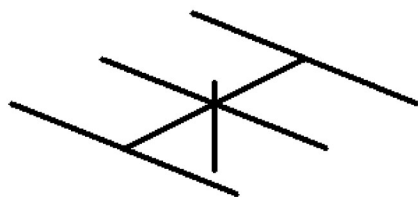
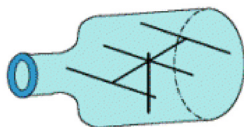


Antenna Modeling Notes



Volume 1



L. B. Cebik, W4RNL

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Table of Contents

Dedication	6
Preface	7
1. Converging Toward Excellence	9
2. Under the Limits: MININEC (3.13)	23
3. Within the Lines: NEC-2	36
4. A Good Start is Half the Trip	50
5. Putting Sources Where You Can Find Them	63
6. Modeling Loads: What Kind, How Much, and Where?	74
7. Maximizing Your Data	88
8. Modeling Wire Arrays	106
9. Modeling Ground Planes	120
10. Tapering to Perfection	134
11. A Ground Is Just a Ground—Unless It Is a Model of a Ground	147
12. Verticals At and Over Ground	161
13. Notes on Reactive Antenna Loads and Their NEC Models A. Some Center Loading Basics	179
14. Notes on Reactive Antenna Loads and Their NEC Models B. Some Mid-Element Loading Basics	195
15. Notes on Reactive Antenna Loads and Their NEC Models C. Some Linear Loading Basics	213
16. Notes on Reactive Antenna Loads and Their NEC Models D. Some Solenoid Loading Basics	237
17. Notes on Reactive Antenna Loads and Their NEC Models E. Some Unfinished Business on Modeling Loads	255
18. Why Tri-Banders Are Hard to Model	269

19. What Can We Learn From Tables?	281
20. The Average Gain Test	296
21. The NEC TL Facility	308
22. Modeling Physical Transmission Lines	320
23. Modeling LPDAs	338
24. The Power and the Source	350
25. Bringing Up the Rear: Front-to-Back Ratios	362
Appendix: Antenna Models	375
Other Publications	377

Dedication

This volume of studies of antenna modeling is dedicated to the memory of Jean, who was my wife, my friend, my supporter, and my colleague. Her patience, understanding, and assistance gave me the confidence to retire early from academic life to undertake full-time the continuing development of my personal web site at the URL, <http://www.cebik.com>. The site is devoted to providing, as best I can, information of use to radio amateurs and others--both beginning and experienced--on various antenna and related topics. This volume grew out of that work--and hence, shows Jean's help at every step.

Preface

Antenna modeling originated with the first mathematical equation that captured a property of antennas. Hence, the idea pre-dates Maxwell's equations. However, when we refer to *antenna modeling* in the 21st century, we access a different concept: the use of modern computers to apply a host of relevant antenna equations in a manner that permits the calculation of the essential performance characteristics of an antenna. Allied to the core calculation programs are appropriate input facilities to allow the user to construct a desired antenna for analysis and output facilities to present the calculated information in the most usable ways. NEC and MININEC are the best-known core programs, although there are others. Some preceded NEC, while others are undergoing independent development (such as FDTD methods). If you wish a concise history of antenna modeling software, see R. P. Haviland, "Programs for Antenna Analysis by the Method of Moments." *The ARRL Antenna Compendium*, Volume 4 (ARRL: 1995), pp. 69-73. MININEC 3.13, NEC-2, and NEC-4 enjoy the widest use within all communities interested in the design and analysis of antennas, antenna systems, and radiation-field situations. In these volumes, the term *antenna modeling* will almost always refer to the use of one or more of these core programs.

The 1980s and 1990s saw new implementations of MININEC and NEC: entry level programs, first in DOS and later in Windows. At a small cost, even radio amateurs could engage in antenna modeling, although usually with a restricted set of the commands available within the cores. The MININEC command set largely set the scope of the entry-level NEC programs that followed. Still, even with a smaller set of options, the beginning antenna modeler had in hand a very powerful vehicle for antenna design and analysis. Programmers strove to make their inter-faces as user-friendly as possible. But users managed to generate a host of questions about using the programs correctly and effectively. Those questions presented me with a challenge that I have tried to meet over the last decade.

Effective basic instruction in the use of antenna modeling programs seemed to call for more than just the entry-level manuals that accompanied modeling programs. Therefore, I started writing on the subject of antenna modeling pretty much as I began to master the programs myself. The sequence of articles began with "A Beginner's

Guide to Using Computer Antenna Modeling Programs,” which initially appeared in *The ARRL Antenna Compendium*, Volume 3 (ARRL: 1992). Since this initial offering on MININEC, I have done a 4-part QST series on the basics of modeling in NEC-2 and two basic modeling tutorials. In 1998, I began a series of monthly columns for *antenneX*, a series still active today. That series originated with information useful to beginning modelers, but has grown in recent times to address as well potentials, limitations, and techniques of interest to advanced modelers. It also began by focusing upon MININEC and especially NEC-2, but has now expanded its scope to include NEC-4.

Over the years, I have had the benefit of working and conversing with many highly talented program developers and antenna modelers. Conversations with Joe Finnerty of Nittany-Scientific and Roy Lewallen of EZNEC taught me much about the complex problems involved in making the NEC core accessible, speedy, easy to use, and data rich. Dick Adler of the U.S. Naval Postgraduate School and Jerry Burke of Lawrence Livermore National Laboratory have shown me what the NEC core can do and how it does its work, as well as correcting numerous misunderstandings. I hope these notes of appreciation will serve representatively to thank the myriad of folks to whom I owe a debt of gratitude. The list includes a host of modelers who have written me from time to time with challenging questions, many of which have formed the basis for columns in the series.

The time has come to collect these columns into a more convenient form for the reader. There is just too much material for a single volume, so I have broken the collection into three 25-column units. I have reviewed the text and graphics for each column to ensure as much accuracy as I can muster. However, I have also reviewed the sample models used in each column. That process permitted me to add something to these volumes that is not available in *antenneX* or at my own web site. The Appendix to each of these volumes contains a collection of antenna modeling files in three formats: .NEC (ASCII), .EZ (EZNEC), and .NWP (NEC-Win Plus). I have revised the text to include a file name for the applicable model in the Appendix. Therefore, should you wish to do so, you will be able to read a column in front of your computer and to test for yourself the ideas involved.

1. Converging Toward Excellence

Whether you are using one of the versions of MININEC or one of the versions of NEC-2, a temptation will overcome you somewhere along the way in your antenna-modeling career. You will ignore the convergence test. And in some cases, you may regret it, especially when your copper or aluminum antenna turns out to be well off the mark set by the model.

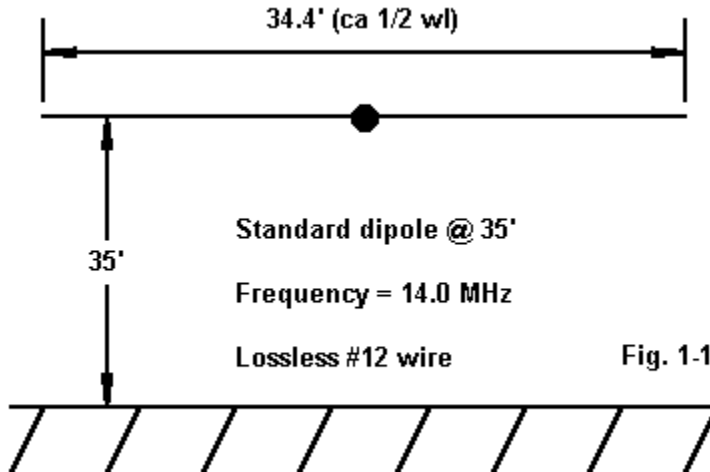
So let's take a patient look at convergence testing. Our aim will be not just to be able to do it, but as well to understand why we should always do it.

Both NEC and MININEC calculate antenna properties based on the division of the antenna elements into a collection of short wire segments. Within the program limits, the shorter, the better. However, both NEC and MININEC make use of complex calculation matrices that take time to fill and resolve. The more segments we have, the longer the run time. Because we are impatient, a bad habit has set in among many program users: to divide the antenna elements into as few segments as possible, again within certain generalizations in the instruction manuals. For MININEC, the recommended minimum number of segments per half-wavelength is 10, while for NEC-2 it is 9 or 11. (When center feeding an antenna element, use an even number of segments for MININEC and an odd number of segments for NEC.)

What we often forget is that neither program is absolutely stable, and the best we can hope for is relative stability. Instability refers to the fact that, if we change the number of segments per half-wavelength, without changing the overall dimensions, the program will produce different output data. We strive for a number of segments per half wavelength that—as we change the segmentation—produces output data differentials that are too small to be operationally significant. Of course, what counts as being operationally significant depends on the modeling goals for a given project.

We can illustrate the instability of the program by making some runs with an ordinary 20-meter dipole (model 1-1), as shown in **Fig. 1-1**. (In all figures, the heavy black “dot” represents the source or feedpoint. It is shown as a dot rather than as a break in the wire to indicate that the wire is continuous and the source is inserted in

series with the wire.) The only factor we shall change from model run to model run is the number of segments into which we divide this half-wavelength of wire.



For NEC-2, we can make a table of the results:

Segments	Gain in dBi	Source Impedance (R+/-jX)
11	7.97	67.51 - j0.22
15	7.98	67.50 + j0.34
21	7.98	67.52 + j0.24
25	7.99	67.53 + j0.32
31	7.99	67.54 + j0.42
35	7.99	67.55 + j0.46
41	7.99	67.56 + j0.53
45	7.99	67.56 + j0.56

This example is the source of the fatal temptation to always use the minimum number of segments. It is operationally stable at the lowest level of segmentation for any purpose I can think of, since I have no instruments capable of trimming an

antenna to the variations shown in the numbers. However, the table does illustrate that the numbers in fact do change with the level of segmentation. As the first 4 lines of the source impedance entries show, the numbers may drift in either direction until some large number of segments is used.

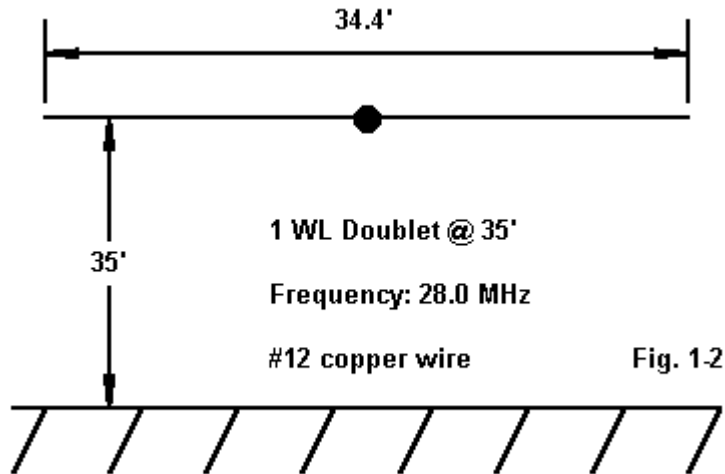
A similar situation exists with MININEC. The segmentation varies from the NEC table to accommodate the source location.

Segments	Gain in dBi	Source Impedance (R+/-jX)
10	7.83	67.51 - j6.78
16	7.85	68.19 - j7.71
20	7.85	68.40 - j7.57
26	7.86	68.59 - j7.22
30	7.86	68.67 - j7.01
36	7.86	68.76 - j6.70
40	7.86	68.81 - j6.50
46	7.86	68.87 - j6.26

The intrinsically higher level of reactance is a function of MININEC's growing error in resonant length as the frequency climbs. Some programs correct for this error; others do not.

Nevertheless, we find a similar drift in values with increasing segmentation. However, for operational purposes, the numbers give us confidence that a dipole constructed from these dimensions would work perfectly well with only such trimming as our yard clutter and local terrain might make necessary.

One common way in which we often fail to observe even the minimum recommended segmentation is when we increase the frequency of operation of an antenna without altering the number of segments. **Fig. 1-2** shows the same antenna (model 1-2), but now operated at 28.0 MHz.



The results for NEC-2 are the following:

Total Segments	Gain in dBi	Source Impedance (R+/-jX)
11	9.58	3813 + j1438
23	9.61	4306 - j 376
33	9.62	4120 - j 945
45	9.62	3928 - j1266
55	9.62	3810 - j1415
67	9.62	3705 - j1529

Although the gain stabilizes as soon as we use at least 11 segments per half-wavelength, the source impedance continues to move around considerably. Since we would likely feed this antenna via a parallel transmission line and antenna tuner, we might easily overlook the variance.

However, the antenna as presently used can be viewed as two end-fed half-wavelength dipoles physically coupled. End feeding near resonance yields large

changes in source impedance with very small changes in physical dimensions. Hence, changes in segmentation yield possibly significant differences in source impedance.

MININEC shows a similar set of figures:

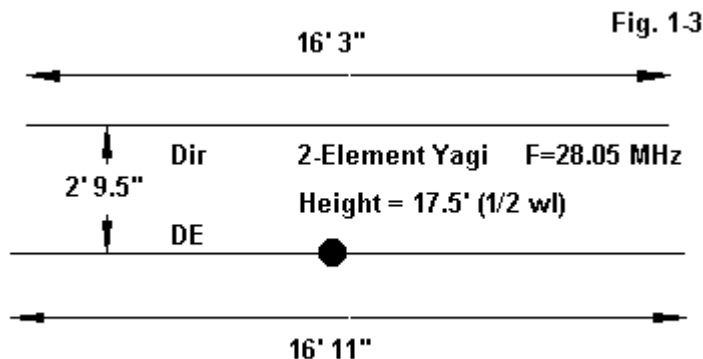
Total Segments	Gain in dBi	Source Impedance (R+/-jX)
10	9.55	3401 - j1646
22	9.56	3537 - j1519
32	9.56	3485 - j1568
44	9.56	3431 - j1616
54	9.56	3399 - j1643
66	9.56	3366 - j1669

Interestingly, the source impedance figures for MININEC appear to be more stable than those for NEC-2. It should be remembered that all MININEC source impedance figures are calculated based on perfect ground, whatever the ground used to determine the far-field figures. NEC-2 calculates source impedance based on the Sommerfeld-Norton ground conditions specified for the antenna.

In determining whether a satisfactory level of convergence has been achieved, the program user must determine how much variance is necessary to make it significant. The following driver-director 10-meter Yagi (model 1-3), shown in **Fig. 1-3** at a height of 1/2 wavelength, might be instructive here.

Special Note: If you run the attached .NEC files, you may find that your data does not agree with the tabulated data beyond about 2 to 3 significant figures. Differences in core compilations among programs and revisions over time may result in slightly different output reports. However, the trends in numbers with the changes shown in the tables should hold good.

Differences among implementations of MININEC may be more pronounced, depending upon the degree to which programmers have corrected various weaknesses in the public domain core. Again, the trends should appear as shown.



NEC-2 tabulates the results as follows:

Segments/Element	Gain in dBi	Front-to-Back in dB	Source Impedance ($R \pm jX$)
11	10.84	14.11	21.11 - j23.57
15	10.88	14.34	20.66 - j22.64
21	10.91	14.55	20.30 - j21.87
25	10.92	14.64	20.15 - j21.54
31	10.93	14.73	19.99 - j21.19
35	10.94	14.78	19.91 - j21.01
41	10.94	14.84	19.81 - j20.78
45	10.95	14.87	19.75 - j20.66

It might appear that, although the gain and the source impedance are fairly stable, especially in models using at least 21 segments per element, the front-to-back ratio continues to change at higher rate. However, from an operational standpoint, changes of a dB in this ratio are not especially detectable, and hence, the model might well be said to converge at about 21 segments per element—or less, if the needs are less critical.

Moreover, since we would likely use a matching system, such as the beta match or a series feedline section, between the feedpoint and the main 50-Ohm coaxial

feedline, the variations in the source impedance are unlikely to fall beyond normal adjustment limits. Hence, the home constructor might well judge the model to be well converged at the lowest level of segmentation. In terms of comparative design analysis, we might hold out for segmentation of at least 20 segments per element.

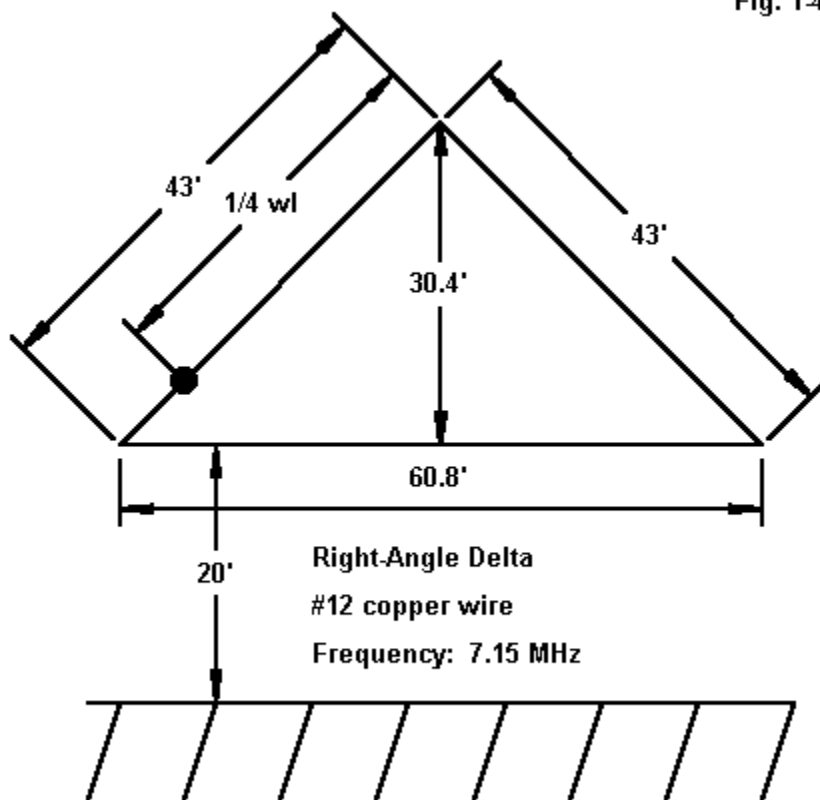
The corresponding MININEC model was run at the same frequency. Hence, its performance figures will be less impressive, since they reach a peak at a different frequency. However, for our purposes, the table shows a similar level of stability as the NEC-2 table above.

Segments/Element	Gain in dBi	Front-to-Back in dB	Source Impedance (R+/-jX)
10	10.49	10.71	21.91 - j29.20
14	10.53	10.90	21.72 - j28.79
20	10.56	11.18	21.29 - j27.75
24	10.61	11.34	21.01 - j27.11
30	10.64	11.47	20.81 - j26.40
34	10.65	11.57	20.63 - j25.79
40	10.68	11.71	20.37 - j25.27
44	10.68	11.75	20.31 - j24.94

Although stable enough for most purposes at the lowest level of segmentation, the figures are highly stable once we use more than 20 segments per elements.

Symmetrical structures, even of some complexity, tend to show a higher level of stability than non-symmetrical structures, even of similar types. To illustrate the point, let's compare a pair of triangular designs, vertically oriented. The first is a right-angle delta for 7.15 MHz (model 1-4), shown in **Fig. 1-4**.

Fig. 14



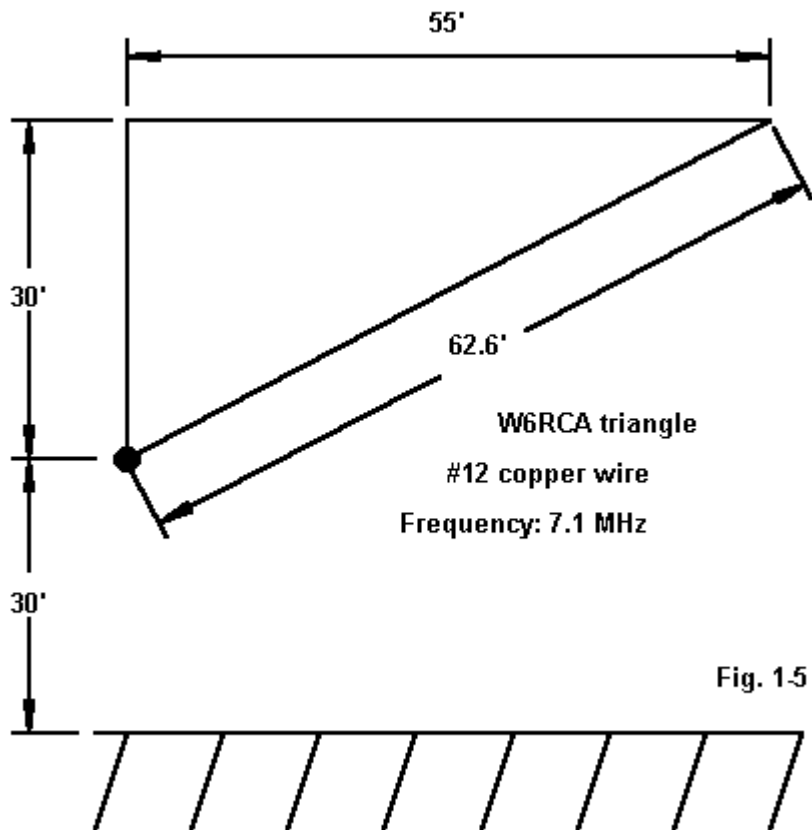
Despite its angularity, the antenna is quite stable once the segments are shorter than the maximum recommended length. The segmentation column shows the number of segments in each angular leg, and then the horizontal leg.

Segments	Gain in dBi	Source Impedance (R+/-jX)
6/6/8	1.73	63.83 + j1.49
12/12/17	1.90	60.95 - j0.29
18/18/25	1.89	60.99 - j0.77
24/24/34	1.89	61.03 - j1.03
30/30/42	1.89	61.06 - j1.18

The lowest level of segmentation results in segments longer than the maximum recommended length. Beyond that level, the model shows excellent stability over many levels of segmentation, despite the fact that the source is moving slightly with each change of segmentation, since it falls on the nearest segment corresponding to the position in the diagram. (There are techniques for placing the source very precisely, but they require more than the 3 wires that we are using in this simple model.)

One might be tempted to also model this antenna via MININEC. However, to do so would be to violate the limits of reliability with respect to the height of the antenna above ground. In general, as an antenna is brought closer to ground than about 0.2 wavelength, the gain increases inaccurately. For example, a fully length-tapered version of this antenna shows a MININEC source impedance of $64.5 - j23.6$ Ohms, somewhat off the NEC-2 mark, but not drastically. However, the antenna also shows a wholly unrealistic gain of 3.63 dBi. We shall examine the limitations of both NEC-2 and MININEC in upcoming columns (#2 and #3).

We began our examination of the vertically polarized right-angle delta loop with an eye toward comparing symmetrical and non-symmetrical structures of relevantly similar general shape. To complete the comparison, let's look at model of another triangle, one developed by W6RCA (model 1-5) and shown in **Fig. 1-5**. The antenna design is intended for multi-band use, but we shall confine our interest to the lowest frequency used. Note in the diagram that we feed the antenna at the lowest corner.



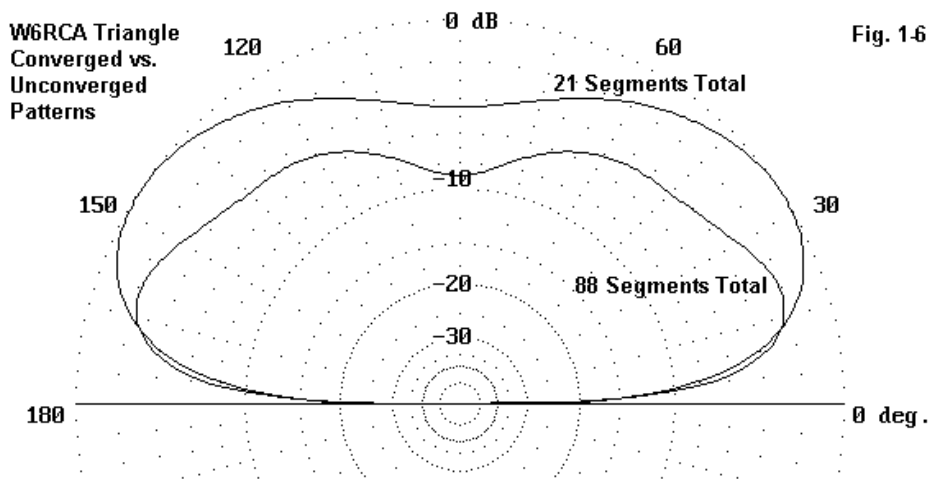
Relative to the plane of the antenna facing us, the antenna is not symmetrical on either side of a line one might draw at to bisect the source. Although the angularity of the antenna might lead us to believe that its model will behave similarly to the right-angle delta, the NEC-2 table tells a different story when using a split-current

source feed. (This feature is available on programs such as EZNEC and can be simulated on other programs by using a dual source and summing the results. The model shows much less instability when using a split-voltage feed.) The segmentation column refers to the number of segments in the vertical leg, the horizontal leg, and the diagonal leg, respectively.

Segments	Gain in dBi	Source Impedance (R+/-jX)
4/8/9	4.33	62.05 - j10.08
7/12/14	2.87	47.99 - j 5.77
9/16/19	2.45	44.10 - j 2.08
11/21/23	2.32	42.65 - j 0.26
13/25/28	2.24	41.83 + j 1.95
16/29/33	2.21	41.37 + j 3.73
18/33/37	2.19	41.19 + j 4.53

As with the right-angle delta model, the initial segmentation yields segment lengths longer than the maximum recommended values, and this fact is reflected in the unrealistically high gain and source impedance. By the point of maximum segmentation in the exercise, the model has stabilized to an operationally useful point—sufficiently so that the dimensions promise to make a good starting point for building such an antenna. However, the rate of change from one level of segmentation to the next is still significantly higher than it is for symmetrical antennas such as the dipoles, the Yagi, or the right-angle delta.

Inadequate segmentation on the split current-fed model does not just affect the gain and source impedance numbers. In fact, the entire far field pattern changes when comparing minimal segmentation models with adequately segmented models. **Fig. 1-6** compares the elevation pattern perpendicular to the face of the triangle for the 21-segment model and the same pattern produced by the 88-segment model. The latter is typical of the patterns of the group of more nearly converged models. The difference is not merely a matter of gain, but as well the correct portrayal of higher-angle radiation from the antenna. (Again, the model does not show the dramatic change in pattern shape when split-voltage-fed.)



Unlike the right-angle delta, the W6RCA triangle can be modeled via MININEC, since the bottom is about 0.22 wavelengths above ground. Even though MININEC advises against modeling with small numbers of segments when an antenna geometry contains right or acute angles, the exercise may be instructive. Again, for this example, the model is split-current fed.

Segments	Gain in dBi	Source Impedance ($R \pm jX$)
4/8/9	3.43	$64.17 - j210.2$
7/12/14	3.28	$50.00 - j125.6$
9/16/19	3.32	$45.72 - j89.5$
11/21/23	3.37	$42.66 - j68.4$
13/25/28	3.38	$42.02 - j54.2$
16/29/33	3.39	$41.68 - j43.3$
18/33/37	3.39	$41.68 - j36.8$

Although the model comes to show reasonable stability, its reported gain is well above that of the NEC-2 model, largely because it is near the border of reliability. This fact is significant, since many MININEC users think of the 0.2-wavelength height

limit as a kind of breaking point. It is not. Instead, we find a region where gain reports on different types of antennas become unreliable at different actual heights above ground.

The large swing of reactance in the source impedance is due to the fact that MININEC models in effect cut off sharp interior corners. The longer the length of individual segments meeting at an acute angle, the more wire is effectively trimmed in the computation of impedance and current interactions. As the segments grow shorter, the antenna gradually approaches its physical full size. A fully length-tapered version of the antenna model showed a gain of 3.38 dBi (still unreliably high), and a source impedance of $40.32 + j5.20$ Ohms, not far off the NEC-2 model.

The lessons taught by the W6RCA triangle are numerous and useful. First, it is unsafe in terms of model reliability to use the minimum number of segments per half-wavelength recommended for wires in the model by the program. These program recommendations are based upon linear elements and apply at most to symmetrical antenna geometries.

Second, it is also unsafe to assume that some arbitrary larger number of segments per half-wavelength will automatically yield a reliable model. The W6RCA triangle remained less stable with 44 segments per half-wavelength than many of the other models with under half that number.

The designer must make a series of convergence tests and reach a decision concerning the adequacy of stability in the model based on standards brought to the modeling exercise from a knowledge of the goals of the overall project. In some cases, stability adequate to one task may not suffice for another.

Third, inadequate convergence holds the potential for producing a model that misrepresents antenna performance in every way: gain, front-to-back ratio (if applicable), source impedance, and field pattern shape and strength. Although convergence alone cannot guarantee the adequacy of a model, it is one necessary condition to that goal.

Antenna modeling can be a very significant shortcut in the design and building of antennas. However, we can achieve the savings only if we do not take shortcuts with the modeling process itself. Convergence testing is one of those steps we

should never omit, lest the minutes we save at the computer testing our models end up costing us hours of frustrating time spent trying to adjust an antenna that is based upon an inadequate model.

The convergence test is only one of the two major tests that we can apply to an antenna model. The other test bears the name “Average Gain Test.” We shall examine that test in column #20 in this volume, with a follow-up column later on. However, both tests present necessary but not sufficient conditions of model adequacy. In other words, they may build confidence in a model, but the tests themselves cannot guarantee the adequacy of a model to its physical counterpart. Full confidence comes only when we have a series of adequate test results to compare with our modeled results. Of course, physical testing also has its own tests of adequacy.

* * * * *

Models included: 1-1 through 1-5.



2. Under the Limits: MININEC (3.13)

All antenna modeling software works within a set of limits. Exceeding the limits may yield a model, but one that is usually unreliable in terms of translating it into reality. Hence, we need to be thoroughly familiar with the limits of the software we use.

Since most of us use a commercial implementation of one of the basic public domain calculation cores (MININEC or NEC-2), we must divide our attention between core considerations and program considerations. There are limitations of the MININEC calculation core, which are initially common to all program implementations. However, some programs use proprietary correctives to overcome some limitations, and these correctives will vary from one program to another. Hence, we shall have to make reference to differences among programs, but without necessarily recommending one program over another.

The main commercial implementations of public domain MININEC (3.13) software are the following:

1. Antenna Model, from Teri Software. URL: <http://www.antennamodel.com>
2. NEC4WIN from Orion Microsystems, 197 Cr. Joncaire, Ile Bizard, Quebec, Canada H9C 2P7. URL: <http://www.orionmicro.com>
3. MMANA (freeware). URL: <http://www.qsl.net/mmhamsoft>
4. ELNEC from Roy Lewallen, W7EL. URL: <http://www.eznec.com>
5. AO (Antenna Optimizer) from Brian Beezley (This program is no longer distributed, but many copies remain in use.)

ELNEC and AO are DOS programs, while Antenna Model, NEC4WIN, and MMANA are Windows programs.

MININEC Limitations: Segmentation

Like NEC, MININEC analyzes an antenna by first analyzing the interaction among the individual straight segments that make up the elements. Segments are subdivisions of straight wires that compose the antenna structure, as shown in **Fig. 2-1**. A dipole may have a single wire subdivided into many segments. A 2-element quad may be composed of many wires.

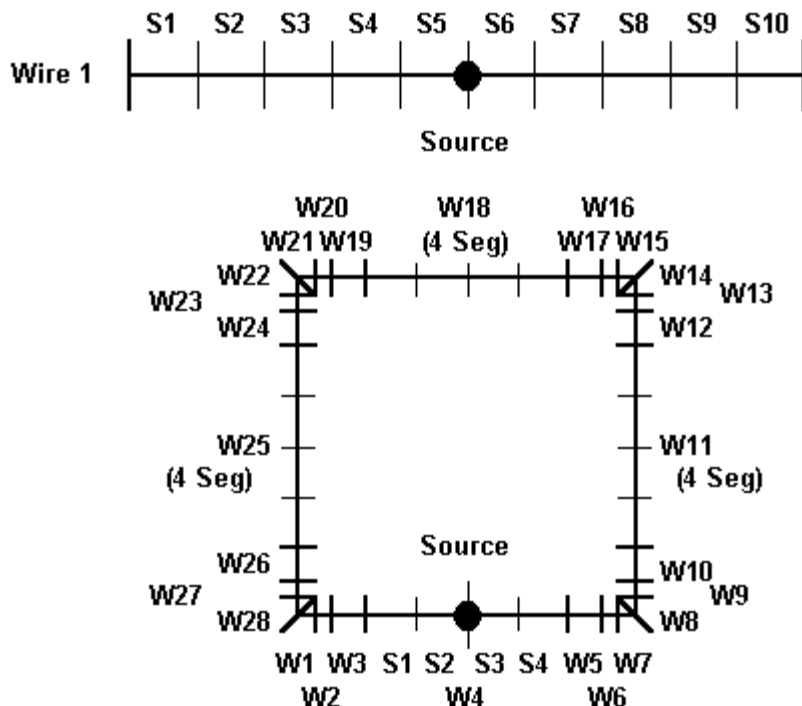


Fig. 2-1 Samples of simple and complex segmentation

The recommended minimum number of segments per half wavelength is 10, although fewer can occasionally be used. The fewer the number of segments per half wavelength, the less reliable the results. Even 10 segments per wavelength may be too few to pass convergence tests described in the last episode. The shortest segment length recommended is 0.0001 wavelength.

Although you can make a wire about as thin as you can imagine, your wires should not exceed a diameter of 0.02 wavelengths.

The following chart provides some guidance into the long and short of segment length, as well as into maximum wire diameter.

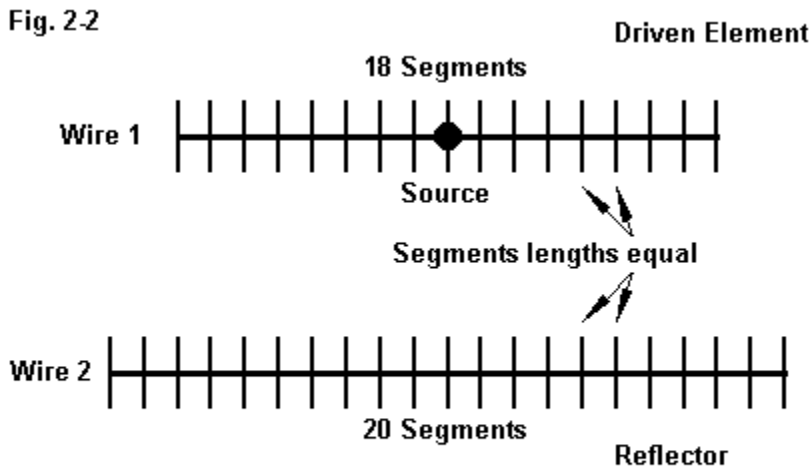
Frequency	Segment Length in “ for 10 seg/0.5 wl	Shortest Segment Length in “	Largest Wire Diameter in “
1.8	327.9	0.6657	133.1
3.5	168.6	0.3372	67.4
7.0	84.3	0.1686	33.7
10.1	58.4	0.1169	23.4
14.0	42.2	0.0843	16.9
18.068	32.7	0.0653	13.1
21.0	28.1	0.0562	11.2
24.89	23.7	0.0474	9.5
28.0	21.1	0.0422	8.4
50.0	11.8	0.0236	4.7
144.0	4.1	0.0082	1.6

The length limitations do not operate in isolation. First, the length should never be less than about 1.25 the wire diameter, and a greater ratio of length to diameter is preferable. Second, the ratio of lengths of adjoining segments should be less than 2:1.

The preceding recommendations represent program limitations. In practice, we should avoid these limits by the greatest possible distance. We can accomplish this goal by adhering to some positive rules of thumb. First, insofar as feasible with linear elements, make the segment lengths equal, even where the element may be composed of wires of different diameter. Second, use at least twice the minimum recommended segmentation for any antenna more complex than a simple half-

wavelength dipole. (We shall soon discover an important situation in which to depart from these rules.)

Fig. 2-2 illustrates these rules of thumb in action for a two-element Yagi.



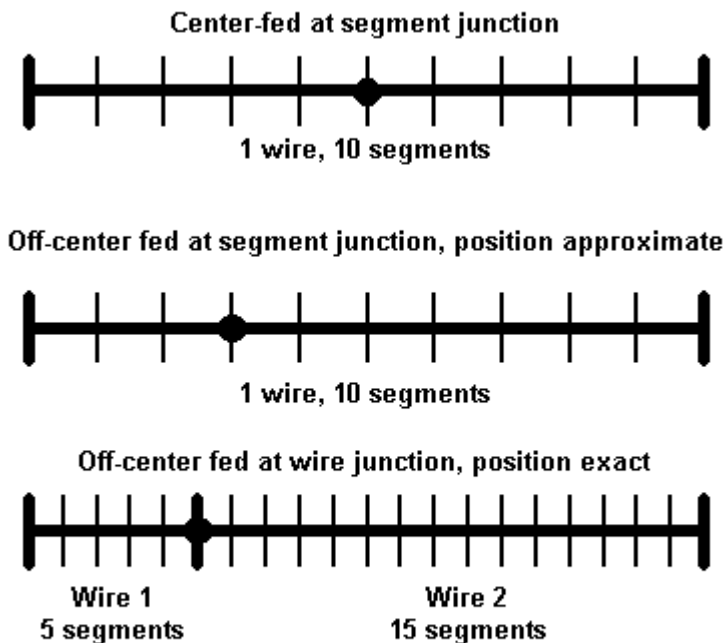
Some "rules of thumb" in operation

Adherence to these rules of thumb has proven less critical in MININEC than in NEC and less critical where wires are widely spaced than when closely spaced. Nonetheless, they are reasonably good habits to get into, except where specifically counter-indicated, as in the case of segment length tapering, to be discussed in a moment. In all cases, increase segmentation as indicated by appropriate convergence testing.

MININEC calculations are based upon placing a current pulse at the intersection of two segments (with wire end pulses added to non-continuous elements). The placement of the pulse (which differs from the manner in which NEC performs its calculations) also dictates the placement of antenna sources: they must be at

pulses. Hence, a center-fed linear element should have an even number of segments so that there is a pulse (segment junction) at the element center on which to place the source. Similar rules apply to the placement of loads: even if off center, they can only be placed at junctions of wire segments. Exact placement of sources and loads which are not at wire centers or ends requires that there be enough segments to position them very close to their physical counterparts. **Fig. 2-3** illustrates the situation.

Fig. 2-3

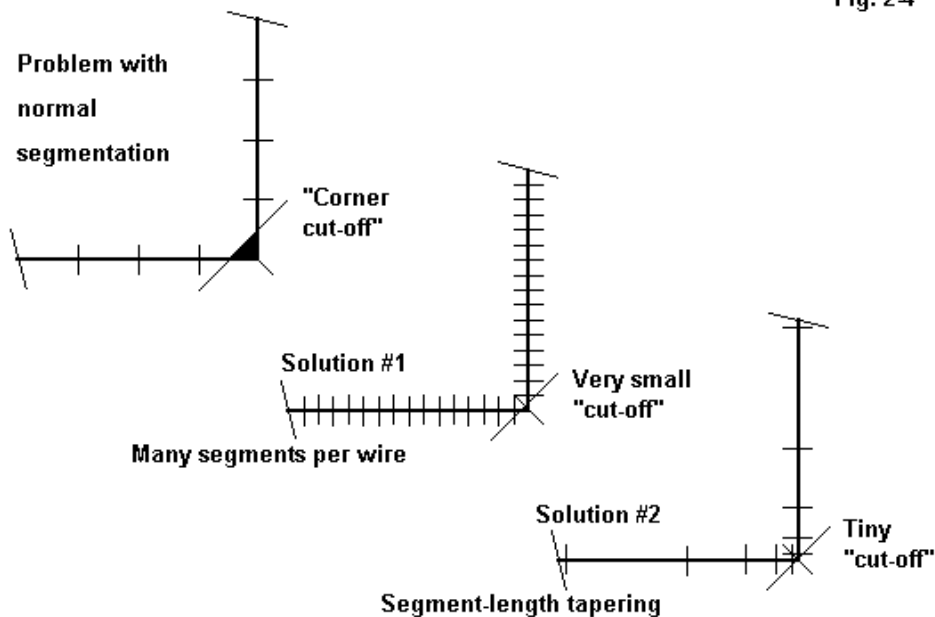


Source/feedpoint considerations

As **Fig. 2-3** shows, when source or load position is critical, you can always subdivide an element into wires of different lengths and place the source or load at the junction of the wires. Some programs limit the position of the source (and loads) to wire ends or centers; hence, the use of separate wires within a single antenna ele-

ment may be necessary in order to place a source (or load) exactly where the antenna dimensions require it. This technique is especially applicable to antennas such as the off-center fed half-wavelength wire, sometimes called a Windom.

Fig. 2.4



Angular wire-junction considerations

The MININEC system of pulse placement puts limitations on the construction of models involving angular geometries, such as the quad loop. The ends of pulses

are roughly at the middle of segments. The effect resembles shortening the antenna by ignoring some of the wire at the corner. Hence, as suggested by **Fig. 2-4**, with only close to the minimum recommended segmentation per half wavelength, a MININEC model quad loop will have to be made larger than in reality to achieve resonance at a desired frequency.

There are two ways to overcome this limitation. One is to increase the number of segments, making each one very short. This technique results in a minimal shortening effect and can yield accurate quad models. A second technique is called segment length tapering. Antenna Model and AO provide an automatic tapering option, while ELNEC provides an option for constructing an alternative model with either preset or user-controlled maximum and minimum segment lengths. In both cases, segments approaching critical angles become very short and gradually increase in length. The technique yields maximum accuracy with a minimum number of segments per wire.

MININEC was created for the PC when the maximum memory available was 640 KB. Memory is the key limitation in the maximum number of segments available for a model. At one time, a maximum of about 128 pulses was standard. However, existing programs have increased this limit by various means. ELNEC uses a special program adjunct to increase the limit to about 256 pulses. AO has altered the code with machine-level routines so that the number of pulses is limited only by memory. Antenna Model and NEC4WIN uses coding within the Windows protocols to remove the limit from the number of available pulses.

Despite these advances, when compared to NEC-2 programs, MININEC models in older implementations have size restrictions that limit some types of modeling. Very large and complex geometries may call for considerable ingenuity. Even then, their accuracy may be suspect in some cases. The latest Windows implementations largely overcome the size restrictions, but with cores that run extremely slowly compared to the speed of compiled Fortran NEC cores. Nevertheless, we shall see when we discuss some of the limitations of NEC-2, that MININEC does model some situations with a good deal more reliability than its FORTRAN counterpart.

Grounds

Another general area in which MININEC shows significant limitations is in its options for grounding and the accuracy of models near ground. MININEC has, in addition to a free space option, a ground system that permits the definition of two ground condition sets. These ground condition specifications affect only the far field pattern of the antenna model. Source impedance and current data is developed as if the antenna were positioned over perfect ground. For many types of antennas, the differences between source impedance figures taken over perfect ground and those taken over less than perfect ground will not be operationally significant. The following table shows figures for a dipole modeled at 3.5 MHz 1 wavelength high in MININEC and also in NEC-2 with high accuracy Sommerfeld-Norton grounds ranging from perfect to very poor. The antenna was first resonated within $\pm j$ 1 Ohm reactance over perfect ground and held at that length for the development of the other figures. MININEC dipoles have 20 segments; NEC dipoles have 21 segments.

Antenna Ground	Cond. (S/m) & Die. Con.	MININEC		NEC-2 (N-S Ground)	
		138.0' #12 copper Gain dBi	Source Z R +/- jX	137.6' #12 copper Gain dBi	Source Z R +/- jX
Perfect		8.12	73.1 + j0.38	8.12	73.2 - j0.42
Very Good	0.0303/20	7.99	73.1 + j0.38	8.05	72.5 + j0.77
Good	0.005/13	7.80	73.1 + j0.38	7.87	72.2 + j2.42
Poor	0.002/13	7.64	73.1 + j0.38	7.69	72.6 + j3.30
Very Poor	0.001/5	7.42	73.1 + j0.38	7.49	72.3 + j4.75

This high dipole shows no operational problem in using the MININEC figures in lieu of the NEC-2 figures. However, the NEC source figures do show a small but possibly significant reactance. To maintain resonance, the antenna requires a change in length amounting to about 0.1' per Ohm reactance. For other types of antennas or heights, the difference might be greater and of higher significance.

MININEC has a second inherent limitation respecting ground. Wires placed roughly parallel to the ground will yield both inaccurate current (and therefore gain) data and also source data when they are below about 0.2 wavelengths. (More accurate data close to the ground would have required a larger code at a time when PCs

had severe memory limitations.) The following table compares MININEC and NEC-2 data for a 3.5 MHz dipole (resonated in free space) at heights from 0.05 to 0.30 wavelengths above medium or “average” earth (conductivity = 0.005 Siemens/meter; dielectric constant = 13).

		MININEC		NEC-2 (S-N Ground)	
Antenna		137.2' #12 copper		136.9' #12 copper	
Height		Gain	Source Z	Gain	Source Z
W/L	Feet	dBi	R +/- jX	dBi	R +/- jX
0.05	14.05	9.4	7.4 - j 4.9	1.2	48.9 + j15.4
0.10	28.10	8.4	23.3 + j20.5	5.1	49.8 + j21.1
0.15	42.15	7.7	62.3 + j37.0	6.5	77.0 + j25.3
0.25	70.26	6.2	87.7 + j28.3	6.2	87.8 + j17.3
0.30	84.31	5.9	97.4 + j13.5	6.1	92.3 + j 6.1

The low altitude gains calculated by MININEC are, of course, unrealistically high, whereas the ones calculated by NEC-2 are more reliable, in light of ground losses. Figures do not begin to coincide between the two programs until the dipole reaches about 0.25 wavelengths in height. Notice that figures for the dipole are not wholly reliable in MININEC at heights of 50 feet, a common amateur backyard dipole height for 80 meters. Similarly positioned 80-meter wire arrays of greater complexity will show equal or greater errors. The upshot is this: if models of antennas around or below the MININEC height limit are needed, then MININEC is not the program of choice for such models.

The limitation just described also makes MININEC unreliable in the evaluation of ground planes for many lower HF verticals, since the ground plane normally consists of wires under, at, or just above the surface of ground and parallel to it. NEC-2 is more accurate using the Sommerfeld-Norton ground and placing wires at their height above the surface. Underground radials can be modeled at very small heights above the surface for reasonably accurate results. (NEC-4 has the ability to model wires underground.)

Miscellaneous Limitations

There are a number of other less problematical limitations within MININEC. Some commercial implementations of MININEC have added corrective measures; others

have not, most likely in the belief that the amount of error introduced is within the boundaries of normal antenna building practice.

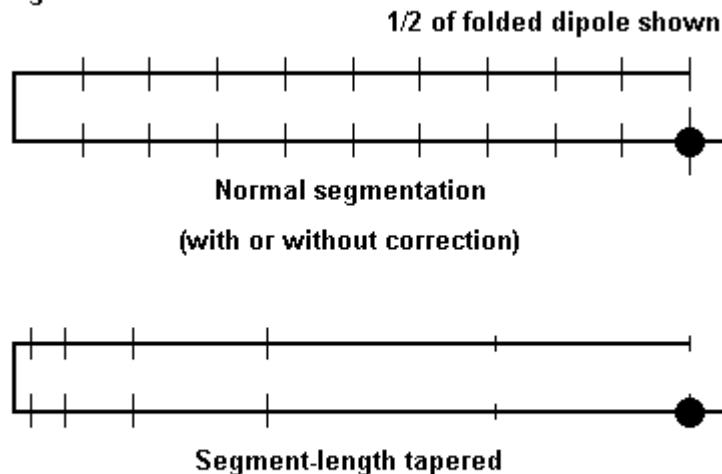
Frequency: As the frequency at which models are set increases, MININEC shows a gradual offset in the direction of showing an antenna to be shorter than its real-world counterpart would be. AO introduces a corrective for this offset to bring the results in line with NEC-2, and some Windows implementations (for example, Antenna Model) have followed suit. To illustrate the phenomenon, the following results were obtained for a 3-element Yagi of uniform 0.5" aluminum elements at 28.5 MHz in free space:

Program	Gain dBi	Front-to-Back dB	Source Impedance R+/- jX Ohms
NEC-4	8.11	27.15	25.7 - j0.8
NEC-2	8.11	27.13	25.7 - j0.8
AO (MIN)	8.09	26.89	25.7 - j0.3
ELNEC (MIN)	8.04	27.21	26.5 - j6.0
ELNEC at 28.65 MHz			25.7 - j0.6

The uncorrected MININEC model is resonant at a frequency about 150 kHz above the resonance shown by the other models or about 0.5% higher in frequency at 10 meters. Similar results are shown by AO when the frequency corrective is switched off. The actual amount of error is a function of both antenna element diameter and frequency and thus may become more significant in the upper HF and VHF region of the spectrum. Be certain to apply this test for frequency correction to MININEC implementations that have appeared after I conducted these tests.

Parallel Wires: Closely space wires, such as in a folded dipole can give erroneous outputs in MININEC unless carefully segment-length tapered toward the junctions with the short end-connecting wires. ELNEC introduced a parallel wire correction feature that permits the use of low numbers of segments per half wavelength without length tapering with quite reasonable output results. The following table illustrates a simple folded dipole (partially shown in **Fig. 2-5**) of #18 copper wire with 1" wire spacing and 16.6' total length at 28.5 MHz in free space.

Fig. 2-5



Closely spaced wires considerations

No. seg./ 0.5 wl	Without correction		With correction	
	Gain dBi	Source Z R +/- jX	Gain dBi	Source Z R +/- jX
22	1.60	11.6 - j106	2.09	287 + j 8.5
32	1.46	11.4 - j 85	2.10	288 + j 2.0
42	1.21	11.0 - j 84	2.10	288 + j 3.1
Full taper			2.09	285 + j31.9

AO results for tapered and uncorrected untapered folded dipoles are similar. The closely spaced parallel wire correction where wires join can be overcome either by a corrective or by segment length tapering. However, where the wires do not join, the modeler must use care in validating the modeled results in the absence of a parallel wire corrective.

On the Positive Side

There are some aspects of modeling where MININEC shows significantly greater accuracy than NEC-2. We shall explore the NEC-2 limitations more extensively in the next episode, but here we can simply list a few MININEC superiorities.

Folded dipoles of dissimilar diameter wires: NEC in all its forms produces rather unreliable models of folded dipoles with dissimilar wires. The 4:1 source impedance ratio for a folded dipole occurs only when both wires are the same size. When the wires differ in diameter, the source impedance is a complex function of the wires' diameter ratios and their spacing. Details of the calculation appear in *The ARRL Antenna Book*. Detailed models in MININEC show very reasonable approximations of the calculated results. For many such folded dipoles models to yield accurate results, the implementation must have a close-wire corrective.

Elements with tapered wire diameters: Yagi elements may start at the center with 1" or larger diameter tubing and decrease the diameter as the progress outward toward the ends. MININEC handles these transitions in stride. In contrast, NEC-2 requires a special type of corrective, usually based on the Leeson correctives, which substitutes an equivalent single-diameter element for the one actually placed in the model. In fact, the Leeson correctives were calibrated using MININEC dipole models.

Closely spaced wires: When wires are very closely spaced—say less than 10" or so at 20 meters (with frequency, wire diameter, and other variables thrown in), NEC results become unreliable, generally showing higher than realistic gain, but with no change in the antenna pattern shape. In contrast, MININEC can, under correction, produce realistic results down to very close wire spacing. (However, never let two antenna-wire surfaces overlap.)

Changing diameters at angular corners: Again, NEC does not handle this situation well, producing unreliable models. Extensive modeling of the folded X-Beam, which normally uses tubular angled arms and wire tails, suggests that, when using segment length tapering at the angular corners, MININEC produces more reliable modeling results. Likewise, tests with quad loops with tubular horizontal sections and wire vertical sections also suggest that MININEC models are more accurate than NEC models. Details on some of these tests will appear in a future column.

This sampling of areas where MININEC produces more reliable models than NEC-2 suggests that for the foreseeable future, MININEC will maintain a significant place in antenna modeling. Indeed, over a broad middle ground that does not approach the limitations of either program, both will produce quite reliable models—sufficiently reliable to be guides of antenna construction. Our discussion of MININEC limitations is not aimed at lessening confidence in the program or its commercial implementations. Rather, it is designed to help us ensure model reliability by keeping our models away from the limits and well within arena where MININEC best does its work.

In a future (very future) column, we shall explore differences among the results obtained with various implementations of MININEC. The currently available versions are highly variable in their corrections of the basic core. Perhaps only Antenna Model has undertaken a full sweep of correctives so that its free-space models hold well when compared to NEC-4 models of known accuracy.

For further information on MININEC limitations, the classic reference is Roy Lewallen, W7EL, “MININEC: The Other Side of the Sword,” *QST* (February, 1991), 18-22.

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Because the implementations of MININEC use such varying model file formats, I have included no models for this episode. Indeed, only NEC models will be included with this volume.



3. Within the Lines: NEC-2

In the preceding episode, we looked at some of the limitations of MININEC and how to model effectively within those limits. Here, we shall give NEC-2 equal time. Virtually all of the limits we shall examine are functions of the core NEC-2 program module and therefore occur on all commercial implementations of NEC-2.

The chief sources of commercial implementations of NEC-2 are the following ones. All are Windows programs, except as noted.

1. EZNEC and EZNEC Pro: Roy Lewallen, W7EL. URL: <http://www.eznec.com>.
2. NEC-Win Plus and NEC-Win Pro: Nittany Scientific. URL: <http://www.nittany-scientific.com>.
3. SuperNEC: Poynting Software. URL: <http://www.supernec.com>.
4. NEC-Wires: Brian Beezley. (This DOS program is no longer distributed, but is still in use by numerous modelers.)
5. 4NEC2: Arie Voors. URL: www.qsl.net/wb6tpu/swindex.html. (freeware)
6. Antenna Solver: Grating Solver Development. URL: <http://www.gsolver.com>. (freeware)

The basic reference for NEC-2 is J. Burke, A. J. Poggio, "Numerical Electromagnetic Code (NEC) Method of Moments, a User Oriented Code," Vol. 2 (Part III: User's Guide), Tech. Doc. 116, Naval Systems Center, San Diego, 1982. For those desiring to create their own input and output systems, NEC-2 is public domain and available in FORTRAN and compiled versions. Ray Anderson, WB6TPU, maintains a site for downloading basic NEC-2 materials, including the 3-part full manual: URL: <http://www.qsl.net/wb6tpu>. There is also a "shell" program in EXCEL by Dan Maguire: MultiNEC that provides input and output facilities for a user-supplied core (URL: <http://www.qsl.net/ac6la/index.html>).

Segmentation Limitations

Although the segmentation limitations of NEC-2 are similar to those of MININEC, they differ just enough so that the transition from one program to the other sometimes creates unanticipated difficulties. First, NEC places the equivalent of a MININEC pulse in the center (or distributed throughout) a wire segment. Hence, in NEC, you should think in terms of segmenting antenna lengths, especially those holding sources and loads, in odd numbers. Although many references speak of segmenting a center-fed half-wavelength dipole into at least 10 segments, 9 or 11 are the proper numbers, as shown in **Fig. 3-1**.



Properly segmenting a fed element in NEC-2

Segment length should be under 0.1 wavelength long, with 0.05 wavelength preferred (about 10-11 segments per half-wavelength). Segments shorter than 0.001 wavelength should also be avoided. For reference, the following table provides a ham band list of the maximum and minimum recommended segments lengths.

Frequency	Segment Length in inches for 0.05-wl segments	Shortest Segment Length in inches
1.8	327.9	6.657
3.5	168.6	3.372
7.0	84.3	1.686
10.1	58.4	1.169
14.0	42.2	0.843

18.068	32.7	0.653
21.0	28.1	0.562
24.89	23.7	0.474
28.0	21.1	0.422
50.0	11.8	0.236
144.0	4.1	0.082

Thin-wire segments are preferred: as with MININEC, the wire circumference divided by the wavelength should be much less than 1 for accurate results. Moreover, the ratio of segment length to diameter should be greater than 4 for errors less than 1%. If the model demands a smaller ratio, it should be approached cautiously by shortening segment lengths gradually with an eye toward results taking off on a tangent.

Maintaining a larger segment length to diameter ratio at corners is also necessary to keep the center of one segment from falling within the radius of the other segment. Again, approaching this limit produces nothing sudden, so that it can be pressed, but cautiously.

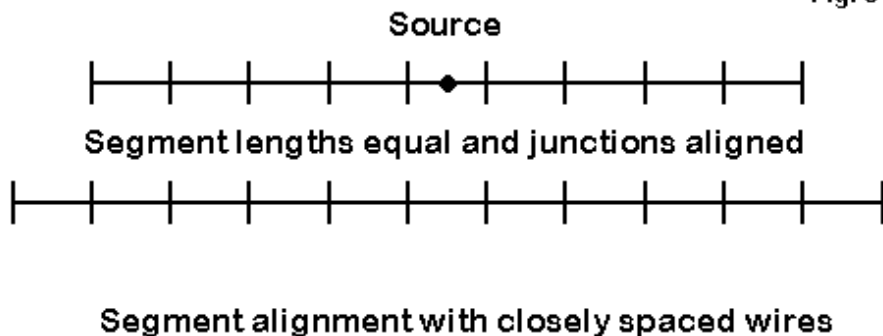
Unlike MININEC, angled antenna elements do not require special treatment other than the warning about very short segment-length-to-wire-diameter ratios. NEC-2 will model equally segmented wires in a quad quite handily.

However, prevent wires from physically touching or coming in very close proximity when crossing. There is no hard and fast rule on where the proximity limits occurs, but separation by several wire diameters is recommended.

Another NEC-2 limitation is the inability to model small loops, less the about 0.1 wavelength in circumference.

NEC-2 documentation specifically recommends that closely spaced parallel wires be arranged so that the segments are carefully matched, as shown in **Fig. 3-2**. As noted in the last episode, this practice is a good one to follow with all models.

Fig. 3-2

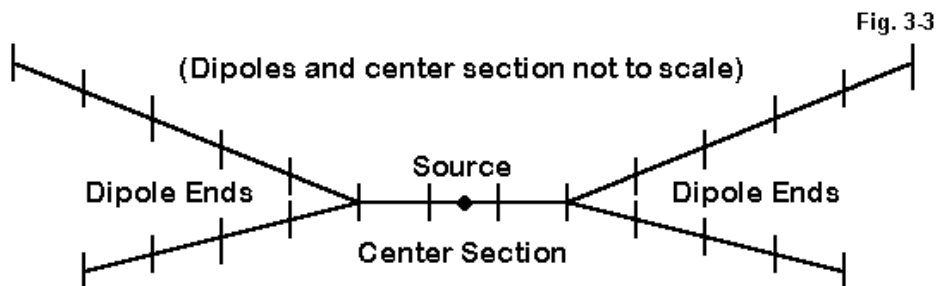


Junction Limitations

There are two junction limitations of note. The first limits the number of wires joined at a signal junction to 30. This limitation is especially applicable to models involving base-fed vertical antennas with radial systems modeled above ground. Extending the ground radial system to larger numbers with a direct connection to the vertical segment holding the source should be approached with caution by examining the sensibleness of the outputs as the limit is approached. A further note on elevated ground radial systems appears later in the discussion of grounds.

The second junction limitation concerns feeding multiple antennas at a common source point. This problem differs from feeding a single bent element (such as inverted Vee) at the apex. In this case, the modeler can often use a split feed, feeding with separate sources the segments on each side of the wire junction. If the segments are short, the resulting sum of the two impedances will yield an accurate overall source impedance for the antenna. Some programs provide for a split voltage or current feed and report the source conditions (voltage, current, and impedance) as a single set of values.

When multiple antenna elements join at a single point, it is no longer possible to employ a split feed effectively. The simplest case is the spread dipole for two bands with a single feedpoint, as shown in **Fig. 3-3**.



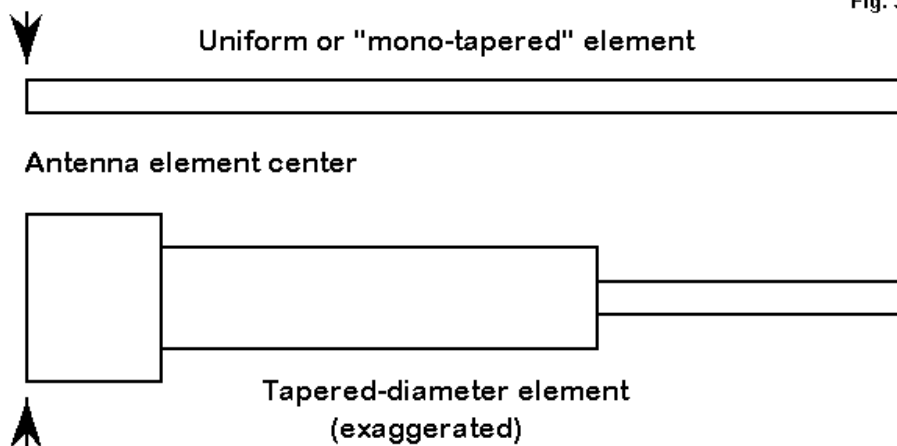
Method for modeling antennas with a common feedpoint

The simple work-around is also shown in **Fig. 3-3**: adding a small common section of wire for the source and joining the diverging elements to the ends of this wire. Recommendations for these segments include a minimum length of 0.02 wavelength and 3 segments. Equally important, the adjoining segments of the diverging wires should be about the same length as those in the center section.

Diameter Limitations

The key "diameter" limitation is much better known by most NEC-2 users: the stepped wire-diameter limitation. NEC-2 produces inaccurate results when an antenna element is composed of wires of differing diameters, as is commonly the case with HF Yagis using several sizes of aluminum tubing. The basic situation is shown in **Fig. 3-4**.

Fig. 3.4



Uniform and stepped-diameter elements

If an element is constructed on each side of center by, say, only 2 wires of different diameters and the junction of these wires is past the mid-point on each side, modeled results will be more accurate than with elements having multiple diameter steps closer to the center. Large steps in diameter also increase the accuracy problem.

Most commercial implementations of NEC-2 have incorporated a technique to overcome this problem effectively. Using equations developed by Dr. David Leeson, the programs calculate the antenna properties with substitute elements having a constant diameter. The resulting models have proven quite reliable. However, you must use caution in constructing the model to ensure that the stepped diameter element is continuous or collinear, with no bends or intervening geometric oddities

along the way. For example, adding a mid-element capacity hat structure will disable the correction feature in some programs. Likewise, the source must be at the center of an element with open ends (such as a dipole), and loads must be symmetrically placed. Transmission lines are sometimes disallowed. Moreover, the element may be required to be within a certain percentage of resonance, which may complicate attempts to model in NEC-2 multi-band HF Yagis with stepped-diameter elements throughout.

Let's take a closer look at the effects of using and not using the substitute elements in place of the stepped diameter elements. The following table shows several aluminum dipole elements for 14 MHz, ranging from a uniform 1" wire to a highly stepped set of wires. Lengths and diameters are from the end to center, with the other side symmetrical to the given side. Except for the uniform element, NEC-2 outputs are shown for both the substitute and the uncorrected elements. (See models 3-1 through 3-5.)

Length(s) from center in inches		Free Space Gain	Source Impedance
Diameter(s) in inches		in dBi	(R +/- jX Ohms)
1. Uniform element (no corrective needed)			
-201.25...0		2.12	71.8 - j 0.6
1.0			
2. One step, far out			
-204...-150...0	No Cor.	2.14	73.0 + j 4.4
0.75 1.0			
-201.45...0	Cor.	2.12	72.0 + j 0.4
0.966			
3. One step, near center			
-204...-50...0	No Cor.	2.22	72.4 + j 5.2
0.75 1.0			
-201.661...0	Cor.	2.13	71.8 - j 0.5
0.792			
4. Two steps, modest taper			
-205.75...-100...-20...0	No Cor.	2.32	72.5 + j10.6
0.75 1.0 1.25			
-201.49...0	Cor.	2.13	71.9 + j 0.1
0.889			

5. Two steps, more extreme taper					
-208.5...-100...-20...0	No Cor.	2.82		67.6 + j17.1	
0.75 1.0 2.5					
-201.63...0	Cor.	2.13		72.1 + j 0.9	
0.895					

The corrected substitute models ("Cor.") are those generated by the program as replacement elements for the original model, which reflects the intended tapered diameter structure. With only moderate levels of diameter stepping, uncorrected NEC-2 reports of gain rapidly become unreliably large, while source impedances improperly show the elements to be too long relative to resonance.

Corner Dissimilar Wire-Diameter Limitations

Although linear stepped-diameter elements are correctable, non-linear elements with changes of diameter are not. In many instances, NEC-2 will produce unusable results. Two such cases are the quad loop and the folded dipole, each with wires of unequal diameter.

Consider the single quad loops (models 3-6 and 3-7) in **Fig. 3-5**. If we construct such a loop for 28.5 MHz of #14 wire, then about 9.13' of wire per side ($1041/f$) will yield a loop with a free space gain of about 3.24 dBi and a source impedance of 126-127 Ohms resistive. This is true whether we model the antenna as a fully length-tapered MININEC item or as a NEC-2 wire antenna with reasonable segmentation. If we make the element 1" in diameter, then 9.5' per side ($1083/f$) in both programs yields a gain of about 3.4 dBi with a resonant source impedance close to 132 Ohms.

Fig. 3.5



Standard and nonstandard corner junctions

However, if we change the construction so that the horizontal portions are “fat” while keeping the vertical portions of “thin” wire, the results are far different. The following table shows how (models 3-6 and 3-7).

Antenna Dimensions	Free Space Gain dBi	Source Impedance (R+/-jX Ohms)
1. 0.5" dia hor/#12 vert: 10.15' per side MININEC tapered	3.61	136.7 - j 2.3
NEC-2	3.57	175.4 + j140
2. 1.0" dia hor/#14 vert: 10.7' per side MININEC tapered	3.80	141.6 - j 5.7
3. 1.0" dia hor/#14 vert: 9.68' per side NEC-2	3.46	138.4 - j 0.5

The first antenna uses the same model in both programs. Compared to the materials of antennas 2 and 3, this model has a less extreme difference in wire diameters between the horizontal and vertical portions. Although the gain figures produced are close, the source impedance values are radically different, as NEC-2 suggests a far smaller loop size for resonance.

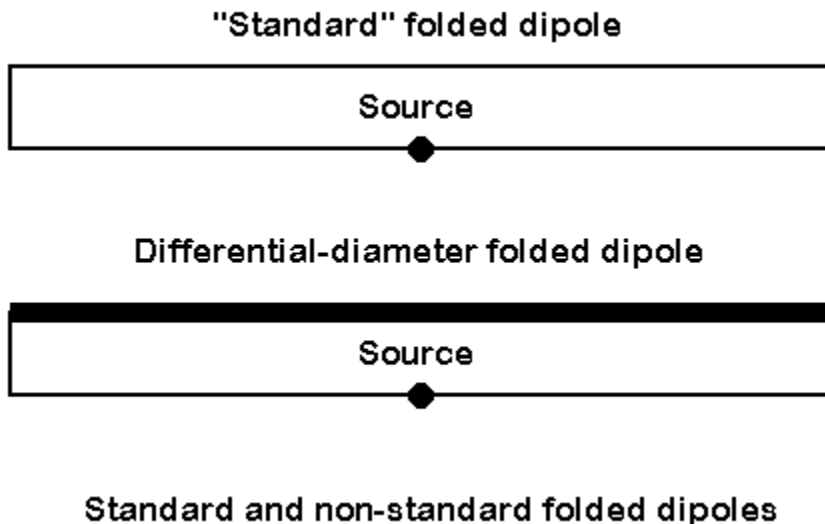
Antennas 2 and 3 use a larger difference between the horizontal and vertical wire diameters. The two models use the same horizontal to vertical wire diameter ratios, and each is each brought close to resonance. Although the source impedances are now comparable between MININEC and NEC-2, the loop sizes for resonance are very far apart.

Divergent results of this order require empirical verification before either modeling system can be trusted for dissimilar diameter materials meeting at the corners of an antenna element. So I modeled a 50-Ohm fat-horizontal, thin-vertical wire loop for 146 MHz. The 0.75" diameter horizontal bars were 16" long. For resonance, NEC-2 called for 29.15" #14 (0.064" diameter) vertical wires, while NEC-4 called for 32.2" wires. Uncorrected MININEC resonated the loop with 33.7" vertical wires. The test antenna resonated with 33.75" vertical wires. The "real" antenna result can be up to about +/- 3/8" in error due to possible variances from the model created by screw heads and short leads to the antenna coax fitting. However, the dimensions are sufficiently accurate to demonstrate the greater reliability of MININEC results and the problems of modeling corner junctions of dissimilar diameter wires in NEC-4 and NEC-2.

An interesting deviation from this pattern occurs when right-angle junctions of dissimilar diameter wires involve symmetrical arrangements of one size of the wires. These models include vertical antennas with elevated ground radial systems, dipoles or verticals with "capacity hats," and similar structures. In these cases, the apparent cancellation of radiation from the elements of the symmetrical portion of the structure yields accurate gain and source impedance reports. A series of experimental models, verified by measurements with antennas built from the models, showed an agreement between NEC-2 and MININEC models within 1 to 2 percent for the radial length in capacity hats on 10-meter dipoles and 2-element Yagis.

Folded Dipole Limitations

Fig. 3-6



Folded dipoles using dissimilar diameter wires add another dimension to the NEC-2 limitations. Consider two folded dipoles, as shown in **Fig. 3-6**. One consists of 2 parallel wires 0.5" in diameter spaced 0.25' apart and 16.1' long for 28.5 MHz. The other consists of one 0.5" diameter wire and one #12 (0.808" diameter) wire, also 0.25' apart and 16.2' long for 28.5 MHz. The results of modeling these antennas in both MININEC and NEC-2 (models 3-8 and 3-9) are as follows:

Antenna Dimensions	Free Space Gain DBi	Source Impedance (R+/-jX Ohms)
1. 2 x 0.5" dia; 16.1' long		
MININEC	2.22	285.7 + j 0.9
NEC-2	2.22	285.9 + j 4.1
2. 0.5" dia and #12; 16.2' long		
MININEC	2.21	530.5 + j 1.5
NEC-2	0.69	375.2 + j25.8

Both systems model the standard folded dipole with very reasonable accuracy. The second, non-standard, folded dipole with dissimilar wire diameters is another matter. Standard textbook equations for calculating the impedance of folded dipoles with dissimilar diameters yield a projected ratio for the source impedance of the folded dipole relative to a single-wire dipole of nearly 7.5:1, or between 530 and 535 Ohms. While the MININEC model falls in the ballpark (considering that the formula does not account for antenna shortening or end connections between the two wires), the NEC-2 model is clearly unusable.

The end result of exploring these limitations is this: wherever NEC-2 is to be used with wire junctions or closely spaced wires of dissimilar diameters, extreme caution must be used to independently check the reliability of the reported performance specifications.

Sommerfeld-Norton Ground

Unlike MININEC, NEC-2 has provision for a highly accurate ground effect calculation system variously called Sommerfeld-Norton, SOMNEC, and other names. Although NEC-2 also has a “fast” ground calculation algorithm (6-8 times faster in solutions than the S-N system), it is far more limited and less accurate than the S-N system. Hence, wherever calculation of ground effects is critical to antenna design, only the more accurate system should be used. This applies to such modeling tasks as low dipoles (under 0.2 wavelengths up), elevated ground planes, and Beverage antennas.

The contrast between the results of the MININEC ground system and the S-N system are sufficiently vivid with low dipoles that I shall repeat a table presented in

the last episode. The following table compares NEC-2 (S-N) and MININEC data for a 3.5 MHz dipole (resonated in free space) at heights from 0.05 to 0.30 wavelengths above medium or “average” earth (conductivity = 0.005 Siemens/meter; dielectric constant = 13) (model 3-10).

Antenna		MININEC		NEC-2 (S-N Ground)	
Height		137.2' #12 copper		136.9' #12 copper	
W/L Feet		Gain	Source Z	Gain	Source Z
		dBi	R +/- jX	dBi	R +/- jX
0.05	14.05	9.4	7.4 - j 4.9	1.2	48.9 + j15.4
0.10	28.10	8.4	23.3 + j20.5	5.1	49.8 + j21.1
0.15	42.15	7.7	45.9 + j35.1	6.4	62.5 + j26.9
0.20	56.20	7.0	62.3 + j37.0	6.5	77.0 + j25.3
0.25	70.26	6.2	87.7 + j28.3	6.2	87.8 + j17.3
0.30	84.31	5.9	97.4 + j13.5	6.1	92.3 + j 6.1

Despite the clearly more reliable figures produced in NEC-2, the use of the S-N ground is not without some limitations. For example, NEC-2 is sometimes used to simulate surface ground radial systems with vertical antennas by placing the radial wires very close to the ground. One recommendation sets the minimum height at 0.0001 wavelength, with segment-length tapering techniques applied between that height up to 0.001 wavelength. For frequencies below the 80-meter ham band, some ground-wave measurements suggest that elevated radial models yield overly optimistic gain figures. Consequently, the limits of the S-N ground system should be approached cautiously.

Another NEC-2 Potential: Transmission Lines

In addition to providing a more accurate ground modeling system than MININEC, NEC-2 also offers the modeler another unique provision: the introduction of transmission lines into the antenna structure. Transmission lines can be used to connect antenna elements, as in phased arrays or LPDAs. They can also be shown as feedlines to an antenna by the addition of a very short, remote wire to provide the line with a terminating point and for placement of the antenna source. Transmission lines can also be shorted or left open by various techniques and thus used as inductively or capacitively reactive stubs.

The transmission line models used in NEC-2 are mathematical and non-radiating, in contrast to the wire elements, which can be classified as “physical” or radiating. Wire elements enter into the matrix calculations and contribute to far-field and other antenna performance specifications. However, transmission lines do not enter into far-field calculations. For example, providing a dipole with a transmission line will not yield results that show any radiation from the line.

In addition, transmission lines in NEC-2 are lossless. Therefore, models using them will not reflect losses incurred in phasing lines, load lines, and similar applications. Determination of those losses must be done by separate calculations. Feedline losses, for example, can be calculated by such programs as N6BV’s TLA, which is readily available.

As with the sampling of MININEC limitations, our purpose in surveying some of NEC-2’s limitations is not to cast doubt on the utility of the program as a highly competent antenna modeling software core. Quite the opposite: by being alert to the program’s limitations, we can avoid producing and relying upon models that cross the limit lines. Staying within the lines is one key to productive and satisfying antenna modeling.

Although NEC-4 is beyond the scope of this column, due to its sparse use in amateur circles, an account of some NEC-4 limitations appears in the May-June, 1998, issue of QEX.

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Models included: 3-1 through 3-10. Note: as models become less reliable, expect to see wider variations in output reports when moving from one version of NEC-2 to another.



4. A Good Start is Half the Trip

Modeling programs are a paradigm of the old saying, “Garbage in, garbage out.” However, I prefer to think of this saying in its more positive form: “Good models in, good information out.” And getting good information into the NEC or MININEC program is fully half the task in modeling antennas.

A good model does not begin at the computer, but on a piece of paper. Not just any piece of paper, but some kind of useful form on which you do all the preliminary work so that you can just transfer the values into the program. (Given the variety of programs, transfer can mean typing up and saving an ASCII input file, entering values in screen tables and entry points, or entering values and selections into a spread-sheet form and pull-down boxes.)

Fig. 4-1 shows a form that I have used for a number of years for common HF and VHF antenna types. Although you can print out the form, you might want to make up your own crisper and more individualized form on a word processor.

Let's divide the form into 4 sections and see how it can help our model input work.

1. *The header:* The top lines of the form provide basic information about the model file and the background values within which the antenna will be modeled. The top line forces you to give the model a name—one that is either specific or generic—along with a file name and directory location. The last two items are important once you begin to develop collections of models. It is aggravating to lose ten minutes trying to remember where you stored a file.

Antenna:_____ Filename:_____ Directory:_____

Band(s): _____ Frequency (ies): _____ Height: _____ No. of Elements: _____ Units: _____

Ground Type: _____ Specs: Cond: _____ Diel. Cons.: _____ Material: _____ Cond.: _____

Antenna Sketch:

Fig. 4-1

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[illegible]

The second line begins the modeling information. The band(s) entry allows you to specify one or more bands on which the antenna may be used, while the frequency entry is for either the antenna design center frequency or the upper and lower frequencies of a sweep you may intend to perform. The antenna height, even though it may be repeated in the wire table, is a reminder of the intended use of the antenna. Alternatively, if the model is preliminary, you may wish to enter “FS” for free space. Recording the number of elements is a check on the wire table entries below, which may consist of many wires. Recording the units of measure for the antenna elements can save you from trying to mix units later.

Recording the ground type and specifications (if other than free space) should be a standard part of your modeling practices. For horizontal antennas, such as dipoles and beams at HF, the use of a ground other than the usual program default may not make significant differences in the data produced by the program for the model. The default for most implementations of NEC and MININEC is “Good” or “Average” soil, with a conductivity of 0.005 S/m and a dielectric constant of 13. These values are taken ultimately from studies performed in the 1930s and often repeated in antenna handbooks.

Some programs provide some preset soil types, and the instruction manual will generally describe their values and characteristics—or refer you to the source handbook from which they came. However you obtain the ground values, be sure to enter them both here and in the program. These values are especially important for vertical antennas, whose performance may vary with either or both the ground type directly under the antenna and the ground type at a distance from the antenna. If you plan to use multiple grounds (for example, one in the immediate vicinity of the antenna and another at a distance), you might make an abbreviated entry here and develop the details on unused lines further down the form.

Modeling programs may specify the antenna material in one of two ways: either as a material type (such as “copper”) or as a specific value. The program may call for either a conductivity or a resistivity value, where one is simply the inverse of the other. For example, copper has a conductivity of 5.8×10^7 (where “E 7” means “ 10 to the power of 7”) S/m, which is equivalent to a resistivity of about 1.7×10^{-8} Ohms/m. (Some programs will still use “mhos/meter” instead of “Siemens/meter.”)

As a modeler, you always have the choice to begin your modeling with zero-loss wire. However, it is usually good practice to specify the antenna wire material in terms of the actual material you intend to use (or are using) in order to take its distributed losses into account. In many cases, the differences between zero-loss wire and actual materials may make only insignificant differences in the output data. However, this fact will not be true for all antenna designs. In addition, some programs require a single material for the entire antenna, while others permit material or resistivity/conductivity specification on a wire-by-wire basis. If limited to a single material for some design combination of copper wire and aluminum tubing, you can choose the material that is quantitatively dominant in the design, the one most prominent near the high current regions of the antenna, or simply the one with the higher losses.

2. *The antenna sketch:* To get an understanding of the importance of sketching the antenna to be modeled, let's look at **Fig. 4-2**, the data sheet for a simple 10-meter aluminum dipole.

On this data sheet, the header is completely filled in with the basic and background data. Note that for this model, "Free Space" has been selected as the modeling medium. Note also that the material is 6063 aluminum, whose conductivity value may differ from other types of aluminum that might have been used.

The sketch itself should be (and this one tries to be) as complete as possible. Not only does the drawing show the overall dimensions, but as well, it displays the lengths and diameters of the individual sections of the dipole elements. Since each change of diameter represents the stop/start point for a wire in the model, complete dimensional data is necessary in order to enter the wire geometry correctly.

Although my CAD program has permitted me to overdo the graphical aspects of the antenna sketch, a pencil sketch will normally suffice. However, strive to make it as clear and complete as possible. Remember that it is the fundamental reference for all other modeling work you do.

3. *The wire chart:* NEC and MININEC operate on the basis of wires that are segmented according to rules. A single antenna element may consist of more than one wire, if the diameter changes along the way. In our example, we have two 3/16" diameter end pieces and a 1/4" diameter center section (model 4-1).

In reality, the center section may consist of 2 pieces of aluminum, slightly separated at a center feedpoint. However, the source (the feedline) is considered to be in series with the segment to which it is connected, and therefore, the wire should be continuous. You can accomplish this by specifying a single 1/4" wire and placing the source at its center. In NEC, you will select an odd number of segments for this wire to ensure source placement at the exact center. In MININEC, you will select an even number of segments to accomplish the same goal. Alternatively, you may choose to create two wires, one for each side of center. By placing a MININEC source at the end of one, it will be correctly placed where the other begins. In NEC, the same task requires selecting or creating a split feed to include the last segment of one wire and the first segment of the next.

In this model, I chose a symmetrical arrangement of the element, using the zero point as the center. I also chose to extend the element length along the Y-axis. Since the antenna has only length, the X-axis is always zero, and since it is in free space, the Z or height axis is also zero. (In most programs, the Z axis can be anything when modeling in free space, and you may want to enter for Z the height of intended use for the antenna.)

Selecting the Y-axis for the antenna length values over the X-axis may or may not be an arbitrary decision, depending upon the program you are using. Some programs, like AO, have a provision for increasing calculation speed via symmetry. If the antenna elements are centered—but only on the Y-axis—the program can calculate only half the antenna and treat the other half as a mirror image. This feature must be disabled for non-symmetrical antennas or for antennas whose length is specified along the X-axis.

In many programs, it makes no difference to calculation speed whether the antenna element lengths appear in either the X-axis or the Y-axis. In some programs, especially when surveying the wire length values of a very complex but linear antenna, it is more visually convenient to scan the left-most X-axis to check for errors. In the end, however, try to adopt one consistent convention for most—if not all—of your models. Swapping axes as you hop from model to model can be a source of unintended errors.

Among the most important conventions to adopt is to begin at one end of each antenna element and to proceed from that end to the other without changing direc-

tions in mid-stream. In the figure, note that the model proceeds from left to right along the sketch, symmetrically with respect to the zero-point of the Y-axis. Hence, a 16.78' antenna becomes $\pm 8.39'$ long.

Notice also that end 2 of the first wire becomes the starting point or end 1 of wire 2. For continuous antenna elements, be certain that the X, Y, and Z values of joined ends are the same. Some programs will indicate when wires are connected; others will rely on you to ensure that the junction ends have the same values. NEC connects wires that are within very tiny distances of each other, so also be certain that, if you do not want wires connected, they are far enough apart.

Wire entries also require that you specify the wire diameter. Many programs allow you to specify common American AWG wire gauges, which is useful for copper wire antennas. These are converted for the program into actual wire diameters. To be more precise, they are converted into the wire's radius. If you look at the NEC input file in programs that permit this, you will see the converted value of the diameter you entered in the input page or file.

You must also specify how many segments the wires will have. In order to make the segments for each wire reasonably close in length, the end wires in Figure 2 have 5 segments (0.78' per segment) and the center wire has 11 (0.82' per segment). Since the center wire containing the feedpoint (source) has an odd number of segments, it is clear that this model is intended for NEC rather than MININEC.

4. *Source and other data:* Missing from the form is any clear specification of the source or feedpoint for the antenna. (Actually, on my own forms, the source specification is on the reverse of the page, along with entry areas for loads, networks, and transmission lines—all subjects of future episodes in this series.) You can and should enter the source specifications on a spare line below the wire lines. For most antennas, this will place the data where it is most usefully compared with the wires themselves.

Enter not only the source location, but also the type (voltage or current, single or split) within the limits allowed by the program. We have already seen the utility of split sources. Single voltage sources are the native environment of most modeling programs. However, specifying a current source and giving it a value of 1 is often

very useful for comparing current magnitude and phase output data for each segment along the antenna element(s).

This model is now ready for conversion into the appropriate program inputs.

Some Other Element Entry Considerations: In our fairly simple dipole entry, we had little difficulty in proceeding from element-end to element-end. However, some antenna designs may not make this process so simple in all cases.

