Chapter 8

Multielement Arrays

The gain and directivity offered by an array of elements represents a worthwhile improvement both in transmitting and receiving. *Power gain* in an antenna is the same as an equivalent increase in the transmitter power. But unlike increasing the power of your own transmitter, antenna gain works equally well on signals received from the favored direction. In addition, the directivity reduces the strength of signals coming from the directions not favored, and so helps discriminate against interference.

One common method of obtaining gain and directivity is to combine the radiation from a group of $\lambda/2$ dipoles to concentrate it in a desired direction. A few words of explanation may help make it clear how power gain is obtained.

In **Fig 1**, imagine that the four circles, A, B, C and D, represent four dipoles so far separated from each other that the coupling between them is negligible. Also imagine that point P is so far away from the dipoles that the distance from P to each one is exactly the same (obviously P would have to be much farther away than it is shown in this drawing). Under these conditions the fields from all the dipoles will add up at P if all four are fed RF currents in the same phase.

Let us say that a certain current, I, in dipole A will produce a certain value of field strength, E, at the distant point P. The same current in any of the other dipoles will produce the same field at P. Thus, if only dipoles A and B are operating, each with a current I, the field at P will be 2E. With A, B and C operating, the field will be 3E, and with all four operating with the same I, the field will be 4E. Since the power received at P is proportional to the square of the field strength, the relative power received at P is 1, 4, 9 or 16, depending on whether one, two, three or four dipoles are operating.

Now, since all four dipoles are alike and there is no coupling between them, the same power must be put into each in order to cause the current I to flow. For two dipoles the relative power input is 2, for three dipoles it is 3, for four dipoles 4, and so on. The actual gain in each case is the relative received (or output) power divided by the relative input power. Thus we have the results shown in **Table 1**. The power ratio is directly proportional to the number of elements used.

It is well to have clearly in mind the conditions under which this relationship is true:

1) The fields from the separate antenna elements must be in-phase at the receiving point.

2) The elements are identical, with equal currents in all elements.

3) The elements must be separated in such a way that the current induced in one by another is negligible; that is, the radiation resistance of each element must be the same as it would be if the other elements were not there.

Very few antenna arrays meet all these conditions exactly. However, the power gain of a directive array using dipole elements with optimum values of element spacing is approximately proportional to the number of elements.

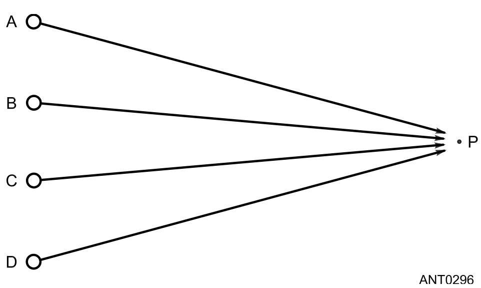


Fig 1—Fields from separate antennas combine at a distant point, P, to produce a field strength that exceeds the field produced by the same power in a single antenna.

Table 1
Comparison of Dipoles with Negligible Coupling (See Fig 1)

Dipoles	*Relative Output Power*	*Relative Input Power*	*Power Gain*	*Gain in dB*
A only	1	1	1	0
A and B	4	2	2	3
A, B and C	9	3	3	4.8
A, B, C and D	16	4	4	6

Another way to say this is that a gain of approximately 3 dB will be obtained each time the number of elements is doubled, assuming the proper element spacing is maintained. It is possible, though, for an estimate based on this rule to be in error by a ratio factor of two or more (gain error of 3 dB or more), especially if mutual coupling is *not* negligible.

DEFINITIONS

An *element* in a multi-element directive array is usually a λ/2 radiator or a λ/4 vertical element above ground. The length is not always an exact electrical half or quarter wavelength, because in some types of arrays it is desirable that the element show either inductive or capacitive reactance. However, the departure in length from resonance is ordinarily small (not more than 5% in the usual case) and so has no appreciable effect on the radiating properties of the element.

Antenna elements in multi-element arrays of the type considered in this chapter are always either *parallel*, as in **Fig 2A** , or *collinear* (end-to-end), as in Fig 2B. Fig 2C shows an array combining both parallel and collinear elements. The elements can be either horizontal or vertical, depending on whether horizontal or vertical polarization is desired. Except for space communications, there is seldom any reason for mixing polarization, so arrays are customarily constructed with all elements similarly polarized.

A *driven element* is one supplied power from the transmitter, usually through a transmission line. A *parasitic element* is one that obtains power solely through coupling to another element in the array because of its proximity to such an element.

A *driven array* is one in which all the elements are driven elements. A *parasitic array* is one in which one or more of the elements are parasitic elements. At least one element must be a driven element, since you must somehow introduce power into the array.

A *broadside array* is one in which the principal direction of radiation is perpendicular to the axis of the array and to the plane containing the elements, as shown in **Fig 3**. The elements of a broadside array may be collinear, as in Fig 3A, or parallel (two views in Fig 3B).

An *end-fire array* is one in which the principal direction of radiation coincides with the direction of the array axis. This definition is illustrated in **Fig 4**. An end-fire array must consist of parallel elements. They cannot be collinear, as λ/2 elements do not radiate straight off their ends. A Yagi is a familiar form of an end-fire array.

A *bidirectional array* is one that radiates equally well in either direction along the line of maximum radiation. A bidirectional pattern is shown in **Fig 5A**. A *unidirectional array* is one that has only one principal direction of radiation, as the pattern in Fig 5B shows.

The *major lobes* of the directive pattern are those in which the radiation is maximum. Lobes of lesser radiation intensity are called *minor lobes*. The *beamwidth* of a directive antenna is the width, in degrees, of the major lobe between the two directions at which the relative radiated power is equal to one half its value at the peak of the lobe. At these *half-power points* the field intensity is equal to 0.707 times its maximum value, or in other words, is down 3 dB from the maximum. **Fig 6** shows a lobe having a beamwidth of 30°.

Unless specified otherwise, the term *gain* as used in this

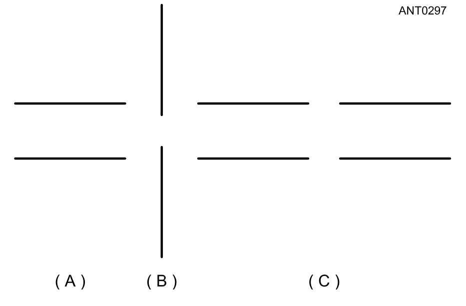


Fig 2—At A, parallel and at B, collinear antenna elements. The array shown at C combines both parallel and collinear elements.

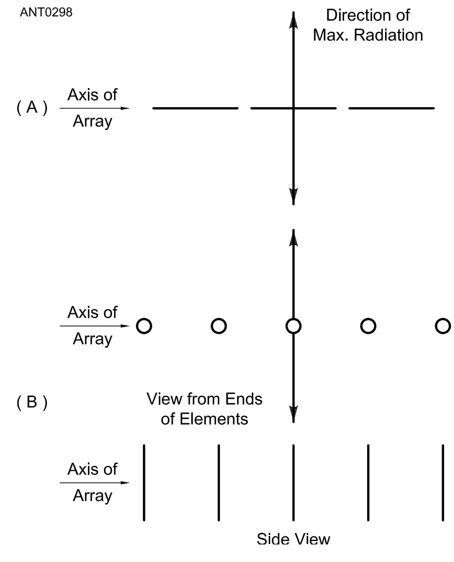


Fig 3—Representative broadside arrays. At A, collinear elements, with parallel elements at B.

