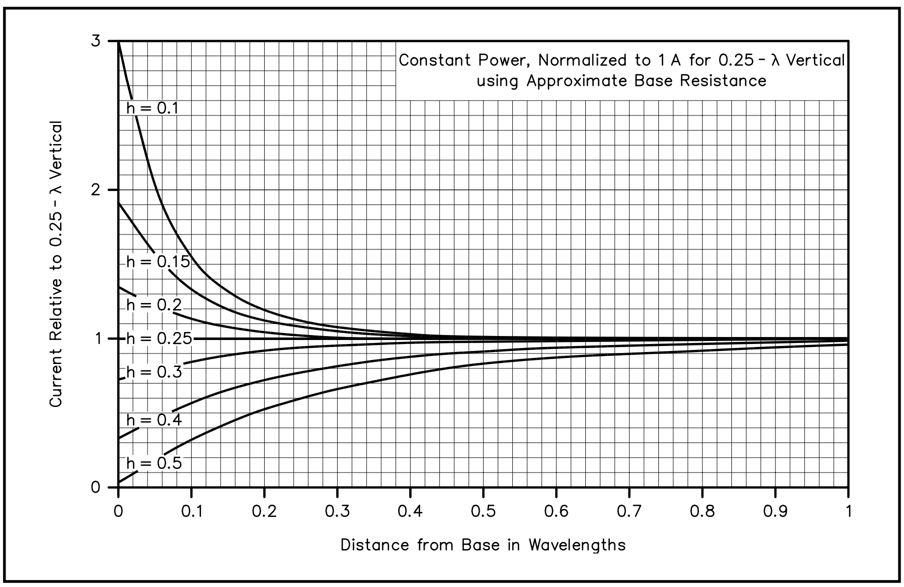

Figure 3—Plot of the current in amperes at the base of a vertical as a function of height and radius in wavelengths. The current in the base of the 0.25-λ antenna is assumed to be 1 A and the currents in the other antennas are adjusted to maintain the same input power.

What follows is a short tour of some of the earlier work that explains some of the lore of verticals and where it came from. I put the math in an Appendix and generated graphs for the discussion. All the graphs were done using a spreadsheet. After reading this article, I recommend you explore for yourself using the equations in the Appendix. The integration of power for Figure 6 was done with *Maple*; *MathCad* or *Mathmatica* would also do fine. You can also do integration with a spreadsheet.[2]

George Brown

In the mid-1930s, radio broadcasting was coming of age and the Institute of Radio Engineers (IRE) proceedings had many papers on vertical antennas and associated ground systems. One of the more influential writers of the time was George H. Brown. A series of papers written by Brown and his colleagues[3-10] at RCA have proved over time to be the most influential. The 1937 IRE paper (see Note 9) has been repeatedly referred to in Amateur Radio publications and is the basis for many later articles. [11-19] (References 16 and 19 have extensive bibliographies for further study.) At the time, these papers were so influential that they became the basis for the FCC standards for broadcast antenna installations! The way we think about verticals today has, in large part, been shaped by this work.

George Brown received his PhD from the University of Wisconsin-Madison in 1933. The core of his dissertation[20] is an analysis of the fields and ground currents associated with a vertical antenna with an extensive buried-radial ground system. This became the basis for much of the work that followed. Brown's work contains a great deal of analysis in addition to experimental results.

Papers on broadcast verticals were not Brown's only contributions to antenna art. He is credited with inventing the ground-plane antenna and wrote numerous other papers on antenna subjects. In later years, Brown was the director of the RCA Sarnoff laboratory. Although not a ham, George Brown contributed enormously to Amateur Radio.

A Closer Look at Verticals

A vertical antenna has two field components that induce currents in the ground around the antenna. Figure 1 shows (in a general way) the electric (E, V/m) and magnetic (H, A/m) field components in the region near the antenna. Because the soil near the antenna usually has a relatively high resistance, both of these field components can induce currents (I_V and I_H) in the ground surrounding the antenna resulting in losses. The worms may enjoy the heated ground, but the power dissipated there subtracts from the radiated power, weakening the signal. As indicated in Figure 1, the tangential component of the H field ($H\phi$) induces horizontal currents (I_H) flowing radially and the normal component of the E field (E_z) induces vertically flowing currents (I_v). Actually, things are a bit more complex than this, but we don't need to thrash that to understand conceptually what's going on. Introducing a system of ground wires, buried or elevated, modifies the current flowing in the ground and (hopefully) reduces loss.

Brown's work was primarily concerned with broadcast antennas in the 0.5 to 1.5 MHz range, although some of his experimental work was carried out at 3 MHz. To make the analysis tractable he made several assumptions:

- The ground system would consist of a large number of radials buried a short distance below the surface.
- The ground characteristics were predominately resistive, ie, dominated by conduction currents, so displacement currents could be ignored.
- Because of the extensive ground screen and its shallow depth, the E-field losses were assumed to be small.

For his work, these assumptions were good approximations, but they are not en-

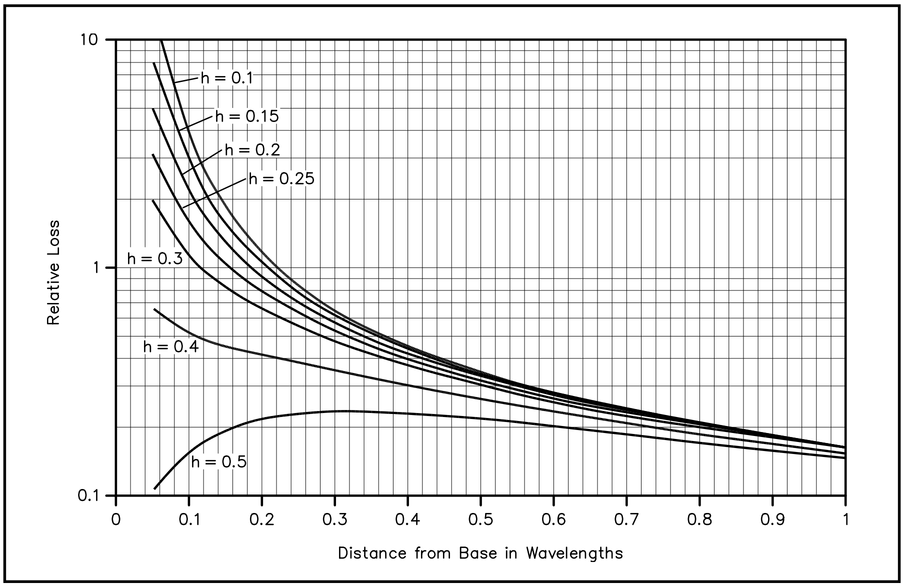

Figure 4—Relative ground loss for several different height verticals. The loss is normalized by allowing the expression which takes into account skin depth and ground conductivity to be equal to 1.

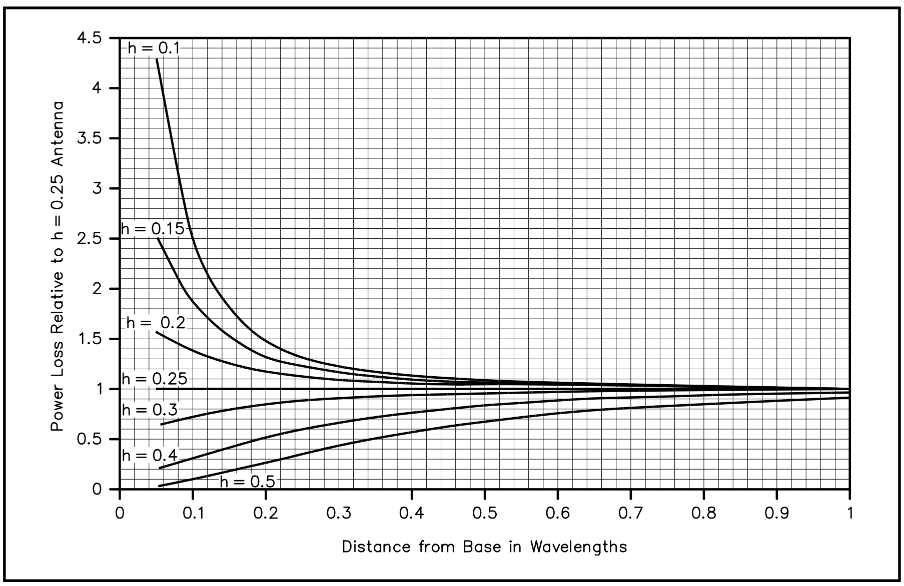

Figure 5—Ground loss at a given radius relative to a 0.25-λ vertical.

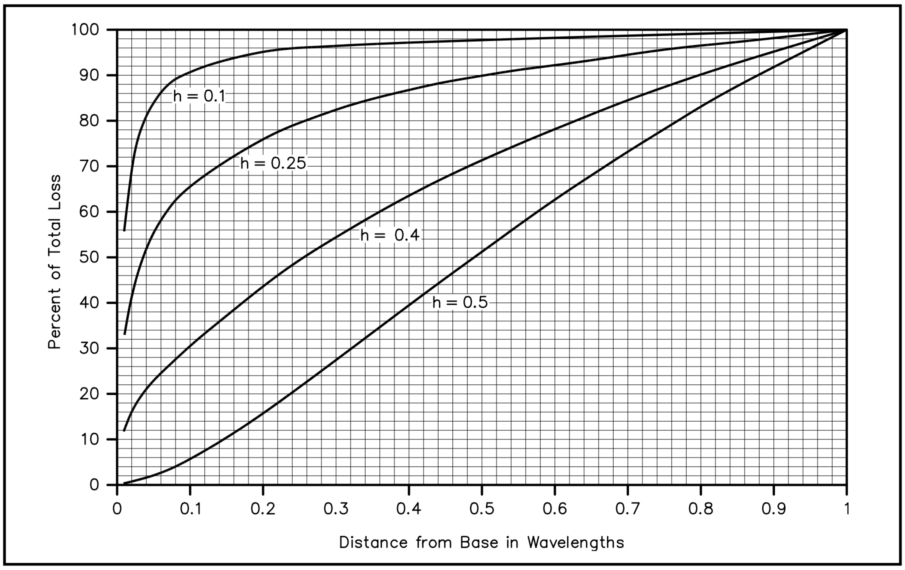

Figure 6--Percent of total ground loss within a given radius (in wavelengths) relative to the total loss at 1-l. This is a measure of the effectiveness of a ground system of a given radius.

