



Modeling Loads and Transmission Lines



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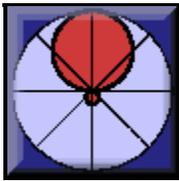
Questions often arise about how properly to model loads and transmission lines within various versions of NEC and its commercial implementations. So I thought a brief primer might be useful. The utility of a primer is unquestionable, but that it can be brief quickly proved false.

Although the information is covered in both the NEC manuals and the instruction manuals that accompany commercial implementations, something more systematic seems needed--something with examples carried from one level to another so that users can see the transitions and get a sense of what is and is not possible--and how to get around what is not possible. I also discovered that we cannot begin in the middle, assuming that readers know the basics. When it comes to antenna modeling, modelers come at the enterprise from many different backgrounds, enter it with different interests, and get into modeling loads and transmission lines with different goals and interests. Hence, everybody knows the basics in a slightly different way.

So we shall begin at the beginning--with simple antennas and simple loads-- and work our way to more interesting examples. Along the way, we shall have occasion to mention some useful adjunct calculation programs and some equations for use on calculators. Modeling programs just do not do everything we need to do to make good models. But there are some utility programs that can take most of the drudge work out of modeling.

In the end, we shall quit far short of exhausting the subject, but a little ways beyond exhausting the writer. However, incomplete, I hope the information is useful to you.

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1. Getting Oriented

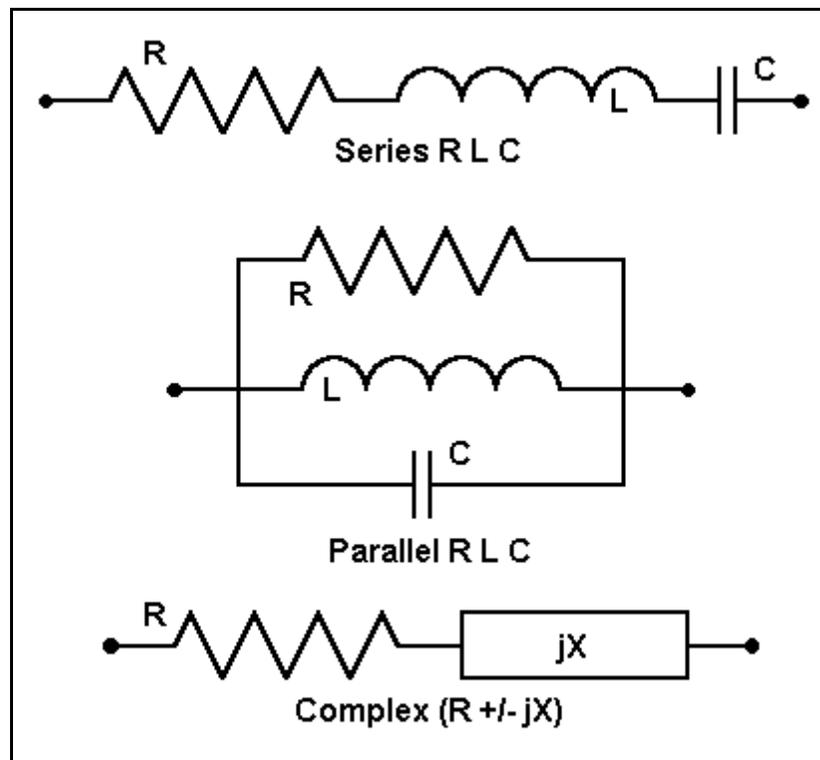
Basic Series Loads and Simple Antennas



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NEC models antennas as collections of straight wire segments. Within NEC programs, we specify the antenna structure as one or more wires and then subdivide each wire into one or more segments. One way in which we can modify one or many segments of an antenna is to load them.

A load is simply a combination of resistance or reactance assigned to a segment. NEC is capable of assigning load as series R-L-C circuits (load type 0), parallel R-L-C circuits (load type 1), or as complex series $R \pm jX$ impedance loads (load type 4). (NEC may also assign loads as series or parallel R-L-C values per unit of length (load types 2 and 3), but we shall not have occasion to deal with this option in this series.) Figure 1 surveys some of the options.



In this part of the series, we shall deal only with series $R \pm jX$ impedance loads. They are among the simplest to implement and manipulate, although they do have some limitations. We shall also mention the load that many users do not think of as a load: wire conductivity (load type 5). (NEC-4 is also able to handle wire relative permeability.)

For many modelers, this first part is a review of the most basic sort. However, within the review there might be a detail or two that will make a more advanced procedure a little bit clearer.

A. The Load Entries: Material Loading

NEC input consists of a series of lines of entry in a model description. Each line corresponds to a "card" in older FORTRAN input systems. It has a label and a sequence of entries that the program interprets as values and instructions. CM entries are comments; GW entries are straight wire specifications; etc. Load entries carry the designation LD.

Unless we wish to use lossless wire in our model, we ordinarily specify a material or materials for the antenna elements. With commercial implementations of NEC, this specification may occur as a special entry apart from program means of introducing other types of loads. Nevertheless, the specification results in LD entries of type 5. Here is a typical entry:

```
LD 5 1 1 101 5.8001E7
Type Tag 1st-Seg Last-Seg Conductivity
```

The entry spacing has been widened to permit labeling the parts. A Type 5 load is a round wire internal impedance specification. The Tag indicates the wire number, here the only wire of a simple dipole antenna. The first and last segment designators set the limits of application of the value. Finally, the conductivity figure is in Siemens(mhos)/meter. The 1-wire antenna has 101 segments, all with the conductivity of copper.

Various implementations of NEC may handle the conductivity figure differently. Some provide named entries (for example, "copper"). Others provide conductivity figures or provisions for conductivity entry so that users can specify exact material properties. Still others provide custom entry in terms of resistance per unit of length, the inverse of conductivity. However entered, the entry is translated to an input of the type shown or its programmatic equivalent.

Unless an antenna uses materials of very specialized sorts, performance will not be altered to a significant degree by specifying either copper or aluminum (in its own various types). EZNEC permits only a single wire material specification, and users may use with confidence in the results either the conductivity of the "worst" material or the conductivity of the most dominant material in the antenna. However, NEC itself permit all manner of combinations of materials to be specified. NEC-Win Pro permits all the element load entries permitted by NEC itself, if modeling changes of material is an important consideration. With access to individual line entries, one may also within NEC-Win Pro load all segments or all wires to the same value in a single line by specifying the tag and the first and last segment entries as 0.

B. The Load Entries: Spot Loads

All spot or specifically placed loads within NEC consist of mathematical objects that do not themselves radiate, whatever the size of their physical implementations. However, the segment(s) on which they are placed do enter into the calculations for mutual impedance and current and are thus a part of the calculations for antenna pattern and other properties. Consequently, the error introduced by treating the loads themselves as non-radiating is normally insignificant. Spot loads do affect the current magnitude and phase along the antenna wire in mathematically correct ways to replicate reality very closely. Hence, they are normally a very accurate way to model loading coils and capacitors, traps, and resistances.

Entering spot loads differs according to the implementation of NEC. Since we shall restrict ourselves in this part of the series to complex impedances composed of resistance and reactance

only, we shall speak only in terms of type 4 loads for the moment. A basic NEC entry cord for such a load might look like the following:

```
LD 4 1 77 77 0 870.4
Type Tag 1st-Seg Last-Seg Resistance Reactance
```

At wire (tag) 1, on segment number 77 (first and last segment), we introduced a load of type R +/- jX with a resistance of 0 ohms and an inductive reactance of 870.4 ohms. (R and jX are always entered in ohms.) Although a 0-ohm resistance is does not coincide with reality, introducing pure reactances as a step in the modeling process can be useful, as we shall see down the line.

Specifying spot loads within commercial implementations of NEC may take a number of forms. In EZNEC, there is a separate load screen, a sample of which is replicated in Figure 2.

----- LOADS -----						
Load	Wire	Wire #/Pct	From End 1	R (Ohms)	X (Ohms)	
	Seg	Actual	(Specified)			
1	25	1/24.24	(1/24.00)	0.000	870.400	
2	77	1/75.74	(1/76.00)	0.000	870.400	

number=load to change **A**dd loads **D**elete loads **G**roup edit
change load **T**ype (Laplace/Z)

<ESC> = Main Menu

Making an entry consists of adding the desired number of loads and then specifying their positions and values. In EZNEC, the user specifies a position in terms of a percentage along the wire. The program places the load on the segment closest to the specified percentage and lists the segment number and the actual percentage along the wire. R and X entries are self-explanatory.

NEC-Win Pro permits three modes of load entry. One is to create a load line of the type shown by typing it into the antenna description. One may also access a "Place Load" box from the description and enter the requisite data, as shown in Figure 3 (which replicates load #1 in the EZNEC figure).

LD - Place Load

Place Load

Tag #:

Starting Segment #:

Ending Segment #:

Load Type

Wire Conductivity
 Series RLC
 Parallel RLC
 Complex Load
 Short All Loads

Wire Conductivity (mhos/meter):

R(Ohms):

L(Henrys):

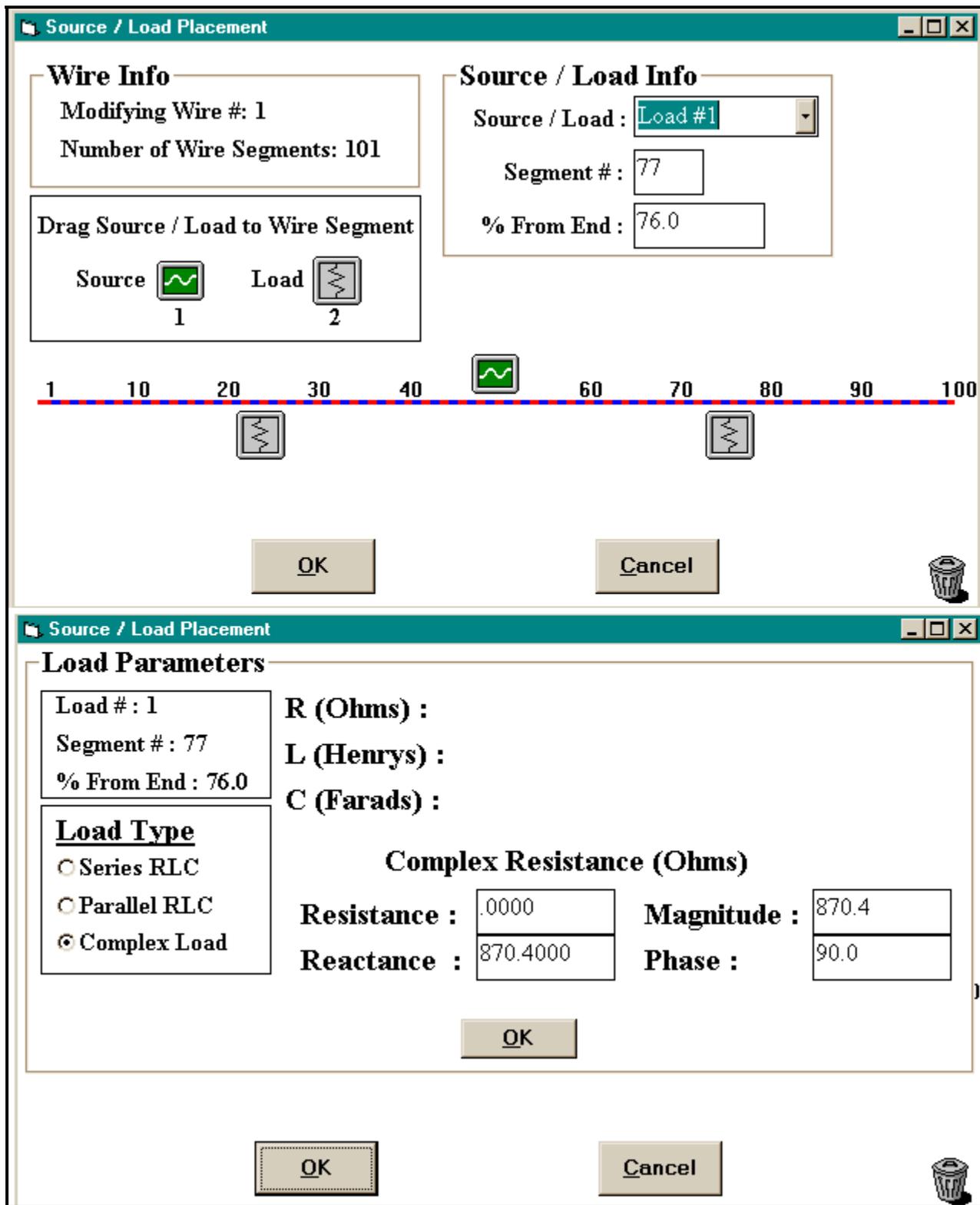
C(Farads):

Complex Resistance (Ohms)

Resistance:

Reactance:

An alternative exists within the NEC-Win Basic module, which is provided for rapid entry of basic antenna information. The user may "drag and drop" loads (and sources) and then access a data box for numerical entries and load-type specification. Although the first two methods refer only to tags, this system also provides a percentage along the wire, which is useful for calculating the precise linear position of loads. Figure 4 shows an example.

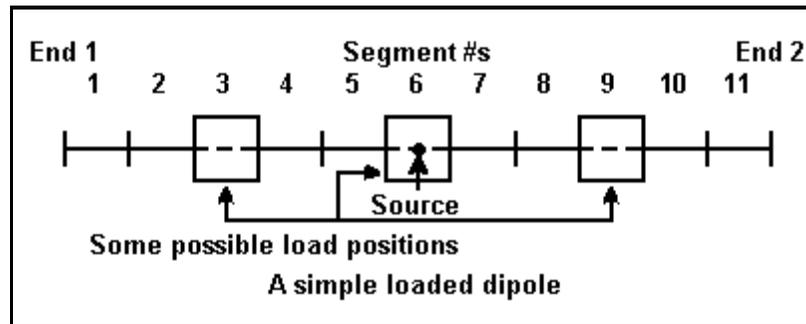


Virtually all other implementations of load placement in commercial implementations of NEC will be a variant on these systems.

C. Simple Loads and Simple Antennas

The Type-4 loads we place on an antenna consist of series R-jX circuits. These circuits are also in series with other features of the wire, including the source. Figure 5 presents a simple dipole,

modeled in 11 segments, with possible loads. One position is at the source, simulating a series loading inductor (or capacitor). Other loading positions are indicated by the outer 2 loads, usually spaced equidistant from the center on symmetrical antennas like the dipole.



1. Loading Short Dipoles: X Only

To illustrate the utility of modeling loads, let us consider a 20-meter wire dipole of #14 copper wire. (I shall not show the LD line for the type-5 element loading information.) A full (1/2 wavelength) resonant dipole in free space would be just about 33.736' long (404.84" or 10.282 meters) at 14.15 MHz. The free space gain for reference is 2.08 dBi. The numerical precision outstrips real building conditions, but is here used as also a measure of resonance. Resonance in an antenna is the condition of having a feedpoint impedance that is purely resistive, that is, with 0 reactance. For modeling purposes, depending upon one's goal, resonance may be approximated for any design frequency with greater or lesser limits. In most pure modeling exercises, I tend to use +/- j1 ohm as the limits of resonance. In this example, the feedpoint impedance happens to calculate to 73.06 - j0.01 ohms. This particular model was done in NEC-4 using 101 segments. (If not pressing the limits of either program, NEC-2 results might still vary from those of NEC-4 by an amount that has no practical significance. Moreover, results from different implementations of NEC may also vary, according to input and output processing--but, again, only by amounts that have no practical significance.)

Like loads, sources are distributed along a segment, although it is often convenient to think of them as appearing in the center of a segment. Hence, for a single-wire element, we need an odd number of segments in order to truly center a single source. (We shall take up split sources in just a little bit.)

This 20-meter full-size dipole will be our reference point. Let us now cut the wire in half, for a length of 16.868' (2.571 m). We shall retain the 101 segments, since each segment turns out to be just about 2" long (about 5.1 cm). At 14.15 MHz, gain is 1.73 dBi, and the feedpoint impedance registers as 12.67 - j870.4 ohms.

Let us introduce a complex or $R \pm jX$ load at the center of the antenna (the 50% mark or segment 51). Let it be a pure inductive reactance of 870.4 ohms ($R=0; jX=870.4$). Loads are considered by NEC to be in series with sources. Therefore, it should be no surprise that rerunning the model produces feed point impedance of 12.67 - j0.04 ohms. The effect is equivalent to introducing a perfect center loading inductor having the require reactance. If the inductor could be perfect, the gain of the antenna would not change: 1.73 dBi.

As an interesting exercise, let us replace the center load with two other loads spaced away from the center. In fact, let us place them at the points where equal inductive reactances of 870.4 ohms

(with no resistance) also produce resonance. These positions happen to be at segments 25 and 77 of our 101 segment model, a little over 24% inward from the ends. In other words, the positions are just beyond the midpoints of each half of the dipole. Interestingly, the antenna gain is 1.76 dBi, an insignificant amount greater than with center loading. (Mid-element loading of dipoles and dipole-based antennas is done not for added gain, but for other antenna properties, some related to the feedpoint impedance on the prime and other frequencies, some related to mechanical considerations).

The feedpoint impedance climbs to $26.08 + j1.65$ ohms. (This value is as close to resonance as one may come in the model without changing the value of the loads.) The higher feedpoint impedance is one of those good reasons for using mid-element loading in preference to center-loading.

As an initial exercise, try moving these mid-element loads, without changing the reactance, one segment closer to center and then one segment farther from center. Notice that a 2" change in placement per load results in over 60-ohms reactance at the feedpoint. You may also replicate this entire sequence using other element wire diameters. Expect to find a different resonant length for the basic reference antenna, as well as changes in the requisite loading to restore resonance using either a center or mid-element loading technique. The patterns will familiarize you with trends that may be useful in the design of more advanced antennas.

As a further exercise, the modeler may also wish to check all of the points along the antenna length between center loading and the point at which mid-element loads reach the value of the center load. One will discover a curve of ever-more rapidly increasing reactance values. In fact, if one continues the exercise beyond the point used in this exercise, there will come a point where adding reactance to achieve resonance is no longer effective in terms of practical inductors. Yet helical structures placed at the ends of short inductors can resonate the antenna. Are they not inductors, that is vehicles of inductive reactance?

In fact, although the structures exhibit inductive reactance, at the ends of antennas they function primarily as lengths of wire necessary to achieve the current distribution along the entire wire to produce a feedpoint impedance that is wholly resistive. Many non-helical structures, including so-called capacitive hats, may achieve similar results. The inductive reactance of end coils is largely incidental to most amateur designs. However, it may be used for other purposes, such as in a trap. (Models suggest that the trap is not an appendage to the antenna length that it terminates, but is indeed a part of that length.)