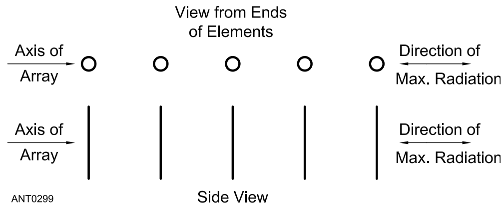


Fig 4—An end-fire array. Practical arrays may combine both broadside directivity (Fig 3) and end-fire directivity, including both parallel and collinear elements.

section is the power gain over an isotropic radiator in free space. The gain can also be compared with a λ/2 dipole of the same orientation and height as the array under discussion and having the same power input. Gain may either be measured experimentally or determined by calculation. Experimental measurement is difficult and often subject to considerable error, for two reasons. First, errors normally occur in measurement because the accuracy of simple RF measuring equipment is relatively poor—even high-quality instruments suffer in accuracy compared with their low-frequency and dc counterparts. And second, the accuracy depends considerably on conditions—the antenna site, including height, terrain characteristics, and surroundings—under which the measurements are made.

Calculations are frequently based on the measured or theoretical directive patterns of the antenna (see Chapter 2). The theoretical gain of an array may be determined approximately from:

$$G = 10 \log \frac{41{,}253}{\theta_H \ \theta_V} \qquad \text{(Eq 1)}$$

where

G = decibel gain over a dipole in its favored direction
θ_H = horizontal half-power beamwidth in degrees
θ_V = vertical half-power beamwidth in degrees.

This equation, strictly speaking, applies only to lossless antennas having approximately equal and narrow E- and H-plane beam widths—up to about 20°—and no large minor lobes. The E and H planes are discussed in Chapter 2. The error may be considerable when the formula is applied to simple directive antennas having relatively large beam widths. The error is in the direction of making the calculated gain larger than the actual gain.

Front-to-back ratio (F/B) is the ratio of the power radiated in the favored direction to the power radiated in the opposite direction. See Chapter 11 for a discussion of front-to-back ratio, and its close cousin, *worst-case front-to-rear ratio*.

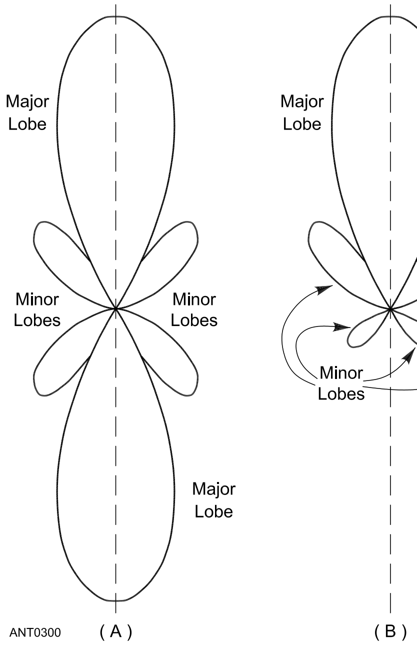


Fig 5—At A, typical bidirectional pattern and at B, unidirectional directive pattern. These drawings also illustrate the application of the terms *major* and *minor* to the pattern lobes.

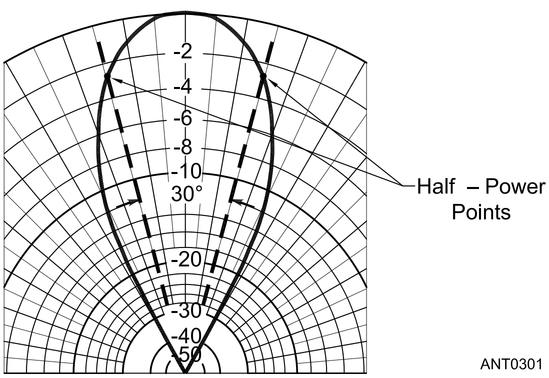


Fig 6—The width of a beam is the angular distance between the directions at which the received or transmitted power is half the maximum power (–3 dB). Each angular division of the pattern grid is 5°.

Phase

The term *phase* has the same meaning when used in connection with the currents flowing in antenna elements as it does in ordinary circuit work. For example, two currents are in-phase when they reach their maximum values, flowing in the same direction, at the same instant. The direction of current flow depends on the way in which power is applied to the element.

This is illustrated in **Fig 7**. Assume that by some means an identical voltage is applied to each of the elements at the ends marked A. Assume also that the coupling between the elements is negligible, and that the instantaneous polarity of the voltage is such that the current is flowing away from the point at which the voltage is applied. The arrows show the assumed current directions. Then the currents in elements 1 and 2 are in-phase, since they are flowing in the same direction in space and are caused by the same voltage. However, the current in element 3 is flowing in the *opposite* direction in space because the voltage is applied to the opposite end of the element. The current in element 3 is therefore 180° out-of-phase with the currents in elements 1 and 2.

The phasing of driven elements depends on the direction of the element, the phase of the applied voltage, and the point at which the voltage is applied. In many systems used by amateurs, the voltages applied to the elements are exactly

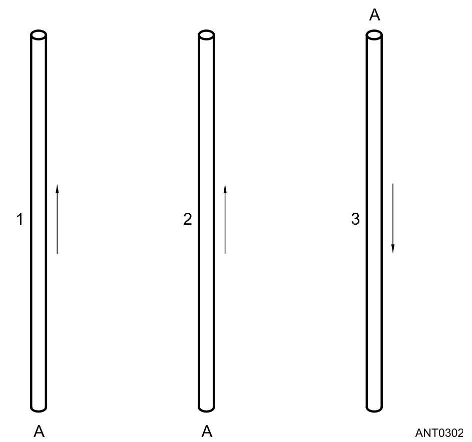


Fig 7—This drawing illustrates the phase of currents in antenna elements, represented by the arrows. The currents in elements 1 and 2 are in phase, while that in element 3 is 180° out of phase with 1 and 2.

in or exactly out-of-phase with each other. Also, the axes of the elements are nearly always in the same direction, since parallel or collinear elements are invariably used. The currents in driven elements in such systems therefore are usually either exactly in or exactly out-of-phase with the currents in other elements.

It is possible to use phase differences of less than 180° in driven arrays. One important case is where the current in one set of elements differs by 90° from the current in another set. However, making provision for proper phasing in such systems is considerably more complex than in the case of simple 0° or 180° phasing, as described in a later section of this chapter.

In parasitic arrays the phase of the currents in the parasitic elements depends on the spacing and tuning, as described later.

Ground Effects

The effect of the ground is the same with a directive antenna as it is with a simple dipole antenna. The reflection factors discussed in Chapter 3 may therefore be applied to the vertical pattern of an array, subject to the same modifications mentioned in that chapter. In cases where the array elements are not all at the same height, the reflection factor for the mean height of the array may be used for a close approximation. The mean height is the average of the heights measured from the ground to the centers of the lowest and highest elements.

MUTUAL IMPEDANCE

Consider two $\lambda/2$ elements that are fairly close to each other. Assume that power is applied to only one element, causing current to flow. This creates an electromagnetic field, which induces a voltage in the second element and causes current to flow in it as well. The current flowing in element 2 will in turn induce a voltage in element 1, causing additional current to flow there. The total current in 1 is then the sum (taking phase into account) of the original current and the induced current.