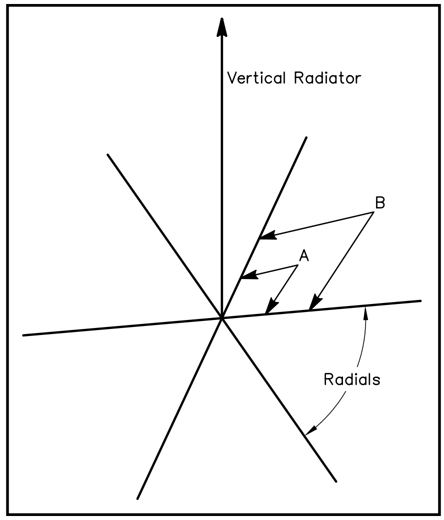

Figure 8—Current entering the ground between radial wires.

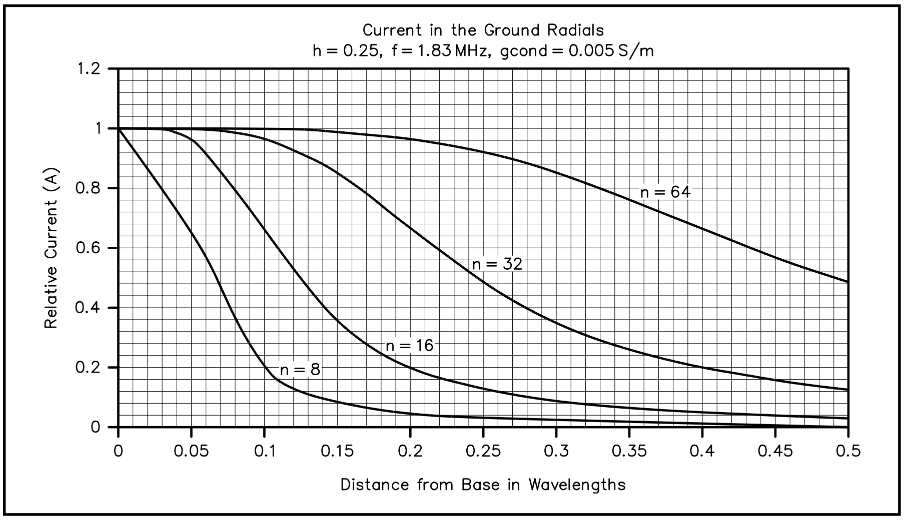

Figure 7—Total current in the radials (I_w) as a function of radius from the base of a 0.25-λ vertical operating at 1.83 MHz and with a ground conductivity of 0.005 S/m (average ground).

tirely valid for HF amateur verticals with small numbers of radials and certainly not valid for elevated radials. Nonetheless, his work is a very good place to start. At the end of the discussion we will look again at these assumptions.

Figure 2 is a sketch of current flow in the antenna and the surrounding ground. I_z represents the total current flowing through a cylindrical zone at a given radius. I_1 represents the current returning to the antenna in addition to the base current. I_o is the current at the base of the antenna. Brown derived an equation (see the Appendix) that describes the ground current as a function of antenna height and distance from the base of the antenna. The heights I will be using in the following discussion are the effective *electrical* heights. For example, if you use some top loading on the vertical, the effective electrical height is greater than the physical height. For the following graphs, I have used simplified expressions that use the effective height. It is important to recognize that simply adding a top hat to a vertical of given physical height can reduce the ground losses. We will be able to see this from the effect of height on ground-current amplitudes. Simply moving a loading coil from the antenna's base further into the antenna reduces ground losses because it reduces ground-current amplitude.

Figure 3 is a graph of this current (I_2)for several effective heights. The currents have been adjusted for constant input power (about 37 W) at the base of the antenna, with 1 A into a 0.25-λ vertical as the reference. This graph clearly shows the high currents flowing in the ground near the base of a short antenna. Compared to a 0.25-λ vertical, the 0.1-λ vertical has three times the ground current; as you further shorten the antenna, the ground current increases rapidly. Keep in mind that the ground loss is proportional to the *square* of the current (I^2R), so the power loss in the immediate region of the base is *much* higher for the shorter antenna.

One way to visualize the relative losses is to calculate them. This is where a spreadsheet really helps. If you take the currents given in Figure 3, square them and divide by the circumference of a circle at a given distance from the base—taking into account the ground resistance and the current's depth of penetration—you know the power loss at a given radius. Figure 4 is a graph of the power loss as a function of the distance from the antenna base. This shows that the losses are high near the base, are greater for shorter antennas and taper off rapidly as distance from the base increases. Note also that for a 0.5-λ vertical, the maximum loss occurs about 0.3-λ away from the base! The ground system in this region may profit from some additional attention. You may ask "Who uses 0.5-λ verticals, especially on 80 or 160 meters?" What about 0.5-λ slopers hung from towers? Even though they are typically not connected directly to ground, they would benefit from a ground system under them. John Devoldere, ON4UN, makes this point in his book (see Note 19). For simplicity, in Figure 4, I have assumed that the depth of current penetration into the soil and the soil conductivity are normalized to 1. For the actual losses in real ground at amateur operating frequencies, the proper equations are in the Appendix if you would like to graph them for yourself. We can also generate a graph showing the loss relative to the 0.25-λ vertical as shown in Figure 5.

Now we can take the next step and integrate the total loss inside a given radius to get a feeling for how large we should make our ground systems. Figure 6 is a graph of the total loss within a given radius, relative to the total loss inside a 1-λ radius for *each* antenna height. I chose the 1-λ radius as the reference because it contains most of the near-field loss and also represents a practical maximum radial length for most installations (560 feet on 160 meters!).The absolute value of the total loss is, of course, higher for a short antenna when compared to a taller one. For the 0.1-λ-high antenna, if we have a good ground screen out to a distance of 0.1-λ, we'll eliminate over 90% of the ground loss! This is where the idea comes from that for short antennas we should concentrate our ground systems inside a short radius. A larger ground system will do no harm; in fact, it reduces the loss even more, but if we have a limited amount of wire, we are much better off to *use many short radials instead of a few long radials*. Note that this graph assumes a large number of radials (more than 100). If only a few radials are used, the effectiveness of the ground system is reduced, although for short antennas it is not necessary to use as large a number of radials.

We can see why this is so by using another of Brown's equations, the one for the current in the radials as a function of radial length and number of radials (see Appendix). Figure 7 is a graph of the current in the radials as you move away from the base of a 0.25-λ vertical with various numbers of radials. The vertical has a 1 A current in the base and (from Figure 3) the total current (I_z) is constant as you move farther out. What we see is the current in the radials (I_w) falling off. The fewer the radials, the more rapidly the current decreases with distance from the base. The total current is still 1 A, but the remainder (I_e) is flowing in the ground and inducing losses. If you use only a few radials it does no good to make them very long because the outer portions of the radials pick up very little current.

What's happening here? Figure 8 is a sketch of a radial system with current entering the ground at two points (A and B). Current reaching the ground at point B has to flow much farther in the soil than current at point A before reaching a radial. The farther from the radiator you go, the greater is the distance between each radial and its neighbor and the farther is the distance the current must flow in the soil. There comes

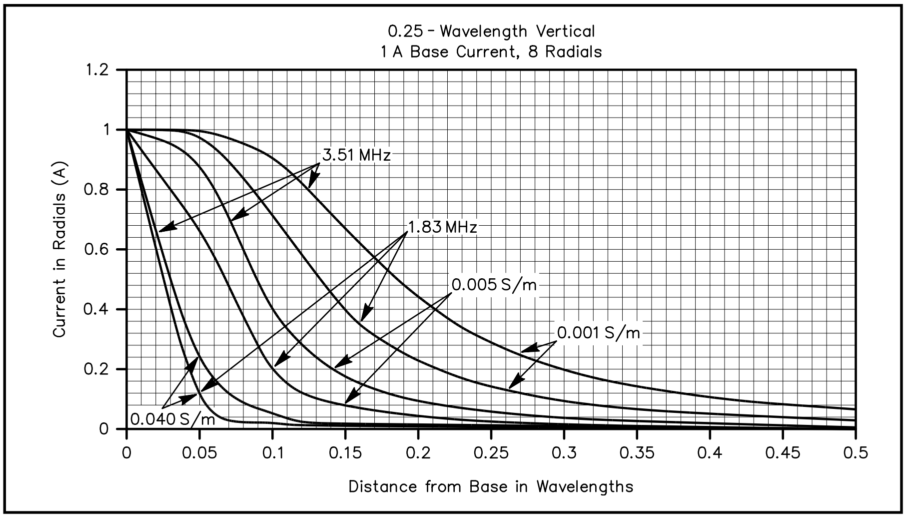

Figure 9—The effect of ground conductivity and frequency on the current in radial wires 1 A of base current and eight radials.

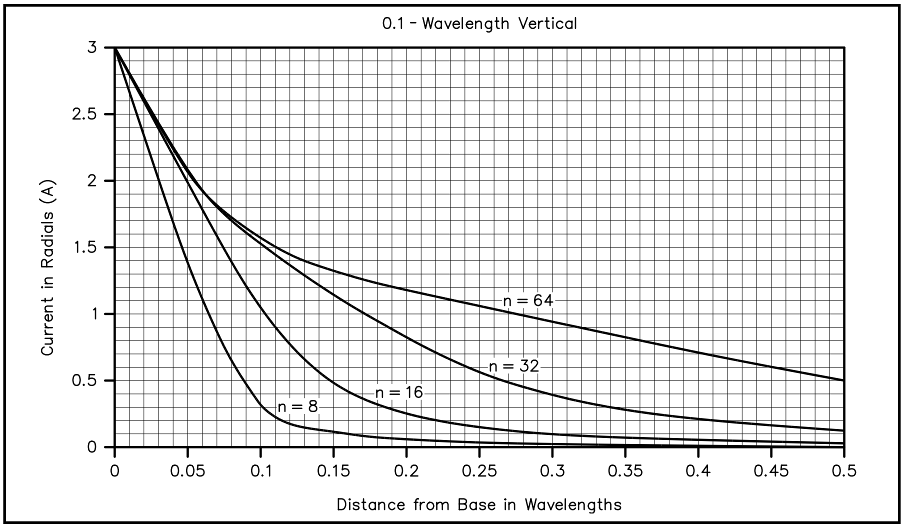

Figure 10—Radial-wire currents of a 0.1-λ vertical for several different numbers of radials (n).