2. Split Sources and Loads (and Alternatives)

The procedures noted so far for applying loads to antennas, especially at sources, are applicable only to linear elements. When elements meet at angles and the apex of the angle is the source, we must use another procedure. The procedure is also applicable to linear elements, although hardly ever used except in exercises like this one.

Our original dipole consisted of a single wire, divided into 101 segments. Let us replace that wire with two wires, each half the length of the original. We shall assign 50 segments to each and let them join in the middle. Unless we use a different feed system, the source will not be centered.

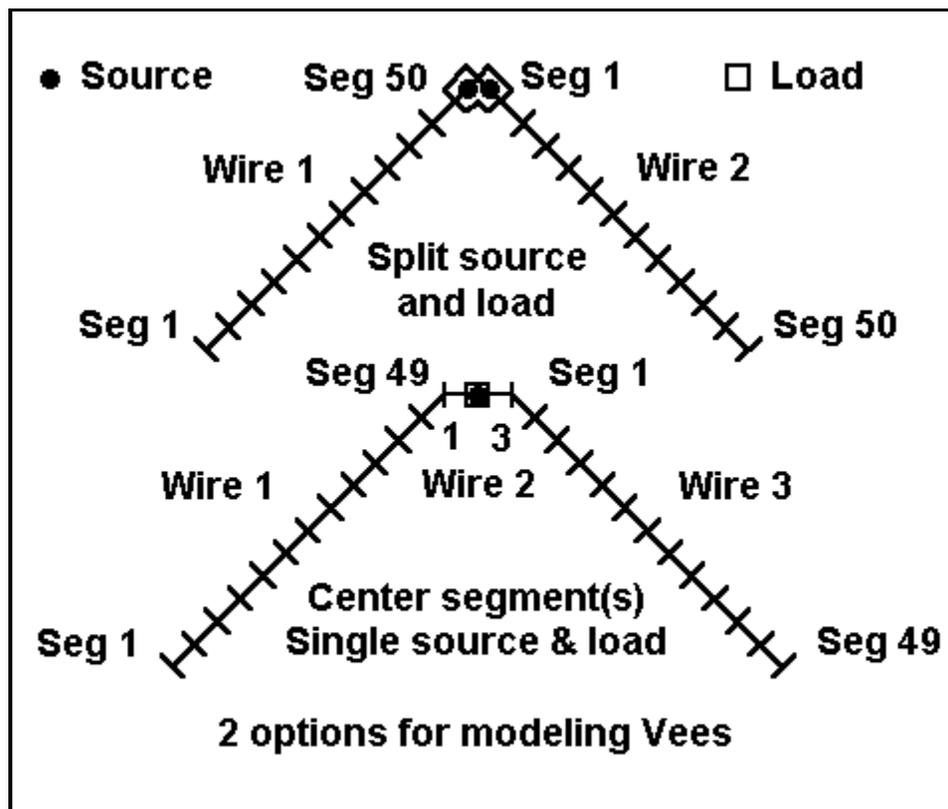
NEC is able to use a split source. If we specify either a split voltage or a split current source, we should place the source on the last (right-most) segment of the first wire--the segment joining the

second wire. It pays to use any graphical antenna viewing system to ensure that the source is properly place. On some systems, the source will show up as two sources, one on the last segment of the first wire and the other on the first segment of the next wire.

NEC treats these two sources as being in series, and the total impedance is simply the sum of the individual impedances of the source. Normally, only the composite values is shown in the outputs of commercial implementations. For our reconfigured dipole (with no change in total length, the split feed system returns a gain of 1.73 dBi (identical to that of the center, single source model) and a feedpoint impedance of $12.93 - j879.4$ ohms. The source impedance is within a percent or so of the earlier model.

Loading the antenna mid-element requires no change except for the precise value of the load assigned to each of the two med-element loads. Moreover, because the segmentation varies very slightly from the single-source model, moving the loads inward or outward by one segment may surround true resonance, rather than hitting it with precision. (Obviously, other models may reverse the effect.)

Loading at the antenna center, however, calls for a change in procedure. Instead of assigning a single load to a single segment, we create two loads, each half the total load. We assign these loads to the segments containing the sources. Hence, the last segment of wire 1 (Seg 50) receives a load of $R=0, X=j439.7$, and the first segment of the next wire (Seg 1) receives an identical load. In our example, the source impedance returned a value of $12.93 + j0.04$ ohms.



The figure shows a schematic representation of the principle of split sources and loads applied to an inverted Vee. For simplicity, I simply rotated the dipole elements downward (still in free space) by 45 degrees. This actually made the elements a bit shorter than half size. But since we are going to load the elements, the difference make no big difference.

Using a split source, the model returned an impedance of $6.84 - j900.9$ ohms. Adding purely reactive loads of $-j450.45$ ohms to each segment containing a source produced a source impedance of $6.843 + j0.05$ ohms, showing that the technique indeed works.

There is an alternative to using split sources and loads. At the junction of the wires where we desire to place either a source or a load (or both), we may do a little substitute modeling. At the center point of the antenna element, create a short wire that is horizontal. To this very short wire, connect the angled wires at their normal lengths or their lengths minus half the length of the new wire.

Recommendations concerning the minimum length of this wire vary. However, if the wire has the same diameter (or radius, in pure NEC terms) as the connecting wires and if the segments are about the same length as the segments in the joining wires, it will make little difference whether the new center wire has 1 or 3 segments or what its precise length is. Place a single source on the center segment of the new wire.

For the inverted Vee with 100 segments, I created first a 3 segment wire 6" long. This preserved the approximate 2" length of all segments. I made no direct changes in the angled-wire lengths, but simply connected them to the new wire. This did alter the overall length of the antenna very slightly. The source was placed at the center of this new wire (Wire 2, Seg 2). The source impedance returned a value of $6.72 - j855.6$ ohms. A single purely inductive load of the same absolute value placed on the same segment (in series with the source) eliminated the reactance.

I also use a single 2" long horizontal segment as the center wire. This changed the antenna length slightly less than the 6" center wire. The source was placed on the only segment of this Wire (Wire 2). The source impedance returned was $6.687 - j875.7$, which a suitable inductive load placed on the same segment also eliminated.

With angled wires, expect minor variations in the values returned between the split source/load technique and the center section technique. In these examples, the values are well within the range of adjustments one normally expects to make with real antennas on site. However, the models with the longer horizontal center sections consistently showed a slightly higher gain than the split source model, and this effect showed up over real ground as well. The effect results from a portion of the antenna at the high current region being immune from the partial canceling of fields that results from the Vee configuration. Expect up to a 0.1 dB difference in gain, depending on the length of the center section.

These notes apply to heavily segmented models, such as the ones used in these examples. Ordinarily, NEC yields accurate results with linear elements with as few as 11 segments per half wavelength. Moreover, the fewer the segments, the faster the run-time for a model. However, where geometries grow complex even to the slightest degree--as in the case of the Vee--attention to segment length is advisable for the most accurate results. For a given configuration, the modeler should always achieve convergence to the degree required by the practical application. Convergence is simply raising the number of segments per wire (usually as a percentage) until further increases yield output figures that do not significantly vary from those produced by the next lower number of segments. These exercises, of course, are unduly segment and precision laden in order to familiarize you with the numbers and movements of numbers you may expect from the program.

D. Reactance, Resistance, and Q

Unfortunately, I know of no way of implementing an inductive (or capacitive) reactance as an antenna load without also having a series resistance. The ratio of the reactance to the series resistance defines the Q of the component used as the loading device.

Knowing or being able to estimate the Q of a loading capacitive or inductive reactance allows us to calculate the resistance needed to complete a realistic load. Most commercial inductors I have encountered have had a Q between 150 and 300. As an exercise, let's go back to the shortened dipole with 101 segments and see what these particular Qs do to antenna performance. The following input line shows a typical entry:

```
LD 4 1 77 77 2.9 870.4
Type Tag 1st-Seg Last-Seg Resistance Reactance
```

With a center load of $j870.4$ and infinite Q (or zero resistance), the antenna gain was 1.73 dBi, with a source impedance of $12.67 - j0.04$ ohms. If we let $Q=300$, then the resistance in series with the reactance is about 2.9 ohms. Under these conditions, the antenna model returns a gain of 0.83 dBi and a source impedance of $15.57 - j0.04$ ohms. The source impedance is clearly the sum of the load loss resistance and the source impedance without a loss. If we let $Q=150$, the load resistance is 5.8 ohms, returning a gain value of 0.09 and a source impedance of $18.47 - j0.02$.

Using the model with two mid-element loads of the same value as the center load and positioned to provide resonance, we come up with comparable figures. With no resistance, the model showed a gain of 1.76 dBi and a source impedance of $26.08 + j1.65$ ohms. With a Q of 300, each load has a series resistance of 2.9 ohms. This model shows a gain of 0.93 dBi and a source impedance of $31.55 + j1.52$ ohms. Lowering the Q to 150 requires a series resistance of 5.8 ohms, with a consequential gain of 0.24 dBi and a source impedance of $37.01 + j1.37$ ohms.

Notice that resistive component of the source impedance is close to but not quite the sum of the load resistors and the source resistance of the unloaded antenna. This difference exists because the current magnitude at the load is not the source current, but a slightly less value of about 97% source current. The loads, of course, are positioned only a little over $1/16$ of a wavelength away from the source.

NEC also returns data on the loads, which is accessible in the outputs of most implementations. Among the direct or calculable data is the power consumed in the loads. With a Q of 300, the center-loaded model consumes nearly 19% of the power in the load, while the mid-element loaded model consumes about 17% of the power between the two loading inductors. Hence, there is about a 0.1 dB difference in the gain of the two models.

At a Q of 150, the center-loaded model consumes over 37% of the power in the load, while the mid-element loads together consume a little under 30% of the power. Hence, there is a greater (0.15) difference in the gain levels of the two models. If mid-element loading begins to show a distinct advantage over center loading (apart from source impedance and mechanical considerations), it is with low Q loads.

As with models using pure reactances, the numbers yielded by your specific implementation of NEC may vary slightly from these. However, you should notice the same trends. As an exercise, you may wish to run load Qs ranging from 100 to 1000 for various load positions and values in order to familiarize yourself with the trends involved. Be sure to make use of all relevant outputs from NEC, including load data and current distribution along the antenna.

E. Capacitive Reactance as a Load

So far, we have dealt only with inductive reactance as a load. This feature of the exercise has been a product of our initial model: a heavily shortened dipole and some angular cousins. However, capacitive loading is also possible and often useful.

Consider a 20-meter dipole of #14 copper wire, which for free space resonance was 33.7' long. The source impedance was just about 73 ohms. Suppose, as an exercise, we make the antenna 36' long. The source impedance now registers as 90.9 + j102.8 ohms: highly nonresonant.

However, using the same techniques as shown earlier, let us insert in the model a load in the same segment as the source. The load will be $R=0$, $jX=-102.8$ ohms.

```
LD 4 1 51 51 0 -102.8
Type Tag 1st-Seg Last-Seg Resistance Reactance
```

Now the model will show a source impedance of $90 - j0.05$ ohms. The lengthening of the element increased the gain by 0.06 dB, but we ignore that small amount.

Although we may divide an inductor into two parts, we must use separate capacitors to place equal amounts of capacitive reactance on each side of the source. Let's revise the load to place a reactance of $-j51.4$ ohms on each side of the load, by using the segments just before and after the source segment.

```
LD 4 1 50 50 0 870.4
LD 4 1 52 52 0 870.4
Type Tag 1st-Seg Last-Seg Resistance Reactance
```

Now the source impedance is $89.25 - j0.32$ ohms.

Although the movement of the loading was small, the trend is evident. Just as placing inductive reactance farther outward on an element moved the source impedance closer to that of a standard dipole (relative to center loading), so too, placing a capacitive reactance farther outward on an element will also move the impedance closer to that of a standard dipole (again, relative to center loading). You can continue to move the capacitive reactance load outward to familiarize yourself with the trends in effects.

Here, the exercise is largely academic. However, the scheme of lengthening a driven element to raise the resistive component of its source impedance and then using capacitive loading to compensate for the resulting inductive reactance at the source has been used by at least one Yagi manufacturer in the past. The case involved a Yagi, whose driven element showed an impedance below 50 ohms. Lengthening and loading it provided the desired source impedance to match a 50-ohm coaxial cable feedline system.

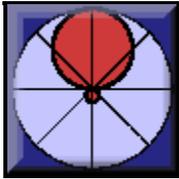
Other applications of intentional capacitive loading include the controlled current distribution antenna, described by WB9RQR and W9WQ in the *ARRL Antenna Compendium*, Vol. 2 (pp. 132-135). A less load-laden antenna using capacitive loading to control antenna characteristics is the modified extended double Zepp (also called the double extended Zepp) described by N6LF in the *ARRL Antenna Compendium*, Vol. 4 (pp. 78-80).

Most capacitors whose ratings are not exceeded by an antenna application have Q_s well in excess of 1000. Consequently, it is much safer to model capacitive loads without adding in a series R value

than it is inductive loads. However, in some instances, capacitor leakage cannot be ignored. In such cases, estimates of Q will result in series resistance values calculated just as for inductive reactances. However, in most cases, the values will be much smaller, since the Qs will still be quite high.

These exercises take us just about to the limit of resistance-reactance (Type 4) loads. Although easy to use--and often the first step in placing loads into amateur antennas--they suffer an important limitation. Real capacitors and inductors show a change of reactance as we change frequency, even slightly. Specified reactances maintain the value assigned. Consequently, they do not accurately represent element loading across a span of frequencies, even as narrow a span as a single amateur band.

In order to more accurately model the behavior of loads, we shall have to learn how to place series and parallel combinations of resistors, capacitors, and inductors as loads. In the process, we shall not only be able to model inductors and capacitors individually, but as well we shall be able to model traps and other tuned circuits placed within antennas. However, all that is for another segment of this series.



2. From Place to Laplace Converting X and R to C, L, and R



L. B. Cebik, W4RNL

Although complex resistance-reactance loads are handy for rapid antenna development, they are limited in application. NEC does not change the value of reactance as the frequency changes. However, real-world components with which we implement reactances do change their reactance value with frequency. If we intend to evaluate an antenna over a range of frequencies, we shall have to employ different kinds of loads that use actual capacitance and inductance values. We have two choices: series R- L-C circuits (type 0) and parallel R-L-C circuits (type 1). Since the series circuit most closely coincides with the complex R +/- jX loads we have been using, let's begin there.

A. Series (Type 0) R-L-C loads: Finding and Entering the Numbers

If we know a set of values for R, L, and C (or any one of them), we can go straight to the data entering process. However, if we have begun our design process with R +/- jX, then we must convert the reactance to a value of inductance or capacitance.

Although the equations for conversion are common handbook material, let's repeat them here for reference (in case you have a calculator handy):

$$X_L = 2 \pi F L \quad L = \frac{X_L}{2 \pi F}$$

where X_L is in Ohms, L is in Henries, and F is in Herz. Incidentally, NEC works almost exclusively in basic units for inductance and capacitance, so using the most basic equations is apt.

$$X_C = \frac{1}{2 \pi F C} \quad C = \frac{1}{2 \pi F X_C}$$

where X_C is in Ohms, C is in Farads, and F is in Herz.

Now we may turn to a definite example. We shall resurrect the 1/2-size #14 copper wire dipole in free space, with its 2 mid-element loads. For the moment, we shall let the resistance be zero and worry only about the load reactance, which was 870.4 Ohms.

Using the appropriate equation, we discover that this reactance, at the target frequency of 14.15

MHz, yields a required inductance of 9.79 microH. However, for modeling purposes, we should get used to specifying this and similar inductances in the computer form of engineering notation, hence 9.79e-06 H. (Likewise, although most HF capacitance values are in pF, for example, 23 pF, we need also to express them in a similar way, that is, as 23e-12 F. Notice that we insert no spaces in any part of the number.)

The following figure shows the antenna description in NEC input file format for the 20-meter dipole with the loads expressed in type 0 format:

```
CM 1/2 size 20m dipole with mid-loads
CE
GW 1 101 -8.434 0 0 8.434 0 0 2.67060367454068E-03
GS 0 0 .3048
GE 0
EX 0 1 51 0 1 0
LD 0 1 77 77 0 9.79e-6 0
LD 0 1 25 25 0 9.79e-6 0
LD 5 1 1 101 5.8001E7
FR 0 1 0 0 14.15 1
RP 0 1 360 1000 90 0 1 1
EN
```

Each of the load lines (ignoring the type 5 wire loading) follows the conventions just noted:

```
LD 0 1 77 77 0 9.79e-6 0
LD 0 1 25 25 0 9.79e-6 0
```

Type Tag 1st-Seg Last-Seg Resistance Inductance Capacitance

For type 0 loads, some of the entries have changed their meanings, specifically, the entries after resistance. Note that the order in which we enter the loads makes no difference to the calculation, although--except for examples like this one--I tend to enter all loads in some logical order. That practice makes modification easier, since I can more easily find the load I wish to modify and less easily select the wrong one. Following the order of the elements in the antenna and working from left to right or from End 1 to End 2 for each element generally works in most cases.

The inductance entry is in the notation we described. In some cases, exponents with leading zeroes will be truncated to a single digit; in others, the leading zero will appear. Where values are missing (we had no resistance and no capacitance in this example, enter zero. NEC knows to treat these as missing values, whether we are dealing with series or parallel loading circuits. NEC does not create inappropriate short circuits for missing values. The user has to do that by specifying very small values other than zero.

Besides allowing the user to enter a load line in the antenna description, NEC-Win Pro provides two other means for entering load data. One is the load screen accessed from the load line.