With element 2 present, the amplitude and phase of the resulting current in element 1 will be different than if element 2 were not there. This indicates that the presence of the second element has changed the impedance of the first. This effect is called *mutual coupling*. Mutual coupling results in a *mutual impedance* between the two elements. The mutual impedance has both resistive and reactive components. The actual impedance of an antenna element is the sum of its self-impedance (the impedance with no other antennas present) and its mutual impedances with all other antennas in the vicinity.

The magnitude and nature of the feed-point impedance of the first antenna depends on the amplitude of the current induced in it by the second, and on the phase relationship between the original and induced currents. The amplitude and phase of the induced current depend on the spacing between the antennas and whether or not the second antenna is tuned to resonance.

In the discussion of the several preceding paragraphs, power is applied to only one of the two elements. Do not interpret this to mean that mutual coupling exists only in parasitic arrays! It is important to remember that mutual coupling exists between any two conductors that are located near one another.

Amplitude of Induced Current

The induced current will be largest when the two antennas are close together and are parallel. Under these conditions the voltage induced in the second antenna by the first, and in the first by the second, has its greatest value and causes the largest current flow. The coupling decreases as the parallel antennas are moved farther apart.

The coupling between collinear antennas is comparatively small, and so the mutual impedance between such antennas is likewise small. It is not negligible, however.

Phase Relationships

When the separation between two antennas is an appreciable fraction of a wavelength, a measurable period of time elapses before the field from antenna 1 reaches antenna 2. There is a similar time lapse before the field set up by the current in number 2 gets back to induce a current in number 1. Hence the current induced in antenna 1 by antenna 2 will have a phase relationship with the original current in antenna 1 that depends on the spacing between the two antennas.

The induced current can range all the way from being completely in-phase with the original current to being completely out-of-phase with it. If the currents are in-phase, the total current is larger than the original current and the antenna feed-point impedance is reduced. If the currents are out-of-phase, the total current is smaller and the impedance is increased. At intermediate phase relationships the impedance will be lowered or raised depending on whether the induced current is mostly in or mostly out-of-phase with the original current.

Except in the special cases when the induced current is exactly in or out-of-phase with the original current, the induced current causes the phase of the total current to shift with respect to the applied voltage. Consequently, the presence of a second antenna nearby may cause the impedance of an antenna to be reactive—that is, the antenna will be detuned from resonance—even though its self-impedance is entirely resistive. The amount of detuning depends on the magnitude and phase of the induced current.

Tuning Conditions

A third factor that affects the impedance of antenna 1 when antenna 2 is present is the tuning of number 2. If antenna 2 is not exactly resonant, the current that flows in it as a result of the induced voltage will either lead or lag the phase it would have if the antenna were resonant. This causes an additional phase advance or delay that affects the phase of the current induced back in antenna 1. Such a phase lag has an effect similar to a change in the spacing between self-resonant antennas. However, a change in tuning is not exactly equivalent to a change in spacing because the two methods do not have the same effect on the amplitude of the induced current.

MUTUAL IMPEDANCE AND GAIN

The mutual coupling between antennas is important because it can have a significant effect on the amount of current that will flow for a given amount of power supplied. And it is the amount of *current* flowing that determines the field strength from the antenna. Other things being equal, if the mutual coupling between two antennas is such that the currents are greater for the same total power than would be the case if the two antennas were not coupled, the power gain will be greater than that shown in Table 1.

On the other hand, if the mutual coupling is such as to reduce the current, the gain will be less than if the antennas were not coupled. The term *mutual coupling*, as used in this paragraph, assumes that the mutual impedance between elements is taken into account, along with the added effects of propagation delay because of element spacing and element tuning or phasing.

The calculation of mutual impedance between antennas is a complex problem. Data for two simple but important cases are graphed in Figs 8 and 9. These graphs do not show the mutual impedance, but instead show a more useful quantity—the feed-point resistance measured at the center of an antenna as it is affected by the spacing between two antennas.

As shown by the solid curve in **Fig 8**, the feed-point resistance at the center of either antenna, when the two are self-resonant, parallel, and operated in-phase, decreases as the spacing between them is increased until the spacing is about 0.7 λ. This is a broadside array. The maximum gain is achieved from a pair of such elements when the spacing is in this region, because the current is larger for the same power and the fields from the two arrive in-phase at a distant point placed on a line perpendicular to the line joining the two antennas.

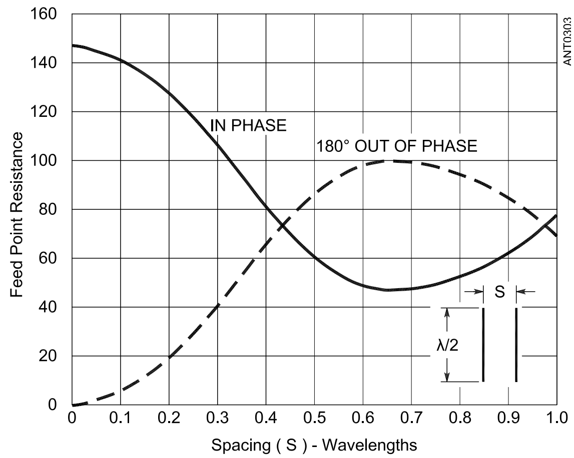


Fig 8—Feed-point resistance measured at the center of one element as a function of the spacing between two parallel ½-λ self-resonant antenna elements. For ground-mounted ¼-λ vertical elements, divide these resistances by two.

The dashed line in Fig 8, representing two antennas operated 180° out-of-phase (end-fire), cannot be interpreted quite so simply. The feed-point resistance decreases with spacing decreasing less than about 0.6 λ in this case. However, for the range of spacings considered, only when the spacing is 0.5 λ do the fields from the two antennas add up exactly in phase at a distant point in the favored direction. At smaller spacings the fields become increasingly out-of-phase, so the total field is less than the simple sum of the two. Smaller spacings thus decrease the gain at the same time that

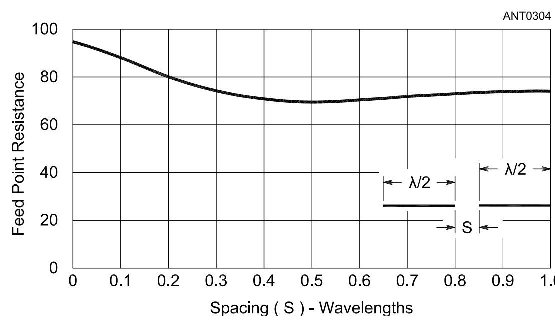


Fig 9—Feed-point resistance measured at the center of one element as a function of the spacing between the ends of two collinear self-resonant ½-λ antenna elements operated in phase.

the reduction in feed-point resistance is increasing it. For a lossless antenna, the gain goes through a maximum when the spacing is in the region of ⅛ λ.

The feed-point resistance curve for two collinear elements in-phase, **Fig 9**, shows that the feed-point resistance decreases and goes through a broad minimum in the region of 0.4 to 0.6-λ spacing between the adjacent ends of the antennas. As the minimum is not significantly less than the feed-point resistance of an isolated antenna, the gain will not exceed the gain calculated on the basis of uncoupled antennas. That is, the best that two collinear elements will give, even with optimum spacing, is a power gain of about 2 (3 dB). When the separation between the ends is very small—the usual method of operation—the gain is reduced.