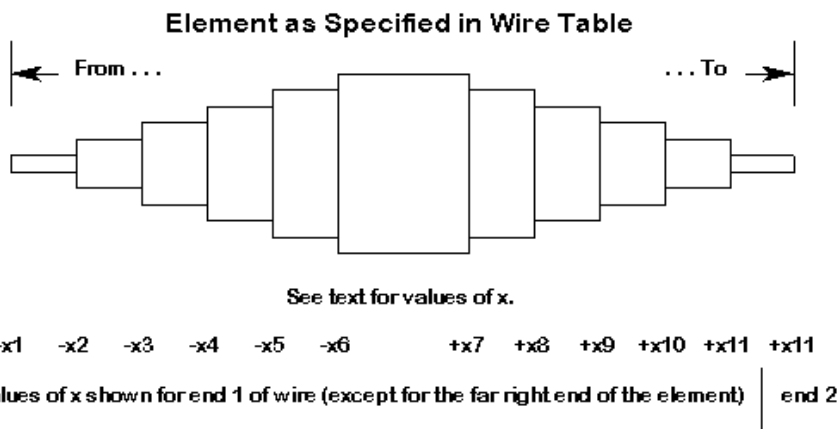
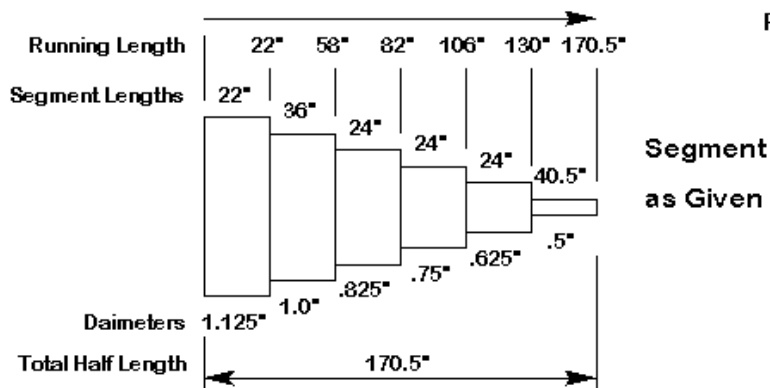
Having examined numerous antenna models from a wide variety of sources, I am struck by the diversity of ways in which modelers create antenna elements and subdivide them into wires and segments. The variety of ways in which wires are placed into the antenna geometry descriptions often makes it difficult for another modeler to read the wire table. Moreover, many of the models will produce correct far-field data and feedpoint information, but will skew the antenna currents. This latter information can become quite important in analyzing why a complex antenna yields its particular set of performance outcomes.

The usual reasons given for the odd collections of wires making up an antenna model are convenience and speed. Many complex antenna element specifications are for half elements, starting at a boom or center point and moving toward the element tip. The specification presumes that other half of the element is a mirror image of the first. So modelers tend to begin in the middle and to work outward. Specification tables usually give the dimensions using positive numbers. Since it is faster to do the positive side, modelers then replicate that side and insert minus signs. Moreover, they tend to go directly from whatever sketch or data sheet they have to the screen entries. I have learned the hard way that slowing down at the beginning saves me a bucket of time later on when I try to troubleshoot my model.

Let me make some suggestions, scratched with chalk, not etched in stone. They began with my suggestion of always working from one end of the antenna element to the other end, and using the same directional sense for each succeeding antenna element.

Now consider the element shown in **Fig. 4-3**. It is one element of a multi-element Yagi. Each element uses many shorter sections of tubing that decreases in diameter as one moves outward from the boom to the end. Let's assume an element that from center outward has these dimensions: 22" at 1.125" diameter; 36" at

1"; 24" at 0.825"; 24" at 0.75"; 24" at 0.625"; and 40.5" at 0.5". The first step is to obtain running totals: 22"; 58"; 82"; 106"; 130"; and 170.5".



If we follow our conventions, the full element will resemble the entries following the figure in the table. It will look the same on paper and in the geometry entry spreadsheet (noting that I have in this case chosen to place length values in the X-axis column).

Wire End 1	x	y	z	End 2	x	y	z	Diameter
1	-170.5	0	0	-130	0	0	.5	
2	-130	0	0	-106	0	0	.625	
3	-106	0	0	- 82	0	0	.75	
4	- 82	0	0	- 58	0	0	.825	
5	- 58	0	0	- 22	0	0	1	
6	- 22	0	0	22	0	0	1.125	
7	22	0	0	58	0	0	1	
8	58	0	0	82	0	0	.825	
9	82	0	0	106	0	0	.75	
10	106	0	0	130	0	0	.625	
11	130	0	0	170.5	0	0	.5	

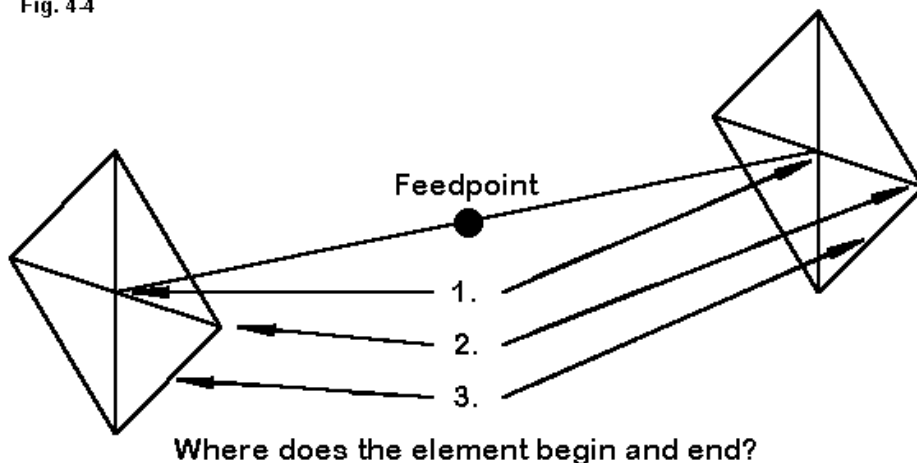
(In some programs that permit symbolic entries, the numbers may be replaced by a set of variables indicated by a letter or combination of letters, with a list of values for these variables set out elsewhere.)

This process may seem like a long way to go, but for an accurate model, all these wires will have to be in the chart anyway. This left-to-right scheme just keeps them well ordered. Note that wire 6 is the center section and can have the feedpoint specified at its center, with correct segmentation. For NEC, I might assign it 5 segments; for MININEC, 4. The sections with dimensions in the 20s might get 2 segments each, with the 36" section getting 3 segments, and the end sections 4 segments. (See model 4-2.) This procedure keeps segment lengths as close to equalized as this structure permits, while holding the segments lengths well below maximum recommended length.

Now let's consider a second example: a shortened dipole with wire hats on each end (**Fig. 4-4**). The hats consist of 4 radial wires and a perimeter wire. Where does the antenna element start and end? Some quick modelers use the horizontal dipole ends and then separately model the wire structures outward and around. However, the wire hat assemblies are part of the antenna. An equally quick second answer is to treat the peaks where the radial wires end as the antenna beginning and ending. Both answers are wrong. The antenna begins and ends where the current goes to zero. MININEC gives a true zero reading, because it takes current nodes to be at the ends of segments. NEC takes current nodes to be at the center of segments,

and so the lowest value will never be zero. However, you can get equally low values either side of a correctly chosen wire junction.

Fig. 4.4



For this example, the antenna begins in four places: at the center of each of the four perimeter lengths. It ends in the comparable places at the other end assembly. Hence, a fully modeled version of this antenna will have, starting on the left end, 8 perimeter wires, each starting mid-length and ending at a radial wire peak. The 4 radial wires come next, working from peak to center, which happens to be the horizontal wire end point. Then we have the horizontal wire, center fed, followed by 4 radial wires working outward from the hub, and finally the 8 perimeter wires, 2 from each peak to common centers.

In my wire tables, I tend to collect these wires by type so that I can adjust all the radials as a group, adjust all the perimeter wire lengths as a group, etc. (Model 4-3)

You can take shortcuts with element structures like the one in the example, but only if a. you can decide in advance that you will never need or want to know the correct current magnitudes and phases on each wire, or b. you are willing to reconstruct the model if you should ever become interested in such matters.

One More Example: Let's close this exercise with one more example. See **Fig. 4-5**. This example is not much more complex than the dipole. However, the differences between the two are interesting. The Yagi design uses inches as the unit of measure and a different type of aluminum from that used for the dipole. Moreover, the antenna is to be modeled over real ground rather than in free space. Both the sketch and the wire table reflect the choice of inches and a ground—the latter evident by the use of a Z-axis value other than zero. (Some programs, such as NEC4WIN, allow you to enter the actual antenna height independently of the Z-axis value.)

In the wire table, element length values appear in the X-axis column. The element spacing is 57.6" but the reflector is set at the Y-axis value of zero, with the driven element set forward of it. This is one of the two most common conventions for single-driver Yagis. The other convention is to set the driven element at zero, with reflectors given a negative value in the spacing column and directors given positive values in that column. Either convention will work, but once more, it pays to select one convention for all (or at least most) of one's work.

Finally, the model is set up for MININEC use, since the driver center wire has an even number of segments and runs symmetrically on either side of the feedpoint at its center. (See model 4-4, a NEC model.)

If racing through the sample models leaves the impression that there are many things to keep track of in correctly setting up even simple models, then the importance of preliminary paperwork should be evident. A good paper model reduces the number of errors and makes the ones that do occur easier to find. The paper model is fully half the trip from antenna idea to useful output data.

* * * * *

Models included: 4-1 through 4-4.

5. Putting Sources Where You Can Find Them

Very often, a model that approximates a desired antenna is good enough. In some case, we need more precision. Making the antenna dimensions precise is easy work: we simply carry out the length and wire diameter to more decimal places.

Placing sources and loads precisely can be more work if they do not happen to fall at the center or end of a wire. How can we place them with high precision relative to an existing or proposed antenna?

There are a few techniques that work very well, although some of them have attached pitfalls that we would want to avoid. We can illustrate the techniques if we contemplate a hypothetical problem demanding great precision. For most of us, the problem will be somewhat unreal, but we will not let that get in the way of seeing how the techniques of source placement work. In this episode, we shall look at sources only, reserving loads and their special requirements for next time.

Consider a wire antenna for 3.5 MHz made up of #14 AWG copper wire. The antenna will be at 70' above average soil. (For modeling purposes, NEC models were run over the high-accuracy Sommerfeld-Norton ground, which is not available in public-domain MININEC 3.13. The 70' antenna height places the antenna about 0.25 wavelength above ground, within the accepted limits for MININEC. However, in this area of transition, some variance in the resonant length of antenna is to be expected between NEC and MININEC.)

If this 80-meter antenna were a standard center-fed type, we would simply place the source in the center. With MININEC, this requires an even number of segments; with NEC we need an odd number of segments. If we further suppose that we want to resonate the antenna so that the source impedance showed less than ± 1 Ohm reactance, we would simply adjust the total length until we achieved this goal. If we set the source at the 0,0,70 point, with equal wire lengths extending to either side of this point, then we would play equally with two end values, but keep them equal to each other.

This is the standard easy case. So let's make the problem a little harder. Suppose we want our wire to be about a half wavelength long, to be resonant within an Ohm or two, and to have the source placed at the point where the resonant impedance is 300 Ohms. Granted, for building purposes, something close to this would do fine, but remember that we are (for some unspecified project reasons) demanding precision.

The 1-Wire Method

"Standard" Source Placement Procedure

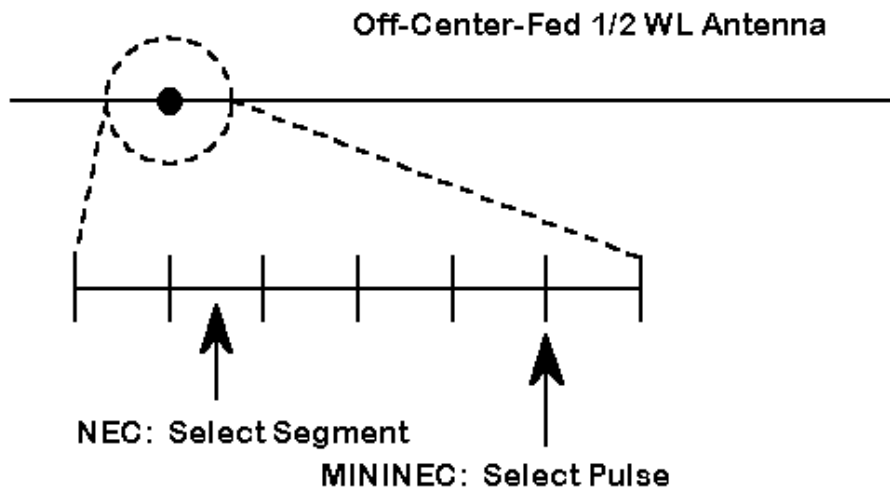


Fig. 5-1

If your program allows, you can approximate the 300-Ohm feedpoint by moving the source off center, as shown in **Fig. 5-1**. In MININEC, we select a pulse number, which is always at the junction of two segments. In NEC, we select a segment and

visualize the source as being at its center. Of course, we use sufficient segments in each case to ensure convergence.

To get ahead of myself, I shall specify that the NEC antenna is 136.55' long. Because uncorrected MININEC does not correlate perfectly with NEC, the MININEC antenna will be 136.01' long. These lengths are based on using one of the techniques to come, but for the moment, let us just accept them. Remember also that different implementations of both NEC and MININEC may return slightly different values, based on a host of reasons related to input and output operations, rounding, and other similar factors.

Since we are trying to be precise, let us choose 136 as the number of segments for our MININEC antenna and 137 segments for the NEC antenna. This move yields segments about 1' long for the two antenna, each segment being about 0.0036 wavelength long, well within limits for each program.

In MININEC, if I place the source at pulse 23, it is 16.91% in from one end (just about 23.0'). The source impedance report is $317.50 - j1.91$ Ohms. Again, this precision is excessive for most practical purposes, but for this example, the numbers will show some things about both the models and the programs. If I move to pulse 24, then the source is 17.65% in from one end (about 24.0'). The impedance report is $295.37 + j1.62$ Ohms. Pulse 24 is close to the goal. Moreover, the antenna is close to resonance, and we would expect that the 300-Ohm point to show under 1 Ohm reactance. So we might well estimate the “true” 300-Ohm source point at about 23.6 to 23.7 feet in from one end. However, unless we were to be very lucky in re-segmenting the antenna to find a pulse very close to that point, we cannot get more precise.

The case with NEC is similar. Using segment 24, we get a source that is 17.15% in from one end or about 23.42 feet (remembering that the NEC antenna model—5-1—is not the same length as the MININEC model). The source impedance report is $314.3 - j3.32$ Ohms. Moving to segment 25, which is 17.88% in from one end or 24.42 feet, we find $292.9 + j0.18$ Ohms. The 300-Ohm point is between the two, perhaps at about 24' in from one end. Moreover, we might expect it to show something over 1 Ohm capacitive reactance. Again, we can juggle segmentation and hope to hit the mark.

Surely there must be an easier way. In fact, there are at least two.

The 2-Wire Method

Some modelers only create new wires when there is a difference between them—perhaps diameter, direction, or something else that corresponds with visible reality. Within MININEC and NEC, there are no rules against creating separate wires for what is in the real world a single length of copper wire. Therefore, let's rebuild our model of the off-center-fed 1/2 wavelength wire in accord with **Fig. 5-2**.

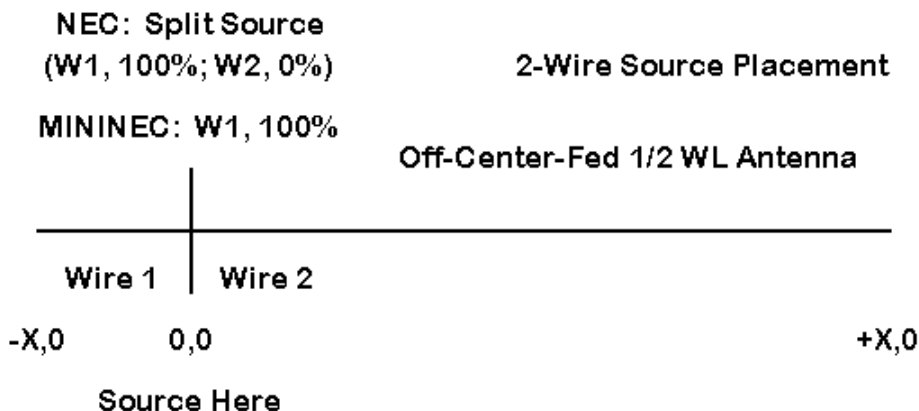


Fig. 5-2

In MININEC, we shall initially let Wire 1 run from about -23.6' to 0, with Wire 2 running from 0 to the +112.4' mark. We shall place the source on Wire 1 at the 100% mark, which forces it to the junction of the two wires. We shall leave our segment total at 136, with 24 on Wire 1 and 112 on Wire 2. The aim is to keep the segment lengths on the two wires roughly equal. We might reduce the number of segments on each wire to proportionately lower numbers for speed of running the model. However, the segment lengths might change more significantly if our first estimates prove to be somewhat off.

Now, by adjusting the end lengths only, we can establish both the 300-Ohm source impedance point and resonance within whatever limits we set for the problem. In fact, the model at hand required that Wire 1 be 23.78' long while Wire 2 had to be 112.23' long to yield a source impedance of $299.87 + j0.54$ Ohms with a voltage source.

Selecting a voltage or a current source in MININEC yields identical source impedance numbers for center-fed (symmetrical) antennas—at least within the limits of the decimal places usually shown. When the antenna is not symmetrical with respect to the source, with identical segment lengths on either side of the source, moving between a voltage and a current source will show a slight, but normally wholly insignificant variance for any desired level of precision. The current source option yielded a source impedance of $299.96 + j0.73$ Ohms for this model. The differential also appears to lessen the closer the model is to absolute resonance (for example, when the reactance reaches 0.1 Ohm). I note these phenomena, but repeat that the differential is insignificant in MININEC.

In fact, the two-wire method is the method of choice for MININEC modeling for precise source placement.

In NEC, the situation is somewhat different. For our 136.55' long antenna, we can set up the wires as we did for MININEC, with wire 1 having an end coordinate of about -24' and using about +112.5' for the end of Wire 2. The two wires meet at 0,0,70. See model 5-2.

Placing the source for this antenna depends somewhat on the implementation of NEC that you are using. Some versions, such as EZNEC, permit placement of the source by specifying Wire 1, 100%, and choosing a Split Voltage source. The effect of this move is the same as setting up two sources, one on the last segment of Wire 1 and the other on the first segment of Wire 2. The program simply adds the two source impedances, which in other implementations, you might have to do manually.

For the model at hand, with 2 sources, the end coordinates for resonance and 300-Ohms proved to be -24.1' and +112.45' respectively. The source impedances returned were $152.4 - j0.45$ Ohms and $147.1 + j0.36$ Ohms. These total to 299.5 -

j0.09 Ohms. (A split-voltage-feed option selection produced 299.5 - j0.06 Ohms as its report.)

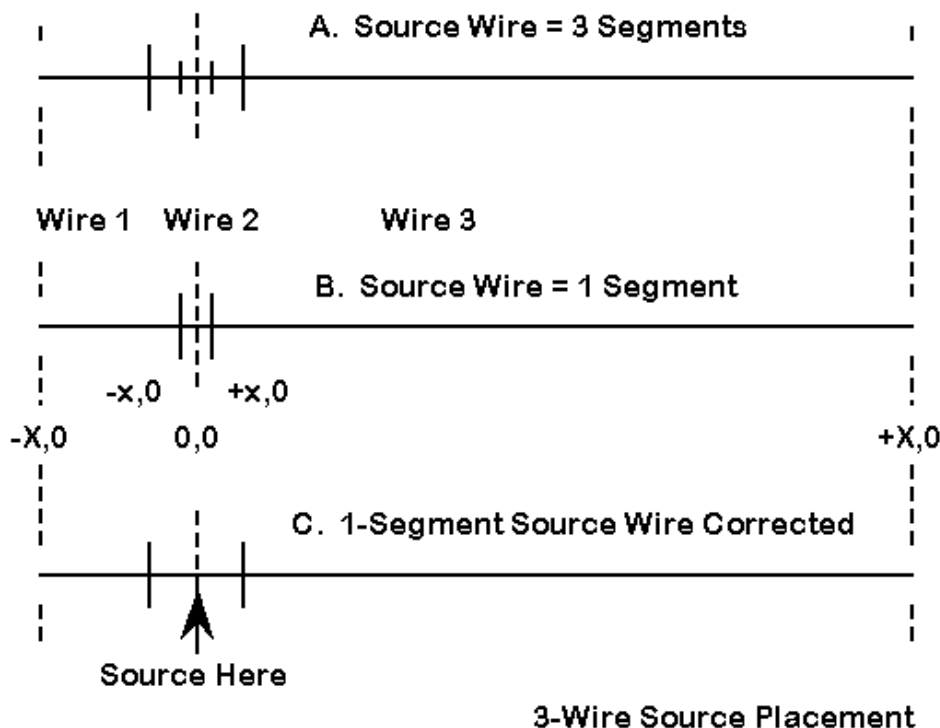
Notice that the two impedances are not identical. Although close in length, the segment lengths for Wires 1 and 2 are not identical. Moreover, the currents on each side of the 0,0 point are also not identical, since the feedpoint is off center. If we choose a current source for this model, we should expect some variance from the voltage-source option, since the current source option is generated via a network. In fact, the 2-current-source option produced values of 32.3 - j760.1 Ohms and 259.1 + j734.4 Ohms, for a net impedance of 291.4 - j25.7 Ohms. For precision work, this result is considerably at variance from the voltage source models. In general, where such variations appear with NEC models, the voltage-source option is usually the more accurate.

The 3-Wire Method

We can avoid any problems in NEC with differentials in reported feedpoint impedances by using a small modification of the two-wire method: the three-wire method. **Fig. 5-3** illustrates the idea, along with some problems and corrections we shall note along the way. (Model 5-3)

First, we shall set up Wire 2 by first settling upon a segment length and then centering a 3-segment wire length around the 0,0 coordinate. Since our model uses segment lengths of just about 1' each, we can set up Wire 2 as running from -1.5 to +1.5. Wire 1 then runs from about -24.1 to the junction with Wire 2. Wire 3 runs from the junction of Wire 2 to +112.45. Thus, we maintain our overall length of 136.55' for the antenna. (Note: once more, there is no rule that prevents the modeler from creating the wires from right to left. Nor is there any rule that requires the short end of the off-center-fed antenna to be on the left. My only suggestion is that, whatever conventions you use, you use them consistently in all similar models for the sake of ease of error detection and prevention.)

Fig. 5-3



Wire 1 and Wire 3 are now shorter than formerly, so they should receive 23 and 111 segments respectively to hold segment length very close to the 1' mark. The total is 137 segments for the model. This arrangement corresponds to part A. of **Fig. 5-3**. We place a single source of either type at the center of Wire 2.

The model reports a feedpoint impedance, using either voltage or current source options of $299.5 - j1.21$ Ohms. The slight capacitive reactance is just what we expected based on our original estimates derived from the 1 wire model where we inspected the source impedance on segments that bracketed the 300-Ohm point.

One dangerous temptation of using the 3-wire method is first to arrange Wire 2 as specified and then to decide that the long wires can use fewer segments. Since the antenna is, overall, only about a half wavelength long, we might easily reduce the Wire-1 segmentation to 4 and decrease the Wire-3 segment number to 19, a proportional drop. However, this move makes the segments adjacent to the Wire 2 segments 6 times longer than the segments in Wire 2. The source impedance now reported is $288.6 - j2.29$ Ohms, a significant difference from the case where all segments along the wires are close to the same in length.

If we maintain equal segment lengths with all three wires, we can reduce the center wire to 1 segment and also shrink its length to 1' by setting its coordinates at -0.5 and +0.5. This move is suggested in part B of **Fig. 5-3**. Bringing Wires 1 and 3 to join this new Wire 2 calls for increasing their segment numbers by 1 each to 24 and 112, for a total of 137 segments. The source impedance reported for this model is $299.5 - j1.10$ Ohms, the same for even the most critical purposes as the 3-segment wire 2 model produced.

However, the temptation to reduce the number of segments on the longer wires produces more radically erroneous results with a center wire having only one segment. If we retain the 1' center wire but decrease the number of segments on the end wires to 4 and 19 respectively, the model reports a source impedance of only $260.3 - 2.02$ Ohms. Notice that the resistive component is far off the reliable mark, although the reactance has not changed significantly.

The segments of the end-wires in our “temptation” model are about 5.7' long. We may retain this segmentation by increasing the length of the center wire to a like amount by setting its coordinates are -2.85 and +2.85. Then, with 4 segments to Wire 1 and 19 segments to Wire 3, the model reports an accurate source impedance of $299.5 + 0.76$ Ohms.

If the smaller model yields reasonably accurate results, why use the highly segmented model? This fair question has a fair answer. With longer segment lengths,

it is easier to develop inequalities of length without either knowing it or being able to correct the inequalities by the simple addition or subtraction of a segment here or there. With short segments, a growing inequality tends to show itself, especially if the lengths have a convenient correlation to the physical dimensions, as they do in this model sequence. Moreover, adding or subtracting a single segment from the end wires has a minimal impact on the length of each segment.

The 3-wire method of source placement is the most accurate and unambiguous for NEC. Moreover, using 3 segments in the source wire resists the effects of unequal segment lengths in adjacent wires more effectively than using only one segment in Wire 2, although the effects even on the 3-segment center wire can upset precision in some instances. As we noted in the beginning, these methods are applicable only where source placement needs to be as precise as possible.

I ran the same exercise, with a slight modification on MININEC, even though the 2-wire method is, in its simplicity and accuracy, the method of choice. For the exercise, the MININEC center wire had 2 segments to ensure placement of the source at the 0,0 point, roughly comparable to part A of **Fig. 5-3**. In this case, the wire ran from -1.0 to +1.0, with the connecting wires reaching -23.78 and +112.23.

MININEC returned a source impedance report of $300.20 + j1.35$ Ohms (compared to $299.87 + 0.54$ for the 2-wire method). Leaving the center wire untouched, the segmentation was reduced to 4 and 19 on the end wires. The source impedance reported was $298.72 + j5.94$ Ohms. Although MININEC shows more variance of reactance under these conditions than NEC, the overall MININEC impedance was considerably less affected by radical re-segmentation than was that reported by NEC.

Returning the center wire to a suitable length relative to the segment lengths on the outer wires, MININEC reported a source impedance of $298.66 - j2.37$ Ohms. As expected, the reactance is now closer to the value reported by the 136-segment model, although the resistive part of the impedance remains below the value reported by the more heavily segmented model.

In general, for this class of cases, MININEC is less sensitive than NEC to changes in wire segmentation for multiple wires making up a single antenna element.

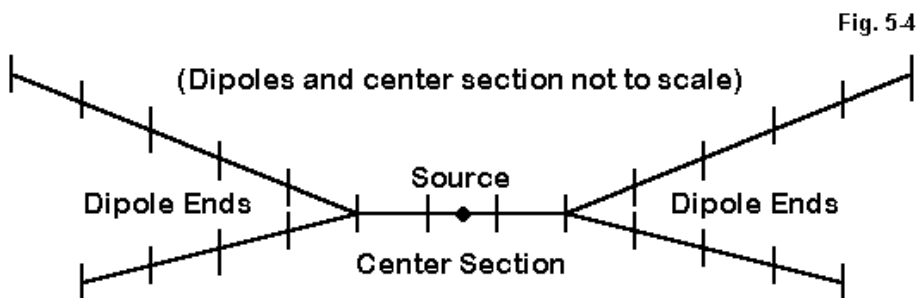
Applications

The off-center-fed $1/2$ wavelength antenna is not the only application requiring fairly precise placement of the source, both in models and in reality. For example, vertically oriented delta loops, used as self-contained vertically polarized antennas on the lower HF bands, call for off-center feeding of one of the vertically angular legs. The proper source point to maximize vertically polarized radiation, relative to the baseline of the antenna, varies with the shape of the delta. (The point is approximately $1/4$ wavelength from the apex of the triangle. This dimension, which may be close enough for many building purposes, requires optimization for vertically polarized use of the loop.)

Moreover, not all uses of antenna modeling software are directly geared toward either the construction of a proposed antenna or the analysis of an existing antenna. There are innumerable exercises one can perform with modeling software to understand better the operation of many types of antennas. For example, off-center-fed $1/2$ -wavelength antennas show interesting source impedance curves as the feedpoint is moved outward along the antenna. Moreover, the resonant length of the antenna changes with the feedpoint position and impedance. This is but one of a host of cases in which we can use our modeling software as a very effective self-instruction tool. The more we know about antenna performance in general, the more sense we can make out of the analysis of specific antennas and the more confidence we can have in the models we convert into real antennas.

In addition, when you pass the $1/2$ -wavelength mark in antenna size, there are ordinarily options in the placement of the feedpoint. A 1.5 -wavelength antenna may be fed at its center or about $1/4$ wavelength in from one end. These collinear arrays demand some precision in source placement not only as a guide to construction, but as well to understand the resultant radiation patterns.

The 3-wire technique of source placement, using a 3-segment source wire becomes virtually mandatory when modeling multiple dipoles or inverted Vees, each with a different frequency, using a common feedpoint. **Fig. 5-4** shows the general situation in NEC terms.



Method for modeling antennas with a common feedpoint

Using 3 segments on the source wire tends to isolate the source segment from current variations created by the pairs of half-dipoles. (Model 5-4 provides a simple example using copper wire dipoles for 20 and 15 meters.) We should strive to make the segment lengths on each wire of the assembly as equal as feasible.

The technique is also applicable to MININEC. The only changes required from the NEC model are to use two segments on the source wire and to place the source on the pulse at the center of this wire.

The techniques that we have explored in this column are also applicable to the placement of various types of loads along an antenna element. What a reactive load size needs to be, relative to typical antenna design goals, depends upon precisely where along the antenna element we place it. Moreover, traps require careful placement and design in order to be effective in an antenna element. We shall find (next time) that the techniques of modeling source placement explored in this episode have extensive application to the precise placement of loads.

* * * * *

Models included: 5-1 through 5-4.

6. Modeling Loads: What Kind, How Much, and Where?

Both NEC and MININEC permit you to add “loads” to the antenna at virtually any point along the wires composing the model. Loads are simply reactances (and associated loss resistances) inserted into the antenna to change its characteristics.

Let's look at loading in several steps, beginning with real-life loads, moving to model loads, and finally applying what we discussed last time (with respect to sources) to place the loads properly in our models.

Loads and Their Applications

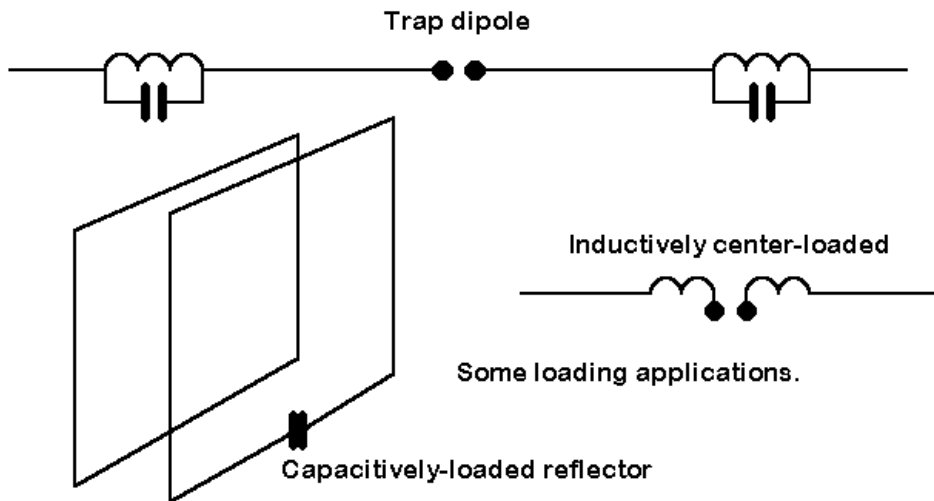


Fig. 6-1

Fig. 6-1 suggests several applications for loads. What they have in common is that capacitively reactive loads make an antenna element electrically shorter than it is without the load, and inductively reactive loads make the antenna element electrically longer than it is without the load.

We often find inductive loads used to shorten linear antenna elements. We place coils along a dipole anywhere from the center outward. Once we pass about the 50-60% point outward along the element, the coils of wire act more like lengths of wire than as inductors with reactance. Hence, we cannot model element end-loading with reactive loads, but must find a way to model those antenna extensions with wires.

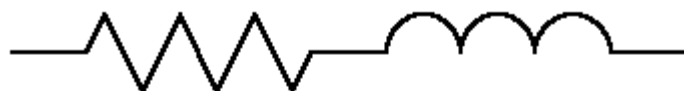
An alternative to the use of coiled or solenoid inductors is the use of parallel transmission line as the inductive loading element. A shorted transmission line less than 1/4 wavelength long will have an inductive reactance that is a function of the length of line and its characteristic impedance. Its characteristic impedance is a function of the wire diameter and spacing. Standard antenna references have the right equations (as does the HAMCALC collection by VE3ERP).

Every inductive load has a Q, a ratio of reactance to resistance (in the wire). Standard coils used in antennas have Qs that run from about 100 to 300 (with some specially made coils running higher), while parallel transmission line loads (also called linear loads) generally have a Q of over 1000. These numbers will have high interest for us in modeling.

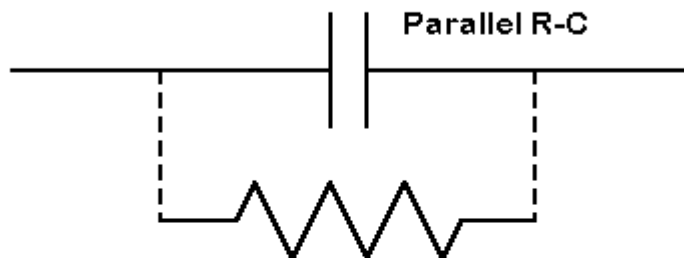
Capacitive loading is less used, since there are fewer cases in which we want to make an antenna physically long and then shorten it. One application is fine tuning the reflector of a quad beam. Capacitive loading is often easier to adjust than the inductive loads on short reflector loops. The Q of capacitors at HF is usually high enough that modelers ignore the resistive losses.

Coil losses are predominantly series resistance losses, so we can think of the coil as a resistor and an inductor (or inductive reactance) in series. Capacitor losses tend to be in the form of parallel resistance across the capacitance (or capacitive reactance). **Fig. 6-2** shows the basic schematic forms that we need to keep in mind as we model loads.

Series R-L (or R - X)



Parallel R-C



some basic R-L-C (X) load circuits

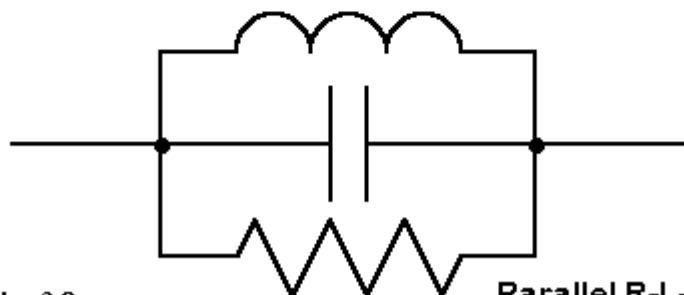


Fig. 6-2

Parallel R-L-C

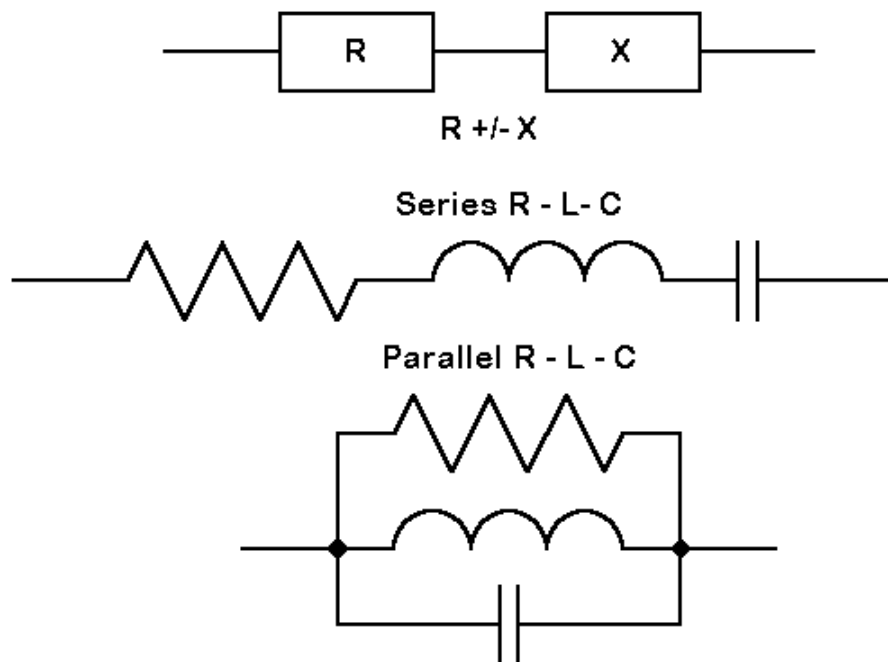
Modeling Loads

In an antenna model, loads are not physical structures. They are mathematical elements factored into the calculations. If an inductor used as a load in a real antenna has a diameter many times that of the element it is loading, we do not try to model the physical structure. Modeling programs view reactive loads as non-radiating elements. For all practical purposes this is true, although there is always some radiation from the physical structure of a coil, capacitor, or linear loading line.

Modeling offers three basic ways to model a loading device: as a series circuit composed of a resistance and a reactance (+ for inductive, - for capacitive); as a series circuit composed of a resistor, an inductor and a capacitor (with units in Ohms, Henrys, and Farads); and as a parallel circuit composed of a resistor, an inductor and a capacitor (with units in Ohms, Henrys, and Farads). Note: the core uses basic units, but the implementation of the core may provide a user interface calling for uH or pF, the most common units used in HF and VHF antenna work. The commercial program will convert your entries prior to the core run.) If you use either of the latter entry modes, and your circuit lacks one of the elements, put a zero in its place. The zero is not a short circuit: instead, the program knows to treat the zero as a missing element. See **Fig. 6-3**.

1. *Resistance-Reactance Loads*: It is tempting to use the series resistance-reactance mode of loading for single loading elements such as coils and capacitors. In many programs, we have direct access to inserting a series resistance and reactance, but inserting values for inductance and capacitance may be more complex. In many cases, the source impedance gives us clues to the reactance values needed. For example, in a shortened dipole, the reactance at the source point gives us the reactance value for a center-loading coil and approximately the value of each of two coils placed near mid-element.

If we know the required reactance for a loading coil, we can derive the series resistance from a measurement or a decent estimate of the coil Q. Never omit the coil resistance. If we need 200 Ohms of inductive reactance and estimate the Q at 200, then the series resistance is 1 Ohm. (As small as it seems, that 1 Ohm does affect antenna performance and needs to be included in the model.) At HF, you may ignore resistance with respect to simple capacitive reactances.



Loads in Models

Fig. 6-3

However, the series resistance-reactance load has a limitation. If we check the performance of an antenna across an amateur band, that is, away from the central design frequency, the load resistance and reactance in the model stay the same. Real coils do not act this way. They change their reactance as we change frequency. Therefore, the series resistance-reactance model may give an unrealistic picture of antenna performance.

For example, here is a 40-meter dipole with mid-element loads modeled with R-X loads and with R-L-C loads when viewed across the 40-meter band. The load design is a coil of about 12.93 microH, which has a reactance of 581 Ohms at the design frequency of 7.15 MHz. The coil Q is about 166, for a series resistance of 3.5 Ohms. (Models 6-1 and 6-2)

Frequency MHz	R + X Load		R-L-C Load	
	Source Z	50-Ohm SWR	Source Z	50-Ohm SWR
7.00	45.4-j33.5	2.003	44.3-j50.2	2.791
7.05	46.5-j20.6	1.536	45.7-j32.1	1.945
7.10	47.5-j 7.5	1.177	47.1-j13.6	1.330
7.15	48.5+j 5.6	1.123	48.6+j 5.4	1.119
7.20	49.7+j18.8	1.455	50.2+j24.8	1.634
7.25	50.8+j32.1	1.873	51.8+j44.8	2.352
7.30	52.0+j45.6	2.383	53.6+j65.4	3.299

Notice the overly optimistic report of the 2:1 VSWR operating bandwidth that the frozen R + X load gives to the antenna. The actual 2:1 VSWR operating bandwidth is not over 250 kHz, but a little over 150 kHz, as reported by using the R-L-C load.

2. Series Resistance-Inductance-Capacitance Loads: Simple loads using a single inductor or capacitor can also be modeled in terms of their inductance or capacitance (along with the resistance, as applicable). This is shown in the right-hand columns of our sample table, which used loads inserted in this way. By using values for inductance and capacitance, the program automatically calculates the reactance for each frequency checked, and a more realistic picture of antenna performance emerges.

The conversion equations for moving from reactance to inductance-capacitance and back to reactance are well known, but they are so important to modeling loads, that I shall repeat them here for reference. L = inductance in Henrys, XL = inductive reactance in Ohms, C = capacitance in Farads, XC = capacitive reactance in Ohms, R = resistance in Ohms, F = frequency in Hertz, and 6.28 is the value of 2 times PI.

$XL = 6.28 F L$ ——from inductance to inductive reactance

$L = XL / (6.28 F)$ ——from inductive reactance to inductance

$XC = 1 / (6.28 F C)$ ——from capacitance to capacitive reactance

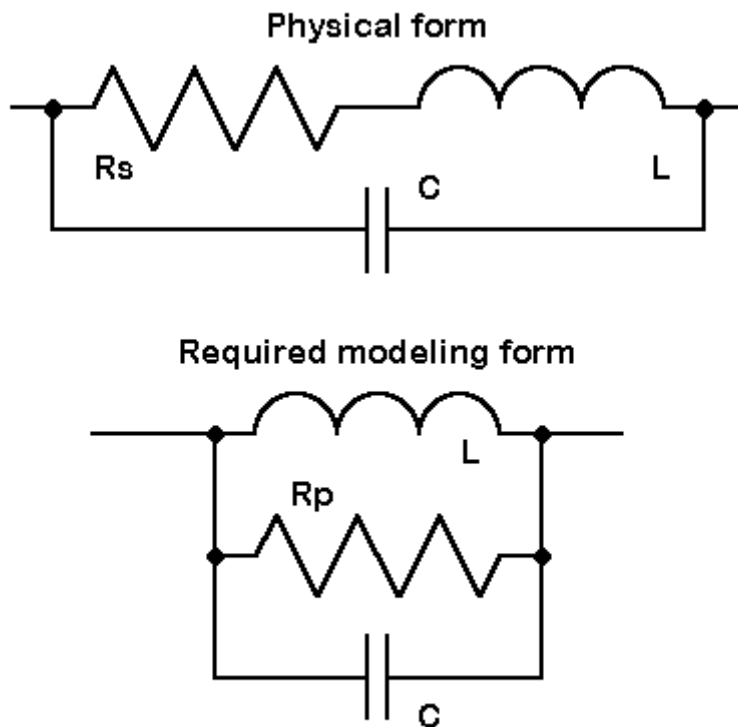
$C = 1 / (6.28 F XC)$ ——from capacitive reactance to capacitance

You may be working in basic units, even though the most common units required by antenna circuits for HF are in terms of microHenrys and picoFarads. Since most modeling programs use engineering notation, get used to expressing 65 picoFarads as 65E-12 and 5.5 microHenrys as 5.5E-6.

Suppose you use mid-element loads for a shortened rotatable 40-meter dipole, coming up with a reactance of 510 Ohms for each coil at the mid- band design frequency of 7.15 MHz. The coils have an estimated Q of 200, so we know the series resistance must be 2.55 Ohms. The required value of inductance is 11.4E-6 Henrys. We would enter the series resistance-inductance-capacitance values as 2.55,11.46E-6,0 (with the zero meaning no capacitor). These are entered as a string in some programs or plugged into spaces in a dialog box in other programs. (Model 6-3)

3. Parallel Resistance-Inductance-Capacitance Loads: For most amateur applications, parallel R-L-C loads have one primary use: for antenna traps. A trap is a parallel tuned circuit used to terminate a portion of an antenna element at a particular frequency by placing a high impedance reactance circuit at a particular place along the element. The circuit acts like an inductive reactance at lower frequencies, where the outer portion of the element comes into play again. Traps are normally tuned to a frequency just below the lower edge of the band for which they function as traps. A 15-meter trap, for example, is normally tuned to about 20.5 to 20.7 MHz or thereabouts.

How to calculate the operative values of resistance and reactance of traps for all frequencies of interest in an antenna will be the subject of a future column. For now, let's focus on getting the trap into the antenna as a load on the band for which it is a trap.



Trap Modeling

Fig. 6-4

Fig. 6-4 shows our problem. The trap consists of a coil and a capacitor in parallel, but the coil is actually an inductor and a resistor (for its losses) in series. Let's set the coil Q at 200 and tune the trap to 28.0 MHz. The coil is 1.1 microH and

the capacitor is 29.4 pF. By the conversion equations above, we get a reactance of 193.5 Ohms, and using the value of Q, we find the series resistance of the coil to be 0.9675 Ohms. We cannot directly enter these values into a model parallel circuit load because not all of the values are in parallel.

Hence, we must convert the series values for the coil reactance and resistance into parallel form. Let RP = equivalent parallel resistance, XP = equivalent parallel reactance, RS = original series resistance, and XS = original series reactance. ^2 means a number is squared.

$$RP = (RS^2 + XS^2) / RS$$

$$XP = (RS^2 + XS^2) / XS$$

For our example, summing the squares of RS (.9675) and XS (193.5) gives 37442. Dividing this number by the series resistance gives 39938 as the parallel resistance. Dividing 37442 by the series reactance gives 193.5 as the parallel reactance. Converting 193.5 into an inductance at 28.0 MHz returns a value of 1.1 microH. (The resistance was so small that it nudged the inductive reactance and inductance figures by too small an amount to show—less than 1/2 of 1%. However, the series and parallel equivalents for resistance are very different.)

Now we can enter our parallel values for R and L, along with the original value for C, for a parallel R-L-C load as 39938,1.1E-6,29.4E-12. (Model 6-4)

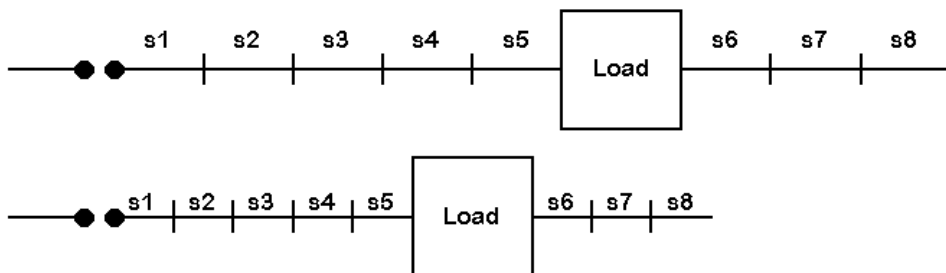
We may use these values at 28 MHz and throughout the 10-meter band. However, if the antenna has an extension for use at 15 or 20 meters, perhaps, the trap will have to be calculated for each of those bands. Although the values for L and C will stand, the fact that the reactance changes for the lower bands means that the equivalent parallel resistance will also change, and that requires a separate set of calculations.

Since our aim here is to create loads we can insert into antenna structures, this example will do until we can spend more time on the subject of traps.

Placing Loads

There is a fundamental difference in the placement of loads in MININEC and in NEC. If you are only a casual modeler and only want a trap approximately correct in its placement, then you may look at the placement question as being the same for both programs. Create a wire and place the load on the nearest pulse or segment, depending on whether you are using MININEC or NEC. Some programs give you an option of placing the load at a certain percentage along a wire, and this is handy. The exact placement will be on the pulse or segment nearest the one specified.

However, changing the length of the wire, for example, to bring it to resonance, will change the length of the segments. The load will be moved—slightly or significantly, depending upon the number of segments and hence on the length of each one. **Fig. 6-5** shows the problem.

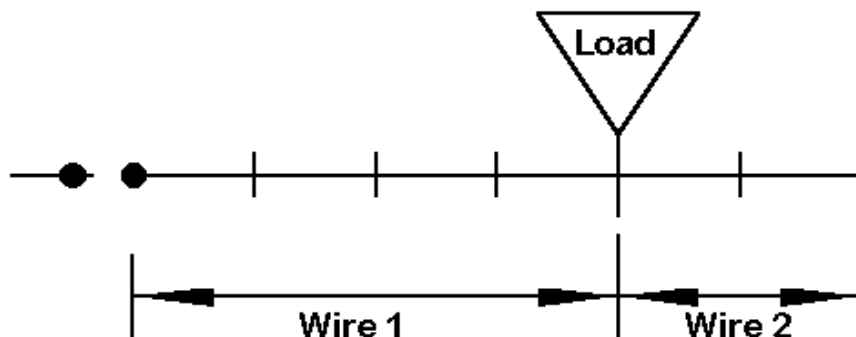


Effects on load placement by changes in single wire length.

Fig. 6-5

For more precise modeling, let's return to separate discussions of MININEC and NEC and see what we can do to set loads more exactly where we want them.

1. *MININEC*: Loads in MININEC go onto pulses, which occur at the junction of segments. This fact tells us more than one thing about loads.



2-wire load placement for MININEC

Fig. 6-6

First, to place a load precisely, we can create two wires that meet at the load position, as in **Fig. 6-6**. This maneuver permits us to independently change the length of the wires. For simple loads, such as loading coils, we can vary the length of the outer wire (or outer wires, in the case of a dipole) without moving the load as we fine tune the antenna model for resonance or some other characteristic.

If the load is a trap, we can separately change the lengths of the inner wire for 10-meters and the out wire for 15 or 20 meters. Only changes in the length of the inner wire will affect the trap position, which is exactly what we want to do. Changes in the outer wire leave the trap position unaffected.

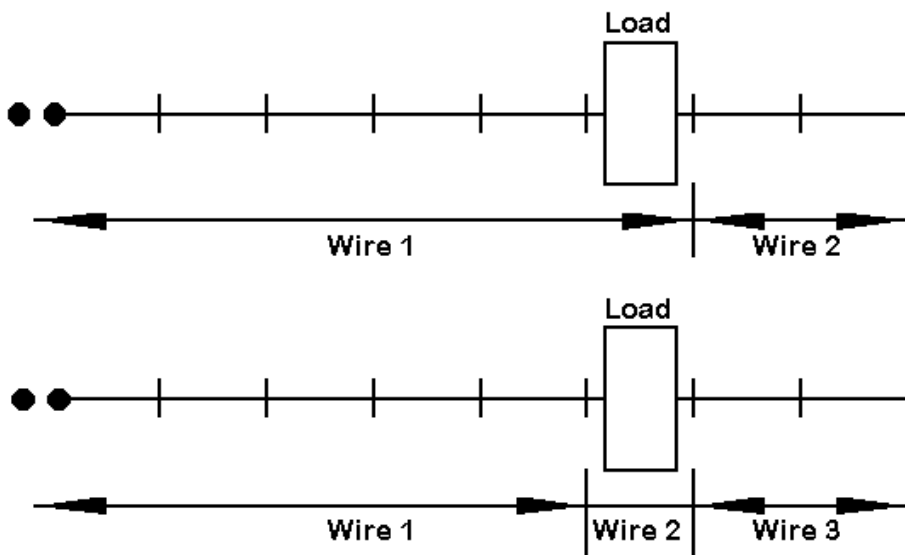
The only caution in the procedure is to makes the segment lengths on each of the wires approximately equal. This is not too critical with MININEC, but it is good practice in all modeling. The procedure we are describing here coincides with the procedures for precisely setting non-centered sources in MININEC.

Second, since MININEC loads occur at points along the modeled wire, they have no dimension. Trying to model a bulge in the antenna to account for the load's physical diameter is a largely superfluous effort, given the normal margins of error within modeling. However, if you wish to try it, you will need to create 2 more wires. End the inner wire by 1/2 the length of the load. Create a new wire of this length and the load's diameter. Set the load at the end of this wire. Create a second wire of the same diameter and length, with one end joined to the first new wire. To the second end of this wire connect the old outer end, which has been shortened by 1/2 the length of the load. Be sure that the inner and outer wires are segmented so that all of the segments are now about 1/2 the length of the load, that is, approximately equal in length all along the wires composing the element.

In MININEC, there are techniques of length-tapering to reduce the total number of segments required, and these are features of ELNEC, AO, and NEC4WIN. Moreover, NEC4WIN can handle the increased segmentation directly. However, in most cases, the exercise will yield no significant improvement in either the analysis of the antenna or in guidance for building a modeled antenna design. Real-world adjustment of the final product will still be required.

2. *NEC*: NEC places a load on a particular segment, and you may view the load as distributed along the segment, even though some modeling program graphics may show the load as a dot in the middle of the segment. In placing the load and structuring our model element, let's be guided by **Fig. 6-7**.

As with our initial attempt to place a load in MININEC, we may use a simple estimate in NEC along a single wire element. Then we simply place the load on the nearest segment to the exact position we may desire. Just as before, if we vary the length of the wire to achieve some antenna characteristics, such as resonance, we inadvertently change the load position slightly with each adjustment. For casual modeling, this procedure may suffice, but it may equally fall far short of the precision we might sometimes require.



2- and 3-wire methods of NEC load placement.

Fig. 6-7

An alternative is the use of the two-wire system used with MININEC. There is a difference, however. Placing the load on the last segment of the inner wire does not position it at the junction of the wires. Instead, the load will be on the last inner wire segment itself.

In most instances, end-of-inner-wire placement will be a correct placement, especially for traps. Traps are a part of the inner wire, which they terminate when operated in the trap frequency band. Whether using a simple load or a trap, the placement on the inner wire will lock the load in place. Adjustments on the outer wire will leave the load placement unaffected.

As with the MININEC version of the two-wire system, good practice dictates that the segment lengths on the inner wire and the outer wire be approximately equal. (Model 6-5)

You may also use a 3-wire system, placing the load on a separate wire of 1-segment length. Here the segment may simulate the length of the load or trap (with due caution for the segment lengths in adjoining wires). As with the two-wire system, the load is locked in position. (Model 6-6)

The natural tendency will be to use this system in order to set the diameter of the 1-segment loaded wire equal to the diameter of the load. This procedure is not disallowed, but must be used with extreme care. If the element diameter tapering correction does not function when the element has loads or when the diameters do not follow some regular progression, then NEC-2 will give erroneous results. In NEC-4, which does not require the correction system for tapered diameter elements, too large a shift in diameter will require either that a very large number of segments be used to achieve convergence as a test of the reliability of the results—or that convergence may not be achieved. In either case, trying to model the load diameter usually proves to hold more problems than it solves.

In the End. . .

For both MININEC and NEC, the 2-wire system of load placement is adequate for almost every modeling situation that may occur with amateur radio antennas. With this settled, the modeler can focus on other important questions surrounding loads. Learn to create series and parallel R-L-C loads in order to go beyond the more restrictive R-X loads—and to use each in its proper place in the development of antenna designs and analyses. Load creation and placement together form a single enterprise that can enrich your modeling work.

* * * * *

Models included: 6-1 through 6-6.

7. Maximizing Your Data

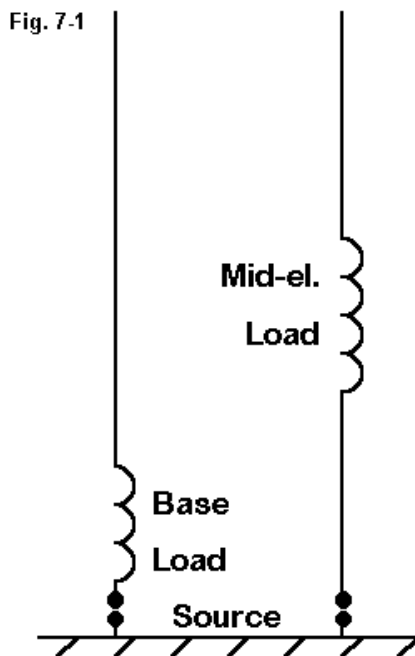
Now that we can pick and place loads to best advantage, do we have a good handle on how best to use them? Many newer modelers restrict themselves to trying to design an antenna, analyzing the performance of a given antenna, or modifying an antenna. Rarely do they develop methods for teaching themselves some very useful things about antennas, using the modeling program as a foundation.

We can teach ourselves much about antennas if we learn to do some systematic modeling. What “systematic modeling” means is best told by way of an example. We can think of it as a model of modeling. It is adaptable to any number situations in which we want to maximize our knowledge and understanding in order to make intelligent decisions about antenna construction options. In this instance, however, we shall focus on the loads we have learned to use.

At root, our exercise is simply an example or case study. In one sense, it is trivial from a modeling perspective. However, for those who grab a pattern and the source impedance information and then shut down the modeling program, it may serve the development of a habit of learning more about an antenna under study. It may also aid in the development of examining some easily overlooked information that modeling programs can provide if we systematically dig for the data.

The Task and Its Start

We want to construct a shortened, loaded antenna element. What is the best loading scheme? Where is the best place to position the load? **Fig. 7-1** illustrates part of our quandary.



Inductive Load Position

To make the task more specific, let's use 7 MHz as our target frequency region. The element will be aluminum. For study purposes, we can set a diameter of 1" as being about midway between beefy construction and wire construction. The exact dimensions may differ if we choose a different material, but the trends will all hold true.

The best way to begin the task is by modeling a $1/4$ wavelength vertical antenna above and in contact with perfect ground. Use about 25 segments, feeding the antenna in the segment closest to ground (in NEC) or at the ground pulse (in MININEC). Adjust the length to resonance. With NEC (from which all of the following numbers were derived), you should obtain from an elevation plot a gain of about 5.14 dBi with a source impedance of about 35.97 Ohms for a 33.7' vertical. Exact numbers will vary from program to program, but they should be very close to these. We shall assume that all cases of resonating an antenna bring the source impedance to less than ± 1 Ohm of reactance.

Even if we plan to use a dipole element, modeling over perfect ground is useful and reliable. Although a real vertical over imperfect ground will show a different gain, the trends derived from our perfect ground model will be accurate. Moreover, a vertical over perfect ground also provides data for a dipole in free space. Simply subtract 3 dB from the gain and double the source impedance. A test 67.4' free space dipole with 51 segments (to place the source at exact center in NEC) yields a gain of 2.13 dBi and an impedance of 71.87 Ohms. From the free space dipole, we can then adjust results for dipole elements over real ground.

So, we shall save half the time it takes to create and modify both the antenna geometry and the load placement simply by modeling the situation over perfect ground with a ground-mounted vertical.

What Happens When We Shorten a Vertical?

Understanding the practical dimensions of loading a shortened antenna element begins by understanding what happens when we just shorten the element. Suppose we take our $1/4$ wavelength vertical and begin shortening it in 10% increments, adjusting the number of segments to prevent them from getting too short.

The graph in **Fig. 7-2** shows what happens for our 7 MHz vertical—and for any other $1/4$ wavelength vertical, as well as for any free space dipole. The gain begins to drop from its peak of 5.14 dBi down to 4.69 dBi at the 10% of original length mark (3.4'). Equally notable is the drop in feedpoint impedance from nearly 36 Ohms down to a fraction of an Ohm. As the resistance at the source drops, the capacitive reactance rises from zero for the full length antenna to over 1400 Ohms for the antenna that is only 10% of that length.

Length vs. Gain and Source Resistance 7 MHz 1/4 WL Vertical

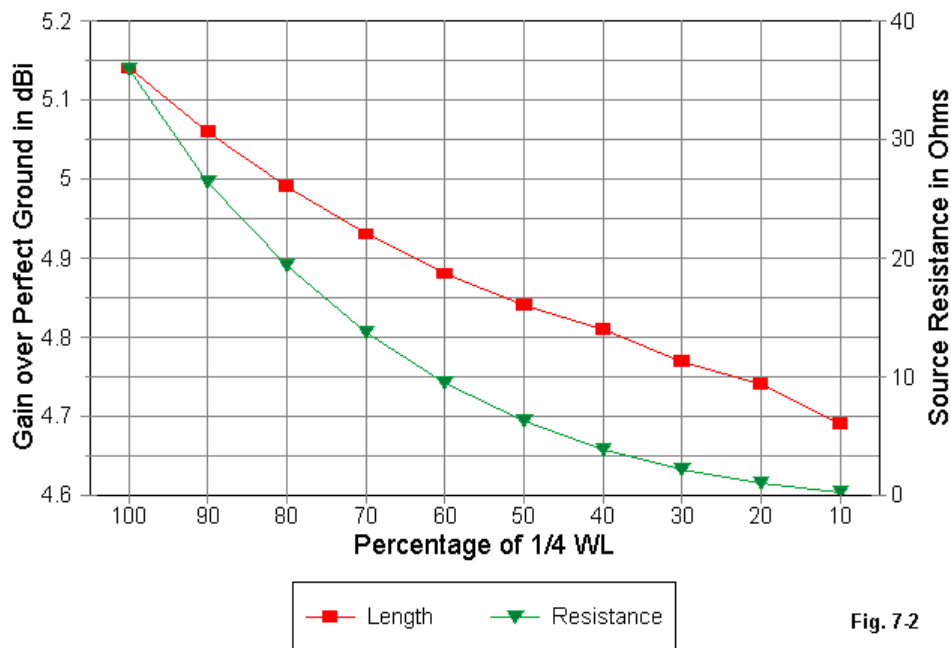


Fig. 7-2

At this point, we must make some study decisions. Ideally, we should explore loading with every length of antenna. Since we do not have room for that ideal in this column, I shall choose a particular antenna length to explore: about 70% of full length. In fact the antenna is 23.4' long—of course, using the same 1" diameter aluminum we started with. With no load, this antenna shows a gain of 4.93 dBi, only about 0.2 dB down from the full size antenna, but with a source impedance of 13.49-j176.3 Ohms.

The capacitive reactance tells us what compensating value of inductive reactance a base loading coil must have to resonate the antenna: 176.3 Ohms (just about 4.008 microHenrys at 7 MHz). However, if we want to look at options for loading the element at other than the base (or, with double the reactance and inductance values, at the center of a dipole), we must look systematically at what values of reactance will do the job. For this task, we can specify a type 4 (complex impedance) load consisting of a resistance and series reactance, and initially set the resistance to zero. (We shall change that zero later.)

Load Position vs. Load and Source R 23.4' 7 MHz Vertical

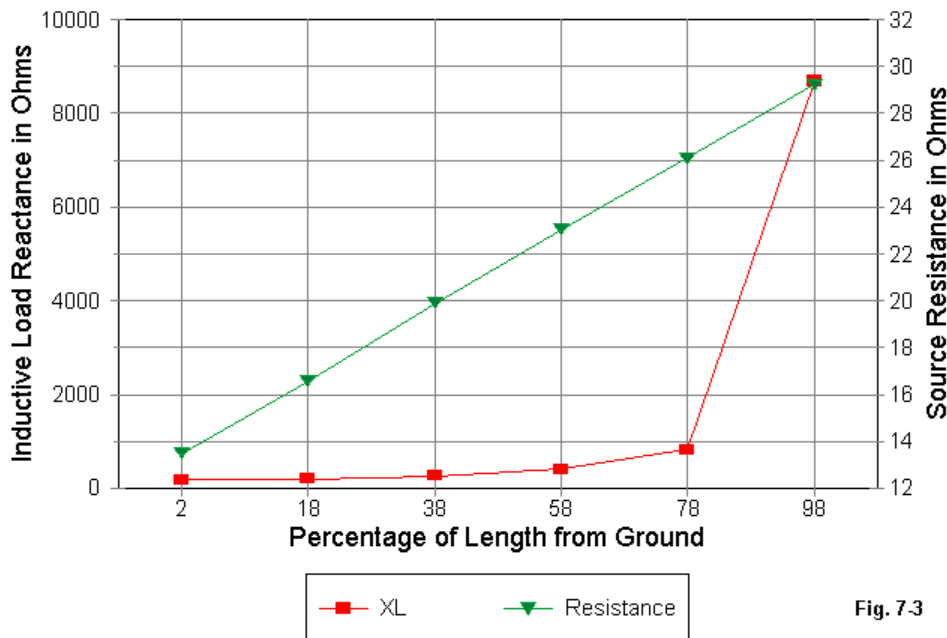


Fig. 7.3

So let's move the load from the antenna base (segment 1 in NEC, or the 2% mark for a 25-segment antenna). We shall for space economy move the load successively to segments 5, 10, 15, 20, and 25 (18, 38, 58, 78, and 98% up the an-

tenna). Then we shall adjust the amount of reactance to bring the antenna to resonance.

The graph in **Fig. 7-3** shows part of the results of our work. The source resistance climbs steadily from its low of 13.49 Ohms up to 29.21 Ohms with the load very near the tip of the antenna element. However, note the curve for the required value of inductive reactance. Near the element tip, the required reactance is 8680 Ohms, or a seeming inductance of 0.197 milliHenrys. The real antenna would be resonant just from added wire length in the solenoid long before that high an inductance was reached.

There is a reason for this situation. NEC loads are pure (mathematical) reactances and not physical coils. In the classic relationships between reactance and inductance, a coil with equal input and output current is assumed. However, on an antenna, the coil replaces a section of linear antenna element. On that “missing” piece of antenna, the standing current magnitude changes so that the input and output current levels are not equal. The differential is an indicator of how much a real coil radiates and to the same degree acts as a length of wire rather than a solenoid.

The further out the antenna that you move the load, the greater the current level change per unit length. A center-loading coil shows far less than 1% change and hence acts strictly as an inductance. At the tip, the coil acts almost exclusively as a length of wire and very little as an inductance.

In addition to the declining effectiveness of coils as inductive reactances as you move away from the source, the required reactance for resonance increases. At the 58% point, the necessary reactance is over double that required at the source. Expressed in other terms, for a dipole, each mid-element loading coil must be as big (in reactance and inductance) as a single center loading coil. For our antenna, a loading coil mounted exactly half way up the element must have 330 Ohms reactance compared to a base-loading coil of 176.3 Ohms.

However, as the load moves outward along the element, two advantageous things happen. The source resistance rises—from 13.49 Ohms for a base-loading coil to 21.79 Ohms for a mid-element loading coil. This rise in source resistance has implications for the ease of matching the antenna to common feedlines. (Remem-

ber to double these values if you are thinking of a dipole.) Second, the antenna gain rises slowly as the load moves outward, from 4.93 dBi with a base load to 5.02 dBi with our impractical load near the tip. With the load exactly mid-element, the gain is 4.95 dBi, which is not a significant difference from the gain of the antenna with a base load.

Q

Once you obtain systematic data about the required reactance to achieve resonance with the load positions you are considering, it is time to factor Q into the effort. Every inductor has a series resistance, and the reactance divided by the resistance is the component Q. However, the resistance is pure loss, and we need to see the magnitude of these losses.

Again, it would be useful to know the effect of Q on loads for every antenna length possible and at every load position along each of the antennas. However, for this example, let's stick with our 23.4' long vertical (corresponding to a 46.8' dipole) and explore the effects of Q upon just the base load and the mid-element load models.

The first question is this: what is a sensible range of Q to explore? It is certainly feasible to construct a loading coil with a sustainable Q of at least 100. ("Sustainable Q" means the Q after the coil has been immersed for months or years in the chemical soup that we call our atmosphere.) At the other extreme, the highest coil Q I have seen claimed for an antenna-loading coil is about 600. Whether or not such a high value is actually feasible or sustainable in use, I cannot say, but exploring the Q-range from 600 down to 100 should be instructive.

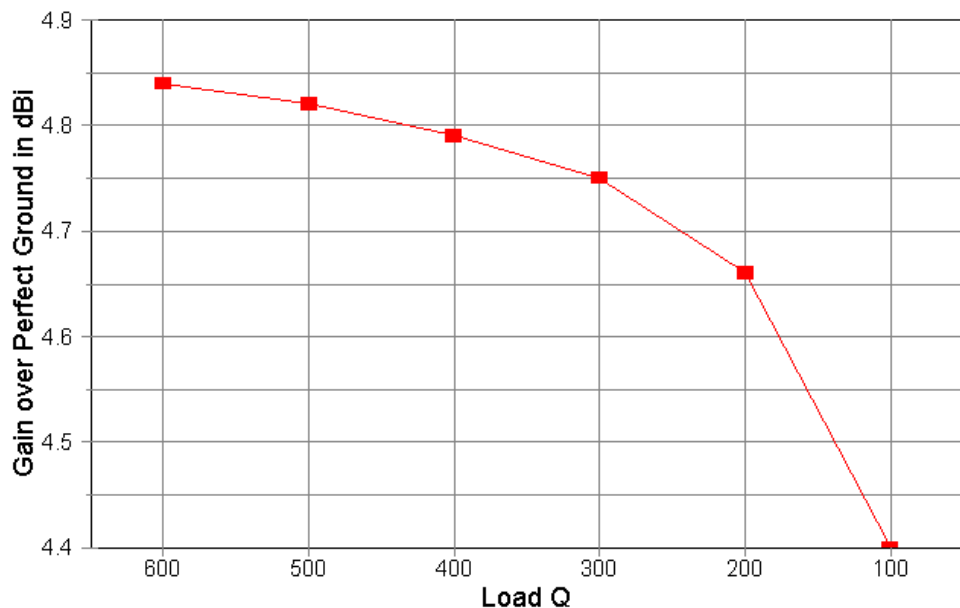
Therefore, in two sets of modeling runs, let's explore what happens to the antenna with base loading and with mid-element loading as we vary the Q. We can do this by simply using the Type 4 load we have been using for reactance insertion and adding a series resistance appropriate to the Q. Since the reactance value of the mid-element load is over double that of the base load, you will have to expect resistance values to parallel that ratio for equivalent Q values.

The graph in **Fig. 7-4** shows the effects of decreasing Q on the antenna gain. From a value of 600 down to a value of 300, the loss is slight. However, from that Q

value downward, the curve gets increasingly steep. As the Q of the base load decreases, the resistance of the coil shows up as an increase in source resistance. The increase is in the range of 0.3 Ohms at $Q=600$ to 1.75 Ohms for $Q=100$.

Base Load Q vs. Gain 23.4' 7 MHz Vertical

Fig. 7.4

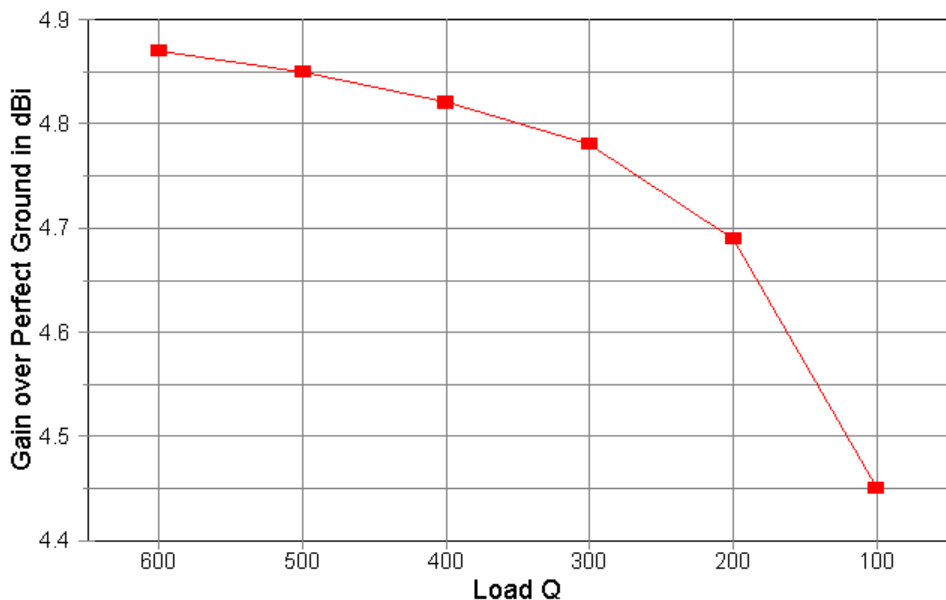


The results for mid-element loading appear in **Fig. 7-5**. The curve is virtually identical to that for base loading. Once more, a Q of 300 seems to be the dividing point between modest losses and more extreme losses with mid-element loading coils. Because the coil is further out, the source resistance changes only by about

the same amount as it does with a base-loading coil, although the series resistances are more than double. The source resistance with a Q of 100 is about 2.5 Ohms higher than it is with an infinite Q.

Mid-Element Load Q vs. Gain 23.4' 7 MHz Vertical

Fig. 7.5



Throughout the range of Q values for both the base loading and the mid-element loading models, the difference in gain between the models is never greater than about 0.3 dB. This very slight difference belies some common recommendations often heard that give mid-element loading grandiosely higher marks than base

loading. We have a case here of remarks being taken out of their original context and becoming unwarrantedly general. Nothing in the physics of a loaded antenna warrants them.

The recommendation for mid-element loading arose from very short mobile antenna practice. In this situation, the antenna must contend with significant amounts of auto body metal and other surrounding objects to absorb or reflect in uncontrolled ways the radiation from the antenna. If the advice has any application, it is not to the general theory of loaded antennas, but to antennas located where external conditions affect their performance. For any given Q , there is no significant difference in the anticipated performance of base loaded or mid-element loaded antennas that are free and clear of such clutter.

Operating Bandwidth

The investigation would not be complete without exploring the operating bandwidth of loaded antennas. Once more, we shall restrict ourselves to just our two examples. We shall specify a Q of 300 for both models, even though a complete study would want to explore operating bandwidth for every increment of Q checked in the preceding section of the study. To check the operating bandwidth of the antennas, we must change the load- type to 0 for a series resistance, inductance, and capacitance (the last of which we set to zero to show that it is missing). The resistance-inductance combination for the base-loaded antenna is .5877 Ohms and 4.008E-6 Henrys. For the mid-element loaded model, the combination is 1.1 Ohms and 7.503E-6 Henrys. This change in load type ensures that the antennas will show the correct complex impedance at every frequency.

We may make a frequency check every 100 kHz for two increments above and below the resonant frequency to ensure that we cover the width of the 40-meter ham band. We can show the results of the sweep in both tabular and graphical forms.

Base-loaded model: $R = 0.5877 \text{ Ohms}$; $L = 4.008\text{E}6 \text{ H}$

Frequency relative to resonance	Gain dBi	Source Impedance $R \pm jX \text{ Ohms}$	VSWR relative to 14.08 Ohms
-0.2 MHz	4.72	$13.16 - j 18.83$	3.645
-0.1	4.73	$13.61 - j 9.36$	1.944
0	4.75	$14.08 - j 0.01$	1.001
+0.1	4.76	$14.56 + j 9.22$	1.886
+0.2	4.77	$15.05 + j 18.37$	3.293

Mid-element-loaded model: $R = 1.10 \text{ Ohms}$; $L = 7.503\text{E}6 \text{ H}$

Frequency relative to resonance	Gain dBi	Source Impedance $R \pm jX \text{ Ohms}$	VSWR relative to 22.68 Ohms
-0.2 MHz	4.76	$20.30 - j 28.17$	3.446
-0.1	4.77	$21.45 - j 14.43$	1.907
0	4.78	$22.68 - j 0.42$	1.019
+0.1	4.79	$24.00 + j 13.88$	1.802
+0.2	4.80	$25.43 + j 28.55$	3.104

Fig. 7-6 graphs the SWR curves relative to the resonant impedance of each antenna. Although the SWR operating bandwidth of the mid-element loaded model is slightly wider than that of the base-loaded model, the difference is unlikely to make a difference in operation. Both antennas have 2:1 operating bandwidths in the vicinity of 200 kHz, or less than the span of the 40-meter band.

The operating bandwidths, of course, are those at the antenna source terminals. The application of matching systems and losses in the feedline may make these bandwidths appear wider by virtue of taking SWR readings at the transmitter location. For our purposes, it is an open question whether that widening of the operating bandwidth comes with a low or high cost in losses.

Base & Mid-el. Load SWR Curves

Load Q = 300

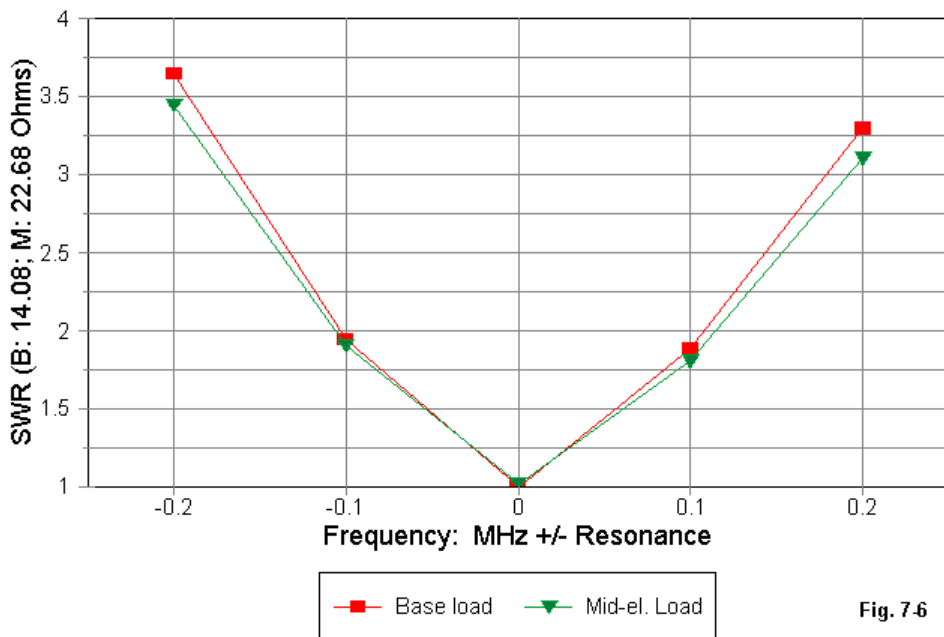


Fig. 7.6

The tabular data show, as a graph cannot, the changes in both the resistance and the reactance at the feedpoint of the respective antennas. The ratio in the source resistances of the two antennas is reflected also in the total change in reactance across the checked bandwidth. Hence, the SWR curves turn out to be very similar.

Interestingly, the key significant difference in the two loading systems appears to be the source resistance. For most purposes, the higher source resistance of mid-element loading would yield higher efficiency with the same set of external losses due to such things as connection resistance and the like. Other variables, such as construction advantages of one system over the other, fall outside what modeling can indicate. However, we have not modeled our last load with our 23.4' vertical over perfect ground.

“Capacity” Hat Loading

As we observed when moving loads outward along the antenna element, by the time we pass the mid-element point, loads act more and more like lengths of wire and less like pure inductances. If we place a coil on the very end of the element, it acts like a simple extension of the antenna element, but with tight coupling and consequent cancellation of some radiation. A more effective end-loading scheme is the so-called “capacity” hat.

A capacity hat is a wholly symmetrical structure added to an element end at right angles to the main element. The symmetry ensures cancellation of any radiation from the structure. The name “capacity” hat arose from a method of calculating the size of the hat, a method which is useful at low and medium frequencies, but which grows increasingly inaccurate throughout the HF range. Physically, the hat is simply an extension of the antenna element that does not produce significant radiation.

Fig. 7-7 shows the general scheme, along with several hat types. The square, hexagon, and octagon shown all have a perimeter wire, which reduces the required size of the support spokes. However, hats can be composed solely of spokes (a few or many), can have intermediate as well as perimeter connecting wires, or can even be disks.

There are some rules for creating hats. The thinner the hat-wire, the longer the spokes will grow, for any given geometry. The fatter the main element, the longer the spokes.

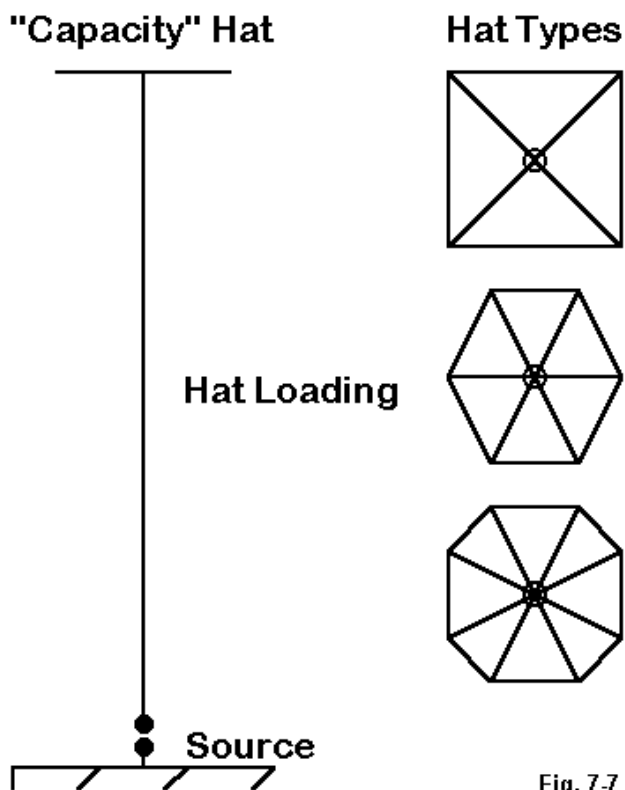


Fig. 7-7

Any hat that passes the symmetry test will work as well as any other such hat. For example, here are gain and source impedance figures at resonance for 3 hatted 23.4' verticals: a square, a hexagon, and an octagon, all with perimeter wires. As you increase the number of spokes further, the rate of spoke length decrease goes down. Spoke length indicates the distance from the main element outward.

	Square:	Hexagon:	Octagon:
	3.11' spokes	2.78' spokes	2.6' spokes
Gain	5.03 dBi	5.03 dBi	5.03 dBi
Source Z	29.5 - j 0.6	29.7 + j 0.9	29.7 + j 0.9

There are two very notable sets of values in this small table. First, the gain of the antenna is higher than with either system of inductive loading at a Q of 300. The gain is about mid-way between the inductively loaded models and the full-length antenna. Second, source impedances at resonance are significantly higher than even those of the mid-element loaded model.

XL vs. Hat Load SWR Curves

XL Load Q = 300

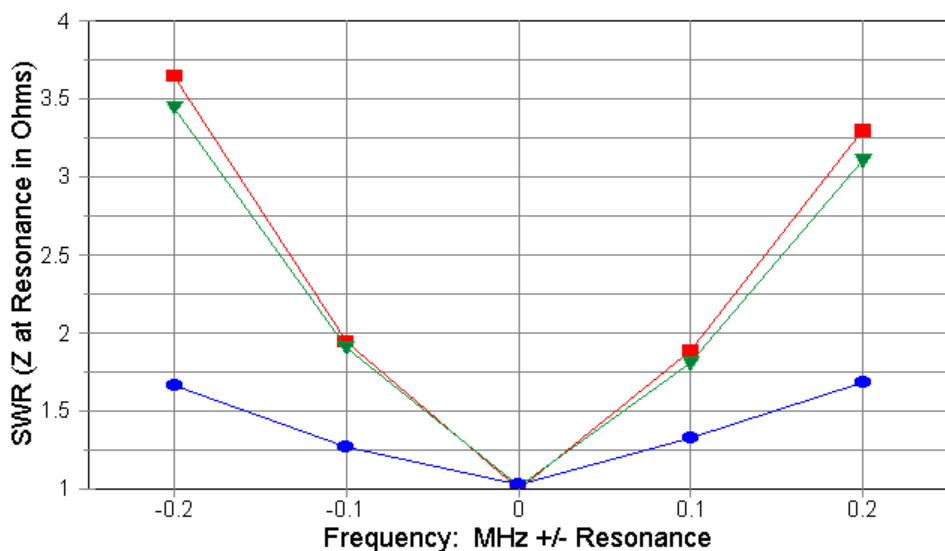


Fig. 7.8

—■— Base load —▼— Mid-el. Load —●— Capacity-hat

The SWR graph in **Fig. 7-8**, based on the resonant impedances of each antenna type, compares the curves of the inductively loaded models with the curve of any one of the hat-loaded models. The 2:1 SWR points are more than 500 kHz apart, well beyond what is needed to cover 40 meters.

Capacity"-hat loaded model: hexagon

Frequency relative to resonance	Gain dBi	Source Impedance R+/- jX Ohms	VSWR relative to 22.68 Ohms
-0.2 MHz	5.01	27.26 - j 14.42	1.663
-0.1	5.02	28.44 - j 6.73	1.265
0	5.03	29.66 + j 0.91	1.031
+0.1	5.04	30.94 + j 8.58	1.330
+0.2	5.05	32.26 + j 16.22	1.690

The chart reveals that the rate of change of reactance across the frequency sweep is the smallest of all three types of loading we have investigated. The tables for the other two types of hats are too close to this one in all values to merit display.

The upshot is this: if mechanical problems do not override performance figures, hat loading deserves serious consideration in the development of shortened antennas.

Summary 1: the Example

Although we have been working with a somewhat contrived example in order to illustrate some techniques of systematically modeling an antenna and its variations, the information that we derived may itself prove useful down the road. Of course, we have explored only a few options for the construction of a shortened vertical monopole, and only for a single frequency region. Within those limitations, we can summarize our work on the shortened 40-meter vertical in a small table.

Loading scheme	Gain dBi	Source Resistance	2:1 VSWR Bandwidth
Base: Q=300	4.75	14.08	<250 kHz
Mid-el.: Q=300	4.78	22.68	<250 kHz
“Capacity” hat	5.03	29.67	>500 kHz
Full-size	5.14	35.97	>500 kHz

A more complete study would yield much more data. However, this much information is enough to reveal some of the design decisions facing us. Gain, bandwidth, and source impedance are the components of the final compromise, after adding in all of the external factors to which modeling cannot contribute. Among these factors are the mechanical potentials and constraints of the antenna alone and within its operating environment.