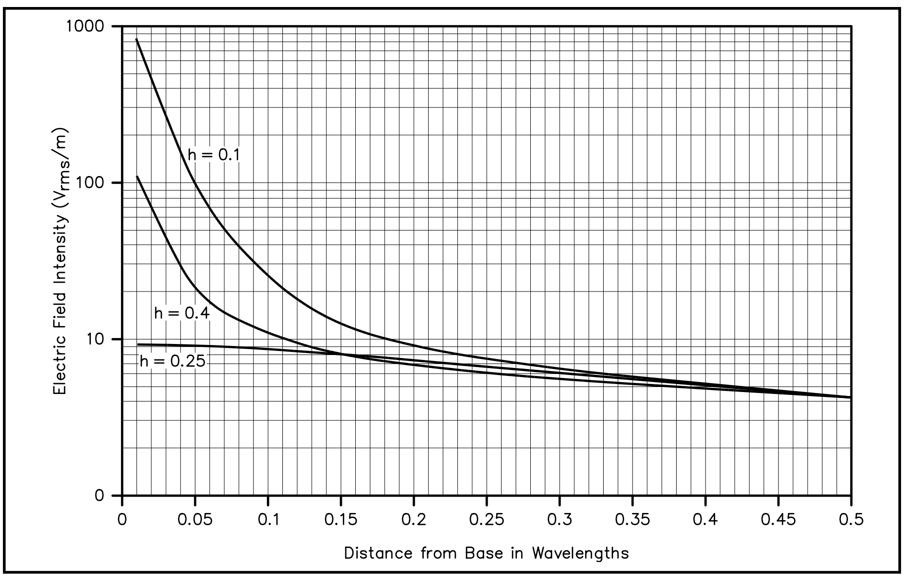

Figure 11—Electric-field intensity near the base of a vertical operating at 1.830 MHz with 1500 W input.

a point where the distance between the radials is so great that the radials are no longer effective. The more radials you use the closer together they will be (at a given radius) and the farther out will be the point at which the radial is no longer effective.

Now that we have Brown's equations in our spreadsheet we can explore further the effects of ground conductivity and frequency on radial number and length. In Brown's time this would have been very laborious, for us it is just a few mouse clicks! Figure 9 is a graph for a 0.25-λ vertical with eight radials, at 1.83 and 3.51 MHz for three different ground conductivities. Notice that as the ground improves (higher conductivity) the current in the radials falls more rapidly. This seems paradoxical: To get the full benefit of the radial system, you have to have more radials as the ground improves! Notice also that as frequency is increased, longer radials can be used effectively.

What about the change in radial current for shorter or longer antennas? That's easy. We just multiply the current values in Figure 3 times the values in Figure 7. Figure 10 is an example for a 0.1–λ vertical. Again we see the advantages of using lots of relatively short radials with a short vertical.

Electric Fields Near the Base

Another consideration is the intensity of the electric field (E) in the region around the base of the antenna. Figure 12 is a graph of E near the base of several verticals of different heights with an input power of 1500 W at 1.830 MHz. Notice how high the field is for the 0.1-λ antenna: about 100 times the value for the 0.25-λ vertical. This is an important consideration for any conductors or structures close to the base of the antenna. Large potentials can be induced into them. These fields can even ignite tall grass! Notice also that as the antenna height exceeds 0.25-λ, the field intensity again increases. The old-fashioned 0.25-λ vertical has many advantages.

A Word of Caution

George Brown's work has proven to be very useful and has been the basis for many articles in amateur publications. However, we have to keep in mind the assumptions Brown made (listed earlier) and remember that his concern was for *broadcast* applications. One assumption he made is that the ground characteristic is primarily resistive. This is a good approximation for most grounds at 160 and even 80 meters, but at higher frequencies, the ground behaves as though there is *capacitance in parallel with the resistance*: ie, there will be displacement as well as conduction currents.

For frequencies above 4 MHz, Brown's equations still give us a good qualitative feeling for what's going on and the overall guidance they offer is still valid. But Brown was careful to point out that you shouldn't rely on the absolute numbers. The need to consider displacement currents can be illustrated by looking at curves for skin depth in soil as a function of frequency and ground characteristics (the generating equations are in the Appendix). Figure 12 is representative of skin depths for typical soils. The graph is an extension of one given in *QST* by Charlie Michaels, W7XC (see Note 18). The dashed lines represent skin depth when conductivity only is considered. The solid lines represent skin depths using the com-

Appendix

Definitions

I_o = current in the base of the antenna or at the current loop in the case of the 1/2λ antenna
I_z = zone current at radius $r_1 = I_w + I_e$
I_e = total current in the earth at radius r_1
I_w = total current in radial wires at radius r_1
f = frequency in Hertz
f_{MHz} = frequency in MHz
E = electric field intensity
h = height of antenna in wavelengths
r_1 = distance from base in wavelengths
s = soil conductivity in Siemens/meter [S/m]
n = number of wires in the radial system
r_2 = radius of radial wires in cm

Zone Currents

$$\frac{|I_Z|}{|I_o|} \equiv I_n = \frac{1}{\sin 2\pi h}\sqrt{\left[\sin 2\pi p \pm \sin 2\pi r_1 \cos 2\pi h\right]^2 + \left[\cos 2\pi p \pm \cos 2\pi r_1 \cos 2\pi h\right]^2}$$

$$p \equiv \sqrt{r_1^2 + h^2}$$

(Equation 1)

Current Distribution in Radial Wires

$$\frac{I_e}{I_w} = j\left(\frac{3.6\sigma\pi^4 r_1^2}{f_{MHz} n^2}\right)\left[\log\left(\frac{3\times 10^4 \pi r_1}{f_{MHz} n r_2}\right) \pm 0.5\right] = j\left|\frac{I_e}{I_w}\right|$$

$$\left|\frac{I_w}{I_z}\right| = \frac{1}{\sqrt{1+\left|\frac{I_e}{I_w}\right|^2}}$$

$$\left|\frac{I_e}{I_z}\right| = \frac{1}{\sqrt{1+\left|\frac{I_w}{I_e}\right|^2}}$$

(Equation 2)

Electric Field Intensity

$$|E| = \left(\frac{2 f_{MHz} I_o}{\sin 2\pi h}\right)\sqrt{\left(\frac{\cos 2\pi \sin 2\pi r}{r_1} \pm \frac{\sin 2\pi p}{p}\right)^2 + \left(\frac{\cos 2\pi h \cos 2\pi r_1}{r_1} \pm \frac{\cos 2\pi p}{p}\right)^2}\left[\frac{V}{m}\right]$$

(Equation 3)

plete equation for skin depth in a general medium. What has been added is the *permittivity of the soil*, which is related to capacitance. For seawater, the conductivity dominates at any frequency below 2 meters. For very good soil, we see that conductivity still dominates over the HF range, but for average or poor soils, the expression for skin depth considering only conductivity gives a depth that is progressively much too large, especially for poor soils. This alters the ground-current distributions from those predicted by Brown; the actual losses may be higher.

If we look at most amateur literature concerning ground characteristics, we see that the emphasis is on measuring ground resistance and the effect of ground resistance on losses, with little said about the permittivity. This is a direct reflection of Brown's work and his concern with broadcast frequencies. We have been following his lead for the last 60 years. In reality, for most soils at HF, we need to take into account the permittivity of ground. Unfortunately, measuring the complex impedance of soil is considerably more difficult than measuring just soil conductivity. W7XC's article partially corrected this and was incorporated in later editions of the *ARRL Antenna Book*, but we still have some work to do.

Brown also assumed that the E-field losses were small. (In his 1935 paper and his thesis, he does compute the electric-field intensity, but then points out that these ground losses are small when a shallow, dense, buried radial system is used with a 0.25-λ vertical. For systems with many buried radials, this is a good approximation. However, when there are only a few radials, or when the radials are elevated above ground, the E-field loss may not be small at all. The importance of E-field losses to amateurs has been pointed out by Clay Whiffen, KF4IX, and Ben Zieg, K4OQK.[21] They showed the increased loss possible when the top of a vertical (where there is a very high electric field) is placed close to a tree. We also know that the outer ends of elevated radials have very high potentials and can induce E-field losses in the ground, grass, shrubs and sod beneath the radial system.

When we compare buried radials with elevated radials we find that the current distribution is very different between the two types of radial systems (see Note 14). Making buried radials longer may not help much if only a few radials are used, but it doesn't hurt. Buried radial systems with a radius greater than 0.5λ can be very effective if enough radials are used. However, as Burke and Miller[22] have shown, making *elevated* radials longer than 0.3λ can lead to greatly increased loss when only a few radials are used. Larger numbers of elevated radials do reduce this loss and allow larger elevated ground systems to be effective. It is important that we *do not directly equate* buried and elevated ground systems on the basis of Brown's work. They are different animals, both of which certainly have their place.