LD - Place Load

Place Load

Tag #:

Starting Segment #:

Ending Segment #:

Load Type

Wire Conductivity
 Series RLC
 Parallel RLC
 Complex Load
 Short All Loads

Wire Conductivity (mhos/meter):

R(Ohms):

L(Henrys):

C(Farads):

Units
 Units/Meter

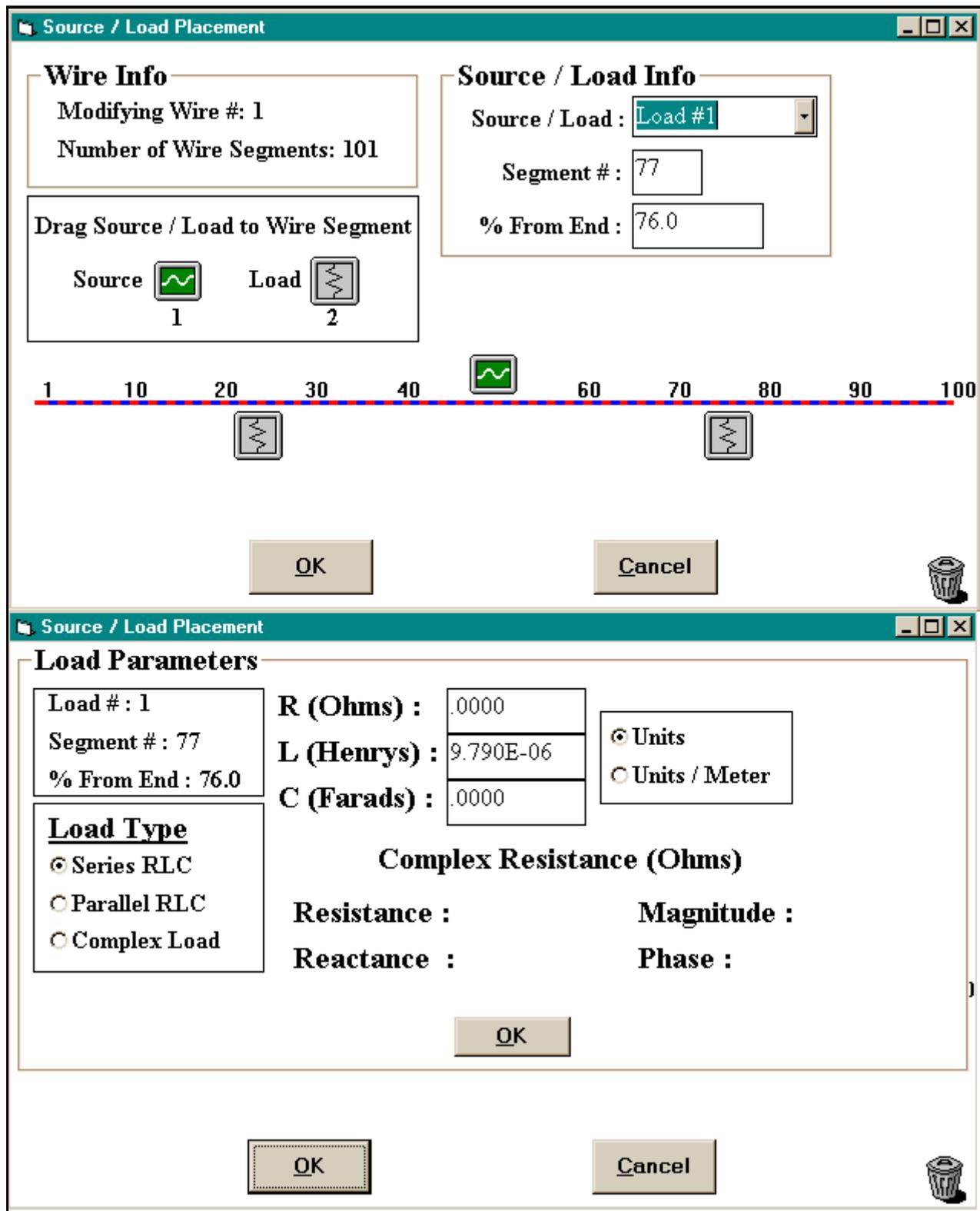
Complex Resistance (Ohms)

Resistance:

Reactance:

On this screen, we specified a Series RLC load and entered zeroes for the missing values and the inductance value. This screen produces the entry line for the load at segment 77.

Within the NEC-Win Basic model, calling up the drag-and-drop screen produces the following screens:



Although the data is the same, this screen set permits us to specify the load position as a percentage of element wire length.

Within EZNEC, entering R-L-C loads is a 3-screen sequence, as shown in the following partial replication of EZNEC screens.

SCREEN 1

-----Loads-----				
Load	Wire Seg	Wire #/Pct Actual	From End 1 (Specified)	Laplace Coefficients
1	25	1/24.26	(1/24)	Select to show values
2	77	1/75.74	(1/76)	Select to show values

(Enter position at wire end or center as 'w#e#' or 'w#c' if desired.)

Note: Loads are placed in series with sources, transmission lines, and other loads.

number=load to change Add loads Delete loads Group edit
change load Type (Laplace/Z)

<ESC> = Main Menu

SCREEN 2

-----Loads-----				
Load	Wire Seg	Wire #/Pct Actual	From End 1 (Specified)	Laplace Coefficients
1	25	1/24.26	(1/24)	Select to show values
2	77	1/75.74	(1/76)	Select to show values

Load 1: Coefficient 'S' or 'P'? S

S=Series R,L,C P=Parallel R,L,C

<PgUp>,<PgDn>=change load number <ESC>=return to main load menu

Load 1:	S^0	S^1	S^2	S^3	S^4
Num	0.000E+00	9.790E-06	0.000E+00	0.000E+00	0.000E+00
Den	1.000E+00	0.000E+00	0.000E+00	0.000E+00	0.000E+00

SCREEN 3

-----Loads-----				
Load	Wire Seg	Wire #/Pct Actual	From End 1 (Specified)	Laplace Coefficients
1	25	1/24.26	(1/24)	Select to show values
2	77	1/75.74	(1/76)	Select to show values

Load 1: Series R,L,C (Q,H,F)? 0,9.79e-6,0

<PgUp>,<PgDn>=change load number <ESC>=return to main load menu

* * * Enter zero for value of missing components * * *

Load 1:	S^0	S^1	S^2	S^3	S^4
Num	0.000E+00	9.790E-06	0.000E+00	0.000E+00	0.000E+00
Den	1.000E+00	0.000E+00	0.000E+00	0.000E+00	0.000E+00

Since EZNEC defaults to $R \pm jX$ values, we changed the load type. If we had been working from a previous model that had $R \pm jX$ loads, these loads would have been deleted. Therefore, we had to enter two new loads in the exact positions occupied by the old loads, as seen on Screen 1.

We begin the process of providing values for these loads on Screen 2, by selecting either a series or parallel type load.

On screen 3, we enter the values for the R, L, and C components of the load, using the engineering-computer notation, as shown.

Once we enter the values, they appear at the bottom of the screen as Laplace coefficients for the equations in the program that supply the NEC core with its input data. I have treated this example as if I were modifying existing values. Had there been no previous entry, all numerator (Num) values would have been $0.000E+00$. However, note that the inductance value appears as the numerator for S^1 .

For simple series loads, the Laplace coefficient for resistance will appear as the numerator for S^0 and the inductance will appear as the numerator for S^1 . In some complex loads, some of the coefficients will not be the same as the values entered. Therefore, it is a good practice to keep a record of the actual R, L, and C values entered for each model.

Now that we have our model, we might as well run it. EZNEC Pro, using NEC-4 returned a gain of 1.76 dBi and a source impedance of $26.08 + j1.65$ Ohms. NEC-Win Pro, running NEC-2 return values of 1.76 dBi and $26.09 + j2.13$ Ohms. In short, no difference.

B. The Significance of R-L-C Loads

Placing loads as series R-L-C circuits results within EZNEC in a number of cases that may initially mystify the user, since these loads are transformed and displayed in the load menu as Laplace coefficients. For the case of the "inductance only load" of $9.79e-6$ H, the coefficient lines looked like this (to S^2 only, since everything beyond that point is zero):

Load 1:	S^0	S^1	S^2
Num	0.000E+00	9.790E-06	0.000E+00
Den	1.000E+00	0.000E+00	0.000E+00

If we add an arbitrary 2 Ohms resistance in series with the inductor, we see the following table:

Load 1:	S^0	S^1	S^2
Num	2.000E+00	9.790E-06	0.000E+00
Den	1.000E+00	0.000E+00	0.000E+00

For inductors and resistors in series, we now know where to look for the values we inserted.

With capacitive loads, the table takes on a different look. Consider once more our slightly long dipole that we center loaded with a capacitor having a reactance of $-j102.8$ Ohms. If we convert the reactance to a capacitance at 14.15 MHz, the value required is 109.4 pF or $109.4e-12$ F. If we now plug this value into the EZNEC load screen, we see a table like this one:

Load 1:	S^0	S^1	S^2
Num	1.000E+00	0.000E+00	0.000E+00
Den	0.000E+00	1.094E-10	0.000E+00

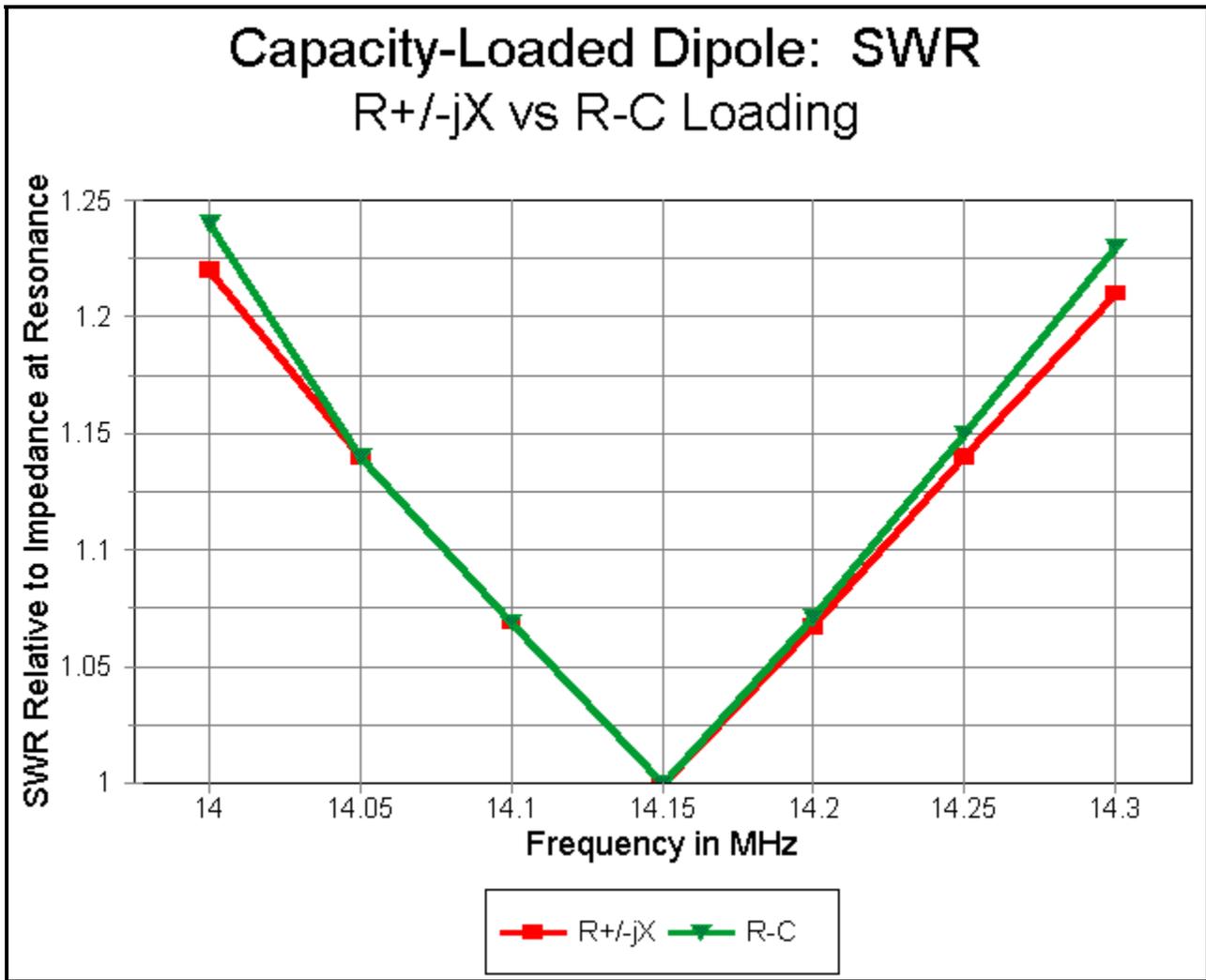
The capacitive value shows itself in the denominator line at S^1. Complex loads having resistance and capacitance are a bit more complex in appearance. Let us place an arbitrary 2 Ohm resistor in series with the capacitor. The coefficients now look like this:

Load 1:	S^0	S^1	S^2
Num	2.000E+00	2.188E-10	0.000E+00
Den	0.000E+00	1.094E-10	0.000E+00

The resistance is in the numerator line at S^0, while the capacitance value remains where it was. The new number in the numerator line at S^1 is the product of the two values. Incidentally, note that all values are transformed to the form x.xxx, with whatever exponent is needed to make the figure equal to the number entered.

The long dipole with capacitive loading is also interesting because it seems (and only seems) to suggest that it makes little difference whether we use a complex $R \pm jX$ or an R-L-C load when evaluating an antenna across a ham band. A frequency sweep of the capacitively loaded dipole from 14.0 to 14.3 MHz produces the following table of source impedance values for each type of load:

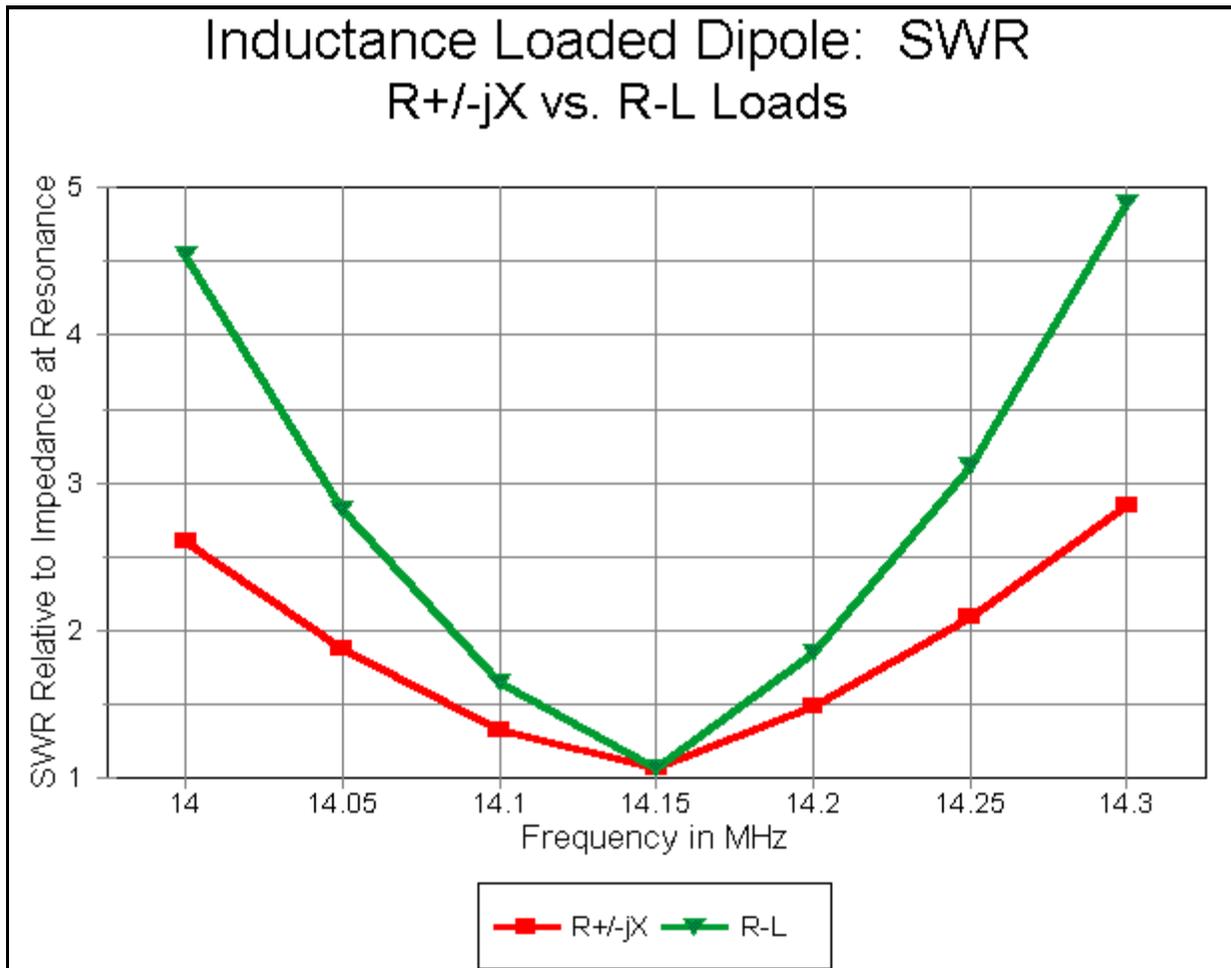
Frequency	Load=R+/-jX Source Impedance	Load=R-L-C Source Impedance
14.00	86.88-j17.46 Ohms	86.87-j18.58 Ohms
14.05	87.90-j11.64	87.90-j12.37
14.10	88.94-j 5.85	88.94-j 6.23
14.15	90.00+j 0.05	90.00+j 0.04
14.20	91.05+j 5.78	91.05+j 6.13
14.25	92.13+j11.70	92.13+j12.42
14.30	93.22+j17.48	93.22+j18.55



The SWR curves are hardly distinguishable, as one might expect from the table of source impedance values. However notice the steeper rise in reactance with the series R-L-C load, which provides the correct reactance of the loading capacitor for each frequency checked.

When the characteristics of an antenna vary more extremely across a band of frequencies, the trend becomes more noticeable and important. Consider our short, mid-element-loaded dipole once more. Just as an example, we shall use the model that does not introduce the series resistance that represent losses due to having a finite Q. The table below provides the source impedance values reported by each type of load, while the graph supplies comparable SWR data.

Frequency	Load=R+/-jX Source Impedance	Load=R-L-C Source Impedance
14.00	25.13-j25.45 Ohms	24.84-j42.26 Ohms
14.05	25.44-j16.46	25.24-j27.82
14.10	25.76-j 7.44	25.66-j13.14
14.15	26.08+j 1.65	26.08+j 1.65
14.20	26.41+j10.62	26.52+j16.47
14.25	26.74+j19.93	26.96+j31.77
14.30	27.08+j29.07	27.42+j47.06



Clearly, the R+/-jX loaded model provides an excessively optimistic report of the antenna operating bandwidth--by nearly a factor of 2. Once more, just to be repetitive, accurate modeling of a loaded antenna across a band of frequencies requires the use of R-L-C loads.