Summary 2: the Method

More important than the particular example that we probed is the method used to probe it. Beginning at a fundamental point in the study—the shortening of antennas before loading them back to resonance—we used the facilities of the program (either NEC or MININEC) to generate a relatively rapid means of developing comparative data. The data trends will carry over to real antennas—both vertical and horizontal—placed over real ground, although the exact gain values that we received are based upon a perfect ground. The use of free space or of perfect ground is not restricted to preliminary antenna design work. It is also an important tool in systematic comparative studies designed to increase our knowledge and understanding of antenna behavior under a wide variety of conditions.

For an individual who needs the information, one could easily expand the study in several directions. One may vary the lengths of the shortened monopole. As well, one may vary the position of the mid-element loading inductor. Next, one might perform studies over a wide range of values of Q for each of these variations. In the arena of “capacity” hats, one may well compare versions with only spokes with versions using a combination of spokes plus a perimeter wire. Both MININEC and NEC shorten the time it takes to gather comparative data. However, nothing in this type of study precludes experimentation to confirm the modeling results.

A little aimless modeling every now and then can be entertaining and occasionally useful. However, the more systematic you make your modeling efforts, the more profitable that the output data and the program become.

* * * * *

Models included: 7-1 through 7-4.



8. Modeling Wire Arrays

Wire arrays present the modeler with some challenges and some opportunities to overlook obvious errors in modeling. They also tend to make us use almost all of the resources available within antenna modeling programs. So let's look at this class of antennas and explore a few useful techniques and cautions.

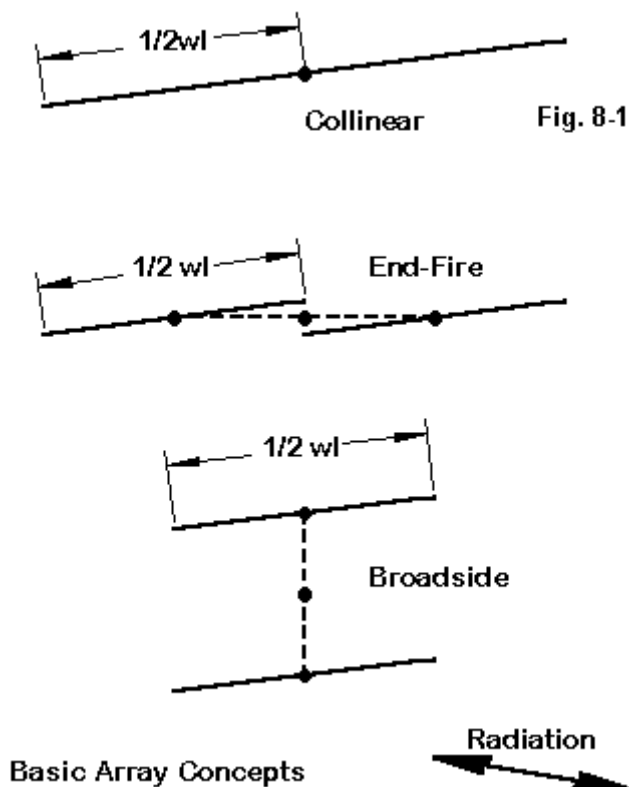
Basic Concepts

Bi-directional wire arrays comprise a class of antennas composed of multiple $1/2$ -wavelength antenna elements arranged to produce antenna patterns that serve specific purposes. If we adopt the $1/2$ -wavelength resonant length of wire as our basic unit, then we can make sense out of the operative vocabulary of bi-directional wire arrays. The key questions are two. a. How are the elements arranged relative to each other? b. With a given arrangement, what is the main axis of radiation?

Fig. 8-1 shows three basic arrangements that yield the fundamental categories within which these arrays are specified. Elements are *collinear* when they are arranged end-to-end so that the main radiation lobes are tangential to the wire. The ends may touch, as in the illustration, or they may be separated by a phasing section, a length of wire or transmission line that alters the current phase and magnitude at the point of junction. Ordinarily, these lines are used to increase the strength of the main radiation lobes.

When elements are parallel to each other, in a plane that is parallel to the ground surface, and the strongest radiation lobes are tangential to the wires, we have an *End-Fire* array. Most commonly for bi-directional arrays, the two elements are fed 180 degrees out of phase with each other.

A *Broadside* array places elements (normally) in a vertical plane, with the main radiation lobe again tangential to the wires. In this configuration, we feed the wire elements in phase with each other.



Most bi-directional wire arrays use combinations of these techniques. For most applications, a 1-wavelength wire is the shortest collinear element having sufficient gain to use in either a broadside or end-fire array. Longer elements, such as the 1.25 wavelength extended double Zepp (EDZ), are commonplace. Moreover, we may combine both end-fire and broadside techniques into single arrays.

For many of the bi-directional array designs, absolute precision of wire length is less important than with other types of antennas. A 5% variation in the length of a 1/2-wavelength element may make little difference to the performance of the array. Indeed, as arrays grow larger, modeling does not so much determine an exact construction version as it looks at trends in both performance and source impedance to guide field adjustment of the physical antenna. In addition, we should not expect source impedances that are ready matches to a 50-Ohm system feedline. Instead, we shall find a wide variety of source impedances that challenge us to develop feed networks.

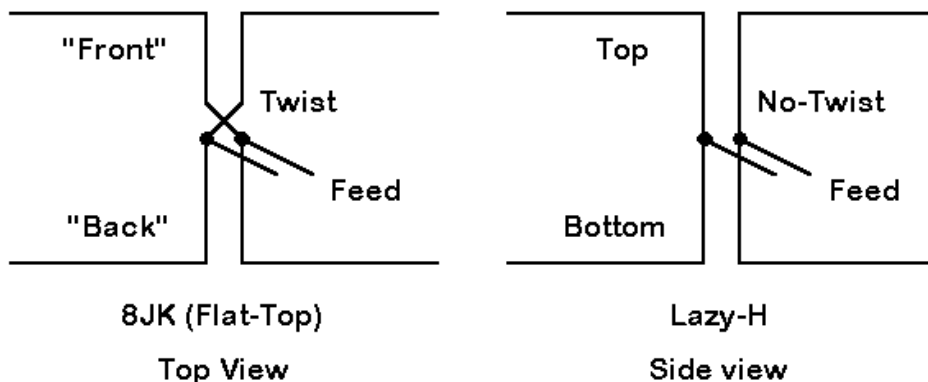
Although bi-directional arrays date from the earliest days of radio communications, they reached a zenith of design effort in the 1930s as short wave broadcasting became commonplace. Great “antenna farms” involving acres of land emerged as stations built multiple fixed-position antennas beaming power with narrow beamwidths at selected reception areas. Although new ideas in bi-directional arrays surface more slowly today, designing an antenna of this type to serve a specific communications need is still a significant engineering enterprise.

In some applications, especially in the upper HF range, the Yagi and other rotatable parasitic antennas have supplanted the bi-directional array. However, up to about 10 MHz, the bi-directional array is still the antenna of choice for most fixed stations. Moreover, even above 10 MHz, these antennas offer a cost-benefit ratio that makes them attractive alternatives to other techniques.

Feeding the End-Fire and Broad Side Arrays

One of the most common errors in modeling common arrays is forgetting which array requires out-of-phase feeding and which requires in-phase feeding. This usually occurs with NEC-2 programs, where we set and forget the transmission lines we use to create the feeding system.

Fig. 2 shows the top view of a flat-top or 8JK and a broadside view of a Lazy-H array.



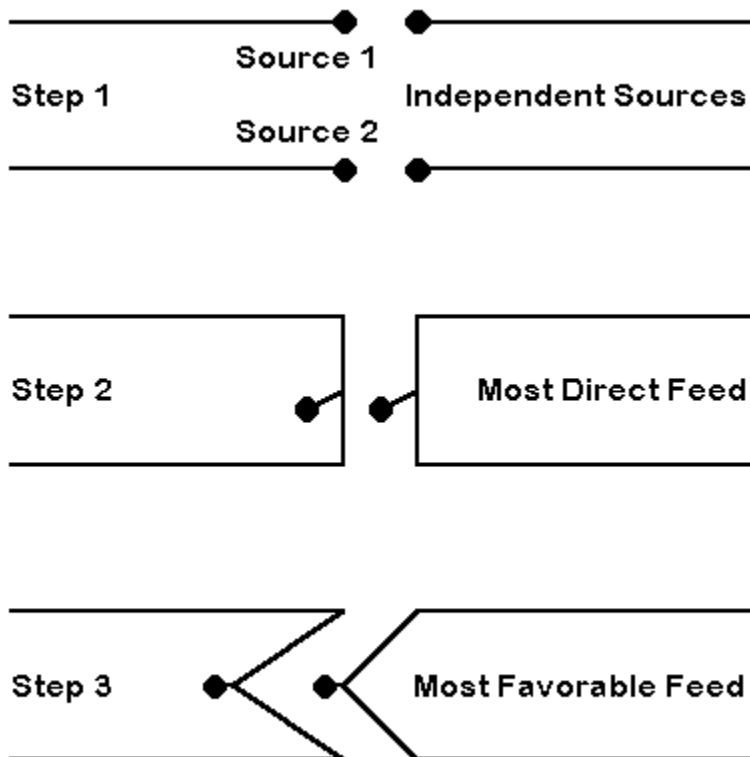
Feeding End-Fire and Broadside Arrays

Fig. 8-2

Virtually all flat-top arrays of 2 elements are fed 180-degrees out of phase. The simplest means of achieving this goal is to create two transmission lines of equal length to an added central source wire (junction). One line is reverse connected to its element. Since the current magnitude and phase shift along each line is identical, the reverse connection on one of the lines ensures the out-of-phase condition.

Virtually all vertically stacked 2-wire arrays are fed in-phase. Once more, the simplest feed system is to use equal lengths of transmission line to an added common junction wire, but normal connections are used for both lines. Ordinarily, the slight differences in the source impedance of the two wires, given their different distances from the earth's surface, are not sufficient to disturb the relative phasing. However, in very sensitive systems, networks may be used to achieve critical phasing adjustments.

For all arrays, it is best to develop models in step-wise fashion, rather than throwing together all the elements and then wondering what might be wrong with the overall result. **Fig. 8-3** shows the steps.

**3 Steps to Feeding Phased Elements****Fig. 8-3**

1. First, create the wires and use a separate source for each. With the simple expedient of alternately feeding one wire 180 degrees out of phase with the other wire and then in phase, you will see the pattern results appropriate to your design.

You will also have a record of the wire source impedances, should you decide to do some independent calculations for a feed system.

2. Next, try the most direct feed system possible. This step involves creating a very short, thin wire in the model physically located exactly between the two antenna wires and along the centerline of those elements. If you are using NEC-2, you can create two transmission lines, one from each antenna wire to the common or junction wire. Place the source on this third wire (remembering to remove the two formerly independent sources). Use the connections appropriate to the antenna type.

Record the composite source impedance. The key question at this point is whether that impedance is acceptable in terms of the system feed from that point back to the transmitting/receiving equipment. Within the scope of this step, you have the option of changing the characteristic impedance of the line, most usually within the 300/450/600 Ohm set of commercially available transmission lines. However, you can always construct your own.

Be certain to allow for the velocity factor (VF) of the line. Some programs permit direct entry of this value; others require that you pre-calculate the longer line length with a VF of 1.0 equivalent to a direct connection with a line whose VF is less than 1.0.

In some cases, you will be “stuck” with the composite source impedance that emerges. Often, the direct line feed is the only physically feasible system of routing the feedlines. In some other cases, you may have the option of running longer lines to a common junction point.

3. If you can manage longer lines, you may experiment with them to see if they yield a more desirable composite source impedance than direct feed.

The following table illustrates the process, using a 1/2-wavelength spaced flat-top with 180-degree out-of-phase feeding. The frequency is 14.175 MHz, resulting in each phase line having a minimum length of about 17.5 feet each side of the feedpoint (35' overall). (Model 8-1)

Line length (each leg)	Composite source impedance	
	600-Ohm line	450-Ohm line
17.5'	44.4 - j 4.4	25.0 - j 1.7
20.0'	46.5 + j 63	26.2 + j 50
22.5'	54.1 + j137	30.7 + j106
25.0'	71.3 + j228	40.7 + j176
30.0'	216 + j556	129 + j444

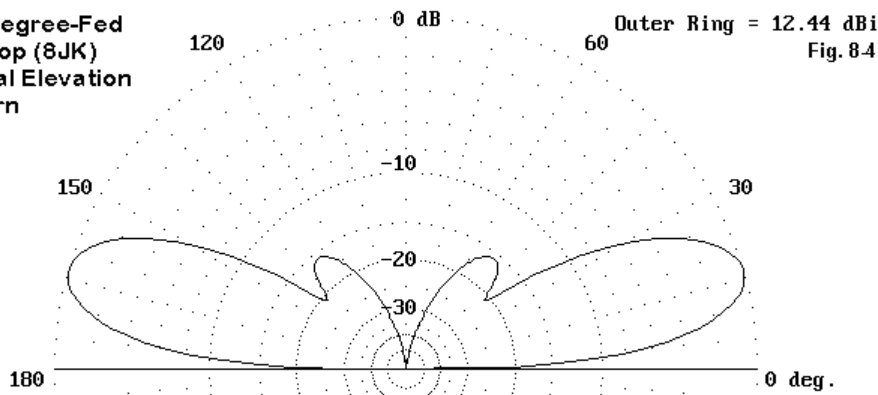
For a coax match, the most direct feed with 600-Ohm coax seems best. However, other feed systems may direct the use of other lengths. While you are at it, do a frequency sweep of the antenna to check the rate of change of both the resistive and reactive components of the composite source impedance. You may also change the wire lengths to place the resonant point of the system at a different resistive value of impedance.

Patterns and Frequency Sweeps

While you are exploring feed systems, also explore the elevation patterns. Many broadside systems have one or more lobes at very high angles, making the system more susceptible to atmospheric noise inputs. Flat-top systems, although generally of lower gain, naturally null the most vertical lobes, reducing sensitivity to high angle radiation. **Fig. 8-4** provides a sample pattern for an end-fire system fed 180 degrees out-of-phase.

On the other hand, in-phase fed broadside systems, such as the Lazy-H, often permit multi-band service with usable gain and patterns extending from 0.5 F to 1.25 F, where F is the frequency for which each element is 2 half-wavelengths. Consider the Lazy-H cut for about 21 MHz, with 44' wires spaced vertically by 22' (about 1/2-wavelength). If the lower wire is at 35' up, the upper wire will be 57' high. (This system might also be called an EDZ Lazy-H cut for 28 MHz and spaced about 5/8-wavelength at that frequency.) The following table records the modeled performance on various amateur frequencies. (Model 8-2)

180-Degree-Fed
Flat-Top (8JK)
Typical Elevation
Pattern



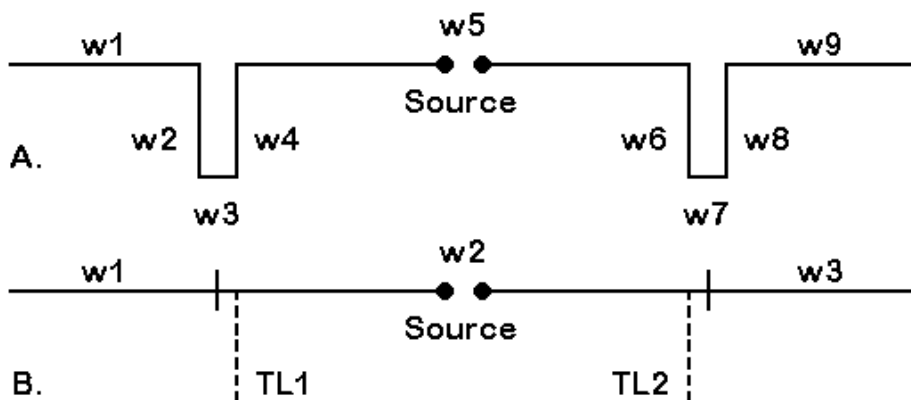
Frequency	T-O Angle	Gain dBi	Source Impedance
10.1	28	7.38	$50 + j\ 91$
14.0	20	9.04	$572 - j\ 394$
18.1	16	10.56	$49 - j\ 142$
21.0	13	12.15	$24 - j\ 36$
24.9	11	14.31	$17 + j\ 94$
28.0	10	14.78	$39 + j\ 291$

If the source impedance is not an issue, this antenna might be useful as a multi-band antenna serving 30 through 10 meters with minimally dipole performance at the lowest frequency. You may verify that all patterns are broadside to the antenna, in contrast to the many-lobed patterns that would appear on 10 meters had we started with a 30-meter dipole. You may also wish to experiment with other line lengths from the center source junction to the antenna elements to see if you can arrive at a length more favorable to your own feed system ideas.

Modeling Collinear Elements with Phase Lines

Collinear antenna elements that employ phasing stubs often tempt us to automatically use NEC-2 transmission line capabilities in our initial models. While this technique sometimes works, it very often leads us into endless trials (and errors)

searching for the correct length to make the antenna model at its performance of record.



Two Methods of Modeling Collinear Elements

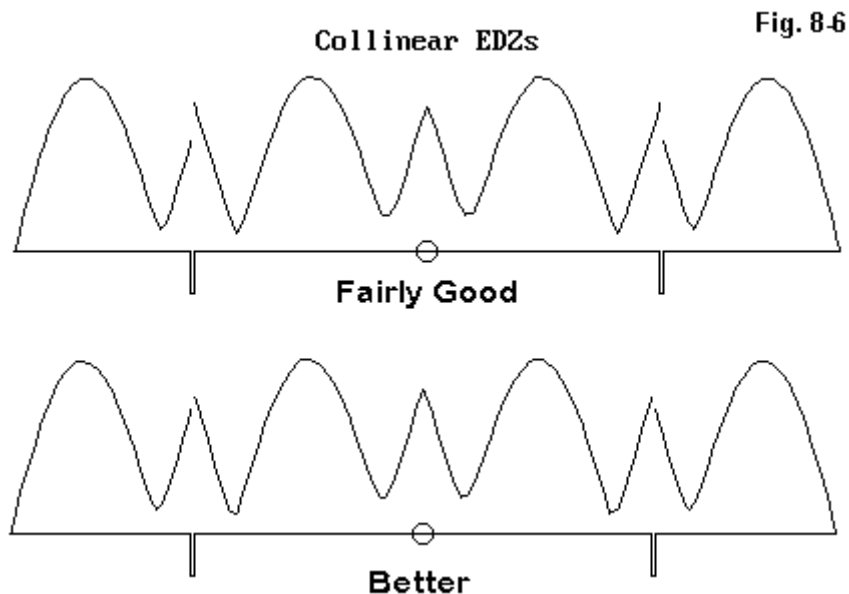
Fig. 8-5

With phasing lines and NEC-2 we always have 2 options, as shown in **Fig. 8-5**, a representation of collinear EDZ antenna elements. (Of course, only the physical modeling option is open to those who use MININEC. However, it is the initial option of choice for all modeling.)

Only some types of arrays are amenable to the substitution of non-radiating transmission-line models for their stub elements. A few depend upon the incomplete cancellation of currents to yield their final performance figures. Therefore, it is usually wise to begin with a physical model of a collinear element and then later to convert it to a transmission line model. (Models 8-3 and 8-4)

Physical modeling is straight forward, as the figure indicates. All wires should run continuously, from end 1 of wire 1 to end 2 of wire 9. Use a segmentation density high enough that the horizontal shorting lines of the stubs are not radically different in length from the wire segments immediately adjacent to them. In devel-

oping the model, make use of either the current tables or a graphical aid to show relative current magnitudes. See **Fig. 8-6**.



The aim of the collinear EDZ design is to place equal magnitude currents at each of the phasing lines. Unequal currents result in radiation from the phasing lines, which can disrupt overall performance. The current magnitude graphics represent two stages on the way to perfecting a model of the antenna in question.

This design, which dates back to the 1930s work of Hugo Romander, W2NB, is actually fairly forgiving, with 2-3% errors in length yielding only a drop of a few tenths of a dB in gain. The beamwidth of this antenna is only about 17 degrees between -3 dB points, so it is best suited to point-to-point communications rather than casual amateur work.

Wide-spaced physical stubs do not change their length rapidly as they are made narrower, since the characteristic impedance of these shorted stubs changes slowly as the spacing is increased beyond a couple of inches. Therefore, they provide a starting point for replacement with transmission line sections. The easiest procedure in many cases is to use 600-Ohm line for initial replacement trials. Once a working length value is found, you may convert it to an inductive reactance and from there to any transmission line value with a length appropriate to its characteristic impedance and velocity factor. Place the transmission line stubs on the first and last segment of the center wire (relative to the model in **Fig. 8-6**), and use a segmentation density high enough to closely approximate the center wire ends.

More Complex Arrays, Such as the Sterba Curtain

Arrays are not always as simple as the models with which we have been working. A few seem to defy adequate modeling, especially if we insist upon using transmission line facilities in our programs. The Sterba curtain is one of these antennas. Yet, it models physically with remarkable ease—if you know some tricks of the trade.

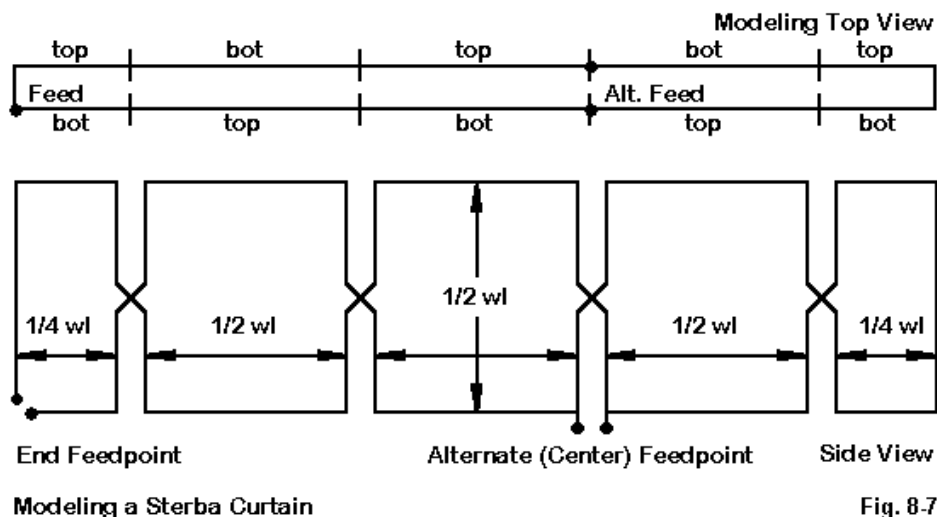


Fig. 8-7

Fig. 8-7 shows a side view of a Sterba curtain, similar to the typical sketch shown in many an antenna manual. The twisted vertical line pairs seem to beg for transmission line treatment. If we contemplate using physical lines, we often bog down in the thought of trying to keep them at a constant spacing and still effecting the “half-twist.”

The top view shows a simple solution. Offset two wires by a small space. 0.5' will do at 40 meters without creating the slightest distortion in antenna performance for this 2.5-wavelength antenna. Run one of the wires from end to end, using vertical lengths to let it alternate between the top and bottom positions. Run the other wire in the same manner, but start at the far end and opposite corner. The result will be a set of wires correctly oriented to each other in both the horizontal and vertical planes. Connect the near and far ends with “near-vertical” wires (offset from the vertical only by the space between the horizontals). Make the connections to keep the wire continuous from its start to finish.

Here is a typical table of wires, with the wire offset in the Y-axis. (Model 8-5)

Sterba curtain: Frequency = 7.2 MHz.

Wire	Conn.—	End 1	(x,y,z : ft)	Conn.—	End 2	(x,y,z : ft)	Dia(in)	Segs
1	W20E2	0.000,	0.500, 67.000	W2E1	0.000,	0.000,134.000	# 12	15
2	W1E2	0.000,	0.000,134.000	W3E1	33.500,	0.000,134.000	# 12	8
3	W2E2	33.500,	0.000,134.000	W4E1	33.500,	0.000, 67.000	# 12	15
4	W3E2	33.500,	0.000, 67.000	W5E1	100.500,	0.000, 67.000	# 12	15
5	W21E1	100.500,	0.000, 67.000	W6E1	100.500,	0.000,134.000	# 12	15
6	W5E2	100.500,	0.000,134.000	W7E1	167.500,	0.000,134.000	# 12	15
7	W6E2	167.500,	0.000,134.000	W8E1	167.500,	0.000, 67.000	# 12	15
8	W7E2	167.500,	0.000, 67.000	W9E1	234.500,	0.000, 67.000	# 12	15
9	W8E2	234.500,	0.000, 67.000	W10E1	234.500,	0.000,134.000	# 12	15
10	W9E2	234.500,	0.000,134.000	W11E1	268.000,	0.000,134.000	# 12	8
11	W10E2	268.000,	0.000,134.000	W12E1	268.000,	0.000, 67.000	# 12	15
12	W11E2	268.000,	0.000, 67.000	W13E1	234.500,	0.500, 67.000	# 12	8
13	W12E2	234.500,	0.500, 67.000	W14E1	234.500,	0.500,134.000	# 12	15
14	W13E2	234.500,	0.500,134.000	W15E1	167.500,	0.500,134.000	# 12	15
15	W14E2	167.500,	0.500,134.000	W16E1	167.500,	0.500, 67.000	# 12	15
16	W15E2	167.500,	0.500, 67.000	W17E1	100.500,	0.500, 67.000	# 12	15
17	W21E2	100.500,	0.500, 67.000	W18E1	100.500,	0.500,134.000	# 12	15
18	W17E2	100.500,	0.500,134.000	W19E1	33.500,	0.500,134.000	# 12	15
19	W18E2	33.500,	0.500,134.000	W20E1	33.500,	0.500, 67.000	# 12	15
20	W19E2	33.500,	0.500, 67.000	W1E1	0.000,	0.500, 67.000	# 12	8
21	W4E2	100.500,	0.000, 67.000	W16E2	100.500,	0.500, 67.000	# 12	1

For this particular model, the source is placed in wire 21, which corresponds to the alternate source point. You may omit wire 21 and employ a split feed system, using the last segment on wire 20 and the first segment on wire 1 as the source positions. There are minor differences in current equality on the horizontal sections between the two sourcing systems, as you may see from the current tables or the graphical current magnitude facility in your program. These differences will alter the gain slightly and also the degree to which the 26-degree beamwidth main lobe is offset from being directly broadside to the wire array. However, the source impedance at the end position may be more convenient, being closer to 600 Ohms. You may undertake many modeling experiments, especially with the vertical spacing between the top and bottom wires of the system, to effect the most equal current distribution possible.

The Sterba curtain has become a rare antenna, being somewhat out of date, compared to some other wire arrays, such as the Bruce. The latter array requires less space for essentially the same performance. However, the Sterba curtain better illustrates the modeling techniques we have explored and so has earned its place in this column.

Conclusion

We might extend these notes indefinitely, since there are dozens more basic and variant types of wire arrays, each with its own set of modeling challenges and each with its seemingly esoteric name. However, those challenges will be mostly variations on those we have encountered in this set of exercises.

If these notes have a message, it is something like this: begin your modeling of a wire array on paper to help ensure that every wire will have its proper place. Model slowly and progressively toward the final complete antenna package (to the degree that your program permits modeling it). Double check all input values before running the model at each stage. Along the way, make use of all of the data provided by the program (and not just gain and source impedance results) to perfect the model.

The more we learn about using the vast array of data provided by antenna modeling programs, the better—and quicker—and more informative our models become.

* * * * *

Models included: 8-1 through 8-5.



9. Modeling Ground Planes

Vertical monopoles require a ground plane for the completion of the antenna, whether at ground level or elevated. Other types of antennas also call for ground radial systems. Modeling radial systems presents a numbers of challenges, especially if we are creating one for the first time. So let's see what is involved.

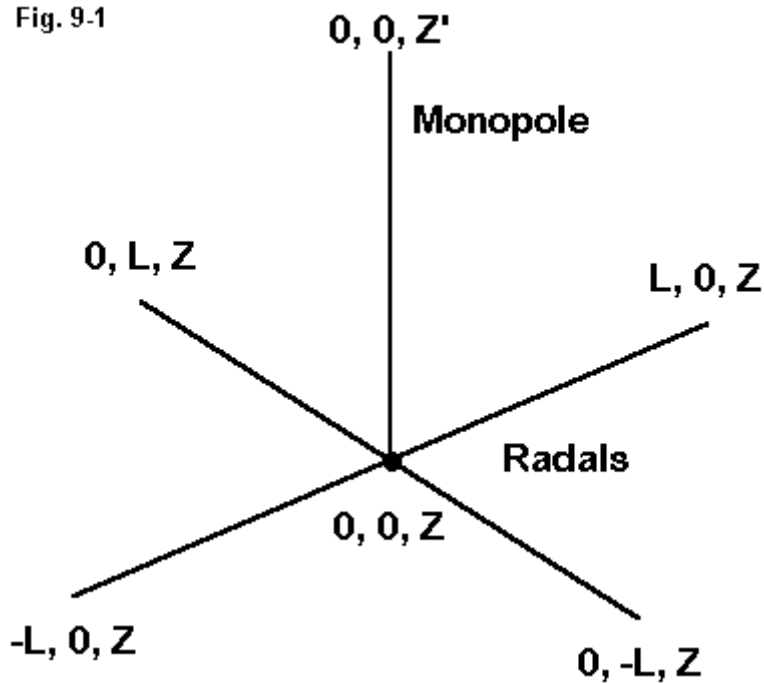
Of course, modeling a vertical monopole over perfect ground requires no ground plane, because the program (NEC or MININEC) creates the requisite image antenna necessary for completion of the overall structure. Hence, our focus will be upon modeling vertical monopoles over real ground.

The Radial System

The standard ground plane system consists of a set of wires arranged radially about a single point. In real systems, the number general ranges from 4 to 120. If the number is only 4, as in **Fig. 9-1** (and model 9-1), the system is simple. Let L be the length of the radial. Then take 4 wires, all with one end at a certain point, usually at the coordinate 0, 0, Z , where Z may be some number greater than zero, representing the height of the radial system above ground. For each of the 4 wires, keep the same value of Z . Then alternatively assign the length to positive and then negative values of alternating X and Y . Our wire chart looks like the following sample.

Wire #	End 1	X	Y	Z	End 2	X	Y	Z
1		0	0	Z		L	0	Z
2		0	0	Z		0	L	Z
3		0	0	Z		$-L$	0	Z
4		0	0	Z		0	$-L$	Z

Fig. 9-1



Beyond 4 radials, the required values for X and Y become a little more complex. Perhaps the greatest temptation in modeling vertical monopoles is always to use perfect ground in order to avoid having to create large models involving many radials. However, modeling a radial system is a straightforward task with simple rules that are amenable to symbolic model entry or to advanced calculation to determine the radial end-points. **Fig. 9-2** shows a typical radial system with 8 radials surrounding a center antenna that passes through the page. (See model 9-2.) To construct or symbolize a set of radials, you may use any system similar to the following simple set-up.

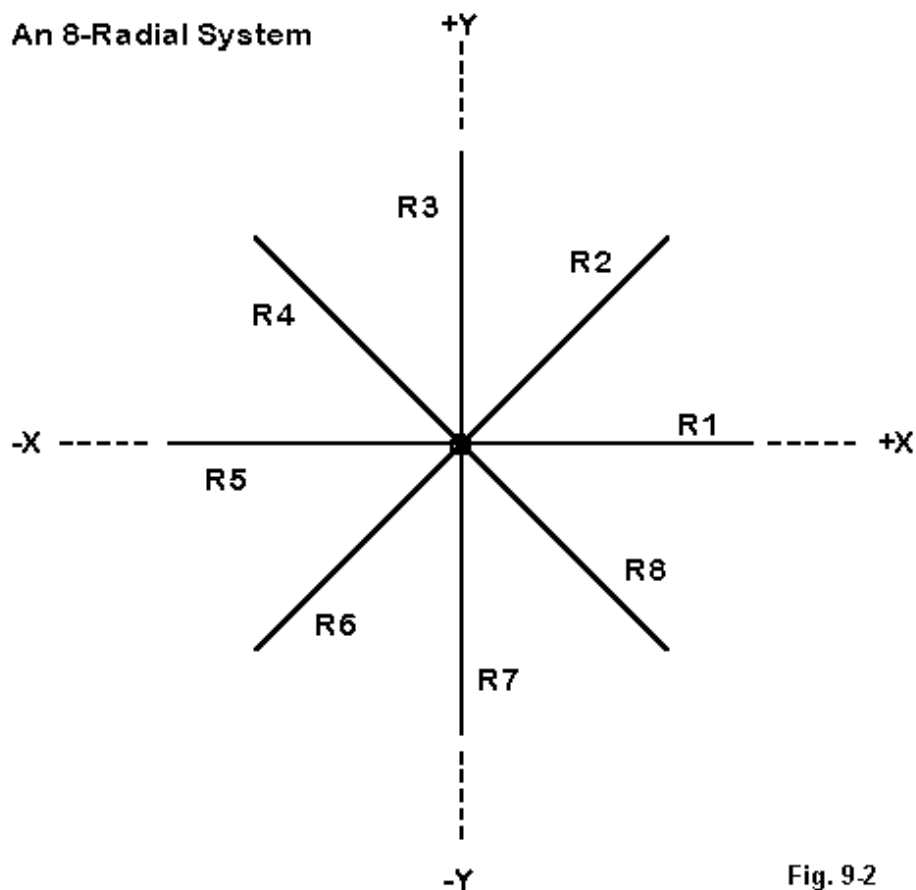


Fig. 9.2

Set the X-axis as the reference line. Let L be the length of each radial (so that a sample radial might have the values at end 1 of $X=0$ and $Y=0$ and at end 2 of $X=L$ and $Y=0$). If 8 is the number of radials, then the angle (A) between each radial is $A=360/8$. The radials will then have end-2 coordinates as follows:

Radial Number	End-2 X	End-2 Y
1	$X1 = L$	$Y1 = 0$
2	$X2 = \cos A * L$	$Y2 = \sin A * L$
3	$X3 = \cos (2 * A) * L$	$Y2 = \sin (2 * A) * L$
* * * * *		
8	$X2 = \cos (7 * A) * L$	$Y2 = \sin (7 * A) * L$

In this example, radial 1 is the one which will extend from the center point to the values $X=L$ and $Y=0$. You may set this table into a modeling program offering symbolic dimensional notation, or you may prepare the table and its results as a preliminary exercise on a prepared form to keep your work systematic and traceable. Symbolic entry does allow revision of the length of the entire radial system with a single change for the value of L. You must, however, have an entry for each radial in the system. Hence, changing the number of radials in a system requires the introduction of the correct number of wires to complete the array.

Once you have modeled radial systems with various numbers of wires, you should save your collection under some system of filenames. You may pull them from the files, scale them to the frequency of a present project as well as to the desired wire size and type, adjust X, Y, and Z values to position them for the new project, and then construct the new radiating wires atop them.

What Some Programs Offer

Various versions of NEC and MININEC offer us different ways of setting up the wires for a radial system. Here are a few examples.

AO offers symbolic wire data entry. Hence, you may simply plug in the equations for each radial coordinate and present the value of A and L. If you have these preset into a number of files for each desired radial set size, you can join the relevant file to your proposed antenna file externally to AO and give the new file an .ANT extension. Then you can complete the radial system by adding 1. an antenna

element wire (or more than one, if needed), 2. a value for L, which itself may be symbolic and determined by some other factor, such as frequency.

EZNEC offers an automated radial maker. From the wire description page, you enter the first wire of the radial set, ordinarily specifying the length along either the X or the Y axis. Then the program will ask which wire forms the basic radial (and it always uses end 1 as the pivot point). Finally, it will ask how many radials. The result that appears will be a set of radials with the correct angular spacing and the correct end coordinates. Because the operation is so fast, it does not usually pay to save radial files as separate entities in EZNEC.

NECWin Plus appeared in 1999. It contains an interesting spread-sheet that permits flipping from a symbolic data page to an actual value page, so that you can see a radial set both ways. In this program, it will definitely pay to save sample radial set models as separate files for later reuse. As well, the program contains a module for synthesizing both radial systems and wire-grid rectangles.

If you have a sloping ground plane system, as illustrated in **Fig. 9-3**, you can back up the calculations in symbolic entry on step and add a rotational component to take the final length at its level coordinates and rotate them downward by the appropriate number of degrees. Within EZNEC, you may group rotate the wires in the radial set downward in one command set. The following table, drawn from a 40-meter model, illustrates the wires of a sloping radial system with the radial length set at 9.8 meters, but sloping downward at a 45-degree angle. (You can verify this length by taking any one of the X or Y values (6.93) and adding its square to the square of the difference between the End 1 and End 2 Z values (6.93). The square root of the sum is the leg length (9.8), via our old friend, the Pythagorean theorem.)

Wire #	End 1 Conx	X	Y	Z	End 2 Conx	X	Y	Z
1		0.000,	0.000,	50.135	W2E1	0.000,	0.000,	40.000
2	W1E2	0.000,	0.000,	40.000		6.930,	0.000,	33.070
3	W1E2	0.000,	0.000,	40.000		0.000,	6.930,	33.070
4	W1E2	0.000,	0.000,	40.000		-6.930,	0.000,	33.070
5	W1E2	0.000,	0.000,	40.000		0.000,	-6.930,	33.070

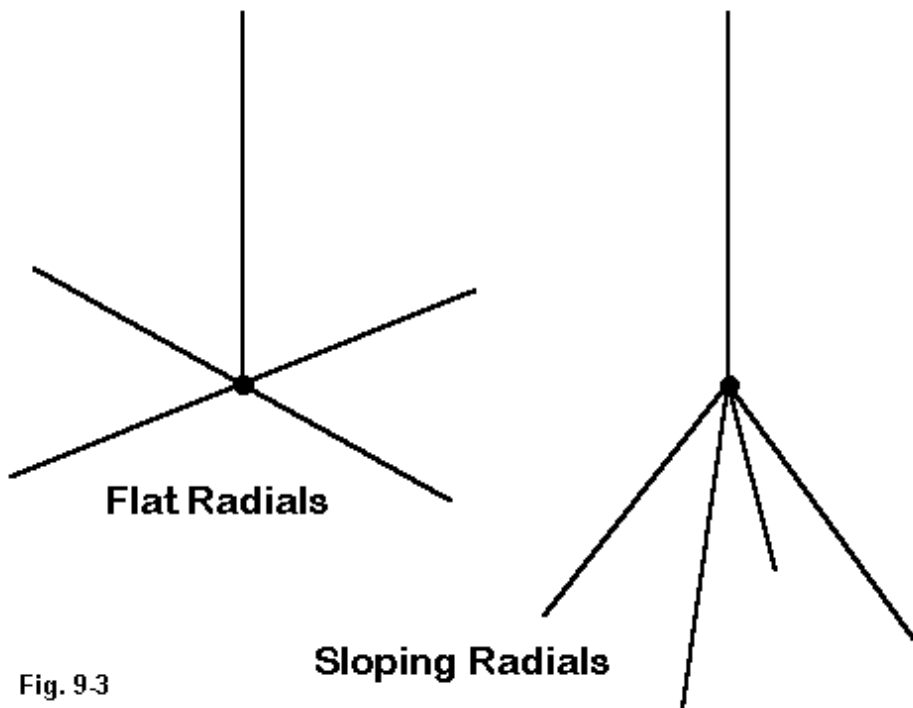


Fig. 9-3

A wise exercise is to familiarize yourself with the coordinate values you might expect using a free-space model centered in the “0, 0, 0” coordinates. Create on paper or as models some flat radials systems using 4, 8, and 16 radials, as well as some 4-radial sloping systems at 30, 45, and 60 degrees. Set the radial length at some round number, such as 10 (any units will do). Make a chart of the coordinate values you obtain for each system. Then, when you construct actual radial systems, you can compare the results to the chart to see if all the values appear to be reasonable. Making a further cross check with the antenna structure viewing sys-

tem of your software will likely ensure a correct model—or at least one that is set up as you intend it to be.

As a sample, my 16-wire reference table (model 9-4), with an arbitrary radial length of 10, looks like this:

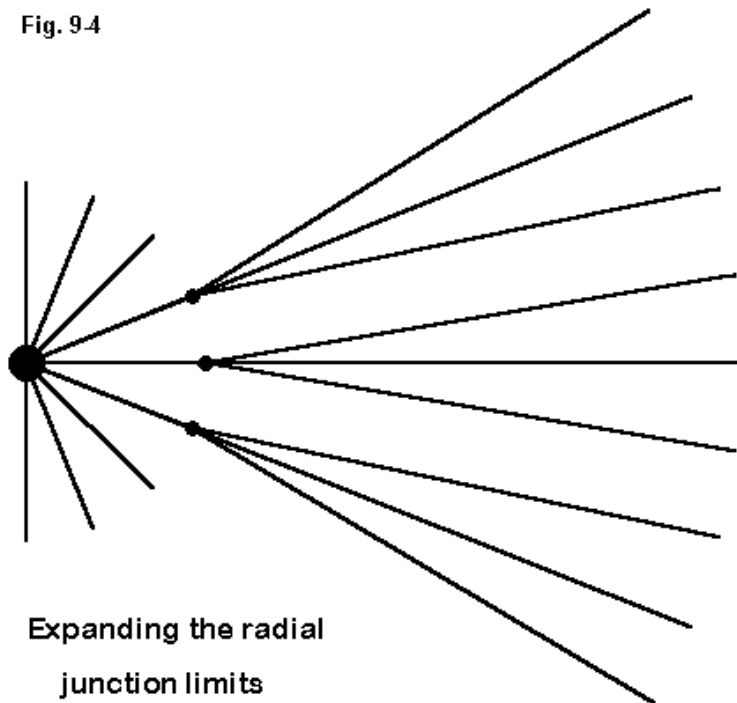
Radial	End 1			End 2		
Wire No.	X	Y	Z	X	Y	Z
1	0.000,	0.000,	0.000	10.000,	0.000,	0.000
2	0.000,	0.000,	0.000	9.239,	3.827,	0.000
3	0.000,	0.000,	0.000	7.071,	7.071,	0.000
4	0.000,	0.000,	0.000	3.827,	9.239,	0.000
5	0.000,	0.000,	0.000	0.000,	10.000,	0.000
6	0.000,	0.000,	0.000	-3.827,	9.239,	0.000
7	0.000,	0.000,	0.000	-7.071,	7.071,	0.000
8	0.000,	0.000,	0.000	-9.239,	3.827,	0.000
9	0.000,	0.000,	0.000	-10.000,	0.000,	0.000
10	0.000,	0.000,	0.000	-9.239,	-3.827,	0.000
11	0.000,	0.000,	0.000	-7.071,	-7.071,	0.000
12	0.000,	0.000,	0.000	-3.827,	-9.239,	0.000
13	0.000,	0.000,	0.000	0.000,	-10.000,	0.000
14	0.000,	0.000,	0.000	3.827,	-9.239,	0.000
15	0.000,	0.000,	0.000	7.071,	-7.071,	0.000
16	0.000,	0.000,	0.000	9.239,	-3.827,	0.000

Core Limitations

Because ground plane wires, even sloping ones, have a horizontal radiation component, MININEC ground planes should not be lower than 0.2 wavelength above ground, the same accuracy restriction imposed on all horizontal wires. Similarly, the quick ground (reflection coefficient) system attached to NEC-2 is limited in accuracy below about the same minimum height. DOS versions of MININEC are further limited by the low number of total segments that a model may include. Using the provisions of symmetry in AO can increase the effective number of segments. Windows implementations of MININEC have broken this barrier within the calculating core.

The Sommerfeld-Norton ground system in NEC-2 permits the modeler to place radials very close to the surface, so long as the radials are at least 0.001 wavelength above the ground and the height is equal to or greater than the diameter of the radial wire. Although most recommendations suggest that no special treatment is needed for the radial wires at this level, some difference in results appear between identical radials with equal segmentation and with tapered length segmentation. (Tapered length segmentation involves using very short segment lengths closest to the junction, with increasing segment length outward until some preset segment length is reached or the radial ends.) If the modeler attempts to place thin radials lower than 0.001 wavelength, but never lower than 0.0001 wavelength, then length-tapering techniques must be used.

Fig. 9.4



NEC-2 recommends a maximum of 30 wires to a junction. Although this limitation can be judiciously violated to a small degree, larger radials systems require special techniques. **Fig. 9-4** suggests a route to overcoming the limit. Some radials join together at a wire junction shy of the main junction, so that the hub sees no more than the recommended maximum number of wires.

Since radials usually form symmetrical patterns, structuring them with tapered diameter materials requires no corrective replacements (such as those we might need with linear Yagi elements). Like a capacity hat at the top of an element, the fields cancel due to the symmetrical arrangement of radials, resulting in the introduction of no errors.

An Example of a “Near-Ground” Radial System Model

Let's examine a typical model of a near-ground radial system (model 9-5). The frequency is 7.05 MHz or a wavelength of 42.55 meters. 0.001 wavelength is about 0.043 meters. Let's round this to 0.05 meters for the example. We can make the ground plane from 6.34-mm diameter (about 0.1") wire. We shall also place the ground plane beneath a 50 mm (nearly 2") diameter 1/4 wavelength monopole and above average Sommerfeld-Norton ground conditions (conductivity = 0.005 s/m; dielectric constant = 13). If we restrict ourselves to a 4-radial model initially, we have room to print the standard NEC file for such a model. It includes 6061-T6 material conductivity constant in the type 5 load (LD) entries. The final entry (RP) is a pattern request for an elevation plot. The file looks like this:

```
CM 1/4 wavelength vertical: 7.05 MHz
CM 4 radials, 0.05 m
CE
GW 1 10 0 0 10.1846 0 0 .05 .025
GW 2 10 0 0 .05 11.796 0 .05 .00317
GW 3 10 0 0 .05 0 11.796 .05 .00317
GW 4 10 0 0 .05 -11.796 0 .05 .00317
GW 5 10 0 0 .05 0 -11.796 .05 .00317
GS 0 0 1
GE 1
GN 2 0 0 0 13 .005 0 0 0 0
EX 0 1 10 0 1 0
```



```
LD 5 1 1 10 2.4938E7
LD 5 2 1 10 2.4938E7
LD 5 3 1 10 2.4938E7
LD 5 4 1 10 2.4938E7
LD 5 5 1 10 2.4938E7
FR 0 1 0 0 7.05 0
RP 0 181 1 1000 -90 0 1 1
EN
```

A 32-radial version of the same antenna would simply be much longer, with 33 wires total (and 33 corresponding LD entries). Interestingly, the gain of the 4-radial antenna is -1.57 dBi with a take-off angle of 64 degrees (zenith, or 26 degrees elevation), and the source impedance is $49 + j 52$ Ohms. In contrast the 32-radial antennas shows a gain of 0.12 dBi at the same take-off angle, with a source impedance of $33 - j 0$ Ohms. As we raise the ground plane to twice the height, both antennas show a rise in gain and a very slight lowering of the take-off angle, although the 32-radial version change is very small indeed. The 4-radial antenna reports a significant drop in both the feedpoint resistance and reactance, while the change in the 32-radial impedance values is about 1 Ohm in each category.

What are we to make of these reports? First, the larger ground plane exhibits much more stable characteristics by nearly a 10:1 factor. Second the reported feedpoint impedance of the larger array is much closer to free space values expected from the antenna than is the report from the 4-radial model. On the other hand, the values continue to change with small adjustments in the height of the ground plane, even for the larger model. We would be hard pressed to know precisely which values at which height to use for more than general guidance.

Is the Model Accurate?

Assessment of a model of a surface or buried radial system using radials that are in close proximity to ground is a complex question. In general, comparative values between different versions of the antenna and its ground radials will be correct. However, absolute values of far field gain and of source impedance are subject to several limitations of the NEC modeling system in addition to the slight variability we encountered with the larger ground plane.

First, the ground type (in terms of conductivity and of dielectric constant or permittivity) specified for the model presumes a uniform or homogenous soil. Since the soil at levels most influential on antenna performance may be stratified, the model may not accurately reflect reality. The potential inaccuracy may be especially acute at lower frequencies, where the field may penetrate the soil to a considerable depth, for example, several feet in the lower HF range.

Second, ground conductivity may vary with frequency. Data on local soil susceptibility to this factor is often difficult to obtain. Even when the soil value is known for one frequency, sweeping the frequency over a large range may introduce errors relative to real conditions.

Note that these limitations are functions of the NEC-2 core and the Sommerfeld-Norton ground calculation system. They are not limitations of any particular implementation of NEC and should not be held against a vendor's product.

So Where Does All of This Leave Us?

For modeling radial systems at least 0.2 wavelength above the ground, both MININEC and NEC-2 will do a very satisfactory job on small ground plane systems. The limitation of segments in DOS-MININEC restricts such radial systems to a quite small size, but highly elevated systems usually are small—3 to 4 radials at most. NEC-2 can safely model systems of up to 30 radials or so without the use of special techniques.

Modeling ground plane verticals close to the ground—as a stand-in for modeling radials systems either on or under ground—is a less certain enterprise. Because we lack assured data for such systems against which to compare the models, our efforts can provide only general guidance and comparative measures. (Even NEC-4 suffers from limitations of soil data.) The complexities of the ground on and in which we place radials preclude the level of confidence we give modeling numbers for well-elevated horizontal arrays.

Nevertheless, do not discount altogether near-ground modeling of vertical monopoles with ground planes. Comparative data and trends provide valuable information that can be used to improve many antenna systems. Meanwhile, a new gen-

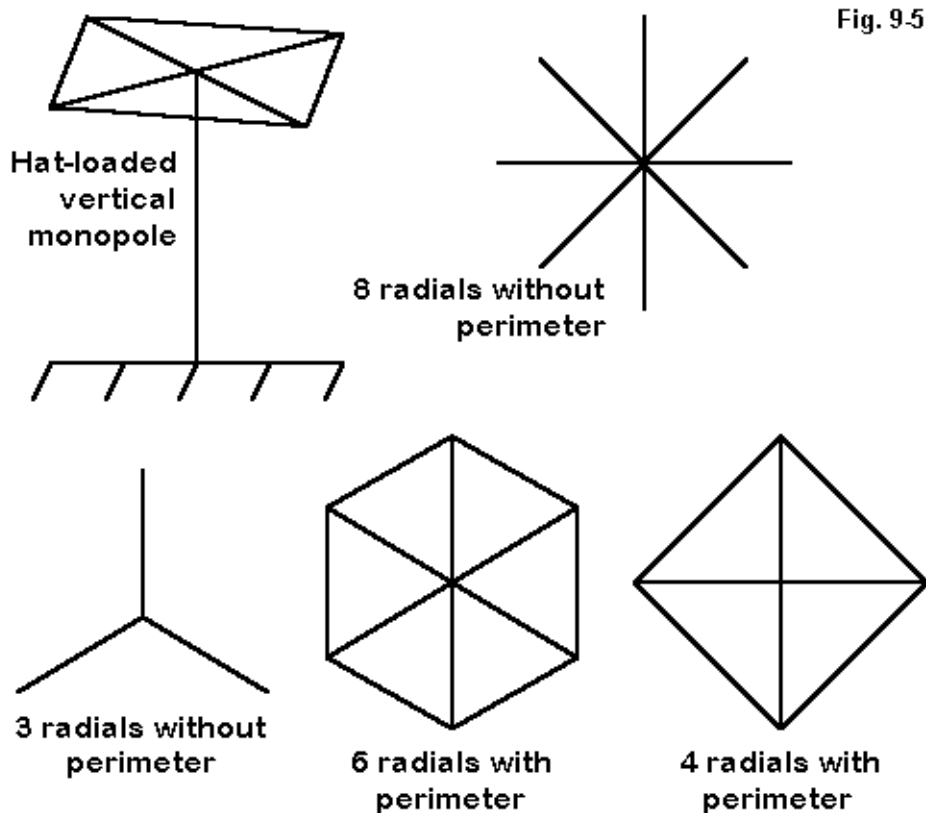
eration of work is underway to tighten the correlation of modeling and experimental results.

Other Applications for Radials

Radials have, of course, their second most common use at the far end of shortened linear elements in structures traditionally called “capacity hats.” These symmetrical, field-canceling structures, shown by example in **Fig. 9-5**, simply supply the linear element with the missing length to achieve resonance or some other specified condition. Although shown applied to the top of a vertical monopole in the sketch, they work in the very same manner at the ends of a shortened dipole element.

The principles of modeling hats are identical to those of modeling radials. Because the fields of a hat cancel, we introduce no significant errors into the model by using wire diameters different from that of the main element. The only construction difference in the model will be that for dipoles, the axes on which we place the radials will not always be X and Y, but will involve Z plus either X or Y.

Perimeter wires are often part of the hat structure. In working with the model to determine the best radial length, we should take one extra step of care to ensure that the perimeter wire remains attached to the spoke ends. If we enter the radial dimensions in symbolic form, as we would with AO or NECWin Plus, the attachment is easy: simply specify the terminating points of each perimeter wire in the same coordinate notation as the appropriate radial ends.



In a program such as EZNEC, where all wires are entered numerically, the task is no harder. When modifying the radial length of an existing hat model that has a perimeter wire, first activate the “preserve connection” feature. Then modify the outer radial end coordinates as a group function, using the “l” or length specification

option. The entire hat system of radials and perimeter wires will grow or shrink according to the selected length value. For model development, it is useful to list all of the radial wires in one batch (which would occur naturally, if you used the automated radial maker) and then add the perimeter wires as a second batch.

Although we have been most concerned with the construction of radial systems for use in ground plane modeling, the same modeling techniques apply wherever we need a radial structure. The automated radial-makers in entry level programs substitute for the GR function built into the NEC core and available on more advanced programs.

* * * * *

Models included: 9-1 through 9-5.



10. Tapering to Perfection

Everyone who models HF (and low VHF) beams is faced with dealing with elements that do not have a constant diameter. We refer to such elements as tapered-diameter (or tapered-radius) elements, in contrast to elements with a uniform diameter, that is, elements that are untapered. The latter have somehow picked up in some literature the self-contradictory name of mono-tapered elements.

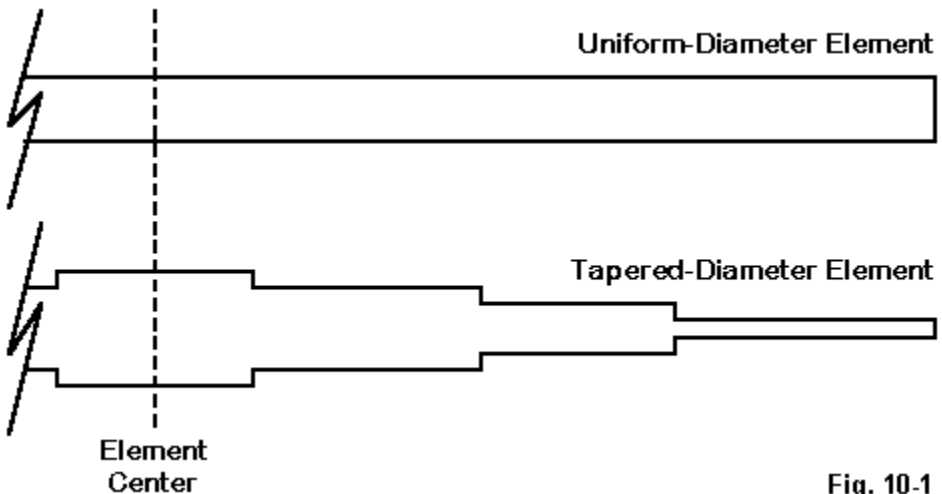


Fig. 10-1

Fig. 10-1 exaggerates—but only to a degree—the amount of stepping involved in most elements composed of several diameters of aluminum tubing. Normally, the maximum diameter shift from one section to another is about 0.25". However, some designers use a short, fat center wire in the element to simulate the effects of ele-

ment-to-boom mounting plates and hardware. In these cases, the diameter shift may amount to as much as 2".

Native NEC-2 cannot handle tapered-diameter elements. It simply yields inaccurate results—ordinarily by over-reporting the gain and under-reporting the source impedance of the element. Consequently, in some implementations of NEC-2, correction factors have been installed to yield more accurate results.

Let's begin by comparing some simple cases of uncorrected and corrected NEC-2 reports for some simple tapered-diameter elements set up as dipoles in the 14 MHz range. The following table repeats an exercise we first encountered in installment 3 of this series. The entries in the table beneath each numbered label represent, first, the distance(s) from the element's center at which a diameter shift occurs and, second, the diameter(s) of the element between shift points. The "Cor." entries give the length from center and the diameter of the substitute element used by NEC-2 to provide corrected results. See models 3-1 through 3-5.

Length(s) in inches Diameter(s) in inches		Free Space Gain in dBi	Source Impedance (R +/- jX Ohms)
1. Uniform element (no corrective needed)			
-201.25...0		2.12	71.8 - j 0.6
1.0			
2. One step, far out			
-204...-150...0	No Cor.	2.14	73.0 + j 4.4
.75 1.0			
-201.45...0	Cor.	2.12	72.0 + j 0.4
0.966			
3. One step, near center			
-204...-50...0	No Cor.	2.22	72.4 + j 5.2
.75 1.0			
-201.661...0	Cor.	2.13	71.8 - j 0.5
0.792			
4. Two steps, modest taper			
-205.75...-100...-20...0	No Cor.	2.32	72.5 + j10.6
.75 1.0 1.25			
-201.49...0	Cor.	2.13	71.9 + j 0.1
0.889			

5. Two steps, more extreme taper			
-208.5...-100...-20...0	No Cor.	2.82	67.6 + j17.1
.75 1.0 2.5			
-201.63...0	Cor.	2.13	72.1 + j 0.9
0.895			

Notice especially the trends. Without correction, the closer the diameter-step to the element center (and the source), the higher the reported gain and the lower the reported source impedance. The more numerous the steps and the greater the step sizes, the higher the reported gain and the lower the reported source impedance. By entry number 5, the figures have grown incredible. In contrast, the corrected elements yield figures that are all quite reasonable for a resonant 14 MHz dipole.

The simple lesson from this exercise is always to use the stepped-diameter correction feature when modeling stepped-diameter elements in NEC-2. However, there are numerous limitations to the tapered-diameter element correction scheme, as well as some cautions to be observed in using it. So let's spend a little time understanding the feature and learning to use it wisely. In addition, let's also pick up the habit of making sure that it is in effect when we want it to be.

A Bit of Background

Wherever the tapered-diameter correction feature is implemented, it creates a substitute element for the tapered one. The substitute element has a uniform diameter and a length that yield essentially the same self-impedance as the tapered-diameter element. Thus, the substitute can be expected to behave—either by itself or in an array of elements—in the same way as the tapered element.

The first method of calculating substitute Yagi elements was developed (to the best of my knowledge) by Jim Lawson, W2PV, in chapter 7 of his classic *Yagi Antenna Design*. Brian Beezley, K6STI, created improve algorithms for use in his NEC-Wires program. A quite different system was developed and explained by Dave Leeson, W6QHS, in chapter 8 of his equally classic *Physical Design of Yagi Antennas*. EZNEC, by W7EL, and NEC-Win Plus by from Nittany Scientific both implement the Leeson correctives.

The Leeson correctives were calibrated against the MN (K6STI) version of MININEC (a part of AO). Indeed, MININEC is considered to be the standard of comparison for all tests of HF tapered-diameter calculations. It exhibits no problems with tapering schedules of all sorts. (A “tapering schedule” is simply where along an element the diameter changes and by how much.) We shall a bit farther on do some comparing of MININEC with both corrected and uncorrected NEC-2 outputs.

The basic principle of the substitute element is to derive an element with a uniform diameter and the same self-impedance as the original tapered-diameter element. The calculation method is accurate at and near element resonance, where the current distribution along the element is nearly sinusoidal (that is, nearly a sine wave in shape). This condition sets a number of limitations for the use of the corrective.

Some Practical Boundaries for Substitute Elements

In practical terms, the limitations of the Leeson substitute element technique are straightforward if we remember what the calculation method requires.

1. The tapered-diameter element must be within about 15% of resonance for the substitute element to be an accurate replacement. The near-resonance requirement applies to the physical length of the antenna without regard to loads, transmission lines, and sources that may be present. EZNEC and NEC-Win Plus, for example, cut off the availability of the correction factor if the element does not meet this standard.

In antennas with many elements, only those meeting this (and other) constraints will have substitute elements calculated. Consequently, a tri-band beam might have the elements for one band (as determined by the frequency selection) substituted, but not those for other bands. This fact invokes a requirement for caution in using the results of such a model. For example, on some tri-band beams, the forward 20-meter director can affect performance on 10-meter, especially in limiting the upper frequency performance on that band. Depending upon beam design, the 20-meter director may carry significant current on 10 meters and be somewhat long as a full wavelength element on the upper band. However, it would not ordinarily have a

uniform-diameter substitute calculated for it when the antenna is modeled at 10 meters.

2. If the tapered-diameter element is open at both ends (a dipole), it must be symmetrical about a center point along (but not necessarily on) one of the axes of the coordinate system. This requirement demands that the steps in a dipole's taper occur at equal distances from the center point of the element and in equal diameter steps for each corresponding step. EZNEC and NEC-Win Plus both report non-symmetries. In modeling, perhaps the chief reason for the absence of symmetry is transposing numbers when entering them in the wire table. The second most common reason for encountering non-symmetry is a second numeric transposition.

3. Loads, transmission lines, and sources are permitted only at the center point. Only one of each is permitted if the midpoint of the symmetrical element has no wire junction. Where the center is at a wire junction, split sources and loads may be used on segments immediately adjacent to the center-point junction. A load at a distance from the center—whether a resistance/reactance load, a resistance/inductance/capacitance load, or a shorted or open length of transmission line—would create an abrupt non-sinusoidal change in current magnitude along the element length, thus violating one of the conditions of accuracy. Center loads, transmission lines, and sources do not ordinarily disturb the current magnitude curve, although they may limit its highest value. Once more, EZNEC and NEC-Win Plus identify misplaced loads, sources, and transmission lines.

Note that a serious loading coil used to shorten an element severely would ordinarily be used with an element already too short to pass the length test. Hence, such antennas could not access the corrective process, even without the load in place.

For monopoles, connection to ground is essential for the substitution system to work. In these cases, the antenna element must be within length boundaries with respect to $1/4$ wavelength resonance, and all source, load, and transmission line attachments must be to the segment immediately adjacent to the ground. For NEC-2, these requirements effectively limit the usable ground type to perfect ground. A monopole connected directly to either the Sommerfeld or the reflection coefficient ground will normally return untrustworthy results. You may wish to review the earlier episode on ground radial systems.

4. The taper-diameter element must be truly linear—that is, all wires making up the tapered-diameter element must be collinear. Ends must be open or, at most, one end may be grounded. Therefore, tapered diameter sections of elements that make up a complex geometry disable the corrective system. As a result, tapered-diameter elements used in Moxon rectangles, quad loops, capacity-hat dipoles/monopoles, and similar antennas cannot be directly modeled in NEC-2.

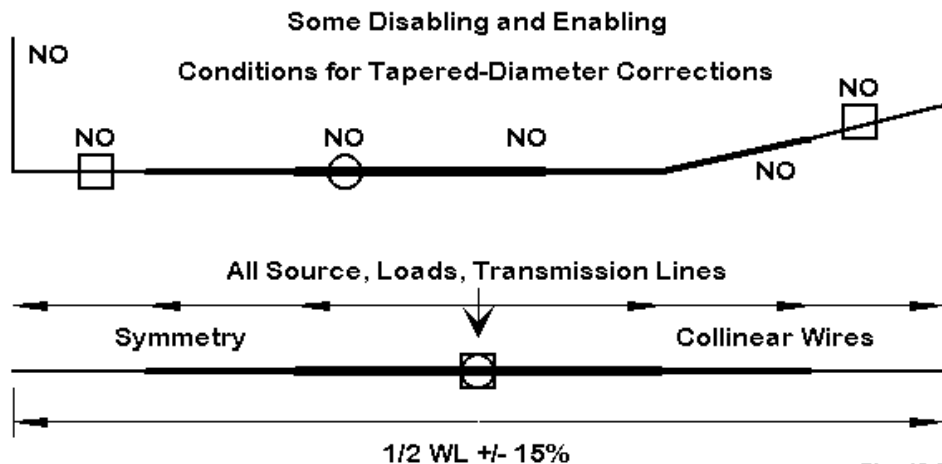


Fig. 10-2

Fig. 10-2 summarizes the rules by calling attention to what goes beyond normal boundaries. The upshot is that the corrective system is quite limited in its applications. However, the design of HF Yagi beams and other arrays with linear tapered-diameter elements is such a widespread application of NEC-2 that the corrective system is considered essential to most users.

Some Practical Examples and a Subtle Limitation

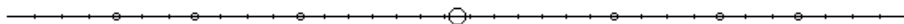
Let us look at some examples of the tapered-diameter correction factor in operation to further our familiarity with it and to gain some sense of what counts as

reasonable expectations of it. We shall look at some wire tables drawn from the EZNEC implementation of NEC-2. Then we shall examine 5 sets of output reports: 1. the results of a MININEC version of the model in AO; 2. the results of uncorrected EZNEC NEC-2; 3. the results of corrected EZNEC NEC-2; 4. the results of creating a single-wire uniform element of the same length, diameter and total number of segments as the substitute group of wires in the corrected NEC-2 report; and 5. the results of (uncorrected) NEC-4. The AO models will have their center sections decreased by one segment to properly place the source at the junction of segments and at the element center. NEC-4 models hypothetically should need no correction. The models are all of the driven elements of large Yagi antennas, so their independent source impedances may not be quite resonant.

1. A 14.175 MHz driven element of W6NGZ design (model 10-1):

W6NGZ Driven Element

Fig. 10.3



W6NGZ Frequency = 14.175 MHz.

Wire Loss: Aluminum - Resistivity = 4E-08 ohm-m, Rel. Perm. = 1

WIRES

Wire Conn.—End 1	(x,y,z : in)	Conn.—End 2	(x,y,z : in)	Dia(in)	Segs
1	-205.95, 79.800, 0.000	W2E1	-156.00, 79.800, 0.000	6.25E-01	4
2	W1E2 -156.00, 79.800, 0.000	W3E1	-120.00, 79.800, 0.000	7.50E-01	3
3	W2E2 -120.00, 79.800, 0.000	W4E1	-72.000, 79.800, 0.000	8.75E-01	4
4	W3E2 -72.000, 79.800, 0.000	W5E1	72.000, 79.800, 0.000	1.00E+00	13
5	W4E2 72.000, 79.800, 0.000	W6E1	120.000, 79.800, 0.000	8.75E-01	4
6	W5E2 120.000, 79.800, 0.000	W7E1	156.000, 79.800, 0.000	7.50E-01	3
7	W6E2 156.000, 79.800, 0.000		205.950, 79.800, 0.000	6.25E-01	4

SOURCES

Source	Wire Seg.	Wire #/Pct Actual	From End 1 (Specified)	Ampl.(V, A)	Phase(Deg.)	Type
1	7	4 / 50.00	(4 / 50.00)	1.000	0.000	I

Ground type is Free Space

This antenna is symmetrical in the X-axis. Leaving the Y coordinate at its original spacing from the reflector is no problem for implementing the correction factor.

The AO model uses 12 segments in the long center element. Note that the center element occupies about 35% of the total element length. Here are the reports.

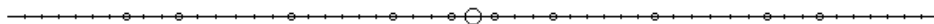
Report Source	Gain dBi	Source Z (R+/-jX)
AO	2.13	77.0 + j 11.6
NEC-2 (No Cor.)	2.21	77.3 + j 23.6
NEC-2 (Cor.)	2.14	74.7 + j 11.9
NEC-2 (1-Wire)	2.14	74.7 + j 11.9
NEC-4	2.17	76.1 + j 12.9

This 35-segment model shows only a slight displacement from the MININEC standard. The key difference is in the reactance report, which seriously displaces resonance in uncorrected NEC-2, while the corrected NEC-2 report coincides with the AO report. The 1-wire substitute element agrees exactly with the subdivided group that composes the substitute model of the element: both show a source resistance that is about 2.3 Ohms lower than the AO report. Finally, note that the NEC-4 report, while closer by far to the corrected NEC-2 and AO reports than to the uncorrected NEC-2 report, is still a bit high in gain and also in the reported source reactance.

2. A 14.175 MHz driven element of WB0DGF design (model 10-2):

WB0DGF Driven Element

Fig. 10-4



WB0DGF Frequency = 14.175 MHz.

Wire Loss: Aluminum - Resistivity = 4E-08 ohm-m, Rel. Perm. = 1

WIRES

Wire Conn.—End 1 (x,y,z : in)	Conn.—End 2 (x,y,z : in)	Dia(in)	Segs
1 -170.00,-209.50, 0.000	W2E1 -170.00,-156.50, 0.000	4.38E-01	7
2 W1E2 -170.00,-156.50, 0.000	W3E1 -170.00,-132.50, 0.000	6.25E-01	3
3 W2E2 -170.00,-132.50, 0.000	W4E1 -170.00,-82.000, 0.000	8.75E-01	6
4 W3E2 -170.00,-82.000, 0.000	W5E1 -170.00,-36.000, 0.000	1.12E+00	6
5 W4E2 -170.00,-36.000, 0.000	W6E1 -170.00,-10.000, 0.000	1.25E+00	3
6 W5E2 -170.00,-10.000, 0.000	W7E1 -170.00, 10.000, 0.000	1.25E+00	3
7 W6E2 -170.00, 10.000, 0.000	W8E1 -170.00, 36.000, 0.000	1.25E+00	3
8 W7E2 -170.00, 36.000, 0.000	W9E1 -170.00, 82.000, 0.000	1.12E+00	6
9 W8E2 -170.00, 82.000, 0.000	W10E1 -170.00,132.500, 0.000	8.75E-01	6

```

10 W9E2 -170.00,132.500, 0.000 W11E1 -170.00,156.500, 0.000 6.25E-01 3
11W10E2 -170.00,156.500, 0.000 -170.00,209.500, 0.000 4.38E-01 7

```

SOURCES

Source	Wire Seg.	Wire #/Pct Actual	From End 1 (Specified)	Ampl.(V, A)	Phase(Deg.)	Type
1	2	6 / 50.00	(6 / 50.00)	0.707	0.000	V

Ground type is Free Space

Two features of this element design (symmetrical in the Y-axis) are notable for our purposes. First, the center section (composed of three wires of 1.25" diameter) is half the length of the W6NGZ design, so that the first diameter step occurs closer to the source. Second, the diameter steps cover a greater range than the W6NGZ design, now ranging from 1.25" down to 0.4375" (in contrast to the W6NGZ range from 1.0" to 0.625"). Now let's look at the reports.

Report Source	Gain dBi	Source Z (R+/-jX)
AO	2.13	76.2 + j 0.8
NEC-2 (No Cor.)	2.33	78.3 + j 30.9
NEC-2 (Cor.)	2.13	72.4 + j 1.7
NEC-2 (1-Wire)	2.13	72.4 + j 1.7
NEC-4	2.21	75.8 + j 6.8

The gain and source reactance reports of AO and corrected NEC-2 are aligned, while the source resistance differs by about 3.8 Ohms, slightly more than the difference reported in the W6NGZ design. Note the high gain and inductive reactance difference for uncorrected NEC-2. Finally, note that NEC-4—for this more highly tapered-diameter model—shows a gain value that is no longer trustworthy.

3. A 14.175 MHz driven element of K6STI design (model 10-3):

K6STI Driven Element

Fig. 10-5



K6STI Frequency = 14.175 MHz.

Wire Loss: Aluminum - Resistivity = 4E-08 ohm-m, Rel. Perm. = 1

WIRES

Wire Conn.	End 1 (x,y,z : in)	Conn.	End 2 (x,y,z : in)	Dia(in)	Segs
1	-203.50, 72.000, 0.000	W2E1	-138.00, 72.000, 0.000	5.00E-01	8

2	W1E2	-138.00, 72.000, 0.000	W3E1	-96.000, 72.000, 0.000	6.25E-01	6
3	W2E2	-96.000, 72.000, 0.000	W4E1	-48.000, 72.000, 0.000	7.50E-01	6
4	W3E2	-48.000, 72.000, 0.000	W5E1	-4.000, 72.000, 0.000	8.75E-01	5
5	W4E2	-4.000, 72.000, 0.000	W6E1	4.000, 72.000, 0.000	3.42E+00	1
6	W5E2	4.000, 72.000, 0.000	W7E1	48.000, 72.000, 0.000	8.75E-01	5
7	W6E2	48.000, 72.000, 0.000	W8E1	96.000, 72.000, 0.000	7.50E-01	6
8	W7E2	96.000, 72.000, 0.000	W9E1	138.000, 72.000, 0.000	6.25E-01	6
9	W8E2	138.000, 72.000, 0.000		203.500, 72.000, 0.000	5.00E-01	8

SOURCES

Source	Wire Seg.	Wire #/Pct Actual	From End 1 (Specified)	Ampl.(V, A)	Phase(Deg.)	Type
1	1	5 / 50.00	(5 / 50.00)	1.000	0.000	V
Ground type is Free Space						

This design uses a short, wide-diameter center section to simulate the effects of element-to-boom hardware. Apart from this short center section, the subdivision of the element is intermediate between the W6NGZ and WB0DGF designs.

Report Source	Gain dBi	Source Z (R+/-jX)
AO	2.12	72.7 - j 13.1
NEC-2 (No Cor.)	4.66	42.1 + j 5.1
NEC-2 (Cor.)	2.16	68.9 - j 11.4
NEC-2 (1-Wire)	2.12	69.6 - j 11.5
NEC-4	3.06	59.1 - j 6.2

This 51-segment element makes every effort to equalize segment lengths from wire to wire within the group, letting the 8" center length determine the segment lengths in the other wires. Even so, the element presses the correction factor limits, showing some difference in gain between the corrected NEC-2 result and the 1-wire substitute element. The effect of a fat short center segment is clearly apparent in the uncorrected NEC-2 result, with its very high gain and very low source impedance report. Note that with this type of segmentation, uncorrected NEC-4 yields quite untrustworthy results.

Despite the fact that even close attention to segmentation density yields corrected NEC-2 results that press the limits of the system, I encounter many carelessly segmented tapered-diameter elements. Perhaps the modelers believe that any old arrangement of segments adding up to more than 10 per half wavelength will do the job. However, the segments adjoining the source segment must be very

close in length to the source segment itself. Moreover, it is always good practice to equalize segments lengths in any element. Let's see what happens if we carelessly segment the K6STI element.

4. A 14.175 MHz driven element of K6STI design (carelessly segmented) (model 10-4):

K6STI Driven Element: Inadequate Segmentation

Fig. 10-6



K6STI Frequency = 14.175 MHz.

Wire Loss: Aluminum - Resistivity = 4E-08 ohm-m, Rel. Perm. = 1

WIRES

Wire Conn.-End 1 (x,y,z : in)	Conn.-End 2 (x,y,z : in)	Dia(in)	Segs
1 -203.50, 72.000, 0.000	W2E1 -138.00, 72.000, 0.000	5.00E-01	4
2 W1E2 -138.00, 72.000, 0.000	W3E1 -96.000, 72.000, 0.000	6.25E-01	3
3 W2E2 -96.000, 72.000, 0.000	W4E1 -48.000, 72.000, 0.000	7.50E-01	3
4 W3E2 -48.000, 72.000, 0.000	W5E1 -4.000, 72.000, 0.000	8.75E-01	2
5 W4E2 -4.000, 72.000, 0.000	W6E1 4.000, 72.000, 0.000	3.42E+00	1
6 W5E2 4.000, 72.000, 0.000	W7E1 48.000, 72.000, 0.000	8.75E-01	2
7 W6E2 48.000, 72.000, 0.000	W8E1 96.000, 72.000, 0.000	7.50E-01	3
8 W7E2 96.000, 72.000, 0.000	W9E1 138.000, 72.000, 0.000	6.25E-01	3
9 W8E2 138.000, 72.000, 0.000	203.500, 72.000, 0.000	5.00E-01	4

SOURCES

Source	Wire Seg.	Wire #/Pct Actual	From End 1 (Specified)	Ampl.(V, A)	Phase(Deg.)	Type
1	1	5 / 50.00	(5 / 50.00)	1.000	0.000	V

Ground type is Free Space

The element is obviously the same as the one just run, but the segmentation has been cut to 25 segments total.

Report Source	Gain dBi	Source Z (R+-jX)
AO	2.12	72.7 - j 13.1
NEC-2 (No Cor.)	5.04	38.2 + j 2.5
NEC-2 (Cor.)	2.68	61.1 - j 10.8
NEC-2 (1-Wire)	2.12	69.5 - j 12.0
NEC-4	3.93	48.2 - j 5.8

The AO numbers have not changed, but the only NEC report that shows sensible values is the 1-wire substitute. Both sets of figures are consistent with those for the carefully segmented version of the element. Both the raw NEC-2 and NEC-4 values are sufficiently off the mark to be completely unusable.

Of great importance is the fact that the corrected NEC-2 figures for this modeled element are also beyond the limits of reliability by a considerable margin. In previous models, we watched the margin of error in corrected NEC-2 reports grow, even in well-segmented elements. In this model, the corrected NEC-2 model and the 1-wire substitute have the same diameter, length, and total number of segment. The only difference between them is the length of segments, especially around the source point: the NEC-2 corrected model has very unequal segment lengths, while the segment length of the 1-wire substitute is uniform throughout the element. This demonstration shows that careful attention to segmentation of tapered-diameter elements is vital to deriving reliable results from the correction system.

Some Practical Conclusions

This episode has been light on the graphics and heavy on the tables, but for a purpose. The tables show the data that can help us improve our modeling techniques when employing the tapered-diameter correction factor implemented in various versions of NEC-2.

The first step is to carefully model the tapered-diameter element so that it meets the basic criteria for implementing the correction system.

The second step is to check to ensure that the correction has been implemented. It is easy—despite programmers' best efforts—to overlook some error in the model set-up that may have prevented the automated correction system from working.

Third, even models that pass the basic tests require careful attention to segmentation to ensure that segments are as equal as the structure will permit.

Remember that errors tend to be cumulative. Unrealistically high gain reports on individual elements usually result in very significant over-estimates of the gain of an array within which the tapered-diameter elements are used. While such over-

estimates may be momentarily pleasing to the designer's ego, they can be most assuredly embarrassing if caught by someone else.

Finally, even if you have access to NEC-4, do not assume that it is accurate in all cases. While NEC-4 improves on the performance of NEC-2 with respect to tapered-diameter elements, it has limitations which suggest that the correction system should also be used with it in all but the simplest tapered-diameter models.

The tapered-diameter element correction system is a valuable feature in any version of NEC-2 that implements it. However, it deserves to be used with both caution and care.

* * * * *

Models included: 10-1 through 10-4. (.NEC and .NWP model dimensions in meters; .EZ model dimensions in inches.)



11. A Ground Is Just a Ground—Unless It Is a Model of a Ground

A perennial collection of questions from newer modelers relates to the use of a ground selection beneath a model. In addition, a number of models that I have checked have shown a variety of ground selection errors. So it seems fitting that we become a little better grounded in grounding.

Ground Types in NEC-2 and MININEC

NEC-2 offers the modeler several different choices of grounds beneath the antenna, including no-ground (free space). Each has its own set of best applications for both speeding preliminary work and for final analysis of greatest accuracy. Let's familiarize ourselves with the types of ground, their typical entry in a .NEC file, and what they mean.

1. *Free Space*: the term “free space” indicates that absence of any surface beneath the antenna. Therefore, the antenna radiates in all directions without reflections (other than those that are a function of the modeled antenna structure itself). Choice of free space or “no-ground” is often the best for comparison of antennas of similar types. Moreover, it usually provides the most rapid calculation speeds and yields the highest accuracy output data.

The absence of ground may appear in a .NEC file as simply no ground entry, that is, no GN card. It may also appear as the following:

GN -1

2. *Perfectly Conducting Ground*: This option places beneath the antenna a perfectly conducting surface with perfect reflection properties—at a distance specified by the entries in the Z-axis column of the wire geometry cards. Although a wire may contact a perfect ground, horizontal wires should be above ground by a factor of at least the following:

$$(h^2 + a^2)^{1/2} > 10^{-6} \lambda$$

where “h” is the wire height and “a” is the wire radius. In addition, h should be several times greater than “a” for valid results.

Since a perfect ground creates an image antenna identical to the original, it requires twice as long to fill the interaction matrix as a free space model. However, it is still faster than the use of either of the ground condition approximation systems. Thus, it has some interesting developmental applications. Selection of a perfect ground results in the following GN card entry:

GN 1

3. *Finite Ground: Reflection Coefficient Approximation:* Sometimes called the “fast” or “real” ground, this system employs an image model modified by Fresnel plane-wave reflection coefficient approximations for near fields. This system loses accuracy as an antenna is brought within several tenths of a wavelength toward the ground or lower and is most applicable to relatively compact antenna structures. Nonetheless, it yields quite reasonable results, especially for preliminary developmental work, and is much faster than the more accurate Sommerfeld-Norton method. (Current computer speeds have made the calculations speed difference between the reflection coefficient and the Sommerfeld ground systems largely irrelevant.)

Every finite ground requires at least the specification of values for the conductivity (in S/m) and the dielectric constant (permittivity) of the ground. A table of typical values for various types of soil appears further on in this column. In general, vertical antennas close to the ground are more sensitive to changes in soil type than are horizontal antennas at their typical heights well above the ground surface. A typical GN card for a “fast” ground using “average” soil would appear as follows:

GN 0 0 0 0 13 .005

4. *Finite Ground: Sommerfeld-Norton Method:* The more accurate but slower (by a factor of 4) Sommerfeld-Norton (S-N, or SOMNEC) ground method uses exact solutions for fields in the presence of the specified ground and is accurate very close to ground. In fact, you may place wires as close to an S-N ground as to a

perfect ground. For the highest accuracy of results for an antenna model above ground, the S-N method is the ground of choice. Except for the entry immediately following the card identification, a simple S-N ground entry looks identical to a reflection coefficient approximation ground entry card. For example,

GN 2 0 0 0 13 .005

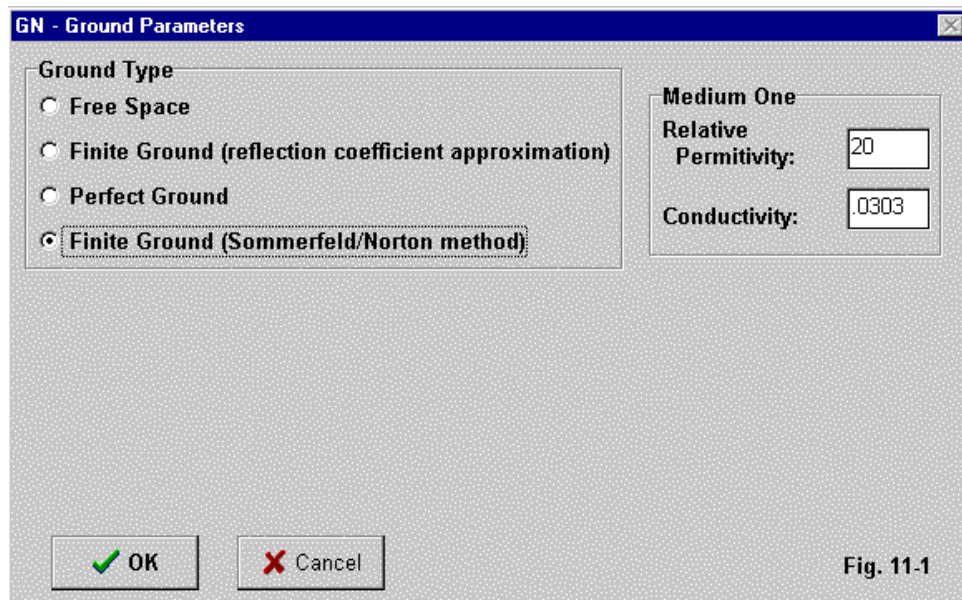


Fig. 11-1

As with all ground surfaces, the S-N ground extends indefinitely to the horizon. The S-N ground method also has provision for a second medium that extends a specified radius within ground specified in the card just shown. For our initial sampling of ground applications, we shall confine ourselves to simple grounds, that is, to grounds whose properties extend from the antenna to infinity.

Selecting a ground type within NEC-2 takes different forms depending upon the way in which the program implements the process. The illustration in **Fig. 11-1** shows the pull-down box used in NEC-Win.

MININEC offers fewer ground type selections. Equivalent to the ones in NEC-2 are free space and perfect ground. The “real” ground system in MININEC is comparable (although not identical) to the “fast” ground system in NEC-2. It becomes less accurate for horizontal wires at heights greater than the accuracy limit for the NEC-2 “fast” or reflection-coefficient ground system, and is generally considered inaccurate below 0.2 wavelengths of antenna height. (Note: these height limitations apply to horizontal or sloped wires with significant horizontally polarized radiation. Vertical antennas will require separate treatment.) In addition, the MININEC implementation of “real” ground applies only to far field calculations. MININEC calculates source impedance data using perfect ground, even when a “real” ground is selected. This feature can lead to source impedance errors for low horizontal wire antennas, such as those commonly used in the lower HF bands.