Skin-Depth Equations

The exact expression for penetration or skin depth in a general material is given by:

$$\delta = \left(\frac{\sqrt{2}}{\omega\sqrt{\mu\varepsilon}}\right)\left[\sqrt{1+\left(\frac{\sigma}{\omega\varepsilon}\right)^2} \pm 1\right]^{\pm 1/2} \qquad \text{(Equation 4)}$$

where:

δ = skin depth in meters

$\omega = 2\pi f$

$\mu = \mu_o\mu_r$

$\mu_o = 4\pi 10^{-7}$ Henry/meter

μ_r = relative permeability

$\varepsilon = \varepsilon_o\varepsilon_r$

$\varepsilon_o = 8.85\times10^{-12}$ Farads/meter

ε_r = relative permittivity

For most soils, $\mu_r \approx 1$ (unless you set up shop in an open-pit iron mine!). For good conductors:

$$\frac{\sigma}{\omega\varepsilon} >> 1 \qquad \text{(Equation 5)}$$

Which allows the equation for δ to be simplified to:

$$\delta = \frac{1}{\sqrt{\pi\sigma\mu f}} \text{ m} \qquad \text{(Equation 6)}$$

where:

f is in Hertz

Ground loss

Ground loss for a ring of soil (dr) at a given radius (r_1) from the base can be calculated with the aid of figure 13. If we assume that the average current is uniform to one skin depth (δ), the loss in the ring will be:

$$\frac{dP}{dr} = \frac{I_e^2}{2\pi\delta\sigma r} = \frac{f_{MHz} I_e^2}{600\pi\delta\sigma r_1} = \left(\frac{f_{MHz}}{300\delta\sigma}\right)\left(\frac{I_e^2}{2\pi r_1}\right)\left[\frac{W}{m}\right] \qquad \text{(Equation 7)}$$

where:

δ and r are in meters and r_1 is in wavelengths (λ).

$$\lambda = \frac{300}{f_{MHz}}\,[\text{m}] \qquad \text{(Equation 8)}$$

The graph in Figure 4 assumes that

$$\frac{f_{MHz}}{300\delta\sigma} = 1 \qquad \text{(Equation 9)}$$

and that r_1 is in wavelengths.

A Final Word

I hope you will find this information useful. If you really want a thorough understanding of the topic, you should graph these equations yourself and read the listed references.[22] The *QST*, *ham radio* and *CQ* articles are quite easy to follow; even Brown's papers are no great chore to read. Some modeling with *NEC* or *MININEC* software will give you even more insight. Particularly on

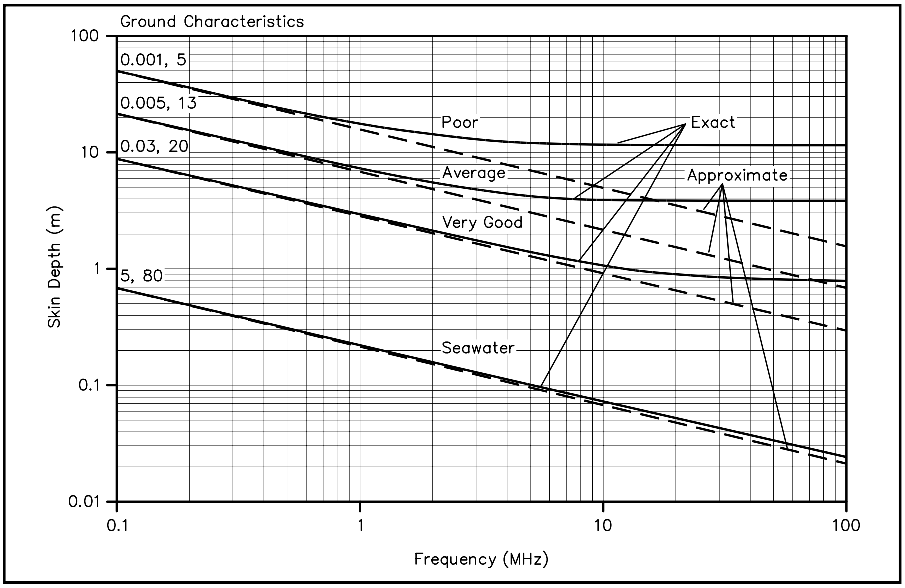

Figure 12—Skin depth in soil of various characteristics as a function of frequency.

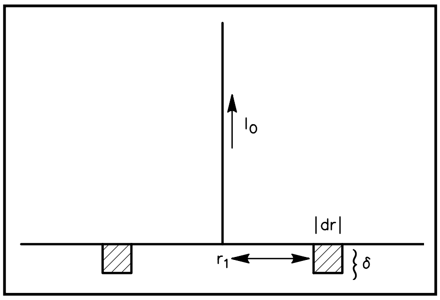

Figure 13—Calculation of ground loss in a small ring of soil at a given radius.

the lower bands, verticals can be very effective, but you have to understand what you are about to get good results.

Acknowledgement

The equations used here have been taken directly from Brown's 1935 (see Note 4) and 1937 (see Note 9) papers. All I have done is to restate them in a form handy for spreadsheet manipulations and to repeat some of his observations and conclusions.

Notes

[1]See also L. B. Cebik, W4RNL, "*NEC-4.1: Limitations of Importance to Hams,*" *QEX*, May/Jun 1998, pp 3-16 and John Rockway and James Logan, N6BRF, "Wire Modeling Limitations of *NEC* and *MININEC* for *Windows*," *QEX*, May/Jun 1998, pp 17-21.

[2]Byron Gottfried, *Spreadsheet Tools For Engineers*, McGraw-Hill Basic Engineering Series Tools, 2000, see Chapter 9. A compressed *Excel* file (*SEVERNS.ZIP*) is available from **http://www.arrl.org/files/qst-binaries/**.

[3]G. H. Brown and Ronold King, "High-Frequency Models in Antenna Investigations," *IRE Proceedings*, Vol 22, No. 4, Apr 1934, pp 457-480.

[4]George H. Brown, "The Phase and Magnitude of Earth Currents Near Radio Transmitting Antennas," *IRE Proceedings*, Vol 23, No. 2, Feb 1935, pp 168-182.

[5]H. E. Gihring and G. H. Brown, "General Considerations of Tower Antennas for Broadcast Use," *IRE Proceedings*, Vol 23, No. 4, Apr 1935, pp 311-356.

[6]G. H. Brown, "A Critical Study of The Characteristics of Broadcast Antennas as Affected by Antenna Current Distribution," *IRE Proceedings*, Vol 24, No. 1, Jan 1936, pp 48-81.

[7]G. H. Brown, "Directional Antennas," *IRE Proceedings*, Vol 25, No. 1, Part 1, Jan 1937, pp 78-145.

[8]G. H. Brown and John G. Leitch, "The Fading Characteristics of the Top-Loaded WCAU Antenna," *IRE Proceedings*, Vol 25, No. 5, May 1937, pp 583-611.

[9]G. H. Brown, R. F. Lewis and J. Epstein, "Ground Systems as a Factor in Antenna Efficiency," *IRE Proceedings*, Vol 25, No. 6, Jun 1937, pp 753-787.

[10]G. H. Brown, "A Consideration of the Radio-Frequency Voltages Encountered by the Insulating Material of a Broadcast Tower Antenna," *IRE Proceedings*, Vol 27, No. 9, Sep 1939, pp 566-578.

[11]F. R. Abbott, "Design of Optimum Buried-Conductor RF Ground Systems," *IRE Proceedings*, Vol 30, No. 7, Jul 1952, pp 846-852.

[12]R. E. Leo, W7LR, "Vertical Antenna Ground Systems," *ham radio magazine*, May 1974, pp 30-35.

[13]John O. Stanley, K4ERO/HC1, "Optimum Ground Systems for Vertical Antennas, *QST*, Dec 1976, pp 13-15.

[14]Jerry Sevick, W2FMI, "Short Ground-Radial Systems for Short Verticals," *QST*, Apr 1978, pp 30-33.

[15]Archibald C. Doty, K8CFU (now W7ACD), John A. Frey, W3ESU, and Harry J. Mills, K4HU, "Efficient Ground Systems for Vertical Antennas" *QST*, Feb 1983, pp 20-25.

[16]Archibald C. Doty, K8CFU, "Improving Vertical Antenna Efficiency," *CQ*, Apr 1984, pp 24-31.

[17]Brian Edward, N2MF, "Radial Systems for Ground-Mounted Vertical Antennas," *QST*, Jun 1985, pp 28-30.

[18]C. J. Michaels, W7XC, "Some Reflections On Vertical Antennas," *QST*, Jul 1987, pp 15-19.

[19]John Devoldere, ON4UN, *Low-Band DXing*, (Newington: ARRL, 3rd ed), 1999, Chapter 9.

[20]G. H. Brown, "A Theoretical and Experimental Investigation of The Resistance of Radio Transmitting Antennas," PhD thesis, University of Wisconsin-Madison, June 29, 1933

[21]Clay Whiffen, KF4IX, and Ben Zieg, K4OQK, "Trees and Verticals," Technical Correspondence, *QST*, Nov 1991, p53

[22]G. J. Burke and E. K. Miller, "Numerical Modeling of Monopoles On Radial-Wire Ground Screens," *IEEE Antennas and Propagation Society International Symposium Proceedings*, Jun 1989, pp 244-247

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The Case of Declining Beverage-on-Ground Antenna Performance

Over the course of two winter seasons, the performance of this antenna dropped off dramatically. Here's why.

Rudy Severns, N6LF

In midsummer 2013, I placed 450 feet of insulated wire in my pasture as a Beverage on the Ground (BOG) receiving antenna. I was using it to receive WSPR signals, which provide signal-to-noise ratio (S/N) estimates. Over time, and especially during two intervening wet winter seasons, I noticed that received signals dropped off by 15 dB and the S/N of the BOG was no better than that of a simple vertical. I carefully checked connections, feed lines, and associated hardware, but found no problems. This rather radical decline in performance seemed a mystery.

There has been skepticism regarding the validity of using *NEC* modeling of antennas with wires close to ground. Over the years, I've often compared my modeling predictions with finished antennas and generally found very good correlation, but the skepticism prompted me to ask if the latest version of *NEC* (*NEC-4.2*) models antennas with wires close to, or buried in, soil well enough to explain my declining BOG performance. Thus I needed to experimentally validate *NEC-4.2* modeling before I could confidently move on to my BOG problem. I performed a series of experiments with actual antennas having wires parallel to the soil at low heights, and wires buried in soil. I measured the feed-point impedances and soil electrical characteristics.