C. Parallel (Type 1) R-L-C Loads and Traps

Any series load having a series resistance as well as reactance can be converted into a corresponding parallel load (type 1). If we begin with series values of R and C or L, we must convert the C or L into a value of +/-jX. For the record, the conversion equations are the following. For series to parallel conversions:

$$R_P = \frac{R_S^2 + X_S^2}{R_S} \quad X_P = \frac{R_S^2 + X_S^2}{X_S}$$

where R_S and X_S are the series values of resistance and reactance and R_P and X_P are the corresponding parallel values of resistance and reactance.

For parallel to series conversions:

$$R_S = \frac{R_P X_P^2}{R_P^2 + X_P^2} \quad X_S = \frac{R_P^2 X_P}{R_P^2 + X_P^2}$$

with the same designations.

Parallel R-L or R-C circuits do not have much utility, since they are only the converted counterparts of series R-L and R-C circuits. However, for the exercise, you can convert the mid-element loads with a Q of 300 into their parallel counterparts, arriving at values of $R_p=261243$ and $X_p=870.41$ Ohms from the series values of $R_s=2.9$ and $X_s=870.4$ Ohms. The reactance, of course, yields a required inductance of $9.79e-6$ H

For NEC-Win Pro, we would simply select a parallel RLC circuit and enter the calculated values of R and L, with zero for the missing C. In EZNEC, we would similarly select a parallel RLC load and enter "261243,9.79e-6,0." In the coefficients table, we would see

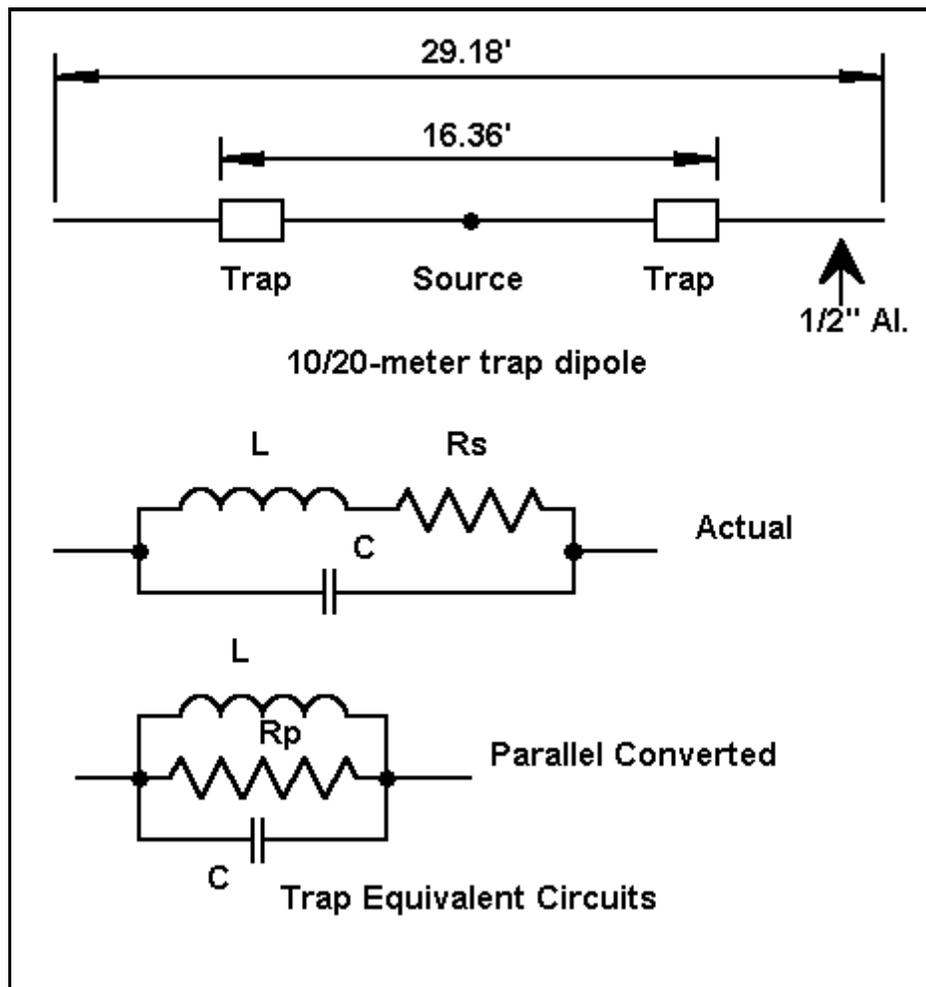
Load 1:	S^0	S^1	S^2
Num	0.000E+00	2.558E+00	0.000E+00
Den	2.612E+05	9.790E-06	0.000E+00

We recognize the values of R and L in the S^0 and S^1 positions of the denominator line, while the value in the S^1 position of the numerator is simply the product of the two.

To further establish the equivalence of appropriately converted series and parallel loads, here are sample results for the mid-element loaded short dipole run with both types of loads in NEC-4 within EZNEC Pro:

Frequency	Series RLC Model		Parallel RLC Model	
	Gain (dBi)	Source R+/-jX	Gain (dBi)	Source R+/-jX
14.00 MHz	0.92	30.07-j42.36	0.94	29.96-j42.38
14.15	0.93	31.55+j 1.52	0.93	31.55+j 1.52
14.30	0.94	33.13+j46.95	0.92	33.26+j46.95

The most useful application of parallel loads is with traps and similar RLC circuits. A trap is a resonant L-C circuit used to provide multiband service from a single element. There are several studies devoted to traps listed in the main index to these articles. Here we shall only cover certain essentials related to traps as computer modeling loads.



The figure shows a typical dual-band trap dipole for 20 and 10 meters, resonant on 28.5 and 14.175 MHz. The element material is 0.5" diameter aluminum. The traps are resonant on 27.75 MHz to equalize performance on the two bands.

The models for this consist of three wires. The first is 6.41' (76.9" or 1.95 m) long with 12 segments. The last wire is identical to the first. Between them is a wire 16.36' (196.3" or 4.99 m) long, with 31 segments. The first and last segments are loaded by parallel R-L-C circuits representing the traps.

The figure also shows trap circuit-equivalents. The upper schematic provides the basis for modeling a trap. The main losses in a trap are normally in the coil, given its finite Q. For this model, a Q of 200 was selected as being slightly more conservative than some commercial traps available. However, Qs from 100-400 have been successfully constructed.

The trap values are 1.2 microH and 27.4 pf to achieve resonance at 27.75 MHz. The reactance of each component is about 210 Ohms at resonance. With a Q of 200, the coil series resistance is about 1.05 Ohms. To create a true parallel circuit, we must convert the series inductive leg to its parallel equivalent. The inductance remains almost identical to its series value, but the equivalent parallel resistor becomes quite large.

However, we must do the conversions at other than the trap resonant frequency. At 28.5 MHz, the net reactance of the circuit is capacitive, and the required parallel resistance is nearly 43 kOhms. At 14.175 MHz, the net reactance is inductive and the required parallel resistance is about 21.3

kOhms. There is a program within HAMCALC and also within another trap article in this collection that will perform the requisite calculations, as well as explain the procedure in detail.

In effect, we have two sets of parallel R-L-C loads, one for 10 meters, on for 20 meters. (The amount that the parallel resistance changes within a band the size of 20 meters is too small to be significant, especially in view of the rounded figures used in most cases for coil Q, although one may calculate the resistor's value for each frequency of interest.)

Entering the parallel R-L-C circuit for which ever band interests us is done in the same way as for series loads, except that we select the parallel option and enter figures for all three types of components. The following figure shows a typical NEC-Win Pro entry screen for one of the loads for the 10-meter model.

With EZNEC, entry procedures are also the same as with a series load. The Laplace coefficient lines that result may seem mysterious at first sight:

Load 1:	S ⁰	S ¹	S ²
Num	0.000E+00	5.155E-02	0.000E+00
Den	4.296E+04	1.200E-06	1.413E-12

S⁰ in the denominator line is the resistance, while S¹ in the same line is the inductance. S¹ in the numerator line is the product of the two numbers. It would appear that S² in the denominator line must have something to do with the capacitance, but the value seems to have no relationship to the 27.4 pF entered. However, if we divide S²(d) by S¹(n), we arrive at our original capacitance. This is handy to keep in mind if you forget to record the capacitance you entered.

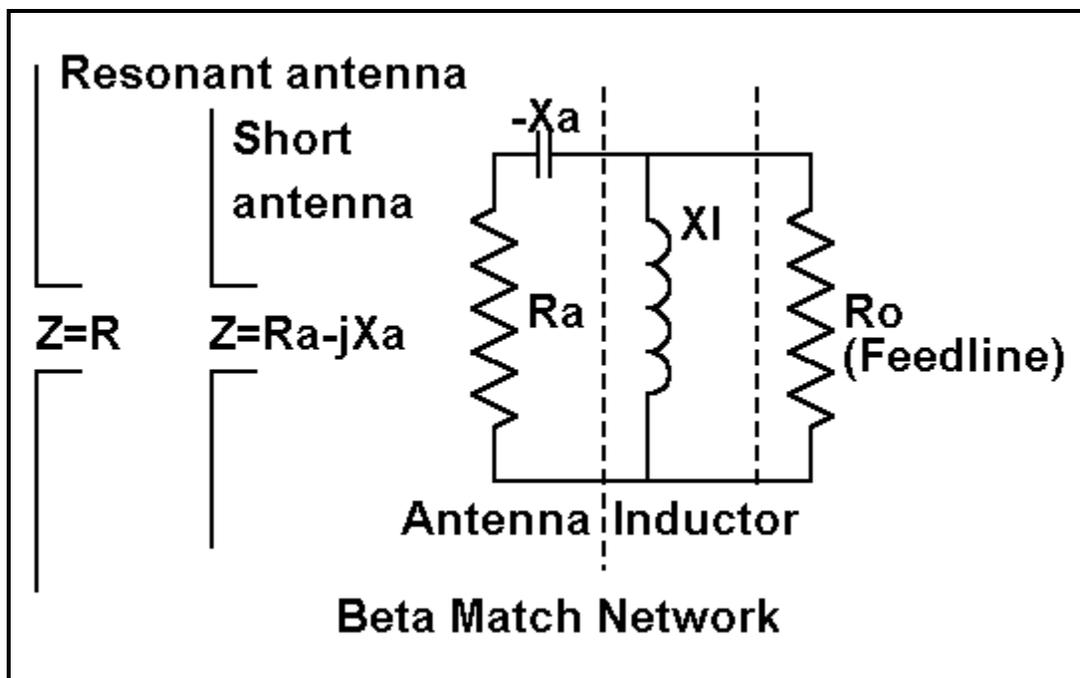
At 28.5 MHz, the model (in NEC-4) returns a gain of 2.08 dBi, with a source impedance of 85.69 + j0.02 Ohms. The same model using NEC-2 returns the same gain and a source impedance of 85.57 - j0.33 Ohms, establishing equal reliability in both programs for handling parallel R-L-C loads. At 14.175 MHz, the model with the appropriate parallel resistor yields a gain of 1.97 dBi with a source impedance of 63.16 - j0.63 Ohms. (Once more, these figures are excessively precise for practice, but are given to illustrate the numbers provided by the programs.)

As with all loads, the parallel R-L-C load is placed in series with sources and transmission lines. There is no direct provision for placing a load in parallel with sources or transmission lines within NEC.

D. R-X and R-L-C Loads in Parallel with the Source

Not all loads that we might use at the same segment as the source go in series with the source. One good example is the beta or hairpin match. Ultimately, this matching system places an inductance or inductive reactance in parallel with the source. The question is this: can we effectively model a beta match as part of the antenna? The answer is a qualified "yes."

The beta match is a modified L-network for transforming a higher (feedline) impedance to a lower (antenna) impedance. An L-network consists, when transforming to a lower impedance, of a shunt or parallel reactance on the high impedance side and a series reactance of the opposite type on the low impedance side.



As the figure shows, the modification consists in letting the antenna feedpoint impedance supply the series reactance and then installing only the shunt reactance. Normally, we make the antenna a bit shorter than resonant to provide a capacitive series reactance and then add a shunt inductive reactance. The inductive reactance may be in the form of an inductor or a shorted transmission line length (the hairpin). However, for special cases, we might also make the antenna long, providing a series inductive reactance and then install a shunt capacitive reactance.

Beta values are easily calculated, either from the following equations or from a program such as found in the HAMCALC collection. Let's sample the normal case, letting the series reactance be capacitive and the shunt reactance be inductive.

The common factor in L-networks is delta, which may be called the working Q or measure of efficiency of the network.

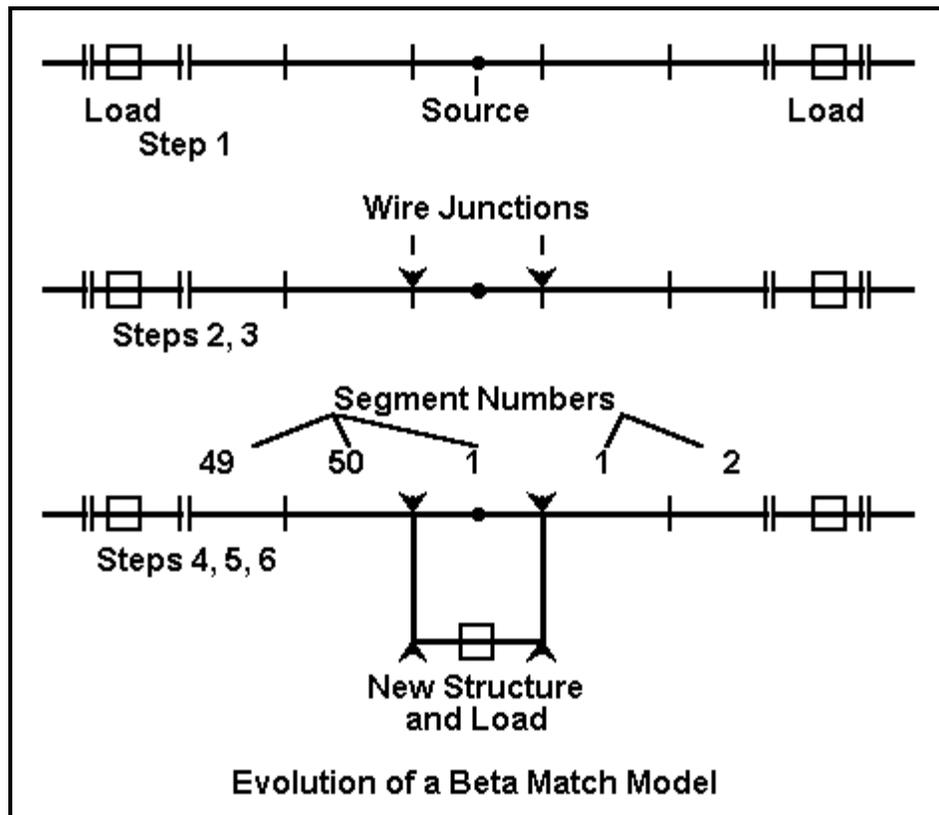
$$\delta = \sqrt{\frac{R_o}{R_a} - 1}$$

From delta, we can calculate the requisite values of X_a (the series reactance at the antenna feedpoint) and X_l (the shunt inductor).

$$X_a = \delta R_a \quad X_l = \frac{R_o}{\delta}$$

Now let's return to one of the antennas in the first episode of this series. Consider the half-length 20-meter dipole with mid-element loading. This model used a single wire of 101 segments 16.868' long. The mid-element loads were on segments 25 and 77 and had values of 870.4 Ohms. The source impedance of this antenna was $26.08 + j1.65$ Ohms.

The transformation from 26 Ohms to 50 Ohms for coax feed is a good job for the beta match. Using the equations above, we find a delta of about 0.95, resulting in an X_a of 24.97 and an X_l of 52.7. Now, let's model this antenna, working toward the beta match.



The figure indicates several steps we need to take in order to ensure an adequate model of the antenna and its beta match. Step 1 we have already taken in modeling the original antenna. The gain, using lossless mid-element loads in this exercise, is 1.76 dBi.

Step 2 consist of preparing the model for a parallel load. Break the single antenna wire into 3 parts, the two outer parts having 50 segments each, with the loads installed where they physically had been in the original model. This would place them on segment 25 (49%) of the left wire and segment 26 (51%) of the right. At the center is a single segment wire the length of the segment that it replaced (about 2" in this example). When we run this model, we obtain a feedpoint impedance of 26.08 +j1.68 Ohms and a gain of 1.76 dBi, establishing that we have not made anything but a cosmetic change in the model.

Step 3 will establish the antenna shortening in order to provide the desired series source capacitive reactance. Here, we have two choices. We may physically shorten the outer wires, or we may simply reduce the mid- element loads a bit. Let's do the latter. Lowering the mid-element load reactances to 857 Ohms each produces a feedpoint impedance of 25.63 - j23.20 Ohms.

Step 4 requires that we add to the physical structure of the antenna. We add 3 1-segment wires to form a box around the center source wire. On the wire parallel to the source wire, we initially install a resistive load of 10 MOhms. This resistive load produces what is essentially an open circuit in the new structure so that we may check the effects, if any, of the structure on the antenna performance. Under these conditions, we obtain a feedpoint impedance of 25.97 -j23.55 Ohms, and an antenna gain of 1.70 dBi.