Selecting a MININEC ground type follows the same procedure as for NEC-2, with fewer selections. The illustration from NEC4WIN provides a sample in **Fig. 11-2**.

EZNEC, which is an implementation of NEC-2, offers the use of the MININEC ground system in addition to the standard NEC-2 selections. This system is intended for use with certain types of models of vertical antennas, a topic we shall reserve for another column. A common error of beginning modelers is to employ NEC-2 with the MININEC ground system for low arrays having a strong horizontally polarized component to the radiation field. Such arrays should ways use the Sommerfeld ground calculation system.

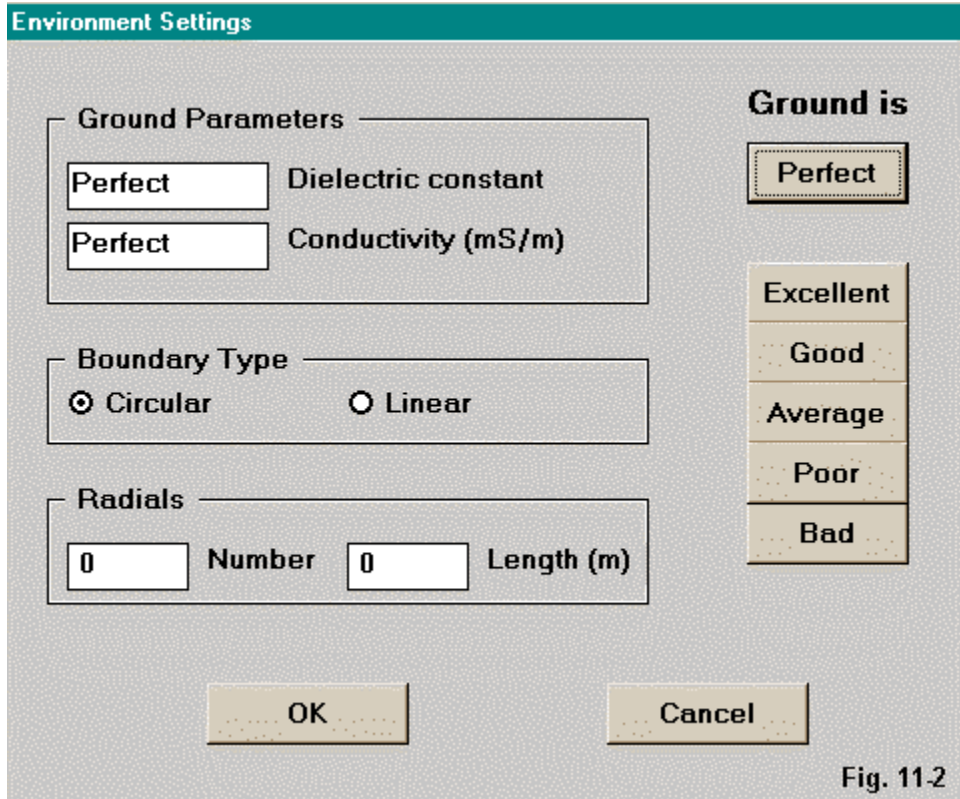


Fig. 11-2

Ground Quality Description

The illustrations show a second facet of ground selection when the choice is “real,” “fast,” or “S-N”: the selection of ground quality. There are two factors involved in filling in this choice: conductivity and dielectric constant (permittivity). Conductivity is measured in Siemens per meter (S/m) and must always be entered as a deci-

mal for anything other than salt water. One can measure soil conductivity, but the use of local or nationwide maps and tables is usually more convenient. The dielectric constant is normally beyond the means of most amateurs to measure and should be derived from a listing. It is virtually always an integer.

NEC and MININEC create from the two numbers a composite used in creating values to correctly derive the reflections that contribute to the far field. Be certain to correctly locate the entry position for each value in the set. Reversing the values can create some interesting but erroneous modeling results. I once encountered a Beverage antenna with an apparent gain of 17 dBi until the ground quality figures were correctly placed.

The following soil descriptions are commonly used in antenna modeling. Always substitute more precise values wherever known. The table represents an adaptation of values found in *The ARRL Antenna Book* (p. 3-6), which are themselves an adaptation of the table presented by Terman in *Radio Engineer's Handbook* (p. 709), taken from "Standards of Good Engineering Practice Concerning Standard Broadcast Stations," *Federal Register* (July 8, 1939), p. 2862. Terman's value for the conductivity of the worst soil listed is an order of magnitude lower than the value shown here.

Soil Description	Conductivity in S/m	Permittivity (Dielectric Constant)	Relative Quality
Fresh water	0.001	80	
Salt water	5.0	81	
Pastoral, low hills, rich soil, typical from Dallas, TX, to Lincoln, NE	0.0303	20	Very Good
Pastoral, low hills, rich soil, typical of OH and IL	0.01	14	Good
Flat country, marshy, densely wooded, typical of LA near the Mississippi River	0.0075	12	

Pastoral, medium hills, and forestation, typical of MD, PA, NY (exclusive of mountains and coastline)	0.006	13	
Pastoral, medium hills, and forestation, heavy clay soils, typical of central VA	0.005	13	Average
Rocky soil, steep hills, typically mountainous	0.002	12-14	Poor
Sandy, dry, flat, coastal	0.002	10	
Cities, industrial areas	0.001	5	Very Poor
Cities, heavy industrial areas, high buildings	0.001	3	Extremely Poor

One of the preset values pairs offered by EZNEC is called “good” but uses the values for average soil from the table above.

Some Ground Modeling Limitations

In general, the best compact source of information on the effects of ground on antenna performance is Chapter 3 of recent editions of *The ARRL Antenna Book*. The information there will help you understand some of the limitations of modeled grounds.

First, modeled grounds consist of flat, uncluttered surfaces. Unless you specifically model ground clutter into the overall antenna model, its effects will not appear in the reported antenna performance figures and graphics.

Second, because modeled ground is flat from the antenna to the horizon (or stepped at a uniform radius or at a linear boundary), the modeling program does not take into account terrain variations that may be very important to a given antenna

situation. Both K6STI and N6BV have developed experimental software to help translate modeled results into reasonable expectations for variable terrain.

Third, the ground beneath an antenna model is uniform to an indefinitely large depth. Although this model of ground is sufficient for the upper HF region and above, at the lower HF region and below, it may not accurately portray the ground conditions for an antenna. The ground beneath an antenna is normally stratified and changes quality with depth. Lower frequency signals can penetrate the earth to considerable depths and thus be variably affected by the changing soil quality. At best, for modeling purposes, one can develop only an average single pair of quality figures for complex soils.

In addition to these limitations, we have noted that some ground modeling systems have accuracy limitations for horizontal antennas. However, we might best view these from a modeled example.

Horizontal Antennas Over Various Grounds

The following tables compare the values reported for a horizontal dipole over the various ground systems for NEC-2 and MININEC. For this test, I used EZNEC Pro in the NEC-2 mode for the NEC test, and ELNEC for the MININEC test. For each test I used a #14 copper wire dipole resonated initially in free space at 28.5 MHz. In NEC-2, the length was 16.70', and the antenna yielded 2.08 dBi gain with a source impedance of 72.6 - 0.9 Ohms. In MININEC, the length was 16.75' which yield a gain of 2.14 dBi and a source impedance of 71.6 - 0.5 Ohms. All tests over ground used average ground (conductivity = 0.005 S/m; dielectric constant = 13). Gain is in dBi, Take-off angle is in degrees, and source impedance is in the form $R \pm jX$ Ohms.

In the tables, "N-Perfect" is NEC-2, perfect ground. "N-S-N" is the NEC Sommerfeld-Norton high accuracy ground system, generally considered to be the present standard of accuracy. "N-Fast" refers to the simplified ground system available in NEC. "N-MIN" points to a feature available only in EZNEC, which is the implementation of the MININEC ground system within NEC-2. "M-Perfect" in the gain table means perfect ground in MININEC, while "M-Real" refers to the real ground system in MININEC. In the source impedance table, "MININEC" covers both, since

the source impedance is always calculated with reference to perfect ground in that system.

Gain and Elevation Angle of Maximum Radiation (Take-Off Angle) (xx/yy)

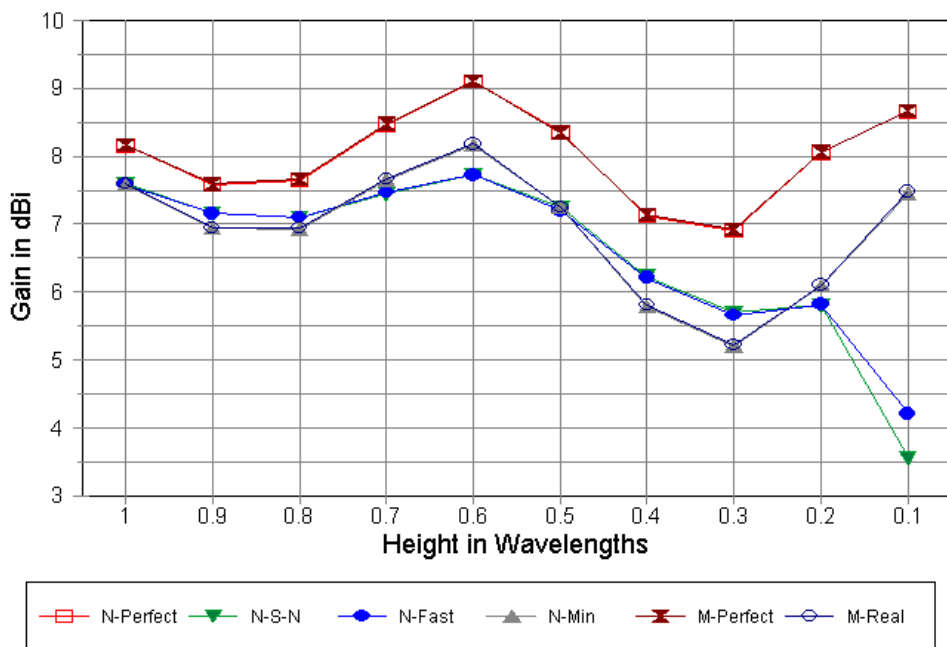
Height (wl/ft)	N-Perfect	N-S-N	N-Fast	N-Min	M-Perfect	M-Real
1.0/34.5'	8.16/14*	7.59/14	7.58/14	7.58/14	8.17/14*	7.60/14
0.9/31.1'	7.58/16	7.16/16	7.16/16	6.94/16	7.59/16	6.94/16
0.8/27.6'	7.65/18*	7.09/17	7.10/17	6.93/17	7.66/18*	6.94/17
0.7/24.2'	8.46/21	7.44/20	7.47/20	7.65/20	8.48/21	7.67/20
0.6/20.7'	9.10/25	7.73/23	7.72/23	8.17/23	9.11/25	8.18/23
0.5/17.3'	8.36/30	7.25/28	7.20/28	7.26/28	8.35/30	7.25/28
0.4/13.8'	7.13/38	6.23/35	6.21/35	5.79/35	7.14/39	5.81/35
0.3/10.4'	6.91/54	5.70/49	5.66/49	5.21/49	6.92/56	5.22/49
0.2/ 6.9'	8.05/90	5.80/90	5.82/90	6.09/90	8.07/90	6.10/90
0.1/ 3.5'	8.66/90	3.55/90	4.21/90	7.45/90	8.67/90	7.48/90

*At the indicated heights, both NEC and MININEC register the second lobe above ground to be the take-off angle (49 degrees and 70 degrees, respectively for the two heights in order). However, there is no significant difference in strength between the two lobes over perfect ground, and the lower one is shown here for tabular consistency.

The maximum gain figures are charted in the **Fig. 11-3** to save a bit of time in tracking the tables. (Detailed tracking is well advised to become familiar with the minor differences in expectations for reported figures.). The variations in gain with antenna height are in general quite natural. They can be tracked against changes in the elevation patterns for the antenna at each height. When we move from the most general level to specifics plotted in the graph, however, certain changes take on significance in terms of modeling accuracy.

Gain vs. Decreasing Height
1/2 WL Horizontal Dipole

Fig. 11.3



The most significant portion of the graph is to the far right. Contrast the rise in reported gain from using the MININEC ground (either in MININEC itself or within the EZNEC version of NEC-2) with the gain decreases reported by the two NEC ground systems. Note also that the NEC “fast” ground reports a higher gain than the Sommerfeld-Norton system. Even though the MININEC ground reports parallel the reports over perfect ground, they are quite untrustworthy over real ground at these low heights.

There are a few less evident anomalies between ground systems. MININEC and NEC perfect ground lines are in excellent accord, as are the S-N and “fast” ground reports above 0.3 wavelength antenna height. However, NEC shows less variation in gain reports over the more reliable span of the graph than does the MININEC ground, whichever system uses it.

Source Impedance ($R \pm jX$ Ohms)

Height (wl/ft)	N-Perfect	N-S-N	N-Fast	N-Min	MININEC
1.0/34.5'	71.0-10.0	71.1-6.0	71.3-6.0	71.0-10.0	70.7-9.8
0.9/31.1'	81.4-6.0	77.3-4.4	77.4-4.2	81.4-6.0	81.1-5.7
0.8/27.6'	81.1+6.8	78.0+2.9	77.8+3.2	81.1+6.8	80.9+6.9
0.7/24.2'	67.6+11.2	70.7+6.4	70.3+6.2	67.6+11.2	67.3+11.5
0.6/20.7'	57.6-2.0	63.9-0.3	64.0-0.9	57.6-2.0	57.3-1.5
0.5/17.3'	67.4-17.8	68.1-10.1	69.0-10.3	67.4-17.8	67.2-17.4
0.4/13.8'	89.8-14.0	81.6-8.8	81.9-9.0	89.8-14.0	89.2-13.9
0.3/10.4'	96.8+12.7	86.0+5.4	86.7+5.8	96.8+12.7	96.5+13.0
0.2/ 6.9'	68.6+36.8	72.1+17.8	71.9+20.1	68.6+36.8	68.8+37.1
0.1/ 3.5'	23.0+20.9	55.7+9.3	48.1+13.7	23.0+20.9	22.4+21.0

The gain inaccuracies at low heights reported by the NEC “fast” ground and the MININEC standard ground are accompanied by inaccuracies in reported source (feedpoint) impedance. Like the gain figures, MININEC figures are more extremely in error than those of the NEC “fast” ground, but both are untrustworthy.

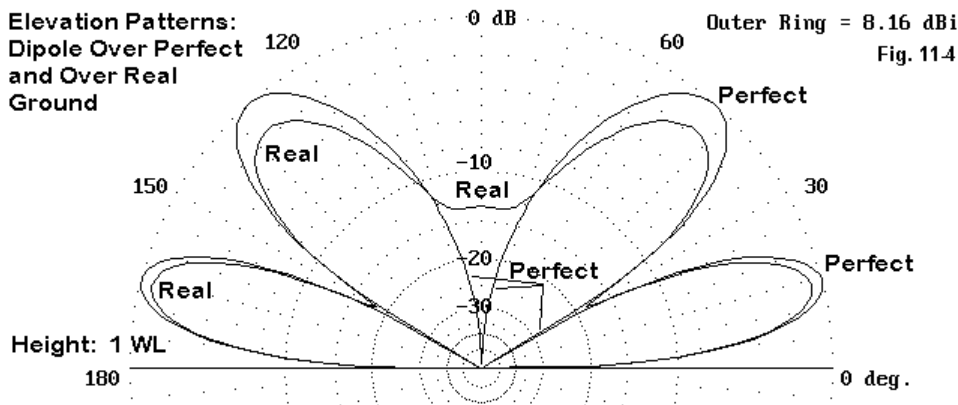
Although I tend to distrust both the MININEC and NEC “fast” ground system below 0.25-wavelength horizontal antenna height, standard recommendations set a limit of about 0.2 wavelength. For reference, here is a chart of heights (in feet and meters) corresponding to 0.2 wavelength for selected frequencies in the amateur bands from 160 to 6 meters.

Frequency	0.2 Wavelength		Frequency	0.2 Wavelength	
MHz	Feet	Meters	MHz	Feet	Meters
1.875	104.9	32.0	18.1	10.9	3.31
3.5	56.2	17.1	21.1	9.32	2.84
4.0	49.2	15.0	24.9	7.90	2.41
7.1	27.7	8.45	28.1	7.00	2.13

10.1	19.5	5.94	29.5	6.67	2.03
14.1	14.0	4.25	52.0	3.78	1.15

Getting Used to Pattern Differences Over Different Grounds

In the early stages of antenna modeling, it pays to explore systematically the ranges of program parameters using simple antennas, such as the sample dipole we used to show the similarities and differences in gain and source impedance reports at various antenna heights. In some cases, you will want to examine tabular data; in other cases, graphical patterns will best illustrate differences.



The first major difference with which to familiarize yourself is how the pattern changes between using perfect ground and using any of the real grounds. **Fig. 11-4** shows a sample for a dipole 1 wavelength up between modeling it over perfect ground and over real ground. The most striking feature, of course, is the higher gain over perfect ground. As well, note the depth of the nulls in the pattern over perfect ground and the shallower nulls over real ground. Straight up, the antenna over real ground has significant radiation, in contrast to the antenna over perfect ground. Note also that the two lobes over perfect ground are very close to equal strength, whereas over real ground, the second lobe is significantly reduced in strength.

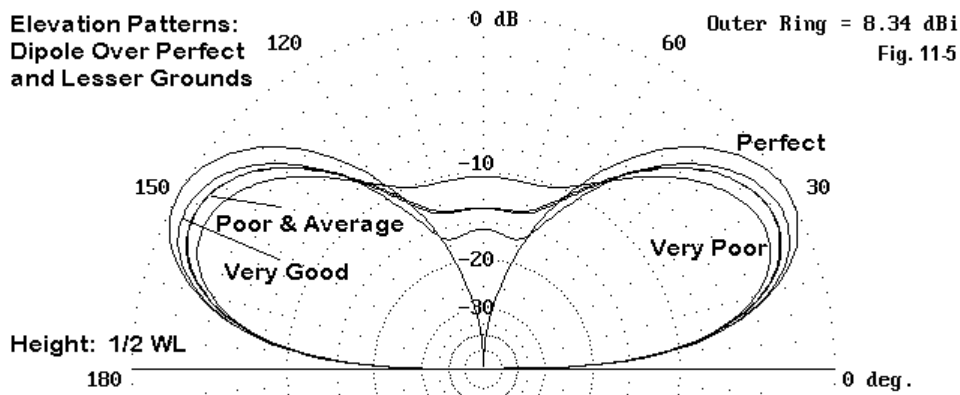
A second set of differences appears if we model a horizontal antenna over various qualities of ground. Although some routinely assert that the ground quality makes virtually no difference for a horizontal antenna, this statement is only relatively true at antenna heights approaching and above 1 wavelength. At lower heights the ground can begin to make a notable numerical difference. Whether that numerical difference makes a significant operational difference requires the modeler to examine the antenna within its intended framework of use.

Suppose we place our 10-meter dipole at heights of 1 wavelength and 1/2 wavelength above ground. Then let us systematically change the ground quality through a few standard variations, as defined earlier in this column. Gain figures are in dBi, while the Take-off angle is in degrees.

Ground Quality	1-Wavelength Height Gain/TO Angle	1/2-Wavelength Height Gain/TO Angle
Perfect	8.16/14	8.34/30
Very Good	7.83/14	7.67/28
Average	7.58/14	7.18/28
Poor	7.56/14	7.13/28
Very Poor	7.16/14	6.43/27
VG to VP gain Difference	0.67 dB	1.24 dB

Increasing the antenna height by double almost halves the gain differential between very good and very poor soils. At low heights, as shown in the elevation patterns in **Fig. 11-5**, typical of horizontal antennas for the lower HF bands, soil quality can make a difference in the potential performance of a horizontal antenna. Above 1 wavelength for the antenna height, the difference becomes ever less significant.

Elevation Patterns:
Dipole Over Perfect
and Lesser Grounds



More to Come

We have looked at the general nature of modeled ground systems, as well as at some of the differences among ground qualities. Still, our look has been limited: we have only used horizontal antennas to draw out some basic distinctions. Everything appears to be quite systematic. Placing horizontal elements close to ground begins to exceed the accuracy limitations of some of the ground modeling systems. Ground quality seems to affect gain—and to a lesser extent, take-off angle—in a regular fashion, with poorer grounds reducing gain more than better grounds.

However, there is another entire class of antennas with which we must contend: the vertically polarized antennas. We should not expect from this unruly antenna group the well-behaved figures we obtained for horizontally polarized antennas. Making sense of modeling verticals at, near, and well above ground will be our task in a forthcoming episode.

* * * * *

Model included: 11-1

12. Verticals At and Over Ground

Last month, we looked at the ground systems available in both MININEC and NEC and at the typical behavior of horizontal antennas over the selected ground system. Horizontal antennas exhibited some interesting general properties. The source impedance of the antenna varied considerably up and down, relative to the free space modeled source resistance of the same antenna, from close to ground to heights approaching or exceeding 1 wavelength. However, gain and the emergence of vertical lobes in the antenna pattern proceeded in a very regular manner. Soil quality beneath the antenna played a very small role in the reported performance of either MININEC or NEC models.

The modeled performance of vertical antennas at or above ground system available in MININEC and NEC shows almost diametrically opposed characteristics. For example, the source impedance will vary by only a small amount, and above about 0.3 wavelength at the bottom of the antenna, hardly any variation at all. In contrast, the antenna gain and its vertical pattern lobes will vary widely and not altogether according to simplistic expectations as we place a vertical antenna at various heights over soils of differing quality.

Apart from any problems associated with the possible accuracy or inaccuracy of models of vertical antennas above ground, the modeler needs to become acclimated to the range of variations that are typical for vertical antennas. This episode is designed to begin that process, with the hope that modelers will carry on with self-generated series of models of antennas and soil types most relevant to their primary activities.

A Short Review of Ground and Ground System Basics

Although we shall give some attention to perfect ground, primary focus will be upon the real ground systems available in modeling programs. In MININEC, there is only one real ground system. In NEC, we have a choice between the so-called “fast” system and the more accurate but slower Sommerfeld-Norton (S-N) system.

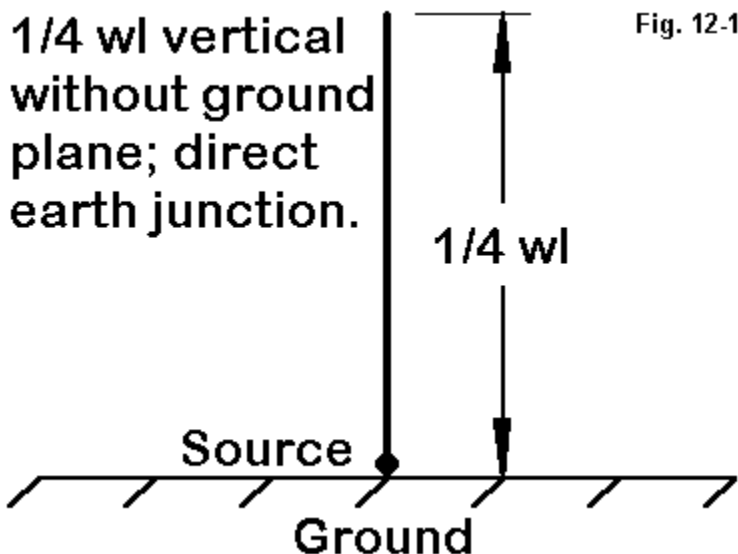
For most purposes, the S-N system is preferable, especially as computer speeds overcome speed differences in ground condition calculations.

As we did last month, we shall employ the ground or soil quality categories commonly used. A relevant portion of the chart we used last month is reproduced for more immediate reference.

Soil Description	Conductivity in S/m	Permittivity (Dielectric Constant)	Relative Quality
Salt water	5.0	81	
Pastoral, low hills, rich soil, typical from Dallas, TX, to Lincoln, NE	0.0303	20	Very Good
Pastoral, medium hills, and forestation, heavy clay soils, typical of central VA	0.005	13	Average
Rocky soil, steep hills, typically mountainous	0.002	12-14	Poor
Cities, industrial areas	0.001	5	Very Poor

Except for the salt-water entry, only the categories we shall use in our sample models are shown here. In calculating ground effects, modeling programs create from the values inserted for the conductivity and the dielectric constant a composite value. Remember that all MININEC and NEC ground calculations assume level terrain without obstructions other than those modeled. Additionally, they presuppose a homogenous soil beneath an antenna and to the horizon, rather than the sort of stratified soil layers we commonly encounter in practice. The effects of soil stratification are more pronounced at lower frequencies, where RF tends to penetrate more deeply into the ground, than at upper HF frequencies and above.

The Simple No-Ground-Plane 1/4-Wavelength Vertical



One typical model employed by nearly all modelers at one or another time in their work is the 1/4 wavelength vertical antenna terminated at the ground, as shown in **Fig. 12-1**. The test antenna is 1" diameter aluminum. No ground plane is used in the model. The source for a MININEC model can be placed at the junction of the wire end and the ground. For a NEC model, the source must be placed on the first segment relative to the ground end of the antenna.

When such antennas are placed over perfect ground, the program creates an image antenna for calculations. Whether in MININEC or NEC, the results are virtually indistinguishable.

Program	Gain	Source Impedance
	DBi	R +/- jX Ohms
MININEC	5.14	35.86 - j 0.75
NEC	5.14	36.01 - j 0.14

The very slight difference in source impedance stems from two sources. First, the source on the MININEC model is at the end of the wire, while in the NEC model, it is inboard, on the first of the 20 segments in this 7.15 MHz model. Second, the required length for resonance (as arbitrarily set to a source reactance limit of +/- 1 Ohm) is 397" in MININEC and 396" in NEC.

When we place these antennas over real grounds of the types available, we obtain far more disparate results. The following table is an illustration. VG is very good, Ave is average, P is poor, and VP is very poor, as defined in the table of ground quality types shown earlier and using the corresponding values of conductivity and dielectric constant in the models.

System	Soil Quality	Gain dBi	TO Angle degrees	Source Impedance R +/- jX Ohms
MININEC	VG	1.93	21	35.85 - j 0.75
Real	Ave	-0.05	26	35.85 - j 0.75
	P	-0.29	27	35.85 - j 0.75
	VP	-1.77	29	35.85 - j 0.75
NEC	VG	2.45	21	31.98 - j 0.95
Fast	Ave	0.97	26	28.61 + j 0.43
	P	0.82	27	28.11 + j 1.73
	VP	-0.18	28	25.14 + j 2.20
NEC	VG	0.88	21	45.85 + j 5.41
S-N	Ave	-1.23	26	47.13 + j 3.65
	P	-1.37	27	46.12 + j 4.09
	VP	-2.37	29	41.23 + j 2.13

Although the elevation angles of maximum radiation (TO angles) show excellent coincidence as we move from system to system, little else does. So let's graph some of the results and see what they might tell us.

Gain Curves: NEC & MININEC 1/4 WL Vertical, No Ground Plane

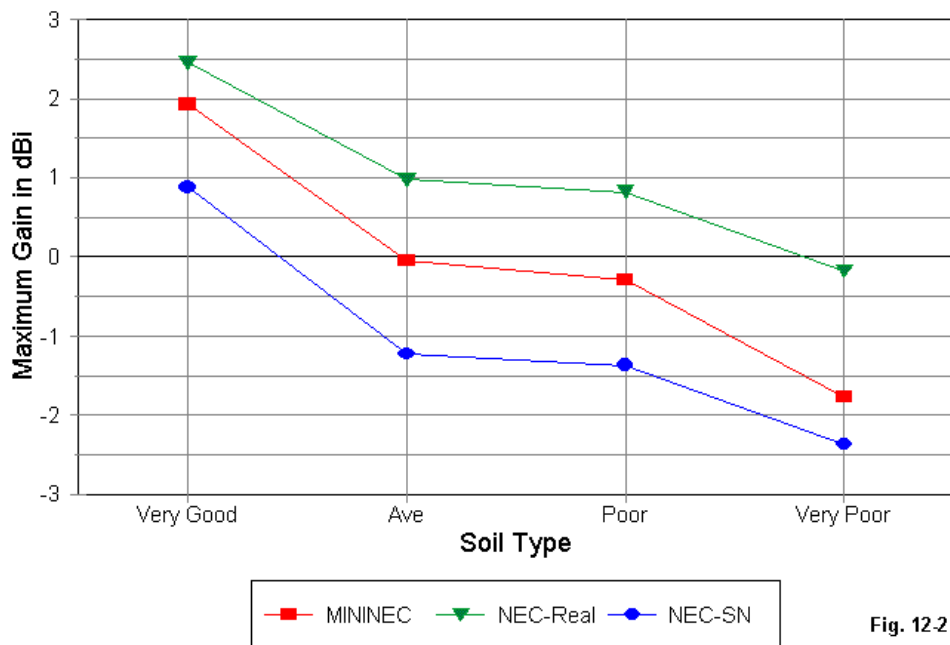


Fig. 12.2

Fig. 12-2 graphs the reported gain of the antenna over the different soils. The intervals among soil types is not regular, and therefore, the lines should be read solely as connectors to keep the families of values well sorted. Each system shows comparable gain differences as we move from soil type to soil type. However, the reported values are distinctly different for any given soil type.

Source Resistance: NEC & MININEC 1/4 WL Vertical, No Ground Plane

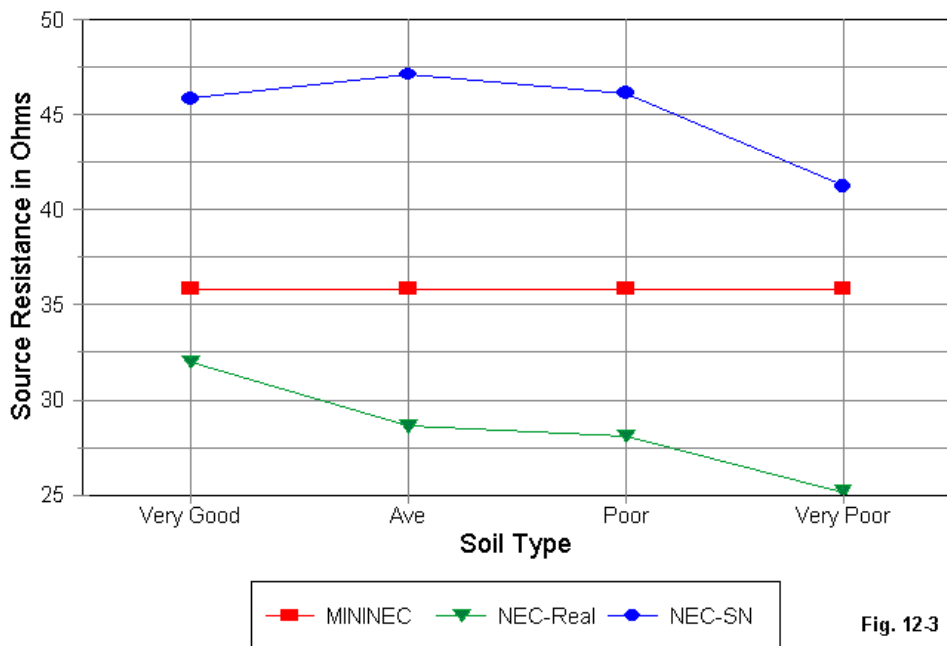


Fig. 12.3

In **Fig. 12-3**, we can view the reported source impedance values for the three systems. The MININEC line is flat because all MININEC source impedance values are calculated from perfect ground. Hence, they do not register any differences occasioned by changes in soil quality. In contrast, the NEC systems provide source impedance values quite distinct from the MININEC value: the “fast” system reporting consistently lower values and the S-N system reporting consistency higher val-

ues. However, the NEC progressions are not consistent with each other with respect either to value or to the progression of values from one soil type to the next.

In general, connecting an antenna model to a NEC fast ground is considered the least accurate of all possibilities. The reported gain values are unrealistically high and the source impedance values unrealistically low for connection to a lossy medium like the earth. In contrast, the values for the S-N ground system yield lower gains and higher source impedance reports, giving them an air of realism.

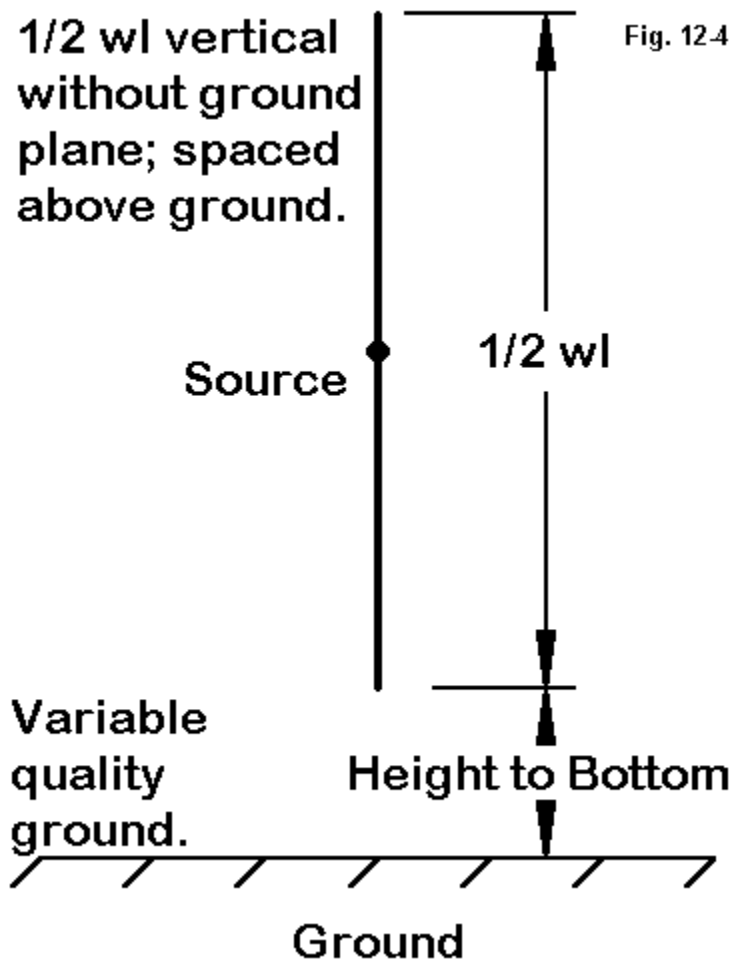
However, the sense of realism may have limited value, since virtually no real 1/4-wavelength antenna is directly connected to ground without a ground plane. Even were we to try to design such an antenna, at lower HF frequencies, there would be numerous variables of soil penetration and stratification to make the modeling results of dubious reliability. Moreover, I know of no definitive test results that would form a benchmark for assessing the reported values of either gain or source impedance.

In most instances, modeling wires directly connected to ground is performed with a perfect ground and is used as a shortcut to modeling symmetrical antennas in free space. Alternatively, some preliminary modeling of phased and parasitic vertical antenna systems is done over perfect ground. Although the results are not especially accurate with respect to actual maximum gain, other factors—such as pattern shape and relative improvements—prove useful.

So why have we spent so much space and time on modeling the 1/4-wavelength antenna touching various grounds? The purpose has been to alert you to what you can expect to see should you attempt such models, whatever the goal and rationale.

A More Realistic 1/2-Wavelength Vertical Dipole Above Ground

We can develop more sophisticated expectations of what to expect by way of reports from models of vertical antennas by exploring the modeled behavior of a 1/2 wavelength vertical dipole at various heights above ground and over differing soil types. **Fig. 12-4** sets the parameters of the exercise.



Because we wish to maximize accuracy and to watch the variations of source impedance with changes in height, we shall confine ourselves to NEC and the S-N ground system. We shall employ a 7.15 MHz 1" diameter aluminum dipole 792" long with 41 segments and fed at the center. For ease of tracking, we shall record the height of the antenna bottom limit as 0.2 through 1.0 wavelength up. For reference, the following table records actual bottom and top heights of the antenna.

Height in wavelengths	Bottom Height "	Top Height "
0.2	330	1122
0.4	660	1452
0.6	990	1782
0.8	1320	2112
1.0	1650	2442

Since our exercise is aimed at eliciting some of the features of modeling programs and not at replicating an actual antenna, it matters little that constructing an antenna meeting the above height conditions is not feasible for most builders. If we run this antenna through the various heights and over the soils we have previously defined, we get a fascinating table of results.

Soil Type	Height wl	Gain dBi	TO Angle degrees	Source Impedance R +/- j X Ohms
VG	0.2	1.75	11	67.97 - j 4.01
	0.4	3.02	42	72.54 + j 1.55
	0.6	4.60	33	72.33 - j 1.79
	0.8	5.25	27	71.26 - j 0.14
	1.0	5.37	22	72.48 - j 0.59
Ave	0.2	0.32	14	68.46 - j 2.93
	0.4	1.84	41	72.59 + j 1.01
	0.6	3.14	32	72.13 - j 1.55
	0.8	3.52	26	71.45 - j 0.20
	1.0	3.44	22	72.35 - j 0.64

P	0.2	1.09	14	68.88 - j 2.80
	0.4	1.29	42	72.46 + j 0.86
	0.6	2.58	33	72.15 - j 1.44
	0.8	2.91	27	71.49 - j 0.26
	1.0	3.21	9	72.30 - j 0.62
VP	0.2	1.04	17	69.37 - j 1.91
	0.4	2.03	15	72.49 + j 0.41
	0.6	3.08	13	71.99 - j 1.24
	0.8	3.87	11	71.65 - j 0.31
	1.0	4.43	10	72.19 - j 0.65

The significance of the values may not be immediately apparent in tabular form, so let us take them by categories in more graphical form in **Fig. 12-5**. The figure graphs the gain values report by NEC over S-N ground for the antenna through its height variations, using separate lines for each soil type. Note that, although certain general progressions are clear with respect to maximum gain, there are great irregularities in the graph as well. Especially notable is the apparent good performance of the antenna over very poor soil when compared to the performance of all but the antenna over very good soil. Indeed, below the level of very good soil, the reported gain of the antenna shows a different “leading soil” as we change antenna heights.

This phenomenon is not restricted to the particular antenna type used in the example. Similar results accrue to most vertically polarized antennas that do not require a ground plane. For example, side-fed delta loops, rectangles, half squares, and bobtail curtains all display the sort of variability with respect to maximum gain shown in the graph for the 1/2-wavelength vertical dipole.

Reported Gain: NEC & S-N Ground 1/2 WL Vertical Above Ground

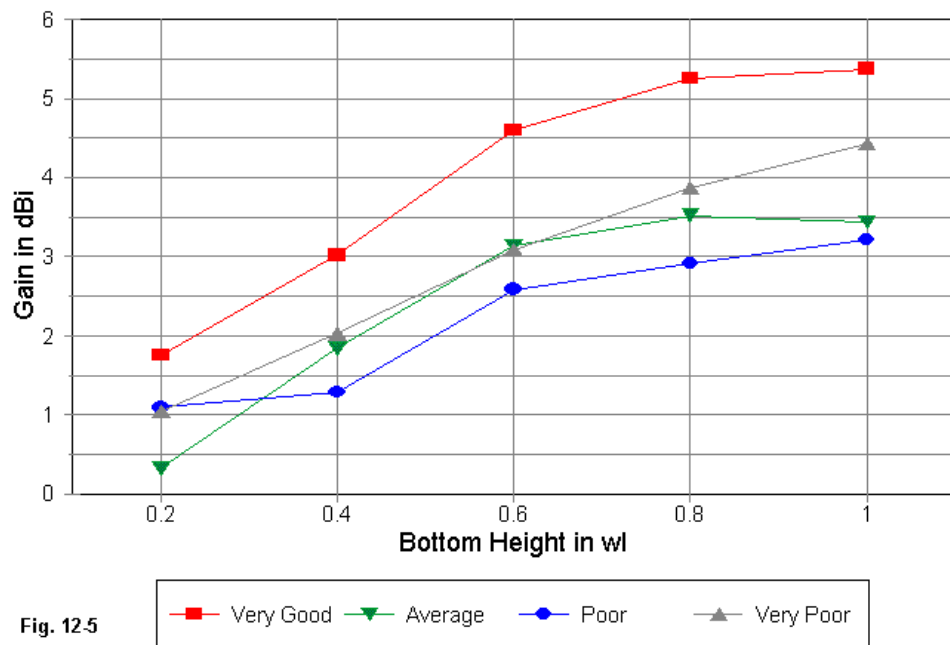
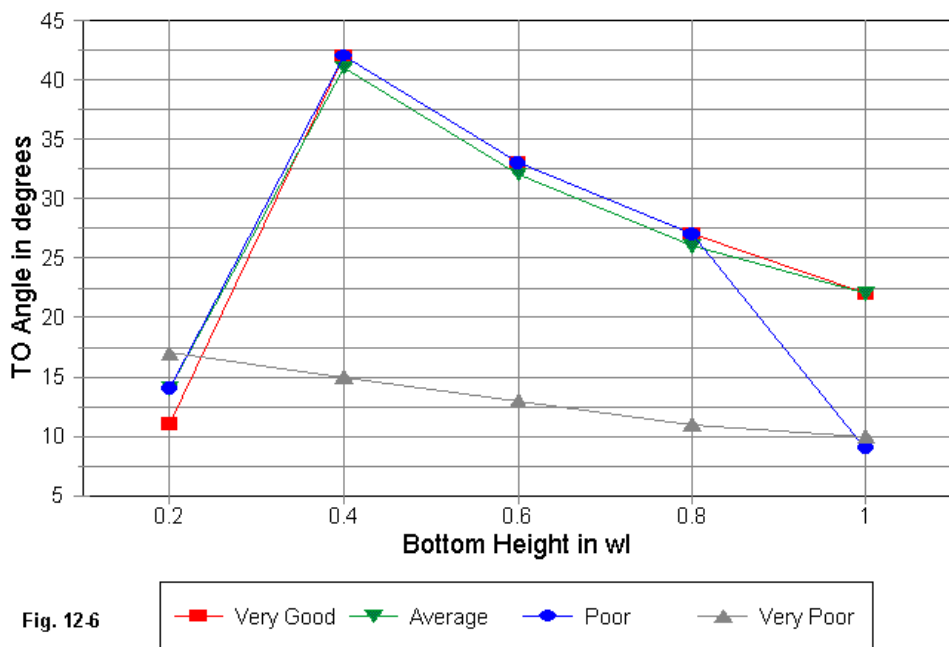


Fig. 12-5

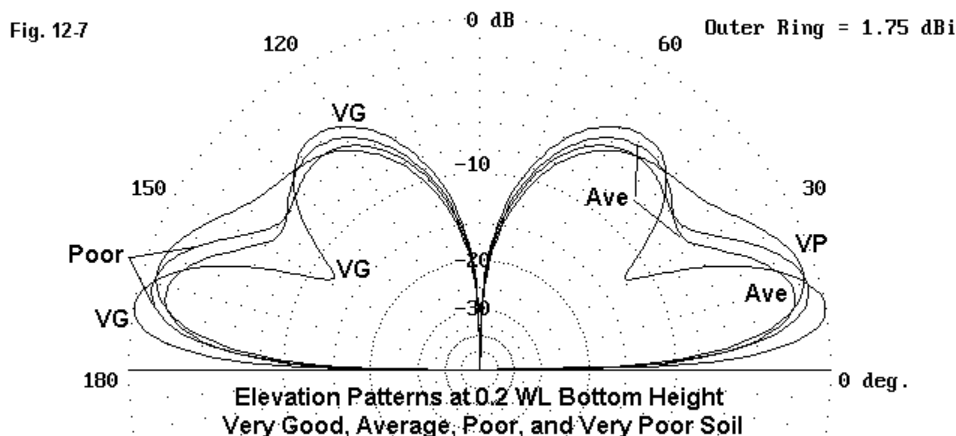
Part of the reason for the gain differentials and irregularities stems from the variability of the elevation angle of maximum radiation, which **Fig. 12-6** demonstrates. Only the antenna over very poor soil exhibits a consistently low TO angle. Note that the antenna, when over poor soil, returns to a low TO angle at the maximum test height at which the bottom of the antenna is 1.0-wavelength up.

Reported TO Angle: NEC & S-N Ground 1/2 WL Vertical Above Ground

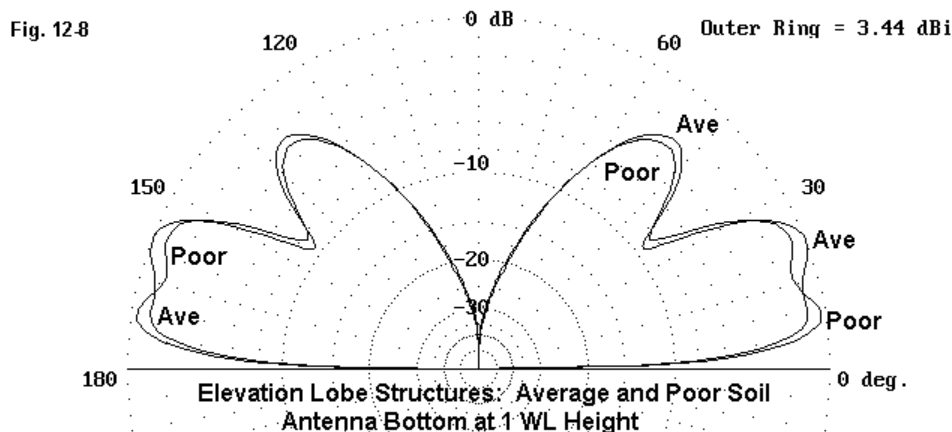


To a degree, graphical results can be misleading and must be supplemented by comparing elevation patterns in order to see and settle many questions about the performance of the antenna over different qualities of soil.

Let's begin with the height at which the antennas show a consistently low TO angle: 0.2-wavelength.



In **Fig 12-7**, the elevation patterns for all 4 soil types are overlaid, with some indicators of which pattern is which. Of immediate importance is the fact that only over very good soil does the antenna pattern show a very marked null between the lower and upper vertical lobes of the pattern. In general, the poorer the soil, the shallower the null between the lobes.



With only this much of a clue, we are in a better position to understand the high-low TO angle question. According to the table and the graphs, when the antenna is at a bottom height of 1 wavelength, it reports a TO angle of 22 degrees over average soil and 9 degrees over poor soil. What the elevation pattern in **Fig. 12-8** shows us is that there is actually little to distinguish the two patterns. Over poor soil, the lower part of the combined (almost null-less) two lower lobes is less than 1 dB stronger than the upper part. In contrast, over average soil, the upper part of the combined 2 lower lobes is about 1 dB stronger than the lower part. The emergent third lobe at an angle of nearly 50 degrees is very similar for both antenna models.

Soils ranging from poor (cond. = 0.002, d.c. = 13) to good (cond. = 0.01, d.c. = 14) can be viewed as middle range soils. They tend to exhibit in models a combined wave front that consists of the two lowest lobes with only an indistinct null between them.

Fig. 12-9

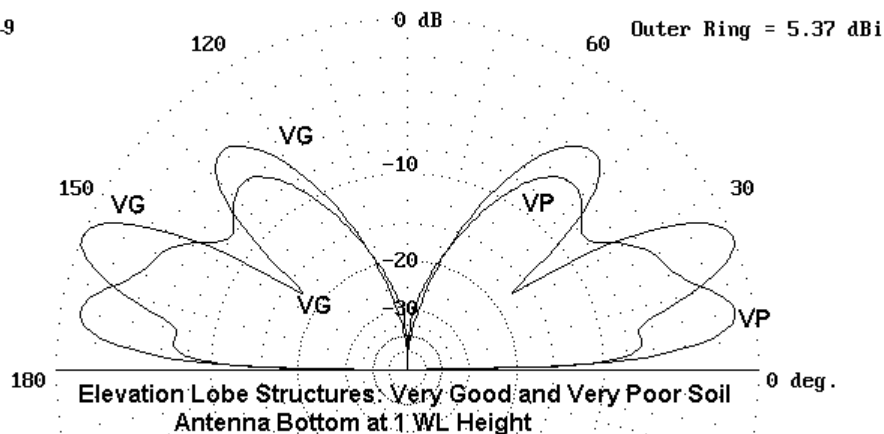
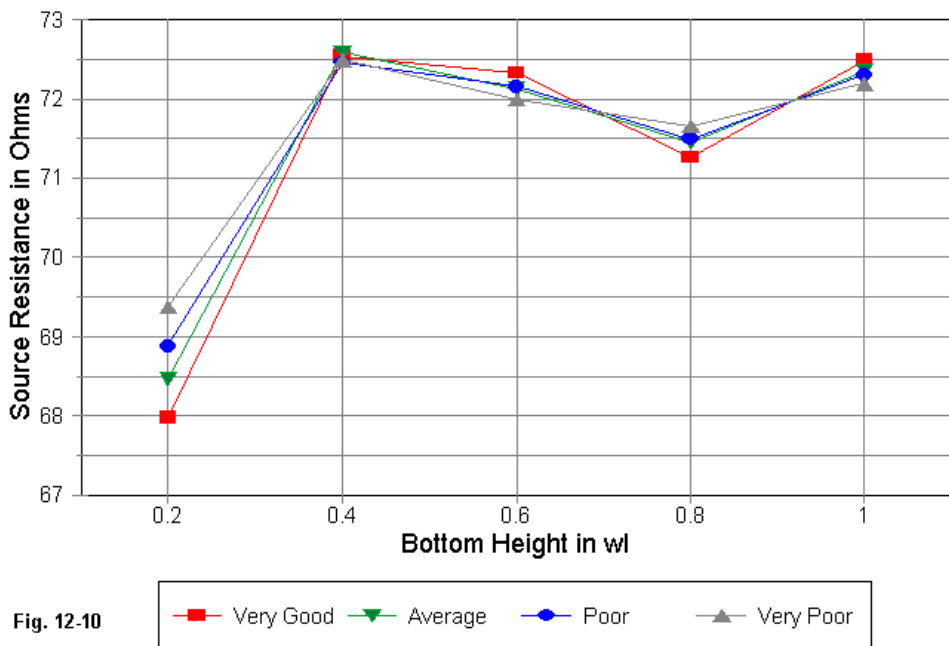


Fig. 12-9 shows the same antenna at the same height, but over very good and over very poor soil. Over very good soil, the lowest lobe is nearly missing, with only the second and third lobes prominent. In contrast, over very poor soil, the upper lobes become indistinctly differentiated, and the lowest lobe is most prominent. Even though the maximum gain of the antenna over very poor soil is not quite as strong

as the maximum gain of the antenna over very good soil, the 7-dB advantage at the lowest angles (in the region of 10 degrees elevation) may be more useful to certain communications needs.

In all of this, we have not mentioned the source impedance of the antenna under any of its conditions of height and soil. **Fig. 12-10** tells us why.

Reported Source R: NEC & S-N Ground 1/2 WL Vertical Above Ground



The graph shows that for any given height, the difference in source resistance is insignificant. The total variation in source resistance for all antenna models under all conditions is under 5 Ohms. Likewise, the total variation in source reactance is about 7-8 Ohms for all cases. With 1/2-wavelength vertical antennas at various heights, the source impedance will be no clue to the type of soil beneath the antenna. More likely candidates as good clues, if the models are at all accurate, will be the antenna's gain and its elevation angle of maximum radiation.

How Accurate Are the Models?

To some degree, there is disagreement about the accuracy of models of the type used in our example relative to real antennas set up under the prescribed conditions. However, disagreement turns to agreement the higher the antenna is placed above ground. Hence, we may consider all but the lowest antenna placements as reflecting real performance.

However, we must once again note the restrictions of even the S-N ground system in its assumptions of level terrain and homogenous soil. At lower VLF through lower HF, soil penetration is great and the nature of the complex stratification that often occurs may alter the performance of a real antenna relative to a given model. With vertical antennas, such as the one used in these examples, if translated into a real structure, modeling for a variety of conditions becomes one route to anticipating the possibilities and to analyzing the actual performance results. Indeed, the goal of modeling is sometimes not so much to exactly replicate a given antenna as it is to anticipate possibilities so that we know what they mean and can take appropriate actions, if the situation calls for any.

A Quick Note on Salt Water

The table at the beginning of this column retained the conductivity and dielectric constant for salt water. The conductivity value is at least two orders of magnitude higher than even very good soil, and the dielectric constant is 4 times that of very good soil. For most purposes, these values are practical limits to the ability of the S-N ground system to yield accurate results.

When the far field has primarily a salt-water reflective ground, the performance of vertical dipoles such as the one in our example is significantly enhanced with

respect both to gain and to TO angle. WWVH uses (as of 1999) vertical dipoles, with the Pacific Ocean as their reflective ground. Boats and ships often use simple vertical antennas, some fairly short for their frequency of operation.

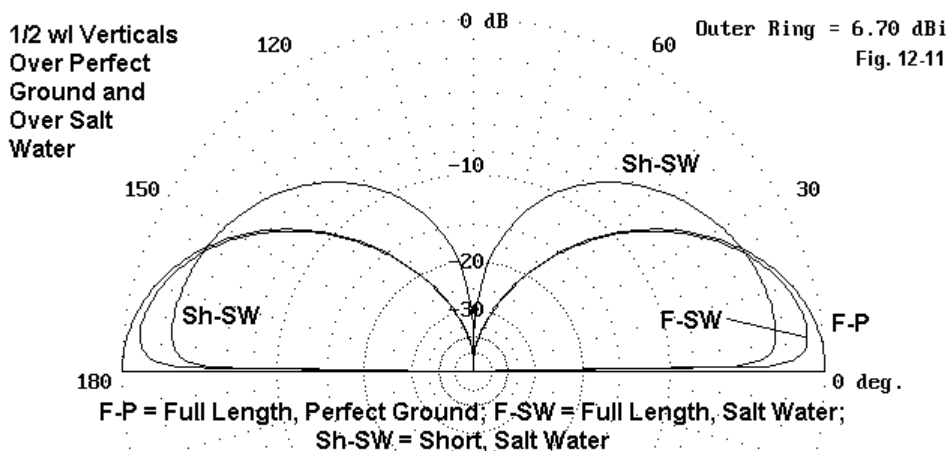


Fig. 12-11 gives us an illustration of the modeled performance of our vertical dipole over perfect ground and over salt water, with the bottom only 5" above the surface. In contrast to our land models, the salt-water dipole loses little: its reported gain is less than 1 dB less than over perfect ground, and the elevation angle of maximum radiation is 7 degrees. The pattern for the short dipole also shown in the figure is about 2.5 dB down from the full-size dipole over perfect ground, and its TO angle is 9 degrees. This performance report comes from an antenna only 14' long, in contrast to the 66' dipole that is resonant at the 7.15 MHz test frequency. The base height of only 5" is a somewhat extreme case, but demonstrates the potential effect of salt water as a reflective ground for vertical antennas.

Using salt water as the ground for our final examples is just a way of suggesting that these notes have only scratched the surface of the ways in which the selection of a ground system and a soil type may interact with vertical antennas to yield reports whose features form quite different regularities, compared to those we anti-

pate for horizontal antennas. Familiarizing yourself with the possibilities and potentials can go far toward allowing you to correctly interpret the modeling reports you get. When you do not have a pressing modeling project to perform, you might consider taking a variety of vertically polarized antennas and making systematic runs over many soil types. The results will be both illuminating and useful when a serious vertical antenna project comes along.

* * * * *

Models included: 12-1 through 12-2.



13. Notes on Reactive Antenna Loads and Their NEC Models

A. Some Center Loading Basics

Although the basics of changing the electrical length of an antenna wire by adding a reactive load to it are seemingly well-known, the precise relationship between real loads provided to real antennas and modeled loads provided to modeled antennas is not so certain as one might think. Therefore, this series of notes will try to look at some of the ins and outs of the matter—all without pretending either to authoritativeness or completeness. It will be enough if we can get a better handhold on the subject.

For antennas up to (but not including) $1/2$ wavelength long, the source impedance is capacitive. To bring the antenna to resonance, we electrically lengthen it by adding an inductive reactance. We can add the reactance at the element center, or we may split the load into two equal parts and place them farther out on the element. The inductive reactance needed to bring an element to resonance increases the farther out along the element we place the reactances.

Since every method of implementing an inductive reactance has some resistive loss, the reactive device has a Q, a ratio of inductive reactance to resistance. The higher the Q, the lower the losses. The most common two means of providing an antenna in the HF region with an inductive reactance are the air-wound solenoid inductor and the shorted transmission line stub less than $1/4$ -wavelength long. Both are physical entities having both basic and functional loss mechanisms. For example, solenoid wire for any given RF frequency has an AC resistance that is a function of skin effect. In addition, the interaction of the individual turns of the coil may yield other losses. Whatever the source of the loss, the solenoids Q or figure of merit expresses them all within the resistive denominator of the ratio.

Modeling Resistance-Reactance Loads

NEC (either -2 or -4) provides several means of modeling inductively reactive loads. The most common type appears on **Fig. 13-1**.

```

CM 7.15 MHz Dipole, 1/4 wl long
CM #8 Wire, Lossless; Center-Loaded
CE
GW 1 91 -206.343 0 0 206.343 0 0 .0642519
GS 0 0 .02540
GE 0
EX 0 1 46 0 1 0
LD 4 1 46 46 0 833.3 0 Type 4 (R +/- X) Load
FR 0 7 0 0 7 .05
RP 0 360 1 1000 -90 0 1 1
FR 0 7 0 0 7 .05
RP 0 181 1 1000 -90 0 1 1
EN

```

Fig. 13-1

Type 4: Complex loads consisting of a resistance and a reactance: This is the native type of load for NEC. All loads are eventually translated into values of resistance and reactance. However, this type of load remains constant. That is to say, for the frequency sweep specified in the simple model in the figure, the values of resistance and of reactance remain the same for each frequency checked.

Type 4 loads are therefore not useful for frequency sweeps, since real solenoid components do not have the same reactance on every frequency. Reactance changes with frequency in relationship to a solenoid inductance. For most purposes in antenna work, the inductance and the resistance of the coil are treated as constants in simple inductive reactance loading problems.

Type 4 loads, however, are especially useful for initial design work. For example, if one has a short center-fed antenna, the reactive component of the source impedance will specify the absolute value of center reactive load necessary to eliminate it. The type of reactance introduced will, of course, be the opposite of the reactive component of the source impedance. Since antenna wires shorter than 1/

2-wavelength show a capacitive reactance at the source, a center reactive load must be inductive.

Type 4 loads are also very useful for setting up or exploring the effects of the load Q on antenna performance. The results will apply only at the frequency for which the values hold good. However, one can easily obtain a family of value curves for comparing various antenna parameters.

Let's set up a test dipole to see the type 4 load at work. See **Fig. 13-2**.

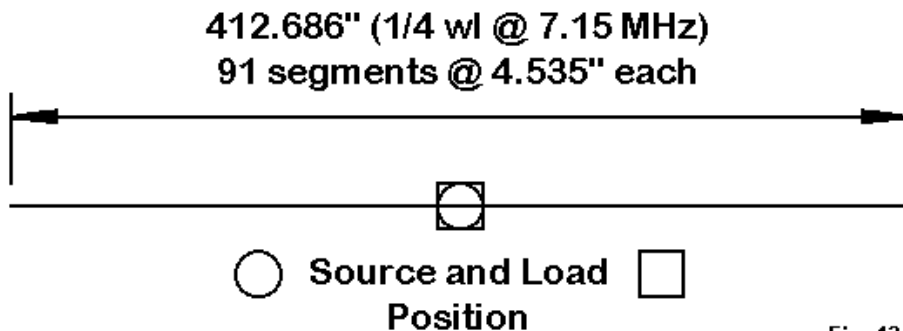


Fig. 13-2

The test dipole (model 13-1) is exactly 1/4-wavelength long at 7.15 MHz, which makes it far short of resonant. We shall introduce a reactive load at the center, in the same position as the source or feedpoint. In NEC models, the source and the load are in series when the modeler specified the same position or segment for both the source and the load. The effect is comparable to splitting the center-loading solenoid at the center to insert the source (normally a transmission line for operational amateur radio stations).

The test model antenna uses lossless or perfect wire so that the only losses that appear will be from the load or loads we add. It uses 91 segments for reasons that will appear in later episodes. If this little test is replicated, be sure to use the same

number of segments as in the example to obtain the same results. Antenna resonance in the modeling exercise is defined as a source impedance of less than ± 1.0 Ohm, and in most cases, the reactance will be less than ± 0.1 Ohm. Since there will be a slight drift in the source impedance as the segmentation of a given wire is changed, even for well-converged models, replication of results requires detailed replication of all of the model parameters.

We may determine the required load for the antenna simply by reading the source impedance before inserting a load. We shall use two diameters of wires for this example: #8 and #12 AWG (with diameters of 0.1285" and 0.0808", respectively). To cancel out the capacitive reactance, the #8 antenna requires a center inductive reactance of 833.3 Ohms, while the #12 wire antenna of identical length requires 889.4 Ohms. For any value of Q , the resistance is simply the reactance divided by the selected Q value. Thus, we can select a range of Q -values to check. In this case, let us check a perfect inductively reactive load ($R = 0$), along with finite Q -values of 1000, 750, 500, 250, and 100. For the two antennas—which differ only in wire diameter—we get two families of values, which the graph in **Fig. 13-3** artificially connects with an identifying line just to keep the two antennas differentiated from each other.

First, let's remember that the two antennas required two different reactive load values solely as a function of the source impedance. Both antennas use lossless wire, so the source impedance difference is a function of wire diameter. Now, if we look at the left-most line, where Q is infinite because the load resistance is zero, the antenna gain is the same for both wires. The fact that the value (1.85 dBi) is less than the traditional values for a resonant dipole (2.14-2.15 dBi) results from the length of the wire. It is much shorter than a resonant dipole. The wire shows the same gain without a load in place. Part of the gain of an antenna is a direct function of its length. In fact, you wish to obtain more gain from a dipole, make it longer—up to about 1.25-wavelength long. You may not like the resulting source impedance, but the antenna will have more gain simply by being longer—until it reaches a length where the pattern begins to degenerate into multiple lobes.

Short Dipole Gain vs. Load Q #8 and #12 Lossless Wires, 1/4 wl Long

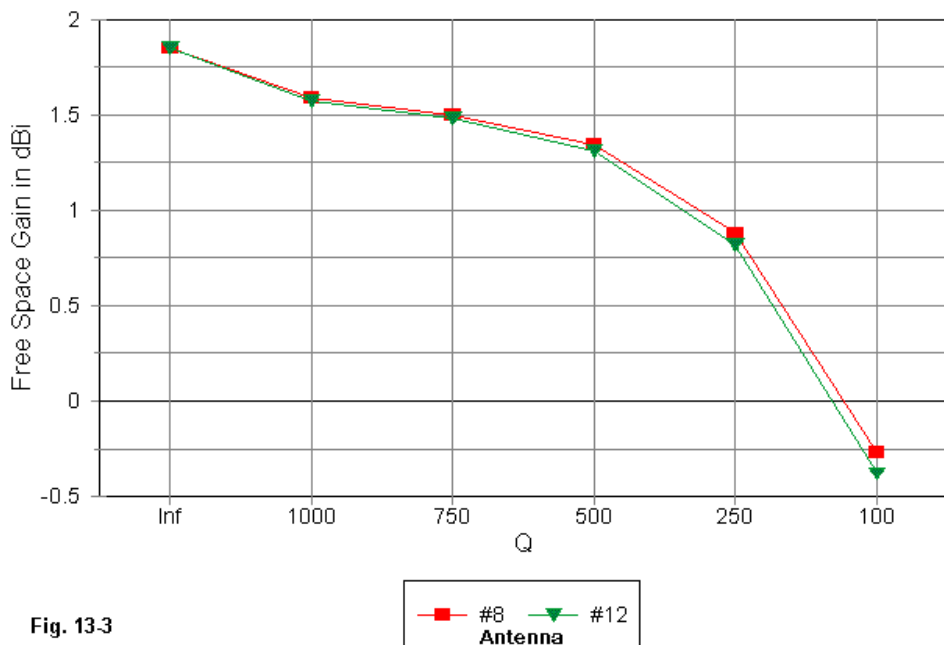


Fig. 13.3

Second, the divisions from left to right following no special pattern. Hence, the actual curve of the values is not apparent. The drop in gain from a perfect load to one with a high Q-value of 1000 is noticeable. The region between a Q of 1000 and a Q of 500 strongly suggests that the curve of gain does not track arithmetically. Indeed, the curve is much straighter if each selected Q-value is of a constant ratio with the immediately preceding value throughout the chart.

Third, the two families of values track each other, with lower gain values for the #12 wire antenna. If we remember that the necessary inductive reactance for the #12 wire antenna was higher than for the #8 antenna, then it follows that for any given Q, the resistance will also be higher. The higher the resistance of a center load, the lower the gain.

The type 4 load from which we derived the modeled data is a mathematical entity only. Whereas the wire is modeled as a physical entity, the load is only a mathematical modification of the results. Hence, any physical properties that the load may have will not play a role in its modeled performance. The length and diameter of the loading mechanism are never modeled in Type 4 loads, even if in reality they may play some kind of role in the performance of the antenna.

Modeling Series Resistance-Inductance-Capacitance Loads

Equally mathematical, that is, non-radiating, are loads of the following type:

```
CM 7.15 MHz Dipole, 1/4 wl long
CM #8 Wire, Lossless; Center-Loaded
CE
GW 1 91 -206.343 0 0 206.343 0 0 .0642519
GS 0 0 .02540
GE 0
EX 0 1 46 0 1 0
LD 0 1 46 46 .8333 .000018549 0
FR 0 7 0 0 7 .05      Type 0 (Series R-L-C) Load
RP 0 360 1 1000 -90 0 1 1
FR 0 7 0 0 7 .05
RP 0 181 1 1000 -90 0 1 1
EN
```

Fig. 13.4

Type 0: Loads consisting of the following in series: resistance, inductance, and capacitance: Missing elements are entered as values of zero (which the program automatically interprets as a missing value, not a zero value). To specify an inductor of 6 microH with a series resistance of 2 Ohms, one enters this: 2, 6E-6, 0. The final zero indicates that there is no capacitor. See **Fig. 13-4** for a sample NEC model in ASCII format using other values for R, L, and C. (See model 13-2.)

The LD line in the input file shows the load type and values for resistance and inductance, again with a missing capacitor. Another load type (Type 1) is available for parallel combinations of resistance, inductance, and capacitance. We shall not need it for this particular set of notes.

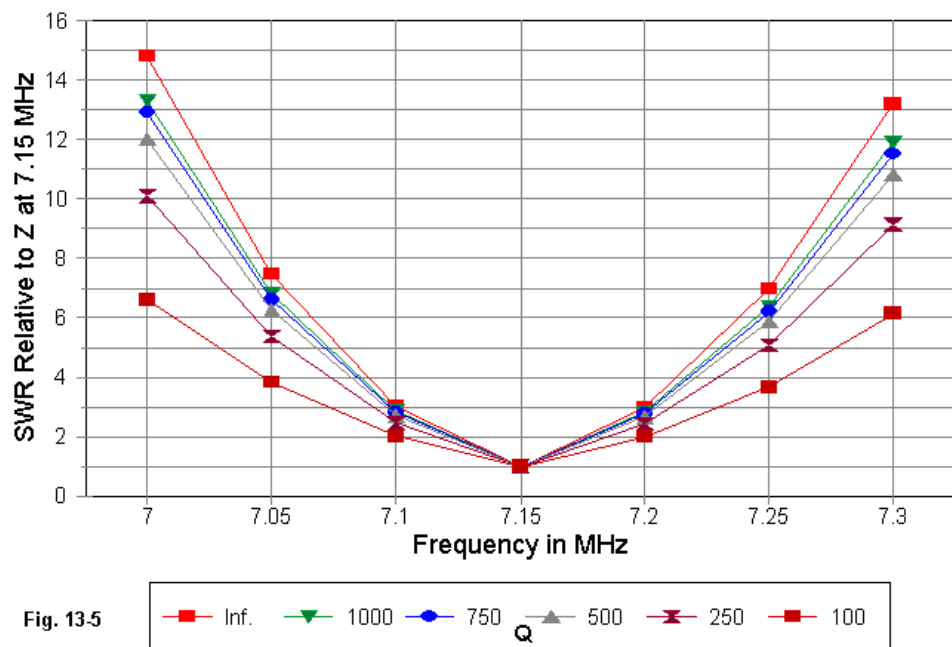
The advantage of using R-L-C loads is that NEC translates them into appropriate values of reactance for each frequency checked in a file request. Hence, with R-L-C loads, one may run a frequency sweep and have the correct reactance at each frequency is the sweep.

However, there is an accompanying limitation. The value of R remains constant. Therefore, the Q of the reactive device changes slightly with each frequency checked, since the ratio of reactance to resistance changes across the range of the frequency sweep. For limited-range frequency sweeps, such as across one of the HF amateur bands, the possible error is too small to be significant.

Of course, one needs to be able to translate reactance to inductance (or, in other cases, capacitance) and back again to set the proper values. If one begins with a value of inductance, one must find the value of reactance for the frequency in question in order to use a value of Q to set the resistance or to find the value of Q from an assigned or measured value of resistance. The standard relationship of reactance to inductance, where inductive reactance is 2π times frequency times inductance in basic units, is an essential utility tool for using R-L-C loads.

A set of frequency sweeps between 7.0 and 7.3 MHz in 0.05 MHz steps for the #8 perfect wire dipole, center-loaded at the design frequency of 7.15 MHz for resonance with an inductance of 18.549 microH will yield a family of SWR curves. For each curve, the source impedance against which the antenna impedance is compared is the resistive impedance of the antenna at resonance on the design frequency. Hence, each curve in **Fig. 13-5** shows a 1:1 VSWR at design center.

Q vs SWR: Center-Loaded Dipole 7-7.3 MHz, #8 Lossless, L=1/4 WL



The graph shows a family of such curves for a perfect inductor (zero Ohms resistance) and for inductors with Q s of 1000, 750, 500, 250, and 100. The requisite values of series resistance for the inductor are 0.8333, 1.1111, 1.6666, 3.3332, and 8.333 Ohms from 1000 down to 100. Once more, as we suspected from the gain curves earlier, the results of altering Q are more geometric than arithmetic, varying more linearly as Q forms a constant ratio from one value to the next.

Apparent in the graph is the fact that as we lower the Q of the loading coil, the operating bandwidth—set by any arbitrary selection of a VSWR ratio—becomes wider in terms of frequency. Often, we simply summarize the effect by saying that the increased losses broaden the operating bandwidth, as if the summary were an explanation. It is not.

In fact, what occurs, as modeled, is interesting in itself—and somewhat obvious if we stop to think about it. Consider the following two tables of values of source impedance for the #8 and the #12 dipoles, each center-loaded. The #12 wire antenna requires a 19.798 microH inductor with suitable adjustments in the values of the series resistor for the Q-values indicated earlier.

#8 Wire Dipole		Source Impedance (R +/- jX Ohms)					
Freq.	Q=Inf.	1000	750	500	250	100	
7.00	12.58-46.28	13.41-46.29	13.69-46.28	14.24-46.29	15.91-46.29	20.91-46.29	
7.05	12.78-30.76	13.61-30.76	13.89-30.76	14.45-30.76	16.11-30.76	21.11-30.76	
7.10	12.98-15.30	13.82-15.30	14.09-15.30	14.65-15.30	16.32-15.30	21.32-15.30	
7.15	13.19-0.02	14.02-0.02	14.30-0.02	14.86-0.02	16.52-0.02	21.52-0.02	
7.20	13.40+15.26	14.23+15.26	14.51+15.26	15.06+15.26	16.73+15.26	21.73+15.25	
7.25	13.61+30.25	14.44+30.25	14.72+30.25	15.27+30.25	16.94+30.25	21.94+30.25	
7.30	13.82+45.29	14.65+45.29	14.93+45.29	15.49+45.30	17.15+45.29	22.15+45.30	
Delta R	1.24	1.24	1.24	1.25	1.24	1.24	
Delta X	91.57	91.58	91.57	91.59	91.57	91.59	

#12 Wire Dipole		Source Impedance (R +/- jX Ohms)					
Freq.	Q=Inf.	1000	750	500	250	100	
7.00	12.61-49.30	13.50-49.29	13.79-49.29	14.39-49.29	16.16-49.29	21.51-49.30	
7.05	12.81-32.78	13.70-32.78	13.99-32.78	14.59-32.78	16.37-32.78	21.70-32.78	
7.10	13.01-16.29	13.90-16.30	14.20-16.30	14.79-16.29	16.57-16.29	21.91-16.29	
7.15	13.22-0.00	14.11-0.00	14.40-0.00	15.00-0.00	16.78-0.01	22.11-0.00	
7.20	13.43+16.31	14.32+16.31	14.61+16.31	15.21+16.31	16.98+16.31	22.32+16.31	
7.25	13.64+32.29	14.53+32.29	14.82+32.29	15.42+32.28	17.19+32.28	22.53+32.28	
7.30	13.85+48.35	14.74+48.34	15.04+48.34	15.63+48.34	17.41+48.34	2.74+48.35	
Delta R	1.24	1.24	1.25	1.24	1.25	1.24	
Delta X	97.65	97.63	97.63	97.63	97.63	97.65	

The limits of accuracy are set by the last digit in each number, plus or minus 1, as a function of the sum of rounding conventions within the program. Within those

limits, which vary only the value in the hundredths column, for each antenna, the range of variation of both resistance and reactance is invariant, regardless of the value of Q. Since Q is a function of changing R, the values for source reactance do not change at all. The values of source resistance are simply the values at infinite Q plus the series resistance of the inductive load.

In other words, the presence of a center load of finite Q does not alter the VSWR operating bandwidth curve by changing the source impedance other than adding the load resistance to the natural source resistance of the antenna without any load at all.

What the addition of the load series resistance does is to change the ratio of resistance to reactance at any of the frequencies swept, essentially raising the ratio of resistance to reactance. For the test antennas, the amount of change ranges from a fraction of an Ohm to under 10 Ohms. However, in each case, the result is a change in the impedance phase angle toward the resistive and away from the reactive.

$$VSWR = \frac{\sqrt{(R_{in} + Z_o)^2 + X_{in}^2} + \sqrt{(R_{in} - Z_o)^2 + X_{in}^2}}{\sqrt{(R_{in} + Z_o)^2 + X_{in}^2} - \sqrt{(R_{in} - Z_o)^2 + X_{in}^2}} \quad (1)$$

The basic SWR equation shows the importance of this simple adjustment in terms of a lowering of the SWR at any frequency within the sweep range, when the value of Z_o is the same as the value of R_{in} at the design center frequency. With values of X_{in} constant, higher values for R_{in} and Z_o will together lower the VSWR relative to a design center frequency value of 1:1.

Transmission Line Stub Loads

Before we turn away from the modeling of center loads with the mathematical models available in NEC, let's look at the alternative method of creating an inductively reactive load: the transmission line stub. These stubs present us with an immediate modeling problem. Modeled transmission lines appear in parallel with sources. So we must move them. Moreover, a single stub would be very long (over

20' long for a 600-Ohm line). Common practice is to split the stub into two equal parts, installing one on either side of the line at the main element feed point. Our simple modeling counterpart would be to place TL entries in NEC on each of the adjoining segments to the source segment.

(Note: Use of transmission-line loads implemented with TL entries is NOT recommended wherever the currents on the two transmission lines are unequal. On some occasions we may not know in advance that the currents are unequal. For this exercise, we are purposely violating the recommendation in order to study the consequences of trying to model load lines using TL entries.)

However, we cannot simply use the value of the center-loading reactance and split it into two pieces. If the reactance is split and moved away from the source point, the sum of the pieces will be greater than the value of the exactly centered single reactance. The center loading reactance of 833.3 Ohms is required for the #8 perfect wire dipole that is exactly 1/4-wavelength long overall in free space. Two reactances placed about 9" apart center-to-center, one on either side of the source, require 426.3 Ohms each (or 852.6 Ohms total).

With a type 4 reactive load, we can move in two loading directions. One of them is a Type 0 series R-L-C load, which requires an inductance at 7.15 MHz of 9.4892E-6 H on each side of the source (model 13-3). The other direction is a pair of transmission line stubs each side of center (model 13-4). Since the models of the stubs are mathematical only and a form of a NEC network, they will not show any interaction, loss, or factors other than reactance. See **Fig. 13-6**.

The reactance of a shorted stub is equal to the characteristic impedance of the transmission line times the tangent of the electrical length of the stub in degrees or radians. Since we know that the stubs will each show 426.3 Ohms reactance, we can work backwards through the relationship to obtain the electrical length for a variety of common lines. Then we can obtain the physical length, since we know that a wavelength at the design frequency is 137.5624' long. For this exercise, we shall look at several common transmission lines with characteristic impedances of 50, 75, 300, 450, and 600 Ohms. We shall let the velocity factor be 1.00, since many of the lines offer options and variability. We have cut off the characteristic impedance at an upper limit of 600 Ohms, since with #8 wire, even 700-Ohm line

requires over 22" of spacing without significantly altering the 600-Ohm values across the band.

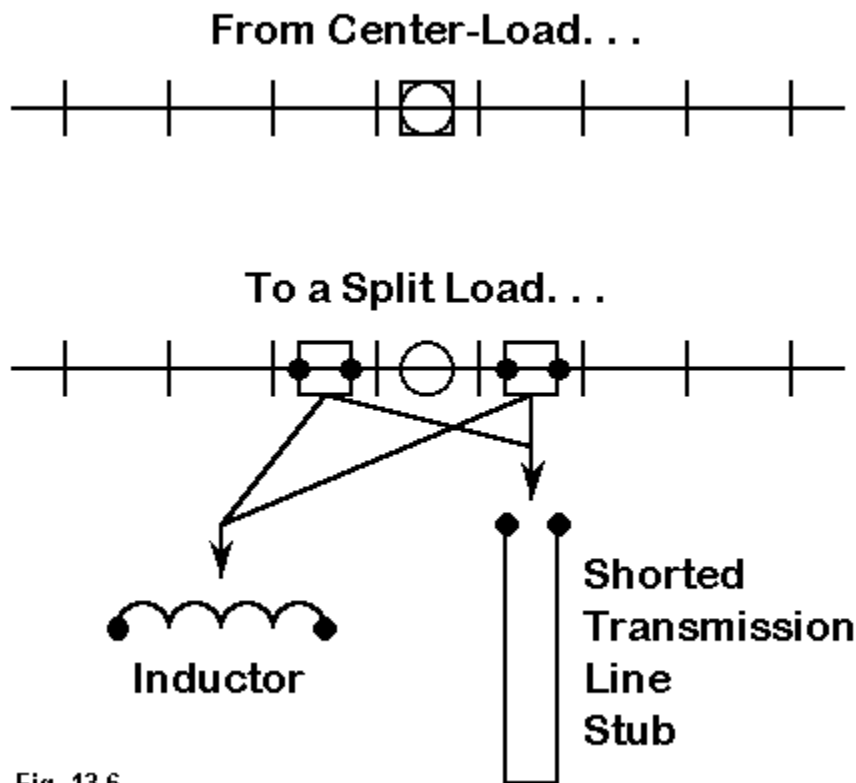


Fig. 13-6

Let's first determine the VSWR for a pair of perfect solenoid coils and then for lines of 300, 450, and 600 Ohms across the 40-meter band. Each lossless coil will

have a 9.4892 microH inductance. The line lengths required by each of the lines (including the ungraphed 50 and 75 Ohm lines) will be as follows:

Line Zo	50	75	300	450	600
L (degrees)	83.31	80.02	54.86	43.45	35.39
L (feet)	31.83	30.58	20.96	16.60	13.52

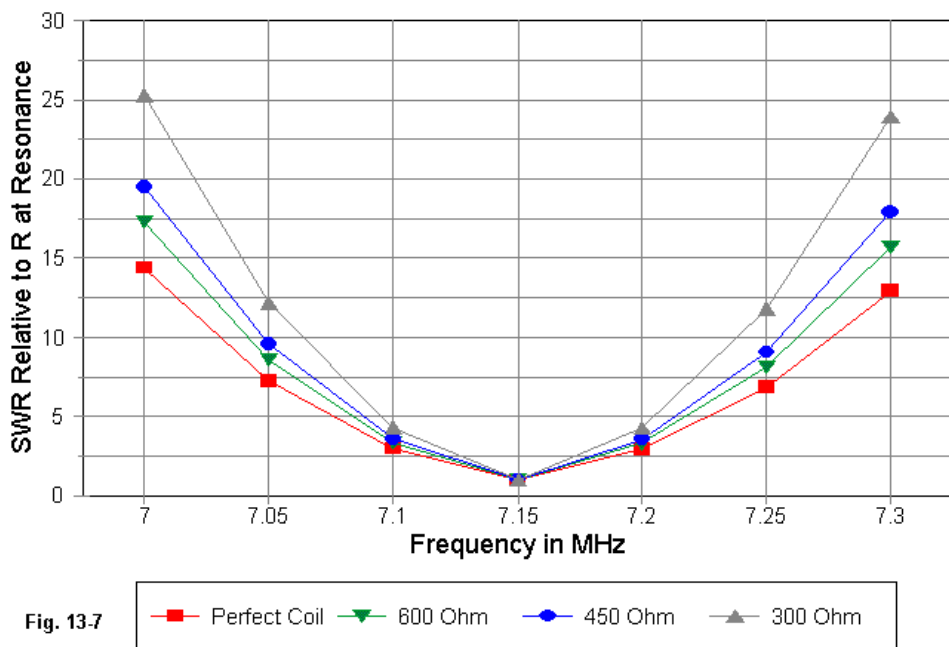
We create on the segments on each side of the source a TL or transmission line entry. Depending on the program we use, we might have to create a distant wire to terminate the stub or the program might do this for us and simply let us specify that we desire a short at end 2. Do not enter the length in electrical degrees, or a frequency sweep will not be accurate. That same electrical length, with a different physical length, will appear at each frequency checked by the sweep. Instead, enter the design center physical length of the stub, along with the characteristic impedance of the line. We shall explain in a moment why the 50- and 75-Ohm lines are not included in **Fig. 13-7**.

With all practical characteristic impedance lines, the SWR bandwidth is significantly narrower for a transmission line stub load than for a coil of indefinitely high Q. Note that the SWR bandwidth decreases as the characteristic impedance of the line decreases. The low impedance lines were omitted because their end values of SWR exceeded 100 and would have obscured the detail of the graph.

The reason for the narrower bandwidth is not Q as ordinarily understood in terms of resistive losses within the reactive loading device. Neither loading means has any resistive losses in the models. Rather, the stubs have reduced bandwidth because the value of reactance is a tangential function of their electrical length as the frequency changes, which yields a curve quite unlike the curve of solenoid coil reactance vs. frequency. The difference shows up vividly in these examples, since the required reactance is so high. For small-value-line-vs.-coil applications, such as the values associated with beta matches, the curves across a ham band will not be significantly different if the stub uses a high characteristic impedance.

To reinforce the differences between single-layer solenoids and transmission lines stubs as sources of reactance in antenna loading, tables of values for the perfect coil and the bank of transmission lines used follow the graph in **Fig. 13-7**.

40-M SWR: Near-Center Load Perfect Coil vs. Linear Loads



		Source Impedance (R +/- jX Ohms) (Values rounded)				
		Transmission Line Zo				
Frequency	Coil	50	75	300	450	600
MHz						
7.00	13.1-47.3	13.0-207	13.0-154	13.1-64.8	13.1-56.3	13.1-52.7
7.05	13.3-31.5	13.2-147	13.3-107	13.3-43.3	13.3-37.5	13.3-35.0
7.10	13.5-15.6	13.5-78.5	13.5-55.9	13.5-21.7	13.5-18.7	13.5-17.5
7.15	13.8-0.0	13.8-0.2	13.8-0.1	13.8+0.0	13.9-0.0	13.8+0.0

7.20	14.0+15.7	14.0+91.5	14.0+61.2	14.0+22.0	14.0+18.8	14.0+17.6
7.25	14.2+31.1	14.3+201	14.3+122	14.2+43.9	14.2+37.4	14.2+33.9
7.30	14.4+46.6	14.7+334	14.6+205	14.5+66.1	14.5+56.2	14.4+52.3

The table shows several features of using transmission line stubs as loading elements at or near the center of the antenna element. First, the range of reactance variation is higher in all cases than for the perfect inductor. Second, although the phenomenon is smaller by far, the range of source resistance values is also greater for transmission line stubs than for perfect (and, by reference to earlier data, for finite-Q) inductors.