GAIN AND ARRAY DIMENSIONS

The gain of an array is principally determined by the dimensions of the array, so long as there are a minimum number of elements. A good example of this is the relationship between boom length, gain and number of elements for an array such as a Yagi. **Fig 10** compares the gain versus boom length for Yagis with different numbers of elements. For given number of elements, notice that the gain increases as the boom length increases, up to a maximum. Beyond this point, longer boom lengths result in less gain for a given number of elements. This observation does not mean that it is always desirable to use only the minimum number of elements. Other considerations of array performance, such as front-to-back ratio, minor lobe amplitudes or operating bandwidth, may make it advantageous to use more than the minimum number of elements for a given array length. A specific example of this is presented in a later section in a comparison between a half-square, a bobtail curtain and a Bruce array.

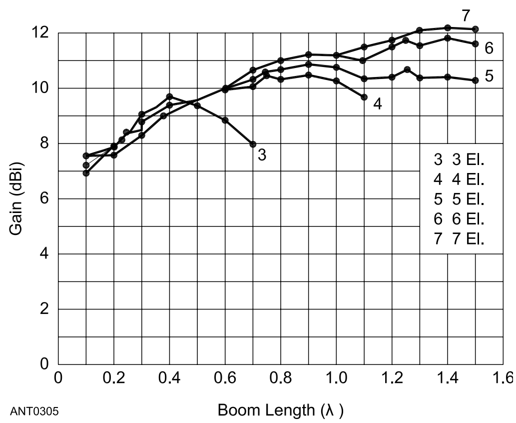


Fig 10—Yagi gain for 3, 4, 5, 6 and 7-element beams as a function of boom length. (From *Yagi Antenna Design*, J. Lawson, W2PV.)

In a broadside array the gain is a function of both the length and width of the array. The gain can be increased by adding more elements (with additional spacing) or by using longer elements (>λ/2), although the use of longer elements requires proper attention to current phase in the elements. In general, in a broadside array the element spacing that gives maximum gain for a minimum number of elements, is in the range of 0.5 to 0.7 λ. Broadside arrays with elements spaced for maximum gain will frequently have significant side lobes and associated narrowing of the main lobe beamwidth. Side lobes can be reduced by using more than the minimum number of elements, spaced closer than the maximum gain distance.

Additional gain can be obtained by expanding the array into a third dimension. An example of this is the stacking of endfire arrays in a broadside configuration. In the case of stacked short endfire arrays, maximum gain occurs with spacings in the region of 0.5 to 0.7 λ. However, for longer higher-gain end-fire arrays, larger spacing is required to achieve maximum gain. This is important in VHF and UHF arrays, which often use long-boom Yagis.

PARASITIC ARRAYS

The foregoing applies to multi-element arrays of both types, driven and parasitic. However, there are special considerations for driven arrays that do not necessarily apply to parasitic arrays, and vice versa. Such considerations for Yagi and quad parasitic arrays are presented in Chapters 11 and 12. The remainder of this chapter is devoted to driven arrays.

Driven Arrays

Driven arrays in general are either broadside or end-fire, and may consist of collinear elements, parallel elements, or a combination of both. From a practical standpoint, the maximum number of usable elements depends on the frequency and the space available for the antenna. Fairly elaborate arrays, using as many as 16 or even 32 elements, can be installed in a rather small space when the operating frequency is in the VHF range, and more at UHF. At lower frequencies the construction of antennas with a large number of elements is impractical for most amateurs.

Of course the simplest of driven arrays is one with just two elements. If the elements are collinear, they are always fed in-phase. The effects of mutual coupling are not great, as illustrated in Fig 9. Therefore, feeding power to each element in the presence of the other presents no significant problems. This may not be the case when the elements are parallel to each other. However, because the combination of spacing and phasing arrangements for parallel elements is infinite, the number of possible radiation patterns is endless.

This is illustrated in **Fig 11**. When the elements are fed in-phase, a broadside pattern always results. At spacings of less than ⅝ λ with the elements fed 180° out-of-phase, an end-fire pattern always results. With intermediate amounts of phase difference, the results cannot be so simply stated. Patterns evolve that are not symmetrical in all four quadrants.

Because of the effects of mutual coupling between the two driven elements, for a given power input greater or lesser currents will flow in each element with changes in spacing and phasing, as described earlier. This, in turn, affects the gain of the array in a way that cannot be shown merely by plotting the *shapes* of the patterns, as has been done in Fig 11. Therefore, supplemental gain information is also shown in Fig 11, adjacent to the pattern plot for each combination of spacing and phasing. The gain figures shown are referenced to a single element. For example, a pair of elements fed 90° apart at a spacing of λ/4 will have a gain in the direction of maximum radiation of 3.1 dB over a single element.

Current Distribution in Phased Arrays

In the plots of Fig 11, the two elements are assumed to be identical and self-resonant. In addition, currents of equal amplitude are assumed to be flowing at the feed point of each element, a condition that most often will not exist in practice without devoting special consideration to the feeder system. Such considerations are discussed in the next section of this chapter.

Most literature for radio amateurs concerning phased arrays is based on the assumption that if all elements in the array are identical, the *current distribution* in all the elements will be identical. This distribution is presumed to be that of a single, isolated element, or nearly sinusoidal. However, information published in the professional literature as early as the 1940s indicates the existence of dissimilar current distributions among the elements of phased arrays. (See Harrison and King references in the Bibliography.) Lewallen, in July 1990 *QST*, pointed out the causes and effects of dissimilar current distributions.

In essence, even though the two elements in a phased array may be identical and have exactly equal currents of the desired phase flowing *at the feed point*, the amplitude and phase relationships degenerate with departure from the feed point. This happens any time the phase relationship is not 0° or 180°. Thus, the field strengths produced at a distant point by the individual elements may differ. This is because the field from each element is determined by the *distribution* of the current, as well as its magnitude and phase.

The effects are minimal with shortened elements—verticals less than λ/4 or dipoles less than λ/2 long. The effects on radiation patterns begin to show at the above resonant lengths, and become profound with longer elements—λ/2 or longer verticals and 1 λ or longer center-fed elements. These effects are less pronounced with thin elements. The amplitude and phase degeneration takes place because the currents in the array elements are not sinusoidal. Even in two-element arrays with phasing of 0° or 180°, the currents are not sinusoidal, but in these two special cases they do remain identical.

The pattern plots of Fig 11 take element current distributions into account. The visible results of dissimilar distributions are incomplete nulls in some patterns and the development of very small minor lobes in others. For example, the pattern for a phased array with 90° spacing and 90° phasing has traditionally been published in amateur literature as a cardioid with a perfect null in the rear direction. Fig 11, calculated for 7.15-MHz self-resonant dipoles of #12 wire in free space, shows a minor lobe at the rear and only a 33-dB front-to-back ratio.

It is characteristic of broadside arrays that the power gain is proportional to the length of the array but is substantially independent of the number of elements used, provided the optimum element spacing is not exceeded. This means, for example, that a five-element array and a six-element array will have the same gain, provided the elements in both are spaced so the overall array length is the same. Although this principle is seldom used for the purpose of reducing the number of elements because of complications introduced in feeding power to each element in the proper phase, it does illustrate the fact that there is nothing to be gained, in terms of more gain, by increasing the number of elements if the space occupied by the antenna is not increased proportionally.