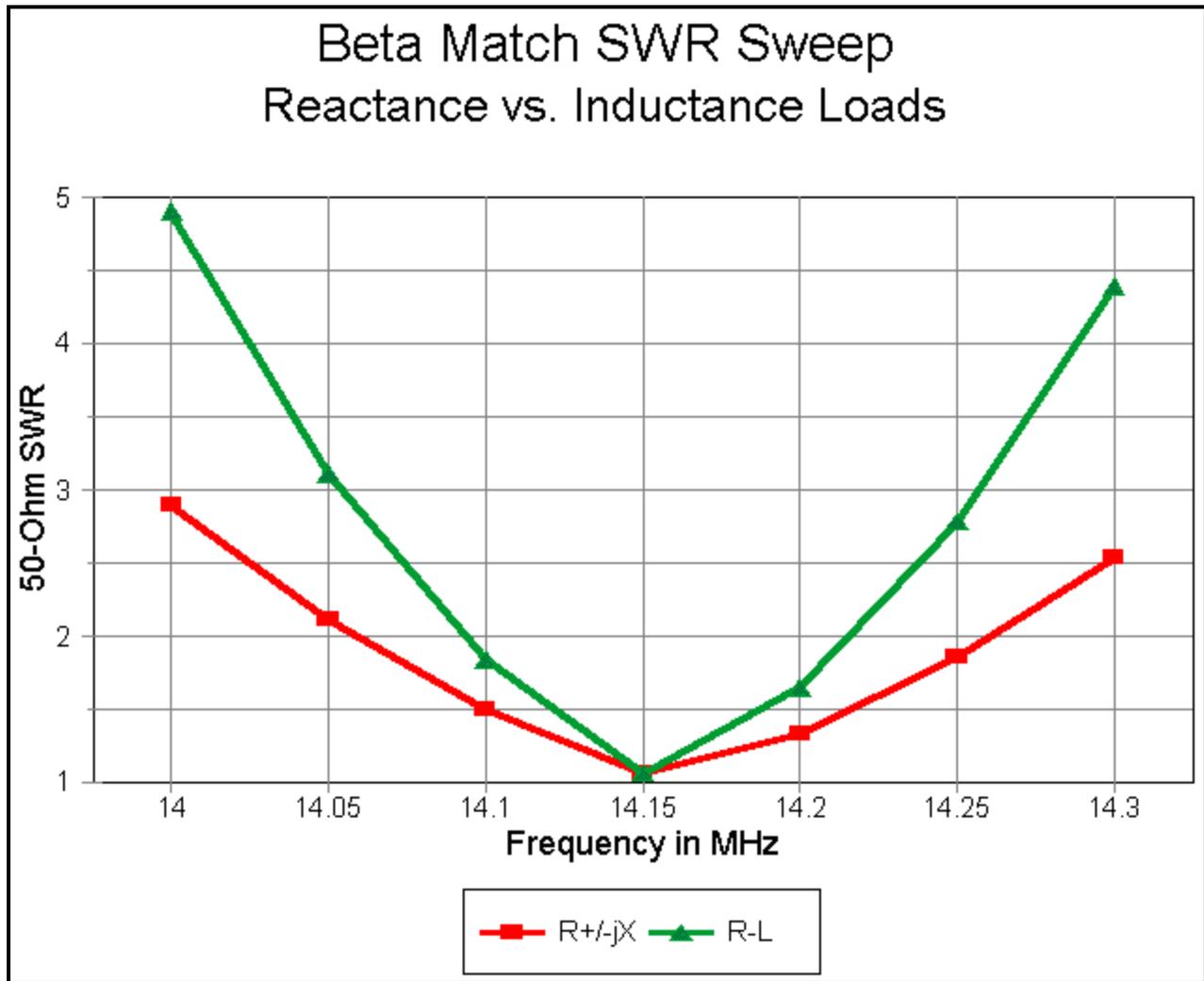
Although these changes are small, they do indicate that the structure is not insignificant. Had we used fewer segments, each would have been longer, and the box wires would have had greater effects on performance of the model. Reversing the positions of the source and the resistive load provide another indication that the structure is not insignificant: the feedpoint impedance and the gain change by very noticeable amounts. In practice, we tend to ignore small lumps and bumps in linear antenna element structure, but careful modeling suggests that their cumulative effects on performance may be more than trivial.

Step 5 consists in replacing the resistive load with the reactive load indicated by our work with the equations. Placing a purely inductive reactance of 52.7 at the parallel load point yields a source impedance of 53.28 - j0.10 Ohms. Using a very small structure to model the parallel load produced results very close to predictions from the equations.

Step 6 takes into account the Q of the beta coil. Let us assume a small inductor with a Q of 300 and hence a series resistance of 0.18 Ohms. This small move produces a source impedance of 53.13 + j0.12 Ohms, all the while making no change in the antenna gain of 1.70 dBi in free space.

If we wish to check the performance of the beta matched antenna away from the design frequency (14.15 MHz), we must make one further move: change the load type from a Type 4 (R +/- jX) load to a type 0 (series R-L-C) load. The resistance remains at 0.18 Ohm, while the requisite inductance is 0.59 microH. At the target frequency, the source impedance is reported by NEC as 52.91 + 0.13 Ohms.

The importance changing to a series R-L-C load becomes apparent the moment we frequency sweep the completed antenna model from 14.0 to 14.3 MHz.



As the graph shows, assuming that the reactance of the coil will remain "close enough" to the value at the design frequency produces an overly optimistic curve of SWR. Using a series R-L-C load for which NEC provides the correct reactance at each frequency checked shows a much narrower (and more correct) operating bandwidth for the antenna. The difference is more than slight.

As the exercise demonstrates, it is possible to model loads in parallel to a source using the smallest feasible structures and paying close attention to details along the way. Among those details are deviations from the modeled performance without the added structure and an evaluation of the degree to which those deviations also apply to the real-world antenna. Equally important is using the right load type to model antenna performance across the relevant band of frequencies.



Although we have not fully covered all of the possibilities for using loads in NEC antenna models, perhaps we have gone far enough to let you continue on your own. If that is true, then we can turn our attention to a related subject, the entry and use of transmission lines in NEC models.





3. Transmission Lines as Lines

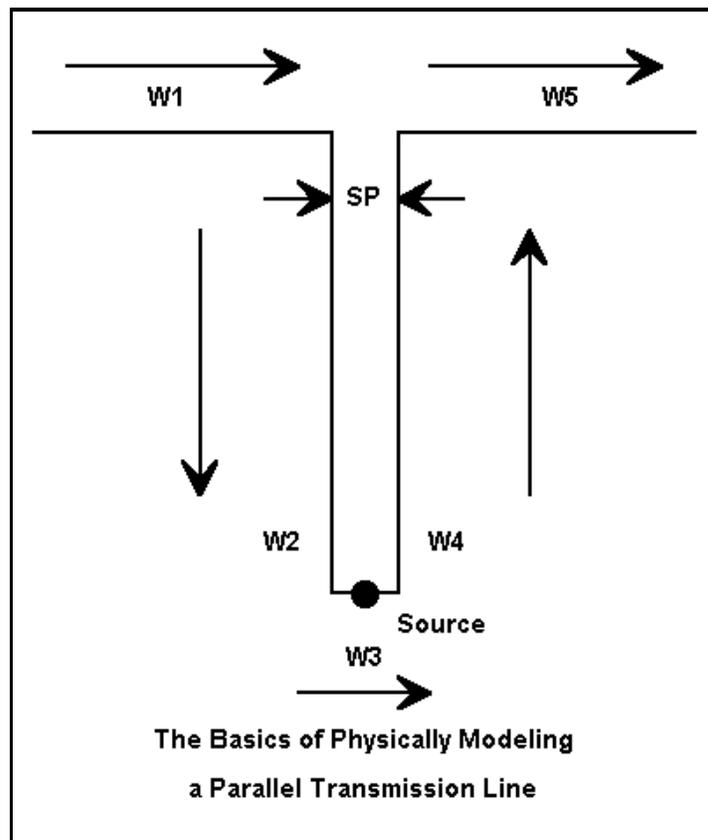
L. B. Cebik, W4RNL



Modeling transmission lines between the source and the antenna can take two forms. We can physically model parallel transmission lines. And we can model them as TL inputs. Each method has its advantages and limitations.

A. Physical Models of Parallel Transmission Lines

Consider a simple antenna fed at the center. 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, etc. 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 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.

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 segments 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 and save from 30 to 130 segments per 1/4 wavelength of each side of the transmission line).

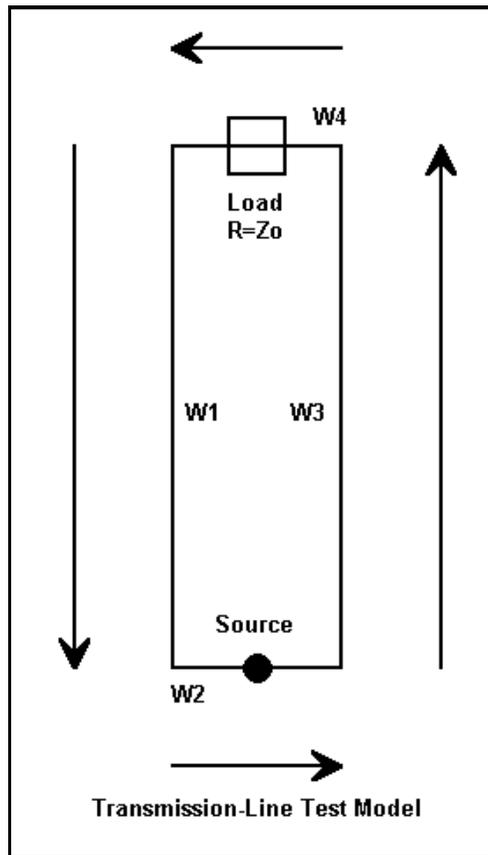
Physical transmission lines are usually calculated according to one of two equations:

$$Z_o = 276 \log \frac{2S}{d} \text{ or } Z_o = 120 \cosh^{-1} \frac{S}{d}$$

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.

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. The figure immediately below shows the test modeling arrangement. 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:

$$Z_0 = \sqrt{R_L Z_s}$$

where Z_0 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 Z_0 Calculated	Wire Dia.	Spacing (inches)	Modeled Z_0 in Ohms	New Wire Size for Z_0
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 segments 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.

B. The NEC TL Card

NEC-2 and up provide an alternative (TL input) means of modeling transmission lines of any type. However, the models are mathematical and not physical. By physical, I mean any portion of the antenna structure for which mutual impedances, currents, and fields are calculated. Although the terminals of transmission lines are real and interact with other parts of the model in regular ways, the line itself is not treated as a physical element. Hence, radiation from transmission lines that are handled by this alternative TL input means is not accounted for by the model.

In exchange for this limitation, the modeler acquires the ability to place one or more transmission lines into the model. The lines may have any Z_0 and any velocity factor. Moreover, the transmission lines do not add to the total wire and segment count. For a given antenna structure, core runs will be faster. Alternatively, where there is a limit to the number of segments allowed by a program, more segments are available for detailing the antenna structure itself.

Transmission lines inserted via the TL input will be treated as lossless. Consequently, if line losses are significant to the analysis, they will have to be separately calculated via a program such as Dean Straw's TL.

```

CM 20-meter dipole with transmission line
CM Source on wire 2 at end of TL
CE
GW 1 15 -17 0 70 17 0 70 2.67060367454068E-03
GW 2 1 -.1 0 1 .1 0 1 2.67060367454068E-03
GS 0 0 .3048
GE 1
GN 2 0 0 0 .002 14 0 0 0 0
EX 0 2 1 0 1 0
LD 5 1 1 15 5.8001E7
LD 5 2 1 1 5.8001E7
TL 1 8 2 1 50 10.6 0 0 0 0
FR 0 1 0 0 14.15 1
RP 0 181 1 1000 -90 90 1 1
FR 0 1 0 0 14.15 1
RP 0 1 360 1000 62 0 1 1
EN

```

The figure shows the NEC input file for a simple 20-meter dipole with a transmission line running from the antenna's normal feedpoint to another point used as the source. The terminating point of a transmission (as well as the starting point) must be a wire. When dealing with transmission lines used to feed an antenna, the "other wire" may be a very short single segment wire. Wire 2 in the sample is the terminating wire for the transmission line. It is only 0.003 wavelength long. The reason for keeping such wires very short is to ensure that they are minimally affected by the main antenna field and likewise minimally affect that field. For most amateur purposes, keeping the current on the secondary wire to 0.1% of the highest level of current on the antenna structure is sufficient to safeguard the accuracy of the reported antenna performance results.

The actual transmission line is specified on the line marked TL. The significance of the entries is as follows:

```

TL 1 8 2 1 50 10.6 0 0 0 0
  1st End Seg. 2nd End Seg. Zo Length Real Imag. Real Imag.
  Wire # # Wire # # Ohms meters Adm. Adm. Adm. Adm.
      W1 shunt W2 shunt

```

The input entry tells us that the transmission line runs from segment 8 (the center) of wire 1 to segment 1 of wire 2. The Zo is 50 ohms. If the entry had been -50, the negative sign would indicate to the program to cross or reverse the terminals at the far end. The length is 10.6 meters, about 1/2 wavelength at the test frequency of 14.15 MHz. Note that there is no provision on the input line for a velocity factor. Any input system that permits entry of a velocity factor pre-calculates the length of an equivalent line with a velocity factor of 1.0 and enters that figure in the length place.

The last four places are for shunt admittances containing conductivity and susceptibility factors for ends 1 and 2 of the transmission line. We shall not use them in this exercise.

Note also that the distance between the two wires specified in the GW, or wires, entries is 69 feet. However, the length of the transmission line used in calculations will be the one specified in the TL card unless a zero is entered for length. In that case, the program will use the straight-line distance between the two terminating points of the transmission line as the line length. This latter entry capability is useful for placing straight line phasing lines between fixed elements and similar applications.

C. Entering Transmission Line Data

Commercial implementations of NEC use a variety of methods for entering transmission line data. Of course, the TL input may be entered in the editor by directly typing in the requisite information.

NEC-Win Basic and the Basic module of NEC-Win Pro use a special window to simplify data entry:

Transmission Line

ptions

No Transmission Lines Defined

Place Transmission Line

From : Wire # : 1 Segment # : 8

To : Wire # : 2 Segment # : 1

Shunt Admittance

Real Imaginary

End One : 0 0 mhos

End Two : 0 0 mhos

Characteristic Impedance (ohms) : 50

Transmission Line Length (m) : 10.6

Transmission Line Type

No Phase Reversal

180° Phase Reversal

Previous Done Next

Although the label says, "No Transmission Lines Defined, this window shows one in the process of being defined. The terminating points are defined in the upper left, while the Zo and Length are immediately below. Note again that there is no provision for a velocity factor: the line length for a VF of 1.0 must be pre-calculated by the user. In the lower left is an entry to specify phase reversal or normal connections, with the unused shunt entries immediately above.

TL - Transmission Line

Place Transmission Line

From Tag # : 1

To Tag # : 2

Segment # : 8 Segment # : 1

Shunt Admittance

Real Imaginary

End One: 0 0 mhos

End Two: 0 0 mhos

Characteristic Impedance (ohms): 50

Transmission Line Length (m): 10.6

Transmission Line Type

No Phase Reversal

180° Phase Reversal

OK Cancel

NEC-Win Pro also accesses a similar window from the TL line of the antenna description. Although styled slightly differently, the data entry scheme is identical to the one just described.

EZNEC Transmission Line Screen Transitions
Note highlighted entries across screen

Line	Wire #/% Actual	From End 1 (Specified)	Wire #/% Actual	From End 2 (Specified)	Length	Z0 Ohms	Vel Fact	Rev/ Norm
1	1/50	(1/50)	2/50	(2/50)	34.775	50	1.00	N

number = t.line to change **Add** t.lines **Delete** t.lines **Group** edit
<ESC> = Main Menu

1	1/50	(1/50)	2/50	(2/50)	34.775	50	1.00	N
---	-------------	----------------	------	---------	--------	----	------	---

Wire number, % of way from end 1; (o)pen; or (s)hort
(Enter position at wire end or center as 'w#e#' or 'w#c', if desired.)
Note: transmission lines are placed in parallel with sources and other
Transmission lines. They are placed in series with loads.

1	1/50	(1/50)	2/50	(2/50)	34.775	50	1.00	N
---	------	---------	------	---------	---------------	----	------	---

Trans. Line length (see below for options)?
(Length in ft, or add 'd' for length in electrical degrees)
Or enter 'a' for length = straight line distance between ends.

1	1/50	(1/50)	2/50	(2/50)	34.775	50	1.00	N
---	------	---------	------	---------	--------	-----------	------	---

Transmission line Z0 (ohms)?
[User may select specific cables from a list at the screen bottom, as
well as entering in a set of values. Entry controls both Z0 and
Velocity Factor. User may also enter 'A' to specify the wire diameter
and spacing of a user-made parallel line.]

1	1/50	(1/50)	2/50	(2/50)	34.775	50	1.00	N
---	------	---------	------	---------	--------	----	-------------	---

Transmission line velocity factor (>0; 1 max)?

1	1/50	(1/50)	2/50	(2/50)	34.775	50	1.00	N
---	------	---------	------	---------	--------	----	------	----------

(N)ormal or (R)eversed connections?

replications in the figure take you through the transitions as you move from one item to the next in creating a transmission line. The top section shows the complete data entry, ready for revision, if one chooses to do so. The remaining partial screen section change as the user highlights one item after the other.

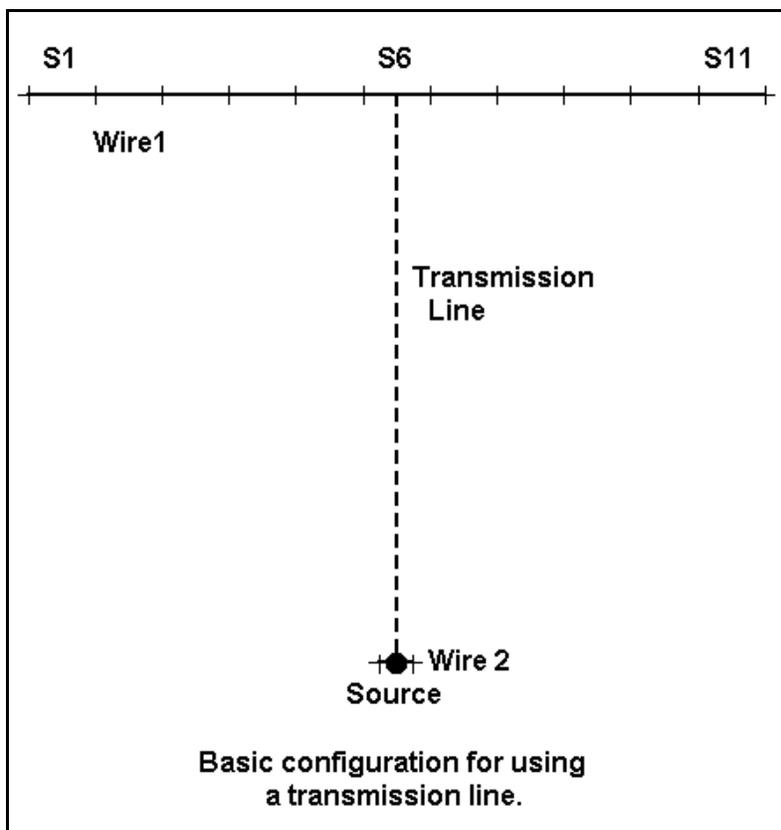
The second screen appears when either end of the transmission line is illuminated. The next screen highlights the length, which in EZNEC is entered in the same units as the wire dimensions. In this case, it is feet. If the user specified the length with a "d" suffix, the entry is interpreted as electrical degrees, which is then calculated into meters for the NEC core. Be careful not to use electrical degrees for a frequency sweep designed to show the properties of a fixed length of transmission line over an amateur band. An entry of "a" for the length will use the straight-line distance between the specified wire segments as the line length.