I'm interested in antennas for 80, 160, and 630 meters, so my test frequencies ranged

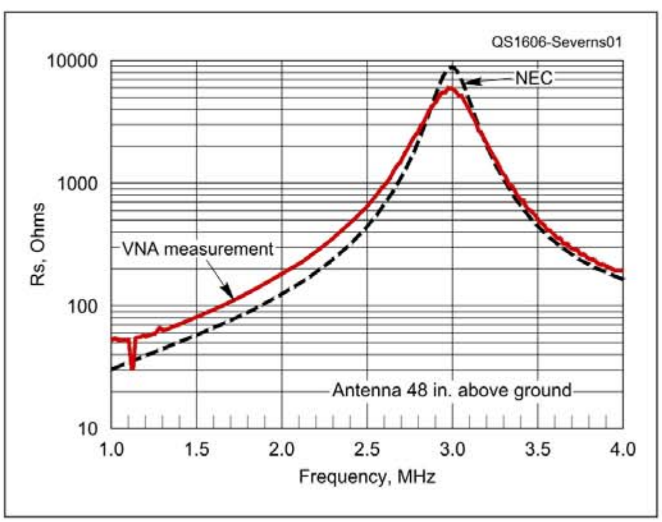

Figure 1 — Measured (red) and modeled (dashed) feed-point resistance of a 300-foot center-fed dipole 48 inches above ground.

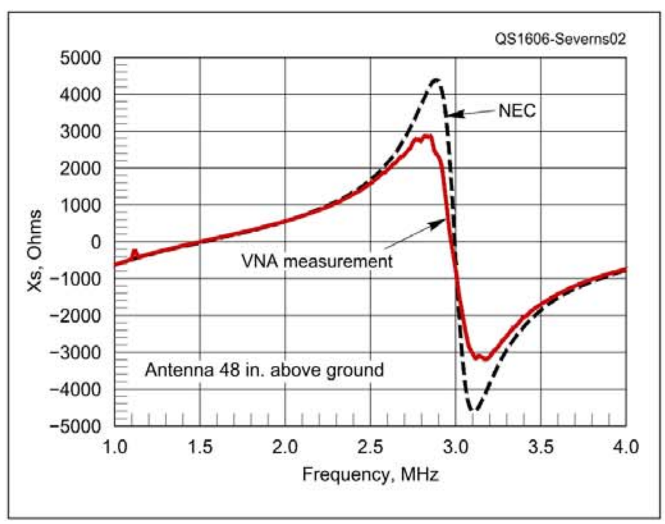

Figure 2 — Measured (red) and modeled (dashed) feed-point reactance of a 300-foot center-fed dipole 48 inches above ground.

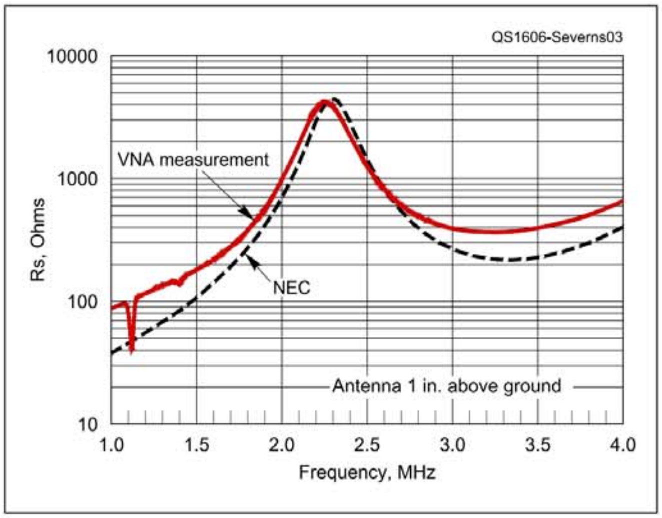

Figure 3 — Measured (red) and modeled (dashed) feed-point resistance of a 300-foot center-fed dipole 1 inch above ground.

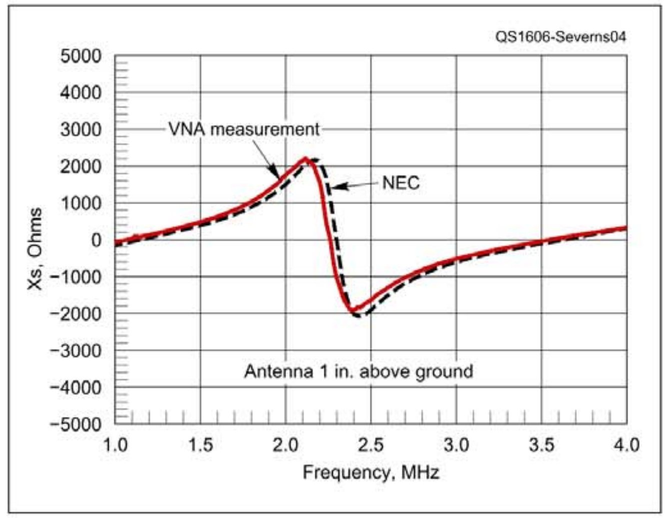

Figure 4 — Measured (red) and modeled (dashed) feed-point reactance of a 300-foot center-fed dipole 1 inch above ground.

from 400 kHz to 4 MHz. The examples do not cover all possibilities, but they're representative. Here is what I found.

Modeling and Instrumentation

NEC solves for the currents on the wires, and from these currents calculates both the feed-point impedance and the radiation pattern. If the impedances calculated from the *NEC* model agree with the values measured on the actual antenna over a wide range of frequencies, you can be reasonably sure the modeling is reliable. In the case of my BOG, which was partially buried, it was helpful to see if *NEC* could predict the current distribution along the wire at a given frequency (1.83 MHz).

For the modeling I used *EZNEC* and *Auto EZ*.[1, 2] I used a vector network analyzer (VNA) to measure impedances. In the graphs, I've made it a point to display the raw measurements without any corrections to the data points. You can see noise present on the graphs, particularly at frequencies associated with local broadcast stations. I measured the soil electrical characteristics concurrently with the feed-point impedances. The following discussion addresses *NEC-4.2* only, because *NEC-2* does not allow buried wires and does not handle wires close to ground very well.

Test Antenna #1

The first test antenna was a low center-fed dipole. I chose a 300-foot length that allowed the antenna to present both odd (like a series resonance) and even (like a parallel resonance) multiples of a half wavelength, presenting a range of high and low feed-point impedances. I varied the height above ground from 48 inches (see lead image) down to 1 inch. Figures 1 and 2 show the comparison between *NEC* predictions and measured feed-point resistance and reactance for the antenna 48 inches above the ground. Figures 3 and 4 repeat the comparisons 1 inch above ground. Evidently, *NEC* does a very good job modeling the impedance of an antenna down to 1 inch above ground.

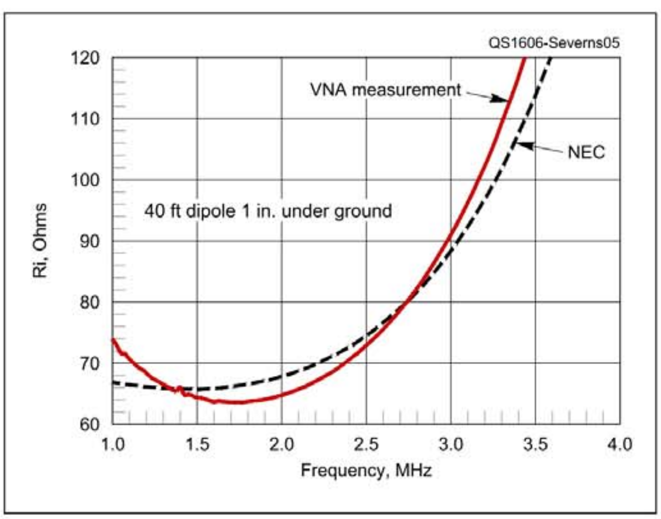

Figure 5 — Measured (red) and modeled (dashed) feed-point resistance of a 40-foot center-fed dipole made from insulated #26 AWG wire buried 1 inch below ground.

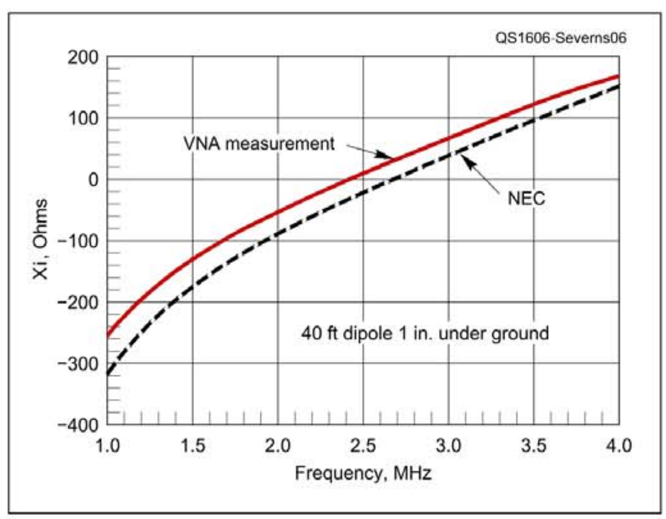

Figure 6 — Measured (red) and modeled (dashed) feed-point reactance of a 40-foot center-fed dipole made from insulated #26 AWG wire buried 1 inch below ground.

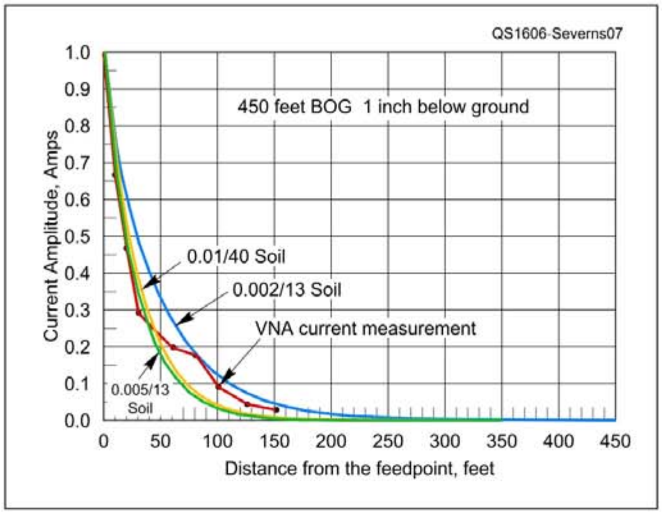

Figure 7 — Measured (red) and modeled (dashes) RF current amplitude at 1.83 MHz along the 450-foot BOG antenna buried 1 inch below ground.