Since all of the loading devices used here have no resistive losses, their Qs as determined by the usual ratio of reactance to resistance are identical. However, their operating bandwidths are not at all equal due to the manner in which the reactance changes with frequency for the two types of reactive loads. Although there are ways around the problem for special cases, there is no direct way to introduce losses into the transmission line model used by NEC. Hence, one cannot by this means explore the effects of accounting for losses in the stubs. In most cases, the losses of stubs will require independent calculation outside the NEC model. That will leave also to independent calculation how to adjust the gain and source impedance figures reported by the NEC model. A centered load's losses call for reasonably straightforward adjustments to the source impedance, but as we shall see in our examination of mid-element loads, losses away from a dipole's center call for not-so-simple adjustments.

Despite this shortcoming, TL models of transmission lines stubs do reveal some properties of their use as reactive loads. The fundamental differences in behavior relative to inductors across a band of frequencies should be clear. Second, for single frequency use, stubs are equivalent in all ways to inductors. Finally, we may note in passing that inductively reactive stub loads have an alternative name: linear loads. That name has, unfortunately, led to a number of persistent misunderstandings about their basic nature. We shall have to return to that name later.

However, lest a small matter go unnoticed, let me set it out in bold relief. When we split and moved the loading reactance only about 9" each side of the source, we found an increase in the required value of loading reactance. At the same time, the 7.15 MHz resonant source impedance for the antenna increased in the perfect re-

actance models from 13.19 Ohms to 13.76 Ohms. Although two cases alone do not constitute a trend, I shall bet that one is afoot. And with this, we have a prelude to looking at inductively reactive loads installed at a distance from the centered source of our very short dipole.

* * * * *

Models included: 13-1 through 13-4.



14. Notes on Reactive Antenna Loads and Their NEC Models

B. Some Mid-Element Loading Basics

In the first installment of these notes on loads, we looked almost solely at short, center-loaded dipoles. Our test dipole is exactly $1/4$ wavelength long (34.39') at 7.15 MHz and uses perfect or lossless wire of various sizes between AWG #8 and AWG #12. For reasons that will become apparent in still later episodes, the test model has 91 segments, each about 4.53" long. For Type 4 (complex $R \pm jX$) loads or for Type 0 (series R-L-C) loads, we placed the load on the center segment (#46).

The one exception to this procedure arose when we substituted shorted transmission line stubs for the series R-L-C load, using the TL function of NEC. Since transmission lines appear in parallel with loads, we were forced to split the load and move the stubs to the segments immediately adjacent to the center segment (Segments #45 and #47). Two phenomena appeared:

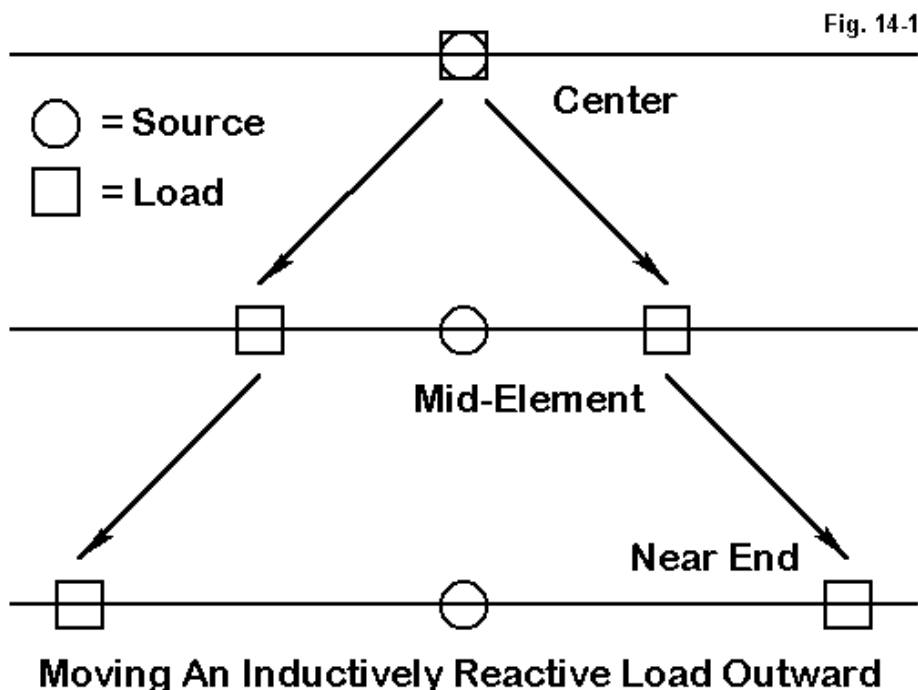
1. Even with such a small spacing between loads, the source impedance of the dipole increased by a measurable amount. Using the #8 AWG wire model, the resonant source resistance climbed from 13.2 Ohms to 13.8 Ohms.

2. Likewise, the required reactance for loading the wire to resonance also increased. For the #8 wire model, the required reactance each side of center climbed from 416.65 Ohms to 426.3 Ohms just by moving the loads to a spacing of about 9" apart.

These two phenomena form part of a set of trends with which all designers of loaded elements must be familiar. We shall look at them initially by using the same load-creation facilities of NEC that we used in looking at centered loads. Before we are finished with these basics, however, we shall formulate some questions about the reliability of models to tell us the entire story using only the "regular" means of load creation: Type 4, Type 0, and transmission line (TL) shorted stubs.

Load Position Trends

In theory, it is possible to split a center inductively reactive load and to place the halves symmetrically anywhere along the wire of our short dipole. As we move the load halves outward, we must increase the value of reactance of each half in order to bring the dipole to resonance. The following exercise is designed to generate an appreciation of how much reactance must be pressed into service for the purpose of re-resonating the dipole as we push the loads further outward.



Let us take the #8 model and place the loads successively 10% further outward from center, as suggested by **Fig. 14-1**. In terms of total antenna length, this movement amounts to placing the load at the 45%-55% positions, the 40%-60% positions, etc. For a 91-segment dipole, we can only approximate these positions, but the results are certainly close enough to reveal the trends involved. The exercise can use simple $R \pm jX$ loads. At each position, we increase the value of inductive reactance to bring the dipole to within ± 1 Ohm reactance of resonance at 7.15 MHz. (See model 14-1)

Initially, we can let the Q of the load be indefinitely high (or infinite, meaning resistively lossless) by setting the value of the load resistance to zero. We obtain the following table of results.

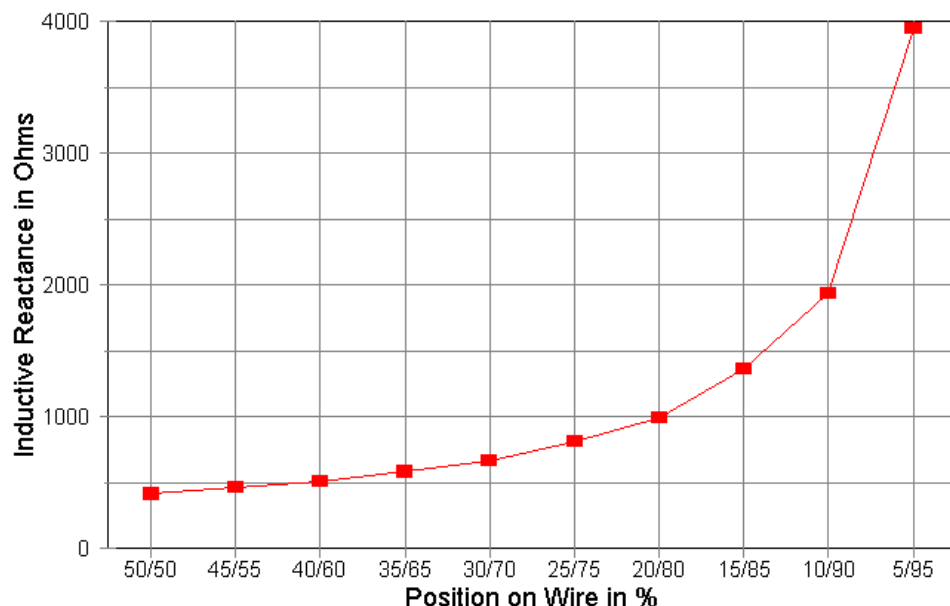
Position Approx %	Position Actual %	Reactance in Ohms per Side	Source Impedance $R \pm jX$ Ohms	Free Space Gain dBi
5/95	4.95/95.05	3948	$37.7 + j 0.3$	1.91
10/90	10.44/89.56	1939	$34.4 - j 0.5$	1.90
15/85	14.84/85.16	1369	$32.0 + j 0.4$	1.89
20/80	20.33/79.67	996	$28.9 + j 0.8$	1.88
25/75	24.73/75.27	817	$26.5 + j 0.8$	1.87
30/70	30.22/69.78	668	$23.6 - j 0.6$	1.87
35/65	44.62/65.38	586	$21.3 + j 0.7$	1.86
40/60	40.11/59.89	510	$18.4 - j 0.5$	1.86
45/55	44.51/55.49	465	$16.2 + j 0.7$	1.86
50/50	50/50	416.65	$13.2 - j 0.0$	1.85

The most significant trend to note is the steady rise in the required reactance to bring the dipole to resonance as the load is moved progressively outward from center. The graph in **Fig. 14-2** shows the trend with some clarity.

The source resistance and gain figures are predicated on the unlikely event of obtaining a perfect (lossless) inductively reactive load. A more realistic value would be a Q of about 300. This requires that for each load, we divide the reactance by 300 to obtain a series resistance for the total load. If we insert the required values of resistance into our $R \pm jX$ loads, we obtain the following values at 7.15 MHz with our #8 lossless wire model. (Reactance values do not change and are not shown below.) (Model 14-2)

Load Position vs. Magnitude Reactance of Inductive Load

Fig. 14-2



Position Approx %	Series R Ohms	Source Impedance R +/- jX Ohms	Free Space Gain dBi	Inductance microH
5/95	13.16	57.4 - j 0.9	0.08	87.9
10/90	6.46	44.7 - j 0.9	0.77	43.2
15/85	4.56	39.6 + j 0.2	0.96	30.5
20/80	3.32	34.9 + j 0.6	1.07	22.2
25/75	2.72	31.6 + j 0.7	1.11	18.2
30/70	2.23	27.9 - j 0.7	1.13	14.9

35/65	1.95	$25.2 + j\,0.6$	1.13	13.0
40/60	1.70	$21.9 - j\,0.5$	1.12	11.4
45/55	1.55	$19.6 + j\,0.7$	1.02	10.4
50/50	1.39	$16.0 - j\,0.2$	1.02	9.3

The inductance that yields the required reactance is shown in the right-most column for convenience, and we shall return to it in a moment. First, however, let us look at the values from left to right. The series resistance values can be compared to the reactance values in the preceding table to confirm a Q of 300. The other values may benefit from a bit of graphing. See **Fig. 14-3**.

Source Impedance vs. Load Position Infinite and Finite Q Inductors

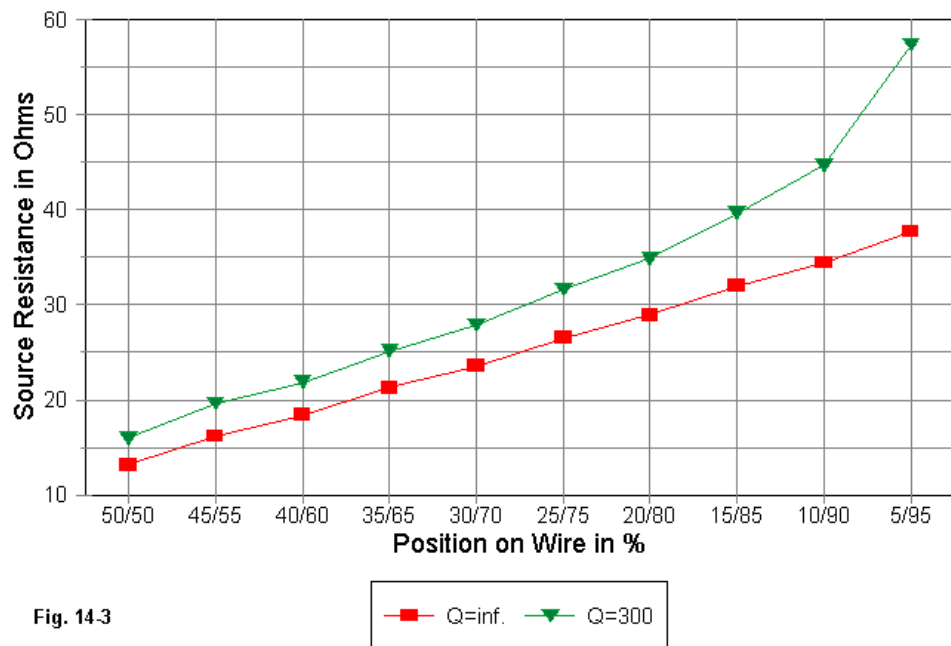


Fig. 14-3

The source resistance increases approximately, but not exactly, as the loss resistance increases for each pair of loads. The small differential between the reported source resistance and the sum of the source resistance with an infinite Q and the total of resistive losses results from placing the loads ever farther from the center or source point of the dipole. Nonetheless, the trend is abundantly clear.

Free Space Gain vs. Load Position Infinite and Finite Q Inductors

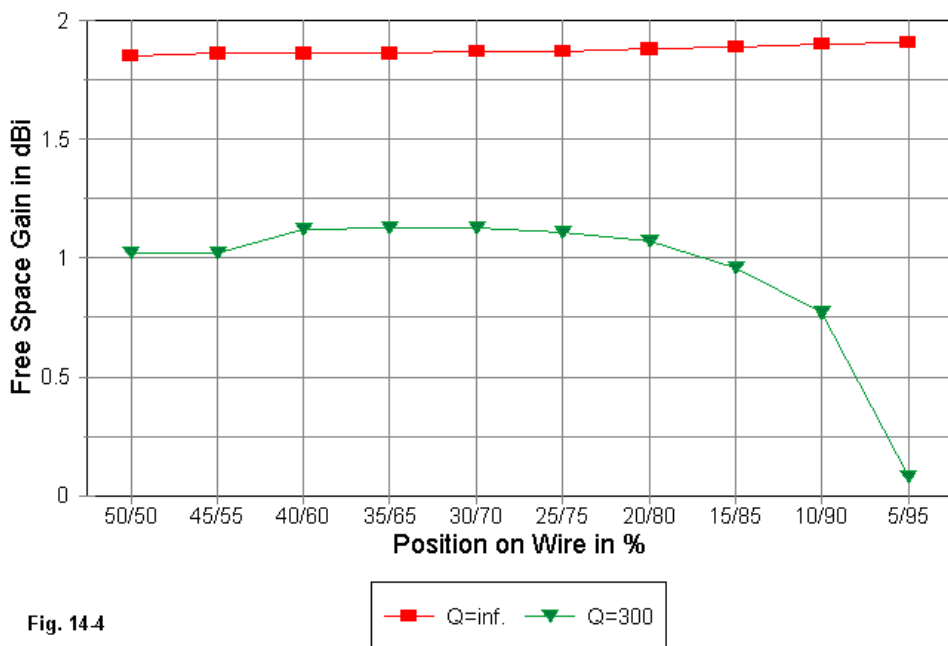
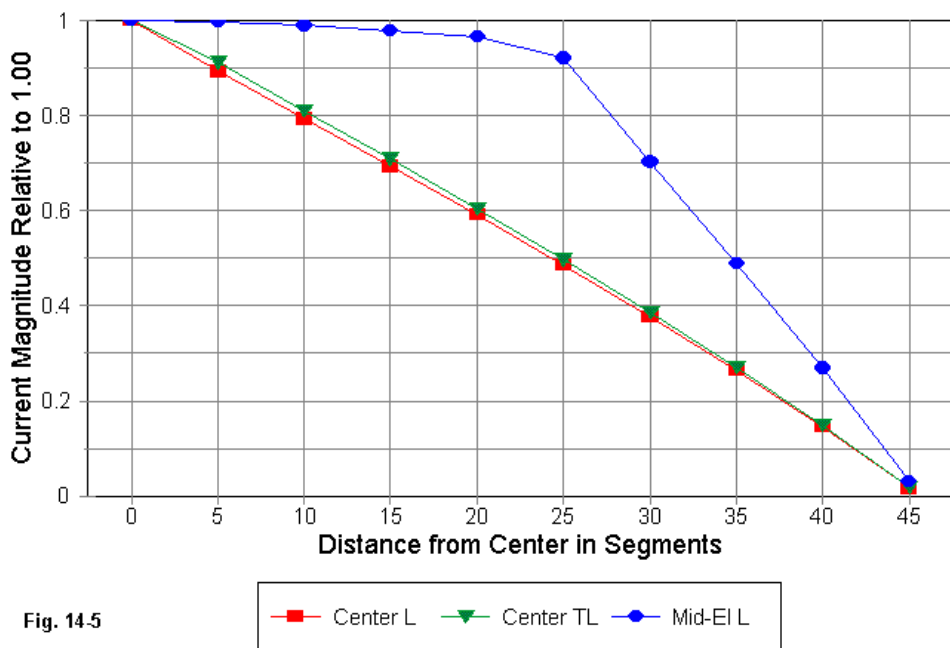


Fig. 14-4, which compares the free space gain of the dipole without load losses and with losses of a $Q=300$ load, is perhaps more interesting. The lossless loads show a wholly insignificant variation of gain, regardless of load position. However, with a $Q=300$ load, the gain of the dipole shows a rise of 0.1 dB when the loads are placed 20% to 60% of the total element length each side of center outward from the center—relative to the gain with the load centered. Beyond the 60% mark (20/80%, in terms of total dipole length), the gain decreases severely and rapidly as a result of the equal rapid increase in resistive loss that goes along with the steep rise in required loading reactance.

Current Profiles of Loaded Elements Inductances and Transmission Lines



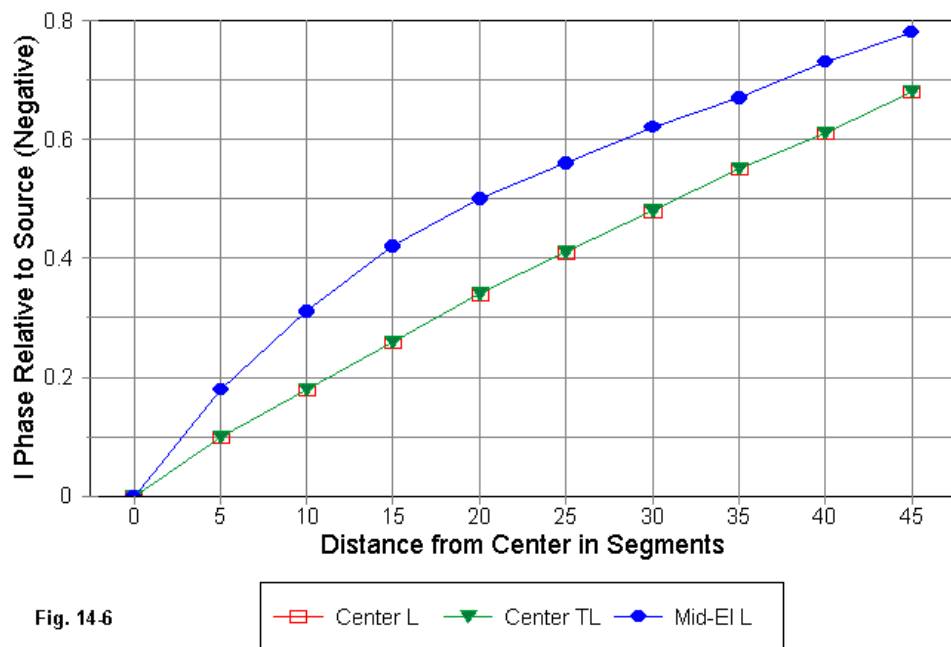
The 0.1 dB gain differential between center and mid-element loading is of questionable operational benefit. In a horizontal dipole, the use of mid-element loading in preference to center loading is for other reasons, most often the higher source resistance of the dipole with mid-element loading. There remains a persistent myth that mid-element loading shows significantly higher gain than center loading. This myth derives from short mobile antenna experience using loaded monopoles. Field strength differences resulting from the environment within which such antennas work and from the very short lengths they use have become unwarrantedly generalized to cover all antennas making use of inductively reactive loads.

A second source of the myth is an uncritical examination of current magnitude curves along loaded elements. **Fig. 14-5** displays the current magnitude for the R-L-C center-loaded element, the TL near-center-loaded element, and the mid-element R-L-C loaded element. Seemingly, the current magnitude (relative to a source value of 1.0) indicates a stronger far field based on the assumption that the field is roughly proportional to the current. However, that assumption needs considerable modification with loaded elements.

One modification is occasioned by the current phase change along the element, as shown in **Fig. 14-6**. For the mid-element loaded model, the rate of phase change along the element is much higher than for either center-loaded model. Moreover, the current magnitude curve is not a direct measure of the undisipated power in the element.

Using the strictly mathematical models of loads of all sorts, we may replace the Type 4 loads with either series R-L-C loads or with transmission lines (TL networks). (Note: as in the preceding episode in this series, we are purposely violating the injunction against using TL loads where current magnitude on the two lines is unequal in order to study the effects of trying to model with them.) The right-most column of the table shows the equivalent inductance values for the reactance loads used in this exercise. Although the weight and gain drop associated with such large inductance values is reason enough not to place loading inductors at the far ends of antenna elements, there is another reason as well. In at least the most extreme case (87.9 microH), for any practical single-layer solenoid coil-winding technique, we would not reach the required inductance before the antenna had passed resonance and had begun to show an inductive reactance at the source. We shall return to this point in a later episode.

Phase Profiles of Loaded Elements Inductances and Transmission Lines



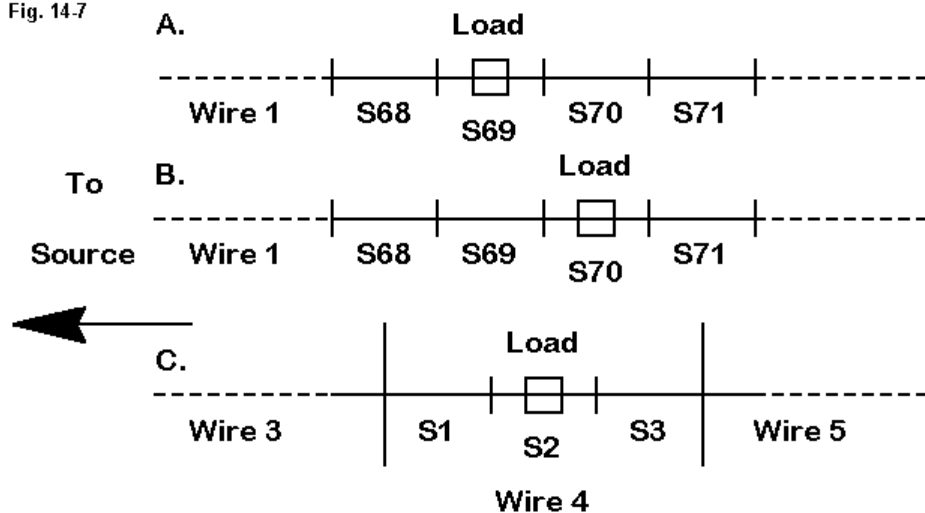
Alternative Means of Placing Loads

Although certainly precise enough to show the important trends associated with mid-element placement of loads, the technique of choosing segments on which to place loads was only approximate. Except at the very center of the dipole, the loads were a bit off their desired marks.

We could so segment the dipole that the loads would appear (as centered on their segments) in the desired places—but only for this exercise. If we chose other percentage values for load placement, further revisions of segmentation would be required. There is an alternative.

Consider the following problem (which, for the moment is academic only): suppose we wish to find the places along the dipole element where each of two mid-element loads is identical in required reactance to a single center load. For our 1/4-wavelength #8 perfect wire dipole, the required reactance of a center load is 833.3 Ohms to achieve resonance at 7.15 MHz. The question then is where along the wire to place two 833.3-Ohm loads also to achieve resonance.

Fig. 14-7



Precise Load Placement (Right End Shown)

The model has 91 segments and is partially sketched in **Fig. 14-7**. If we place the loads on segments 23 and 69, the source impedance reports as $27.11 + j 31.7$ Ohms, which indicates that the loads are too far inboard. If we place the loads one segment further outward—on segments 22 and 70—the source impedance reports

as 26.34 - j 39.04 Ohms, which indicates that the loads are too far outboard. Within the limits of the model, neither placement satisfies the requirement of achieving resonance within +/-1 Ohm. (Model 14-3)

These unsatisfactory situations are indicated in parts A. and B **Fig. 14-7**. The figure also indicates an alternative modeling method that permits precise load location without sacrificing model accuracy. The model in C. is a 5-wire model, all composed of the same type of wire. Two outer sections and a center section are long wires having many segments. Two special wires have 3 segments each, with the load placed on the center segment.

The use of 3 segments on the load wires serves to ensure that the segments immediately adjacent to the load segment are equal in length to the load segment. The center wire and the outer wires are segmented so that each segment length is as close as possible—within the limits of the total segmentation of the antenna—to the length of the load wire segments.

```
CM 1/4 WL Loaded Dipole #8 Lossless Wire
CE
GW 1 20 -206.34 0 0 -113.1 0 0 .064251968503937
GW 2 3 -113.1 0 0 -99.5 0 0 .064251968503937
GW 3 45 -99.5 0 0 99.5 0 0 .064251968503937
GW 4 3 99.5 0 0 113.1 0 0 .064251968503937
GW 5 20 113.1 0 0 206.34 0 0 .064251968503937
GS 0 0 .02540
GE 0
EX 0 3 23 0 1.414214 0
LD 4 2 2 2 0 833.3 0
LD 4 4 2 2 0 833.3 0
FR 0 1 0 0 7.15 0
RP 0 1 360 1000 90 0 1 1
EN
```

**5-Wire Dipole Model
For Mid-Element Load
Placement**

Fig. 14-8

The .NEC model in **Fig. 14-8** shows the dimensions of the final model in terms of the wire geometry for the dipole with Type 4 reactive loads. For the #8 wire model, a center wire 199" long (+/- 99.5") used 45 segments, each 4.42" long. The outer wires are each 94.24" long and use 20 segments, each 4.71" long. The 3-segment load wires are 13.6" long, with segments 4.53" long. The total number of segments for the antenna remains 91. (Model 14-4)

The loads are centered 106.3" each side of the source. With this configuration, the reported source impedance is $26.78 + 0.53$ Ohms, well within the limits set for resonance. The figures are for a pure reactance, with no series resistance.

The technique used here is subject to many variations, including re-segmentation to achieve even closer segment length equality along the antenna. Even more precise placement is possible to achieve resonance within narrower limits (for example, +/- 0.1 Ohm). However, for the purposes of this demonstration, the techniques used are sufficiently precise for both placement and resonance.

Center vs. Mid-Element Loading

Having placed the 833.3-Ohm load correctly at mid-element positions, it useful to take a preliminary look at the performance curves that result from applying the equivalent Type 0 (series R-L-C) load at both center and mid-element positions. The inductive equivalent of 833.3 Ohms at 7.15 MHz is 1.8549×10^{-5} H. We can look at the curves both for infinite Q and for some representative Q, say, 300. The series resistance required for a Q of 300 at the design frequency is 2.778 Ohms. (Model 14-5 and 14-6)

Once more, we shall use our #8 lossless wire 34.39' dipole.

#8 Wire Dipole		Source Impedance (R +/- jX Ohms)		
Freq.	Center Load	Center load	Mid-El. Load	Mid-El. Load
MHz	Q=inf.	Q=300	Q=inf.	Q=300
7.00	12.58-46.28	15.35-46.29	24.33-82.17	29.07-82.25
7.05	12.78-30.76	15.56-30.76	25.11-55.15	29.98-55.23
7.10	12.98-15.30	15.76-15.31	25.93-27.54	30.94-27.64
7.15	13.19-0.02	15.97-0.02	26.78+0.53	31.93+0.40

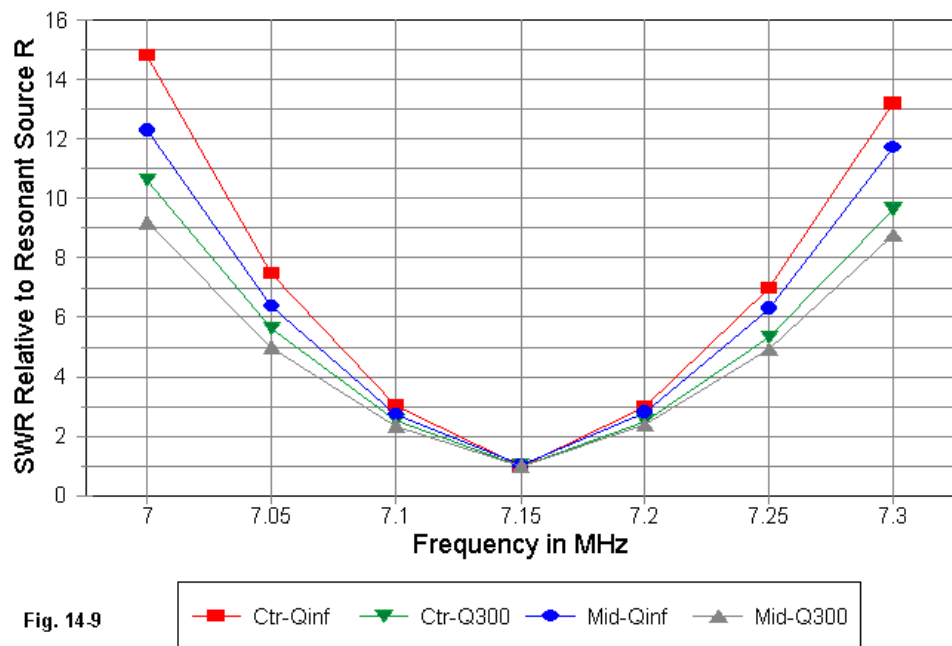
7.20	13.40+15.26	16.18+15.26	27.67+29.19	32.98+29.08
7.25	13.61+30.25	16.39+30.25	28.60+58.35	34.07+58.54
7.30	13.82+45.29	16.60+45.29	29.58+87.98	35.21+87.84
Delta R	1.24	1.25	5.25	6.14
Delta X	91.57	91.58	170.15	170.09

Mid-element loading results in a larger swing of both source resistance and source reactance across the span of frequencies swept in this exercise. The result for resistance is larger not only in terms of Ohms, but also when taking the swing as a percentage of the design center frequency value. The result for reactance, however, is the opposite: as a percentage of the design center frequency resistance, the reactance swing is smaller for the mid-element load case than for the center loading case. Moreover, for the mid-element loads, there is not the virtual identity of swings for either resistance or reactance that holds over the frequency span swept by the models with center loads.

The smaller change of reactance relative to the resonant source resistance also suggests that the SWR curves for the mid-element loading cases might be broader than for their center-load counterparts. Of course, SWR is plotted relative to the design center-frequency resistance value, where the antenna is resonant within the prescribed limits for this exercise.

The graph in **Fig. 14-9** confirms our suspicions. The upper two curves are for the test runs at an indefinitely high Q , while the lower curves are for the value $Q=300$. In both pairs, the curve for the mid-element loaded dipole is broader than its center-load counterpart. Two cautions attend these curves. First, the effect of wire losses are not accounted within the overall curve, since we are using perfect wire in order to isolate, as best we can within the model, the phenomena that may ascribed directly to the load itself. Second, the conditions of modeling must be remembered: the model is a free space model that, of course, does not take any ground effects into account. The modeled loads assign both the series resistance and reactance and do not take into account any further effects that might accompany a real single-layer solenoid inductor placed in center or mid-element loading service. Although specified in a Type 0 load in terms of a series combination of resistance and reactance, the loads are strictly mathematical entities whose physical properties are not calculated in the overall antenna evaluation.

SWR Curves: Center & Mid Loading For $Q = \text{Infinite}$ and $Q = 300$



The antenna free space gain for the various cases corresponds well to the curves earlier presented. The infinite- Q antenna shows a gain of 1.85 dBi for the center-loaded case and 1.87 dBi when mid-element loaded. When each loading inductor is set at a Q of 300, the gains are 1.02 dBi and 1.11 dBi, respectively. Again, these gain figures do not include any wire loss associated with the main element.

While we are exploring these modeling curves, we might as well compare transmission line loading when placed at the same mid-element point with the case de-

veloped in Part 1. In that instance, we placed a split load in the segments immediately adjacent to the source segment, because placement exactly on the source segment would not have yielded valid series load results. Even that small move of the load position required that we use two 426.3-Ohm loads, rather than a pair of 416.65-Ohm loads ($833.3/2$). The resultant 600-Ohm (velocity factor = 1.0) transmission line stubs were each 13.52' long (model 13-4). To replace the mid-element 833.3-Ohm loads with transmission line stubs with the same characteristic impedance requires lengths of 20.73' each (model 14-7). Once more, remember that transmission lines used in NEC models are lossless, non-radiating, and mathematical. Within those constraints, the results are as follows:

#8 Wire	Source Impedance ($R \pm jX$ Ohms)	
Freq.	Near-Center	Mid-Element
MHz	Placement	Placement
7.00	13.09-52.66	23.84-110.0
7.05	13.31-35.04	24.76-74.67
7.10	13.53-17.46	25.74-37.85
7.15	13.76-0.03	26.78+0.42
7.20	13.98+17.57	27.90+40.47
7.25	14.21+34.88	29.09+82.25
7.30	14.44+52.30	30.38+126.0
Delta R	1.35	6.54
Delta X	104.96	236.0

The mid-element transmission line load source impedance value verifies that the transmission line stub is the correct replacement for the reactance load of 833.3 Ohms. However, even with a 600-Ohm transmission line, the variance of reactance across the band is very large. As we have noted, with transmission lines of lower impedance, the variance will be larger still.

SWR Curves: Center & Mid TL Loads Transmission Line $Z_0 = 600$ Ohms

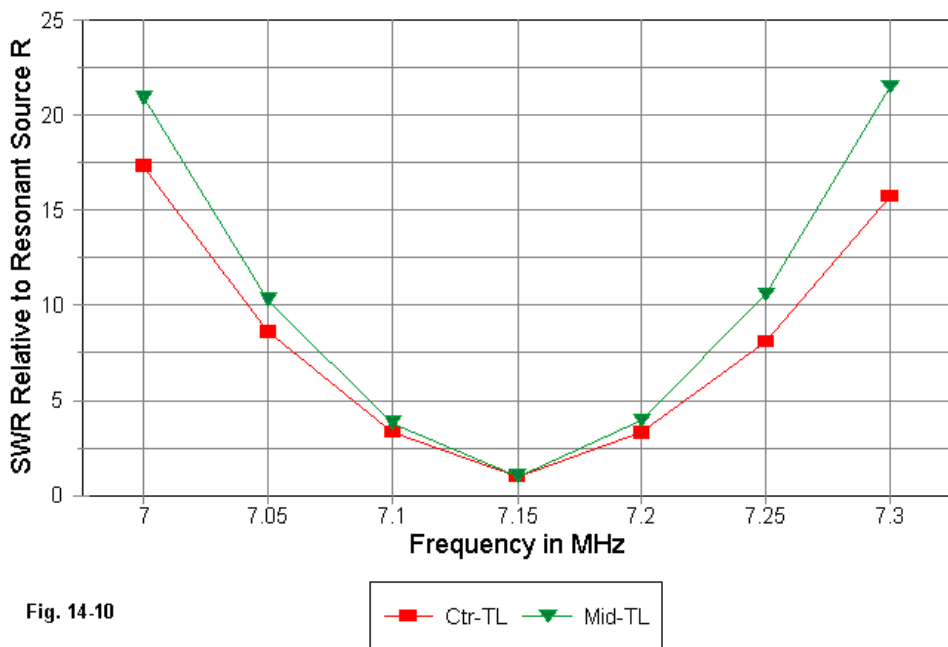


Fig. 14-10

Comparing the SWR curves (relative to an SWR of 1:1 at resonance) for the near-center and the mid-element transmission line loads (**Fig. 14-10**) shows an interesting result that is opposite to results from using series R-L-C loads. The mid-element loading stub system yields a narrower SWR curve than the near-center loading system. This consequence follows from the much high values of reactance required of the stub. Since the reactance of the stub is a tangent function of the

electrical length of the stub, stubs over 45 degrees long change their reactance values rapidly. The required stub length in this case is 54.2 degrees.

Concerns About Modeled Loads

Commonly, series R-L-C loads are considered to be adequate models of single-layer solenoid inductor performance in a loaded dipole or similar element. Despite their strictly mathematical nature, that is, the fact that their physical structure does not play a role in the calculation of predicted antenna performance, the resulting antenna performance figures are considered accurate enough for most practical design and analysis purposes. The key loss factor in a solenoid inductor is the series resistive loss, and other losses and factors that might cause variations from the predicted outcomes are considered too small to be of significance. Consequently, only in the most critical cases are solenoid loading coils physically modeled—and usually only where such coils are very simple. An example of such a case is the solenoid in the middle of many automobile UHF antennas for cellular telephone service.

Transmission line loads, on the other hand, are widely distrusted as models of linear loads used in HF antennas. Most obviously, they do not permit the accurate introduction of resistive losses in the transmission line. Nor do they permit, except as an external calculation, the introduction of the line's velocity factor. (Some programs permit velocity factor information to be introduced into the interface between the user and the core calculation, subsequently converting it to the values needed by the core's network calculations.)

Most antennas employing linear loads use load wire having a diameter different from that of the main element. Although MININEC can directly handle these types of models, the required model often presses the segmentation limit of the program. (Some recent MININEC developments, such as NEC4WIN, have broken the 256-segment barrier, and one version permits virtually unlimited numbers of segments, but at the cost of slow calculation speeds.) MININEC has additional requirements in order to reduce errors at sharp corners; NEC (in either version 2 or version 4) is inaccurate when faced with angular junctions of wires having dissimilar diameters. Consequently, physically modeling existent antennas with linear loads is not normally attempted.

However, it is possible to physically model linear loads using a uniform wire diameter throughout the model. While the technique does not yield an accurate model of a particular antenna, it does produce quite accurate results that can reveal some interesting properties of linear loads. The technique may also permit some preliminary sorting of different functions performed by a linear load and the conditions under which it performs them. Physically modeling some linear load configurations will therefore be our next task.

* * * * *

Models included: 14-1 through 14-7.

15. Notes on Reactive Antenna Loads and Their NEC Models

C. Some Linear Loading Basics

Linear loads do not model accurately using the TL facility of NEC. There are two major factors (and a number of minor ones) that contribute to the inaccuracy. First, the currents on the wires of the linear load are rarely, if ever, exactly equal in magnitude and opposite in phase. Consequently, the linear load does not act as a pure transmission line. Second, most applications of linear loading snug the parallel load wires close to and parallel to the main radiating element. This placement creates complex interactions between the linear load wires and the antenna wire, often to the point that the contributions of each wire to the total far field and to the total element length cannot easily be distinguished.

Nevertheless, NEC is able to physically model linear loads with a good level of accuracy, where “good” might be provisionally defined as providing reasonable guidance for building or providing reasonable calculations for analysis. Good accuracy, however, is subject to certain restrictions. For example, for best accuracy, the wires comprising the antenna element and the wires comprising the linear load must be the same diameter to avoid both closely spaced wires of differing diameters and wires of different diameters meeting at an angular junction. In both cases, NEC results can be quite inaccurate. One consequence of this restriction is that NEC cannot directly model most of the existing commercial antennas employing linear loading, since these antennas generally use element and load wires of radically different diameters.

MININEC is subject to neither of these restrictions. With due attention to the need for very short segment lengths at angular junctions and to the overall segment limitation of many versions of MININEC, the program is capable of directly modeling virtually any linear loaded element. MININEC was the program of choice in my 10-meter study of linear loaded dipoles and Yagis in “Modeling and Understanding Small Beams: Part 4: Linear-Loaded Yagis.” *Communications Quarterly*, (Summer, 1996), pages 85-106. These antennas used 0.75" aluminum elements and #12 copper wire linear loads, and MININEC provided excellent building guidance. Despite the different materials used, no construction correctives could be directly at-

tributed to the program limitation of specifying a single material for the entire modeled structure. For conservative predicted results, the higher loss material (aluminum) was used throughout the modeling.

Nonetheless, NEC (either -2 or -4) can model linear loaded elements where all wires are of the same diameter. This much capability is sufficient to reveal a number of linear load properties that are fundamental to this loading technique. Consequently, we may for this exercise employ some of the models already developed for other parts of this series and simply add to them various linear load configurations. Central to the work in this episode will be the 34.39' (412.68") short dipole that we have loaded in several ways. We shall principally use the #8 wire version. Throughout, we shall retain the use of 91 segments for the overall antenna main element, with an average segment length of about 4.53". This segment length will determine several other parameters in the course of physically modeling transmission lines or linear loads.

In previous exercises, we have used perfect or zero-loss wire for the dipole element. So long as we could specify load losses in Type 0 or Type 4 loads (series R-L-C and series R-X loads, respectively), we could isolate losses directly attributable to the load. In this exercise, we shall physically model linear loads as wires in the antenna structure. In order to leave the "main" element perfect and register the material losses of the load wires, we shall have to use either NEC-Win Pro or GNEC (NEC-2 and NEC-4 programs from Nittany Scientific). These are the only commercial versions of NEC commonly available that permit specification of different materials for each wire in the antenna.

Physically Modeling Transmission Lines

Modeled transmission lines must be parallel wire assemblies, since there is no practical way to model coaxial cables. (It is possible to simulate the cable braid with several wires of the same diameter as the center wires, but the cost in wire segment total makes the technique impractical here.) However, since a parallel wire line with significant spacing between wires represents the most common form of linear loading, this restriction will impose no problems.

The linear loads we shall use will have a spacing of 4.53" center-to-center between wires. The diameter of a #8 AWG wire is about 0.1285". We can calculate

the anticipated characteristic impedance of the resulting parallel line from this standard equation:

$$Z_o = 276 \log_{10} \left(\frac{2S}{d} \right) \Omega$$

where Z_o is the characteristic impedance in Ohms, S is the center-to-center spacing of the wires, and d is the wire diameter, with both S and d in the same units. For the #8 line, the Z_o is 510.1 Ohms.

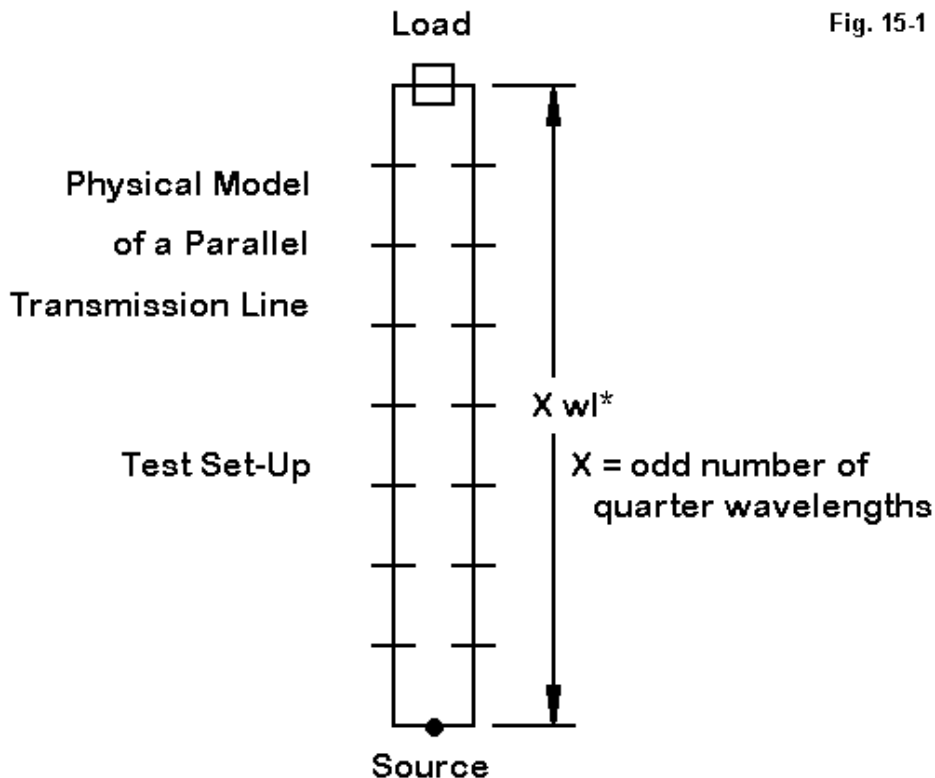
The standard equation presumes an indefinitely long line without connection wires to either a source or load. Moreover, it is invariant for any wire type and thus does not account for variations in the materials one might use for an open line. (In addition, it does not account for spacers, insulation, and other line variables that are not a part of this exercise.) In fact, wires physically arranged as parallel transmission lines do not model with precisely the same characteristic impedance as the value yielded by the standard equation.

We can model transmission lines to determine their characteristic impedance using **Fig. 15-1** as guidance. The “top” and “bottom” wires are single segments, each 4.53" long. The parallel wires will be an odd multiple of a quarter wavelength for the frequency tested. In our case, the frequency is 7.15 MHz, and we shall check the line using lengths of 1/4, 3/4, and 5/4 wavelength.

By adjusting the line to a length that yields resonance at the source for a given load, we can find the characteristic impedance of the line from the standard equation,

$$Z_o = \sqrt{Z_L Z_S}$$

where Z_o is the characteristic impedance of the line, Z_L is the load impedance, and Z_S is the source impedance. If we use a purely resistive load and bring the line length to resonance (defined as usual as less than +/- 1 Ohm reactance), the calculation is simplified. Let us initially specify a load of 5,000 Ohms resistance with a zero-loss (perfect) #8 wire line and see what we get.



Test Length in wl	Resonant Length "	ZS R +/- jX	Zo Ohms	Segs/Length	Delta L
1/4	409.9	59.2 - j 0.1	544.1	91 / 4.50"	2.75"
3/4	1236.6	59.3 - j 0.0	544.6	271 / 4.56"	1.45"
5/4	2063.0	59.5 + j 0.0	545.3	451 / 4.57"	0.40"

The “Delta L” figure is the departure of the resonant length from a perfect odd multiple of a quarter wavelength. The model (model 15-1) has two wires at top and bottom, which contribute to the overall line length and become error sources with respect to having a perfect parallel transmission line. However, the error they introduce decreases as the test line length increases.

The selection of 5000 Ohms as the resistive load for this construction was arbitrary. For any given test length for the line, selection of a different load value will give slightly different results. Let’s look at a 3/4-wavelength (1236.6") line and examine the characteristic impedance yielded by various load values.

ZL	ZS	Zo
Ohms	R +/- jX	Ohms
50	5920.0 - j 40	544.1
100	2961.0 - j 10	544.2
250	1185.0 - j 2.0	544.3
1,000	296.3 - j 0.0	544.3
2,500	118.6 - j 0.0	544.5
5,000	59.3 - j 0.0	544.6
10,000	29.7 - j 0.1	544.6

With perfect wire, the range of variation in this second test of the characteristic impedance derived from the test model is not especially significant, with a maximum variance of about 0.5 Ohm for the range tested. The key variance in this exercise is between the modeled-derived Zo and the value calculated by the standard equation: about 33 Ohms—or a 6% variance, with the modeled impedance higher than the calculated value.

If we specify the #8 wire to be copper (Resistivity = 1.72E-08), we end up with slightly different tables relative to those for perfect wire:

Test Length	Resonant	ZS	Zo	Segs/Length	Delta L
in wl	Length "	R +/- jX	Ohms		
1/4	409.7	60.1 + j 0.1	548.1	91 / 4.50"	2.98"
3/4	1235.6	61.7 - j 0.0	555.4	271 / 4.56"	2.44"
5/4	2061.0	63.4 - j 0.1	563.1	451 / 4.57"	2.40"

The range of variation in derived Z_o for the different test lengths, using the arbitrary 5000-Ohm load, is almost 15 Ohms. This represents a range of change of around 3%, which is well within the range of variation in real lines, especially those with insulation and a consequential velocity factor. However, the question remains as to whether the test yields accurate results. The second test provides the clue we need.

ZL	ZS	Z_o
Oms	R +/- jX	Ohms
50	5669.0 - j 80	532.4
100	2901.0 - j 22	538.6
250	1179.0 - j 5.3	542.9
1,000	298.5 - j 0.5	546.3
2,500	121.0 - j 0.1	550.0
5,000	61.7 - j 0.0	555.4
10,000	32.0 - j 0.0	565.9

Note: when using low values of load impedance, it may not be possible to remove all of the reactance to the arbitrary standard of resonance that we have been using (+/- 1 Ohm reactance). However, if the reactance value is less than about 2% of the resistive value and slight line length changes to achieve resonance do not change the resistive value significantly, then the resulting resistive value will be usable.

The test using different load values shows a wide range of values both above and below the values yielded by the perfect wire test. Since the approximate value of the characteristic impedance is also approximately the geometric mean between the adjacent load values (250 and 1000 Ohms), we can zero in on the actual line impedance by using that mean as a new load. If the first try does not yield a source impedance equal to the load we chose, we can adjust the load slightly until the results are equal within about 0.1 Ohms (to allow for rounding conventions in the software).

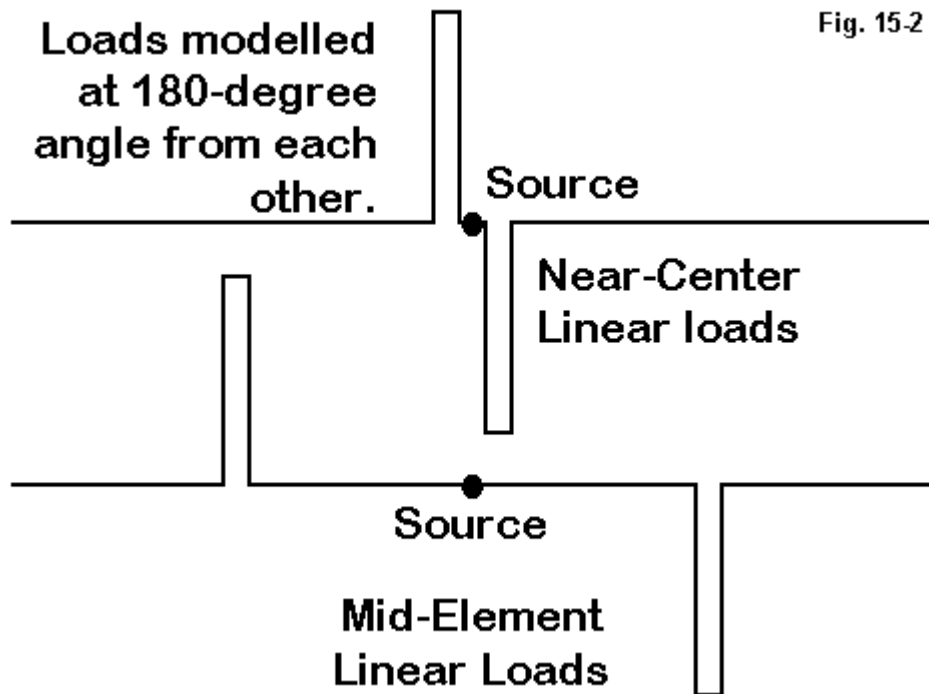
Using this method, the final value for the perfect wire Z_o is a little over 544.3 Ohms, while the final value for the copper wire is a little over 544.8 Ohms: very close indeed. In fact, we have extended the number of significant digits with respect to the characteristic impedance of the line too far for practical purposes in order to

ensure clarity of the mathematical progressions. A value of 544 or 545 Ohms would suffice for all practical enterprises. This conclusion means that running the characteristic impedance modeling test with perfect wire with any reasonable load would have produced an acceptable result with a single test run.

Modeling Transmission-Line Loads

We have lingered over the transmission line modeling test not only to show that transmission line characteristic impedance may differ in modeling software from values derived from standard equations, but as well to ensure that when modeling transmission line loads, we do not draw the wrong conclusions from the results we encounter. In order to get a handle on drawing the right conclusions, the next step is to model some transmission line loads using our #8 parallel transmission line. Because the principles do not change whether or not we account for wire losses and because the models are for demonstration purposes and do not represent structures anyone would actually build, we may use perfect wire transmission lines for these tests.

The models we shall construct use the basic 34.39' #8 perfect wire dipole. We shall insert transmission line loads at the segments adjacent to the feedpoint or source segment in the center and later at the points in the lines calling for 833.3-Ohm mid-element loads. From previous episodes of this exercise, we determined that loads placed immediately adjacent to the source for the 91-segment model required a reactance of 426.3 Ohms. We also used the 5-wire model to arrive at the positions of loads requiring 833.3 Ohms (the value of a single center load for this antenna).



For our model, we may place the transmission lines parallel to each other or at 180 degrees apart, as shown in **Fig. 15-2**. I have chosen the latter type of model to minimize interactions between the transmission lines, especially in the close-spaced model. Whichever model we choose, there will be a significant radiation component at right angles to the radiation from the main element.

Let us take the Z_o of the transmission line as 544 Ohms. Using the equations shown in earlier episodes, the 426.3-Ohm load requires a transmission line length

of 174.6" while the 833.3-Ohm load is 260.7" long. Modeling these lines via the TL facility, in which the lines are mathematically but not physically modeled, would require these calculated values.

Physically modeling the lines tells a quite different story, one that reveals why using the TL facility of NEC yields inaccurate results. To achieve resonance with the closely spaced transmission line loads required a length of 155.3" for each line, almost 20" shorter than called for by the equation. The mid-element model required loads that were 157" long, more than 100" shorter than called for by the standard equations. (For alternate models using lines that parallel each other, the values were 151.2" and 156.5" respectively.) Each loading transmission line replaced one segment of the main element and the spacing was set at 4.53" between transmission line wires. (Models 15-3 and 15-3)

A transmission line acts as a pure transmission line if and only if the currents everywhere along the line are equal in magnitude and opposite in phase on the two wires. An examination of the current tables produced by the NEC core reveals the degree to which this condition is not met by physical lines used as loads. In constructing models to perform this test, it is essential that the segments on each wire align perfectly.

The current phase differential along the transmission line wires does not exceed 0.4 degrees in either case, so current magnitude comparisons are sufficiently accurate for the level of analysis required here.

The graph in **Fig. 15-3** shows the current magnitude differential between the transmission line wires for both cases tested. Current differential is greatest where the transmission line joins the main element. It does not reach zero at the far end of the line. The level of current differential is far less along the closely spaced or "center" load lines than along the mid-element load lines. Note, however, that there is nothing like a linear relationship between the degree of current magnitude differential and the degree of shortening of the lines.

Current Magnitude Differentials Between Wires Along Linear Loads

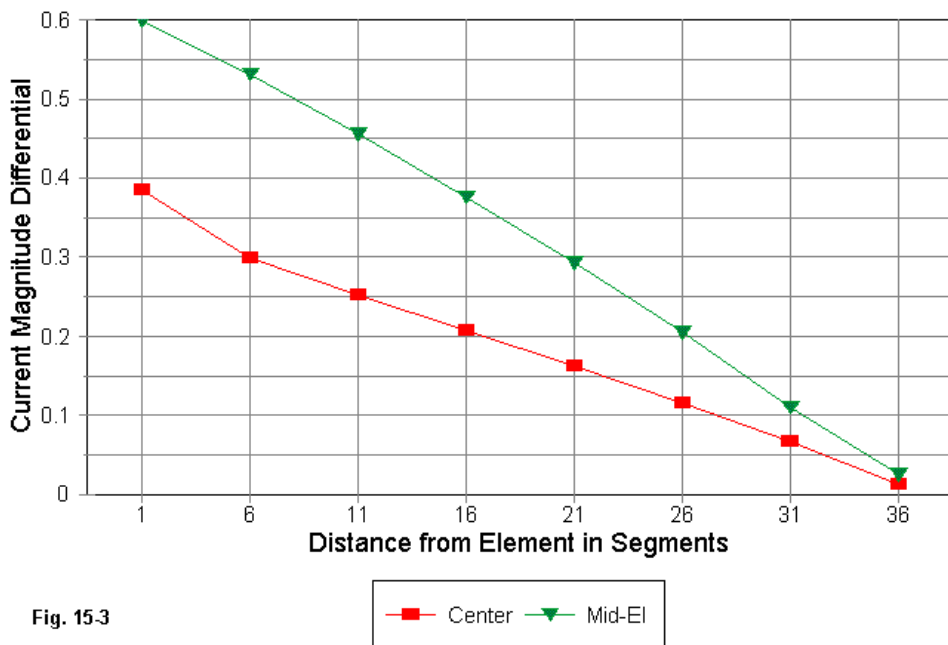


Fig. 15.3

The current differential is an indicator of—but not a measure of—the fact that the lines have significant radiation. As radiating lines, they contribute linearly to the length of the antenna. To the degree that the wires in the load contribute to the linear length of the antenna wire, less reactive loading is required, and the lines required will be shorter than called for in standard equations. The higher the current magnitude differential, the greater the contribution to the antenna's linear length and the higher the level of load shortening relative to equation-derived values. By mid-

element, the contribution to the wires to the antennas linear length is so great that the load lines may be over a third shorter than their calculated value.

In designing this overall exercise, I specifically chose a very short dipole (about 1/2 normal length) in order to set the effects we have noted in bold relief. Antennas only mildly shortened will show lesser current differentials in each of the models, and the degree of load shortening will be correspondingly less. Nonetheless, I know of no way of eliminating the fact that transmission line loads act as both reactive loads and as part of the linear antenna length—short of eliminating transmission lines as loads.

The placement of the loads in this drill is far from optimal for maximum performance for the antenna. The gain of the close-spaced or center load model is 1.67 dBi, while the gain of the mid-element model is 1.88 dBi. The latter value is comparable to the values for zero-loss center and mid-element R-L-C loads (1.85 and 1.87 dBi, respectively), but there is considerable cross-polarized radiation that reduces the normal side nulls of a free space dipole model.

Before we leave these informative but impractical models, we should briefly examine the SWR curves (relative to the source impedance at resonance for each model) in **Fig. 15-4**. The close-spaced or center loaded model shows a curve consistent with those we examined earlier in connection with transmission line loads generated with the TL facility. The mid-element line, in contrast, shows a very shallow curve, indicating a much wider operating bandwidth. This curve is in stark contrast to curves associated with TL-generated loads.

The source of the broader SWR curve for the mid-element transmission line load lies mostly in the fact that the load lines act to a much higher degree as part of the linear or radiating length of the antenna. Consequently, the SWR curve approaches that of a standard full-size dipole. The major difference is that the shortened dipole has a resonant impedance of about 20 Ohms. (The resonant impedance of the close-spaced model is under 10 Ohms. The impedance of a perfect center-loaded dipole in this series of models is about 13 Ohms, while a perfect mid-element loaded version shows a resonant impedance of about 27 Ohms.)

SWR Curves: Center & Mid TL Loads

TL $Z_o = 544$ Ohm; Zero-Loss Wire

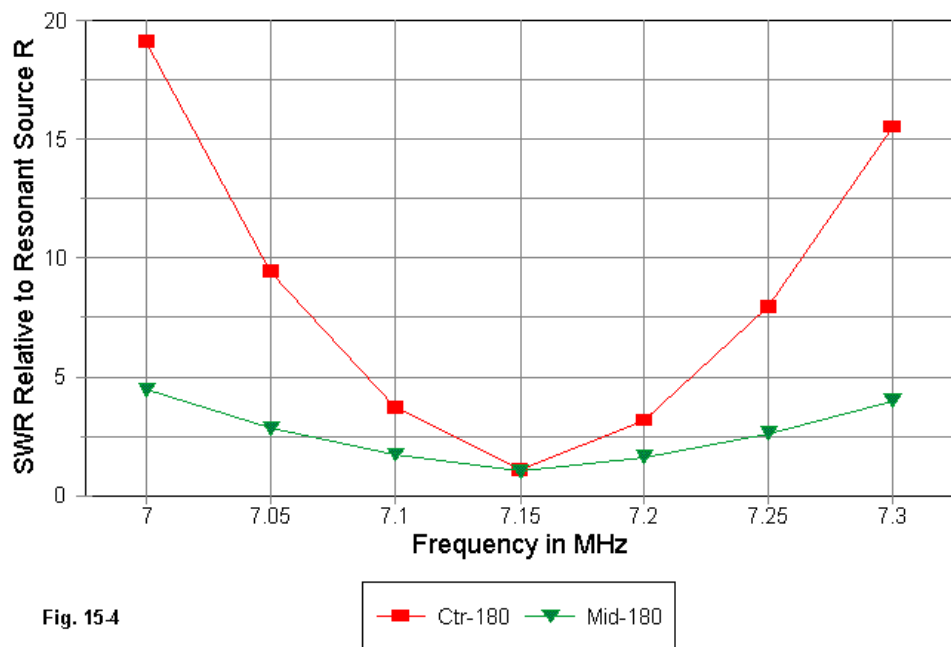


Fig. 15-4

Linear Loads

The transmission line loads that we have just explored place the load lines at right angles to the main element in order to minimize interactions. Conventional linear loads tend to place the loading lines parallel to and close by the main element.

Main element = uppermost wire in each sketch.



Triangular Center Linear Load



In-Plane Linear Center Linear Load



**Triangular Pseudo-Mid-Element
Linear Load**



**End
Views**

Fig. 15-5

The sketch in **Fig. 15-5** shows three typical linear load line assemblies. Actual antennas tend to fall into one of the three categories, although details may vary. The triangular load at the top feeds the center junction point of the linear load shorted lines, which then attach to the main element. Although we cannot accurately model actual structures in this exercise, the main element of most antennas of this type will be much larger in diameter than the load lines. For many purposes, the load lines may be placed in a plane with the main element, as shown in the middle sketch. Once more, the main or upper element will be larger in diameter than the load lines.

A third popular assembly is shown at the bottom as a mid-element linear load. In this case, a “main” or larger diameter element is the feedpoint. The apparent load lines begin at mid-element and return toward the center, terminating and returning

to the completion of the main element. Ordinarily, the apparent load lines are placed symmetrically about the main element, forming a triangle or a single plane. Why the sketch refers to this schema as a “pseudo-mid-element linear load” will become apparent as we proceed.

The models for the tests were carefully constructed to achieve several goals. The segments throughout are all close to 4.53" long, and the spacing between lines is exactly that figure. Hence, the triangular models form equilateral triangles so that the influence of the main element is as equal upon both load lines as possible. The center separation between load ends (shorting segments) is also 4.53" or one segment's length. This procedure permitted alignment of all segments both between the load wires and on the main segment as well. All of the models are #8 AWG wire.

In this sequence of tests, however, we shall look both at load lines that use perfect wire and lines that use other materials. **Fig. 15-6** shows an in-plane model in which only the load lines (wires 2 through 8) are copper. In this way, we may isolate the losses due to the load from the losses due to the main element wire, which remains a perfect or zero-loss wire like the ones used in earlier R-X and R-L-C load models.

The table below **Fig. 15-6** presents the load line lengths, measured from the source point for each of the three types of loads modeled. The “triangular mid-element” load line is measured from the source point both to the start and to the finish of the line outward. For the center triangular and in-plane models, line lengths for perfect, copper, and aluminum load lines are recorded, while the mid-element line is recorded only for the perfect wire version.

Antenna	Load Material	Load-Line Length ("
Triangular Center	Perfect	175.2"
	Copper	174.5"
	Aluminum	174.35"
In-Plane Center	Perfect	188.0"
	Copper	187.3"
	Aluminum	187.15"
Mid-Element	Perfect	176.43" (outer)
		171.90" (inner)

```

CM In-plane loaded dipole
CE
GW 1 44 -206.343 0 0 -2.265 0 0 .064251968503937
GW 2 1 -2.265 0 0 -2.265 0 4.53 .064251968503937
GW 3 41 -2.265 0 4.53 -187.3 0 4.53 .064251968503937
GW 4 1 -187.3 0 4.53 -187.3 0 9.06 .064251968503937
GW 5 83 -187.3 0 9.06 187.3 0 9.06 .064251968503937
GW 6 1 187.3 0 9.06 187.3 0 4.53 .064251968503937
GW 7 41 187.3 0 4.53 2.265 0 4.53 .064251968503937
GW 8 1 2.265 0 4.53 2.265 0 0 .064251968503937
GW 9 44 2.265 0 0 206.343 0 0 .064251968503937
GS 0 0 .02540
GE 0
EX 0 5 42 0 1.414214 0
LD 5 2 1 1 5.8001E7
LD 5 3 1 41 5.8001E7
LD 5 4 1 1 5.8001E7
LD 5 5 1 83 5.8001E7
LD 5 6 1 1 5.8001E7
LD 5 7 1 41 5.8001E7
LD 5 8 1 1 5.8001E7
FR 0 7 0 0 7 .05
RP 0 1 360 1000 90 0 1 1
EN

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Sample In-Plane Linear-Loaded Element

Note that the "main" element is perfect wire, while only the wires for the linear load are copper (LD 5, wires 2-8).

Fig. 15-6

The decrement in length for the loads with the use of materials having a reduced conductivity is identical for both the triangular and the in-plane models. The load length for the in-plane model is longer than for the triangular or the mid-element model. However, the mid-element model load structure is virtually identical in length to the equally triangular center loaded model. Note that the triangular model load lengths correspond very closely to lengths for a split center load reactance when calculated from standard equations using the 544-Ohm Z_0 of the load line. For the required 416.7-Ohm load, the required load line calculates to 171.7" or within 2% of the actual modeled line lengths.

Current Magnitude Differentials Parallel Linear Loads; Zero-Loss Wire

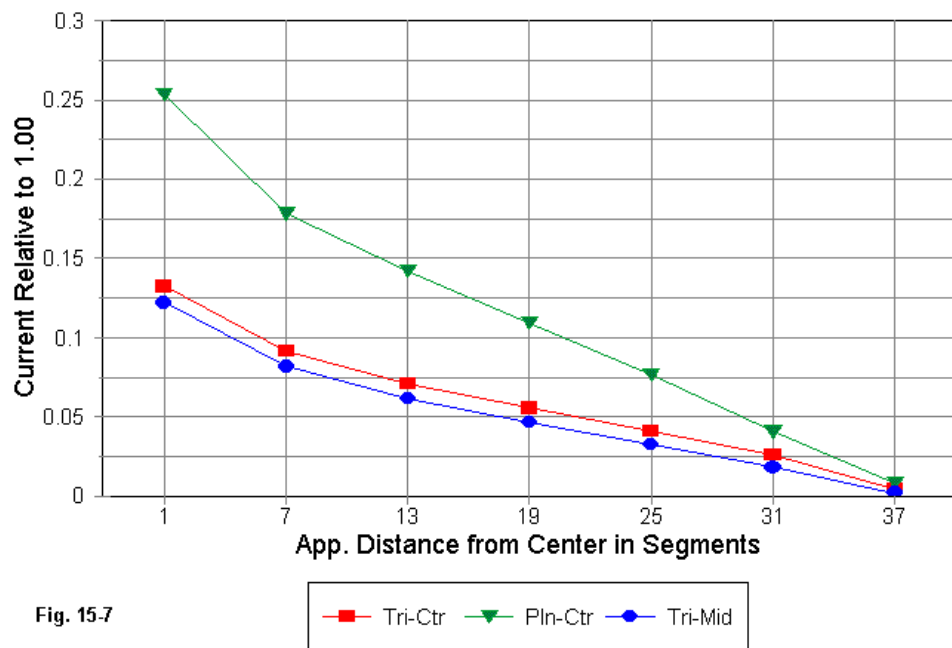
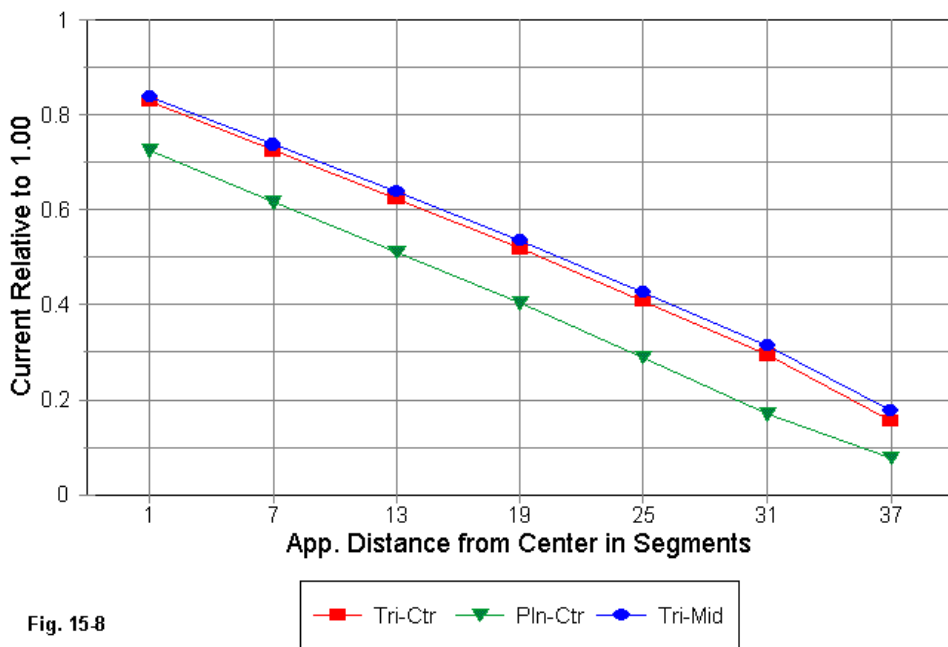


Fig. 15-7

Although the triangular models are physically set to minimize unequal influence from the main element, the current differential on the load wires is not zero. **Fig. 15-7** shows the two values of current differential to be almost precisely the same for the zero-loss wire models, with closely corresponding phase differences as well. With the exception of a 4.53" change of wire alignment, the two models are the same model in every operational respect. In the mid-element model, the so-called fed main element is actually one leg of the loading line, and the apparent second leg of

the loading line corresponds to the main element in the triangular center-loaded model.

Outbound Element Current Magnitude Parallel Linear Loads; Zero-Loss Wire



In contrast, the current differential between the load wires of the in-plane model is much higher. The higher differential is due to the differential interaction between the main element and the individual wires that are at different distances from the main element.

A further insight into the difference between the triangular and the in-plane models can be gathered from examining the currents on the main element outward from center. See **Fig. 15-8**. Note once more the close coincidence between the triangular models. In contrast, the current magnitudes along the outbound wire for the in-plane model are lower everywhere. Part of this reduction is due to the increased length of the in-plane linear load, which results in an electrical shift in the position of the beginning of the main element at the antenna's center.

As might be expected, the current distribution on the in-plane model results in a lower gain than for either triangular model. With perfect wire, both triangular models report an anticipated gain of 1.86 dBi, while the in-plane model gain is 1.79 dBi—not a big reduction, but noticeable. With copper loads, the triangular model gain is 1.35 dBi, while the in-plane gain is 1.30. An aluminum (6061-T6) loading wire set yields a gain of 1.11 dBi for the triangular models and 1.06 dBi for the in-plane model. The values for aluminum wire load lines (with perfect wire main elements) correspond closely to the values for center and mid-element load dipoles using R-L-C loads with a Q of 300 (1.02 and 1.11 dBi, respectively).

Equally revealing about the behavior of linear loads are SWR curves for the three models, each predicated on the source impedance at resonance at the design center frequency of 7.15 MHz. See **Fig. 15-9**. The resonant source impedances are 12.0, 12.1, and 12.2 Ohms for the triangular center, in-plane center, and triangular mid-element models. The coincidence of the mid-element linear load model source impedance with the center loaded models and not with the mid-element transmission line load model examined earlier is further confirmation that the mid-element model is simply a minor variation of the triangular center loaded model. The SWR curves confirm this further by overlaying each other, while the in-plane curve is slightly broader, as one might expect of a very slightly lossier structure. The mid-element linear load model nowhere approaches the broadness of the mid-element transmission line load we explored earlier.

SWR Curves: Triangular & In-Plane TL $Z_0 = 544$ Ohms; Zero-Loss Wire

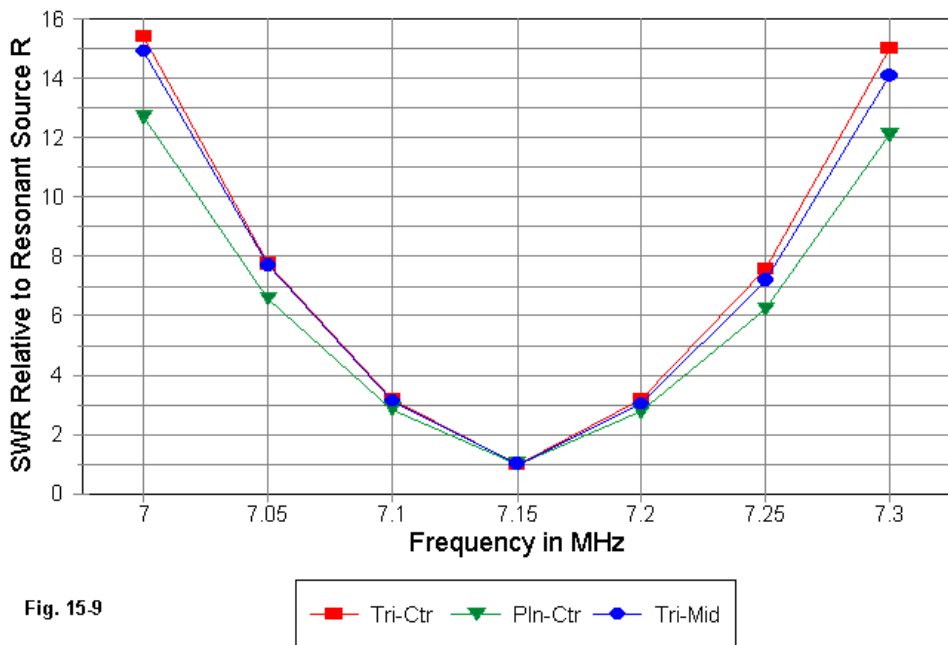
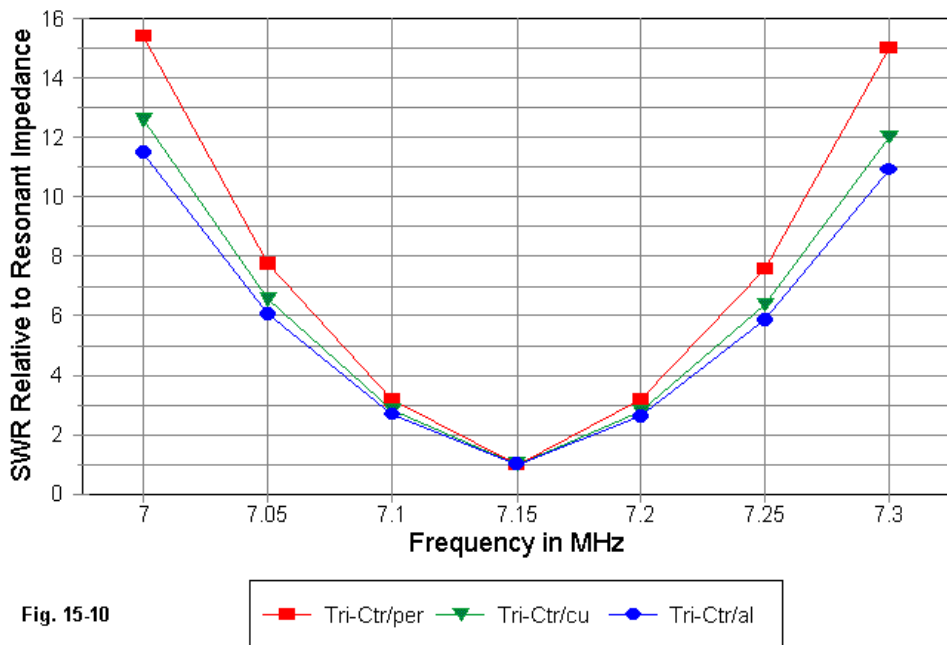


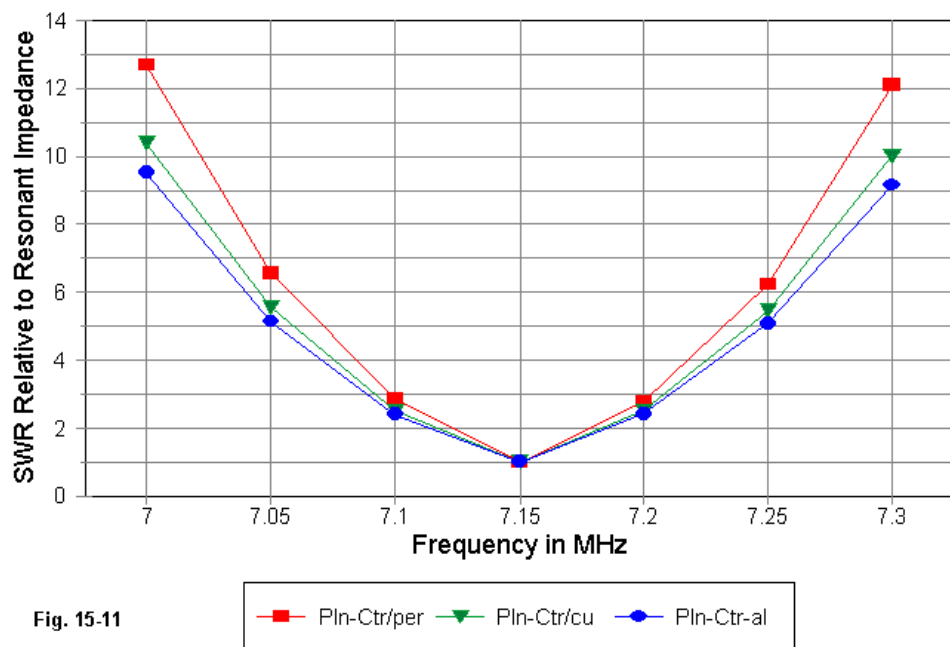
Fig. 15-9

The SWR curves for the triangular model for the three load line materials, shown in Fig. 15-10, show the expected broadening of the curve with the use of materials with a lower conductivity. Similar results accrue to the in-plane model, as shown in Fig. 15-11.

SWR Curves: Triangular Linear Load Perfect, Copper, & Aluminum Wire



SWR Curves: In-Plane Linear Load Perfect, Copper, & Aluminum Wire



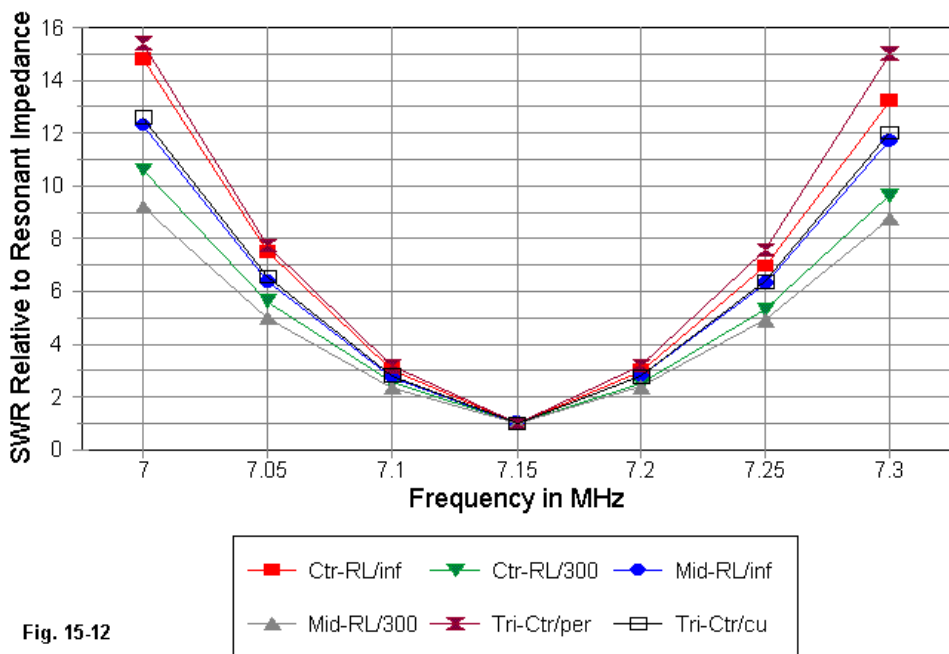
Conclusion

The linear load is essentially a shorted transmission line stub placed in line with the main element of the antenna it loads. As an antenna load, the stub cannot act as a pure transmission line. The current can never be equal in magnitude and exactly opposite in phase on the two wires making up the load line, since the current magnitude and phase on the antenna element into which it is inserted is not the

same at any two points along a path from the antenna center outward. However, the condition is most closely approximated when loads are as exactly at center as possible and symmetrically placed about the main element to equalize its influence on each load wire.

Since linear loads are transmission lines and answer in part to the equations for calculating length from required loading reactance—especially when placed at the center of the antenna structure—their operating bandwidth curves, as reflected in the SWR curves shown, are narrower than for corresponding R-L-C loads.

SWR Curves: Linear vs. RL Loads Triangular LL, RL Center & Mid-Element



The graph above plots curves for center and mid-element loads using perfect loads and loads with a Q of 300. Additionally, the graph records the SWR curves for the triangular linear load model using perfect and copper wire. Note that the curve for the copper wire linear load is sharper than any of the R-L-C load curves, except the center load with infinite Q . The broadest lines in the curve are for R-L-C curves with a Q of 300.

Because the operating bandwidth of a linear load rests in large part on transmission line stub behavior and not on solenoid inductor behavior, we cannot directly correlate gain reductions with operating bandwidth. Within each type of load, increases in bandwidth might point to increasing losses in the loads. However, cross-load type correlations are far less certain.

To the degree that the currents on the linear load transmission line stubs are unequal, the wires also contribute to the linear length of the antenna. If placed at right angles to the main element wire, the radiation appears as a cross-polarized field relative to the radiation from the main element. Moreover, the larger the current magnitude differential between load line wires, the more the wires act as linear contributions to the overall antenna length and the shorter the required stub.

As with any complex electrical structure inserted into an antenna and modeled physically, the modeling calculations cannot directly account for all phenomena. For example, some theory exists to suggest that transmission line wires have greater losses for a given wire size and frequency of RF current than that same wires in an electrical circuit, due to the effects of the intense field between the wires. However, these effects turn out to be small, even for the long linear loads used in the very short modeled dipole. MININEC tests and other marginally valid tests within NEC suggest additional total field antenna gain losses between 0.1 and 0.2 dB. These losses, if present, have no noticeable effect on the operating bandwidth of the linear-loaded antenna models.

While the bulk of the transmission-line view of linear loads has been previously established, presenting it here is useful if—for no other reason—than because it raises an interesting question. Just as linear loads have unequal current magnitudes at their junction with the main antenna element as a function of the current distribution along an antenna element, it would appear that solenoid inductors used as loads would share the same behavior. If this turns out to be the case, then the

mathematical specification of loads within NEC would be equally incomplete, despite the high trust antenna designers and analyzers place in R-L-C loads of Type 0 and Type 1.

It may be worth one more episode just to test the hypothesis and confirm (or disconfirm) the trust we place in Type 0 and Type 1 R-L-C loads.

* * * * *

Models included: 15-1 through 15-6. .EZ models in inches; .NEC and .NWP models in meters.



16. Notes on Reactive Antenna Loads and Their NEC Models

D. Some Solenoid Loading Basics

The final question in our inquiry is whether Type 0, 1, or 4 mathematical or non-radiating loads (each within their application limitations) fully or adequately specify inductively reactive loads used in a shortened antenna. Answering this question via antenna construction would seem natural, were it not for our customary building practices and presumptions. When we install a loading coil and it proves too large relative to the wire length, we simply shorten the wire to match the coil. Then we fail to interrogate the situation further to discover why the coil was too large. Most of the time, we assume that we miscalculated or that the usual coil equations are not sufficiently accurate to yield a coil that requires no subsequent wire pruning.

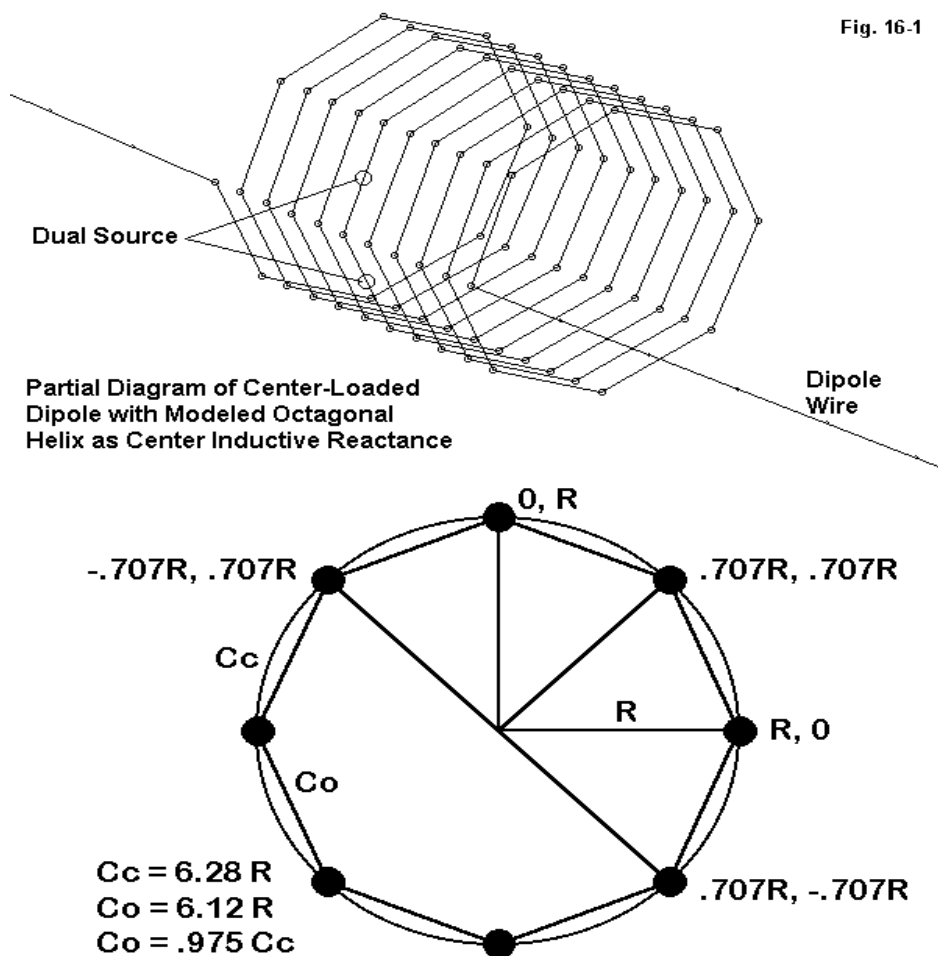
As a consequence, few modelers have explored the behavior of helically wound single layer solenoid inductors in their models—simply because winding a coil is too laborious a task relative to using the loading facilities built into NEC. We have presumed that the loads—especially Type 0 series R-L-C loads—adequately specify the loads we need.

Yet nagging suspicions remain about such loads. The current differential that is present at each end of the load suggests that—like their linear load counterparts—these loads do not act like pure inductances, but may also function as simple current bearing wires. Models of helical dipole elements show very usable gain, which parallels the experience of those who have constructed such antennas. I guess we shall not know whether mathematical models of loads accurately reflect the operation of single layer solenoid inductances unless we actually wind some model coils and replace the Type 0 loads with them.

Winding a Helix and Making a Model

The basic model for the half-size (1/4-wavelength) 7.15 MHz dipole uses #8 (0.125" diameter) zero-loss wire that is 412.68" (34.39') long and divided into 91 segments. Each segment is about 4.53" long. For least error, the segments within the helix will be as close to this value as helix construction permits.

Fig. 16-1



The required center loading reactance for the antenna is 833.3 Ohms or 18.549 microH. We may let the coil be 3 segments long or 13.6" overall. Using the standard Wheeler equations (either the common 1928 version or the 1982 version), a circular, single-layer, solenoid inductor of this length will require 10 turns and an 11.852" diameter.

The most reasonable approximation of a circular solenoid is an octagonal structure. **Fig. 16-1** shows the resultant structure and the basic relationships that determine the coordinates of the points around each loop. For this exercise, let's adopt the convention of extending the turns along the X-axis and using the Y-axis and the Z-axis for the radial points of each loop. Under this convention, the point references in the lower part of the drawing refer to Y and Z coordinates. Beginning, perhaps, with half the diameter as the value of R (5.926"), we might place the first entry at 0, 5.296, 0 (assuming a free-space modeling environment). Intermediate points between the horizontal and the vertical will require values of 0.707 R (3.744"). (Note: rounding to one less decimal place will make no significant difference in the long run.)

Each point progresses along the X-axis by the total length (13.6") divided by the number of turns (10), divided by the number of points in each turn (8), for a value of 0.17" in this case. Hence, the second point of the turn will be located at 0.17, 3.744, 3.744. We can then proceed to complete the first turn. Then, using whatever copy function may be present, we add progressively more turns, increasing the value for the X-coordinate by 1.36 with each turn.

The resulting solenoid will be about 2.5% short in overall wire length used compared to a truly circular solenoid, suggesting a slightly lower inductance than provided by the Wheeler calculation. Since the wire is small compared to the turns spacing, a slight additional variance may also be encountered. Of greater significance for the model, each wire in the helix is about 4.56" long, a close match to the length of the segments in the remaining portions of the dipole wires.

Once the coil is complete, it should be saved as a file, since we may use it later. For the initial test of center-loaded short dipole, we may replace the center 3 segments of the basic wire with the helix. The best coordinates for the coil may be -6.8 to + 6.8 on the X-axis to obtain a useful symmetry in the now enlarged geometry sheet—useful for visual error trapping. The formerly single wire is now two wires

extending to + and to - 206.343" respectively, and each with 44 segments. (Model 16-1)