When either the Zo or the velocity factor are illuminated, the user has the option of selecting a transmission line from a table of both coaxial cables and parallel lines. Alternatively, the user can enter the required values. Finally, the user can construct an air-insulated parallel line by supplying the wire diameter (in measured units or as an AWG value) and the wire spacing.

The normal-reverse flag is activated in EZNEC by specifying N or R as the final line entry.

D. Basic Transmission Line Applications

The most basic transmission line application is as a feedline from a power source to the antenna, as shown in the figure.



Although the figure shows 11 segments, the antenna wire can have any (odd) number of segments (if a true center connection is desired). The second wire is very short and has one segment. The transmission line goes from the center of one wire to the next.

Consider a 20-meter dipole similar to the one in the figure, but with 101 segments of #14 copper wire. The total length is 33.696' (5.141 m). The second wire is 0.02' (0.006 m) long and has 1 segment. (This length at the test frequency of 14.15 MHz is about 0.0003 wavelength long, while the recommended NEC-4 minimum wire length for such applications is about 0.0001. For NEC-2, the wire should be at least 0.001 wavelength long.)

The NEC-4 source impedance of the antenna alone is $73.06 - j0.01$ ohms. Now let's do a familiarization exercise based on the fact that a lossless feedline repeats the antenna source impedance every multiple of $1/2$ wavelength. First, we move the source from the long antenna wire to the short second wire. Then, we specify a 50-ohm, $VF=1.0$ transmission line between the center of the antenna (the former feedpoint or segment 51) to the center of the second wire.

Since TL-type transmission lines are in parallel with sources, the effect of placing both the source and the transmission line on the second wire is to place a source in series with each wire of the transmission line, just as we would model the system physically. Except for the fact that the TL-type transmission line will not even show copper wire loss, the results will be identical for physical and TL transmission lines.

For example, let us assign to the transmission line a length that is exactly 180 electrical degrees. The resulting reported source impedance will be $73.06 - 0.08$ ohms in NEC-4. (As with all models in the exercises in this series, minor adjustments will have to be made for NEC-2, adjustments that are only apt to yielding the excessively precise outputs used for numerical comparison, but which have little or no practical application to real-world antennas.)

EZNEC permits the specification of transmission line lengths in electrical degrees (and automatically specifies a velocity factor of 1.0). For other programs, we may have to calculate precisely the length of the transmission line that is exactly 180 electrical degrees long at the test frequency. At 14.15 MHz, this length is 34.7552' or 10.593 m. Using either of these values, the source impedance reported is the same as just noted.

Now let us substitute a truer 50-ohm coaxial cable, perhaps RG-213, with a velocity factor of 0.66. The requisite length of the cable is simply 0.66 times the value used with the assumed velocity factor of 1.0, or 22.9384' (6.9914 m). Using this transmission line, the reported source impedance is $73.06 - j0.00$ ohms.

A real coaxial cable, even brand new, will not deliver lossless performance. If we use a program like TL, we shall discover that RG-213 at the specified length of nearly 23' (nearly 7 m) with an antenna impedance of $73.06 - j0.01$ ohms will show that the source end an impedance of $71.91 - 0.03$ ohms. This value represents the apparent SWR decrease (from 1.46 to 1.42) due to line loss with which most antenna users are familiar. The line also creates a 0.195 dB power loss, which amounts to 4.4%. The loss is a combination of basic line loss plus the additional loss due to the SWR level.

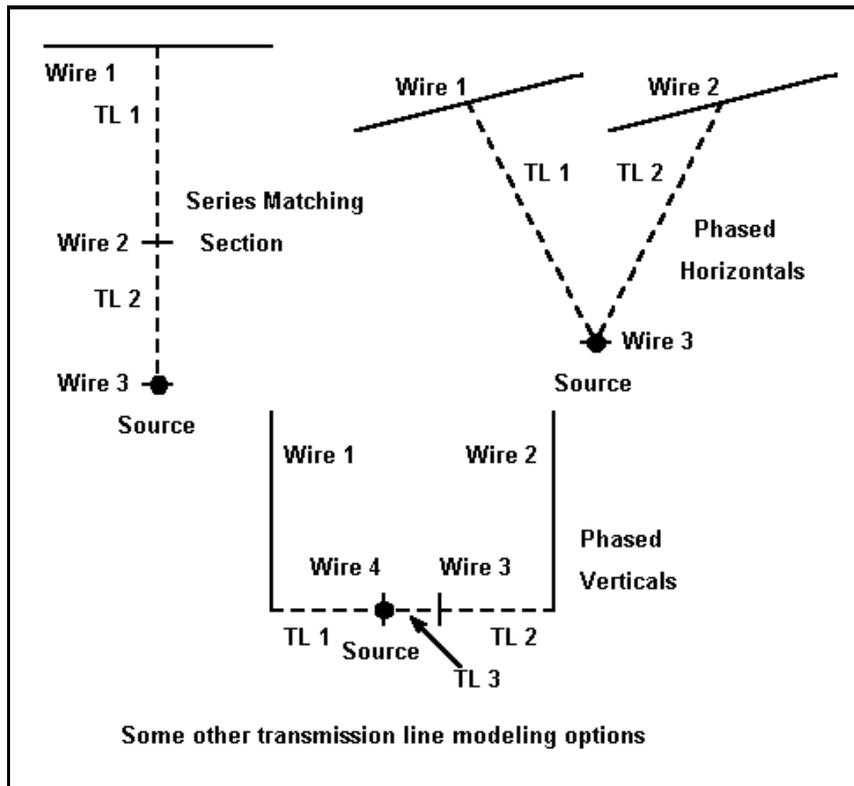
There is no convenient way to incorporate this loss into the antenna wire at either end of the model. Nor should we want to. Whatever the power delivered to the antenna, its gain and other characteristics remain constant (assuming that material junction losses can be neglected). If the losses and gains within an overall antenna-plus-transmission line-plus- source system are important, they can easily be calculated externally to modeling the antenna itself.

This exercise has little practical value other than familiarizing the modeler with the placement of TL transmission lines within a model. The model used here can be extended by changing the length of the transmission line in increments of 10 electrical degrees and observing the change in source impedance across a full half-wavelength. One might also use 300-ohm and 450-ohm transmission lines in order to observe the greater excursions of resistance and reactance with each 10-degrees of electrical length. All of these are familiarization exercises, designed to develop a set of reasonable expectations for models employing TL transmission lines.

In each case, we must remember that the values given are for a lossless line. The TL transmission line will return the same value of source impedance for 0, 180, 360, and 540 electrical degrees, or multiples of 34.7552' (10.593 m) for velocity factors of 1.0. For comparison, run each length through a program such as TL to determine the losses for each length and the anticipated impedance presented at the source end of the line.

E. Sample Practical Applications

There are many practical combinations of antenna elements plus feedline(s) for which modeling is very apt. The figure below illustrates only three of them.



At the upper right is a pair of horizontal antenna elements spaced some distance apart. The spacing might be anything from 1/8 to 1/2 wavelength. We may feed the antennas either in phase or out of phase. To explore the various situations, we need only create a third wire. Then we create transmission lines from wire 1 and from wire 2 to wire 3, where we also locate the source. In order to reverse the phase of feeding, we simply specify one of the two transmission lines as "reversed" or "180-degree phase reversal," which enters a minus sign in front of the Zo we specify for the two lines. We can even experiment with different Zos for the transmission lines and different lengths. The models, even though subject to slight inaccuracies due to the lossless nature of the transmission lines, will teach much about phasing lines with a common feed point.

At the bottom is the KB8I phasing system for feeding two 1/4 wavelength verticals that are spaced 1/8 wavelength apart. TL1 and TL2 are 50-ohm coaxial cables 157 degrees long, each terminating with a wire in the model. Between the wires is another cable, TL3, which is 29 degrees long. To create this model, we might place the verticals over perfect ground. If we wish to model over real ground, we would elevate the verticals and their ground planes for NEC-2. In either event, wires 3 and 4 should be elevated. There are methods for calculating the phasing lines to achieve a desired current phasing of -135 degrees for one of the verticals with the other at zero degrees, giving a broad, nearly cardioid pattern to the array. Of course, we can replicate Al Christman's dual directional array switch simply by moving the source from wire 3 to wire 4 and back again. For this array, which permits a direct 50-ohm coaxial cable feed from the model source to the shack, it is wise to look at the current magnitude and phase at the base of each antenna as well as the overall current and source impedance. You might even wish to experiment by replacing the 50-ohm cable with 75-ohm cable, just to see what happens. Be sure to remember to calculate for the velocity factor when translating the cable lengths in electrical degrees into feet or meters.

At the upper left is an interesting system for improving the operating 2:1 SWR bandwidth of an antenna-feedline system for dipoles. This can be operationally important to amateurs using gear with sensitive SWR shut-down systems with certain antenna situations. Consider a half-wavelength center-fed dipole for 40 meters, which we might model as #14 copper wire 67.2' long with 101 segments. Let us place this antenna at a height of 50' above medium earth. We would obtain the following values of source impedance and 50-ohm SWR from NEC-4:

Freq.	7.0	7.05	7.10	7.15	7.2	7.25	7.3
R	83.28	84.88	86.44	88.07	89.72	91.38	93.06
X	-32.5	-21.3	-10.1	-0.9	12.0	23.0	34.0
SWR	2.03	1.85	1.76	1.76	1.84	1.99	2.19

Even allowing for the lower SWR that appears at the shack end of the line due to line losses, the band edge values may activate power reduction circuitry in some gear. As a step toward resolving the problem, let us create a short wire spaced well away from the dipole. From the antenna, run a 50-ohm coaxial line exactly 1/2 wavelength long at the band center, 7.15 MHz. RG-213, with a velocity factor of 0.66, would be 45.40' long for this application. Be sure to move the source to the new wire (wire 2 in the diagram). Because the transmission line is less than 1/2 wavelength long at 7.0 MHz and longer than 1/2 wavelength long at 7.3 MHz, we obtain the original feedpoint impedance only at the band center, 7.15 MHz. At other frequencies across the band, we obtain the following values for the lossless line:

Freq.	7.0	7.05	7.10	7.15	7.2	7.25	7.3
R	90.09	87.80	87.13	88.07	90.58	94.75	100.8
X	-27.1	-17.3	-8.0	0.9	9.5	17.9	26.0
SWR	2.03	1.85	1.76	1.76	1.84	1.99	2.19

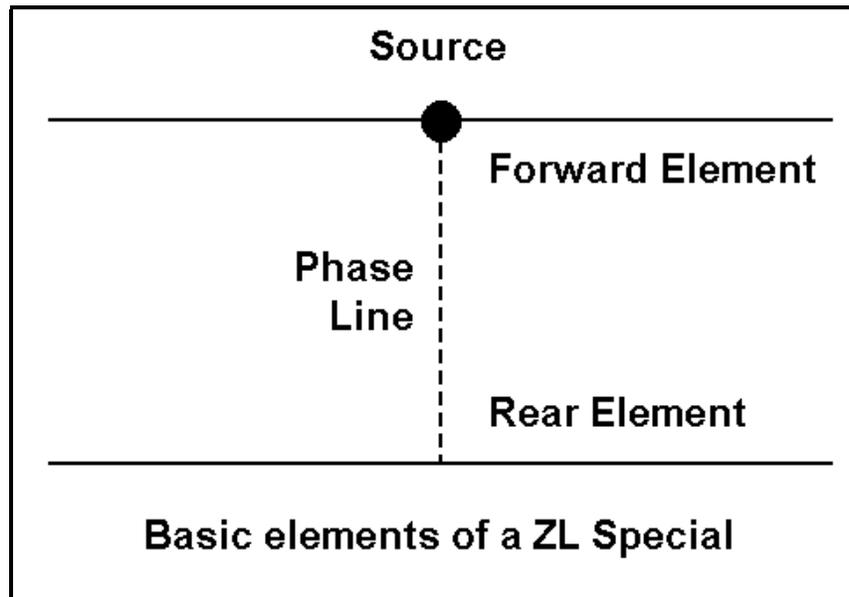
Note that the SWR has not changed, but the impedance values have, except for mid-band. The resistive component has gone up, and the reactive component has gone down. In fact, the values are sufficiently higher than 50-ohms to make a 1/4 wavelength 75-ohm matching section useful to transform them to lower values. So let's add a new short wire (wire 3) and move the source to it. Then let's create a 75-ohm, 0.66 velocity factor transmission line from wire 2 to wire 3. The line will be 22.70' long. From it--again allowing for the fact that the model uses lossless line--we obtain the follow source values:

Freq.	7.0	7.05	7.10	7.15	7.2	7.25	7.3
R	56.45	61.26	63.92	63.86	61.31	56.96	51.72
X	16.0	11.5	5.6	-0.7	-6.2	-10.1	-12.2
SWR	1.38	1.34	1.30	1.28	1.26	1.26	1.27

The result is a set of SWR values across 40-meters that should satisfy any piece of amateur equipment. The line losses at any point in the band may be calculated from TL and compared to the use of a single 50-ohm cable or to the use of a single 75-ohm cable with impedance transformation (and network losses) at the shack. There is not likely to be a significant difference.

For developing the general picture of what occurs all across the band with this system of matching, modeling both the antenna and the transmission lines provides one of the most efficient methods. I am indebted to Dave Leeson, N6NL, for bringing the matching system to my attention. Dave attributes it to Frank Witt, AI1H and other sources. All in all, it provides a good exercise in seeing the modeling capabilities of NEC in handling multiple lengths of transmission line on the way from the source to the antenna.

Not all transmission lines must have a short wire at one end. Consider the following sketch of a ZL Special. Pioneered, although apparently not invented, by George Prichard, ZL3MH (later ZL2OQ), the antenna consists of two elements spaced about 1/8 wavelength apart. The feedpoint is the junction of the forward element and the phase line (transmission line) to the rear element.



Let's build one of these antennas as a model. For 28.5 MHz, we might choose 3/4" diameter aluminum tubing and make two identical elements, each 16' long with 51 segments. We shall space them 3.46' apart, which is just about 0.1 wavelength. We shall set one element as the forward element and place the source at its center. We create the phasing line by placing a transmission line between segments 26 of each of the two elements. Let us follow Prichard's lead and use 71-ohm parallel line, which had a reported velocity factor of 0.67. Let's be sure to place the traditional half twist in the line by reversing the phase. If we place the antenna at 35' above medium earth, NEC will return a gain of about 11.8 dBi, a front-to-back ratio of about 25.7 dB, and a source impedance of about $7.2 + j8.8$ ohms. (The low source impedance did not bother Prichard, since he fed his arrays with parallel feeders and an antenna tuning unit.) These facts are perhaps the least interesting in connection with the ZL Special.

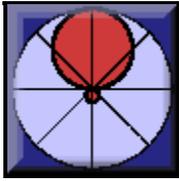
Now let's examine the current tables available as an output from NEC. If we used a current source of 1 for the antenna, we shall find the current at the center of the forward element to be about 0.455 at -16.69 degrees. At the center of the rear element, we find the current to be about 0.462 at 159.78 degrees. The rear element has a magnitude of 1.015 and phase of 143.09 degrees relative to the forward element. Yet, the phase line length was only 0.1 wavelength long physically and 0.15 wavelength long electrically. How is this possible?

To find out, let's make a slight change in the model. If the forward element was modeled from left to right (-8.0' to +8.0'), then let's make the rear element in the reverse direction (+8.0' to -8.0'). Next, let's change the transmission line back to its normal or non-reversed orientation.

When we run this model, we obtain identical performance figures to those cited above. However, the current tables show another story. The forward element remains as it was, but the rear element shows a center current of 0.462 at -20.22 degrees. The rear element is -36.91 degrees relative to the forward element, which is equivalent to a phase angle of 323.15 degrees. Of course, subtracting 180 degrees from this figure yields 143.15 degrees, which we obtained from the first version of the model.

The current phase change of (-)36.91 degrees along the 71-ohm, 0.67 VF phase line of 3.46' from the rear element to the forward element is precisely what we need to provide the two elements with the relative current magnitude and phase angles so that the array achieves a front-to-back performance far in excess of what a similar driven element-reflector Yagi might obtain with the same spacing. The details of the analysis of ZL Specials go way beyond our purpose here, but the ZL Special is an excellent example of modeling transmission lines between two elements.

There are numerous other feedline uses of feedlines, but these examples should get you started. We may also use feedlines for other things, for example, as substitutes for lumped components. However, that set of applications will require that we search for shorted and open lines, and they can only be found in the next part of this series.



4. Transmission Lines as Lumped Constants



L. B. Cebik, W4RNL

Lengths of transmission line can be used as "lumped constants," that is as inductive or capacitive reactances. In many cases, the use of a transmission line to provide a desired reactance is preferable to or at least as equally good as the use of a capacitor or inductor. The reasons are many and depend on the particular circumstances of the application. For some purposes, transmission line lengths have a higher Q; in others, they are more durable or better shaped. In antenna work, transmission line lengths--or stubs--are very useful, and therefore worth modeling in place.

A. Stub Principles

The general principles of creating reactances from transmission line stubs is well known. The figure shows the general technique.

using.

For open stubs, we use the equation

$$X = -Z_0 \cot \beta l \text{ or } X = -\frac{Z_0}{\tan \beta l}$$

where the terms have the same meaning. The value of X will be either positive or negative, indicating inductive or capacitive reactance. At exactly 90 degrees, the tangent of the angle climbs without limit, resulting in a value of reactance that is zero for open stubs and that is indefinitely high (some folks like to say infinite) for shorted stubs.

At zero degree lengths and at multiples of 180 degree lengths, the value of the tangent of the angle is zero. The stubs reverse their roles with their 1/4 wavelength counterparts.

There are a number of applications of shorted and open stubs in antennas, and we can model most of them. However, we must always remember that transmission lines (TL entries) in NEC represent lossless lines. For all but short lines, we should externally calculate line losses in the final evaluation of any antenna design where the line or stub is a part of the antenna.

B. Entering Stub (TL) Data

Every transmission line or TL entry in NEC requires two wire segments to act as terminations for the line. In effect, the TL entry opens the wire segment and inserts the transmission line segment in series with it at that point. Without wire segments at either end, the transmission line cannot be formed.

This process is invisible with EZNEC.