Generally speaking, the maximum gain in the smallest linear dimensions will result when the antenna combines both broadside and end-fire directivity and uses both parallel and collinear elements. In this way the antenna is spread over a greater volume of space, which has the same effect as extending its length to a much greater extent in one linear direction.

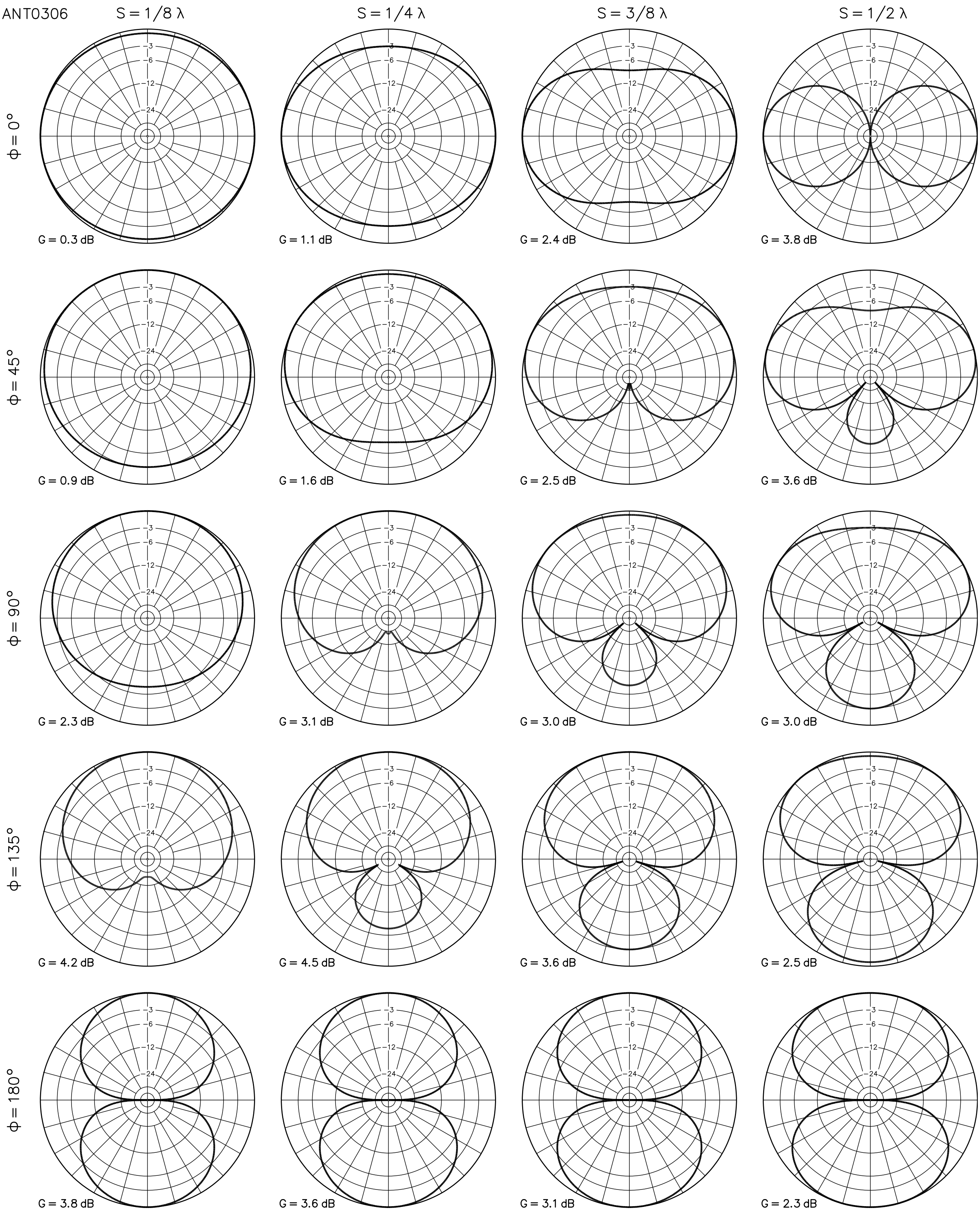
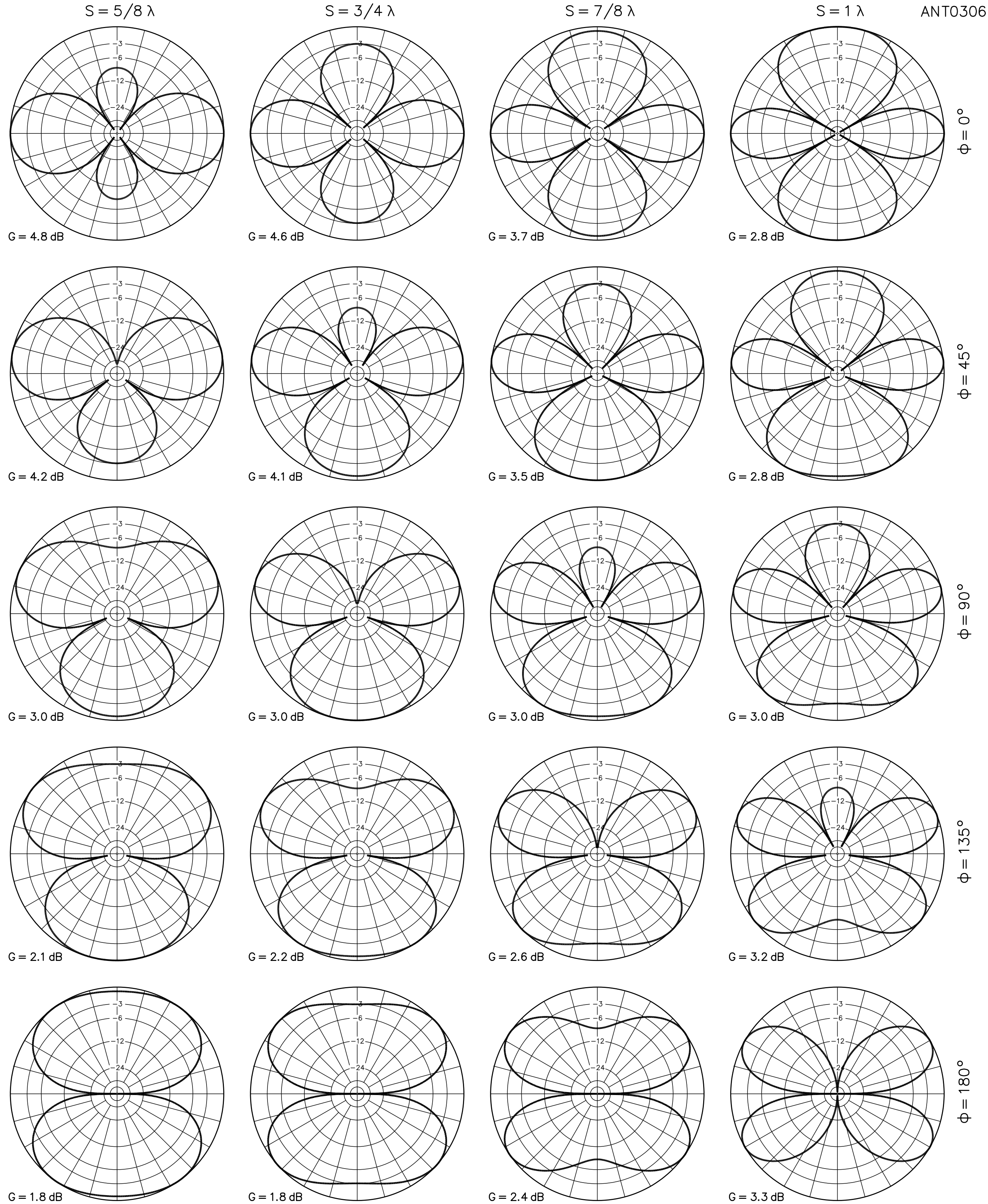