```

GW 58 1 .06909 .14122 0 .07341 .09982 .09982 .00163
GW 59 1 .07341 .09982 .09982 .07772 0 .14122 .00163
GW 60 1 .07772 0 .14122 .08204 -.0998 .09982 .00163
GW 61 1 .08204 -.0998 .09982 .08636 -.1412 0 .00163
GW 62 1 .08636 -.1412 0 .09068 -.0998 -.0998 .00163
GW 63 1 .09068 -.0998 -.0998 .095 0 -.1412 .00163
GW 64 1 .095 0 -.1412 .09931 .09982 -.0998 .00163
GW 65 1 .09931 .09982 -.0998 .10363 .14122 0 .00163
GW 66 1 .10363 .14122 0 .10795 .09982 .09982 .00163
GW 67 1 .10795 .09982 .09982 .11227 0 .14122 .00163
GW 68 1 .11227 0 .14122 .11659 -.0998 .09982 .00163
GW 69 1 .11659 -.0998 .09982 .1209 -.1412 0 .00163
GW 70 1 .1209 -.1412 0 .12522 -.0998 -.0998 .00163
GW 71 1 .12522 -.0998 -.0998 .12954 0 -.1412 .00163
GW 72 1 .12954 0 -.1412 .13386 .09982 -.0998 .00163
GW 73 1 .13386 .09982 -.0998 .13818 .14122 0 .00163
GW 74 1 .13818 .14122 0 .14249 .09982 .09982 .00163
GW 75 1 .14249 .09982 .09982 .14681 0 .14122 .00163
GW 76 1 .14681 0 .14122 .15113 -.0998 .09982 .00163
GW 77 1 .15113 -.0998 .09982 .15545 -.1412 0 .00163
GW 78 1 .15545 -.1412 0 .15977 -.0998 -.0998 .00163
GW 79 1 .15977 -.0998 -.0998 .16408 0 -.1412 .00163
GW 80 1 .16408 0 -.1412 .1684 .09982 -.0998 .00163
GW 81 1 .1684 .09982 -.0998 .17272 .14122 0 .00163
GW 82 44 .17272 .14122 0 5.24111 0 0 .00163
GS 0 0 1 (Coordinates in Meters)
GE 0
EX 0 41 1 0 .707107 0
EX 0 42 1 0 .707107 0
LD 5 2 1 1 5.8001E7
LD 5 3 1 1 5.8001E7
LD 5 4 1 1 5.8001E7

```

Fig. 16-2

Part of the extended exercise in looking at NEC loads, both mathematical and physical, has involved sorting out the losses attributable to the load from those which may be associated with the linear wire portion of the antenna. For this purpose, we shall need to make the linear-wire lossless and the coil-wire of a suitable material, namely, copper. At present, only Nittany-Scientific software among commercial implementations of NEC permits a wire-by-wire material selection. **Fig. 16-2** shows a partial .NEC file, with the dimensions in meters.

The portion of the model shown shows the last few turns of the helix, wire by wire. The last wire (82) is the linear dipole wire in the +X direction. Skipping down to the Type 5 load lines (material loading), we see that wire 1 (and also 82) is skipped, being lossless, while the conductivity value for copper appears for each wire in the helix.

Finally, because the turns of the helix have an equal number of wires per turn, a split feed is used at the wires forming the very center of the inductor.

The Center-Loaded Half-Length Dipole

The model just constructed in a free space environment yielded an inductive reactance of over 86 Ohms, indicating not that our octagonal coil was slightly small, but rather that it was considerably too large. Achieving resonance required a significantly smaller helix. In order to preserve the length of the coil that was equal to 3 segments of the linear wire, the coil was reduced in diameter. With a diameter of 11.12" the loaded antenna achieved resonance (as defined in this study as less than $\pm j 1$ Ohm source reactance). Wheeler calculations for a circular coil of this diameter, length, and number of turns yielded an inductance value of about 16.6 microH, roughly 10% less than the value called for by the mathematical load systems provided in NEC. The length of the coil wires reduces to about 4.25" each, which is unlikely to induce any significant error in the model.

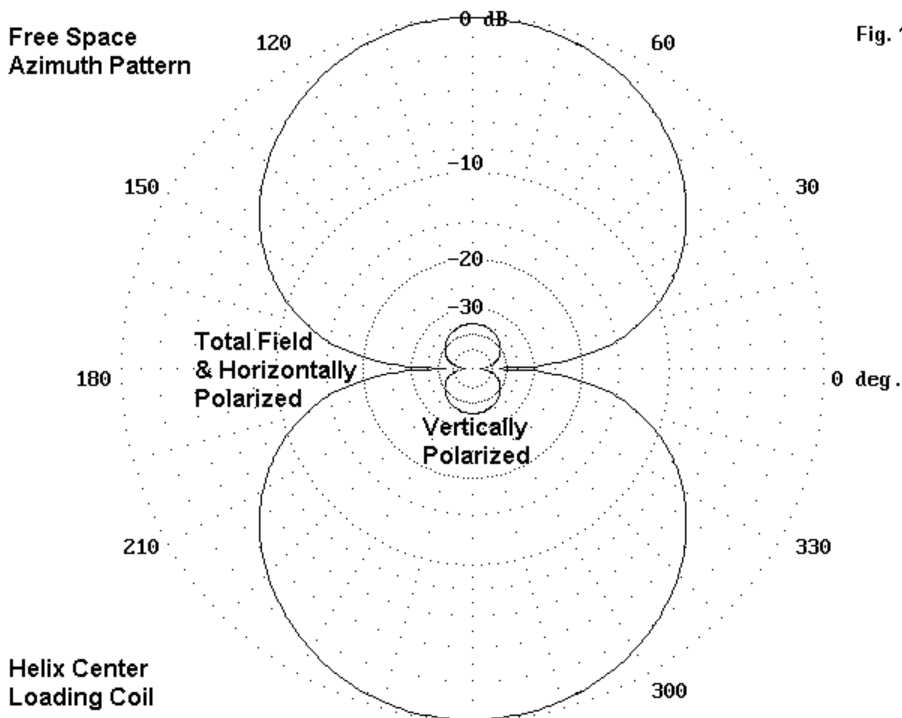
The following table provides some basis for making comparisons between the physically modeled load and the Type 0 loads used in earlier parts of this exercise. In all cases, the linear portion of the antenna is composed of lossless wire.

Load	Free Space Gain dBi	Source Impedance R +/- jX Ohms
Lossless Type 0	1.85	13.19 - j 0.02

Q=300 Type 0	1.02	$15.97 - j 0.01$
Copper Helix	1.35	$10.61 + j 0.64$

The higher gain provided by the physically modeled center-load should not be surprising, considering that the helix contains wire that radiates. See **Fig. 16-3**.

Fig. 16-3

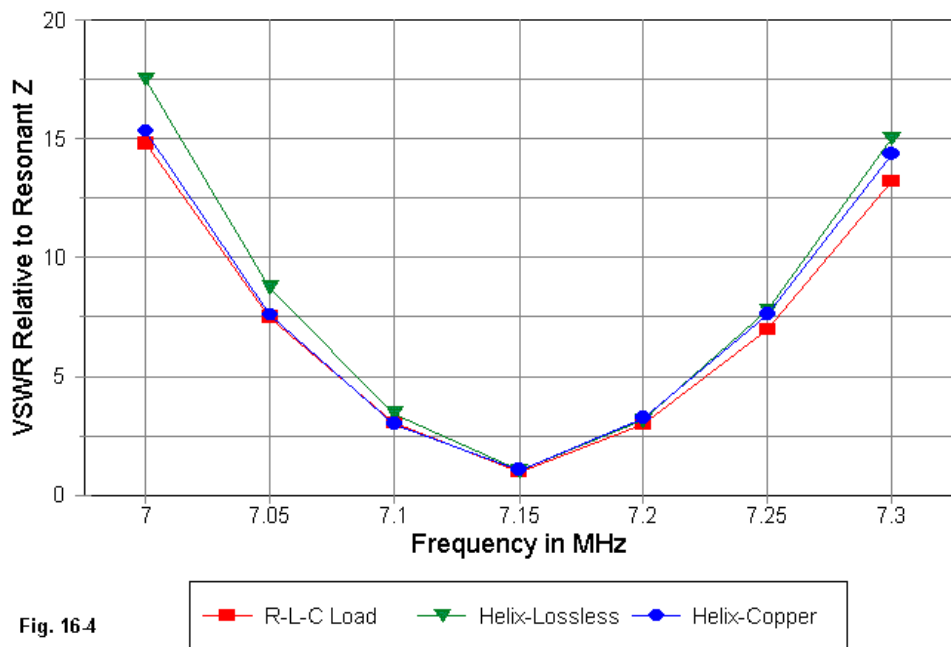


The small but definite vertically polarized radiation field results from radiation from the portions of the helix that are vertically oriented. This field alone would not suffice to provide the added gain over the Q=300 center-loaded model, but the

entire 13.6" length of the inductor would. More precisely, that portion of the overall inductor wire beyond the 13.6" it replaces in the overall antenna length would suffice to provide additional gain.

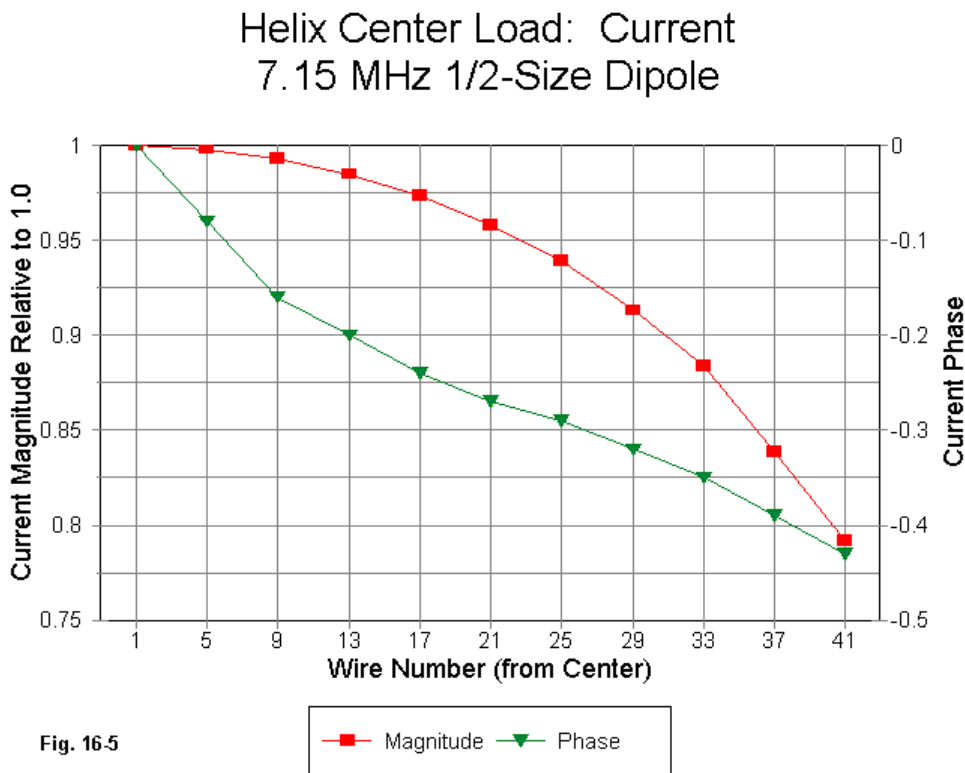
The actual Q of the inductor cannot be ascertained with accuracy. Some calculations estimate the Q at over 400, but only under the conditions that the wires be closer to half the spacing between turns. The modeled helix uses much smaller diameter wire, resulting in a decrease in Q, perhaps to the 200-300 range at best.

Helix Center Load: SWR 7.15 MHz 1/2-Size Dipole



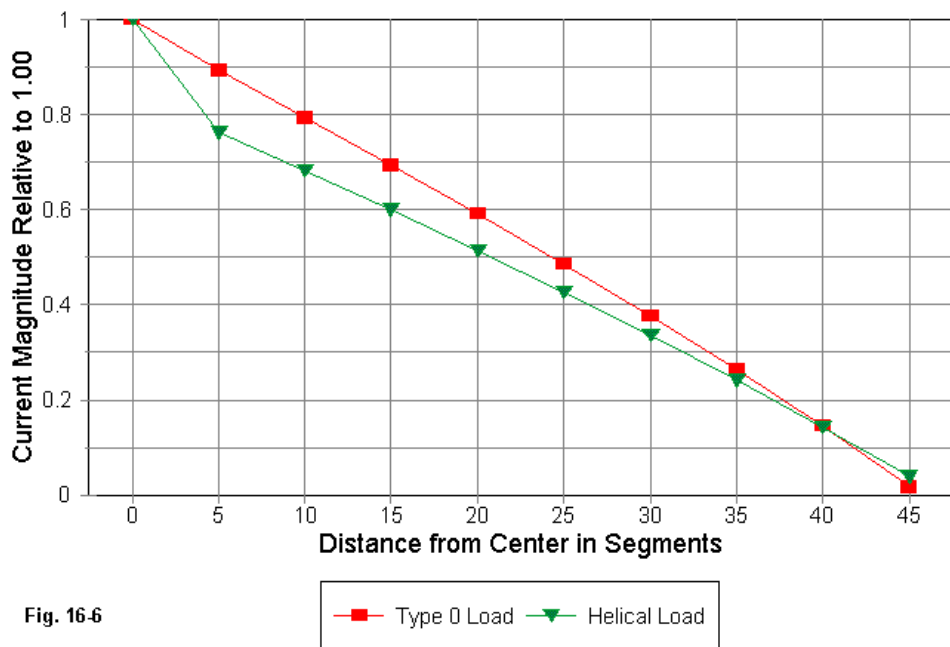
Despite the lower Q estimate, the source impedance of the physically modeled helical loaded dipole is lower than even the Type 0 model at zero load loss. As **Fig. 16-4** shows, the SWR curve is slightly steeper than even the R-L-C lossless load model—and a lossless helix yields a steeper curve yet.

The actions of the physically modeled helix load are in part a function of the current magnitude and phase profile along the coil. **Fig. 16-5** shows the two curves from the center point outward toward the end.



The current magnitude differential between the source and the coil end is approximately 20% of the maximum value. Since the inductance of a helix is in theory predicated upon equal current magnitudes at both ends of the coil, the differential currents dictate that in part the inductor does not act solely as an inductor. The physical model of the center load provides indications that the helix also radiates in the manner of any length of wire included within the total length of the dipole—in terms of gain over the Type 0 load, the vertical radiation field, and the reduction in the required inductance to effect loading to resonance.

Current Profiles: Center Loads Type 0 vs. Helical Loads



There are potential error sources within the model. First, the octagonal inductor with less than optimal wire diameter relative to turns-spacing casts some doubt on the precision of the calculations relating inductance to turns, diameter, and length. However, most of these factors suggest that the octagonal coil's inductance would be low, when in fact, the model showed it to be very much higher than required for resonance.

Second, the helix design, although apt to the modeling task, is very large and open compared to coils one might ordinarily apply as center loads in a shortened dipole. However, as an error source, the helix configuration would be a matter of degree and not of fundamental principle of operation. What the dimensionless load of a Type 0 load does not show is the relatively large early drop of current magnitude along the total length of the dipole. **Fig. 16-6** shows the more dramatic drop occasioned by the physically modeled coil. Although different coil configurations may alter the amount of drop somewhat, the current magnitude curve is unlikely to resemble that for the Type 0 load.

The plausibility of the physical model might be put to still another test. The original short dipole model was designed to permit one to place load values equal to the center load value at specific mid-element points. If the center load helix as modified to achieve resonance acts like a Type 0 load, then placing the same coils at the specified points should also result in resonance.

The Mid-Element-Loaded Half-Length Dipole

The original model was modified into a 5-wire model having the same operational characteristics as the original. The inserted 3-segment wires were placed at points between 99.5" and 113.1" from the center so that the Type 0 load was positioned 106.3" from the center. At these positions either side of center, a load of 833.3 Ohms (18.549 microH) resonated the antenna. The value of these two loads was the same as the value of the single center loading reactance.

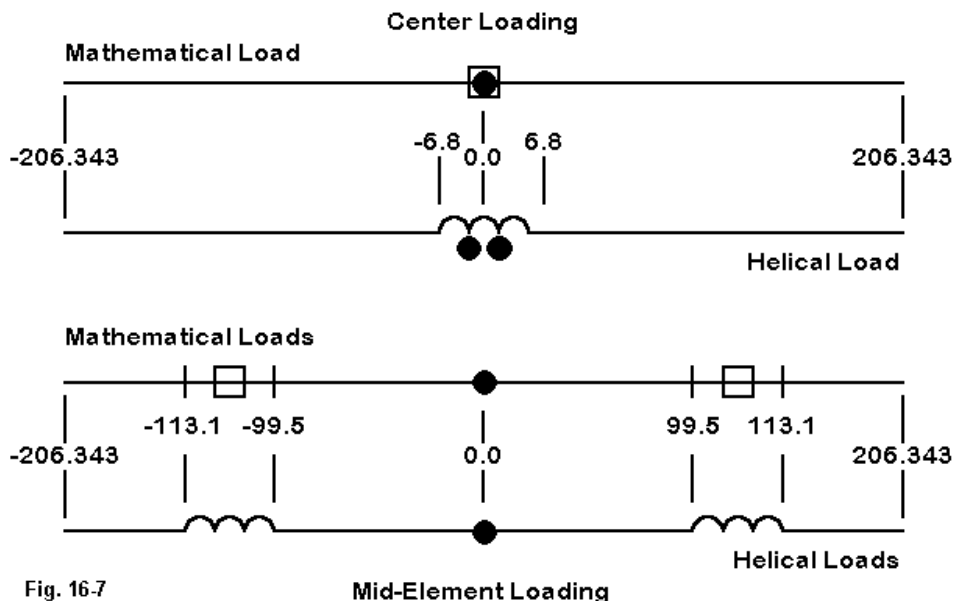


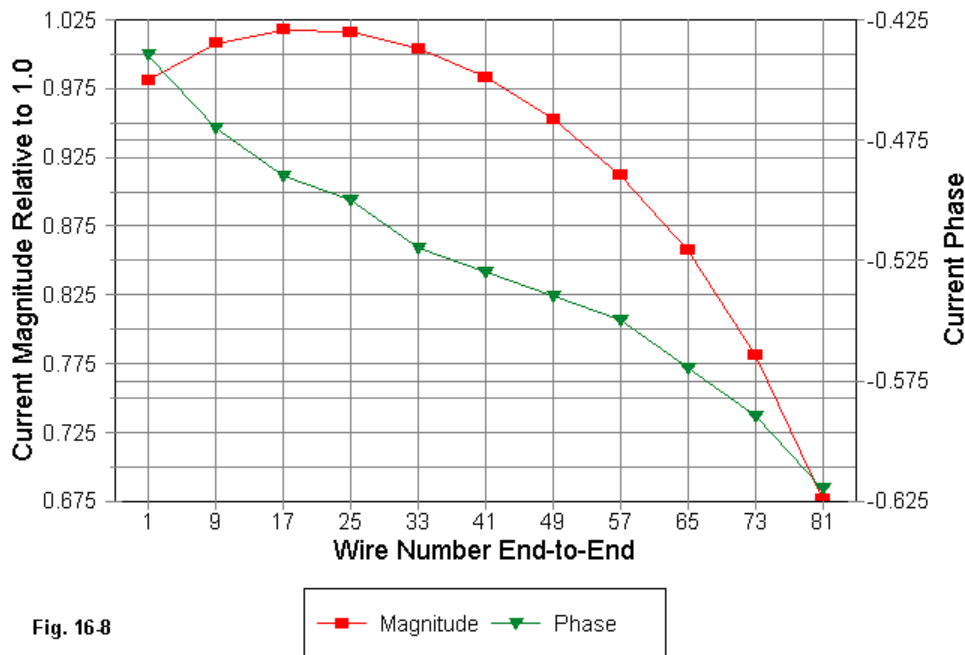
Fig. 16-7

As shown in **Fig. 16-7**, the helix developed for the center load may directly replace the inserted 3-segment wires in the 5-wire model. By copying the original helix and changing the values for all of the X coordinates by the same amount, each coil may be placed in the correct position. The connection points to the coils are very slightly displaced relative to the axis line, but not by an amount to show up in a movement of the antennas resonant frequency. The 11.12" diameter ($R = 5.56"$) helix was copied and transferred to these points and the model run.

The result was a source impedance of about $29 + j 229$ Ohms, indicating that the inductor size (16.6 microH) was much too large for resonance. This result was consistent with but considerably larger than expected. To create helices that allowed the overall antenna to achieve resonance required a reduction in the coil

diameter to 10.2" ($R = 5.10''$). By Wheeler calculations, the inductance was reduced to about 14.3 microH for a truly circular coil. The wire length from coil point to coil point diminished to 3.90" compared to linear wire segment lengths of 4.53" each.

Helix Mid-Element Load: Current 7.15 MHz 1/2-Size Dipole



The amount of reduction in coil size was about 14% relative to the final center loading coil and almost 23% relative to the inductance required by a Type 0 load. This reduction is consistent with several factors involved in inductive loading, including the fact that the further out along an element one places the inductor, the less it acts as an inductor and the more it acts as a compact form of wire length to increase the overall wire length to resonance. By the outer ends of a shortened dipole, helices and others forms of element extension cannot be viewed as inductances almost at all. See **Fig. 16-8**.

The current magnitude and phase profile within the helix may be of some surprise to some. Along the mid-element loading helix, after size adjustment to resonance, the current does not show a steady progression lower in magnitude. Instead, there is a current peak about a fourth of the way outward from the inboard end of the coil. From that point onward, the current magnitude descends rapidly. The phase transition, although not linear, generally follows the progression along the entire antenna element.

Compared to the center-loading coil, the current magnitude excursion within the inductor is about 34% of the maximum value, and the end to end difference is about 30% of maximum current magnitude. This compares to the 20% differential from center to end for the center-loading coil. The greater differential of current magnitude along the mid-element inductor wire is another indicator that it is functioning to a greater extent than the center loading coil as part of the element length and hence requires less inductance and inductive reactance to effect loading to resonance.

Like the physically modeled center load, the physically modeled mid-element load shows a much larger drop in current at its outer end than does a corresponding Type 0 load. The dramatic decrease in current immediately outside the loading helix appears vividly in **Fig. 16-9**, with the Type 0 load curve shown for comparison.

Despite the close correspondence of current up to the load position and the dramatically lower current beyond that point for the physically modeled helix, the model shows a significantly higher gain than either the corresponding Type 0 load model or even the physically modeled center load. The table following **Fig16-9** summarizes the gain situation for various models.

Current Profiles: Mid-Element Loads Type 0 vs. Helical Loads

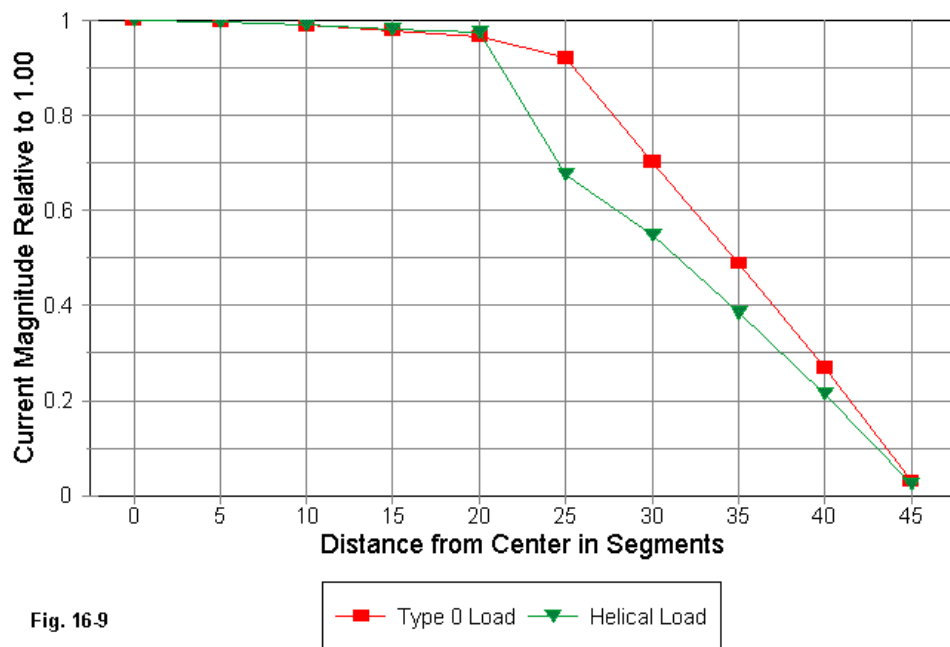


Fig. 16.9

Antenna	Free Space Gain dBi	Gain Relative to a Full Length Dipole
Full-Length	2.14	—
Half-Length Lossless Wire	1.85	-0.29 dB
Type 0 Center Load, Q=300	1.02	-1.12 dB
Helical Center Load, Copper	1.35	-0.79 dB
Type 0 Mid-Element loads, Q=300	1.11	-1.03 dB
Helical mid-Element Loads, copper	1.71	-0.43 dB

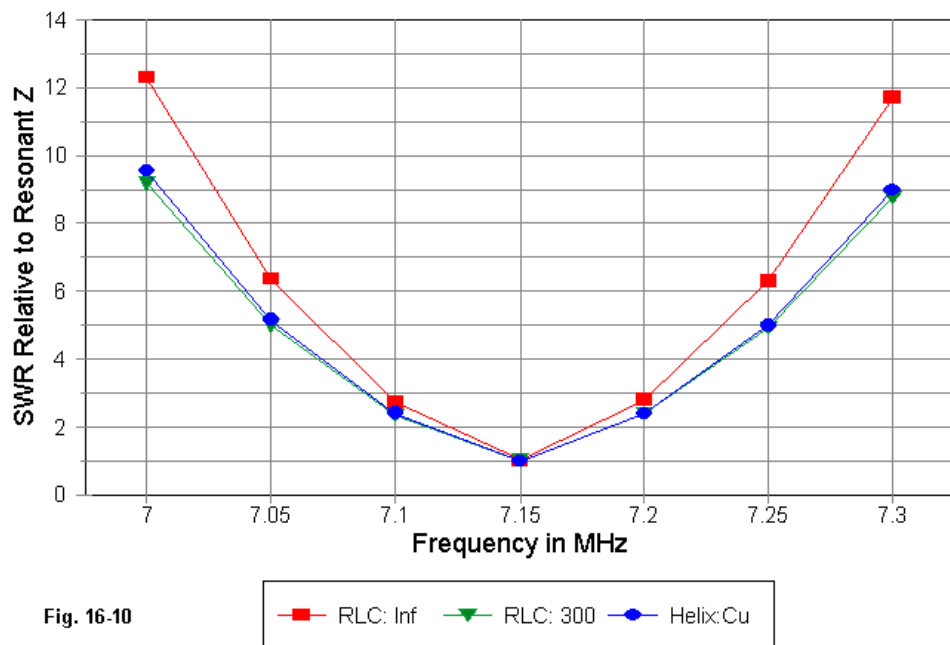
The gain figures strongly suggest that the current within the turns of the helix is not going wholly to waste within a tightly coupled inductive field. That part of the current not devoted to the field, as indicated by the current differential at the coil ends and among the turns is contributing to the total field of the antenna.

The use of Type 0 loads to characterize both center and mid-element loading strongly suggests that there is little difference between the gain potential of two antennas of equal length loaded in each manner. However, physically modeled loading helices tell a quite different story. Not only do both types of loaded antennas potentially perform better than mathematically inserted loads would indicate, but as well the empirical experience giving the nod in gain to mid-element loading gains some support from the physically modeled loading coils.

Although the physically modeled center-loaded antenna showed a steeper SWR curve than its Type 0 counterpart, the mid-element loaded antennas show a good correspondence between mathematical and physical loads. In fact, as shown in **Fig. 16-10**, the mid-element physical helices correspond nicely to mathematical loads with a Q of 300. This curve corresponds reasonable well with estimates for the Q of the coil at or just below 300, based upon reductions from the calculated Q of over 400 that presumed an ideal wire diameter-to-spacing ratio. The resonant source impedance of about 24.6 Ohms is a bit lower than the Type 0 model at Q=300, where the source impedance was about 31.9 Ohms.

It should be noted for accuracy and completeness that the actual resonant frequency for the model using physically modeled mid-element coils with a 10.2" diameter was 7.16 MHz, and that the curves for this antenna actually run from 7.01 to 7.31 MHz. This 10 kHz displacement did not seem great enough to warrant reconstruction of the coils through the iterations necessary to move the resonant frequency. The estimated diameter would have been something of the order of 10.205" +/- 0.002" for resonance at 7.15".

SWR Curves: Mid-Element Loads 7.15 MHz Half-Size Dipole



Potential Errors and Conclusions

The potential error sources for the mid-element load model remain the same as for the center-loading coil. The octagonal helices may in fact show a lower inductance than the values calculated by the Wheeler equations. Moreover, the widely spaced open helix may prove a better radiator than some more tightly wound loading coils. Unique to the modified mid-element loading coils is the shortening of the

wires and a somewhat greater differential to the lengths of the segments in the linear wire portions of the antenna.

Some of these possible error sources work in the wrong direction relative to the results, for example, the reduced inductance of octagonal helices relative to truly circular ones. The sum of the remaining error sources appears to be far short of what would be required to account for the radical reduction in required inductance for the loading coils. Indeed, the most telling part of the demonstration is the further reduction in coil size required in moving the center-loading inductor out to the position where the same value should have served as a mid-element loading inductor. Although absolute values of the inductors involved may be off the mark by some small amount, the trends are in all probability accurate reflections of anticipated antenna performance.

As a consequence of these data and considerations, the bottom line seems inevitable: Type 0 (and by extension, Type 1 and Type 4) loads do not fully characterize the performance of inductively reactive loads in the form of single layer sole-noids. For many purposes, that characterization may be adequate to a design task. For example, where design criteria permit adjustment of the antenna outer section length, designing to the Type 0 load value may cause no harm.

However, using mathematical loads alone can give a potentially misleading impression of the gain of the loaded antenna. The gain for physically modeled loads is considerably higher than the gain for mathematical loads. Where multi-element arrays—such as Yagis—may employ center or mid-element loads in all elements, Type 0 load models may seriously underestimate the overall array gain.

This exercise, along with the previous portions of this study, should be viewed as only a beginning to the investigations into modeling loaded shortened antennas. Further work might yield additional insights into the behavior of linear loads examined in the last episode. Likewise, for helical inductive loads, much remains to be done with respect to the use of more compact forms of inductance. Even using these models, additional data can be gathered for the use of linear wires with non-zero losses and for the use of other helix materials, such as aluminum.

Nonetheless, if the present demonstration holds up as reasonably valid, then the suspicion raised at the outset has been established. In several important re-

spects, the non-radiating mathematical loading features provided within NEC do not fully characterize the action of these loads when physically modeled. Nevertheless, modeling actual loading inductors that form junctions with elements of radically different diameter from the coil wire and modeling small-diameter coils without violating adjacent segment length restrictions may prove to set major challenges for the amateur and the professional modeler. This study was carefully designed to use models that stayed well within all NEC restrictions. Most real antennas that use inductive reactance to electrically lengthen a physically short element will press those restrictions or result in quite large models.

In other words, what we can establish in principle in a straightforward manner might prove to be an exceptional modeling challenge when real antennas and their parts are involved. Loads are not quite as simple as casual modeling might lead us to believe.

* * * * *

Models included: 16-1 through 16-6. .EZ model dimensions in inches; .NEC and .NWP model dimensions in meters.



17. Notes on Reactive Antenna Loads and Their NEC Models

E. Some Unfinished Business on Modeling Loads

In the first four sections of this series on inductively reactive loads used to electrically lengthen physically shortened antennas, we uncovered some interesting disparities between the standard mathematical or non-radiating models of loads and loads that are physically modeled. If the demonstration models are valid, then the following summary points are true in principle.

1. Because the current magnitude and phase in a transmission line used to load a short antenna do not show a true transmission line relationship in a physical antenna, the use of the TL facility in NEC is not a reliable model of the real line. The modeler should confirm the results with a physically modeled transmission line load.

2. Linear loads are a variety of transmission line stub load, but their behavior should be physically modeled for greatest reliability, especially when folded back from a mid-point along the so-called main element. Within the length of the linear load, which wire comprises the “main” element is ambiguous, since the radiation from that portion of the antenna is a function of the closely coupled fields from the element and the load wires in the stub.

3. To a lesser but still significant degree, physical models of single layer solenoids, when modeled as helices, depart from the Type 0, 1, and 4 NEC mathematical models of them when placed as loads along an antenna element. In general, such loads show composite properties as inductances and as added wire length, the latter of which properties contributes to the radiation field of the antenna. The effect increases as the inductors are moved outward from the center of a dipole, with less inductance being required for the load than for a center loading coil. Even a center-loading coil requires less inductance in physical models than called for by the NEC mathematical load.

Although the exercises have demonstrated the effects, they fall far short of establishing a set of universal principles, mostly due to the limitations of NEC. Some of

the potential error sources in NEC-4.1 have been reviewed in "NEC-4.1: Limitations of Importance to Hams," *QEX* (May/June, 1998), pages 3-16.

Throughout the exercises, NEC-4.1 was used in the EZNEC Pro and the Nittany-Scientific GNEC implementations. In all models, unless otherwise specified, the linear wire was modeled as lossless in order to sort out losses and performance curves due to the load from those attributable to the linear wire. Hence, all gain figures in previous exercises will be higher than with any real antenna using copper or aluminum elements. All models were constructed in free space.

The model short dipole for 7.15 MHz was 1/2 full size at 34.39' (412.7") long and used #8 AWG (0.1285" diameter) wire. All physically modeled load structures used the same wire in order to avoid possible errors resulting from angular junctions of wires having dissimilar diameters. The original antenna was assigned 91 segments, each about 4.53" long. The selection of segment length served a number of purposes. First, it set the spacing of linear load wires from each other at a reasonable distance for accurate modeling. Second, it permitted the construction of octagonal single layer solenoid inductors that were exactly 3 segments long and whose wire lengths for each portion of each turn were close in length to the adjacent segments of the linear wire.

The original model was revised for mid-element loading exercises into a 5-wire model. I placed 3-segment wires on each side of the center wire so that the center segment of the insert occurred exactly where a mid-element load was equal in size (reactance and inductance) to a corresponding center-load. Either loading system brought the antenna to resonance with no change in the overall length of the dipole. The overall segmentation was not changed, so that segment lengths among models remained close to equal. A 3-segment long (13.6") physical inductor model replaced the center 3 segments of the model for center loading and replaced the two inserts for mid-element loading tests.

Subject to continuing review, it is believed that these models fall well within the limitations of NEC-4.1's ability to provide accurate results. It should be unnecessary to add that the relationship of the NEC-4.1 modeling system to real antennas is not like the relationship of a child's plastic toy to a real automobile. Rather, NEC-4.1 (and other versions of NEC and MININEC) are complex mathematical systems based upon both fundamental antenna theory and fundamental mathematical principles

for calculating the various performance parameters of many types antennas. Those who dismiss antenna modeling in general largely reveal only a general lack appreciation of the role of mathematics in both antenna theory and antenna practice.

Those who develop, implement, and work with any version of NEC are also conscious of its limitations. All versions of NEC have been extensively tested against experimental results for many fundamental antenna types. However, it is possible to create antenna geometries that exceed the limitations of the calculation system. In such cases, NEC yields results that do not tally with either theory or practice. Since no program developer or development team can fully predict all of the antenna geometries (or environmental circumstances) that users will model, they can only enumerate some of the program limitations. Users will discover other limitations in the course of their work, and these limitations set the challenges for further development.

This note is about some of the limitations that apply to the extension of the models used in this exercise set to various real antenna designs.

Linear Load Limitations

The difficulty of extending the models used for linear loads in these notes to real antennas can best be illustrated by taking up a few concrete examples of extant antennas.

Fig. 17-1 shows a sketch that outlines the basic design features of one commercially manufactured antenna element using a linear load. To this point, I have been unable to construct a model of this element that falls within NEC limitations.

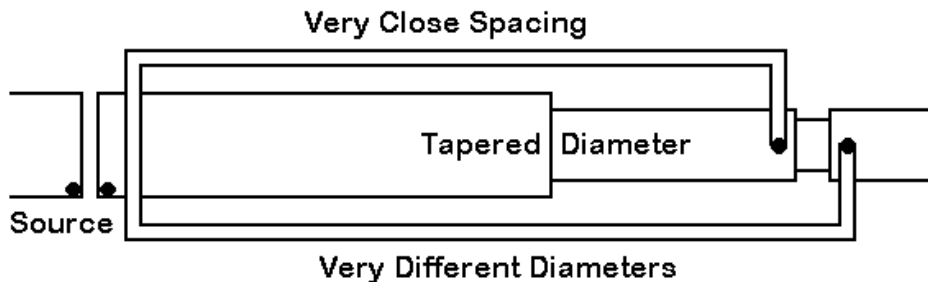


Fig. 17-1

Tightly Spaced Mid-Element Linear Load

The element uses a series of tubes of decreasing diameter as one moves from the center point outward. The NEC-2 limitation with tapered diameter elements is largely, but not completely, overcome in the NEC-4.1 algorithms. Modeling this element in NEC-2 with Leeson corrections (or, more properly, conversion of the element into its equivalent uniform diameter equivalent) is not possible, since the linear load junctions would block correction implementation (and the implementation of the correction would be inaccurate, since the element is not continuous).

In this particular design, the linear load wires are very thin compared to the center element, and they are very closely spaced to the center element. All versions of NEC yield erroneous results when wires of different diameter are closely spaced. Moreover, there is a right-angle junction of wires of different diameter, another case in which NEC yields erroneous results.

A second design of commercial origin appears in **Fig. 17-2**. In this case, the previously noted limitation involving angular junctions of wires having dissimilar diameters also applies to this design, since the load wires are very much thinner than the center element tubing. The tubing also uses a diameter-tapering schedule, adding a definite limitation to modeling with NEC-2 and a minor limitation for NEC-4.1.

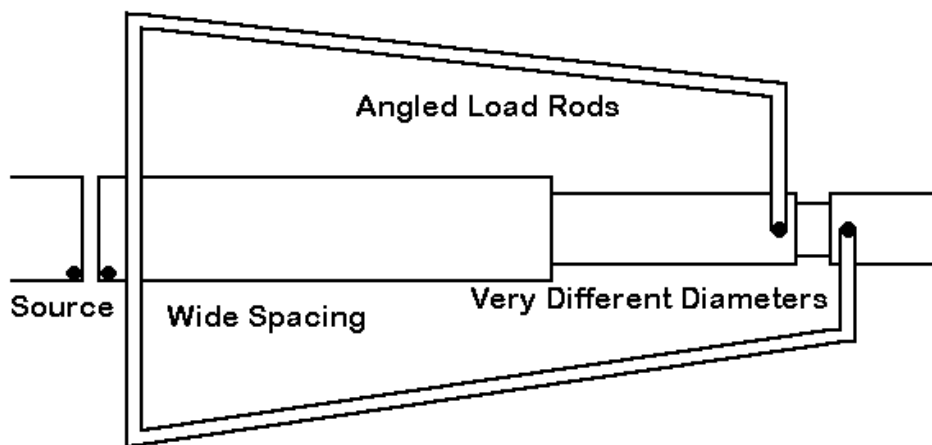


Fig. 17-2

Widely Spaced & Angled Mid-Element Linear Load

The wires of the load are not parallel to each other. This feature would make the correlation of the load to transmission line stubs more complex, although for general purposes, the averaged characteristic impedance of the line might be used with fair accuracy. However, to the degree that physical models of parallel transmission lines yield results at variance from standard calculations, determining the characteristic impedance of the physically modeled line will prove more difficult.

A third feature of this design is the placement of the long shorting bar near the center of the element. Physical models of shorted transmission line stubs take into account the current magnitude and phase of the shorting bar. In this instance of a long bar, the bar itself becomes an integral part of the line, one having a rapidly decreasing spacing between opposing segments. The design also may place the shorting bar at a different distance from the element center point than the value used in the exercise models. Hence, the correlation drawn in the exercise between mid-element loads folded back to the center and loads properly called center linear loads may be less certain for this case.

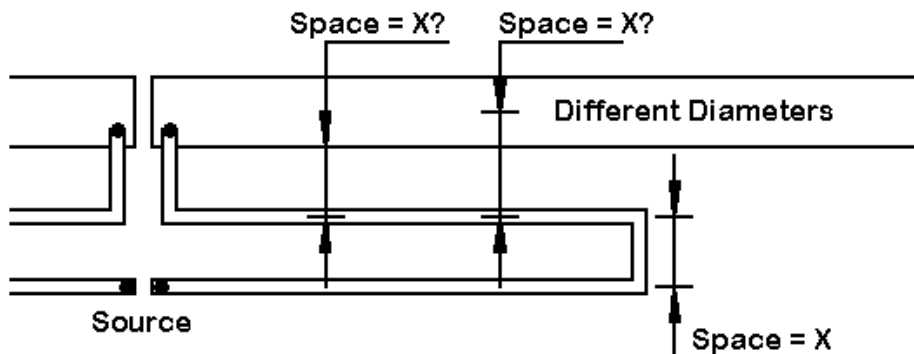


Fig. 17-3

In-Plane Construction

We can illustrate the difficulties of modeling a real linear load in NEC by looking once more, in **Fig. 17-3**, at the model of the in-plane center linear load, discussed in part C. In that model, using #8 wire uniformly throughout the model, several facets of real antenna construction were simplified. The spacing between the three wires was set at 4.53" center-to-center. This spacing produced identical surface-to-surface spacing values that were 0.1285" less than the center-to-center spacing.

If we increase the "main" element to a 1" diameter, center-to-center spacing values no longer correlate to surface-to-surface spacing values, changing the relationship among the elements. In addition, we encounter the potential problem in NEC of closely spaced wires with different diameters.

Fortunately, this model just barely fits within the MININEC segmentation limit and can be evaluation on that system. MININEC does not show a limitation with respect to closely spaced wires with different diameters. However, MININEC models tend to clip corners with angular connections, and we should expect (without invoking length-tapering) some required change in the length of the loading wires if we hold the overall length constant.

The following table shows the results of modeling the in-plane antenna in both NEC-4.1 and in MININEC. Load line lengths (from center to short) are given to indicate the require modifications to achieve a value close to resonance (in this demonstration, loosely defined as under $\pm j 5$ Ohms). In addition, values are shown for models using lossless, copper, and aluminum wire throughout.

Main Element	MININEC		NEC-4.1	
	Free Space Gain dBi	Feed Z R+/-jX	Free Space Gain dBi	Feed Z R+/-jX
#8 AWG	(Load = 194.5")		(Load = 188")	
Lossless wire	1.85	11.9-j1.2	1.79	12.1-j0.0
Copper	1.30	13.5-j1.2	1.25	13.7+j1.5
Aluminum	1.04	14.3-j1.2	0.99	14.5+j2.4
1"	(Load = 158")		(Load = 165.3")	
Lossless wire	1.85	11.9+j4.5	1.25	13.7-j0.5
Copper	1.39	13.2+j4.5	0.83	15.1+j0.9
Aluminum	1.17	13.9+j4.5	0.62	15.9+j1.6

In the table, gain figures are shown to 2 decimal places in order to show numerical trends that would have been erased by excessive rounding. (For operational purposes, figures to a single decimal place would suffice for many comparative cases, and the decimals might be dropped from certain generalized discussions. However, those who believe that showing gain to multiple decimal places should be dropped in all cases simply lack an appreciation for all of the purposes for which the antenna modeling programs can be put to effective use. Each degree of precision in reportage has its proper context. My own preference would be to see modeling reports overly precisely reported and then to do my own rounding rather than to see such figures pre-filtered.)

The gain and impedance figures for the #8 model show excellent correlation between the modeling systems, with a maximum gain variance of 0.05 dB and a maximum resistive variance of 0.2 Ohms. If we look at the MININEC column and read downward, then the reported performance of the 1" main element version of the modeled antenna shows very sensible results. The lossless version of the model shows minuscule differences from the #8 version (actually none in the truncated

decimals used here). The copper and aluminum versions of the 1" version show increases in gain over the #8 version as the material becomes more lossy. This is a function of the increased surface area of the larger element, and a similar trend shows up in full size dipoles. Likewise, the feedpoint impedance of the 1" antenna decreases more rapidly from the #8 version as the material loss increases, again for the same reason of lower overall losses. Once more, this trend also shows itself in unloaded antennas.

Regardless of the absolute accuracy of the MININEC figures, the trends in numbers are not only intuitively sensible, they correlate with the behavior of models of unloaded antennas. In contrast, the NEC-4.1 figures for the 1" diameter main element model report decreases in gain and increases in feedpoint impedance for each material assignment. These numbers indicate either that the wire spacing for the element sizes involved and the frequency of test (7.15 MHz) have crossed the threshold at which NEC-4.1 no longer delivers accurate results, or that the angular junction of the #8 wire with the 1" wire is yielding less than accurate results—or both. It is common in either error mode for NEC to report a gain that is too high/low while reporting a source impedance that is too low/high.

Allowing for corner junction errors, MININEC remains the program of choice for modeling complex geometries that involve junctions of wires having dissimilar diameters and/or closely spaced wires of differing diameters. Recent developments in MININEC to break the segment limit, such as in Antenna Model and NEC4WIN, will go far to make more complex MININEC models possible. However, not until the other MININEC limits are overcome—namely, raising the speed of matrix execution and grafting on the Sommerfeld-Norton ground calculation system—will the program be fully competent for all purposes.

Solenoid Coil Limitations

In the physical models of single-layer solenoid inductances used as loads, the constraints imposed on the exercise dictated the use of helices with a relatively large diameter (11.85") and a long length (13.6"). The wire size (0.1285") was relatively thin compared to the turns spacing (1.36"), which yielded a coil Q somewhat less than the theoretical maximum for the overall configuration.

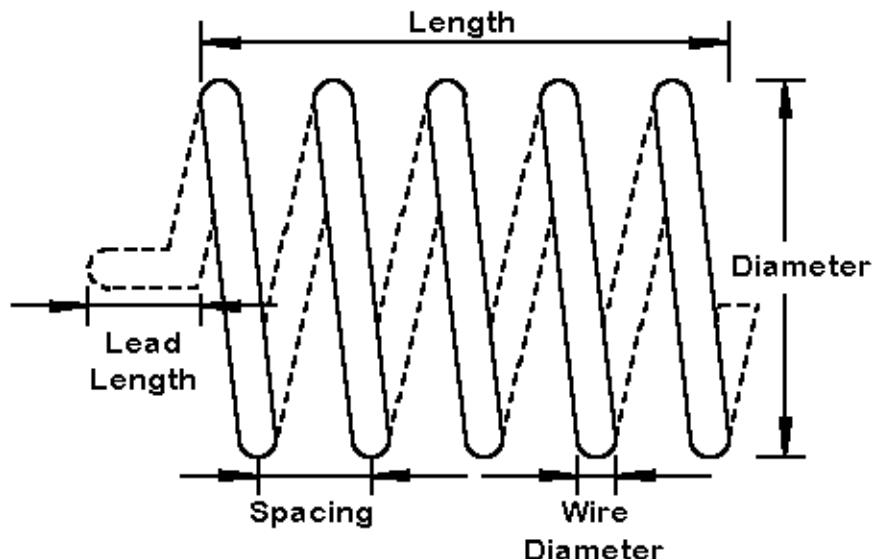


Fig. 17-4 Inductor Parameters

Fig. 17-4 shows some of the key parameters of coil construction that play a role in physical determination of coil inductance. Length, diameter and turns spacing are parts of the classic 1928 Wheeler approximation for single-layer solenoids (as found in ARRL *Handbook* since almost time immemorial):

$$L_{(\mu H)} = \frac{d^2 n^2}{18d + 40l}$$

where L is the inductance in microHenries, d is the diameter, l is the length, and n is the number of turns, and where d and l are in the same units. The number of turns can be determined from the length and turns spacing (or vice versa). For coils used at upper HF and higher, the lead length becomes a significant factor in the coil's inductance. In antenna modeling, the lead length is generally absorbed by the linear element length.

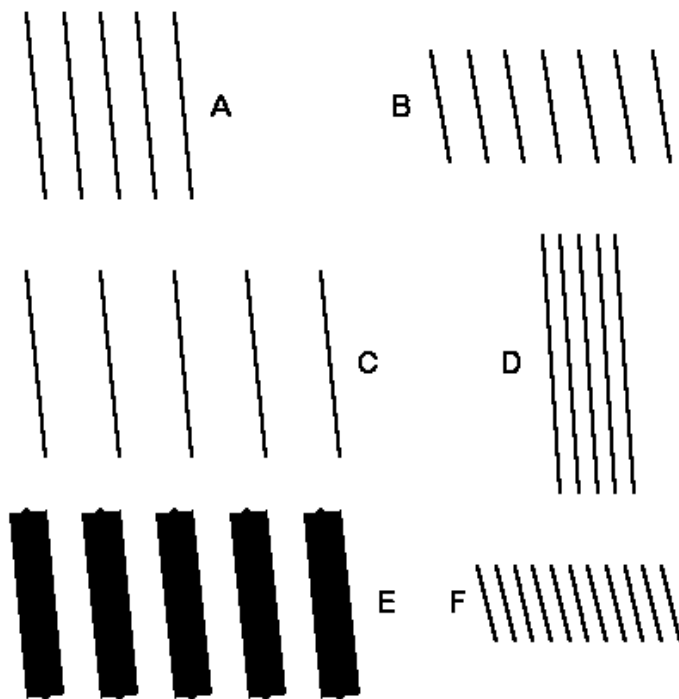


Fig. 17-5 Some Solenoid Configurations

In **Fig. 17-5**, we find some variations on the coils used in the physical models. If we let A represent the coil configuration used, we might obtain the same inductance (and reactance) using a coil like that in B, which uses the same spacing, but a smaller diameter and more turns. C spreads the turns over a longer length, while D uses a shorter length, closer spacing, and a larger diameter, perhaps to effect the highest Q configuration that might be practical. E employs fatter wire to achieve a high Q by optimizing the wire-diameter-to-turn-spacing ratio. Each of the preceding variants on the original coil was based on theoretical factors. However, in practical antenna design, one might also accept a somewhat lower Q in order to achieve a design that slipped wind by virtue of its small diameter, as suggested in F.

Each coil configuration is likely to affect antenna performance slightly, due to different radiation field additions that depend on diameter and turn spacing. Some of the variants are more easily modeled than others. With proper attention to segment lengths, coils that use the same size wire as the main or linear element can be modeled with fair ease. However, length-tapering of the linear element as it approaches the helix may be necessary for coils with small diameters. Where coils use wire diameters that differ from that of the main element, NEC (in any version) will exhibit a tendency toward erroneous results.

The consequence is that modeling solenoids as physical loading elements in an antenna may not in all cases be practical, although the task itself may become design-specific, especially in view of all of the variations possible. Thus, it may be some time before there is a more general profile of the affects of coil design on the performance of shortened antenna elements that use them. However, whatever differential there is between a physical coil and a mathematical load will apply equally to both NEC and MININEC. Both use essentially the same mathematical loading schemes, allowing for the fact that NEC places the load on a segment and MININEC places a load on a pulse or junction of segments.

It is perhaps a curiosity (since multiple types of coils have not been checked) that the loading coil used as the center load in part 4 of this series—and also used as the mid-element load—can be trebled to yield a 30-turn helical dipole only 40.8" long by less than 12" diameter. The self-resonant frequency of this experimental design was 7.105 MHz, only slightly lower than the standard design frequency used throughout the demonstration. If the exercise has any utility at all, it lies in the con-

firming the soundness of the advice to model solenoids using the material that will actually form the coil. (Model 17-1)

Material	Free Space Gain dBi	Feedpoint Impedance R +/- jX Ohms
Lossless wire	1.76	0.3 + j 0.5
Copper	-5.46	1.3 + j 1.6
Aluminum	-6.99	1.9 + j 2.2

The impracticality of the design, as indicated by the source impedance, is less important than a comparison of the anticipated gain of the antenna from the lossless wire to normal materials. Lossless wire makes the antenna look like a promising performer, while the gain figures for real materials tell the true story. There is no magic in this system. Indeed, it packs over 93' of wire into the helix, more than required for a standard 40-meter dipole. Yet, at 7-9 dB below the performance of a full size dipole, signals would be down by only about 1.5 S-units. Such performance would be usable under certain kinds of operating conditions, although the losses associated with feeding the antenna might attenuate signals even more than the basic design itself.

Obviously, those interested in helical antennas will always benefit from making them as long and open as possible. An adequate model of a helical monopole or dipole will always take the form of a physical model.

One final note on solenoids seems necessary, based on **Fig. 17-6**. A common practice, especially in commercial antenna manufacture, is to place a solenoid within a shield, and the shield may be connected (or not) to either of the linear wires extending from it. I am unaware of any technique of modeling this situation, which has been correlated to experimental results. One can surround the solenoid with a series of wires or even a wire grid—with one end connected to an antenna wire if desired. To what degree antenna fields arising from the solenoid will interact with the shield, and what the consequences for the overall antenna field will be, remains an area yet to be fully explored. It is likely to remain a project-specific task for designers. If we add the variables of coil construction that might be present within the shielding structure, the task is no small one. Note that the term “shield” is here used by convention. To whatever degree the casing plays a role in radiation-field

production, its shielding effects may be only mechanical. Its electrical role may turn out to be either simple or complex.

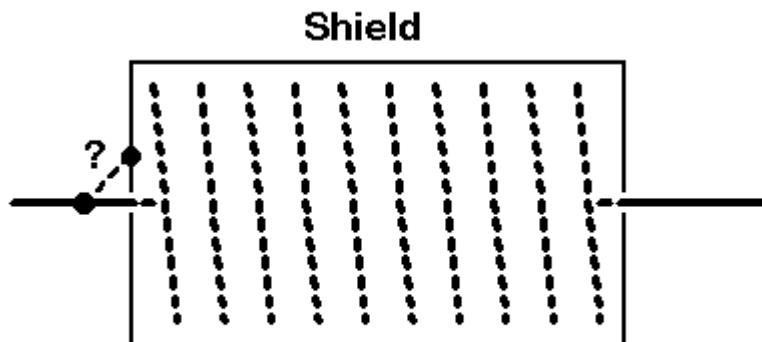


Fig. 17-6

Shielded Coils

Conclusion

These notes are intended to provide a reasonably fair view of the limitations of the modeling studies undertaken in the first 4 parts of this series. They provide some cautions against taking the results of the work as numerical general principles, which they are not. What remains valid are the trends shown by the numbers cataloged in tables and graphs. The notes also caution against hasty modeling that might unwittingly cross the boundaries of NEC capabilities. There are many ways to cross the boundary between accurate and inaccurate modeling results, and only experience (or a detailed checklist) will safeguard against wandering into illicit territory. Finally, the notes also outline a host of work that might be done to extend this study and to make its conclusions both more general and more precise.

Some of the further work can be done with some modeling ingenuity. Other parts of the work might be undertaken in alternative programs, such as MININEC. However, there may be a residue of the effort that may have to await the next gen-

eration of modeling cores before fully adequate modeling of all reactive loads, whatever their type and configuration, is possible.

* * * * *

Model included: 17-1. .EZ model dimensions in inches; .NEC and .NWP model dimensions in meters.



18. Why Tri-Banders Are Hard to Model

One of the most popular amateur radio antennas is the tri-band Yagi. It also happens to be one of the hardest antennas to model adequately, whether one is using NEC or MININEC. As a result, there are few, if any, reliable models available to the amateur modeler. The models that do exist tend to be done in one of the implementations of MININEC, unless they use a log-cell driver system. However, most of these models are proprietary. Moreover, many of them are used to guide final antenna design, but they do not represent exact models of the final design.

It may be useful to understand some of the limitations of trying to model tri-band Yagis. The account—which is very general and not tied to the specific circumstances surrounding any particular commercial antenna—may give some insight into why tri-bander models are hard to come by. In addition, it may give the owner of one of these antennas some cautionary notes lest he or she give undue trust to a hasty home-generated model of the station beam. Modeling tri-band Yagis requires the greatest of care.

Fig. 18-1 presents a sketch of a hypothetical tri-band Yagi. Its sole purpose is to register some of the features we shall look at in more detail. First, notice that the elements come in different lengths, each consisting of several different diameters of aluminum tubing. Second, notice that this tri-bander has traps, which are not universal in such antennas. Wherever they do occur, they present modeling problems of several sorts. Third, notice that some of the elements having different lengths are closely spaced. Finally, notice that this particular beam uses a driver system called the log-cell driver.

Each of these features presents either NEC or MININEC with some problems. Not all of the problems are insurmountable, but some may require the use of substitute models, while others simply demand that there be a slight difference between the modeled structure and the physical structure. Still others may leave a few unanswered questions about the precision of the model relative to the real antenna.

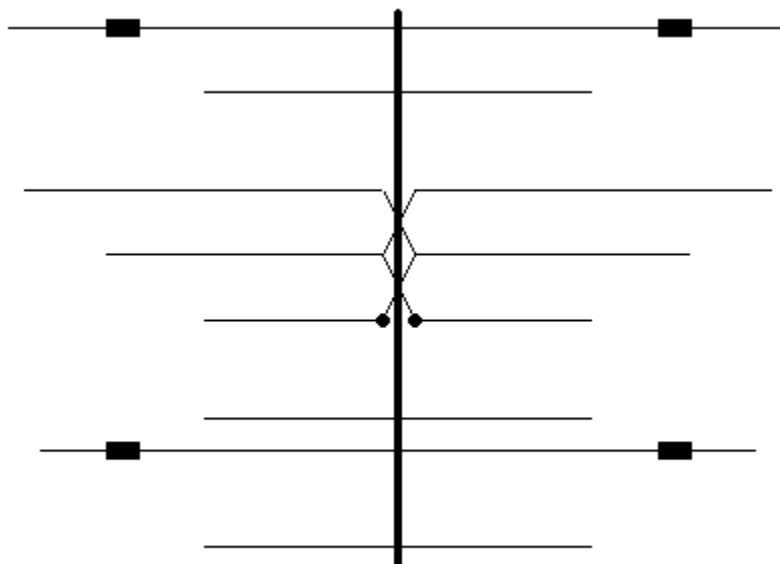


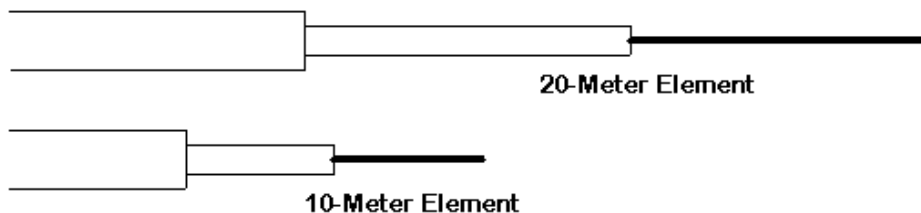
Fig. 18-1

A Composite Tri-band Yagi

1. Tapered-Diameter Elements

Most tri-band Yagis use tapered diameter elements composed of successive sections of aluminum tubing having a decreasing diameter from center to end. These element types present no special problem to MININEC. With NEC-2, the results of directly modeling such an element will be quite erroneous. NEC-4 makes an improvement upon NEC-2 for directly modeling tapered diameter elements, but tends to fall short of some requirements for precision. It grows increasingly inaccurate when the diameter step grows larger. This last condition is often encountered when modelers simulate boom-to-element mounting plates with large diameter, short-length segments.

For a mono-band beam, these limitations can be overcome by careful segmentation of the wires making up an element and through the use of correction factors built into some of the commercial implementations of NEC-2. Variant correction schemes have been developed by Brian Beezley and by David Leeson. These schemes substitute a uniform diameter element of the correct length to give the same electrical performance as the tapered diameter element, thus eliminating the situation that creates errors in NEC-2 outputs.



Tapered Diameter Elements for Each Band

Fig. 18-2

In a tri-band beam, we introduce a new problem: tapered diameter elements of different lengths, as shown in **Fig. 18-2**. MININEC would take this change in stride, since it requires no correction factor for tapered diameter elements. However, NEC-2—even with corrected elements—has a problem. The equations for a corrected element are only accurate within a reasonable margin from 1/2 wavelength resonance—about $\pm 15\%$. Commercial implementations of NEC-2 ordinarily provide corrected elements only if the element passes this length vs. frequency test. Elements cut for frequencies other than the band in question will ordinarily not undergo correction for their tapered diameter.

How much error may be introduced into the model by virtue of having uncorrected tapered diameter elements may vary from model to model. My own modeling suggests that 10-meter elements play little role in the 20-meter performance of a tri-bander and have low current levels. Varying their length has in most cases (but not in all cases) little or no effect on any significant 20-meter parameter. Hence, there is little error from having uncorrected 10-meter elements in the model.

The opposite case—the influence of 20-meter elements on 10-meter operation—is quite different. Although the level of influence varies from element to element in various designs, 20-meter directors often carry considerable current on 10-meters and may play a large role in 10-meter performance. In some designs, changes in the length of a 20-meter director can alter the gain vs. frequency curve for 10-meters. Thus, an uncorrected 20-meter element may become an error source when a tri-bander is modeled on 10 meters.

2. Traps

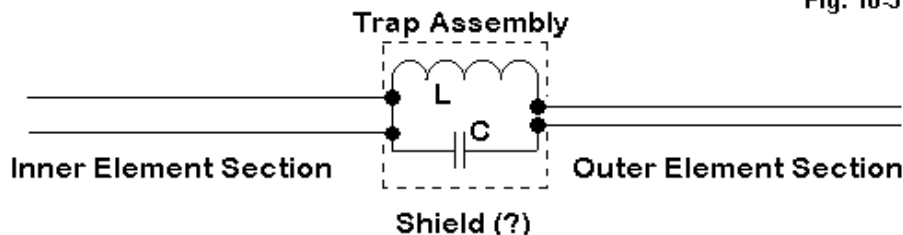
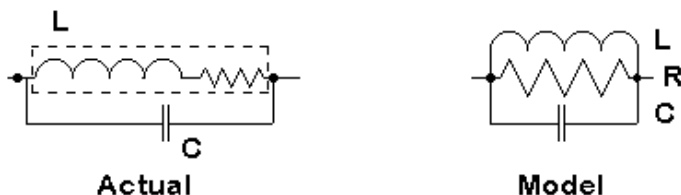


Fig. 18-3



Trap Considerations

Some tri-band Yagi designs make use of trapped elements. Although many tri-bander designs have eliminated traps from the driven element, they remain in some reflectors and directors. **Fig. 18-3** introduces us to some aspects of the difficulties of modeling trapped elements in Yagi designs

The upper sketch shows a typical trap in schematic form as inserted into an element. Normally, the trap is configured in the model as a parallel R-L-C “load,” which both NEC and MININEC allow.

In NEC-2, the only permissible loads for symmetrical elements are those positioned at the center of the element. The implementation of the tapered diameter correction requires that the curve of current magnitude along the element be continuous, and a load that is not at the center of the element creates a current magnitude discontinuity. Hence, most implementations test for loads outside the center-most segment, and the existence of such a load will defeat use of the correction.

Although the placement of traps along an element is not a concern with MININEC, it is a potential source of error with NEC-2 models. The element remains uncorrected at any test frequency.

The trap itself may call for careful treatment, since the parallel R-L-C mathematical model of the load, shown on the right in the lower sketch in **Fig. 18-3**, is a substitute for the circuit shown at the left. For most purposes, the resistance that limits the Q of the trap circuit can be viewed as a series resistance with the inductance (or inductive reactance) of the coil. Setting a value for the resistance in the modeled trap load requires a conversion of the series inductive arm into its parallel equivalent.

Such conversions are straightforward calculations when a parallel resonant circuit is operated at its resonant frequency. However, traps are rarely operated in a truly resonant mode. A trap tends to be resonant at a frequency just below the band edge for which it is a trap, giving a net capacitive reactance to the parallel circuit. At lower frequencies, the reactance is inductive, but not simply the inductance of the coil. There are standard handbook equations for calculating the reactance of any parallel tuned circuit when operated off resonance.

The net effect of the change of reactance off the resonant frequency of the trap is to alter the required value of the parallel resistance in the modeled trap with every change in frequency of test. Utility programs exist to ease the calculation of these values (for example, in the HAMCALC suite of utility programs). This requirement is in no way fatal to any tri-bander model (where traps are used in the design), but it does require care on the modeler's part lest the trapped element show greater or lesser losses than exist on the physical element itself. (EZNEC Version 3.0 and upward has a special "trap" load entry provision that performs the required transformation into a parallel R-L-C load. The entry calls for a trap frequency. Thus, for each change of frequency, the entry not only calculates the change of reactance, but as well, the change of resistance.)

Series and parallel modeled loads using the LD facilities of either NEC or MININEC are mathematical, non-radiating models. As such, they do not perfectly model the physical elements they represent. Inductors do not show perfectly equal currents along their length when physically modeled, suggesting that they perform radiating as well as loading functions. The radiating function becomes more pronounced as a coil is moved further from the element's center. When coils are small enough in length and in diameter, the radiating function may approach negligible proportions, but is never zero.

In **Fig. 18-3**, there is a "Shield (?)" notation. A trap (or a loading coil) may be placed inside a metal shell, and the shell may be connected to the inner portion of the element, to the outer portion of the element, or to nothing. To this time, I have not seen any reports on the effects of shielding—however connected—on the overall element performance. The presence or absence of a trap shield may or may not constitute a further potential error source for a trapped element.

3. Close-Spaced Elements

Fig. 18-4 shows a situation that is becoming more common in tri-bander designs: two elements of different lengths in close proximity. These elements may occur in designs which use open-sleeve coupling, where the "slave" 15- and 10-meter drivers are closely spaced to the fed or "master" 20-meter element. The exact spacing and element lengths determine the impedance seen at the feedpoint for each of the bands of operation.

Fig. 18.4

Close Spaced Elements

A second situation in which we can find similarly close-spaced elements of different lengths is in the director set. The 20-meter directors may function as part of the 10-meter director set, showing a single current magnitude curve which peaks at the center of the element. However, the 20-meter directors may not be the correct length to provide a 10-meter gain curve where the designer desires it. Placing a 10-meter director of the correct length on the driver side of the 20-meter element can permit the designer to better control the 10-meter performance curves. This new element may be as close as 4-8 inches from the old 20-meter element.

Although MININEC has no trouble with this situation, NEC-2 does. Two elements of significantly different lengths that are closely spaced can introduce errors into the NEC-2 output reports. The effect is subject to a considerable number of variables, including frequency, diameter of the elements, element lengths, and spacing. The closer the spacing, the greater the effect on the model.

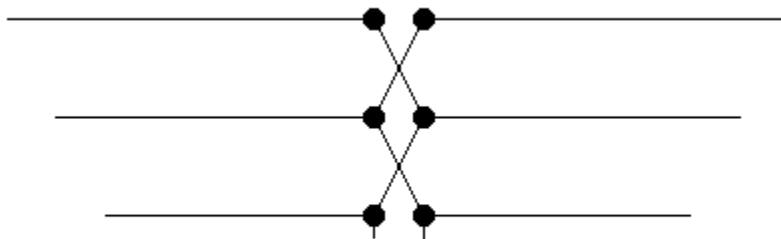
Whether or not a given spacing poses a possible problem can be determined by the Average Gain Test. Although the test results have been available in NEC, the test is just now appearing in new versions of commercial implementations of NEC. New versions of EZNEC (for Windows) and NEC-Win Plus both make the test available. One may test an entire complex antenna. However, where a specific question arises—as in the case of a pair of elements like the ones we have described—we may perform the test on just the element pair by making an isolated model of the elements and placing a source at the center of the longer element.

For the present case, the test would be performed in free space, with the resistive component of any loads disabled, including the resistivity of the antenna materials. For any antenna in these conditions, the average gain—or ratio of radiated power to power provided—will be 1, taking into account radiation in all possible directions. Ordinarily a fair sampling of directions, as indicated by taking gain readings every so many degrees over the sphere implicit in free space models, will provide a clear indication of how close the average gain of the model comes to 1. Values that deviate too far from 1 make the reliability of the model suspect. (Note: there is no absolute level at which a given model becomes untrustworthy. For some purposes, values above 0.8 and below 1.2 may indicate a usable model, while other modeling tasks may require values within about 0.05 of the ideal 1.0.)

The Average Gain Test is a necessary but not a sufficient condition of model adequacy. For example, the Average Gain Test may not indicate all of the possible error sources we have described for tri-band Yagis. See column 20 for further details of the test.

4. Log-Cell Drivers

A typical log-cell driver set is roughly shown in **Fig. 18-5**.



Log-Cell Drivers

Fig. 18-5

The log-cell consists of a driver set for the tri-band Yagi so that the set has a single point of connection with the antenna feedline. A true log-cell driver set has

crossing lines of connection among the elements. Another driver-set type may use uncrossed connections and has been loosely dubbed “closed-sleeve” coupling. In this latter case, both direct connections and open-sleeve coupling (which depends on element spacing and length for a given source impedance on each band) may be at work.

Modeling both log-cells and closed-sleeve drivers requires the construction of transmission lines from the source to each driver in the set. NEC-2 has fewer problems in creating such a model, since the transmission line is at the element center where currents are balanced. Since the TL facility of NEC-2 is purely mathematical, connection reversal presents no model construction difficulties over and above those of a closed-sleeve driver set. Indeed, the only difficulty for NEC lies in possible accuracy problems that might arise if the elements of different length are spaced too closely, as described earlier.

Modeling the element-connecting transmission line in NEC-2 as a physical set of wires may be more difficult. The connecting wires are normally of a smaller diameter than the elements. Among its limitations, NEC has difficulties with junctions of wires having dissimilar diameters. Consequently, for most purposes (including the modeling of full LPDA antennas), use of the TL facility is recommended. (However, the technique described below as a work-around for MININEC can also be applied to some NEC models.)

These same driver sets can also create problems for MININEC, which has no TL facility. Therefore, one can only physically model transmission lines as wires. The log-cell is particular difficult, since the wires must maintain a constant distance between them to effect a transmission line.

Apart from the line spacing, MININEC suffers another difficulty: at sharp angles, the core will “clip” corners and shorten elements unless the segment lengths at the angular junction are very short. Length-tapering routines are provided in some implementations of MININEC to provide short elements at the junction but segments of increasing length away from the junction. These routines are designed to use the least number of segments that can adequately model the element and still hold the total model within the 256-segment limit of the core in DOS versions. More recent Windows implementations of MININEC 3.13 have removed the segment limitation, but still run very slowly when models are very large.

The log-cell and the closed-sleeve driver sets have a considerable number of right-angle junctions, and the potential for overrunning model size or run-time patience limitations is very good. The development of a MININEC model that can handle such driver-set complexities requires considerable ingenuity.

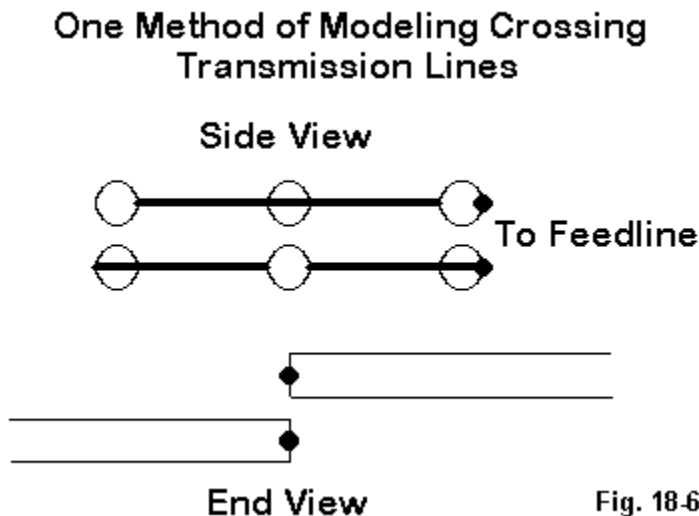


Fig. 18-6

When physically modeling crossing transmission lines, you can avoid extra line bends by using the set-up shown in **Fig. 18-6**. Set up two transmission lines as 4 wires (for a 3-driver set). The wires maintain a constant distance from each other, as determined by transmission line construction calculations (available in HAMCALC). Each element is offset from the centerline to intersect the correct wire junction of the transmission line. Alternate offsets as you move from one element to the next. The displacement of the element wires is less likely to create significant modeling errors than trying to maintain a constant distance between undulating wires of a transmission line that tries to bend its way to the connections.

This is only one sample of the variety of modeling techniques that may be required to overcome complexities of design when they run up against limitations of the modeling software.

Conclusion

These notes are titled “Why Tri-Banders Are Hard to Model,” not “Why Tri-Banders Are Impossible to Model.” Certainly there are models that are sufficient guides to construction and that permit commercial designers to make final prototype adjustments in field tests. Whether these models suffice to confirm performance claims is an issue beyond the scope of these notes. However, to the extent that models require the use of substitute elements or leave open questions about the adequacy of the model to the physical antenna, they may be accounted by some to be less than fully acceptable. Nonetheless, these models will have served a very useful purpose in guiding the development of an antenna. The exactitude we demand of a model is a function of a well-defined set of task goals. (On the other hand, a sloppily defined task does not justify acceptance of a sloppy model or its over-optimistic performance predictions.)

In terms of modeling one’s own tri-bander, these notes may serve as a reminder of the caution one must use in building the model. There are limiting boundaries for both MININEC and NEC that one may cross without noticing them. Use great care in trying to model a tri-band Yagi—the job is harder than you may initially think.

* * * * *

Models included: 18-1 through 18-2. .EZ model dimension in inches; .NEC and .NWP model dimensions in meters.

There are no text references to the included models, because they do not correspond to any particular portion of these notes. The two models are of the same tri-band Yagi, the first using tapered diameter elements (for use with Leeson corrections), the second consisting of uniform-diameter substitute elements. The substitute elements derive from the corrected elements, using the prescribed total length and diameter after sampling the antenna on each band. This technique compensates for possible erroneous reports that may arise because the correction factors for the tapered-diameter version do not apply to all elements at any given test frequency. Since all elements have uniform diameters from end-to-end, there is no

need for the correction system during test runs. The arrays do have a special feature not included in the antenna on which the model is based: a free-floating short tube that simulates the effects of the element-to-boom hardware on the closed-sleeve driver system. The source of the models will remain unnamed, since the model is a private construction and in no way licensed by the manufacturer. Its purpose is simply to provide you with a sample of a complex model of a tri-band Yagi and some of the techniques that one may use to overcome the challenges presented by the tri-bander.

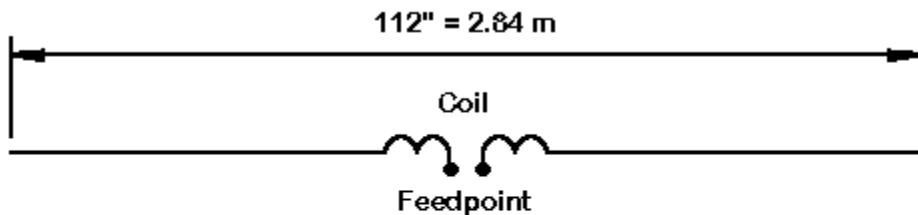


19. What Can We Learn From Tables?

Every commercial implementation of either NEC or MININEC tries to anticipate the needs of its users. Therefore, each tries to provide crucial data in the most usable form. Polar plots, SWR sweeps, etc. appear in graphical form, with supplementary numerical data. Some data is available in tabular form, often modified to make it more readable. Thus, a typical NEC output value of 2.1356E-03 will be rounded to 0.0021.

Behind every form of data made available to the user lies a basic NEC core output table—or, more correctly, a collection of tables. Once you have generally mastered the readily available information, it is time to look at the core output in order to gather new data, refine data already in hand, and even to get a new slant on existing data.

So let's look at a typical NEC core output file for a simple antenna: the 10-meter center-loaded dipole shown in **Fig. 19-1**.



Coil: 644 Ohms inductive reactance
3.60 microH
Q = 600

Element: #12 Copper wire

Fig. 19-1

Center-Loaded Dipole

The antenna consists of a single modeled wire with 41 segments (each about 2.75" long for the overall 112" antenna length). The antenna wire is copper, with its characteristic loss factor. The center-loading coil is specified as a resistance and reactance (1 Ohm and 644 Ohms), which is equivalent to a 3.6 microH coil with a Q of about 600. Since we are using only one frequency for this model, the impedance load and the R-L-C load are equivalent. We shall set the antenna in free space, with no transmission lines, networks, or other complexities. This much of a model will already introduce us to an episode's worth of information from the NEC core output.

Let's take up the core output a bit at a time. In many cases, the old "green sheet" output width of the data will be too wide for our screen, so I shall omit some columns of less interest to us here. (In other contexts, they may hold great interest.) Note that NEC operates only in meters, and converts everything to that measure. For simplicity, I have pre-converted the data from our 112" antenna. Also note that NEC uses a radius, although conventional input systems let you input wire diameters or AWG wire sizes.

```

- - - STRUCTURE SPECIFICATION - - -

COORDINATES MUST BE INPUT IN
METERS OR BE SCALED TO METERS
BEFORE STRUCTURE INPUT IS ENDED

WIRE
NO.      X1      Y1      Z1      X2      Y2      Z2      RADIUS  SEG.
  1  0.00000  0.00000  0.00000  2.84480  0.00000  0.00000  0.00103  41
    STRUCTURE SCALED BY FACTOR  1.00000
TOTAL SEGMENTS USED= 41 NO. SEG. IN A SYMMETRIC CELL= 41 SYMMETRY FLAG= 0
- MULTIPLE WIRE JUNCTIONS -
JUNCTION  SEGMENTS  (- FOR END 1, + FOR END 2)
NONE

```

This portion of the file essentially replicates your wire entry for the antenna element(s). It is a good crosscheck against what you thought you entered. However, NEC goes further and sorts out every wire segment and its connections. Only the first 5 segments are shown below—with some columns omitted.

- - - - SEGMENTATION DATA - - - -

COORDINATES IN METERS

I+ AND I- INDICATE THE SEGMENTS BEFORE AND AFTER I

SEG. NO.	COORDINATES OF SEG. CENTER			SEG. LENGTH	WIRE RADIUS	CONNECTION DATA			TAG NO.
	X	Y	Z			I-	I	I+	
1	0.03469	0.00000	0.00000	0.06939	0.00103	0	1	2	1
2	0.10408	0.00000	0.00000	0.06939	0.00103	1	2	3	1
3	0.17346	0.00000	0.00000	0.06939	0.00103	2	3	4	1
4	0.24285	0.00000	0.00000	0.06939	0.00103	3	4	5	1
5	0.31223	0.00000	0.00000	0.06939	0.00103	4	5	6	1

Note that each segment is located by the coordinates of its center, which is the focal point of NEC calculations. (MININEC focuses upon pulses, which are located at segment junctions or ends.) The connection data can be important in checking to ensure that complex wire structures that you devise are connected where you want them and not connected where you wish the wires to be separated.

Many implementations of NEC have error checking routines. However, they cannot read the modeler's intentions. If you transpose numbers in a coordinate (for example, 42 instead of 24), the error checker will not catch the error unless it results in a standard error, such as wires that connect within a segment. Reviewing the coordinates of the segments and their connections is a good supplemental error check routine.

```

INPUT LINE 1 EX 0 1 21 0 1.00000E+00 0.00000E+00
              0.00000E+00 0.00000E+00 0.00000E+00 0.00000E+00
INPUT LINE 2 LD 4 1 21 21 1.00000E+00 6.44000E+02
              0.00000E+00 0.00000E+00 0.00000E+00 0.00000E+00
INPUT LINE 3 LD 5 1 1 41 5.80010E+07 0.00000E+00
              0.00000E+00 0.00000E+00 0.00000E+00 0.00000E+00
INPUT LINE 4 FR 0 1 0 0 2.85000E+01 0.00000E+00
              0.00000E+00 0.00000E+00 0.00000E+00 0.00000E+00
INPUT LINE 5 RP 0 1 361 1000 9.00000E+01 0.00000E+00
              1.00000E+00 1.00000E+00 0.00000E+00 0.00000E+00

```

Each of the "Input Line" Entries is actually a single report line (split here to ensure that it fits the page). These lines register the control cards. In this model, we have the following entries:

EX or source information (wire and segment of placement and the magnitude and phase of the source voltage);

LD 4 or the center-loading-coil load (Type 4 impedance load showing wire and segment of placement and the resistance and reactance);

LD 5 or the wire conductivity (showing the wire of placement and the segment range, as well as the conductivity of copper);

FR or the frequency request (a single frequency: 28.5 MHz); and

RP or the pattern request (an azimuth pattern of 360 degrees at zero degrees elevation, which NEC expresses as a 90-degree theta angle).

These entries are once more useful as cross checks on what you intended to put into your model. To read values, you must grow accustomed to engineering notation (for example, 5.80010E+07) and to the positions of information along each line.