EZNEC Transmission Line Screen									
Note highlighted entries across screen									
Line	Wire #/%	From End 1	Wire #/%	From End 2	Length	Z0	Vel	Rev/	
	Actual	(Specified)	Actual	(Specified)		Ohms	Fact	Norm	
1	1/25	(1/25)	Short ckt	(Short ck)	11.735	450	.95	N	
2	1/75	(1/75)	Open ckt	(Open ckt)	7.3270	450	.95	N	
Wire number, % of way from end 1; (o)pen; or (s)hort									
(Enter position at wire end or center as 'w#e#' or 'w#c', if desired.)									
Note: transmission lines are placed in parallel with sources and other Transmission lines. They are placed in series with loads.									

As the screen replica shows, we have inserted 2 stubs in Wire 1. The first is a shorted stub, which we specified simply by keying "s" for end 2 of the transmission line. The remainder of the entry of length, Zo, and velocity factor are normal. We use a similar procedure for the second open stub by

specifying "o" for end 2 of the line.

What the entry does not show you is a second wire created by the program to act as the other termination of the line specified. The line is created about 100 wavelengths away from the antenna in all 3 axes, X, Y, and Z. Additionally, the line is about 0.002 wavelengths long and divided into two segments, one to act as the terminal point of each of the two loads. The wire has a radius of about 0.0001 wavelength, the smallest permissible under NEC. Each of these measures ensures that the wire will interact minimally with the other wire segments, thus not disturbing the accuracy of the impedance, current, and field calculations.

You can easily recreate this situation with programs, such as NEC-Win, where you may have to create your own stub conditions.

The screenshot shows the 'TL - Transmission Line' dialog box. It has a title bar with a close button. The main area is divided into four quadrants. Top-left: 'Place Transmission Line' with 'From' (Tag #: 1, Segment #: 25) and 'To' (Tag #: 2, Segment #: 2). Top-right: 'Shunt Admittance' with 'End One' (Real: 0., Imaginary: 0.) and 'End Two' (Real: .E+10, Imaginary: 1.E+1), both in mhos. Bottom-left: 'Characteristic Impedance (ohms):' 450. and 'Transmission Line Length (m):' 3.687. Bottom-right: 'Transmission Line Type' with 'No Phase Reversal' selected. At the bottom are 'OK' and 'Cancel' buttons.

The TL window in NEC-Win presumes that you have created the second wire. In this case, we are creating a shorted stub at wire 1, segment 25, and it goes to wire 2, segment 2. We are using 450-ohm transmission line with a length of 3.687 meters (12.09'). Note, however, one final necessary element of the stub. End 2 must specify the conditions for a short circuit across the terminals, which means an indefinitely high value of "shunt admittance" (the inverse of a shunt impedance). Since the boxes are small, I have move the "real" (conductance) component to the right to show the exponent value, while the "imaginary" (susceptance) component remains shifted left. Values of 1E+10 are so high as to be indistinguishable from a true short circuit.

The corresponding input line in the text editor would appear as follows:

```
TL 1 25 2 2 450 3.687 0 0 1E+10 1E+10
  1st End Seg. 2nd End Seg. Zo Length Real Imag. Real Imag.
  Wire # # Wire # # Ohms meters Adm. Adm. Adm. Adm.
      W1 shunt W2 shunt
```

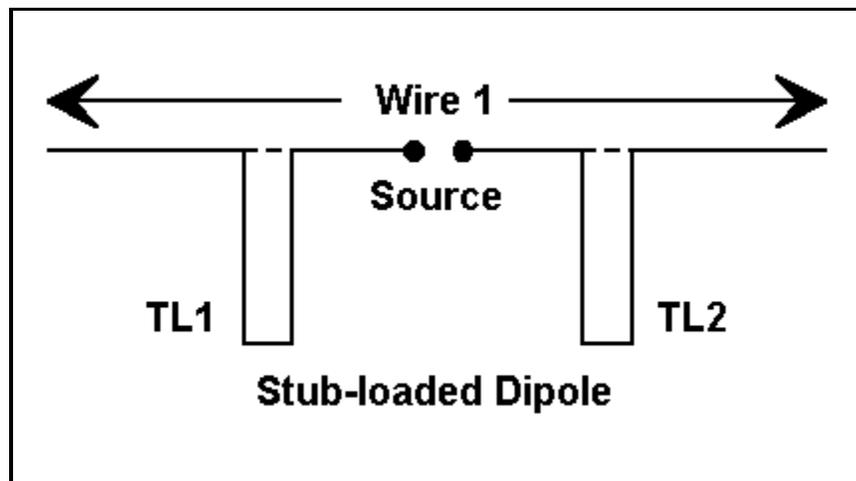
For an open circuit, we would use the same procedure, but place values of 1E-10 in each of the boxes. This low a value of admittance corresponds to a value of 1E+10 for impedance, indistinguishable from an open circuit. Although such values are manually placed in this NEC-Win example, EZNEC places them automatically and invisibly to the user.

Now that we can create transmission line stubs, we can litter our antennas with them. Unlike physically modeled transmission lines and stubs, these TL entries occupy no wire segments, speeding core calculation runs and allowing greater antenna detailing. They appear in series with loads on the same segments they occupy, but in parallel with sources on the same segment.

C. A Basic Stub Use

One basic use of transmission stubs is to replace inductors as loads along antenna elements. To understand their use within models, let us return to our running model: the 101 segment #14 copper wire model in free space of a dipole shortened to about half-normal length. In past exercises, we loaded this dipole at segments 25 and 77, first with a reactance of 870.4 ohms, and then with an inductor of $9.79e-6$ H, at 14.15 MHz. For a lossless inductor, we obtained a gain of 1.76 dBi in free space, with a feedpoint (source) impedance of $26.08 + j1.65$ ohms.

Now let's replace the R-X and R-L-C type loads with a transmission line stub. In fact, let's try a couple of different stubs.



First, we can replace the loads with TL entries at the same segments using 450-ohm, 0.95 VF transmission line. The standard equation tells us that the line needs to be 11.494' (3.503 m). This will show up in the TL line in the text editor as 3.687 m, the equivalent length of line with a VF of 1.0. (The second wire is assumed to have been placed either manually or automatically by the program.)

At our test frequency, the model returns (in NEC-4) a source impedance of $26.09 + j1.72$ ohms, and a gain of 1.76 dBi in free space. The slight numerical differences from the R-x and R-L-C load models, of course, make no difference at all.

Now let's replace the 450-ohm line with 50-ohm coaxial cable with a VF of 0.66 (RG-213). The required line length will be 11.05' (3.369 m), which will translate to 5.104 m of 1.0 VF line in the TL entry.

Using this line in the model, it returns a source impedance value of $26.09 + j2.147$ ohms and a free space gain of 1.76 dBi. Again, no difference.

Although we recognize that these lines are lossless in the model, losses in transmission line lengths--when used within their recommended limits--tend to be negligible for such short lengths.

We would expect no changes in the real antenna that are attributable to line loss adjustments, in contrast to those occasioned by wire and line cutting and assembly precision. However, transmission lines have at least wire losses which given them finite Qs in applications such as these. Although the Qs may be higher than those of inductors, we may not expect true lossless performance.

An exercise such as this one can mislead as well as instruct. At first sight, it would appear that we might easily replace our loading inductor with transmission line stubs and have the same antenna. This would be true if and only if we operated the antenna only at 14.15 MHz.

D. Frequency Sweeping Inductive and Stub Loads

Let us take the trouble to run frequency sweeps of our half size dipole using all three loads: the inductor, the 450-ohm stub, and the 50-ohm stub. Across 20 meters from 14.0 to 14.3 MHz, we would obtain the following sets of values. The values are grouped according to source resistance, source reactance, and SWR relative to 26.08 Ohms (the value at resonance).

Source Resistance (in Ohms):

Type Load	FQ: 14.0	14.05	14.1	14.15	14.2	14.25	14.3
Inductive	24.84	22.24	25.66	26.08	26.52	26.96	27.42
450-ohm Stub	24.38	24.92	25.49	26.09	26.70	27.36	28.04
50-ohm Stub	20.25	21.65	23.50	26.09	29.98	36.50	49.50

Source Reactance (in Ohms):

Type Load	FQ: 14.0	14.05	14.1	14.15	14.2	14.25	14.3
Inductive	-42.3	-27.8	-13.1	1.65	16.47	31.77	47.06
450-ohm Stub	-68.9	-46.1	-22.5	1.72	26.59	52.64	79.38
50-ohm Stub	-320.	-240.	-136.	2.15	198.8	502.9	1038.

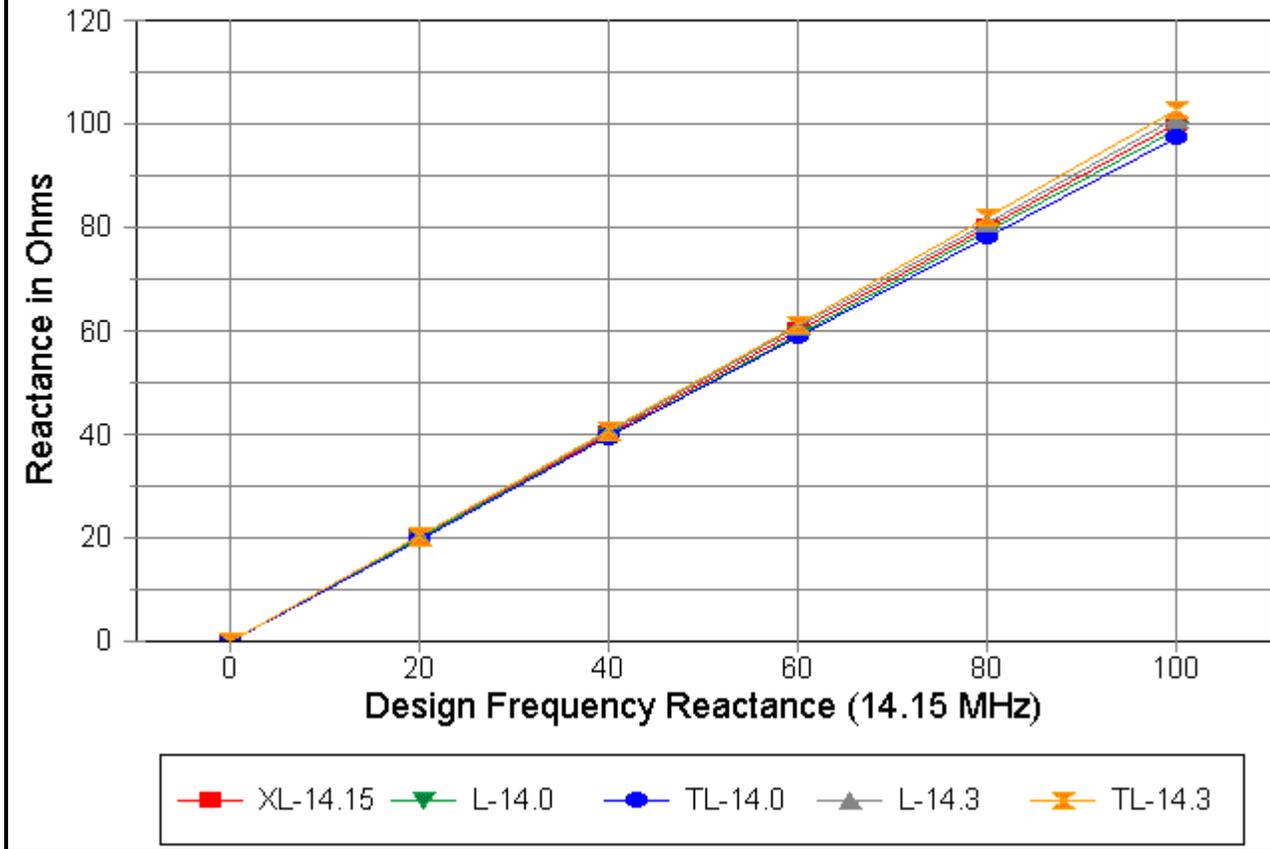
SWR relative to 26.08 Ohms:

Type Load	FQ: 14.0	14.05	14.1	14.15	14.2	14.25	14.3
Inductive	4.54	2.82	1.65	1.07	1.85	3.12	4.90
450-ohm Stub	9.35	5.07	2.34	1.07	2.65	5.71	10.50
50-ohm Stub	>100.	>100.	32.3	1.08	52.5	>100.	>100.

Clearly, something is different about the stubs relative to the inductor-- and between the stubs themselves. Although the rate of change in source resistance is moderate as we move from one load to the next, the rate of change in the source reactance is much higher for the stubs--and extremely high for the low-impedance transmission line. The 2:1 SWR operating bandwidth for the inductive load is less than 200 kHz, for the 450-ohm stub less than 100 kHz, and for the 50-ohm stub less than 20 kHz.

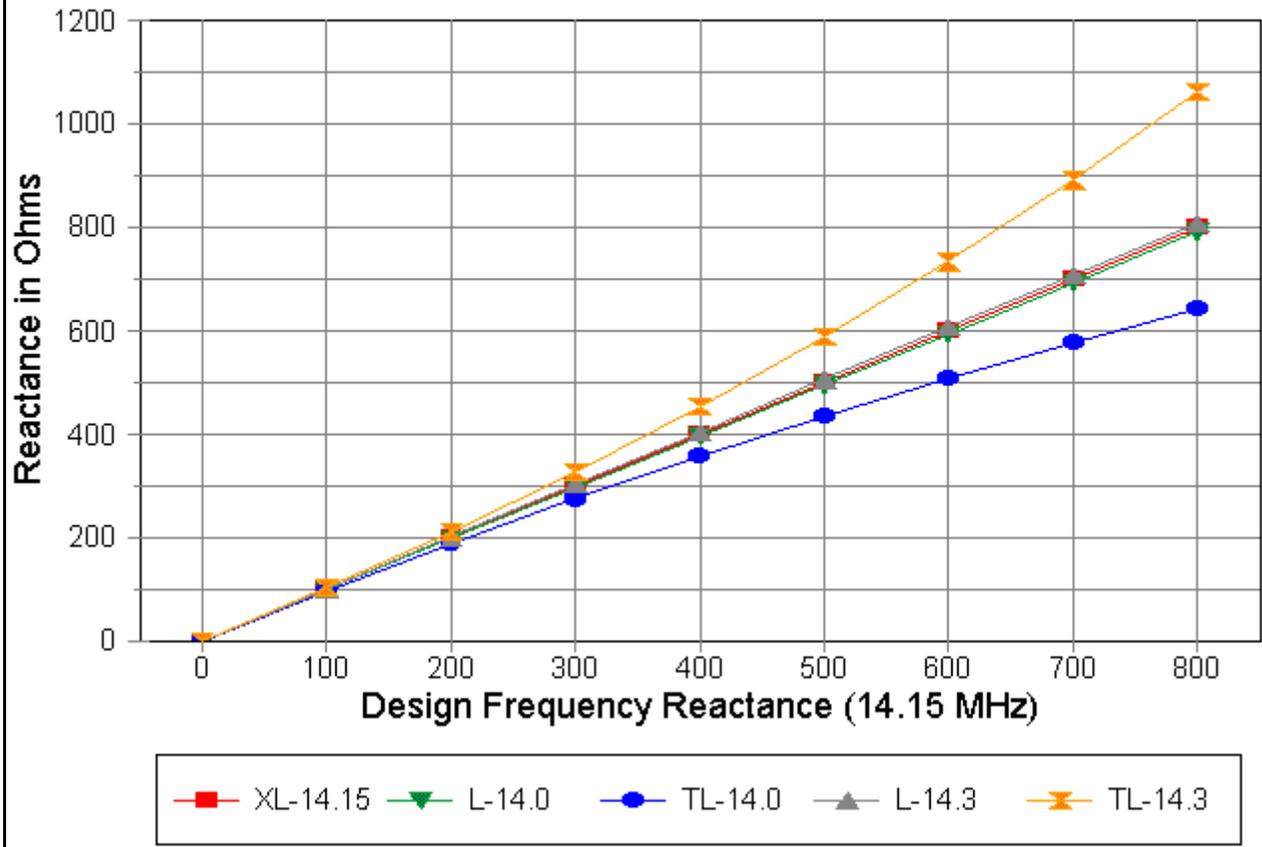
Stubs change reactance according to changes in the TAN function of the electrical length in degrees or radians and relative to the characteristic impedance of the transmission line. Inductors, however, change reactance according to changes in frequency. The differences in the way each changes reactance do not show up if the required inductive reactance of a shorted stub is low--in the 0 to 100 ohm range.

Coax Stub vs. Inductor Reactance Away from Design Frequency



The graph compares change in reactance at 14.0 and at 14.3 MHz for an inductor (inner lines) and 50-ohm transmission stubs (outer lines) when the load at 14.15 MHz is assigned values between 0 and 100 ohms. Even at the upper end of the graph, the differential between 50-ohm stubs and an inductor is a matter of a few ohms. However, when the required inductive reactance is greater--between 100 and 800 ohms--a more radical divergence appears.