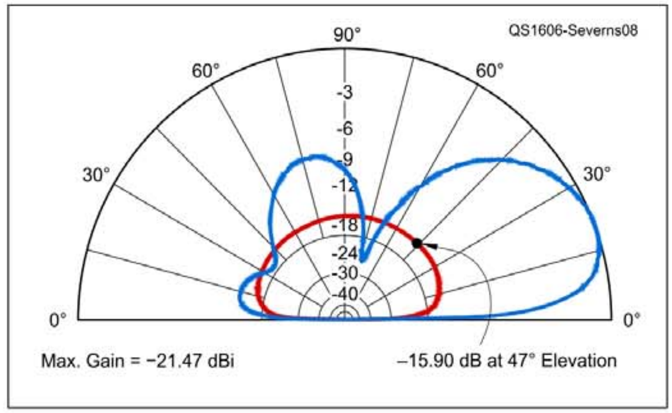

Figure 8 — Elevation radiation patterns for the BOG 1 inch above ground (blue) and 1 inch below ground (red).

Test Antenna #2

My second test antenna was a 40-foot-long dipole made from #26 AWG insulated wire, and buried 1 inch below ground surface. Although only 13% as long as test antenna #1, the resonant frequency is still close to that of the 300-foot dipole above ground. Placing an antenna very close to or in the soil drastically reduces the resonant frequency. Figures 5 and 6 show the resistance and reactance results for antenna #2. The correlation between *NEC* calculations and VNA based measurements indicate that the modeling provides reasonable predictions.

Test Antenna #3

I wanted to test an antenna that incorporated a ground rod to get a feeling for how well ground rods are modeled. I have a pair of tall wooden support poles, so I simply suspended a vertical 77-foot length of insulated wire from the midpoint of a Dacron line stretched between the poles. Overall agreement was very good and the predicted resonant frequency was particularly close.

Test Antenna #4

This entire exercise was prompted by the mysterious declining performance of my BOG, so my final test antenna was the BOG itself. I measured feed-point impedances from 400 kHz to 4.4 MHz, as well as the current amplitude and phase along the wire at 1.83 MHz. Again, the impedance measurements were close to predictions. I added the current measurements as a further confirmation of *NEC* modeling. I modeled the antenna 1 inch below ground to account for its actual burial over time. A comparison between modeling and measurement for the current distribution is seen in Figure 7. The correlation between measurements and modeling is good despite the inherent uncertainties in the ground surface modeling and the variation in soil electrical characteristics along the 450-foot length of the antenna.

If the impedances calculated from the *NEC* model agree with the values measured on the actual antenna over a wide range of frequencies, you can be reasonably sure the modeling is reliable.

The rapid exponential decay of the antenna current predicted by *NEC* was a surprise to me, but field measurements confirmed it. This goes a long way towards explaining why the performance became so poor. Functionally, the partially buried BOG behaves more like a radial than an antenna! Disconnecting the ground rod at the far end had no effect on either current distribution or feed-point impedance, because there was very little if any current at the far end of the antenna.

At this point, I modeled the BOG at heights of 1 inch below and 1 inch above the ground to approximately represent the changes from the time the BOG was first installed to the present, and to see how the radiation patterns might have changed. Those patterns are compared in Figure 8.

Figure 8 solves the initial mystery. The larger (blue) pattern represents the initial condition of the antenna 1 inch above ground. The receiver directionality factor (RDF) is 12 dB and a peak gain Gp is –21.5 dB. The smaller pattern (red) represents the present BOG condition 1 inch below the ground. The RDF is 6 dB and Gp is –37.4 dB. These patterns make it clear just how severely the performance declined as the BOG gradually sank into the sod and soil through two winters. The differences in the pattern shown in Figure 8 correlate well with the S/N estimates observed over time.

Conclusions

Correlation between measurements and modeling was excellent in the four examples. These representative examples cover a range of practical cases using very low and buried wires. From this and earlier work, I conclude that *NEC* modeling gives reliable results if we use *NEC-4.2*, follow *NEC* modeling guidelines closely, make sure that the model is dimensionally as close as possible to the actual antenna, and we perform careful soil measurements.[3, 4, 5] The practical limitations of *NEC* modeling are not computational shortcomings in the *NEC* code. What limits us is our knowledge of the details of the actual antennas and the associated soil characteristics, and our ability to replicate these in a model. A more detailed study is available on the *QST* in Depth web page.[6]

As a practical matter we can never be perfect, but careful modeling will get us close. The BOG can be a very useful receiving antenna, but your results may vary. If your BOG is slowly covered by what grows around it, or what falls from the sky, you might see significant degradation over time. The cure is simple. Inspect the antenna regularly and pull it out of the weeds as needed.

Acknowledgments

This work was prompted by an e-mail from Al Christman, K3LC, relaying a question from Carl Luetzelschwab, K9LA, regarding the reliability of *NEC* modeling for wires close to or on the surface, or buried. I express my sincere appreciation to both Roy Lewallen, W7EL, and Dan Maguire, AC6LA, for the use of their latest software. I also thank Don, N4DJ; Greg, W8WWV, and Carl, K9LA, for reading and commenting on earlier versions of this article.

Notes

[1]Several versions of *EZNEC* antenna modeling software are available from developer Roy Lewallen, W7EL, at **www.eznec.com**.

[2]*AutoEZ* automates use of *EZNEC*; see **www.ac6la.com**.

[3]Rudy Severns, N6LF, "Experimental Determination of Ground System Performance for HF Verticals," *QEX*, in seven parts, Jan/Feb 2009 pp 21 – 25 and pp 48 – 52, Mar/Apr 2009 pp 29 – 32, May/Jun 2009 pp 38 – 42, Jul/Aug 2009 pp 1 – 3, Nov/Dec 2009 pp 19 – 24, and Jan/Feb 2010 pp 18 – 19.

[4]Rudy Severns, N6LF, "An Experimental Look at Ground Systems for HF Verticals," *QST*, Mar 2010 pp 30 – 33.

[5]Rudy Severns, N6LF, "A Closer Look at Vertical Antennas With Elevated Ground Systems," *QEX*, Part 1 Mar/Apr 2012 pp 32 – 44, Part 2 May/Jun 2012 pp 24 – 33.

[6]**www.arrl.org/qst-in-depth**

Photo courtesy of the author.

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New Products

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The Pico Balun from SOTABEAMS is a 1:1 current balun or 4:1 matching transformer and antenna feed point assembly available in kit form. The transformer is mounted on a printed circuit board that allows easy attachment of the antenna and feed line. Specifications include a common mode impedance of more than 1000 Ω from 5 MHz to 30 MHz, 10 W power rating, and weight of about 1 ounce. Price: about $9. For more information, visit **www.sotabeams.co.uk.**

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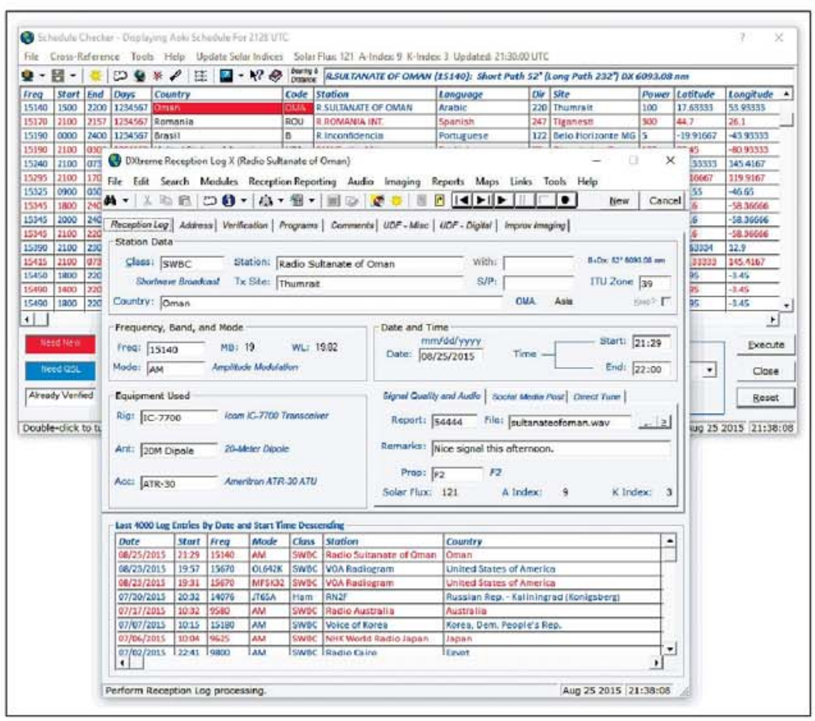
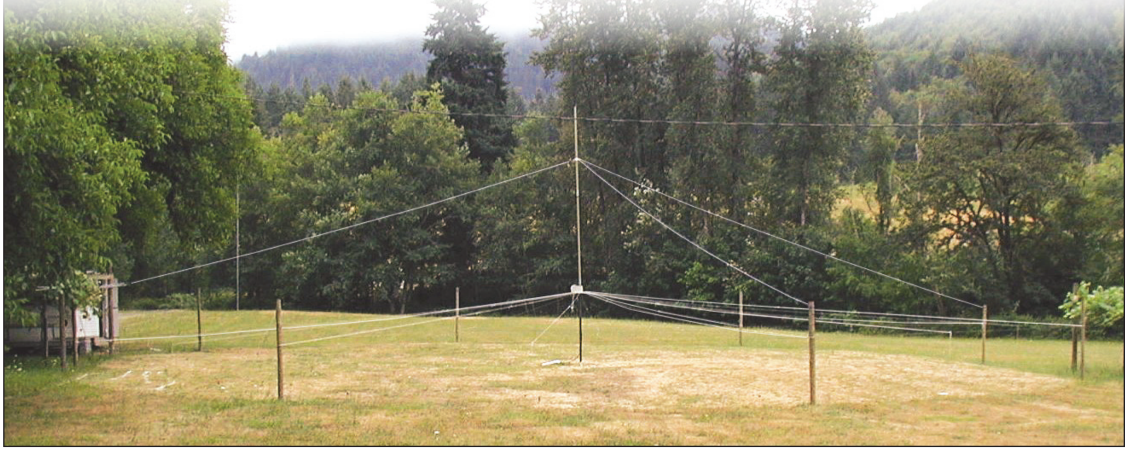
DXtreme Reception Log X (version 10) is a *Windows* logging program for radio monitoring enthusiasts. In addition to the usual logging functions, the software integrates Afreet *Ham CAP* for propagation predictions. The program also features an integrated QSL imaging facility that allows users to scan physical QSL cards and capture electronic QSLs received over the Internet. A dedicated tab on the Reception Log window allows users to associate photos or other images with log entries. An embedded audio feature lets users maintain an audio archive of stations heard. The Schedule Checker lets users import schedules from Aoki, EiBi and FCC AM websites and filter and display the data. The software includes a customized paper and e-mail reception report generator. Price: $89.95. Upgrade pricing is available. For more information, or to order, visit **www.dxtreme.com**.