```

- - - - - FREQUENCY - - - - -
      FREQUENCY= 2.8500E+01 MHZ
      WAVELENGTH= 1.0519E+01 METERS

```

NEC reads the frequency in MHz and provides its equivalent in wavelengths. If you multiply the two figures, you will find that NEC uses 2.997915E+06 m/s as the speed of electromagnetic radiation.

Since our antenna is in free space, we can skip over the “Antenna Environment” entry to the loading lines.

```

- - - STRUCTURE IMPEDANCE LOADING - - -

```

LOCATION			IMPEDANCE (OHMS)		CONDUCTIVITY	TYPE
ITAG	FROM	THRU	REAL	IMAGINARY	MHOS/METER	
1	21	21	1.0000E+00	6.4400E+02		FIXED IMPEDANCE
1	1	41			5.8001E+07	WIRE

NOTE: SOME OF THE ABOVE SEGMENTS HAVE BEEN LOADED TWICE - IMPEDANCES ADDED

These lines have unused columns omitted (those used for R-L-C loads). NEC uses the mathematical terminology of “real” and “imaginary” in place of the electrical terms “resistive” and “reactive” to characterize not only impedance, but as well every other quantity which may also have a phase angle. Note also that segment 21 (the element center segment) has both the impedance load and the wire conductivity load. Since conductivity is a measure per unit length, one cannot simply arithmetically add resistances to obtain the value used by NEC on that segment. The “Tag” and ITag” notations refer to the wire number in this particular model. (The use of the “tag” number as identical to the wire number is only one way to use tag numbers.)

- - - ANTENNA INPUT PARAMETERS - - -

TAG NO.	SEG. NO.	VOLTAGE (VOLTS)		CURRENT (AMPS)		POWER (WATTS)
		REAL	IMAG.	REAL	IMAG.	
1	21	1.00000E+00	0.00000E+00	5.80233E-02	1.91669E-03	
		IMPEDANCE (OHMS)		ADMITTANCE (MHOS)		
		REAL	IMAG.	REAL	IMAG.	
		1.72157E+01	-5.68686E-01	5.80233E-02	1.91669E-03	2.90117E-02

The “Antenna Input Parameters” report is actually a single line entry. Because of its length, I have split it to ensure a page fit.

NEC calculates the mutual impedances of all segments in a total antenna structure. With these figures, plus the source values (1 volt at 0-degrees phase angle in our example), the source or feedpoint collection of values can be calculated from Ohms Law and its derivative power equation. Admittance, of course, is simply the reciprocal of impedance. Had we set the voltage at a value other than 1.0, the other values (except the impedance and the admittance) would have changed accordingly.

Some implementations permit the use of current sources by a network from a distant true voltage source to the user-designated source segment. Other implementations scale these figures so as to allow the user to specify a constant power value.

Among the most important information provided by the NEC core output file is the current data. What follows is the table for 1/2 of the antenna we are using as our

example. The other half of this symmetrical single element would show the same values in reverse order.

- - - CURRENTS AND LOCATION - - -									
LENGTHS NORMALIZED BY WAVELENGTH (OR 2.*PI/CABS(K))									
SEG. TAG	SEG. CENTER	SEG.			- - - CURRENT (AMPS) - - -				
NO. NO.	X	Y	Z	LENGTH	REAL	IMAG.	MAG.	PHASE	
1	1	0.0033	0.0	0.0	0.00660	2.1353E-03	3.4527E-05	2.1356E-03	0.926
2	1	0.0099	0.0	0.0	0.00660	5.6942E-03	9.6119E-05	5.6950E-03	0.967
3	1	0.0165	0.0	0.0	0.00660	8.9690E-03	1.5790E-04	8.9704E-03	1.009
4	1	0.0231	0.0	0.0	0.00660	1.2127E-02	2.2233E-04	1.2129E-02	1.050
5	1	0.0297	0.0	0.0	0.00660	1.5203E-02	2.8992E-04	1.5205E-02	1.093
6	1	0.0363	0.0	0.0	0.00660	1.8213E-02	3.6086E-04	1.8216E-02	1.135
7	1	0.0429	0.0	0.0	0.00660	2.1165E-02	4.3528E-04	2.1169E-02	1.178
8	1	0.0495	0.0	0.0	0.00660	2.4062E-02	5.1323E-04	2.4068E-02	1.222
9	1	0.0561	0.0	0.0	0.00660	2.6908E-02	5.9476E-04	2.6915E-02	1.266
10	1	0.0627	0.0	0.0	0.00660	2.9703E-02	6.7991E-04	2.9711E-02	1.311
11	1	0.0693	0.0	0.0	0.00660	3.2448E-02	7.6871E-04	3.2457E-02	1.357
12	1	0.0759	0.0	0.0	0.00660	3.5144E-02	8.6124E-04	3.5154E-02	1.404
13	1	0.0825	0.0	0.0	0.00660	3.7791E-02	9.5760E-04	3.7803E-02	1.452
14	1	0.0890	0.0	0.0	0.00660	4.0392E-02	1.0579E-03	4.0406E-02	1.500
15	1	0.0956	0.0	0.0	0.00660	4.2951E-02	1.1625E-03	4.2966E-02	1.550
16	1	0.1022	0.0	0.0	0.00660	4.5472E-02	1.2717E-03	4.5490E-02	1.602
17	1	0.1088	0.0	0.0	0.00660	4.7967E-02	1.3861E-03	4.7987E-02	1.655
18	1	0.1154	0.0	0.0	0.00660	5.0456E-02	1.5070E-03	5.0478E-02	1.711
19	1	0.1220	0.0	0.0	0.00660	5.2978E-02	1.6367E-03	5.3003E-02	1.769
20	1	0.1286	0.0	0.0	0.00660	5.5668E-02	1.7830E-03	5.5697E-02	1.834
21	1	0.1352	0.0	0.0	0.00660	5.8023E-02	1.9167E-03	5.8055E-02	1.892

I have truncated the 0.00000 entries for the Y and Z coordinates to make the table fit the space available. Note that the coordinate positions for the current placements are the segment centers. For this single element antenna, you can simply compare the current placement coordinates with the coordinates given in the segmentation data section of the report.

Current values are based on the source voltage and the impedance, accounting for losses within a given segment. NEC reports these currents in two ways: as real and imaginary components and as a current magnitude and phase angle. For our center-loaded dipole, the curve of the current magnitude and phase are shown in **Fig. 19-2**.

The current magnitude is shown on the left Y-axis, while the current phase value is shown on the right Y-axis. Note that neither curve is linear. Many users find that the current values are more meaningful when normalized to a source current of 1.0 with a phase angle of zero degrees.

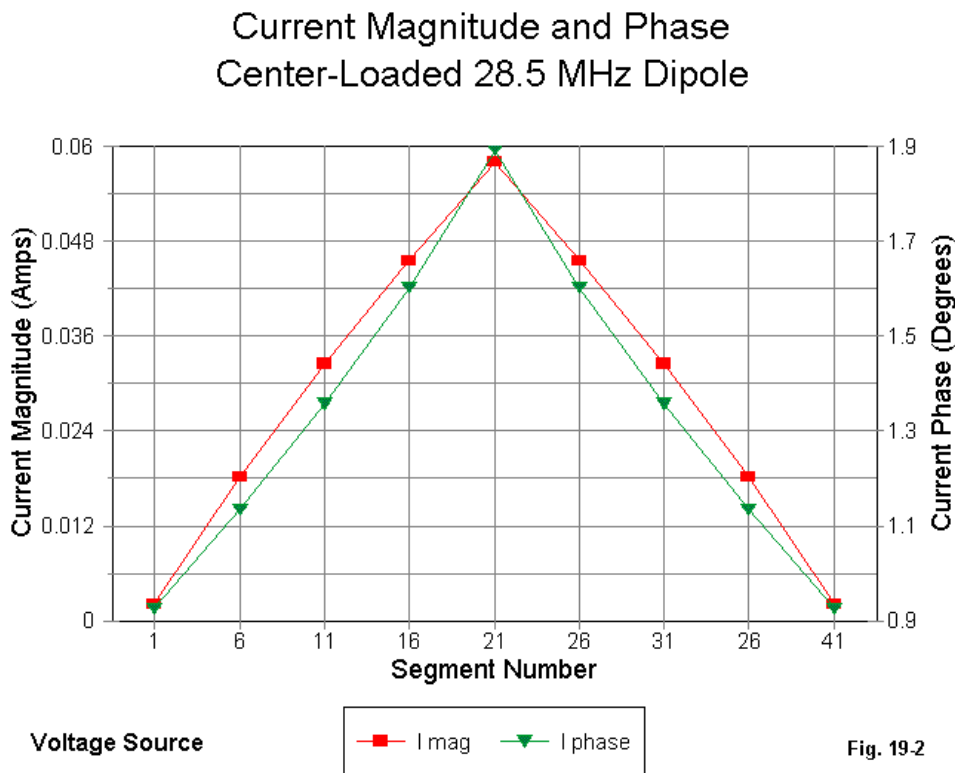


Fig. 19-2

Fig. 19-3 re-plots the data using a current source as just specified. Now it is clear how the current levels at various points along the antenna wire compare with

the source current. As well, with the phase angle referenced to zero, the amount of phase shift along the element is also clearer. (The nearly 1-degree phase shift is far more interesting than it is significant to antenna performance.)

Current Magnitude and Phase Center-Loaded 28.5 MHz Dipole

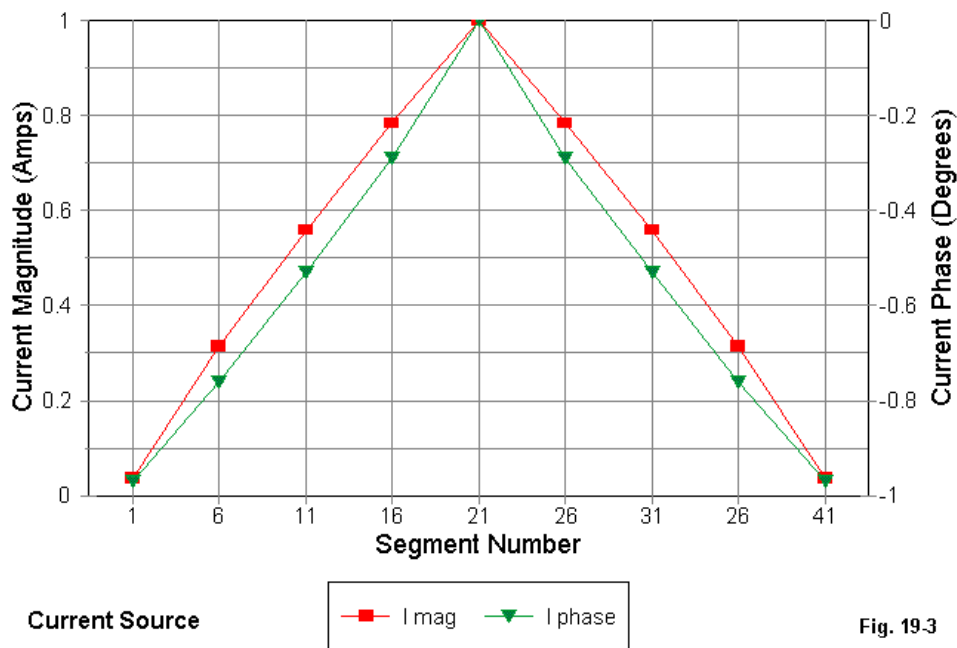


Fig. 19-3

Different implementations of NEC provide differing ways of plotting the current data. Curves may be overlaid on outline drawings of the antenna elements—or they

may be shown in color transitions rather than as curves. Some programs contain rectangular plot capabilities for plotting current. Virtually all programs make the current location and level data available as an immediate program output, thus saving the user the time of searching it out in the large NEC output file.

Detailed current data can be instructive about antenna performance. Our single loaded element is far too simple to hold any surprises. However, I was surprised when examining a model of a tri-band beam to discover, first, that a 20-meter director had a very significant current magnitude on it (compared to the other directors) when the operating frequency was within the 10-meter band. The second surprise came from examining the current magnitude along the element. Although about 1 wavelength long at the 10-meter operating frequency, the 20-meter element showed a single current maximum at its center with ever-decreasing levels outward from the center. The 20-meter element was very actively involved in the formation of the pattern from the Yagi on 10 meters. In fact, its length—set for 20-meter performance—provided a limit on the operating bandwidth on 10 meters, pushing the performance curves closer to one end of the band than the other. For this particular model, changing the length of the 20-meter director could enhance performance across the first MHz of 10 meters, but at the expense of performance on 20 meters.

```
- - - POWER BUDGET - - -  
  
INPUT POWER      = 2.9012E-02 WATTS  
RADIATED POWER= 2.6950E-02 WATTS  
WIRE LOSS        = 2.0613E-03 WATTS  
EFFICIENCY       = 92.90 PERCENT
```

The “Power Budget” report in the NEC output file can be very instructive when considering an antenna by itself and when comparing more than one competing design. For our center-loaded dipole, the efficiency (the ratio of radiated power to input power X 100) is considerably less than that of an unloaded full-length dipole. The wire loss is the sum of all resistive losses in the antenna, the major portion of which occurs in the loading coil.

Some programs make the power budget immediately available. Others provide alternative useful data, such as losses in loads of Type 0, 1, and 4. In many instances, it is important not just to know the overall losses in an antenna, but as well to identify the sources of these losses.

- - - RADIATION PATTERNS - - -

- - - ANGLES - - -			POWER GAINS -			- - E(THETA) - -		- - E(PHI) - -	
THETA	PHI	VERT.	HOR.	TOTAL	MAGNITUDE	PHASE	MAGNITUDE	PHASE	
DEGS	DEGS	DB	DB	DB	VOLTS	DEGS	VOLTS	DEGS	
90.00	0.00	-157.31	-999.99	-157.31	1.79796E-08	-39.79	0.00000E+00	0.00	
90.00	1.00	-157.31	-34.18	-34.18	1.79772E-08	-39.80	2.57884E-02	140.20	
90.00	2.00	-157.31	-28.16	-28.16	1.79700E-08	-39.82	5.15721E-02	140.18	
90.00	3.00	-157.32	-24.64	-24.64	1.79581E-08	-39.86	7.73462E-02	140.14	
90.00	4.00	-157.33	-22.14	-22.14	1.79415E-08	-39.91	1.03106E-01	140.09	
90.00	5.00	-157.34	-20.20	-20.20	1.79200E-08	-39.98	1.28847E-01	140.02	
90.00	6.00	-157.35	-18.62	-18.62	1.78938E-08	-40.06	1.54563E-01	139.94	
90.00	7.00	-157.37	-17.29	-17.29	1.78629E-08	-40.16	1.80251E-01	139.84	
90.00	8.00	-157.38	-16.13	-16.13	1.78271E-08	-40.27	2.05905E-01	139.73	
90.00	9.00	-157.40	-15.11	-15.11	1.77866E-08	-40.39	2.31520E-01	139.61	
90.00	10.00	-157.42	-14.20	-14.20	1.77413E-08	-40.53	2.57091E-01	139.47	

The only radiation pattern requested for our center-loaded dipole is an azimuth pattern at an elevation angle of zero degrees. Only the first 10 degrees of the report are shown here. When the entire data series is plotted, the pattern has the appearance of **Fig. 19-4**.

Because graphical plotting involves the use of lines having some thickness on a finite overall graph size, some angles are difficult to resolve visually. Consequently, the radiation table report—often made separately available by implementations of NEC—is a useful source of data. The portion of the 360-degree report shown here includes the first 10 degrees, beginning with a direction that is at the side null of the dipole pattern. The depth of a free-space dipole null is indicated by the total field power gain. By comparison with the value at the next degree of pattern resolution, one can see to what extent this deep null is a point value and what sorts of values are more realistic to expect of an antenna that does not quite stand still in a small breeze.

Very often, you can make important data more graphically accessible via rectangular graphs. Some software packages have internal rectangular graphing capabilities. However, in all cases, you can transport the data to a spreadsheet, where even more powerful graphing capabilities are normally available. In the present case, we can easily graph the radiation pattern on a rectangular chart to show the deep nulls, as in the top half of **Fig. 19-5**.

Free Space Azimuth Plot: 28.5 MHz Center-Loaded Dipole
Maximum Gain: 1.55 dBi; Front-to-Side Ratio: 158.86 dB

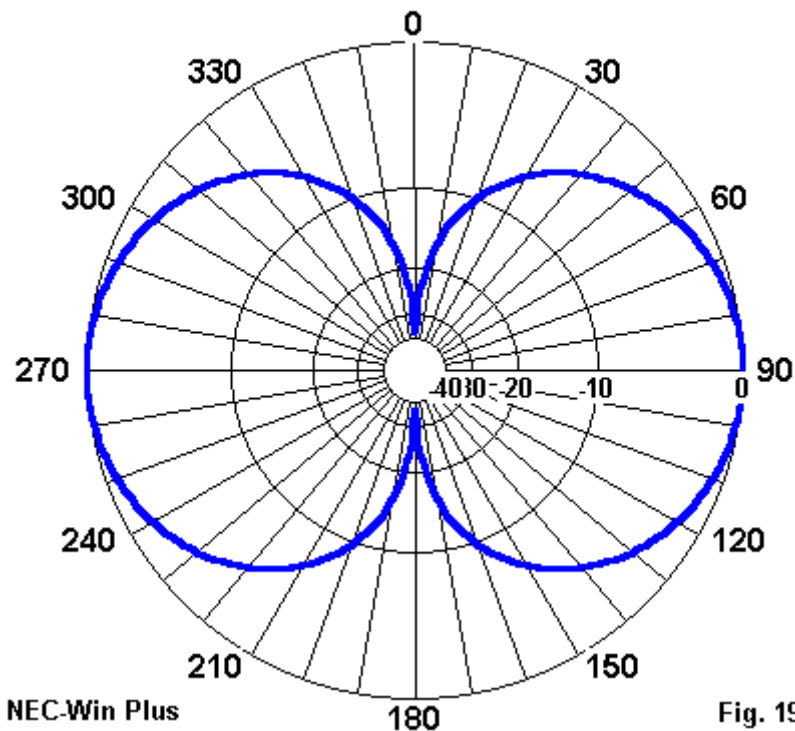
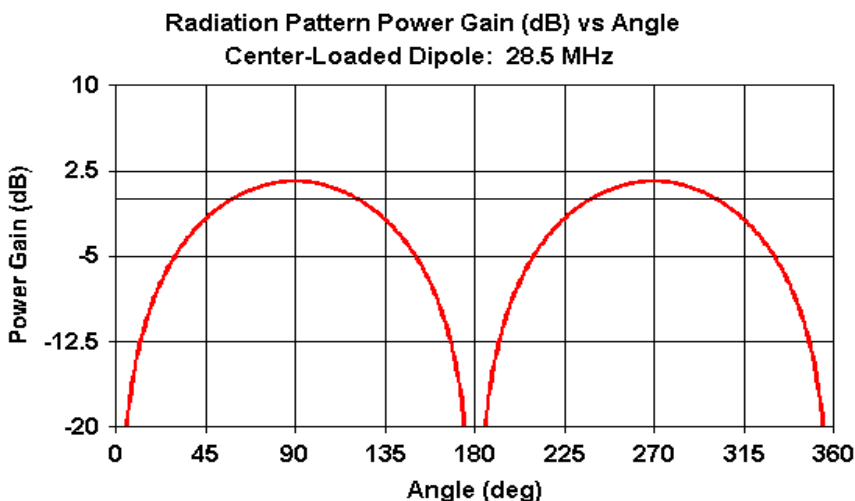
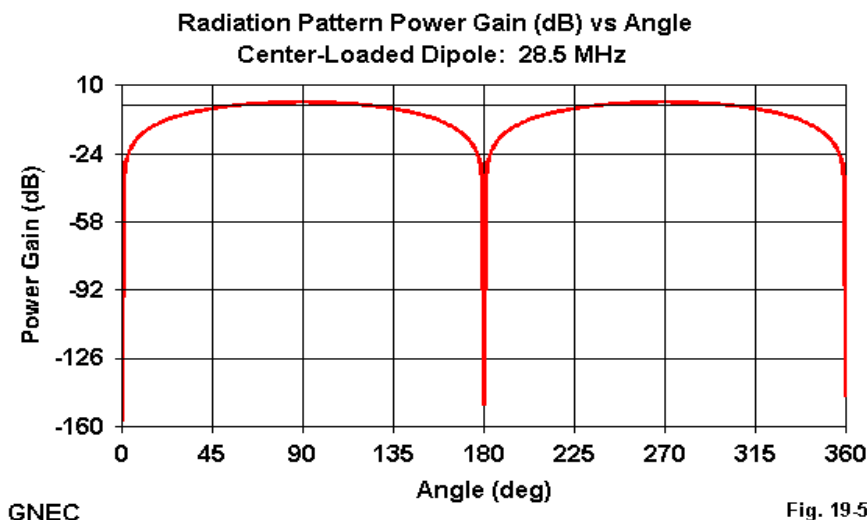


Fig. 19.4



What we gain in the ability to see the fine detail of the side nulls, we lose in terms of possibly misinterpreting the higher gain portions of the antenna pattern. The flat tops of the gain peaks in the upper graph do not permit resolution of important gain changes as we change the azimuth heading. However, by changing the Y-axis scale, we can enhance the presentation of these changes, as shown in the lower half of **Fig. 19-5**.

In the end, translating tabular data into graphical form requires great care in order to present data accurately and without leaving misleading impressions. In all cases, be prepared to refer to the tabulated data in the NEC output report to resolve questions left by any graphic presentation.

To make effective use of the radiation pattern report, it is important to understand the conventions used by NEC. The conventional elevation angle counts from the horizon upward, while the theta angle used by NEC counts from the zenith downward. Hence, a zero-degree elevation angle is a 90-degree theta angle. In addition, phi angles count in a counterclockwise direction, while azimuth angles use compass headings clockwise around the circle. (For most symmetrical antennas, the azimuth-phi distinction makes little or no difference in the pattern viewed, and I have seen software that simply uses the phi-angle as is while calling it an azimuth angle.)

The report provides power gains in dBi (isotropic) for both vertically and horizontally polarized radiation, as well as the power gain for the total field. Ordinarily, all three fields are available graphically as well as tabularly.

NEC also provides theta (vertical) and phi (horizontal) values of the electrical field of an antenna, in terms of voltage magnitude and phase (in volts per meter). Since no range is specified in the most typical pattern specifications (like the one we have been using in our example), the voltage values shown are the voltage as the range approaches infinity. I have shown them to illustrate some of the further data available for special purposes, and even further available columns of polarization information has been omitted.

Had we chosen a different example, the resulting NEC output file would have looked somewhat different, but mostly by way of additions. Over real ground (fast or Sommerfeld-Norton), there would be a ground calculation entry of considerable interest. Additional sources and loads would have enlarged the entries for these

items. Requests for alternative outputs—available in some implementations of NEC—would have produced substitutes for or additions to the pattern report we used in the sample.

For complex antenna structures and requests for further output data, a NEC output file can grow to very significant proportions—proportions that defy printing except in cases where preservation of the data is vital. For example, a frequency sweep of free-space azimuth patterns for an antenna on 10 meters from 28 to 29 MHz at 0.1 MHz intervals would have yielded 11 pattern reports, each with 360 lines for a 1-degree resolution setting. In some implementations, it is also possible to make multiple pattern requests. For the example, the added request might consist of one or more elevation patterns or of azimuth patterns taken at different elevation angles. The number of pattern reports in the frequency sweep would more than double. Now suppose we had chosen as our model antenna a 5-element quad beam, each element of which consisted of 4 wires with about 11 segments per side. The 880 segments would each appear in the segmentation data portion of the report.

For reasons such as these, entry-level implementations of NEC rarely make all of the data in the NEC output file available, although some make the file itself available for consultation. Instead, they extract the most crucial information from the file and present it in separate tables, often in formats that make reading easier. The most common change is to convert the engineering notation to plain decimal notation, with a limit set to the number of digits shown.

At the same time, implementations of NEC will often provide additional information that can be calculated from the NEC output file. The most prominent example is SWR data, which is normally referenced to 50 Ohms and to user-selected values. In some software, load loss figures are also available. These are only samples of many possible pre-NEC and post-NEC calculations that can enhance the basic core functions for the user.

We have passed over a number of NEC conventions that either are not relevant to our sample model or can be somewhat confusing at first encounter. Among the latter group is the fact that the NEC core registers all voltages and currents in peak values rather than in RMS values. For certain simple calculations, such as those directly derived from Ohm's Law, the difference makes no difference. Resistance

equals $E_{\text{peak}}/I_{\text{peak}}$ or $E_{\text{rms}}/I_{\text{rms}}$. However, when dealing with power, you must first convert the peak values to RMS values of voltage and current to arrive at the correct result.

If you move from one implementation of NEC to another, you may encounter a further potential confusion. Programs like EZNEC use the input and output interface to provide conversion so that as a user, you always enter RMS values and receive RMS values. However, other, such as NEC-Win Plus, use the native NEC convention of inputting peak values and receiving peak values in the output. Either system will produce the same patterns and tabular impedance and current data, assuming that the input values are numerically the same. However, the power report in the tables will differ.

There will be other inter-program differences as well. For example, one program may set the azimuth polar plot to show zero degrees at the top of the graph, while another will set the same position to the right. Moving from one program to another requires that you know the conventions of the implementations in order to ensure that you interpret the output clearly and accurately.

As you develop your modeling skills, you will discover that reference to the basic NEC output file becomes a normal part of the modeling routine. Familiarizing yourself with the types of data available and the ways in which the data is formatted can enhance the utility of the output file in your work. This note has at most pointed the way to the file, but has not yet penetrated significantly into what is available there. Penetration comes with the well-defined tasks that you will develop as you continue to model antennas.

* * * * *

Model included: 19-1.



20. The Average Gain Test

We have noted in a past column (#1 of this series) the convergence test for model adequacy. In that test, we increase the number of segments per wavelength until the output reports no longer vary from one level of segmentation to the next by a significant amount. For some antenna, such as dipoles, converge may occur with as few as 10-15 segments per wavelength, while other antenna structures may require many times that figure. There are some antenna models that never converge. Convergence is considered a necessary but not a sufficient condition of model adequacy. A model may have its results converge and still be a poor model for reasons that convergence testing cannot reveal.

In addition to the convergence test, there is another important test of model adequacy: the Average Gain Test. Like convergence testing, it is a necessary but not a sufficient condition of model adequacy. Although the raw data necessary for performing the test has always been in the NEC output file, only recently has it found its way into commercial implementations of NEC. It is available in current releases of NEC-Win Plus and EZNEC for Windows.

Essentially, we only need two numbers to perform the test: the input power and radiated power. For a lossless antenna, the input power and the average radiated power should be equal. Whatever the gain in one or more favored directions, it will be offset by nulls in other directions. Over the entire sphere of free space, the total amount of radiated power can never exceed the power supplied to the antenna. Hence, the ratio of average radiated power to supplied power should be 1. If the ratio differs by more than a small amount from 1, then the model may be considered suspect.

The conditions under which an adequate model will show an Average Power Gain (G_{AVE}) of 1 also establish the conditions for performing the Average Gain Test. The model is set in free space. (We shall look at setting the model over perfect ground in a moment.) The wire material must be perfect or lossless. All “real” or resistive parts of loads, networks, and transmission lines must also be set to zero

(which may require in a parallel R-L-C load a very high value for the parallel resistance).

For test purposes, the model is run by taking a regular sample of the radiation pattern every few degrees, and the results are averaged. (Note: for these tests, the sample is taken as a power and not as a power ratio, although one can be easily converted to the other.) The result is a statistically fair reading of the average radiated power. To calculate the average power gain, we simply apply the following simple equation:

$$G_{AVE} = \frac{k P_{RAD}}{P_{IN}} \quad (1)$$

where G_{AVE} is the average power gain, P_{RAD} is the radiated power as averaged, and P_{IN} is the input power as calculated from source information.

What about k ? For a free space model, $k = 1$. However, if the test lossless model is placed over perfect ground, then $k = 2$.

The results will not vary by much if the only loss in the antenna is wire loss for high conductivity materials of reasonably large diameters. However, for the most reliable figure of merit, the test is best run on a wholly lossless version of the model being tested.

The average gain figure that results from the test may be higher or lower than 1.0. One proposed gradation of model merit uses the following dividing points:

G_{AVE} Value Range	Significance
0.95 - 1.05	Model is considered to have passed the test and is likely to be highly accurate.
0.90 - 0.95 & 1.05 - 1.10	Model is quite usable for most purposes.
0.80 - 0.90 & 1.10 - 1.20	Model may be useful, but adequacy can be improved.
<0.80 and >1.20	Model is subject to question and should be refined

The user may develop more strict limits for the adequacy of a model based on the specific tasks within which the model plays a role.

Most models that deviate in the test from an average gain of 1 show an inverse correlation between errors in gain and in the resistive component of the source impedance. As the gain climbs, the source impedance decreases, and vice versa. For limited purposes, the average gain value derived from the test can be used to correct both figures, using the following equations:

$$\text{Corrected Input Resistance} = \frac{\text{Computed } R_{IN} \times G_{AVE}}{k} \quad (2)$$

and

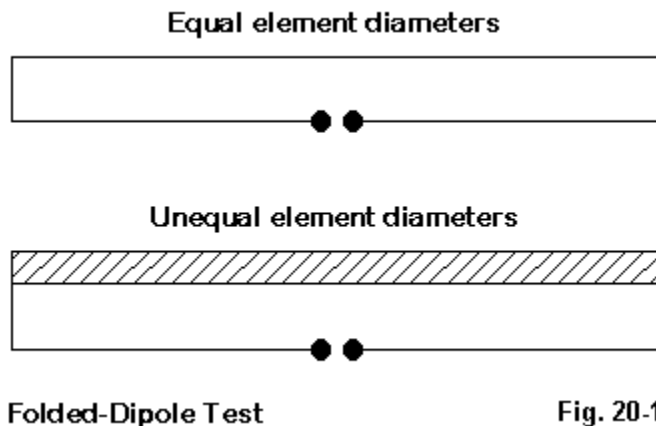
$$\text{Corrected Gain} = \frac{\text{Computed Gain} \times k}{G_{AVE}} \quad (3)$$

Obviously, an average gain value that is greater than 1 will increase the input resistance and decrease the gain. Values less than 1 will do the opposite. In no case should such corrections be applied and used unless the model has first been established as a good model relative to tests other than the Average Gain Test. As we shall see, the Average Gain Test can identify some questionable models, but certainly not all of them.

Some Test Cases

To familiarize ourselves with the Average Gain Test and its utility, let's look at a few test cases. All of them were tested for average gain using the NEC-Win Plus implementation of the test, which rewrites a given model to establish the necessary lossless conditions. All of the models were run in free space.

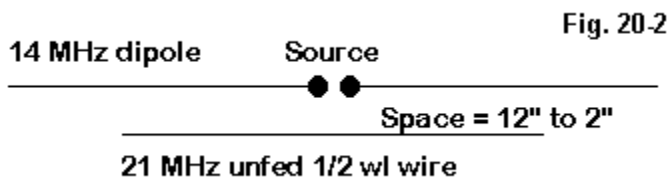
1. *A folded dipole with elements having different diameters.* **Fig. 20-1** shows the folded dipole we shall test, using the model shown in the lower portion of the drawing.



The test model uses a #12 wire for the source with a 0.5" wire as the thick element. With the ratio of wire diameters given, the source impedance should be transformed by a factor of over 7 relative to a standard dipole, or something around 530 Ohms. (Standard equations for calculating the impedance transformation of a folded dipole—taking into account both wire diameters and the spacing between wires—can be found in the *ARRL Antenna Book*.) The free space antenna gain should be that of a standard folded dipole, in the 2.1 to 2.2 dBi range. NEC-2 returns a gain of 0.76 and a source impedance of under 400 Ohms for a model that is resonant and which reports the approximately correct values when modeled carefully in MININEC. (MININEC returns a free space gain of 2.2 dBi and a feedpoint impedance of about 530 Ohms for this test case.)

The Average Gain Test gives a value of 0.75, indicating a highly untrustworthy NEC model. Note, however, that in this case, NEC returns values for gain and for impedance that are both less than one might reasonably anticipate from models of other folded dipoles and from calculations of the impedance transformation for the wire diameter ratios and wire spacing. Nonetheless, as a figure of merit, the Average Gain Test correctly characterizes the model.

2. *Close spaced elements of unequal length.* The basic set-up for this condition appears in **Fig. 20-2**.



A resonant 1" diameter 14 MHz dipole is placed close to an unfed wire, also 1" diameter, approximately resonant on 21 MHz. NEC reports the 14 MHz free-space gain and feedpoint impedance of this configuration in the following pattern:

Spacing in inches	Gain in dBi	Feedpoint Z (R +/- jX Ohms)
12	2.2	75 - j 0.3
4	2.4	74 + j 2.8
2	3.3	61 - j 0.1

Passing all three versions of the model through the Average Gain Test yields the following set of values:

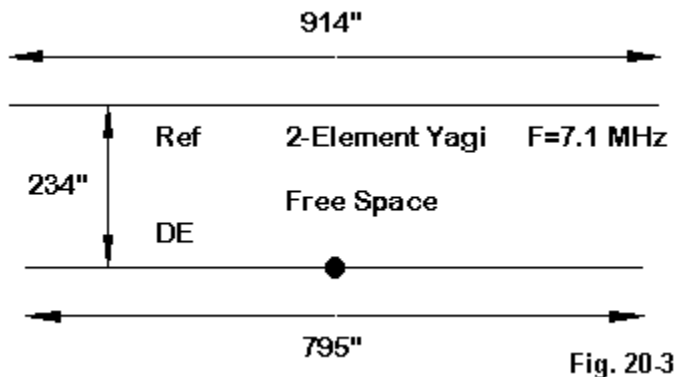
Spacing in inches	Average Gain
12	1.0000
4	1.0449
2	1.2547

With a spacing of 12", the model completely passes the test. The gain figure, which is very slightly high for a resonant dipole, is the maximum gain in the direction of the secondary element, and the configuration shows a front-to-back ratio of nearly 0.1 dB. Hence, everything is quite normal and usable.

The figure for the 4" spacing version suggests that the model is quite usable for most purposes. Here is a case where a user should feel free to reach a more strict requirement if the modeling task requires quite high precision—as do many studies

that systematically track performance predictions from NEC. The model with 2" spacing is clearly unusable for anything except accounts such as this one that point out its dangerously over-estimated gain and underestimated feedpoint impedance. In this example, the use of the Average Gain to correct the numbers yields quite good results.

3. *A 2-element 7.1 MHz Yagi with highly tapered element diameters.* The basic outline of the subject Yagi appears in **Fig. 20-3**.



From the center of each element outward, the prescribed element taper schedule is as follows (with all dimensions in inches):

Element Diameter	Section Length
4.7	12
2.25	108
2.0	120
1.0	66
0.75	69
0.5	to tip

The very large diameter center section simulates the effect of an element-to-boom mounting plate.

Without the Leeson corrections, NEC-2 returns for this 2-element Yagi a free space gain of about 6.2 dBi and a front-to-back ratio just above 8 dB. The feedpoint impedance is just about 39 Ohms. The Average Gain Test returns a value of 1.21, indicating a model of highly dubious adequacy. By way of contrast, the model with substitute uniform diameter elements produced via the Leeson equations yields a gain of 5.8 dBi with a front-to-back ratio of about 10.6 dB. The feedpoint impedance is about $41 - j36$ Ohms. Although the resistive component of this impedance appears to defy the correction function of the Average Gain Test, it does not. Since the driver is highly reactive, it must be lengthened to achieve resonance, and this action will increase the resistive component of the impedance well above the level of the uncorrected source impedance. Likewise, shortening the uncorrected model's driver to yield about the same capacitive reactance as the corrected model will also lower the value of the resistive component of that impedance.

Some Other Test Cases

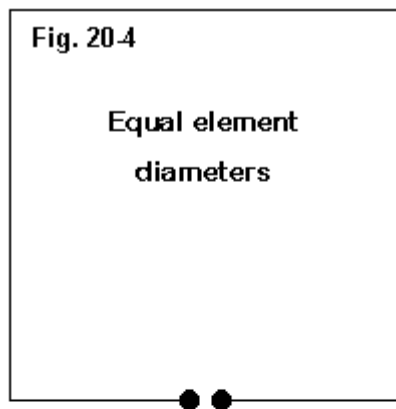
Although the three test cases we just ran generally confirm the utility of the Average Gain Test, that test does not reveal every inadequate model. A few illustrations may be useful.

4. *A quad loop with “fat” horizontal wires and “thin” vertical wires.* Let's construct a 10-meter quad loop like the one shown in the lower half of **Fig. 20-4**.

Quad loops constructed entirely of 1" “fat wires” or of #12 “thin wires” both have bi-directional maximum free space gain values close to 3.3 dBi, with resonant feedpoint impedances between 125 and 130 Ohms. Suppose we construct a composite quad consisting of horizontal fat wires and vertical thin wires. NEC-2 reports the gain as about 3.6 dBi and the feedpoint impedance as $180 + j110$ Ohms. By way of contrast, a MININEC model shows a feedpoint impedance closer to 135 Ohms, which is more in line with the mono-taper versions of the quad loop.

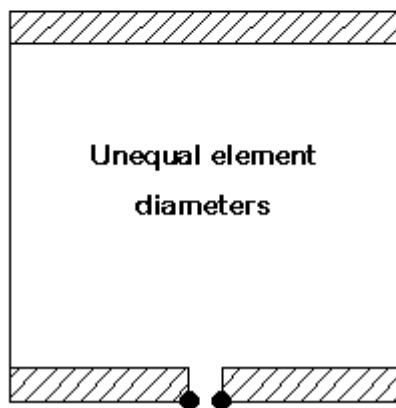
Fig. 20-4

Equal element
diameters



Corner-Junction Test

Unequal element
diameters



Interestingly, the Average Gain Test produces a value of 1.006, indicating a highly reliable model—at least so far as that test can indicate. However, the quad with corner junctions of dissimilar diameter wires is an odd NEC model to be sure. Although its reported gain remains stable, the source impedance varies widely according to the number of segments per side. Some modelers believe that the most accurate results come not from the highest segmentation density, but from the lowest. Whatever the final verdict on the model, it is a case of a dubious model that the Average Gain Test cannot catch.

5. *A corner-fed non-symmetrical 40-meter triangle.* Suppose we construct a right-angle triangle of #12 AWG wire and feed it at the lower corner, as shown in **Fig. 5**.

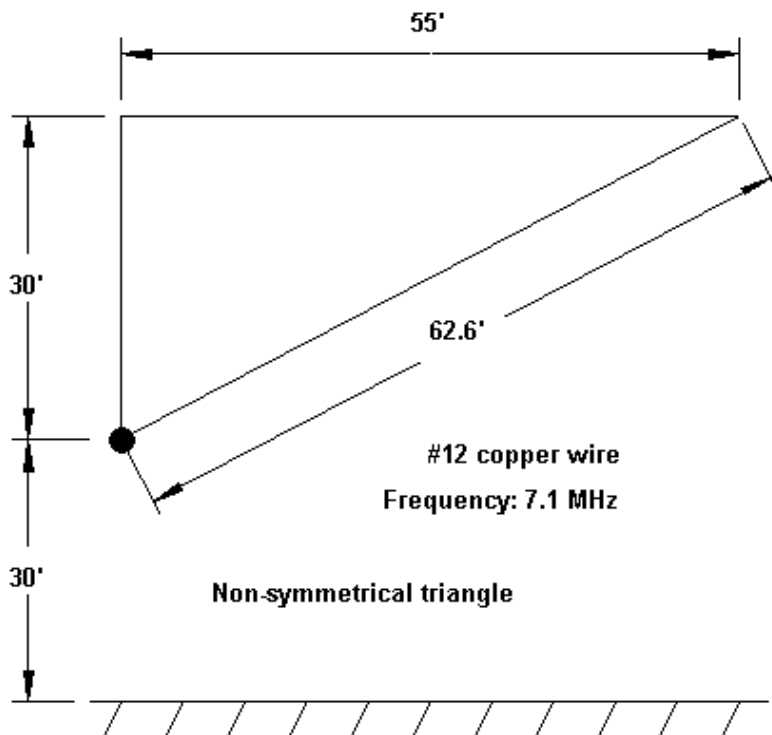


Fig. 20-5

Even at low segmentation densities, the Average Gain Test returns values such as 0.96, indicating a highly reliable model. However, this model requires many segments for results to convergence, due largely to the lack of symmetry in current magnitude adjacent to the feedpoint. Unfortunately, the Average Gain Test can give no clue as to what this model requires for reliable results.

6. *A folded-X beam for 28.5 MHz.* Let's construct a folded-X beam using 1" wire for the X-portion and #18 wire for the "tails," as suggested in **Fig. 20-6**.

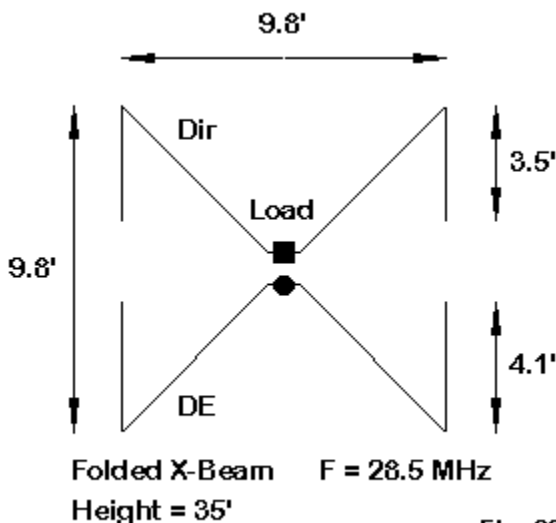


Fig. 20-6

The folded X-beam is a highly unstable model in NEC-2. It will not converge at any level of segmentation density tried. The highly angular junctions of wires having very disparate diameters are the most likely source of the convergence failure. NEC-2 cannot provide any guidance for the dimensions of this antenna that will yield resonance and desired levels of gain and front-to-back characteristics. MININEC

will provide the necessary guidance so long as the modeler gives due care to the sharp corners by using segment length tapering techniques.

However, the Average Gain Test certifies the folded X-beam model as reasonably reliable with a value of 0.916.

The failure of the Average Gain Test to reveal the defects of these last three examples is not a flaw of the test. Instead, these cases simply illustrate that passing the test is a necessary but not a sufficient condition of model adequacy. Adequate models will pass the test, but bad models will not always fail it.

Conclusion

The Average Gain Test is a valuable addition to the collection of tests for evaluating the structural adequacy of our models. When combined with the convergence test, most models that go bad structurally can be detected. However, since each is a necessary but not sufficient condition of adequacy, there is no guarantee that some few models will pass both tests and still prove to be inadequate as models.

Notice that I qualified the idea of adequacy by referring to “structural” adequacy, that is, the basic geometry of the antenna. There are additional inadequacies that can result from using networks, transmission lines, and loads in less than optimal ways. Because any resistive component of these mathematical additions to the model will be eliminated for the purpose of running the Average Gain Test, that test may not show flaws resulting from networks, transmission lines, and loads.

In the end, there may be no single test or battery of tests that will automatically detect all inadequate models. To the best of my knowledge, there has been no systematic attempt to identify and isolate the specific properties of antenna models which are likely to fail either the convergence test or the Average Gain Test (or both). The small sample in this note is hardly even a beginning in that direction. In the end, knowing from experience when values are climbing or falling outside the boundaries of good sense and basic antenna theory is likely the last line of defense against an inadequate model. Nonetheless, the Average Gain Test is a very useful addition to our repertoire of evaluative tools in modeling.

* * * * *

Models included: 20-1 through 20-6. Note: the less reliable the model in terms of its AGT score, the more that the AGT values might vary between EZNEC and NEC-Win Plus.



21. The NEC TL Facility

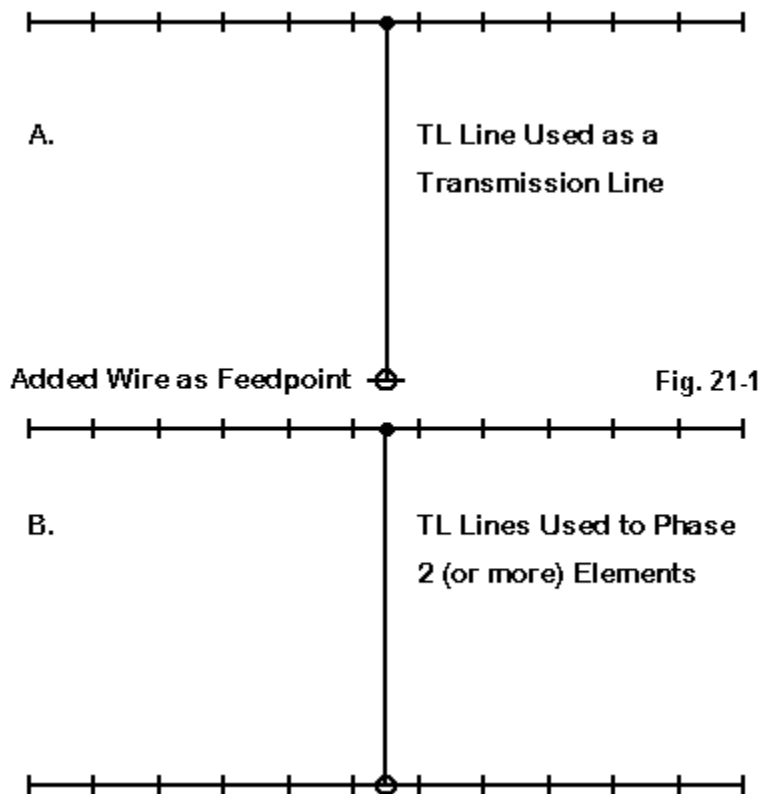
We have mentioned the use of the NEC TL facility along the way, but perhaps a concentrated dose of TL might get us better oriented to using mathematical transmission lines in NEC. Note that this facility is not available in public domain versions of MININEC.

First of all, let's look at the rules for when it is apt to use transmissions lines and what their limitations may be. The TL input for NEC (-2, -3, -4) is a special form of network input, designed specifically to enable modelers to place transmission lines into a model. Since we are dealing with a network, we must have two wires on which to place the ends of the line we create.

For applications where the line connects two antenna elements, the requirement for 2 wires poses no problem, as shown in **Fig. 21-1**, part B. However, when we wish to connect a transmission line to a source to see the effects of line length on the impedance, or when we wish to use transmission line stubs as reactances, then we must create the second wire, as shown in **Fig. 21-1**, Part A.

For the types of cases just noted, the second wire should be positioned and sized so that it plays no significant role in the far field of the antenna itself. This requirement forces the wire to be at some considerable distance from the antenna—the farther the better. Moreover, the wire should be thin and short—a single segment wire just meeting the lower limits of radius and length works fine. The wire should itself be lossless, if that option is available.

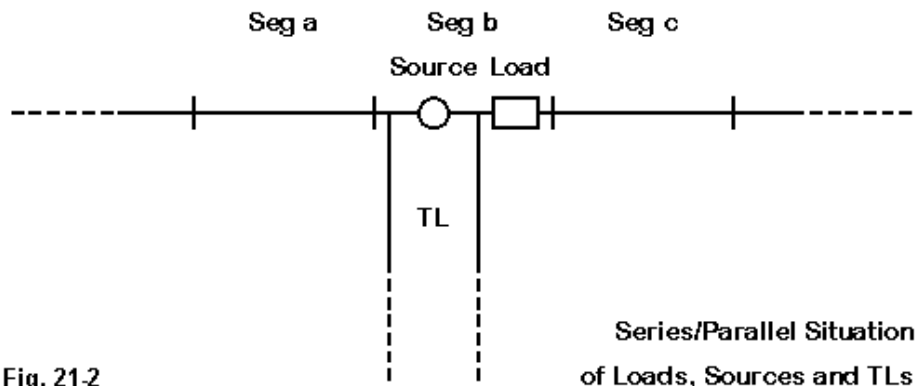
The position of the terminating wire makes no difference to the length of the transmission line in the mathematical model. The user will always specify the length of the transmission line, and this value will supersede the distance created by the modeled placement of the terminating wire.



The user will specify the segments of wires that represent the terminal points of the transmission line. As well, the user will specify the length and the characteristic impedance of the line. The basic length is the electrical length, assuming a velocity factor of 1.0. Some programs allow the user to specify the physical length and the

velocity factor: the program then uses those numbers to calculate the electrical length for actual NEC calculations.

Since the line created by the TL entry is non-radiating and mathematical only, the line itself plays no role in the generation of the antenna's far field. This feature may be a modeling convenience in many cases, but in some other cases, it may represent a limitation on the model. If the transmission line plays a role in the antenna pattern, as is the case in many off-center-fed wire antennas, a physical model of the transmission line may be preferable to use of the TL facility.



Sources and loads on the same segment appear in series with each other. TL entries are different. They appear in parallel with sources and loads that are placed on the same segment, as shown in **Fig. 21-2**. This factor must be kept in mind when trying to add a TL transmission line to a source segment that is already loaded. This feature is both a limitation and an opportunity—as we shall see further on.

Since TL transmission lines are mathematical, they show no losses. If you wish to know the losses involved in a transmission line application, you must hand calculate them or use other software, such as TL by N6BV. These calculations may also

reveal slight differences in the 2:1 SWR bandwidth of some models relative to the lossless line in the model.

The TL line will be accurate only if applied under balanced conditions, that is, where the current on each side of the line is the same. This restriction dictates that TL lines be used at such places as the center segment of a linear horizontal element. Applying a TL line off center will yield inaccurate results. Theoretically, one might apply a TL transmission line as a phasing section between two collinear dipoles. However, the current and voltage at the junction point are changing so rapidly that equalized currents are rarely feasible. For such applications, modeling a physical parallel transmission line is advisable.

How to Implement TL Transmission Lines

Depending upon the level of user help in a given implementation of NEC, the user may have to do little calculation—or a lot. EZNEC, for example, allows you to specify the two ends of the line—and offers preset options if you wish a shorted or open stub. These options eliminate the task of specifying a new wire in the model that represents the TL termination, since the program automatically generates the distant wire. For a shorted stub, the program specifies a very low resistance value as a load on the far wire—perhaps 1E-10 Ohms. For an open stub, the program loads the wire with a very high value of resistance, perhaps 1E10 Ohms.

With raw NEC and some programs, you must introduce the stub-terminating wires as added wires in the model description. Some programs require that you give them tag or wire numbers over a certain value so that they will not appear in views of the antenna.

You must then specify the characteristic impedance of the transmission line being mathematically simulated. EZNEC permits the entry of a velocity factor, and it also has a catalog of common transmission lines with their associated velocity factor values. Other software will require that you enter the electrical length of the line as well as the characteristic impedance. If this requirement applies to your software, you may simply divide the physical length by the velocity factor to determine the electrical length. Use care: even though you may have entered the antenna geometry in inches or feet, some programs will require that TL lengths be entered in meters, thus requiring one more conversion before finalizing the entry.

TL - Transmission Line

Place Transmission Line

From Tag #: 1 Segment #: 8

To Tag #: 2 Segment #: 1

Shunt Admittance

Real **Imaginary**

End One: 0 0 mhos

End Two: 0 0 mhos

Transmission Line Type

☒ No Phase Reversal

☐ 180° Phase Reversal

Characteristic Impedance (ohms): 50

Transmission Line Length (m): 10.6

☒ OK ☐ Cancel

Fig. 21-3

Fig. 21-3 shows a data entry box where no short-cuts are available. Here, standard NEC language is used. The “Tag” numbers are the wire numbers in the wire description portion of the model description, while the segment numbers are referenced to the segments themselves and not to percentages of the way along a wire. Hence, you must be careful to identify the correct segment number for exact placement of the TL. If you change the segmentation of the wire, you must return to this box and change the segment number for the TL on that wire if you wish it to be placed at a certain position, for example, at the wire center.

The line characteristic impedance is given as 50 Ohms, with an electrical length of 10.6 meters. This might represent 34.78' of line, if you did your original calculations in feet. If the velocity factor of the line is 0.66, then the physical length of the line would be only 22.95'. Be sure to keep records on paper of all calculations that lead up to NEC entries so that you can check your work later for errors.

The standard NEC entry for networks is in terms of admittance rather than impedance. Therefore, had we been building a shorted or open stub, we would need

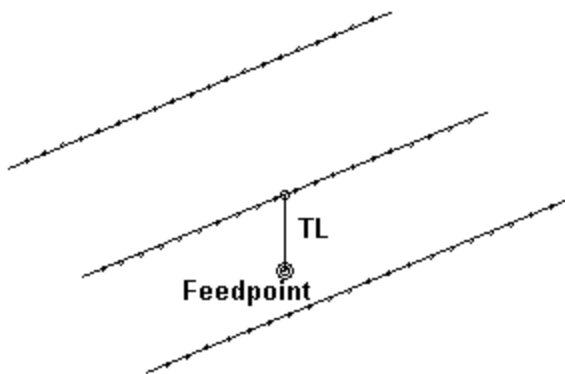
an entry here. A shorted stub would have virtually no resistance. In terms of admittance, we would insert a very high value, perhaps 1E10 mhos (now called siemens). An open stub would require a very high resistance and require a very low value of admittance, perhaps 1E-10 in the real column. For stubs, you may ignore the “imaginary” column, which would take values of susceptance, the inverse of reactance.

A Few Real Examples

To gain a better handhold on the use of TL transmission lines, let's look at a few examples involving real design questions. For this exercise, we shall use a few sketches and some model descriptions taken from EZNEC.

1. *A 3-Element 15-Meter Yagi With a Quarter Wavelength Matching Section:*

Fig. 21.4



**3-Element 15-Meter Yagi
with 1/4 WL TL Matching Section**

Fig. 21-4 shows the outline of a 3-element 15-meter Yagi of good performance (8+ dBi free-space gain and 26 dB front-to-back ratio at 21.2 MHz). The elements have been sized so that the antenna is resonant at the design frequency. The feedpoint impedance is $24.8 + j\,0.4$ Ohms, too low for a match to the standard 50-Ohm coaxial cable.

However, we can insert a 35-Ohm quarter-wavelength matching section to effect the match. To do so, we must add a 4th wire that is very short and of small diameter. Then, we add a quarter wavelength section of feedline according to the entry rules of the software we happen to be using. The line would be 11.6' or 3.54 m long electrically. If the velocity factor is taken into account, the line would be 7.655' long physically.

Below is the model description of the final arrangement:

3 el Yagi 3/4" el Frequency = 21.2 MHz.

Wire Loss: Aluminum - Resistivity = 4E-08 ohm-m, Rel. Perm. = 1

WIRES

Wire Conn.	End 1 (x,y,z : ft)	Conn.	End 2 (x,y,z : ft)	Dia(in)	Segs
1	-11.554, 0.000, 0.000	11.554, 0.000, 0.000	7.50E-01	21	
2	-11.032, 6.991, 0.000	11.032, 6.991, 0.000	7.50E-01	21	
3	-10.381, 15.074, 0.000	10.381, 15.074, 0.000	7.50E-01	21	
4	-0.100, 6.991, -5.000	0.100, 6.991, -5.000	# 14	1	

SOURCES

Source	Wire Seg.	Wire #/Pct Actual	From End 1 (Specified)	Ampl.(V, A)	Phase(Deg.)	Type
1	1	4 / 50.00	(4 / 50.00)	1.000	0.000	V

No loads specified

TRANSMISSION LINES

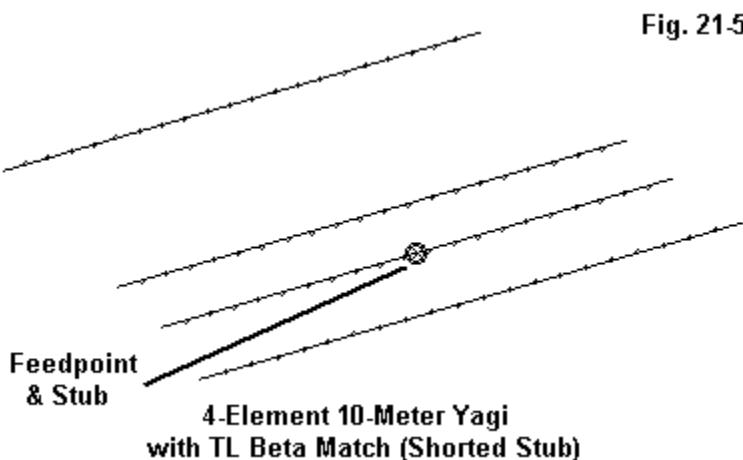
Line	Wire #/% Actual	From End 1 (Specified)	Wire #/% Actual	From End 1 (Specified)	Length	Z0 Ohms	Vel Fact	Rev/ Norm
1	2/50.0	(2/50.0)	4/50.0	(4/50.0)	7.655 ft	35.0	0.66	N

Ground type is Free Space

The feedpoint impedance at the design frequency with the line in place is $49.4 - j0.8$ Ohms. Remember that TL entries are lossless, so there may be a small loss in the section that must be independently calculated.

Most interesting is the placement of the tiny wire (#4) used to terminate the transmission line and also to serve as the model source. I purposely placed it close to the antenna and vertically in line with the driven element. You should move this wire around in order to get some feel for how close or distant such added wires must be to have no significant effect on the antenna performance. In this case, the wire can be quite close without changing any of the performance figures. However, a distant wire placement is always in order.

2. A 4-Element 10-Meter Yagi With a Beta Match (Shorted Stub):



In **Fig. 21-5**, we have the outline of a 4-element Yagi. At 28.5 MHz, the design frequency, the antenna shows a free-space gain of 8.6 dBi with a 22 dB front-to-back ratio. The natural feedpoint impedance of the antenna is $25.8 - j28.8$ Ohms, a good candidate for a beta match. The simplest form of the standard beta match

used to compensate for capacitive reactance is a hairpin or shorted stub having the correct inductive reactance according to standard L-circuit calculations. We can simulate the hairpin by creating a shorted stub of a reasonable characteristic impedance at the source junction.

For the model in this exercise, I used a 5" stub of 450-Ohm line with a velocity factor of 1.0. The EZNEC model description looks like this:

```

4 el 13' boom                                Frequency = 28.5 MHz.

Wire Loss: Aluminum - Resistivity = 4E-08 ohm-m, Rel. Perm. = 1

      ----- WIRES -----
Wire Conn.- End 1 (x,y,z : in)      Conn.- End 2 (x,y,z : in) Dia(in) Segs

1      -103.73, 0.000, 0.000      103.731, 0.000, 0.000 5.00E-01 21
2      -97.942, 35.811, 0.000      97.942, 35.811, 0.000 5.00E-01 21
3      -97.226, 65.653, 0.000      97.226, 65.653, 0.000 5.00E-01 21
4      -91.078,152.195, 0.000      91.078,152.195, 0.000 5.00E-01 21

      ----- SOURCES -----
Source      Wire      Wire #/Pct      From End 1      Ampl.(V, A)      Phase(Deg.)      Type
            Seg.      Actual      (Specified)

1           11       2 / 50.00      ( 2 / 50.00)      1.000            0.000            V

No loads specified

      ----- TRANSMISSION LINES -----
Line      Wire #/%      From End 1      Wire #/%      From End 1      Length      Z0      Vel Rev/
            Actual      (Specified)      Actual      (Specified)