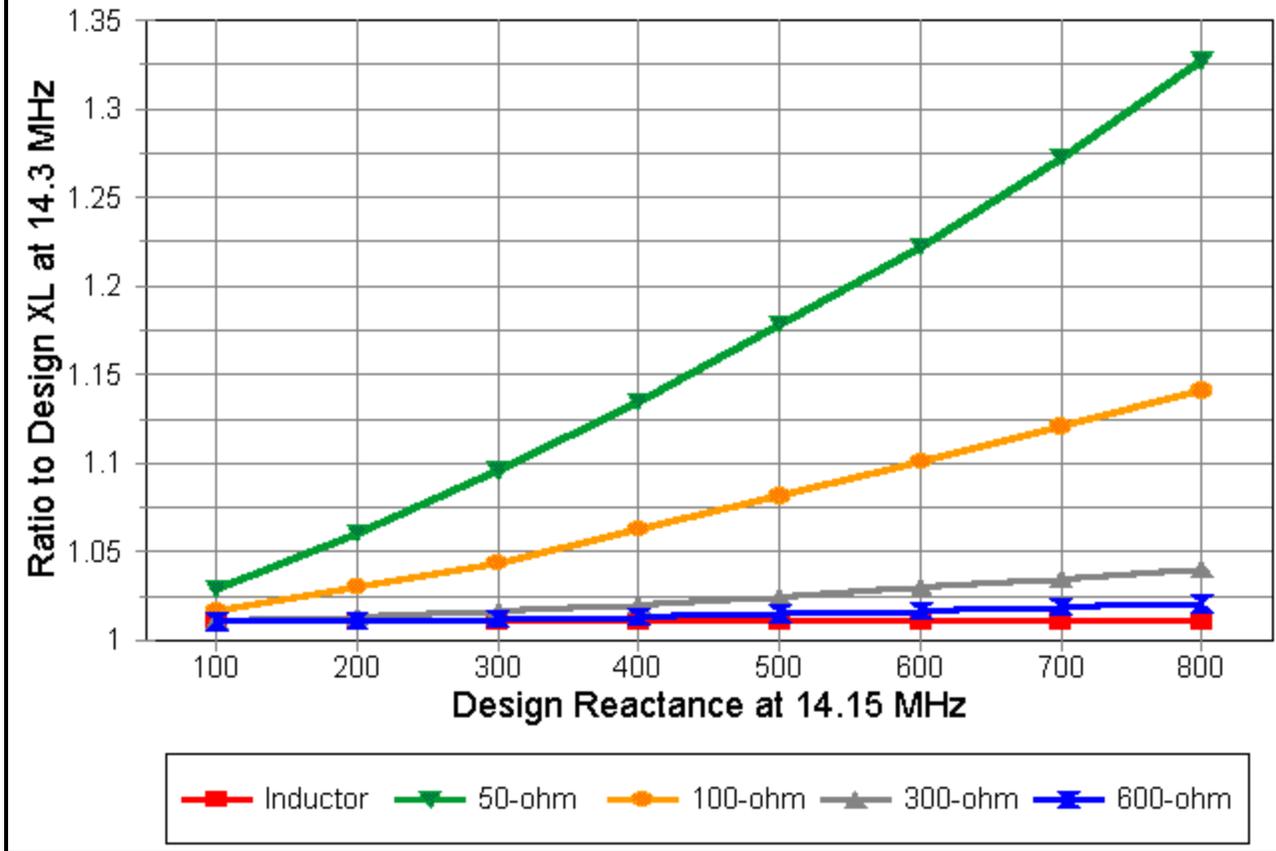
Coax Stub vs. Inductor Reactance Away from Design Frequency



As the graph shows, the inductor changes its reactance by only small amounts all the way up to 800 ohms. However, the 50-ohm transmission line stub changes its value more rapidly as the frequency departs from the design center as the load value increases. The result is that for antennas requiring high values of inductive reactance as loads, the 50-ohm stub permits only a very narrow operating bandwidth.

The 50-ohm stub is a very extreme case, used here to dramatize the differences in the rate of change of value between transmission line stubs and inductors with frequency movement from the design center. Fortunately, the higher the impedance of the transmission line stub, the smaller the difference between its change of reactance and the change of reactance of an inductor over the same frequency deviation from design center.

Rate of Departure from Design XL Inductor, 50-ohm to 600-ohm Lines



The graph above is developed on a different premise. Let us use the change of reactance of the inductor as a base line. If the design center is 14.15 MHz, then at a frequency of 14.3 MHz, the percent of change in reactance will be constant over the entire range of reactances. In contrast, the percent of change of transmission line stubs will vary, increasing with the reactance required, and so too will the ratio of the change in reactance in stubs to the change in inductors.

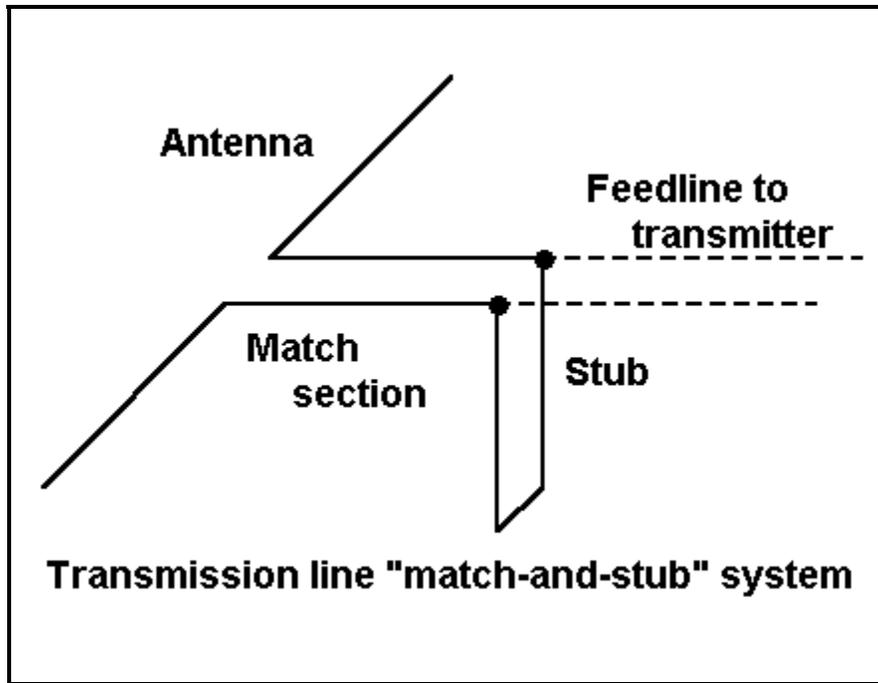
The graph shows the ratios for 50, 100, 300, and 600 ohm lines, relative to the inductor's baseline. The lesson is clear: for the widest bandwidth when using transmission line stubs, use the highest feasible characteristic-impedance transmission line. Even so, a lossless inductor will always show a wider operating band width than an equally lossless transmission line stub.

For inductors, Q_s of 150 to 300 are common in antenna loading service. These Q_s generally further increase the operating bandwidth at the cost of slight losses in antenna gain due to the resistive losses. In contrast, transmission line stubs less than $1/4$ wavelength long rarely have Q_s less than 300, and often they can reach 1000 to 1500. Hence, stubs offer slightly higher gain at the expense of operating bandwidth for any given antenna design that is inductively loaded.

For the antenna modeler, this exercise should reinforce the importance of making frequency sweeps across every the range of frequencies for which the antenna is designed. The sweep must be run using the correctly modeled load for the application. Even if losses are not apparent from the modeling exercise itself, they can be separately calculated and factored into the design.

E. Other Applications of Stubs

Element loading is not the only possible application for modeled transmission line stubs. We can also combine stubs with normally terminated transmission lines to explore the properties of the combined array. A common application in this vein is the matching-section-and-stub system used to match an antenna's odd source impedance to a standard feedline.



Consider the following example, corresponding to the sketch. We have an extended double Zepp antenna with a design center frequency of 28.5 MHz which we wish to feed with 50-ohm coax. Can we design a match and stub system to do the job? If we can design the system, what will be the 2:1 SWR operating bandwidth of the resulting antenna and match system?

The antenna model consists of one wire, 44.2' (13.48 m) long, with 31 segments. We shall place this antenna at a height of 35' (1 wavelength) above medium ground conditions. The program returns a gain of 10.76 dBi and a source impedance of 135.3 -j680.1 ohms. The source impedance becomes the starting point for our match-and-stub system. We shall use 450-ohm, 0.95 VF parallel line for the transition to our 50-ohm coax.

Because transmission lines are impedance transformers, we can find 2 points along each half-wavelength of the initial 450-ohm line at which a properly sized stub (either open or shorted) will yield an impedance of $50 \pm j0$ ohms. (The limiting case is where the impedance along the line never reaches 50 ohms, in which case, we must select a different Z_0 line for the match and stub system.) We can use either a Smith Chart or a utility program, such as the one included in the HAMCALC collection, to simplify the calculation process.

For the case in hand, we find that lengths of A. 4.983' and B. 5.441' are able to support stubs for the transformation. At point A., a shorted stub 1.176' long or an open stub 9.373' long will do the job. At point B., a shorted stub 12.216' long or an open stub 7.020' long will also do the job.

Ordinarily, we select the combination of matching section and stub that is shortest. In this example, the 4.983' + 1.176' stub combination is the most compact.

To place these lines in the model, we create a second wire, 0.02' long at a height of about 30' up. The exact height is not critical, since we shall control the length of the match line with the TL entry. If we must manually create the distant wire for the termination of the shorted stub, we do so at this time.

Then we enter two transmission lines, both 450 ohms and 0.95 VF. The first goes between wire 1, segment 16 and wire 2, segment 1. The length is 4.983 feet. The second is a shorted stub from wire 2 segment 1, with a length of 1.176 feet. (Adjust lengths for a VF of 1.0 if directly entering values into the TL line: 1.599 m for the match line and 0.378 m for the stub.) Be certain to move the source from the center of wire 1 to wire 2, segment 1.

Running this revised model yields at 28.5 MHz a source impedance of 49.97 - j0.01 ohms, showing the coincidence between the external calculations and our modeling. The gain, of course, does not change. However, once we have established the fact that our external calculations and our modeling work coincide, is there any use to installing the transmission lines in our model?

The answer is affirmative, for we can use this installation to determine the anticipated 2:1 SWR operating bandwidth of the antenna. If we perform a frequency sweep across the first MHz of 10 meters, we obtain the following data:

Frequency in MHz	Source Impedance R +/- jX Ohms	50-Ohm SWR
28.0	111.0 - j26.13	2.37
28.1	93.43 - j22.74	2.02
28.2	79.03 - j17.74	1.71
28.3	67.30 - j11.98	1.43
28.4	57.75 - j 5.97	1.20
28.5	49.97 - j 0.01	1.00
28.6	43.57 + j 5.74	1.20
28.7	38.29 + j11.21	1.45
28.8	33.89 + j16.36	1.73
28.9	30.22 + j21.18	2.07
29.0	27.12 + j25.70	2.47

The data tell us that we are likely to be able to operate the antenna without equipment indisposition over at least 0.8 MHz bandwidth. Moreover, using the curve as a guide, we are likely to be able to make field adjustments of the actual match and stub system to bring it into line with the portion of the band we most wish to use. Since we are using a mere 6' (2 m) of line for the match and stub system, line losses will be negligible within it. In any event, the losses will be far less than had we designed an inductor at the feedpoint to compensate for the capacitive reactance, followed by a 2:1 transformer to bring the resistive component of the source impedance closer to 50 ohms.

This example, I hope, illustrates the utility of judiciously employing stubs in antenna models beyond their function to load antenna wires. I recommend that you rebuild the model using each of the other three combinations of match line and stub just for the practice in using TL entries in models (if not out of curiosity over whether those values "work").

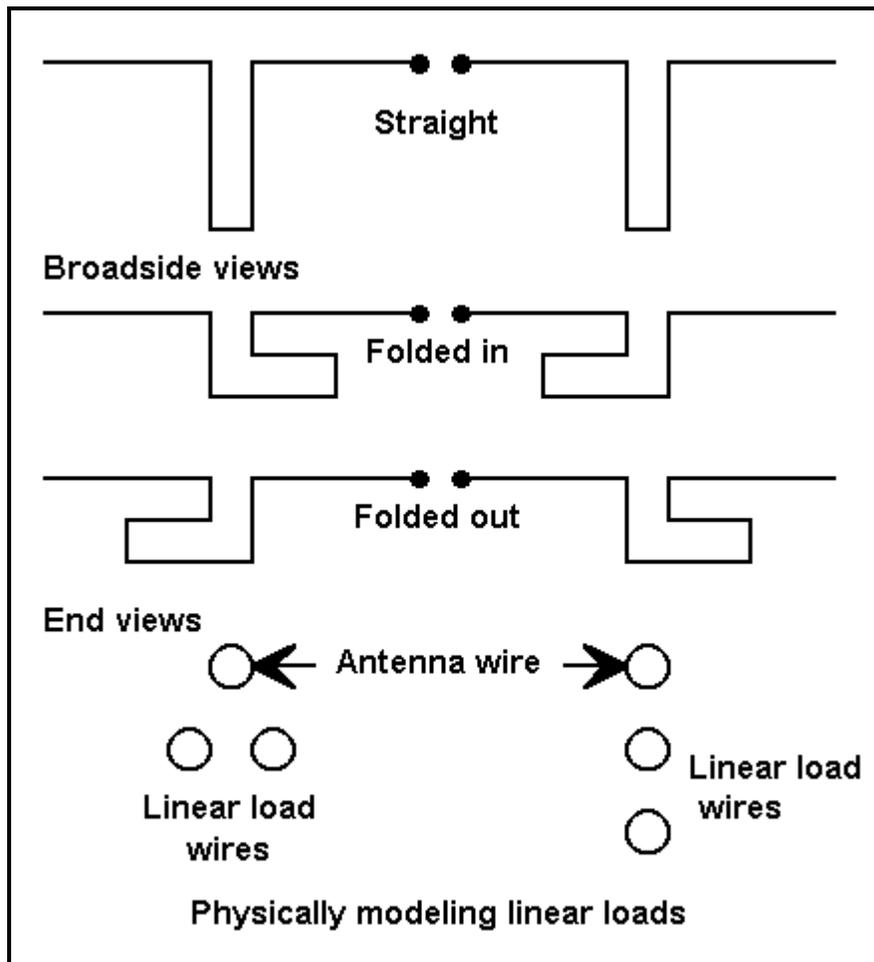
E. Physically Modeling Stubs

Like regular transmission lines, we can also model stubs as physical elements within the antenna design. The advantage is that any interactions of the stubs with other parts of the antenna structure will show up in the model's output data. The disadvantage is that modeling the stubs will occupy

wires and segments, thus slowing the core calculations. For complex designs, modeling all stubs can drive the model size to the program limits.

In addition, we must always be aware of basic NEC limitations. Primary among these limits is the possible unreliability of results where wires of different radii (diameters) meet, especially at angular junctions. If possible, one should model such geometries using the same wire size throughout, and this may require the development of several different models to obtain all of the desired output data, such as element lengths, stub lengths, gain, and source impedance. In addition, segments at and near angular junctions should be consistent in length and short.

Let's sample the process by replacing the loads and transmission lines in the half-length dipole with physically modeled transmission line stubs. To do this, we must restructure the entire model. Instead of one wire, we shall need 9, in accord with the upper diagram of the following figure.



The wires follow the following table:

Wire	Conn.	End 1 (x,y,z : in)	Conn.	End 2 (x,y,z : in)
1		-101.21, 0.000, 0.000	W2E1	-54.000, 0.000, 0.000
2	W1E2	-54.000, 0.000, 0.000	W3E1	-54.000, 0.000, 76.300
3	W2E2	-54.000, 0.000, 76.300	W4E1	-50.000, 0.000, 76.300
4	W3E2	-50.000, 0.000, 76.300	W5E1	-50.000, 0.000, 0.000
5	W4E2	-50.000, 0.000, 0.000	W6E1	50.000, 0.000, 0.000
6	W5E2	50.000, 0.000, 0.000	W7E1	50.000, 0.000, 76.300
7	W6E2	50.000, 0.000, 76.300	W8E1	54.000, 0.000, 76.300

8 W7E2 54.000, 0.000, 76.300 W9E1 54.000, 0.000, 0.000
9 W8E2 54.000, 0.000, 0.000 101.210, 0.000, 0.000

The source is at the center of wire 5. Dimensions have been translated into inches for ease of seeing the modeled structure. The stubs are composed of the same #14 copper wire as the antenna element and are 4" wide by 76.3" long. The stubs are centered at the approximate points where the loads and transmission lines had been assigned in previous models.

The model returns a gain of 1.47 dBi and a source impedance of $18.02 - j0.55$ ohms. The gain figure is less than the 1.76 dBi returned by the earlier models. However, remember that the stub now accounts for wire resistance losses. In fact, if we add a resistance of 0.93 ohms to the reactive load of 870.4 ohms in previous models, we find that the R-X model returns the same gain figure of 1.47 dBi. 0.93 ohms resistance in series with a reactance of 870.4 ohms represents a Q of about 936, well higher than what we can achieve with real inductors.

The model is also instructive with respect to the length of the stub, which is considerably shorter than what a standard calculation of a shorted stub would yield. #14 wire with 4 inch spacing has an approximate Z_0 of 580 ohms, which would call for a length of about 130" for the stub. The actual stub is 76.3" long. However, the theoretical stub presumes equal but opposite currents in each leg. Examining the current table returned for this model will show the currents for any given position across the stub lines to be significantly different.

(Note: Modelers should be very cautious at this point with respect to the accuracy of the stub length as modeled in NEC-2 or NEC-4. Even 4" apart at 14.15 MHz, the stub wires are very close. NEC may become inaccurate with very closely spaced wires, even when the modeler has observed all cautions regarding using the same wire diameter at angular junctions and regarding the alignment of segments in the parallel wires. This same antenna was modeled in MININEC 3.13 (which has limitations of its own, but not in this area) via ELNEC, and the requisite stub lengths were 1.3 to 1.6 inches longer, depending upon the precise technique used to construct the model. Since the stub length differential approaches 2%, the variance must be considered significant.)

Stubs close to the feed point, where the current changes slowly as one moves along the antenna wire, will model physically in close accord with standard parallel transmission line calculations. However, for closest equality of current magnitude on the wires, they must be positioned symmetrically relative to the antenna wire, as shown in the lower left of the figure. If they are positioned linearly, the currents will be unequal due to the unequal distances from the main antenna element. Likewise, stubs positioned further out along the antenna wire, even if positioned symmetrically relative to the antenna wire, will show considerable differences of current on the two parallel wires. At a certain indefinite point along the wire, stubs act less like transmission line sections and more like simple wire extensions necessary to lengthen an antenna element to resonance.

The differences in currents along the two stub wires tend to be somewhat less if the stub wires are folded back or folded outward, parallel to the antenna wire and equally spaced from it. When positioned straight down, the unequal currents in the stubs yield a small vertically polarized field that in one plane is down from the main lobe by only -18 dB. It is largely the interaction of this field with the main element field that reduces the source impedance to 18 ohms from the 26 ohms found in the other models.

A physical model of the stub-loaded half-length dipole also provides an interesting lesson in operating band width. Relative to the resonant source impedance of 18 ohms, the SWR at 14.0

MHz is 2.50 and at 14.3 MHz is 2.30, well below the modeled figures for lossless transmission lines and lossless inductors. However, the physical model includes its losses, which increase the operating bandwidth.

The point of generating this sample physical model of transmission line stubs--usually called "linear loads" among antenna designers--is to illustrate some, but by no means all, of the differences between using lossless and non-interacting mathematical antenna parts and physical lossy and fully interactive antenna parts. Each provide valuable information about anticipated antenna performance and each withholds valuable information. Physical modeling is limited by certain program limitations that force the modeler to depend in part on the use of mathematical substitutes. However, only by thorough modeling in both modes can one develop a good sense for what to expect by way of field adjustment in real antennas.



This last point is as good as any for bringing to a close this inexhaustive but hopefully useful set of exercises in modeling with loads and transmission lines. One could multiply the examples almost endlessly to draw out innumerable subtleties regarding both antennas and models of them. However, I hope that the examples and notes in this series allow you to carry on the process on your own. There is much we can learn about antennas, and antenna modeling can go a little way toward instructing us-- even about its own limitations.