An Experimental Look at Ground Systems for HF Verticals

In this groundbreaking work we obtain definitive results on ground system effectiveness.

Rudy Severns, N6LF

It's been over 100 years since Marconi used vertical antennas. With such a long history it would seem unlikely that anything new could be said about them. The way Amateur Radio operators use and implement vertical antennas often differs from commercial or military practice leaving amateurs with unanswered questions.

These questions can be addressed analytically or through the use of modeling and simulation, but for most of us neither is quite convincing. Actual measurements on real antennas are a lot more satisfying, at least to verify the modeling.

Some years ago, Jerry Sevick, W2FMI, (SK) published exactly this kind of information in *QST*.[1-5] Reading his articles inspired me to take another experimental look at HF ground systems. The result was an 18 month effort, partly replicating Jerry's work, but also addressing other questions such as the comparison between ground surface and elevated radial systems. These experiments have been covered in detail in a series of seven *QEX* articles. Since not everyone wants all the gory details, this article is a summary of the more interesting results.[6]

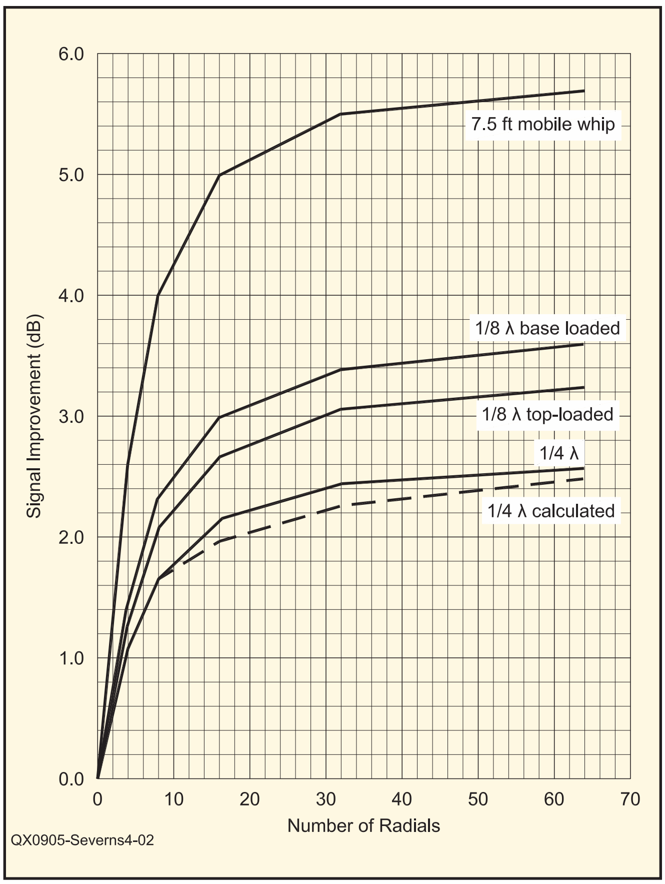


Figure 1 — Typical improvement in signal as ¼ wave radials are added to the basic ground system of a single ground stake.

Near and Far

It is important to keep in mind the role of the ground system associated with the radiation from a vertically polarized antenna. The radiation pattern for a vertical is strongly influenced by the characteristics of the soil in the neighborhood of the antenna. This is particularly true at lower angles for which the pattern is determined by soil characteristics out to a great distance (many wavelengths), often referred to as the *far-field* region.[7] As a practical matter we can't usually do much about conditions beyond perhaps ½ wavelength from the base of the vertical, other than select our location — we simply have to accept what's out there. We can, however, do a lot to reduce the losses in the immediate vicinity

[1]Notes appear on page 33.

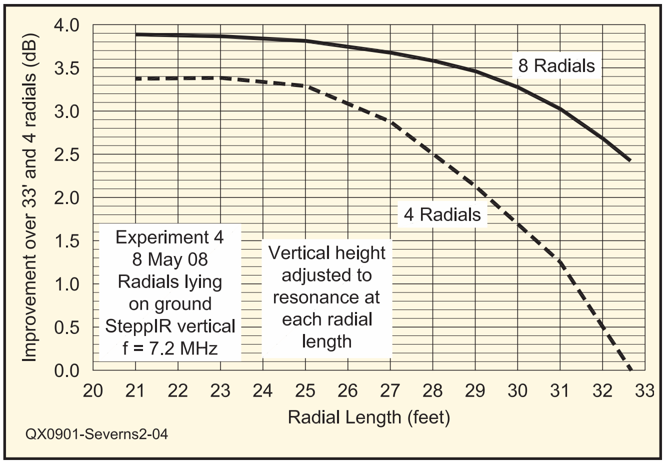


Figure 2 — Effect on signal strength of shortening radial lengths. The 0 dB reference is four 33 foot radials.

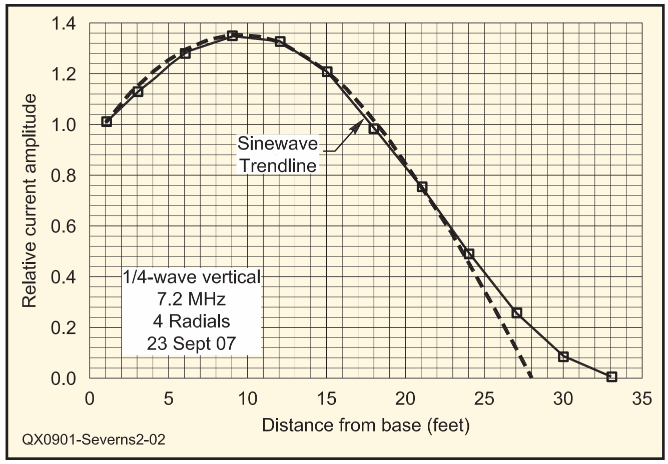


Figure 3 — Measured current distribution on a radial.

of the antenna (the *near-field* region), where the losses can be very high.[8] The purpose of the ground system is to reduce these near-field losses, increasing efficiency and allowing us to radiate as much of the antenna input power as possible, which ultimately improves our signal.

Overview of the Experiments

This work started with a 160 meter vertical with which I varied the number of ¼ wave radials and measured the change in signal strength for a fixed input power. This was interesting and educational but I realized that repeatedly laying down and picking up some 8000 feet of #12 AWG wire was not practical for more extensive investigations. I thus changed the test frequency to 7.2 MHz initially, and later added experiments for multiband ground systems (40 through 10 meters). This initial experiment also stimulated me to use the much more accurate measurement procedure that is outlined in the sidebar on the QST In Depth Web site.[9]

I went through several rounds of experiments, each one answering some questions but, of course, always generating more. In the following three sections we'll consider radials for vertical monopoles — on and above the ground and finally, radial systems for multiband verticals.

Round One — Radials on the Ground

This set of experiments used four different antennas: a ¼ wave vertical, an ⅛ wave vertical with base loading, an ⅛ wave vertical with sufficient top loading to be resonant at 7.2 MHz and a 40 meter mobile whip. I started with a single 4 foot ground stake (zero radials) and then progressively added ¼ wave radials, measuring the changes in signal strength with each increase in radial number. The results are shown in Figure 1. Note that the graph is in terms of the *improvement* in signal for a given input power for *each* antenna over the single ground stake with no radials. The graph does *not* compare the relative merit of each antenna. Obviously a short, lossy mobile whip will yield less signal, typically 10 dB less, than a full size ¼ wave vertical. The signal improvement metric gives us a direct idea of how much is gained for a given improvement in the ground system.

How Many Radials?

This graph shows several things. First it makes clear just how important a radial system is. It can make a difference of many dB in our signal strength. Keep in mind that the soil over which the experiments were done would be classified as good to very good. Over average or poor soils the signal improvements could be many dB greater than shown here. The second thing the graph shows is the point of diminishing returns. Laying down a system with at least 16 radials will give you most of the obtainable improvement. As we go to 32 and then 64 radials the improvement gets progressively smaller. It's arguable that the improvement from going from 32 to 64 radials is worth the cost and clearly the standard 120 radial BC ground system would be overkill.

A final point the graph makes is that *the shorter and more heavily loaded your vertical, the more you have to gain from improving the ground system.* The shorter the vertical, the higher will be the field intensity (for a given input power) in the near field of the antenna and the lower will be the radiation resistance. This leads to much higher ground losses, which translates to more improvement when you reduce these losses by improving the ground system.

How Long Should They Be?

Radials ¼ wave in length are known to be effective in ground systems, but I wondered what the penalty would be from using shorter radials. I was expecting to see a fairly uniform decrease in signal strength (due to an increase in ground loss) as the radials were shortened. That is *not* what I found. Figure 2 shows the results of an experiment in which I measured the signal strength while progressively shortening the radials in four and eight radial systems.