Fig 11—H-plane patterns of two identical parallel driven elements, spaced and phased as indicated (S = spacing, φ = phasing). The elements are aligned with the vertical (0°-180°) axis, and the element nearer the 0° direction (top of page) is of lagging phase at angles other than 0°. The two elements are assumed to be thin and self-resonant, with equal-amplitude currents flowing at the feed point. See text regarding current distributions. The gain figure associated with



each pattern indicates that of the array over a single element. The plots represent the horizontal or azimuth pattern at a 0° elevation angle of two ¼-λ vertical elements over a perfect conductor, or the free-space vertical or elevation pattern of two horizontal ½-λ elements when viewed on end, with one element above the other. (Patterns computed with *ELNEC*—see Bibliography.)

Phased Array Techniques

Phased antenna arrays have become increasingly popular for amateur use, particularly on the lower frequency bands, where they provide one of the few practical methods to obtain substantial gain and directivity. This section on phased-array techniques was written by Roy Lewallen, W7EL.

The operation and limitations of phased arrays, how to design feed systems to make them work properly and how to make necessary tests and adjustments are discussed in the pages that follow. The examples deal primarily with vertical HF arrays, but the principles apply to VHF/UHF arrays and arrays made from other element types as well.

OVERVIEW

Much of this chapter is devoted to techniques for feeding phased arrays. Many people who have a limited acquaintance with phased array techniques believe this is a simple problem, consisting only of connecting array elements through "phasing lines" consisting of transmission lines of the desired electrical lengths. Unfortunately, except for a very few special cases, this approach won't achieve the desired array pattern.

Other proposed universal solutions, such as hybrid couplers or Wilkinson or other power dividers, also usually fail to achieve the necessary phasing. These approaches sometimes produce—often more by accident than design—results good enough to mislead the user into believing that the simple approach is working as planned. Confusion can result when an approach fails to work in different circumstances. This section will explain why the simple solutions don't work as often thought, and how to design feed systems that do consistently produce the desired results.

Very briefly, the reason why the simple phasing-line approach fails is that the delay of current or voltage in a transmission line equals the line's electrical length only if the line is terminated in its characteristic impedance. And in phased arrays, element feed-point impedances are profoundly affected by mutual coupling.

Consequently, even if each element has the correct impedance when isolated, it won't when all elements are excited. Furthermore, transmission lines that are not terminated in their characteristic impedance will transform both the voltage and current magnitude. The net result is that the array elements will have neither the correct magnitudes nor phases of current necessary for proper operation except in a few special cases. This isn't a minor effect of concern only to perfectionists, but often a major one that causes significant pattern distortion and poor or mislocated nulls. The problem is examined in greater depth later.

Power dividers and hybrid couplers also fail to achieve the desired result for different reasons, which will be discussed below, although in one common application hybrid couplers fortuitously provide results that are acceptable to many users. This chapter will show how to design array feed systems that *will* produce predicted element currents and array patterns.

Various *EZNEC* models are provided to illustrate concepts presented in this chapter. They can all be viewed with the *EZNEC-ARRL* software furnished on the CD included with this book. Step-by-step instructions for the examples are given in **Appendix A**.

FUNDAMENTALS OF PHASED ARRAYS

The performance of a phased array is determined by several factors. Most significant among these are the characteristics of a single element, reinforcement or cancellation of the fields from the elements and the effects of mutual coupling. To understand the operation of phased arrays, it is first necessary to understand the operation of a single antenna element.

Of primary importance is the strength of the field produced by the element. The field radiated from a linear (straight) element, such as a dipole or vertical monopole, is proportional to the sum of the elementary currents flowing in each part of the antenna element. For this discussion it is important to understand what determines the current in a single element.

The amount of current flowing at the base of a ground mounted vertical or ground-plane antenna is given by the familiar formula

$$I = \sqrt{\frac{P}{R}} \qquad \text{(Eq 2)}$$

where

P is the power supplied to the antenna

R is the feed-point resistance.

R consists of two parts, the loss resistance and the radiation resistance. The loss resistance, R_L, includes losses in the conductor, in the matching and loading components and dominantly (in the case of ground-mounted verticals) in ground losses. The power "dissipated" in the radiation resistance, R_R, is actually the power that is radiated, so maximizing the power dissipated by the radiation resistance is desirable. However, the power dissipated in the loss resistance truly is lost as heat, so resistive losses should be made as small as possible.

The radiation resistance of an element can be derived from electromagnetic field theory, being a function of antenna length, diameter and geometry. Graphs of radiation resistance versus antenna length are given in Chapter 2. The radiation resistance of a thin resonant λ/4 ground-mounted vertical is about 36 Ω. A resonant λ/2 dipole in free space has a radiation resistance of twice this amount, about 73 Ω. Reducing the antenna lengths by one half drops the radiation resistances to approximately 7 and 14 Ω, respectively.

The radiation resistance of a large variety of antennas can easily be determined by using *EZNEC-ARRL*, which is included on the CD in the back of this book. The radiation

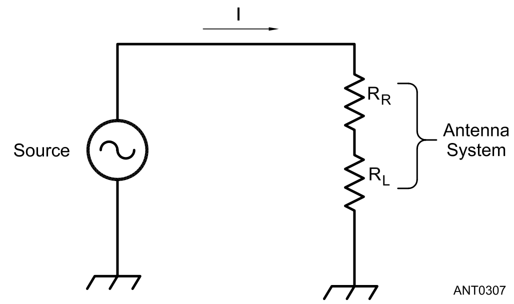


Fig 12—Simplified equivalent circuit for a single-element resonant antenna. R_R represents the radiation resistance, and R_L the ohmic losses in the total antenna system.

resistance is simply the feed-point resistance (the resistive part of the feed-point impedance) when all losses have been set to zero.

Radiation Efficiency

To generate a stronger field from a given radiator, it is necessary to increase the power P (the brute-force solution), decrease the loss resistance R_L (by putting in a more elaborate ground system for a vertical, for instance) or to somehow decrease the radiation resistance R_R so more current will flow with a given power input. This can be seen by expanding the formula for base current as

$$I = \sqrt{\frac{P}{R_R + R_L}} \qquad \text{(Eq 3)}$$

Splitting the feed-point resistance into components R_R and R_L easily leads to an understanding of element efficiency. The efficiency of an element is the proportion of the total power that is actually radiated. The roles of R_R and R_L in determining efficiency can be seen by analyzing a simple equivalent circuit, shown in **Fig 12**.