1      2/50.0 ( 2/50.0) Short ckt (Short ck)      5.000 in      450.0      1.00

Ground type is Free Space

```

In some software, you would need to create the distant wire for the shorted end of the TL stub and give it a very high admittance value. You may even have to convert the 5" length to its equivalent in meters.

With the stub specified, the feedpoint impedance is $43.4 + j25.2$ Ohms, something less than a perfect match at the design frequency of 28.5 MHz. However, the

stub length was not chosen for a perfect match at one frequency. Rather, the length was experimentally determined based on the goal of obtaining the widest 2:1 SWR operating bandwidth between 28 and 29 MHz. You should experiment with stub lengths (and characteristic impedances) to see the effects on SWR, both at the design frequency and across a desired band segment. Other values might easily improve on the operating bandwidth. You may also change the length of the driver to change the amount of capacitive reactance for which the beta match must compensate. The results may make full coverage from 28 to 29 MHz either easier or more difficult.

At a specific design frequency with the correct value of capacitive reactance to accompany the resistive portion of the feedpoint impedance, a beta match can be exceptionally efficient. However, as one moves away from the ideal conditions, small losses may accrue as the match must not only effect an impedance change, but must also compensate for non-ideal reactance values.

3. *A 3-Element 10-Meter Phased Array With Director and a Beta Match (Open Stub):*

Fig. 21.6

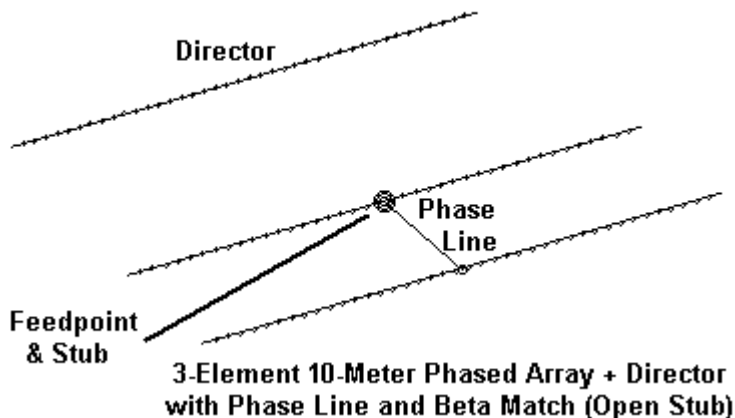


Fig. 21-6 presents two uses of TL transmission lines in one example. There is a reversed phasing line between the rear 2 elements of the array. In addition, there is a beta match to compensate for the natural $15.5 + j23.0$ Ohms of the feedpoint impedance. Because the feedpoint reactance is inductive, an open stub will be required, calculated according to standard L-circuit techniques.

3-el ZL Sp 10m exp.

Frequency = 28.5 MHz.

Wire Loss: Copper - Resistivity = $1.74E-08$ ohm-m, Rel. Perm. = 1

WIRES									
Wire Conn.-	End 1 (x,y,z : ft)	Conn.-	End 2 (x,y,z : ft)	Dia(in)	Segs				
1	-7.526, 7.876, 0.000	7.526, 7.876, 0.000	6.56E-01	41					
2	-8.313, 0.000, 0.000	8.313, 0.000, 0.000	6.56E-01	41					
3	-8.401, -4.375, 0.000	8.401, -4.375, 0.000	6.56E-01	41					

SOURCES						
Source	Wire Seg.	Wire #/Pct Actual	From End 1 (Specified)	Ampl.(V, A)	Phase(Deg.)	Type
1	21	2 / 50.00	(2 / 50.00)	1.000	0.000	V

No loads specified

TRANSMISSION LINES								
Line	Wire #/% Actual	From End 1 (Specified)	Wire #/% Actual	From End 1 (Specified)	Length	Z0 Ohms	Vel Fact	Rev/ Norm
1	2/50.0	(2/50.0)	3/50.0	(3/50.0)	5.776 ft	50.0	0.66	R
2	2/50.0	(2/50.0)	Open ckt	(Open ckt)	3.201 ft	50.0	0.66	

Ground type is Free Space

The potential performance of this modeled antenna shows (at 28.5 MHz) a free-space gain of about 8.6 dBi and a front-to-back ratio of about 33 dB.

The phasing line in the model is evident, since it runs between wires 2 and 3 of the antenna structure. Note that the phase line has been reversed to achieve the correct current magnitude and phasing on the rearmost element. (For this example, we need not consider whether the use of 50-Ohm transmission line with a 0.66 velocity factor is feasible for a practical antenna.)

The open circuit stub forming the beta match is about 3.2' of the same material. The wire for the termination, invisible in EZNEC, is distant from the array and supplied with a very low value for admittance to effect the open circuit condition.

The final array shows a feedpoint impedance at 28.5 MHz of $46.1 + j12.7$ Ohms. Once more, the stub length was not chosen for the lowest SWR at the design frequency. Instead, the length provides the widest possible 2:1 SWR operating band width for the array. You should experiment with other lengths and transmission line types to see the effect that these changes have on the SWR operating bandwidth.

These examples should be sufficient to get you started on the road to effectively using the TL facility within NEC. Remember the requirements for balance and the fact that TL transmission lines are in parallel with sources and in series with loads. Next time, we shall look at a few occasions where the physical construction of transmission lines may provide superior modeling results to the use of TL lines.

* * * * *

Models included: 21-1 through 21-3.

22. Modeling Physical Transmission Lines

Modeling transmission lines between the source and the antenna can take two forms. We can physically model parallel transmission lines. And, as we saw in the last column, we can model them as TL inputs.

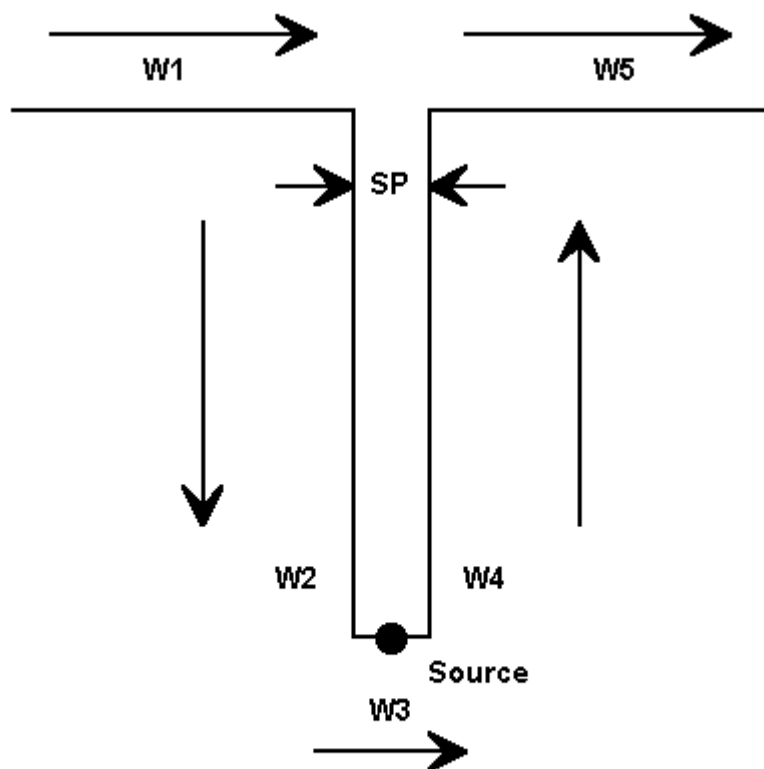
Physical Models of Parallel Transmission Lines

Consider a simple antenna fed at the center, as sketched in **Fig. 22-1**. The source is in series with the antenna wire. Any transmission line we use to feed the antenna thus forms a series circuit with the antenna wire. At the base of the transmission line, the actual source is now in series with the two wires making up the line.

These simple facts determine the basic rules for modeling a parallel transmission line as a parallel set of wires. First, we break the single antenna wire element into two segments, with the centermost ends spaced the same distance apart as the spacing of the wires in the transmission line.

In the figure are a set of arrows indicating ends 1 and 2 of each wire, using a left-to-right convention for this example. Note that we keep the series of wires continuous so that End 2 of Wire 1 connects to End 1 of Wire 2. The model progresses continuously from one end to the other. This is crucial to ensure that the transmission line has equal and opposite currents at every point. (For many placements of transmission lines, the “equal and opposite currents” situation will not fully materialize in reality. However, we should always model the transmission line in the manner shown.)

The source wire may simply be a one-segment wire from one side of the transmission line to the other. This wire will only be as long as the spacing between transmission line wires, and its segment length sets some specifications for the transmission line segmentation.



**The Basics of Physically Modeling
a Parallel Transmission Line**

Fig. 22-1

Segments that meet at angular junctions should be about the same length and in any event within about a 2:1 length ratio. Closely spaced transmission lines require much shorter segments due to the close spacing of their wires. The similar junction at the top of the assembly, where transmission lines wires and antenna wires meet, dictates short antenna wire segment lengths. Even for simple antennas, the entire assembly can quickly grow past the overall segment limit (usually about 500) for modestly priced commercial NEC packages.

One possible solution is to taper the segments lengths along the transmission line, especially at the end closest to the source. EZNEC provides for element length tapering. However, the shortest segment length selected should not be the standard value (0.0025 wavelength), but should match the element spacing. However, NEC-2 recommends that no segment be shorter than 0.001 wavelength long, which the spacing of a 300-ohm line will violate at many HF frequencies. Errors for lines in the 450-ohm to 600-ohm range run about 5 ohms difference from full segmentation (at the test frequency of 14.15 MHz), but length-tapered transmission line models may save from 30 to 130 segments per 1/4 wavelength of each side of the transmission line.

Physical transmission lines are usually calculated according to one of two equations:

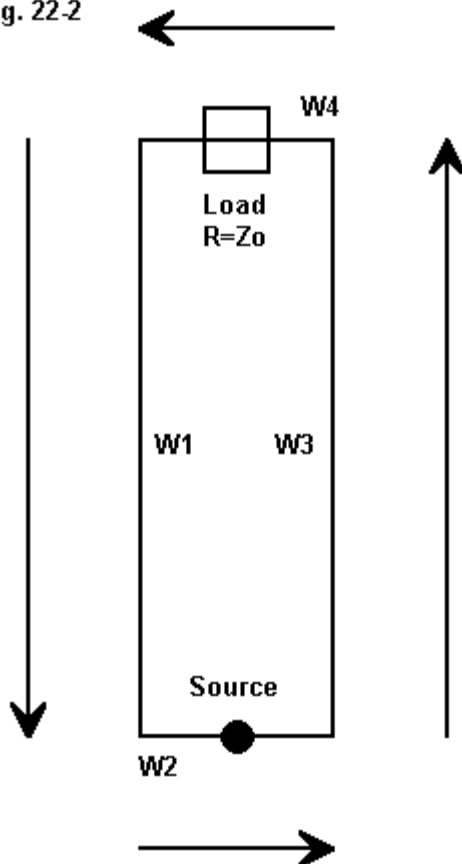
$$Z_o = 276 \log \frac{2S}{d} \text{ or } Z_o = 120 \cosh^{-1} \frac{S}{d}$$

where Z_o is the characteristic impedance of the resulting air-insulated parallel feedline, S is the center-to-center wire spacing, and d is the wire diameter in the same units as S . The first form is handiest for calculators, while the second form is somewhat more accurate for closely spaced lines.

Program suites such as HAMCALC have modules for calculating the Z_o from the wire size and spacing of air-insulated lines—or for calculating any third factor from the other two. Note that these equations do not account for insulation other than air, and the velocity factor is either 1.0 or very close to it. Therefore, any models of physical transmission lines should be based on a velocity factor of 1.0

and the results later readjusted for the actual velocity factor of the line used. Likewise, any losses due to the dielectric must also be accounted for after modeling.

Fig. 22-2

**Transmission-Line Test Model**

Interestingly, an extensive series of models using NEC-4 reveals that transmission lines between 300 and 600 ohms show a higher impedance than that calculated for them with the equations. **Fig. 22-2** shows the test modeling arrangement (see column #15). Instead of an antenna structure at the top of the transmission line, the lines were bridged with a resistive load.

The following table shows the results of using the arrangement with a carefully adjusted 1/4-wavelength line of #14 AWG copper wire at 14.15 MHz (210"). Theoretically, the source impedance should be the same as the load resistance only when both are equal to the line impedance, according to the standard 1/4 wavelength line equation:

$$Z_o = \sqrt{R_L Z_s}$$

where Z_o is the characteristic impedance of the line, R_L is the load resistance, and Z_s is the reported source impedance, all in Ohms.

First, there is a small and uneliminable reactance at the source from the test model. Second, the wires are very close together, approaching the limits of NEC-4 to handle, even when segments are carefully paralleled. In all cases, the modeled impedance is higher than the calculated impedance by from 5% to 7%.

Line Zo Calculated	Wire Dia.	Spacing (inches)	Modeled Zo in Ohms	New Wire Size for Zo
600	0.064"	4.78"	635	0.086"
450	0.064"	1.37"	483	0.084"
300	0.064"	0.39"	316	0.075"

All results used segment lengths equal to the wire spacing. As the segment lengths were lengthened, the 600-ohm line was least affected, but the 300-ohm and 450-ohm lines changed values for each new segment length tried—without a definitive pattern. As the segment lengths increased, the far field from the test model increased from nearly -40 dBi to values in the low 30s. Increasing the wire size brought the line impedance close to the calculated value. Equally, one might widen the spacing.

NEC-2 results with the same models correlated well with NEC-4 results with segment lengths greater than 0.005 wavelength. With shorter segments, the two programs apparently respond somewhat differently to closely spaced wires. Tapered-length transmission line models correlated well between the two programs, especially in the 450-ohm to 600-ohm Z_0 range.

For many applications, the exact line impedance may be unimportant. However, when physically modeling lines where their impedance does make a difference, running a test model to determine the line Z_0 is recommended. Physical models of transmission lines are at best somewhat cumbersome. However, they are important to any application where the lines may experience unequal currents and therefore radiate. Typical of such applications is the off-center-fed wire antenna. Comparing the far fields (including both vertical and horizontal components) of the antenna alone and the antenna with its parallel feedline will provide an estimate of such things as the field shape modification occasioned by the feedline, any change of elevation angle of maximum radiation, and any far-field gain changes. An examination of the currents along the transmission line may also be useful in determining the degree of unbalance occasioned by the off-center feed point. Similar exercises may prove useful with other antenna types.

Within their limitations, physically modeled parallel transmission lines can provide useful information, even if in some cases exactitude is sacrificed. To date, I have found no useful technique to model physically a coaxial cable such that the cable model fits within the segmentation limitations of versions of NEC aimed at radio amateurs.

Some Applications for Physically Modeled Transmission Lines

1. *The Behavior of Off-Center-Fed Antennas*

In **Fig. 22-3**, we see three on many possible configurations of an off-center-fed 1/2-wavelength wire antenna (modeled with #12 copper wire at 7.15 MHz for this example at a height of 100' above average ground). We can model this antenna without a transmission line or with any length of line we might choose.

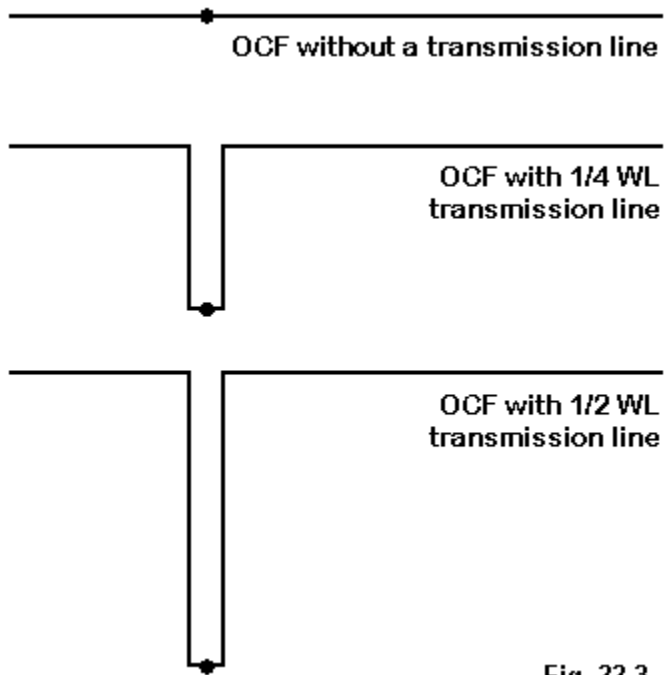


Fig. 22-3

For the example, I have used 1/4-wavelength and 1/2-wavelength transmission lines, each composed of #12 copper wire spaced 4.78" apart. The characteristic impedance of the line is approximately 600 Ohms. The segment lengths throughout the model adhere to this length, so that the model with a 1/2-wavelength line is sizable. Its description appears below as a sample: dimensions are in inches.

40 dipole 7.15 MHz

Frequency = 7.15 MHz.

Wire Loss: Copper - Resistivity = 1.74E-08 ohm-m, Rel. Perm. = 1

WIRES							
Wire Conn.-	End 1 (x,y,z : in)	Conn.-	End 2 (x,y,z : in)	Dia(in)	Segs		
1	-244.00, 0.000,1200.00	W2E1	-2.390, 0.000,1200.00	# 12	51		
2	W1E2 -2.390, 0.000,1200.00	W3E1	-2.390, 0.000,374.620	# 12	172		
3	W2E2 -2.390, 0.000,374.620	W4E1	2.390, 0.000,374.620	# 12	1		
4	W3E2 2.390, 0.000,374.620	W5E1	2.390, 0.000,1200.00	# 12	172		
5	W4E2 2.390, 0.000,1200.00		557.000, 0.000,1200.00	# 12	117		

SOURCES						
Source	Wire Seg.	Wire #/Pct Actual	From End 1 (Specified)	Ampl.(V, A)	Phase(Deg.)	Type
1	1	3 / 50.00	(3 / 50.00)	1.000	0.000	I

No loads specified

No transmission lines specified

Ground type is Real, high-accuracy analysis

Conductivity = .005 S/m Diel. Const. = 13

MEDIA				
Medium	Conductivity(S/m)	Dielectric Const.	Ht(in)	R Coord(in)
1	5.000E-03	13.00	0 (def)	0 (def)

I purposely chose the line lengths to illustrate that physically modeling transmission lines can reveal antenna system behavior changes that would not show up without the line. Here is a brief table of results.

Antenna	Gain DBi	TO Angle degrees	Feedpoint Z R +/- jX Ohms
No line	7.46	19	107 + j 0
1/4 wavelength line	2.78	17	95 + j23
1/2 wavelength line	5.77	18	156 - j61

Fig. 22-4 shows the differences in the azimuth patterns for the three versions of the antenna, each taken at the elevation angle of maximum radiation. The change

in feedpoint impedance is partially a function of the mismatch between the transmission line and the characteristic impedance of the line. However, it is also a partial function of the imbalance of current along the line.

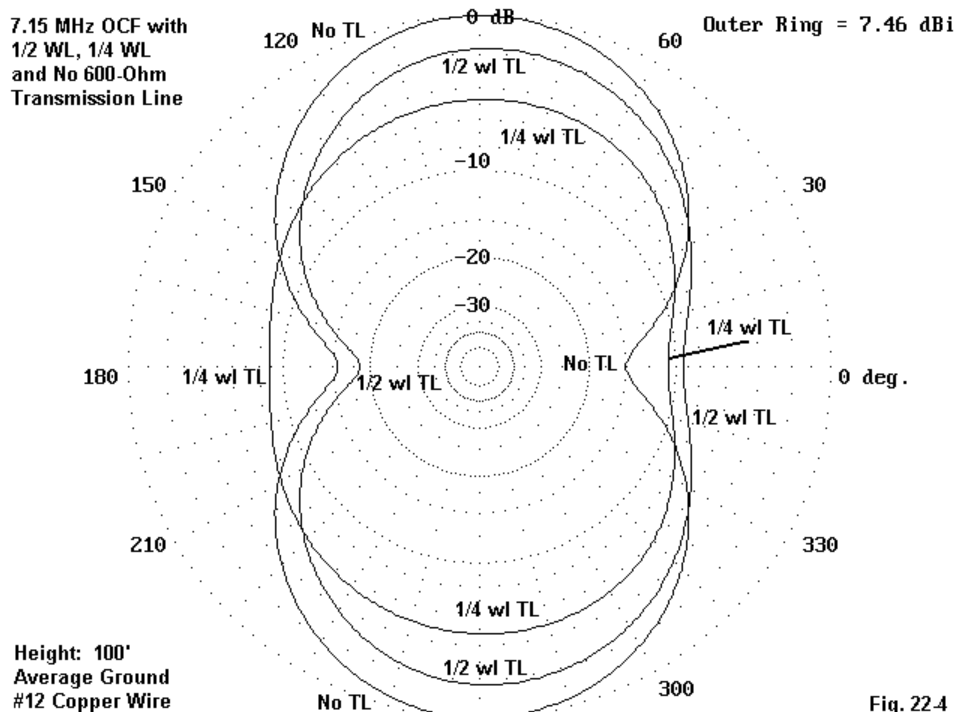


Fig. 22-5 shows the current magnitudes along both the antenna and the transmission line for the version with a 1/2 wavelength line. Feedline radiation is not solely a result of current magnitude on the line. Instead, it stems from the difference of current magnitudes and the amount by which they fail to be 180 degrees out of phase. Consequently, it pays to examine the current tables for antennas such as this one in order to uncover the degree of imbalance on the line.

As an exercise, you might well try other line lengths, especially those that are not an easy fraction of a wavelength.

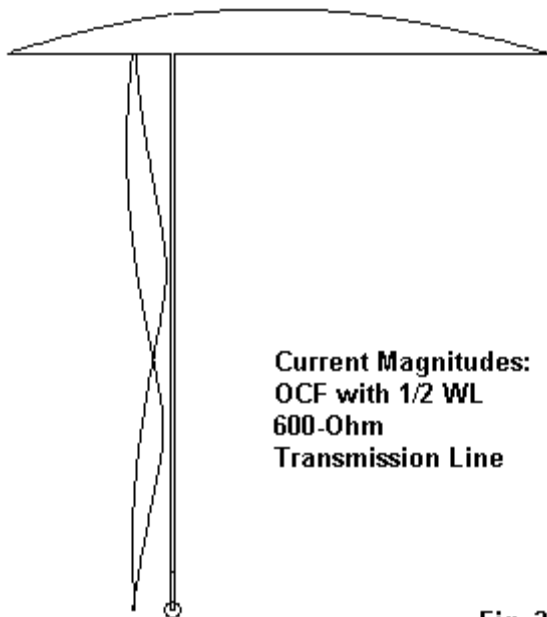


Fig. 22-5

2. A Collinear EDZ Array With Phasing Lines<

In **Fig. 22-6**, we have a collinear extended-double-Zepp (EDZ) for 21.225 MHz. It is made 129' of #12 copper wire and is 46' above average earth—about 1 wavelength up. For the moment, we may ignore the current magnitude lines on the graph except to note the multiple occurrences of the typical EDZ current distribution.

Fig. 22-6 shows the construction of physical models of the phasing lines, composed here of 7' lengths of #12 wire spaced 6" between wires. These shorted stubs have a characteristic impedance of about 650 to 700 Ohms. The description file below shows the antenna geometry fully.

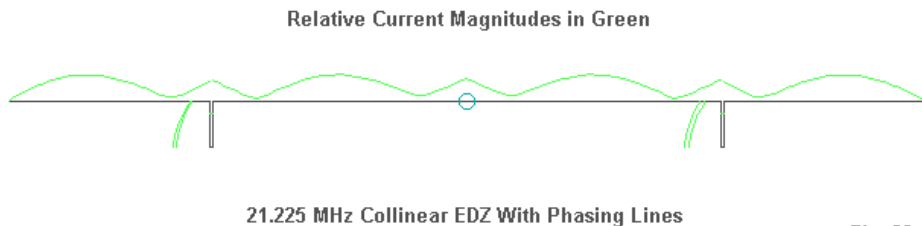


Fig. 22-6

Collinear EDZs

Frequency = 21.225 MHz.

Wire Loss: Copper - Resistivity = 1.74E-08 ohm-m, Rel. Perm. = 1

WIRES									
Wire Conn.-	End 1 (x,y,z : ft)			Conn.-	End 2 (x,y,z : ft)			Dia(in)	Segs
1	-64.500,	0.000,	46.000	W2E1	-36.250,	0.000,	46.000	# 12	33
2	W1E2	-36.250,	0.000,	46.000	W3E1	-36.250,	0.000,	39.000	# 12 5
3	W2E2	-36.250,	0.000,	39.000	W4E1	-35.750,	0.000,	39.000	# 12 1
4	W3E2	-35.750,	0.000,	39.000	W5E1	-35.750,	0.000,	46.000	# 12 5
5	W4E2	-35.750,	0.000,	46.000	W6E1	35.750,	0.000,	46.000	# 12 61
6	W5E2	35.750,	0.000,	46.000	W7E1	35.750,	0.000,	39.000	# 12 5
7	W6E2	35.750,	0.000,	39.000	W8E1	36.250,	0.000,	39.000	# 12 1
8	W7E2	36.250,	0.000,	39.000	W9E1	36.250,	0.000,	46.000	# 12 5
9	W8E2	36.250,	0.000,	46.000		64.500,	0.000,	46.000	# 12 33

SOURCES						
Source	Wire Seg.	Wire #/Pct Actual	From End 1 (Specified)	Ampl.(V, A)	Phase(Deg.)	Type
1	31	5 / 50.00	(5 / 50.00)	1.000	0.000	I

No loads specified

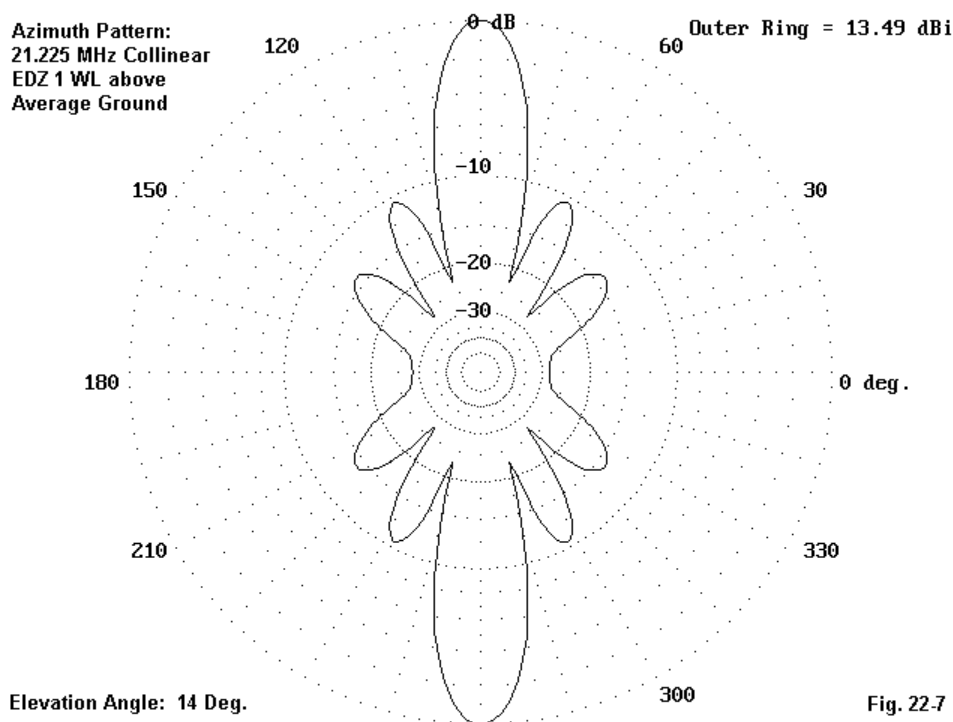
No transmission lines specified

Ground type is Real, high-accuracy analysis

Conductivity = .005 S/m Diel. Const. = 13

Medium	MEDIA		Ht(ft)	R Coord(ft)
	Conductivity(S/m)	Dielectric Const.		
1	5.000E-03	13.00	0 (def)	0 (def)

The potential performance of this antenna at an elevation angle of 14 degrees is a gain of 13.5 dBi with a beamwidth between -3 dB point of about 16 degrees. The feedpoint impedance is about $250 - j735$ Ohms. **Fig. 22-7** shows the azimuth pattern for the antenna model.



As important as this pattern is the current table. Below is a portion of that table showing the segments just before and after the phasing lines, and the lines themselves. Only one side of the antenna is shown, since the other side has identical current magnitude and phase values.

Collinear EDZs**Frequency = 21.225 MHz.**

 CURRENT DATA

Wire No. 1:

Segment	Conn	Magnitude (A.)	Phase (Deg.)
31		.53941	-7.26
32		.66831	-3.83
33	W2E1	.79846	-1.52

Wire No. 2:

Segment	Conn	Magnitude (A.)	Phase (Deg.)
1	W1E2	.97776	0.57
2		1.1851	2.37
3		1.3554	3.70
4		1.4793	4.79
5	W3E1	1.5519	5.81

Wire No. 3:

Segment	Conn	Magnitude (A.)	Phase (Deg.)
1	W2E2	1.5713	6.55

Wire No. 4:

Segment	Conn	Magnitude (A.)	Phase (Deg.)
1	W3E2	1.568	7.25
2		1.5154	8.19
3		1.4087	9.19
4		1.2527	10.36
5	W5E1	1.0567	11.91

Wire No. 5:

Segment	Conn	Magnitude (A.)	Phase (Deg.)
1	W4E2	.8601	13.91
2		.68562	16.47
3		.50825	20.48

In the table, wires 2, 3, and 4 compose the phasing line. We might be drawn to the fact that the current magnitudes on the facing wires (2 and 4) are very close to equal. Hence, we might erroneously conclude that a TL entry might well have substituted for the physically modeled phasing line. However, note that the relative current phase between the lines is neither in nor out of phase with each other. In fact, substituting TL entries for these physically modeled lines would have yielded an excessive gain figure and a quite different feedpoint impedance.

Phasing lines between collinear sections of an antenna are almost always best modeled physically.

3. A Linear-Loaded Dipole

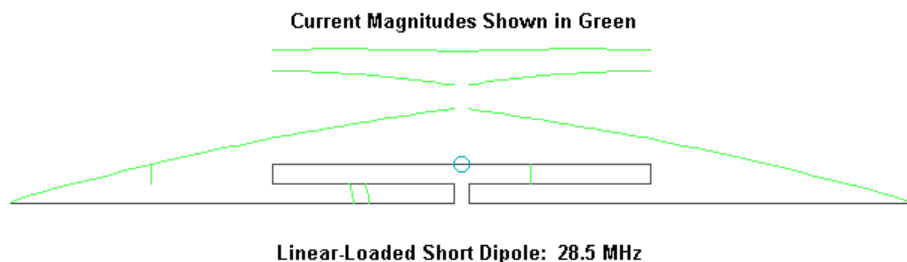


Fig. 22-8

In **Fig. 22-8**, we find a linear loaded #12 copper wire short dipole for 28.5 MHz. The total length of the dipole is under 135", about 65% of a full size dipole.

Loading is accomplished with linear loads, which resemble shorted transmission line stubs. In fact, when linear load wires are arranged symmetrically relative to the main dipole wire (for example, in a triangle), transmission line calculations for loading work quite well in figuring the required length for the load lines each side of the center feedpoint. If we omit the linear load, we would find a high value of capacitive reactance at the source point. An equivalent inductive reactance would compensate for the capacitive reactance. We can calculate shorted stubs, each having half the required full value, and they would come very close to lengths required when building an actual linear loaded dipole.

However, the arrangement of lines in this model is linear, with each line a different distance from the main element. The model description tells the story succinctly.

linear-loaded short dipole

Frequency = 28.5 MHz.

Wire Loss: Copper - Resistivity = 1.74E-08 ohm-m, Rel. Perm. = 1

WIRES									
Wire Conn.-	End 1 (x,y,z : in)	Conn.-	End 2 (x,y,z : in)	Dia(in)	Segs				
1	-67.300, 0.000, 0.000	W2E1	-1.200, 0.000, 0.000	# 12	34				
2	W1E2 -1.200, 0.000, 0.000	W3E1	-1.200, 0.000, 3.000	# 12	2				
3	W2E2 -1.200, 0.000, 3.000	W4E1	-28.200, 0.000, 3.000	# 12	13				
4	W3E2 -28.200, 0.000, 3.000	W5E1	-28.200, 0.000, 6.000	# 12	2				
5	W4E2 -28.200, 0.000, 6.000	W6E1	28.200, 0.000, 6.000	# 12	27				
6	W5E2 28.200, 0.000, 6.000	W7E1	28.200, 0.000, 3.000	# 12	2				
7	W6E2 28.200, 0.000, 3.000	W8E1	1.200, 0.000, 3.000	# 12	13				
8	W7E2 1.200, 0.000, 3.000	W9E1	1.200, 0.000, 0.000	# 12	2				
9	W8E2 1.200, 0.000, 0.000		67.300, 0.000, 0.000	# 12	34				

SOURCES						
Source	Wire Seg.	Wire #/Pct Actual	From End 1 (Specified)	Ampl.(V, A)	Phase(Deg.)	Type
1	14	5 / 50.00	(5 / 50.00)	1.000	0.000	I

No loads specified

No transmission lines specified

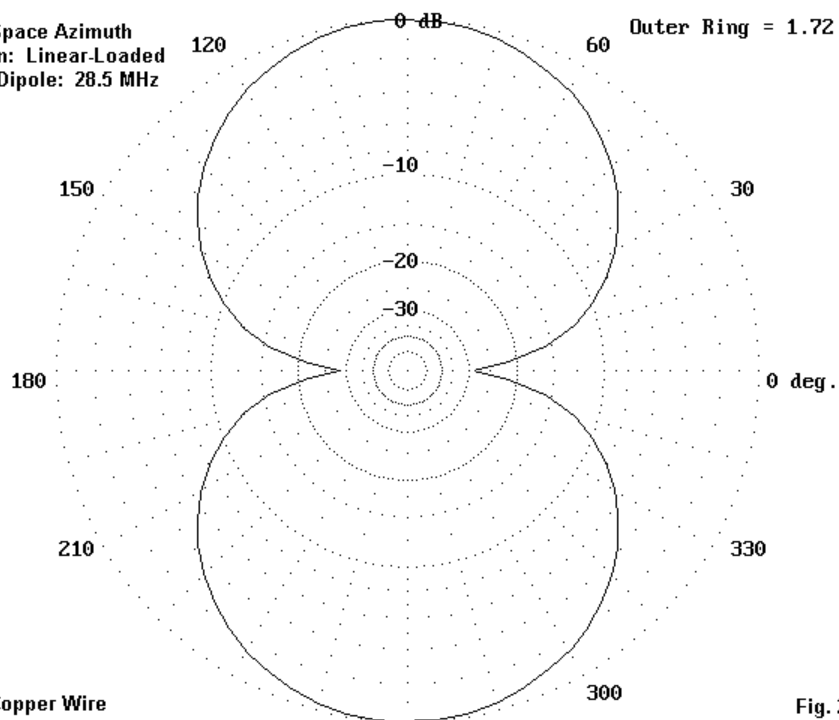
Ground type is Free Space

The linear load lines are spaced 3" apart and 3" from the main element. It makes no difference in free space that the load lines are above the element rather than below it. The free-space gain for the dipole is 1.7 dBi with a feedpoint impedance of 21 - j0 Ohms. **Fig. 22-9** shows the expected dipole pattern.

More significant for model construction is the current table, a portion of which appears below.

Free-Space Azimuth
Pattern: Linear-Loaded
Short Dipole: 28.5 MHz

Outer Ring = 1.72 dBi



#12 Copper Wire

Fig. 22.9

linear-loaded short dipole

Frequency = 28.5 MHz.

———— CURRENT DATA ————

Wire No. 1:

Segment	Conn	Magnitude (A.)	Phase (Deg.)
21		.57137	-1.50
22		.5929	-1.46
23		.61385	-1.42
24		.63428	-1.38
25		.6542	-1.34

26		.67366	-1.30
27		.69266	-1.25
28		.71122	-1.21
29		.72946	-1.16
30		.74741	-1.11
31		.76522	-1.06
32		.7831	-1.01
33		.8014	-0.95
34	W2E1	.82098	-0.88

Wire No. 2:

Segment	Conn	Magnitude (A.)	Phase (Deg.)
1	W1E2	.84048	-0.81
2	W3E1	.85779	-0.75

Wire No. 3:

Segment	Conn	Magnitude (A.)	Phase (Deg.)
1	W2E2	.87723	-0.68
2		.89596	-0.62
3		.91193	-0.58
4		.926	-0.54
5		.93844	-0.51
6		.94961	-0.49
7		.9595	-0.46
8		.96827	-0.45
9		.97598	-0.43
10		.98255	-0.41
11		.98806	-0.40
12		.99266	-0.39
13	W4E1	.99633	-0.38

Wire No. 4:

Segment	Conn	Magnitude (A.)	Phase (Deg.)
1	W3E2	.99895	-0.37
2	W5E1	1.0021	-0.37

Wire No. 5:

Segment	Conn	Magnitude (A.)	Phase (Deg.)
1	W4E2	1.0066	-0.36
2		1.0109	-0.35
3		1.0141	-0.34
4		1.0162	-0.33
5		1.0174	-0.31
6		1.0176	-0.29
7		1.017	-0.27
8		1.0155	-0.25

9	1.0134	-0.22
10	1.0105	-0.19
11	1.0073	-0.15
12	1.004	-0.11
13	1.0012	-0.06
14	.99999	0.00

Wires 1, 3, and 5 are the parallel sections of wire forming one side of the dipole. These are the wires on which to compare the current magnitudes and phases relative to each other. The current inequalities tell us that the shorted stub is not acting as a pure transmission line. Radiation is not solely a function of the main wire, but of the net current magnitude and phase of the three closely spaced wires composing the linear load and the main element. The current magnitude line in **Fig. 22-8** gives a more graphical view of the rate of change of current along each of the relevant wires.

These examples are samples of when physically modeling transmission lines will provide a more accurate model than using the TL facility. Wherever the current at the terminals of the line is not equal, the line will contribute to the far field radiation pattern—either adding to or subtracting from the overall pattern. Under those conditions, physically modeling the transmission line will always be more accurate than using the TL facility in NEC.

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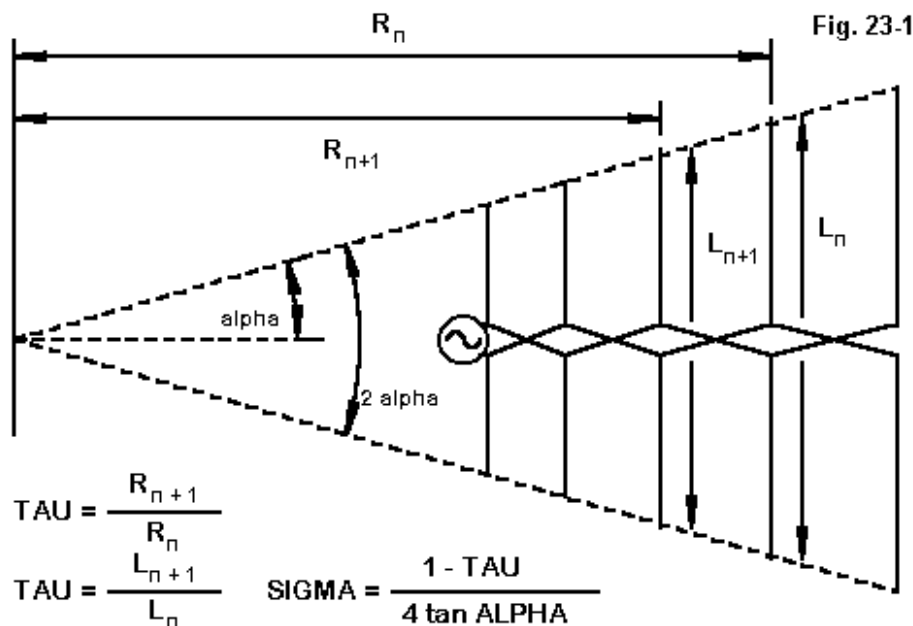
Models included: 22-1 through 22-3.

23. Modeling LPDAs

Log-Periodic Dipole Arrays are becoming more popular for amateur use and have become a mainstay of governmental and commercial broadcast and communications systems. When well engineered, they eliminate the necessity of re-matching a main coaxial cable feedline to the antenna over a wide frequency range. The frequency range may vary with the application. There are log-cell Yagis for monoband amateur use, LPDA arrays with a 1.0 to 1.5 octave range for the upper HF ham bands, and also arrays with over a 3 to 4 octave range for other services.

The design of LPDAs is covered well in the literature—at least with respect to long-standing design principles. There are some proprietary algorithms that vary from the standard, but most amateur radio LPDAs are still designed using the basic relationships shown in **Fig. 23-1**.

The antenna array consists of a series of dipoles whose self-resonant frequencies and lengths are set up in a periodic fashion and whose spacing is equally periodic. The elements are interconnected with a transmission line that reverses connections at each new dipole element, with the feedpoint normally at the junction with the shortest dipole of the array. We shall not dwell on the design parameters, since they appear in Johnson and Jasik as well as in the *ARRL Antenna Book* (Chapter 10). Our concern will be limited to modeling LPDAs.



Some LPDA Design Basics

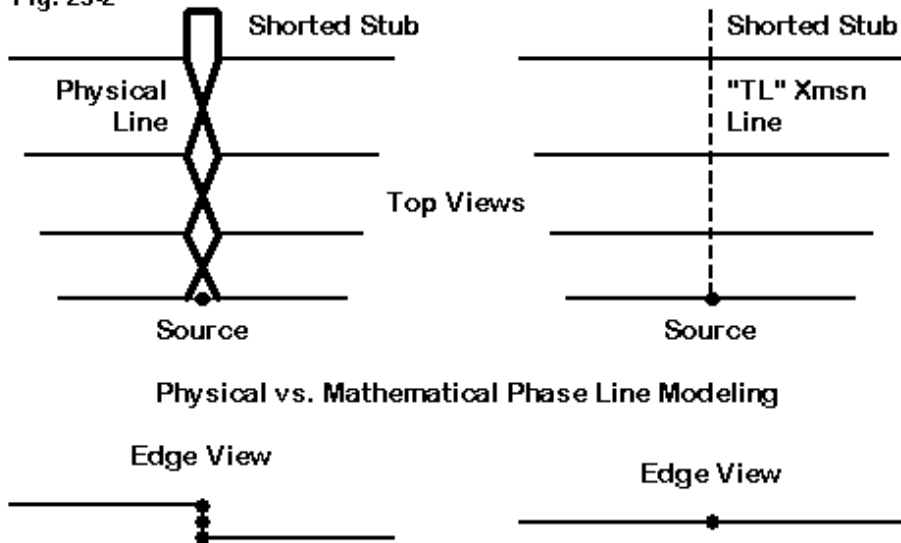
(Adapted from Johnson, 3rd. Ed. p. 14-36)

Essentially, we have two choices. We can attempt to model physically all aspects of the antenna array, as shown on the left in **Fig. 23—2**. Unfortunately, for NEC, this strategy quickly meets with some of the geometry limitations within the program. These limitations are functions of the calculation core and not of any particular implementation of NEC.

Since the elements are rarely of the same diameter as the transmission line wires, we encounter NEC's difficulties with angular junctions of dissimilar diameter

wires. Results will rarely be accurate. Of course, every physical implementation of an LPDA has to deal with the crossing transmission line wires. The simplest scheme that works well appears in the left edge view. Let the inter-element transmission line (or phase line for short) be set up vertically. Then the left and right sides of each dipole can intersect alternate upper and lower wires of the phase line. The misalignment of up to a few inches of the two sides of the dipole will create no significant errors in the resulting antenna pattern or Performance figures. However, the angular junction problem can only be overcome for a few designs using a constant diameter for all portions of the antenna.

Fig. 23-2



Physical vs. Mathematical Phase Line Modeling

MININEC (3.13) does not have the same angular junction limitation that troubles NEC, but MININEC does have limitations of its own. Sharp angular junctions require the use of high levels of segmentation or the use of length tapering to ensure

that each junction is met with very short segment lengths. The short segment lengths minimize errors created by MININEC's tendency to "cut off" corners. The end result of overcoming the inherent MININEC limitation is a model that will overrun the maximum segment limitation of most DOS versions of the program. Those programs that have extended the number of available segments will run very slowly with very high numbers of wires and segments in the model. One version of MININEC 3.13 (Antenna Model) has proven that its considerable collection of correctives to the core weaknesses does allow it to effectively model LPDAs in the manner indicated.

The most promising way to model an LPDA is to follow the lead of the right hand sketch in **Fig. 23-2**, which applies only to NEC (-2 or -4) models. Set up each dipole in its proper position. Use an odd number of segments on each dipole wire. (This note applies to the center wire of elements that have a tapered element diameter schedule, which in turn only applies to programs having corrections for NEC's difficulties with tapered diameter elements.)

From one dipole to the next, run a TL transmission line of the desired characteristic impedance. Reverse the connection of each transmission line installed. Place the source on the center of the shortest dipole. Many LPDA designs employ a shorted transmission line stub at the center of the longest element, and this can also be put in place in the model using the TL facility.

The following model description, adapted from a test file in EZNEC, illustrates the model construction principles.

17-10m Log Per - ARRL Ant Book Frequency = 28 MHz.

Wire Loss: Aluminum - Resistivity = 4E-08 ohm-m, Rel. Perm. = 1

———— WIRES ————

Wire Conn.—	End 1 (x,y,z : in)	Conn.—	End 2 (x,y,z : in)	Dia(in)	Segs
1	0.000,-163.46, 0.000		0.000,163.460, 0.000	1.25E+00	41
2	39.230,-130.76, 0.000		39.230,130.760, 0.000	1.00E+00	33
3	70.620,-104.62, 0.000		70.620,104.620, 0.000	7.50E-01	25
4	95.720,-83.690, 0.000		95.720, 83.690, 0.000	6.25E-01	21
5	115.810,-66.950, 0.000		115.810, 66.950, 0.000	5.00E-01	17

SOURCES

Source	Wire Seg.	Wire #/Pct Actual	From End 1 (Specified)	Ampl.(V, A)	Phase(Deg.)	Type
1	9	5 / 50.00	(5 / 50.00)	1.000	0.000	I

TRANSMISSION LINES

Line	Wire #/% Actual	From End 1 (Specified)	Wire #/% Actual	From End 1 (Specified)	Length	Z0 Ohms	Vel Fact	Rev/ Norm
1	1/50.0	(1/50.0)	2/50.0	(2/50.0)	Actual dist	490.1	1.00	R
2	2/50.0	(2/50.0)	3/50.0	(3/50.0)	Actual dist	490.1	1.00	R
3	3/50.0	(3/50.0)	4/50.0	(4/50.0)	Actual dist	490.1	1.00	R
4	4/50.0	(4/50.0)	5/50.0	(5/50.0)	Actual dist	490.1	1.00	R
5	1/50.0	(1/50.0)	Short ckt	(Short ck)	6.000 in	490.1	1.00	

Ground type is Free Space

Although this is a good illustration of the modeling technique, the model does not show exceptional performance. (Model 23-1)

Use of the TL facility avoids most of the difficulties of physically modeling the LPDA. However, the use of TL phasing lines has some limitations of its own. Theoretically, the phase line does not enter into the antenna's radiation on any frequency. However, physical placement of the stub can sometimes alter the antenna's performance on certain frequencies. As mathematical entities created by a network placed at a large distance from the antenna model proper, the TL phase line cannot show the potential effects of physical placement.

In addition, there are construction variables connected with the assembly of real LPDAs, and some of these cannot be captured by the suggested NEC modeling technique. For example, K4EWG presented a 20-10 meter LPDA in Vol. 3 of the *ARRL Antenna Compendium*. He used a number of interesting construction techniques, sketched in **Fig. 23-3**.

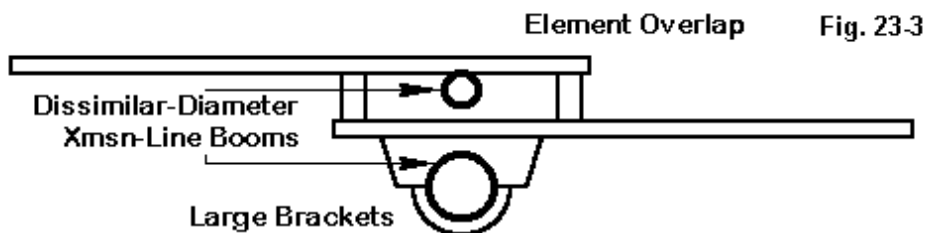


Fig. 23-3

Some Potential Modeling Difficulties With LPDAs

His phase line consisted of two tubes having different diameters. Ordinarily, the impedance of the line created by the two should be calculable from the diameter of the smaller tube, but the physical effects of the arrangement would be missing from the suggested modeling technique. In addition, he used muffler clamps to connect the lower element set to the larger boom. Most significant are the overlapping element ends—8" each side of centerline. Overlapping the ends of dipole elements does affect antenna impedance in ways that the simplified model cannot fully capture.

For the most part, none of these construction techniques—and others that might be comparable—affects the antenna pattern with respect to gain, shape, or front-to-back ratio. When trying to model an LPDA using construction methods that have physical significance, it is important to establish before finalizing a model that these physical factors do not have distorting affects relative to the proposed final model. You can do this by modeling individual elements with all physical aspect taken into account and comparing the results with simplified single elements.

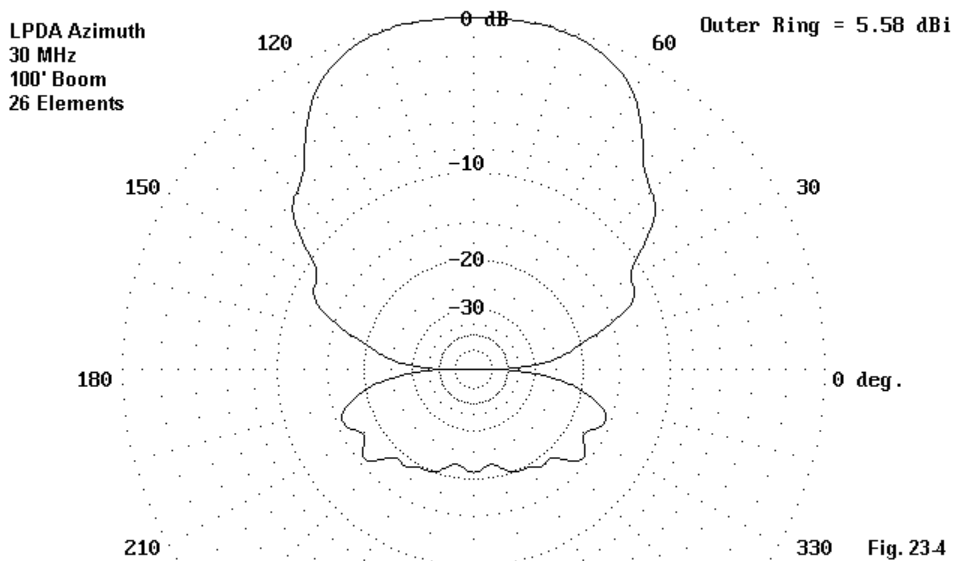
Where constructed elements of the sort illustrated have their main effect is on the performance of the phase line. Its net effective impedance may not match the design impedance that we calculate from standard equations and simple round wires. The most straightforward way to deal with phase line variables is to survey the antenna performances at each check point using a variety of phase-line charac-

teristic impedances from about 75-80 Ohms at the lower end of the spectrum to about 200-250 Ohms at the upper end.

The resulting chart of performance figures should include at least gain (initially done as free-space gain in dBi), front-to-back ratio, and source impedance. You may encounter phase-line characteristic impedance values that are more favorable to a smooth VSWR curve than other characteristic impedance values. You may also encounter interesting anomalies. For example, at one frequency within the antenna design range, gain may increase with the characteristic impedance, while at another frequency, the gain may increase inversely with increases in the characteristic impedance.

In short, do not expect a smooth set of performance figures that vary only slightly with variations in the phase-line characteristic impedance or frequency. Indeed, do not expect all resulting radiation patterns to closely resemble those obtained from Yagis. There is an old over-simplification that suggests that at any frequency, relative to the dipole that is closest to resonance, the elements affecting the overall antenna performance at that frequency are those immediately adjacent to the near-resonant dipole. Actually, at least two elements to the rear of the near-resonant dipole and almost all elements forward carry significant current and thus have major consequences for the resulting antenna radiation pattern. Since all of the elements carry at least some current, antenna patterns can vary significantly from the smoothly rounded norm that we have come to expect from Yagis.

Fig. 23-4 shows one mild example, drawn from a model of a 3-30 MHz LPDA at 30 MHz. The array is 100' long and has 26 elements. Clearly apparent are the incipient lobes created by harmonic operation of elements distant from resonance at 30 MHz. Two extra notes are needed here. First, this is not the only type of pattern one might encounter. Diamond-shaped forward lobes are common. So, too, are split forward lobes with peaks separated from the centerline by up to 20 degrees or so. Moreover, more than one of these effects may occur at once. Fortunately, most of the pattern aberrations do not occur on common single octave (for example, 20-10 meter) LPDAs of standard design.



Second, the characteristic impedance of the phase line has a bearing on the occurrence of these less-controlled azimuth pattern shapes. In general, the higher the impedance of the phase line, the smoother the outline of the azimuth pattern. However, the rear lobes may show some incipient lobes in the form of a slightly squared shape.

The gain reported by NEC is very likely to be less than the gain reported by designers who use some of the early gain equations associated with basic LPDA design instructions. The gain of an LPDA is a function of the value of τ , one of the constants selected by the designer. Only values of τ in the vicinity of 0.95 are capable of very high gain values, and only when these values are associated with very long boom lengths determined by an optimal value of the spacing constant, σ .

To see how this works, let us take a sample LPDA using 7 elements for 14 to 30 MHz and having a Tau of just over 0.85. We shall place these elements on three different booms of 20, 30, and 40 feet, respectively. These booms correspond to values of Sigma of 0.07, 0.10, and 0.14. The last value is closing in on the optimum value one might calculate for Sigma (about 0.16). Although there are many ways of calculating potential gain in advance, Roger Cox, WB0DGF, reports in his program, LPCAD, that the gain values associated with the Tau of 0.85 should range from about 5.5 dBi to about 7.5 dBi. (Values are interpolated from his table.) We should expect the lower values to be associated with the shortest version of the antenna and the highest values to be connected to the longest version.

Here is a table of NEC-4 modeled results for the antenna. Gain is free-space gain in dBi, while front-to-back is the 180-degree ratio in dB. The shortest antenna uses a 75-Ohm phase line, while the longer models use 100-Ohm phase lines. All use a shorted stub from the longest element, and the stub is set to optimize performance and the feedpoint impedance. (Model 23-2)

Gain/Front-Back Ratios of 7-Element LPDAs for 14-30 MHz

Boom Length	20'	30'	40'
Frequency			
14	5.1 / 8.2	5.7 / 9.7	6.2 / 10.4
18.12	5.8 / 9.8	7.3 / 13.1	7.2 / 22.6
21	6.2 / 14.0	6.7 / 21.5	7.3 / 22.4
24.95	6.3 / 16.8	6.6 / 17.2	7.5 / 27.9
28	6.2 / 16.9	6.2 / 17.2	6.2 / 17.7

This table demonstrates several properties you are likely to see in LPDA designs of modest proportions. First, the longer the boom (up to the limits of optimal Sigma), the higher the gain for the same number of elements. Second, for any given design, some gain or front-to-back figures may well occur that are higher or lower than we would expect from smoothing the curve of results. For example, the 30' version gain at 18.12 MHz is higher than expected from the other values, and the feedpoint impedance turns out to be lower than expected at that frequency. Here is a case that might be “normalized” by adjustment of the length of the shorted stub—if that operation does not disturb performance on another frequency.

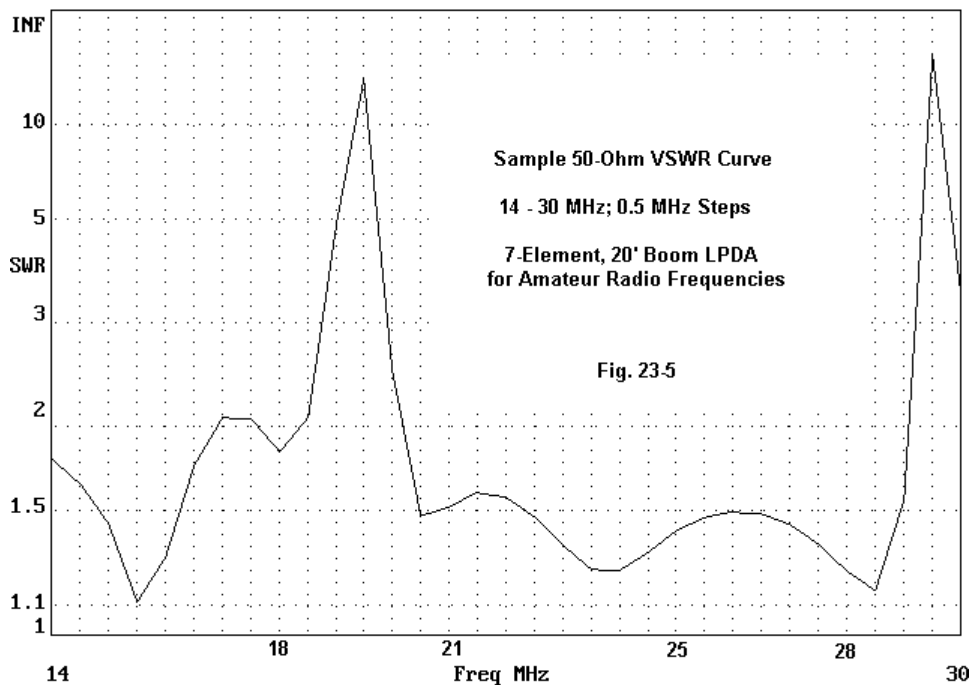
Third, performance at the upper and lower limits of the design range tends to fall off unless the LPDA is designed to have wider frequency limits surrounding the frequencies actually used. There appears to be a balance one might draw between too high an element density and too low a density. Higher densities tend to increase low frequency performance, but can actually detract from higher frequency performance when carried too far. Likewise, too low an element density can yield lower gain and front-to-back performance at the frequency limits of the design.

The present design does not correspond exactly with any published design, although it is related to several. It uses constant diameter elements—all 1" in diameter. Some designs may benefit from staggering the element diameters by increasing the values by the inverse of Tau, counting from the shortest element to the longest.

The table of values represents no judgment about the three variants of one design other than this: the values are typical of NEC reports and largely coincident with expectations raised by WB0DGF's suggested table. However, they are significantly lower than values we often see reported for LPDA designs in amateur literature. Therefore, each modeler who anticipates more than casual modeling of LPDA designs should model many designs and develop appropriate expectations for himself or herself.

We are often led to believe that LPDA antennas exhibit relatively smooth VSWR curves across the range of their design. If you model LPDAs, develop other expectations, as illustrated by **Fig. 23-5**.

The SWR sweep in the graph is a function of NEC-4 calculations. Note that the SWR is well below 2:1 relative to a 50-Ohm standard for each of the amateur bands. While it modestly climbs above 2:1 at 17 and 17.5 MHz, it shows very major peaks in the 18.5 to 20 MHz range and again in the 29 to 29.5 MHz range. One of the functions of the shorted stub is to control the placement of the peaks so that the SWR is acceptable at the anticipated operating frequencies. Careful selection of the dipole lengths—as determined by the selection of the value for Tau—also contributes to lower SWR values at the desired frequencies.



As we lengthen the boom for the same set of elements, the SWR peaks decrease and the SWR becomes more manageable without critical adjustment to the value of Tau and the shorted stub. However, the curve will never be absolutely smooth. Moreover, the SWR value may never settle at a certain value, even when referenced to the most optimal impedance standard. What can happen with some designs is that the resistive component of the feedpoint impedance deviates most from the standard when the reactance is the lowest. Conversely, when the resistance is closest to the standard, the reactance reaches peak capacitive or inductive values. However, in some designs, the mean resistive component may shift as you move to different parts of the array's frequency range.

I have sampled some of the phenomena you may encounter when modeling LPDAs in order to remove some of the surprises that might shock a relatively inexperienced modeler out of further work with this kind of antenna design. Not all of the phenomena listed occur with every design—indeed, some designs will be very tame and well controlled. However, the more phenomena with which you are acquainted, the less likely it will be that you will be discouraged from making optimizing adjustments to a model.

Nonetheless, even when you have achieved the best possible NEC model of an LPDA, expect to make field adjustments to compensate for construction factors that cannot be included in the model. NEC can be highly accurate in predicting the performance of a design, but it cannot adjust the shorted stub or decide for you whether or not to use a broadband-matching device between the antenna feedpoint and the coax.

* * * * *

Models included: 23-1 and 23-2.



24. The Power and the Source

A continuing source of confusion among newer modelers is the data that NEC produces concerning power and the source conditions. Few modelers realize that they have both limitations and opportunities in these parts of modeling. So let's look a little closer at both arenas.

Power

NEC provides a "power budget" report as part of its standard report. Although this report is accessible to users, few ever look at it in detail. A typical power budget appears in **Fig. 24-1**.

```
NEC-Win Plus+ OUTPUT DATA
-----
C:\NEC-Plus\examples\10-1.NOI

                                Fig. 24-1
FREQUENCY
-----
FREQUENCY= 2.8500E+01 MHZ
WAVELENGTH= 1.0519E+01 METERS

POWER BUDGET
-----

INPUT POWER    = 1.2494E-03 WATTS
RADIATED POWER= 1.2494E-03 WATTS
STRUCTURE LOSS= 0.0000E+00 WATTS
NETWORK LOSS   = 0.0000E+00 WATTS
EFFICIENCY     = 100.00 PERCENT
```

Without supplementary data, the numbers may be of little use. The model involved in this report is a folded dipole for 28.5 MHz. The actual power budget consists of 5 entries.

The input power is that supplied to the antenna at the source (or sources). It is a direct function of two factors: the source specified by the modeler and the source impedance calculated by the program. Let's assume that we opted for a 1 volt, 0-degree source—the one most commonly used with NEC. NEC itself uses peak volts and reports output voltages and current in peak values. Programs like NEC-Win Plus use these values directly. Programs like EZNEC accept only r.m.s. values as inputs and report r.m.s. values as outputs. Adjustments needed to work with the core are made internally within the program.

NEC calculated for the segment on which the source happens to be placed an impedance of $398.8 + j23.5$ Ohms, or 399.5 Ohms at a phase angle of 3.375 degrees. NEC calculated a peak current of 0.0025 A at a phase angle of 3.375 degrees. This is a simple application of Ohms law, dividing the specified voltage by the impedance of the source segment.

Now let the power, P , equal the traditional I^2R . If we take only the current, which is 0.0025031 A (to be overly precise), square it, and multiply by the real or resistive part of the source impedance (398.815 Ohms), we get a power report of 0.0024988 watts, which is double the power reported in **Fig. 24-1**. However, **Fig. 24-1** reports the NEC core peak value for current. The r.m.s. value of current is 0.707 times the peak value. In other terms, the power (always an r.m.s. value) is half the value calculated from the peak current value. The result is 0.0012494 watts, just the value that **Fig. 24-1** reports in engineering notation.

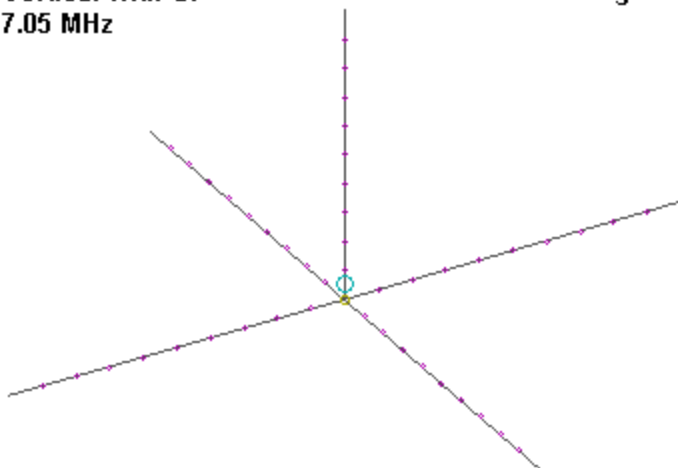
Notice that the radiated power has the same value. The structure losses, here reported as zero, are a function of the material we specify for the wires of the model. This model evidently specified perfect conductors and no load with resistive components. There are also no network losses, meaning that the model has no networks with resistive components.

When the Radiated Power equals the Input Power, the efficiency is 100%. NEC always calculates efficiency as a ratio of radiated power to input power (times 100, of course). The efficiency will include only those losses internal to the antenna

model structure, including material losses, load losses, and network losses. Not included are any external factors that affect the antenna's environment. Let's illustrate using the simple ground-plane vertical for 7.05 MHz in **Fig. 24-2**. We shall initially use no loads and specify perfect conductors for all 5 wires. Also, we shall use the Sommerfeld-Norton ground system and average soil.

Vertical with GP
7.05 MHz

Fig. 24-2



Following is the basic description for this small model

1/4 wavelength w/GP: 7.05 MHz

Frequency = 7.05 MHz.

Wire Loss: Zero

WIRES

Wire Conn.—	End 1 (x,y,z : m)	Conn.—	End 2 (x,y,z : m)	Dia(mm)	Segs
1	0.000, 0.000, 10.135	W2E1	0.000, 0.000, 0.000	5.00E+01	10
2	W3E1 0.000, 0.000, 0.000		11.796, 0.000, 0.000	6.35E+00	10
3	W4E1 0.000, 0.000, 0.000		0.000, 11.796, 0.000	6.35E+00	10
4	W5E1 0.000, 0.000, 0.000		-11.796, 0.000, 0.000	6.35E+00	10
5	W1E2 0.000, 0.000, 0.000		0.000, -11.796, 0.000	6.35E+00	10

SOURCES

Source	Wire Seg.	Wire #/Pct Actual	From End 1 (Specified)	Ampl.(V, A)	Phase(Deg.)	Type
1	10	1 / 95.00	(1 /100.00)	56.984	0.000	V

No loads specified

No transmission lines specified

Ground type is Real, high-accuracy analysis

Conductivity = .005 S/m Diel. Const. = 13

MEDIA

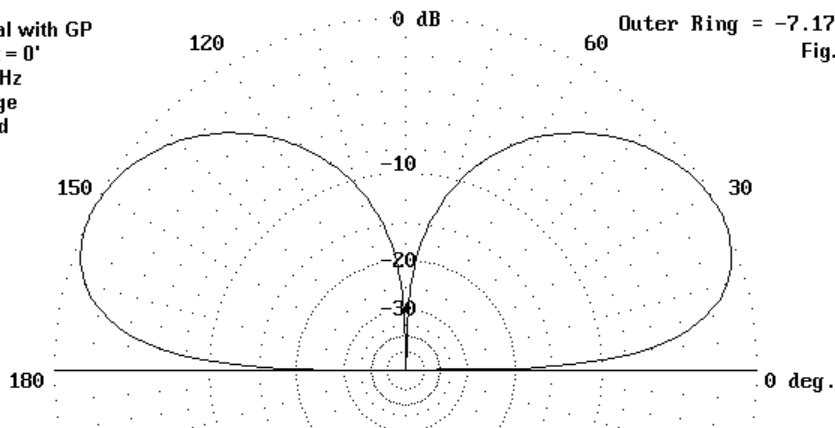
Medium	Conductivity(S/m)	Dielectric Const.	Ht(m)	R Coord(m)
1	5.000E-03	13.00	0 (def)	0 (def)

Using NEC-4, let's place the ground plane directly on the ground, that is, at $Z=0$. (We cannot do this with NEC-2, which demands that the base of the antenna be elevated slightly above ground.) The resulting elevation pattern for the antenna appears in **Fig. 24-3**.

Now let's place the antenna at a height of 42.5 m (1 wavelength) above ground, changing nothing in the model except the values for the Z-coordinates. The result is an elevation pattern like that in **Fig. 24-4**. (Model 24-1)

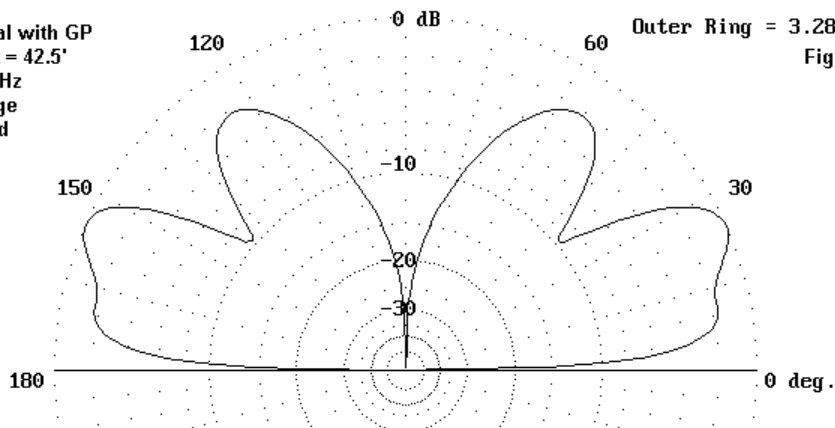
Vertical with GP
Height = 0'
7.05 MHz
Average
Ground

Outer Ring = -7.17 dBi
Fig. 24.3



Vertical with GP
Height = 42.5'
7.05 MHz
Average
Ground

Outer Ring = 3.28 dBi
Fig. 24.4



Despite the differences between the two patterns, both antennas show an efficiency of 100%, that is, a 1:1 ratio between the input power and the radiated power.

In fact, we can look at the power budgets for three conditions: 1. the perfect (lossless) wire model with no loads, 2. the same model with aluminum wires, and 3. a shortened aluminum wire version with a loading coil of 200 Ohms reactance and 1 Ohm resistance ($Q=200$). The power budgets look like this:

Height M	Input Power W	Radiated Power W	Wire Loss W	Efficiency %
No wire or load losses				
0	9.9992E+00	9.9992E+00	0.0000E+00	100
42.5	1.5162E+02	1.5162E+02	0.0000E+00	100
Aluminum wire, no load				
0	9.9979E+00	9.9952E+00	2.7137E-03	99.97
42.5	1.5079E+02	1.4993E+02	8.5922E-01	99.43
Aluminum wire, load (1+j200 Ohms)				
0	1.2004E+01	1.1927E+01	7.7168E-02	99.36
42.5	2.8589E+02	2.5892E+02	2.6971E+01	90.57

No entry for network loss will appear unless we introduce a network to the antenna system. However, we read in the chart differences of efficiency. Some are easy to interpret. For each height, we see a reduced efficiency as we add losses to the system—first the wire material, then the loading coil. However, we also see a difference in efficiency between ground and elevated mounting of the antenna. The efficiency goes down as the height goes up, although the maximum gain goes up with height.

The efficiency differences for each case result from the fact that we are dealing with I^2R losses. The ground-mounted model shows an impedance over ten times that of the version 1 wavelength up. There is also a similar difference in current, with the elevated model showing the higher value. When each of the two currents is squared, a fixed resistance shows a greater loss in dissipated power. Nonetheless, the difference made by material losses is small in terms of far field radiation (<0.02 dB). The shorter, loaded version of the antenna shows about 2.2 dB lower strength in the ground mounted version and about 0.4 dB lower strength in the elevated model.

The maximum gain difference between the two sets of models—well over 10 dB—is, with respect to NEC, a function of the interaction of the antenna's radiation

with its environment. NEC does not take the environmental factors into account when calculating efficiency. Analyses of the relative effectiveness of two antenna models must be done externally, normally beginning with the relative gain values, pattern shape, and similar factors.

What NEC does do when it has a constant 1 volt, 0-degree phase angle source is to account for the effects of the environment in the impedance of the source segment. The perfect-wire elevated vertical shows an input impedance of $21.4 - j0.6$ Ohms, while the version planted on the ground has a source impedance of $191.1 - j159.8$ Ohms. The result is different values of current on the source segment: 0.0467 A vs. 0.0040 A. (Since the values for the current vertical model are taken from EZNEC, they are r.m.s. values.) The result is a large difference in input power: 0.0467 W for the elevated vertical vs. 0.0031 W for the ground-mounted version. In each case, the antennas radiate all of the input power and have efficiencies of 100%.

Note the fact that the power for the elevated vertical has a numerical value that is the same as the numerical value of the current. Since the impedance is well over 99% resistive, the value of I^2R and I^2Z are virtually identical. However, for the ground mounted vertical, the impedance is highly reactive. Only the resistive or real portion of the impedance enters into the power calculation, and 0.0040^2 times 191.1 yields the value for the input (and the radiated) power.

The Convenience of 1 Amp

When using a voltage source in NEC, the current values that emerge along each wire tend to have inconvenient values. It is difficult to sense the slope of a curve, even though graphical representations of the relative current are helpful visual aids to this process. What can we do to obtain more convenient numerical values?

If all segment currents are referenced to a source current value of 1.0, the user can much more readily see the relationships among values at various segments along the wires making up the antenna. One easy way to do this is to specify a current source and give it a value of 1.0 at phase angle zero. However, a current source in NEC is created by a remote network normally invisible to the program user. Under certain conditions, a current source may not be as reliable as a voltage

source. One case in which this is true is where the currents on wire segments adjacent to the source segment are not equal to each other. This phenomenon can occur where the feedpoint occurs at the corner of an element and the two sides are not of equal length or form a non-symmetrical structure relative to the source.

There is way to provide a source current of 1.0 A at a phase angle of 0 degrees and to stay within the voltage source mode. This technique applies to single source antenna models. First, assign the model a normal voltage source value of 1.0 V at 0-degrees phase angle. Run the model and note the current magnitude and phase. Let us say that we get a report of 0.024 at -16.12 degrees.

Now take the inverse of the current magnitude, in this case 41.67 and enter it as a new voltage source value. For the source phase angle, use the negative of the current phase angle. In this case, we began with a negative phase angle, so we would enter a positive 16.12 degrees. At the specified frequency for the model, the current will now read 1.0 A at 0-degrees phase angle.

Fig. 24-5

A Simple 11-Segment Dipole



Fig. 24-5 can let us quickly demonstrate the utility of using 1 Amp as a source reference. Consider the current profile along the dipole segments. The following table will show only one half of the model, since the other half is a mirror image. You may decide which set of numbers provides the intuitively clearer portrait of the antenna's current levels.

Source Type	Segment Number and Current Report					
	1 (End)	2	3	4	5	6 (Center)
1 Volt	.00237	.00645	.00984	.01241	.01403	.01458
1 Amp	.0126	.4425	.6746	.8514	.9622	1.000

This technique is useful only at the design center frequency for the antenna model. Since the impedance will change as we alter the frequency of interest, so

too will the required voltage source values to achieve a 1 A current. However, for single frequency investigations of current along antenna elements, the technique is a quick and easy way to view all current magnitudes in relationship to a source value of 1 and current phase angles in relationship to a source value of 0 degrees.

Power Specification

For some applications, it is useful to provide an antenna source with a set power level. Some programs, such as EZNEC provide this option. The program simply takes the core output and recalculates relevant values in terms of the power its sets based on what NEC reports using a standard source input set of values.

For programs that do not have this provision, we can do the job ourselves. Let's work through an example, using our ground-mounted 7.05 MHz vertical. If we provide the model with a voltage source of 1.0 at 0-degrees phase angle, it returns the following source data.

Voltage: 1.0 V at 0.0 degrees
Current: 0.004014 A at 39.90 degrees
Impedance: 191.1 - j 159.8 Ohms
Power: 0.003079 Watts

This is all the data we need to set a power level. Since these are EZNEC reports, the values are r.m.s. Our goal is to find the voltage magnitude needed for some specific power level. Let's set 10 watts as the required level.

First, we need to find the impedance magnitude, which is

$$Z = \sqrt{R^2 + X^2}$$

where Z, R, and X have their usual meanings. For the example, the impedance is 249.11 Ohms.

The value of voltage to use at the source is simply

$$E_{r.m.s.} = \sqrt{\frac{PZ}{\cos\theta}}$$

where P is 10, the power we selected, and Z is the impedance magnitude we just calculated. Theta is the phase angle, which for the example is simply 39.90 degrees. Since the collection of values is highly interrelated, there are other ways to arrive at the same numbers, but this one will certainly work. For the example, the desired voltage to supply the source with 10 watts is 56.98 Vr.m.s.

If program needs demand a peak voltage value, we may simply multiply the r.m.s. value by 1.414 to arrive at 80.57 Vpk. Since the input voltage derivation uses values of power and impedance, the resulting calculation will always yield an r.m.s. voltage value which requires conversion for programs needing the peak value.

There are numerous applications for which the specification of a specific power level is either necessary or desirable. One of the most important is the analysis of ground wave values. The RP1 output of NEC yields the total ground wave radiation. The user specifies—in the most usual type of case—a distance from the antenna and a height above ground for the reading. (This assumes that the antenna is centered at a location where the X and Y axis values are both zero.) NEC will report calculated values of radiation in volts per meter (which might be reported as mV/m in the user output report).

The ground wave output is very useful in estimating the results of making changes in the antenna, changes that in each case might change the source impedance. With changes in source impedance and a simply voltage source, the input power to the antenna may be different for each modification. Recalculating the source voltage to give a uniform power input for each version of the antenna will make the ground wave output reports comparable. A similar utility to using a constant power input to the antenna applies when trying to compare different antennas at the same position.

Fig. 24-6

Double Rectangle



Fig. 6 illustrates a second instance where a constant power level can make instantly sensible current values that might otherwise seem initially confusing. The antenna is an open-double rectangle, fed either on the end vertical or on the center vertical. The array produces vertically polarized energy in the main, and is useful in the MF through lower HF region.

If we provide the end-wire source with 1 A, using a current source, then feeding the end element produce a current on the end vertical center-point of 1.0 A and at the midpoint of the center vertical of 1.84 A. If we move the source to the center-wire, the current readings are 0.54 on the end wire and 1.0 on the center wire. The ratios are the same, but the values differ. Of course, an ideal double rectangle with no losses and no ground effects would show vertical element current ratios of 1:2:1 from end to end. So too would any three equal length verticals spaced in a line at $1/2$ -wavelength intervals.

The current value differences result from the change of source impedance when we move the source position. The ratio of impedances ($28.4 - j11.0$ to $8.6 - j3.0$ Ohms) between end and center feeding is roughly the square of the ratio of currents. However, if we select a constant power—say, 10 watts—we get the following numeric current values. End feed: end wire = 0.59; center wire 1.09. Center feed: end wire 0.58; center wire 1.08. Had we used lossless wire, the numeric values would have been even closer. Nonetheless, the use of a constant power level relieves use from having to make further calculations to correlate current values on the elements of the open-double rectangle.

These samples only begin to demonstrate the advantages of being able to transform source voltage, current, impedance, and power numbers from their initial input and report values into other values of more instant utility. Becoming more familiar

with the power and source values used and reported by NEC can make the program (and its commercial implementations) more productive for even the casual user.

* * * * *

Models included: 24-1 and 24-2.



25. Bringing Up the Rear: Front-to-Back Ratios

When we work with models of directional arrays, such as Yagi-Uda beams, we take great interest in the front-to-back ratio reported by either NEC or MININEC software. However, I am not certain we always fully appreciate the data provided to us by our models. Let's look at **Fig. 25-1** as an illustration.

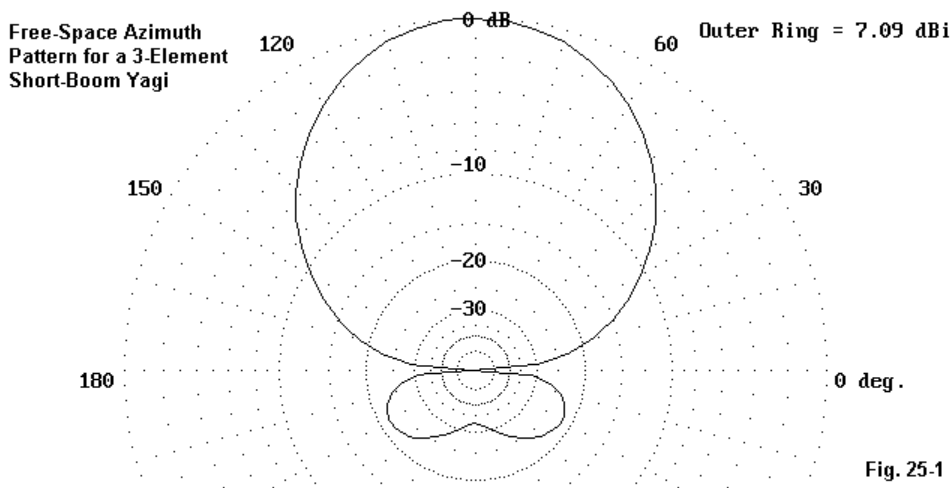


Fig. 25-1 is the free-space azimuth pattern of a short-boom 3-element Yagi. Both EZNEC and NEC-Win software report the gain at 28.4 MHz as 7.09 dBi. However, EZNEC reports a front-to-back ratio of 32.56 dB, while NEC-Win reports that the front-to-back ratio is 21.63 dB. (Model 25-1)

Is one of the software packages deeply in error? Not a bit. Each is simply reporting the front-to-back ratio according to a different but valid definition. Actually,

there are three working notions of front-to-back ratio generally current in antenna work. Let's begin with the two that are relevant here.

180-Degree front-to-back ratio: This ratio is simply the difference in gain between the maximum forward gain bearing and another bearing 180 degrees opposite.

Worst-case front-to-back ratio: This ratio uses the maximum gain anywhere in the rear quadrants as the basis for establishing the ratio relative to the maximum forward gain. In most instances, the side points, as defined by the 90-degree angle from the maximum forward gain heading, determine where the rear quadrants begin.

If we compare the azimuth pattern in **Fig. 25-1** with the two definitions, we begin to suspect that EZNEC is reporting the 180-degree ratio, while NEC- Win is reporting the worst-case ratio. However, we need not guess; we can find out easily enough from the far-field radiation tables, a portion of which we can view here.

Azimuth Pattern Elevation angle = 0 deg

Deg	V dB	H dB	Tot dB	Deg	V dB	H dB	Tot dB
90	-99.99	7.09	7.09	180	-99.99	-99.99	-99.99
95	-99.99	7.03	7.03	185	-99.99	-24.24	-24.24
100	-99.99	6.83	6.83	190	-99.99	-19.24	-19.24
105	-99.99	6.50	6.50	195	-99.99	-16.80	-16.80
110	-99.99	6.03	6.03	200	-99.99	-15.47	-15.47
115	-99.99	5.42	5.42	205	-99.99	-14.78	-14.78
120	-99.99	4.67	4.67	210	-99.99	-14.54	-14.54
125	-99.99	3.78	3.78	215	-99.99	-14.65	-14.65
130	-99.99	2.73	2.73	220	-99.99	-15.05	-15.05
135	-99.99	1.52	1.52	225	-99.99	-15.70	-15.70
140	-99.99	0.12	0.12	230	-99.99	-16.59	-16.59
145	-99.99	-1.49	-1.49	235	-99.99	-17.69	-17.69
150	-99.99	-3.34	-3.34	240	-99.99	-18.97	-18.97
155	-99.99	-5.50	-5.50	245	-99.99	-20.40	-20.40
160	-99.99	-8.08	-8.08	250	-99.99	-21.88	-21.88
165	-99.99	-11.28	-11.28	255	-99.99	-23.28	-23.28
170	-99.99	-15.57	-15.57	260	-99.99	-24.44	-24.44
175	-99.99	-22.41	-22.41	265	-99.99	-25.21	-25.21
				270	-99.99	-25.47	-25.47

For the antenna orientation in the model, 90 degrees represents the peak forward gain heading and shows a gain of 7.09 dBi. Directly opposite, at an angle of 270 degrees, we read a gain of -25.47 dBi. The 180-degree front-to-back ratio is the forward gain minus the gain directly opposite: $7.09 - (-25.47) = 32.56$ dB.

The rear quadrants begin at an angle of 180 degrees, 90 degrees to the rear of the maximum forward gain heading. At an angle of 210 degrees, we find the peak gain in the rear quadrants: -14.54 dBi. The worst case front-to-back ratio is the highest forward gain minus the highest gain in the rear quadrants:
 $7.09 - (-14.54) = 21.63$ dB

Of course, the tables can always be used to find any set of gain relationships we may desire to know. They are a too-little used portion of the data available from NEC or MININEC models.

For the present case, one might wish to ask which of the two front-to-back figures is “right,” “more accurate,” or “more important.” Actually, both figures are equally right and accurate, since both calculate with equal precision from the data. Importance is a judgment to be gleaned from the concerns we bring to the investigation. If we are concerned about rear quadrant QRM in general, then the worst-case figure may be a more useful indicator of the likely levels we might experience when using this antenna. If we are seeking the best figure we can obtain relative to the immediate rearward direction, then the 180-degree figure may be more useful.

In some cases, the difference may not make much of a difference. **Fig. 25-2** shows the free-space azimuth pattern for a very long 1-octave LPDA. For most practical purposes, any measure would suffice, since the radiation to the rear does not exceed -40 dB in any rearward direction. Nevertheless, the pattern does reveal another facility of NEC and MININEC programs that we use too little—the ability to expand the plotting scale to investigate small pattern features.

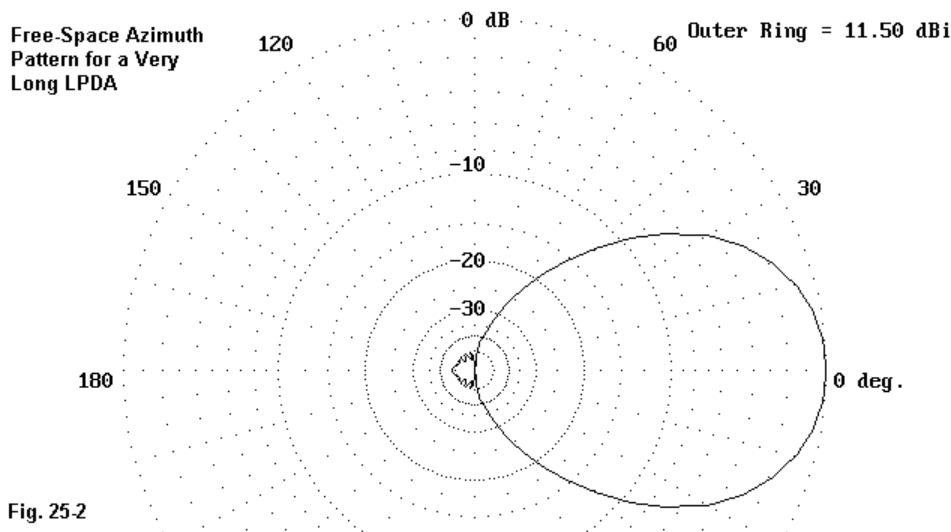
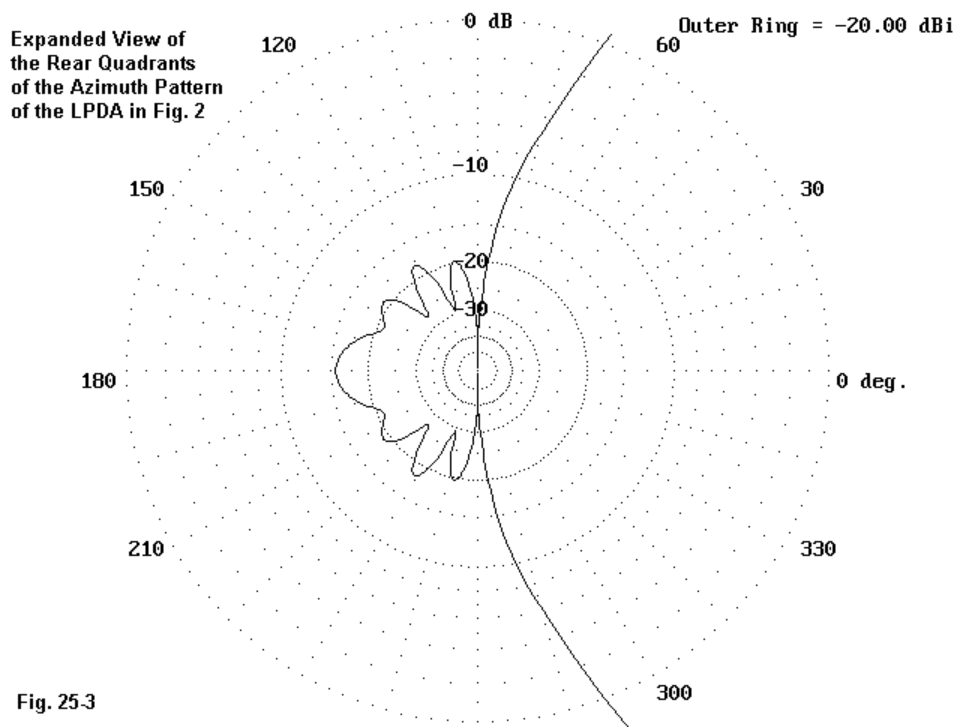


Fig. 25-3 sets the outer limit of the plot at -20 dB. This move automatically expands the rear quadrant lobes to much more visible proportions. Here, we can clearly see that no lobe to the rear exceeds the one that is 180 degrees opposite to the forward lobe. However, the gain value may seem odd: about -15.5 dB from the outer ring value. If we add this value to the outer ring value of -20 dB, we find the 180-degree lobe to have a gain of -35.5 dB. If we then use the 11.5-dBi gain of the forward lobe as our reference, we obtain a 180-degree front-to-back ratio of 47.0 dB. Since the 180-degree rear lobe is the strongest within the rear quadrants, this value is also the worst-case front-to-back ratio.



The very high front-to-back ratio exhibited by the long LPDA may be very significant to commercial applications. Indeed, the long LPDA has in many arenas supplanted other wire arrays, and the high potential front-to-back ratio is one important reason (along with the absence of side lobes). Most amateur radio antenna applications tend to settle for a minimum front-to-back ratio of 20 dB, with anything more being a bonus. Hence, the 180-degree and the worst-case ratios might equally satisfy amateur antenna users. (Still, we know which of the two figures an advertiser is more likely to use.) Nonetheless, we might adjudicate the differential we often find between the two numbers by appealing to a third front-back notion.

Front-to-rear ratio: To overcome misleading aspects of both 180-degree and worst-case front-to-back ratios, some engineers have adopted the front-to-rear ratio. This ratio is based on averaging the power gain of the antenna over the rear quadrant and using the resulting figure as the basis for a ratio with the forward gain. There is no general standard on exactly how many data points to use or where to locate them.

Let's do a simple calculation relative to the antenna in **Fig. 25-1** and its far-field table. Since our antenna pattern is symmetrical on each side of the center line from front-to-rear, let's just add up the gain values at 5-degree intervals from the side point to the rear point and then obtain their average by dividing the total by the number of data point. There are statistically more sophisticated ways to obtain an average value, but this simple exercise can be easily checked by consulting the table above.

Skipping the -99.99 side inset value, we obtain a total of -344.40 for the 18 points from 185 degrees through 270 degrees. The average gain for the quadrant is -19.13 dBi. Since the forward gain is 7.09 dBi, the front-to-rear ratio is $7.09 - (-19.13) = 26.22$ dB.

The front-to-rear ratio becomes a reasonable evaluation of overall rearward performance as we compare it a. to the smooth rearward lobes of the antenna pattern in **Fig. 25-1** and b. to the 180-degree and the worst-case values, which are higher and lower, respectively, than the front-to-rear ratio we just roughly calculated. In this case, using any one of the three ratios would not seriously mislead a potential user of the antenna.

There are cases, however, in which no single figure will adequately evaluate an antenna's rearward performance. **Fig. 25-4** is the free-space azimuth pattern of a folded X-beam. Below is a partial set of gain values.

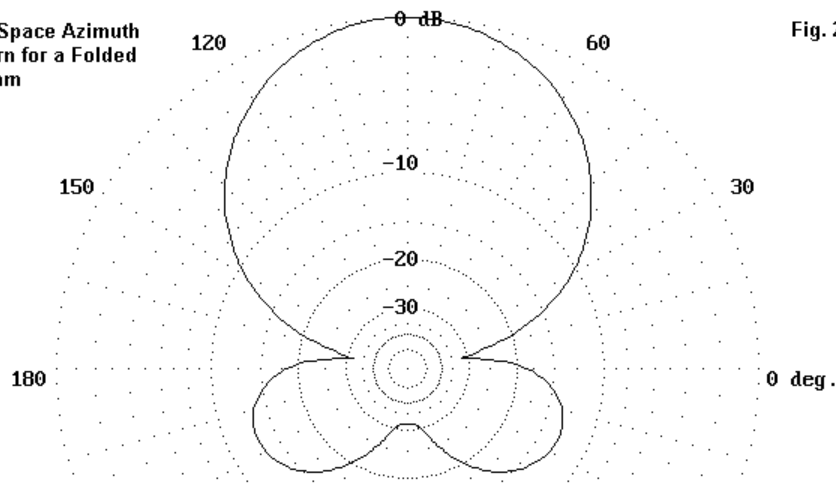
Free-Space Azimuth
Pattern for a Folded
X-Beam

Fig. 25.4

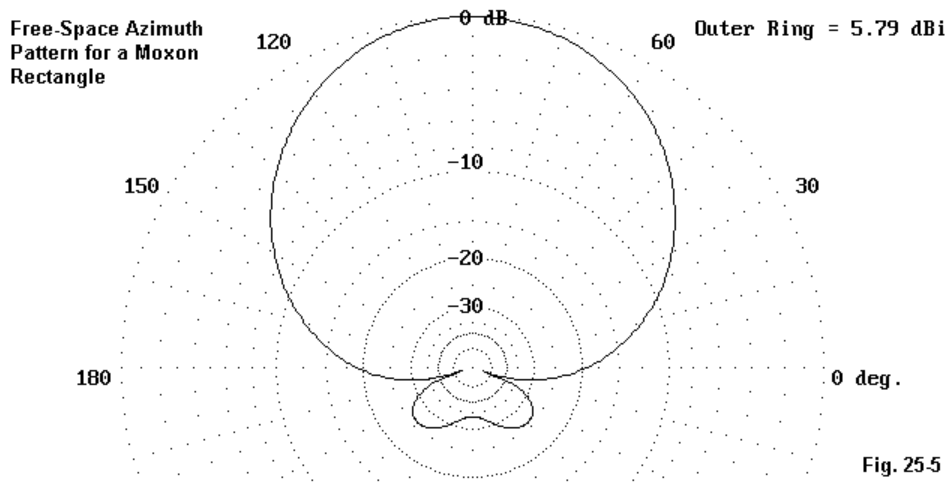
Azimuth Pattern Elevation angle = 0 deg

Deg	V dB	H dB	Tot dB	Deg	V dB	H dB	Tot dB
90	-99.99	7.34	7.34	185	-99.99	-8.64	-8.64
95	-99.99	7.28	7.28	190	-99.99	-7.10	-7.10
100	-99.99	7.11	7.11	195	-99.99	-6.17	-6.17
105	-99.99	6.82	6.82	200	-99.99	-5.67	-5.67
110	-99.99	6.41	6.41	205	-99.99	-5.52	-5.52
115	-99.99	5.86	5.86	210	-99.99	-5.68	-5.68
120	-99.99	5.18	5.18	215	-99.99	-6.12	-6.12
125	-99.99	4.34	4.34	220	-99.99	-6.84	-6.84
130	-99.99	3.33	3.33	225	-99.99	-7.86	-7.86
135	-99.99	2.11	2.11	230	-99.99	-9.18	-9.18
140	-99.99	0.65	0.65	235	-99.99	-10.86	-10.86
145	-99.99	-1.14	-1.14	240	-99.99	-12.93	-12.93
150	-99.99	-3.35	-3.35	245	-99.99	-15.46	-15.46
155	-99.99	-6.21	-6.21	250	-99.99	-18.41	-18.41
160	-99.99	-10.20	-10.20	255	-99.99	-21.47	-21.47
165	-99.99	-16.67	-16.67	260	-99.99	-23.64	-23.64
170	-99.99	-23.93	-23.93	265	-99.99	-24.26	-24.26
175	-99.99	-15.57	-15.57	270	-99.99	-24.26	-24.26
180	-99.99	-11.14	-11.14				

Using the 90 and 270 degree points, we obtain a 180-degree front-to-back ratio of 31.60 dB. Using the 90-degree point and the 205-degree point, where rearward gain is highest, we obtain a worst-case front-to-back ratio of 12.86 dB. For the third figure we shall again average the values between 185 degrees and 270 degrees (even though the minimum gain bearing is 170 degrees rather than 180 degrees) and divide by the 18 points to get a -12.23 dBi average rearward gain. The resulting front-to-rear ratio is 19.57 dB.

The very high disparity among the three values makes it dubious whether one of those values would fully inform a potential antenna user. The very high 180-degree front-to-back value is very misleading unless QRM from quartering rear directions is judged in advanced to be an insignificant consideration. Even the front-to-rear ratio overrates the antenna's rearward performance by at least a receiving S-unit (6 dB) relative to the gain of the antenna at the 205-degree bearing and its counterpart to the other side. Nonetheless, using only the worst-case value can under-estimate antenna performance overall.

In the end, obtaining all three figures and applying them in conjunction with careful observation of the azimuth pattern may be the only means of arriving at a satisfactory evaluation of the design. Of course, this evaluation only makes sense in the context of a set of performance objectives and needs.



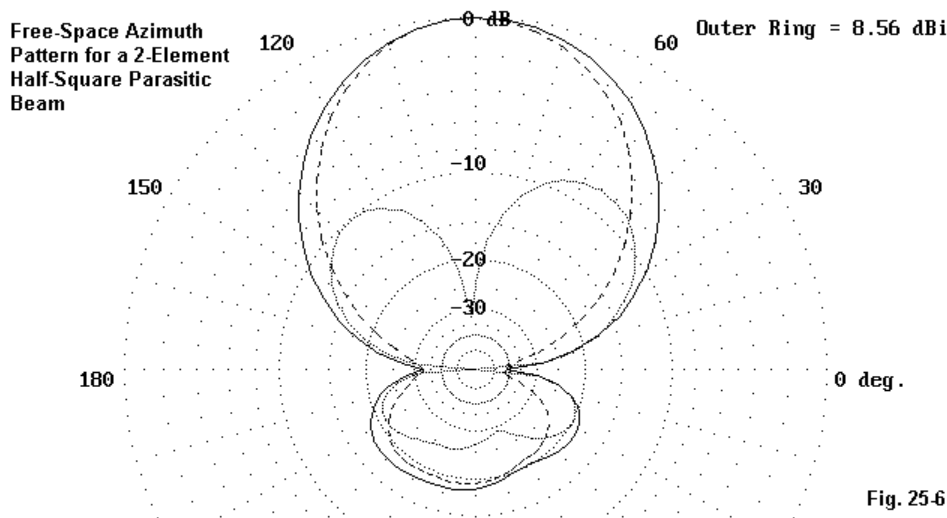
Simple techniques of calculating front-to-rear ratios, such as the ones that we have used to illustrate the principle, suffer limitations. One such limitation is illustrated in **Fig. 25-5**. The free-space azimuth pattern of the Moxon rectangle seems innocent enough until we try to calculate a front-to-rear ratio. Here is the table for our task. (Model 25-2)

Azimuth Pattern				Elevation angle = 0 deg			
Deg	V dB	H dB	Tot dB	Deg	V dB	H dB	Tot dB
90	-99.99	5.79	5.79	180	-99.99	-13.37	-13.37
95	-99.99	5.74	5.74	185	-99.99	-16.95	-16.95
100	-99.99	5.60	5.60	190	-99.99	-21.96	-21.96
105	-99.99	5.35	5.35	195	-99.99	-31.20	-31.20
110	-99.99	5.01	5.01	200	-99.99	-36.45	-36.45
115	-99.99	4.57	4.57	205	-99.99	-26.38	-26.38
120	-99.99	4.03	4.03	210	-99.99	-22.85	-22.85
125	-99.99	3.38	3.38	215	-99.99	-21.12	-21.12
130	-99.99	2.63	2.63	220	-99.99	-20.33	-20.33
135	-99.99	1.78	1.78	225	-99.99	-20.15	-20.15
140	-99.99	0.80	0.80	230	-99.99	-20.43	-20.43
145	-99.99	-0.29	-0.29	235	-99.99	-21.10	-21.10
150	-99.99	-1.52	-1.52	240	-99.99	-22.10	-22.10

155	-99.99	-2.89	-2.89	245	-99.99	-23.34	-23.34
160	-99.99	-4.44	-4.44	250	-99.99	-24.73	-24.73
165	-99.99	-6.20	-6.20	255	-99.99	-26.07	-26.07
170	-99.99	-8.21	-8.21	260	-99.99	-27.15	-27.15
175	-99.99	-10.55	-10.55	265	-99.99	-27.82	-27.82
				270	-99.99	-28.04	-28.04

The 180-degree front-to-back ratio is 33.83 dB. If we use our standard simple front-to-rear calculation for headings from 185 degrees through 270 degrees, we obtain a value of 30.13 dB. However, this value includes points within the main lobe as well as the deep side null—which occurs at 200 degrees. In some easy cases, the higher main lobe values may offset the side null, but in other cases, the inclusion of these values may skew the overall performance. If we use only those values from 205 degrees onward, the average rear gain becomes -23.69 dBi for a front-to-rear ratio of 29.48 dB. Although the differences from the simple calculation is not great in this case, the inclusion of main lobe and side null values reduces the degree of trust we can place in the first result. The fact that the revised front-to-rear ratio is different from the simple result should suffice to illustrate the care that must be used in developing such front-to-rear values.

So far, we have worked with symmetrical azimuth patterns and could therefore use only one of the rear quadrants to calculate a front-to-rear ratio. Consider **Fig. 25-6**. (Model 25-3)



The azimuth pattern for the half-square 2-element parasitic array has a feedpoint that is necessarily offset from center. Hence, as the vertical and horizontal components of the overall pattern show, there is good reason for the total azimuth pattern to be slightly non-symmetrical. The forward lobe is scarcely affected, but the rear pattern shows great unevenness.

Had we calculated the front-to-rear ratio by our simple method, the result would be too high or too low, depending on from which side of the centerline we draw our data. Obviously, an operational method of developing front-to-rear ratios should include both halves of the rear radiation pattern. Very likely, the operational method should also include a much finer resolution (perhaps to 1 degree) to obtain many more data points than we have used in our illustrations.

So far, all of our sample cases have used free-space azimuth patterns. We have a dangerous tendency to assume that, when we take azimuth patterns at some antenna height over ground, the maximum rear gain occurs at the same elevation angle as the maximum forward gain. Therefore, the resulting front-to-back ratio

becomes a safe estimate for the antenna's performance for signals from the rear of the antenna.

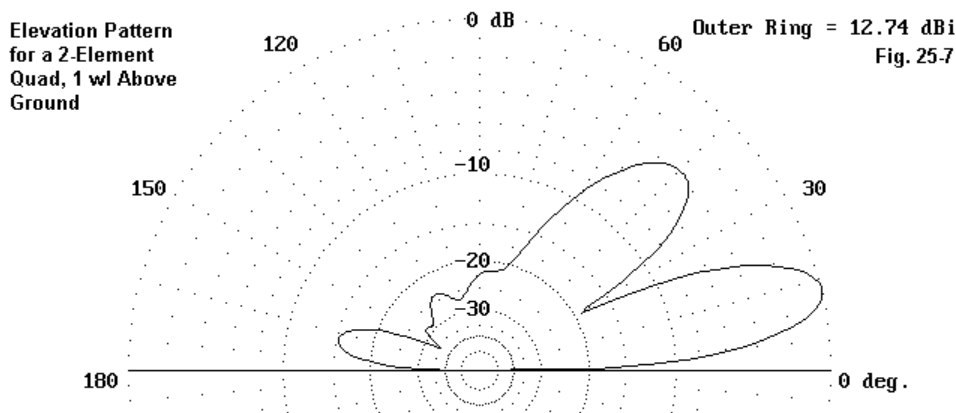
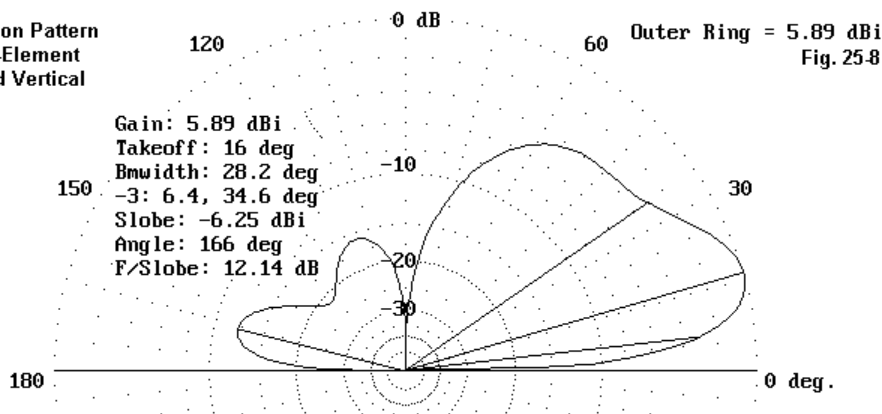


Fig. 25-7 illustrates our presumptions in action. I purposely chose a quad model with significant rearward radiation to make the rear lobe clear in the elevation pattern. In this case, and in a large number of other cases involving horizontally polarized antennas, the rear lobe angle of maximum radiation is the same as the forward lobe angle of maximum radiation: about 13 degrees above the horizon for this illustration.

However, it is never safe to assume that this phenomenon is universal. For example, see **Fig. 25-8**. The elevation pattern for this phased array of vertical elements shows a forward elevation angle of maximum radiation of 16 degrees. The corresponding rear elevation angle of maximum radiation is two degrees lower, at 14 degrees above the horizon. Although the lobe-angle difference may not make a very significant operational difference with this array, it does illustrate the need to locate such angles when evaluating the front-to-back or front-to-rear performance of an antenna placed over real ground. In this case, the elevation pattern supplements simple azimuth patterns. Of course, we can always turn to the far field tables for both elevation and azimuth patterns to complete our data needs.

Elevation Pattern
for a 2-Element
Phased Vertical
Array



Many antenna modelers approach their software with limiting assumptions. Perhaps the most common are oversimplifications of the idea of a front-to-back ratio. In this set of exercises, I hope we have gone some distance in showing this basic idea to be somewhat more complex than some of us suspected. But more importantly, I hope the exercises have shown some of the resources within NEC and MININEC with which we can do a full analysis and evaluation of an antenna's rearward performance.

* * * * *

Models included: 25-1 through 25-3.

Appendix: Antenna Models

This volume—the first of 3 volumes—of antenna modeling notes comes with 93 antenna models, almost all of which have text references (Model xx-x). I have included all of the models in 3 different formats: .EZ for users of EZNEC, .NWP for NEC-Win Plus users, and .NEC for NEC-Win Pro, GNEC, and generic NEC-2 core users. The folder (directory) structure simply follows this scheme.

N:\models\ez	for EZNEC format files
N:\models\nec	for generic ASCII NEC files
N:\models\nwp	for NEC-Win Plus format files

I recommend that you copy the most relevant set of files into your hard drive for use. NEC-Win Plus, for example, will store its output files in the same directory as the basic model file, and that requires a space to which the computer can write. Each file name follows the text by starting with the column number, followed by the model number within the chapter. So Model 14-3 is the third model used in conjunction with column #14.

The files are not likely to add to your collection of models in the sense of providing new or interesting antenna designs. For that purpose, I have assembled collections of interesting models from my own storehouse. These collections are available from *antenneX*. The files that go with this volume of antenna modeling notes are those referenced in the text. As such, they are illustrations of the principles discussed in the text. Hence, you may read along with your own modeling software active and investigate further the model under discussion.

You will find some discrepancy between many of the model outputs and the performance figures cited in the text. This situation has a number of sources, all related to the fact that I began the series in the last century (the late 1990s). In some cases, I simply could not find the file used for a column, and so I had to reconstruct as best I could the model under discussion. Sometimes, the text did not provide complete modeling data, so I approximated the text model as closely as

possible. Although the exact figures may not jive between text and model, the trends certainly do.

In other cases, software developments are the source of slight numerical deviance between the model as used when I wrote and the same model when you run it on your current software. When I began the series of columns, EZNEC was a DOS program, and now uses Windows. NEC-Win Plus did not yet exist. In the course of time and software development, the NEC cores have undergone customizing and enhancement for speed—with special reference to the latest Fortran compilers. In the process, there have been changes in the order of operations and rounding conventions, enough to create slight output differences. With respect to guiding construction, showing performance trends, or yielding reliable analysis, the changes make no noticeable difference relative to older outputs. However, in order to show trends as sensitively as possible, many parts of the output data cited in the text will be overly precise. That fact will create an illusion of difference where no operationally significant difference exists.

I regret that I cannot include in this collection of models samples of MININEC files. I have and use several MININEC programs, including ELNEC, AO, MMANA, NEC4WIN, and the most recent and able version, Antenna Model. Each version uses a different file format, and there are few means of converting a MININEC file from one format to another except by writing the model from scratch. In contrast, I was able to convert files from one to another of the NEC formats. Conversion is not perfect, and so some EZNEC files given in English measures will appear in metric form in the generic NEC and NEC-Win Plus formats.

With these limitations in mind, I hope that the attached files enhance your safari through the topical jungle within this volume of antenna modeling notes.

Other Publications

We hope you've enjoyed this first of several volumes of the **Antenna Modeling Notes: Volume 1**. You'll find many other very fine books and publications by the author L.B. Cebik, W4RNL in the ***antenneX Online Magazine BookShelf*** at the web site shown below.

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