Surprisingly, shortening the radial lengths *increased* the signal strength — not by just a little bit, but by more than 3 dB. This is certainly counterintuitive, but I was seeing clues that helped explain what was happening. I noticed that with only the ground stake the resonant frequency of the vertical was much lower than expected and, as I added more radials, the resonant frequency increased slowly. Most of the change occurred between 4 and 16 radials and had pretty much leveled out by the time I had 64 radials. This suggested to me that the radials might be self-resonant below 7.2 MHz. To check this out I measured the current distribution on a radial and found it to be sinusoidal. The results are shown in Figure 3.

The maximum current point has been moved from the base of the antenna out onto the radials and this substantially increases the ground loss. The radials are resonant below the band and this affects the antenna. A wire, close to ground, can be heavily loaded by the ground, decreasing its resonant frequency. The extent of the loading will depend on the characteristics of the soil. Figure 3 shows that the maximum current point is 10 to 11 feet away from the base. Looking at Figure 2 we see that the maximum signal occurs when we have shortened the radial by this amount.

Figure 3 also illustrates a difference

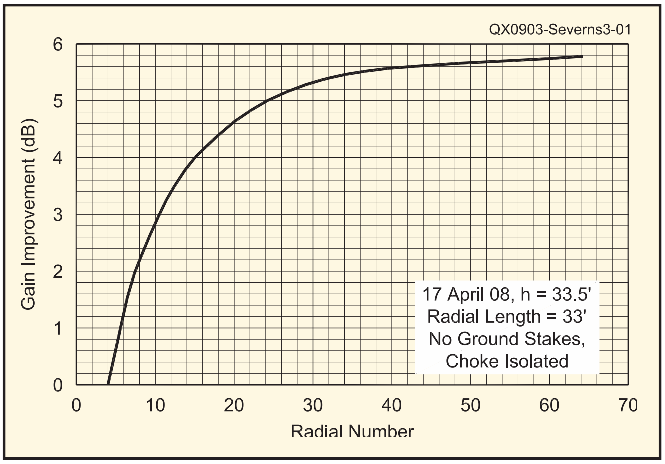


Figure 4 — Signal improvement as a function of radial number. All radials lying on the ground surface, F = 7.2 MHz.

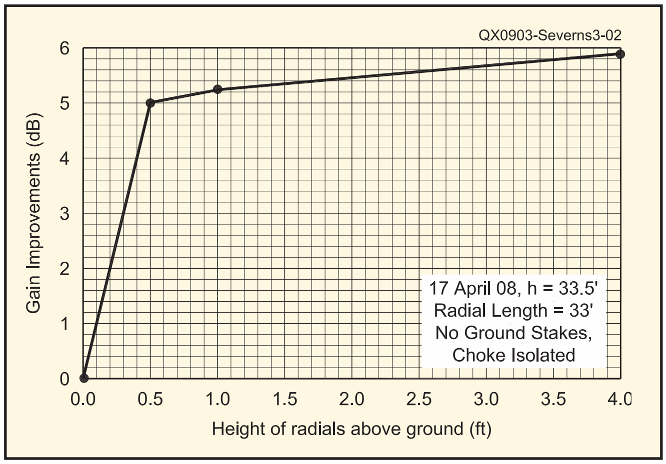


Figure 5 — Signal improvement with four radials and the antenna base at different heights. F = 7.2 MHz.

Table 1

Relative Signal Strengths for 4, 8, 16 and 32 Radials, Comparing Lengths of 33' and 21'

Number of Radials	*Normalized to Four 33' Radials (dB) 33' Radials*	*Normalized To Four 33' Radials (dB) 21' Radials*	*Gain Change (dB)*
4	0	3.08	+3.08
8	2.26	3.68	+1.42
16	3.76	3.95	+0.19
32	4.16	4.04	–0.12

between buried bare wire radials and radials lying on or very near the surface of the soil. The current distribution on a buried bare radial will usually decrease exponentially from the base regardless of its length.[10] You will not see the standing wave shown in Figure 3 except in very poor soils. The insulated radial lying on the ground surface behaves much more like a radial in an elevated radial system in that it has a sine wave-like current distribution. A buried insulated wire will be somewhere in between these two cases depending on the burial depth and soil characteristics.

You can also see in Figure 2 that the signal increases as the radial numbers increase. To check this out I extended the experiment to 32 radials, comparing 33 to 21 foot radials. The results are given in Table 1.

The results in Table 1 indicate that the excess loss due to radial resonance has pretty much disappeared by the time you reach 16 radials. This leads to some advice — rather than trying to determine the optimum radial length, which will vary with every installation due to soil differences, just use at least 16 radials. If you are limited by the total amount of wire available, you're better off to use a larger number of shorter radials rather than a few long ones.

I didn't have time to run an extensive set of experiments comparing different radial length and radial number combinations (each with the same total length of wire), but I did model that situation with *EZNEC*.[11] The modeling predicted, particularly with short verticals, that it was often advantageous to reduce the length of the radials and increase their number. The modeling showed that there is a correlation between vertical height and optimum radial lengths. More details can be found in the modeling report and in the work of others.[12-15]

Round Two — Elevated Radials

Over the past few years there has been a lot of discussion about the relative merits of ground systems using a large number of surface or buried radials versus only a few elevated radials. This stems from *NEC* modeling that indicated that four radials elevated 8 feet or so above ground could be just as effective as 120 buried radials. Many of us, including me, simply could not believe that.

I decided the best way to address this question would be to directly compare two antennas, one with a large number of ground radials and the other with only a few elevated radials. The same antenna was used in both cases, a simple ¼ wave vertical. For the surface tests I used ¼ wave radials and varied the number from 4 to 64. For the elevated tests I used four ¼ wave radials. The elevated radials were placed at 0, 6, 12 and 48 inches above ground. The results are shown in Figures 4 and 5. The 0 dB point in the graphs is normalized to the signal strength for the case of four ¼ wave radials lying on the surface (0 dB). What you see in the graphs is the improvement as you either add more surface radials or elevate the antenna and the four radials above ground.

The most striking thing shown by the graphs is that four elevated radials at a height of 48 inches are within 0.2 dB of 64 radials lying on the ground. This would seem to support the predictions from *NEC* modeling. A detailed view of the results with different elevated configurations is provided on the QST In Depth Web site.

Round Three — Multiband Ground Systems

While single band verticals are frequently used, multiband verticals are even more popular but I'd not seen any experimental work related to multiband ground systems. So I did some. The experiments were performed in two phases. The first was for radials lying on the ground and the second was for elevated radials. These represent two typical scenarios for amateurs, helping to answer a related question: "Do I put the antenna in the backyard or up on the roof?" For this series of tests I used a SteppIR III vertical.[16] The motor driven SteppIR can be adjusted to be resonant anywhere between 40 and 6 meters.

For these experiments I made up four sets of thirty-two ¼ wave radials, one set for each band (40, 20, 15 and 10 meters). I then tried several different configurations starting with sets of 32 single band radials, *one set at a time*. In this way I had a ¼ wave vertical over a ground system of thirty-two ¼ wave radials on each band. These antennas were then measured individually on each band. I then tried groups of four and eight (32 total) ¼ radials for each band, connected all at the same time. Next I tried 32 radials each 32 feet long, followed by 16, 8 and 4 at 32 feet each.

Obviously with a multiband antenna you would not run out to the antenna and change the radials whenever you changed bands! But this data can give us a feeling for any compromises resulting from the shift from monoband to multiband ground systems.

Four radials per band (16 radials in a four band system) probably represents the most common multiband ground system in general use both for elevated and ground surface radial systems, and we will use this as one measurement standard. I could have chosen many other possible combinations but those I did choose are at least reasonable. In particular I wanted to show that a few long radials don't work very well whether on the ground or elevated.

Radials Lying on the Ground

A comparison of the relative signal strength of each configuration with radials lying on the ground was made in comparison to the four radials per band case. The detailed results of this and following cases are shown on the QST In Depth Web page. In summary, however, there was little to choose among the cases (1 dB or less) until we came to the four 33 foot case that was down 2 to 4 dB from the standard four radials per band. The best performer is found with the 32 radials of 33 feet each, which is 0.4 to 1 dB better than our standard depending on the band. This case does require almost four times as much wire, however.

In the final analysis it appears that the standard ground system works just fine, but you can add more wire and get some improvement.

Vertical and Radials Elevated 48 inches

Once again the standard multiband radial system of four elevated radials appears to work well, nearly as well as the 32 radials of 33 feet each, although it has an edge of about 1.1 dB on 10 meters. As we move to fewer long radials, however, we found a problem on 20 meters in which the gain starts to fall quickly. This is related to the fact that the 33 foot, ¼ wave, radials on 40 meters are close to ½ wave radials on 20 meters, presenting a high impedance. At eight 33 foot radials the 20 meter response is down 4 dB, and at four 33 foot radials the performance was so poor I wouldn't consider it a multiband ground system. The four long radials didn't even work well on 15 meters, on which they were close to ¾ wave long.

Elevated Versus Ground Surface Radials

How do elevated multiband and ground surface radial systems compare to each other and to a large number of radials on the ground on each band? While the details are tabulated in the In Depth Web page, some conclusions can be summarized.

The differences between a 32 radial monoband system on the ground and a four radial elevated monoband system on each band are small, as we would expect from our earlier results.

If we compare a 16 radial multiband system on the ground with the same configuration elevated, the elevated system has about a 1 dB advantage on all bands. Doubling the number of radials on the ground will reduce the differences by 0.2 to 0.3 dB. The standard multiband system works just fine if elevated, but when the radials are lying on the ground it's not quite as good. If a radial system lies on the ground, the rule is you should use more radials to achieve comparable performance.

Acknowledgments

I want to acknowledge the helping hands that Mark Perrin, N7MQ, and Paul Thompson, W8EIB, provided in the field during these experiments. My thanks also to Mike Mertel, K7IR, for the loan of a SteppIR vertical for these experiments.

In addition to creating the design for the VNA used in these experiments, Paul Kiciak, N2PK, originally suggested to me the use of a VNA for these experiments when I was moaning and groaning about more conventional techniques. Paul also provide important criticism at several points to keep me on the straight and narrow

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