The power dissipated in R_R (the radiated power) equals I^2R_R. The total power supplied to the antenna system is

$$P = I^2 (R_R + R_L) \qquad \text{(Eq 4)}$$

so the efficiency (the fraction of supplied power that is actually radiated) is

$$\text{Eff} = \frac{I^2 R_R}{I^2 (R_R + R_L)} = \frac{R_R}{R_R + R_L} \qquad \text{(Eq 5)}$$

Efficiency is frequently expressed in percent, but expressing it in decibels relative to a 100%-efficient radiator gives a better idea of what to expect in the way of signal strength. The field strength of an element relative to a lossless but otherwise identical element, in dB, is

$$\text{FSG} = 10 \log \frac{R_R}{R_R + R_L} \qquad \text{(Eq 6)}$$

where FSG = field strength gain, dB.

For example, information presented by Sevick in March 1973 *QST* shows that a λ/4 ground-mounted vertical antenna with four 0.2-λ radials has a feed-point resistance of about 65 Ω (see the Bibliography at the end of this chapter). The efficiency of such a system is 36/65 = 55.4%. It is rather disheartening to think that, of 100 W fed to the antenna, only 55 W is being radiated, with the remainder literally warming up the ground. Yet the signal will be only 10 log (36/65) = –2.57 dB relative to the same vertical with a perfect ground system. In view of this information, trading a small reduction in signal strength for lower cost and greater simplicity may become an attractive consideration.

So far, only the current at the base of a resonant antenna has been discussed, but the field is proportional to the sum of currents in each tiny part of the antenna. The field is a function of not only the magnitude of current flowing at the feed point, but also the distribution of current along the radiator and the length of the radiator. Nothing can be done at the feed point to change the current distribution, so for a given element the field strength is proportional to the feed point current. However, changing the radiator length or loading it at some point other than the feed point will change the current distribution.

More information on shortened or loaded radiators may be found in Chapters 2 and 6 and in the Bibliography references of this chapter. The current distribution is also changed by mutual coupling to other array elements, although for most arrays this has only a minor effect on the pattern. This is discussed later in more detail. A few other important facts follow.

1) If there is no loss, the field from even an infinitesimally short radiator is less than ½ dB weaker than the field from a half-wave dipole or quarter-wave vertical. Without loss, all the supplied power is radiated regardless of the antenna length, so the only factor influencing gain is the slight difference in the patterns of very short and λ/2 antennas. The small pattern difference arises from different current distributions. A short antenna has a very low radiation resistance, resulting in a heavy current flow over its short length. In the absence of loss, this generates a field strength comparable to that of a longer antenna. Where loss is present—that is, in practical antennas—shorter radiators usually don't do so well, since the low radiation resistance leads to lower efficiency for a given loss resistance. If care is taken, reasonably short antennas can achieve good efficiency.

2) Care has to be taken in calculating the efficiency of folded antennas. (See Chapter 6.) Folding transforms both the radiation resistance and loss resistance by the same factor, so their ratio and therefore the efficiency remains the same. It's easy to show that in a ground-mounted vertical array, folding reduces the current flowing from the feed line to the ground system by a factor of two due to the impedance transformation. However, the folded antenna has an additional connection to ground, which also carries half the original ground current. The result is that the same amount of current flows

into the ground system, whether unfolded or folded, resulting in the same ground system loss. Analyses purporting to show otherwise invariably transform the radiation resistance but neglect to also transform the loss resistance and reach an incorrect conclusion.

3) The current flowing in an element with a given power input can be increased or decreased by mutual coupling to other elements. The effect is equivalent to changing the element radiation resistance. Mutual coupling is sometimes thought of as a minor effect, but often it is not minor!

Field Reinforcement and Cancellation

The mechanism by which phased arrays produce gain, and the role of mutual coupling in determining gain, were covered earlier in this chapter. One important point that can't be emphasized enough is that all antennas must abide by the law of *conservation of energy*. No antenna can radiate more power than supplied to it. The total amount of power it radiates is the amount it's supplied, less the amount lost as heat. This is true of all antennas, from the smallest "rubber ducky" to the most gigantic array.

Gain

Gain is strictly a relative measure, so the term is completely meaningless unless accompanied by a statement of just what it is relative to. One useful measure for phased array gain is *gain relative to a single similar element*. This is the increase in signal strength that would be obtained by replacing a single element by an array made from elements just like it. In some instances, such as investigating what happens to array performance when *all* elements become more lossy, it's useful to state gain relative to a more absolute, although unattainable standard: a lossless element.

And the most universal reference for gain is another unattainable standard, the *isotropic radiator*. This fictional antenna radiates absolutely equally in all directions. It's very useful because the field strength resulting from any power input is readily calculated, so if the gain relative to this standard is known, the field strength is also known for any radiated power. Gain relative to this reference is referred to as dBi, and it's the standard used by most modeling programs including *EZNEC-ARRL*. To find the gain of an array relative to a single element or other reference antenna such as a dipole, model both the array and the single element or other reference antenna in the same environment and subtract their dBi gains. Don't rely on some assumption about the gain of a single element—many people assume values that can be very wrong.

Nulls

Pattern nulls are very often more important to users of phased arrays than gain because of their importance in reducing both man-made and natural interference when receiving. Consequently, a good deal of emphasis is, and should be, placed on achieving good pattern nulls. Unfortunately, good nulls are much more difficult to achieve than gain and they are much more sensitive to array and feed-system imperfections.

As an illustration, consider two elements that each produce a field strength of, say, exactly 1 millivolt per meter (mV/m) at some distance many wavelengths from the array. In the direction in which the fields from the elements are in-phase, a total field of 2 mV/m results. In the direction in which they're out-of-phase, zero field results. The ratio of maximum to minimum field strength of this array is 2/0, or infinity.

Now suppose, instead, that one field is 10% high and the other 10% low—1.1 and 0.9 mV/m, respectively. In the forward direction, the field strength is still 2 mV/m, but in the canceling direction, the field will be 0.2 mV/m. The front-to-back ratio has dropped from infinity to 2/0.2, or 20 dB. (Actually, slightly more power is required to redistribute the field strengths this way, so the forward gain is reduced—but by only a small amount, less than 0.1 dB.) For most arrays, unequal fields from the elements have a minor effect on forward gain, but a major effect on pattern nulls. This is illustrated by ***EZNEC* Example: Nulls** in Appendix A.

Even with perfect current balance, deep nulls aren't assured. **Fig 13** shows the minimum spacing required for total field reinforcement or cancellation. If the element spacing isn't adequate, there may be no direction in which the fields are completely out-of-phase (see curve B of Fig 13). Slight physical and environmental differences between elements will invariably affect null depths, and null depths will also vary with elevation angle.

However, a properly designed and fed array can produce very impressive nulls. The key to producing good nulls, like producing gain, is controlling the strengths and phases of the fields from the elements. Just how to accomplish that is the subject of most of the remainder of this section. But be sure to keep in mind that producing good nulls is generally a much more difficult task than producing approximately the predicted gain.

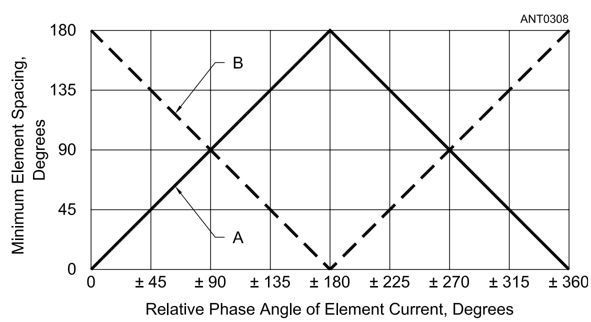


Fig 13—Minimum element spacing required for total field reinforcement, curve A, or total field cancellation, curve B. Total cancellation results in pattern nulls in one or more directions. Total reinforcement does not necessarily mean there is gain over a single element, as the effects of loss and mutual coupling must also be considered.

Mutual Coupling

Mutual coupling was discussed briefly earlier in this chapter. Because it has an important and profound effect on both the performance and feed system design of phased arrays, it will be covered in greater depth here.

Mutual coupling refers to the effects which the elements in an array have on each other. Mutual coupling can occur intentionally or entirely unintentionally. People with multiple antennas on a small lot (or car top) often discover that a better description of their system is a single antenna with multiple feed points. Current is induced in conductors in various antennas by mutual coupling, causing them to act like parasitic elements, which re-radiate and distort the antenna's pattern. The effects of mutual coupling are present whether or not the elements are driven.

Suppose that two driven elements are many wavelengths from each other. Each has some voltage and current at its feed point. For each element, the ratio of this voltage to current is the element self-impedance. If the elements are brought close to each other, the current in each element will change in magnitude and phase because of coupling with the field from the other element. The field from the first element changes the current in the second. This changes the field from the second, which alters the current in the first, and so forth until an equilibrium condition is reached in which the currents in all elements (hence, their fields) are totally interdependent.

The feed-point impedances of all elements also are changed from their values when far apart, and all are dependent on each other. In a driven array, the changes in feed-point impedances can cause additional changes in element currents, because the operation of many feed systems depends on the element feed-point impedances. Significant mutual coupling occurs at spacings as great as a wavelength or more.

Connecting the elements to a feed system to form a driven array does not eliminate the effects of mutual coupling. In fact, in many driven arrays the mutual coupling has a greater effect on antenna operation than the feed system does. All feed-system designs must account for the impedance changes caused by mutual coupling if the desired current balance and phasing are to be achieved.

Several general statements can be made regarding the effects of mutual coupling on phased-array systems.

1) The resistances and reactances of all elements of an array generally will be substantially different from the values when the elements are isolated (that is, very distant from other elements).

2) If the elements of a two-element array are identical and have equal currents that are in-phase or 180° out-of-phase, the feed-point impedances of the two elements will be equal. But they will be different than for an isolated element. If the two elements are part of a larger array, their impedances can be very different from each other.

3) If the elements of a two-element array have currents that are neither in-phase (0°) nor out-of-phase (180°), their feed-point impedances will not be equal. The difference will be considerable in typical amateur arrays.

4) The feed-point resistances of the elements in a closely spaced, 180° out-of-phase array will be very low, resulting in poor efficiency due to ohmic losses unless care is taken to minimize loss. This is also true for any other closely spaced array with significant predicted gain.

It's essential to realize that this is *not* a minor effect and one that can be overlooked or ignored. See ***EZNEC* Example—Mutual Coupling** in Appendix A for an illustration of these phenomena.

Loss Resistance, Mutual Coupling and Antenna Gain

Loss reduces the effects of mutual coupling because the feed-point impedance change resulting from mutual coupling is effectively in series with loss resistance. If the loss is great enough, two important results occur. First, the feed-point impedance becomes independent of the presence of nearby current-carrying elements. This greatly simplifies feed system design—the simple "phasing-line" or hybrid-coupler feed system described below is adequate *provided that all elements are physically identical and the feed point of each element is matched to the Z_0 of the feed line and, if used, the hybrid coupler.*

The impedance matching restrictions are necessary to insure that the phasing line or hybrid coupler performs as expected. Identical elements are needed so that equal element currents will result in equal fields from the elements.

In the absence of mutual coupling effects, the maximum gain of an array of identical elements relative to a single (similarly lossy) element is simply 10 log(N), where N is the number of elements—providing that spacing is adequate for the fields to fully reinforce in some direction. If spacing is less, maximum gain will also be less. Of course, the array gain relative to a single *lossless* element will be very low, most likely a sizeable negative number when expressed in dB. So intentionally introducing loss isn't a wise idea for a transmitting array. It is sometimes an advantageous thing to do for a receiving array, however, as explained in the following section.

High-gain close-spaced arrays, such as the W8JK phased array (see *EZNEC-ARRL* example file **ARRL_W8JK.EZ** and accompanying Antenna Notes file), and most parasitic arrays depend heavily on mutual coupling to achieve their gain. Introduction of any loss to these arrays, which reduces the mutual coupling effects, has a profound effect on the gain. Consequently, parasitic or close-spaced driven arrays often produce disappointing results when made from grounded vertical elements unless each has a fairly elaborate (and therefore very low-loss) ground system.

If you place two low-loss elements very close together and feed them in-phase, mutual coupling reduces the array gain to essentially that of a single element, so there's no advantage to this configuration over a single element. However, if you have a single lossy element, for example a short vertical having a relatively poor ground system, you can improve the gain by up to 3 dB by adding a second, close spaced, element and ground system and feeding the two in-phase. Another way to look at this technique is that you're putting two equal ground system resistances in parallel, which effectively cuts the loss in half. The gain you can realize in practice depends on such things as the ground system overlap, but it might be a practical way to improve transmitting array performance in some situations.

FEEDING PHASED ARRAYS

The previous section explains why the fields from the elements must be very close to the ratios required by the array design. Since the field strengths are proportional to the currents in the elements, controlling the fields requires controlling the element currents. Since the desired current ratio is 1:1 for virtually all two-element and for most (but not all) larger amateur arrays, special attention is paid to methods of assuring equal element currents. But we will examine other current ratios also.

The Role of Element Currents

The field from a conductor is proportional to the current flowing on it. So if we're to control the relative strengths and phases of the fields from the elements, we have to control their currents. We usually do this by controlling the currents at the element feed points. But because the field from an element depends on the current everywhere along the element, elements having identical feed-point currents will produce different fields if they have different current distributions—that is, if the way the current varies along the lengths of the elements is different.

A previous section explained that mutual coupling alters the current distribution, so in many arrays the current distributions will be different on the elements and consequently the relationship between the overall fields won't be the same as that between the feed-point currents. Fortunately, this effect is relatively minor in thin, λ/4 monopole or λ/2 dipole elements. The most common arrays are made from elements in this category, so we can generally get very nearly the desired ratio of fields by effecting the same ratio of feed point currents. Exceptions are detailed immediately below.

Feed-point vs Element Current

For most antennas, environmental factors are likely to cause greater performance anomalies than current distribution differences, and both can be corrected with minor feed system adjustments. The difference between field and feed-point current ratios can become very significant, however, if the elements are very fat and/or close to λ/2 (monopole) or 1 λ (dipole) long. In those cases, most of the feed systems described here won't produce the desired field ratios without major adjustment or modification, except in the special cases of 2-element arrays with identical elements having feed point currents in-phase or 180° out-of-phase. In those special cases, the element current distributions are the same for the same reason the feed-point impedances are equal. This is explained later in the feed system sections.

To get an idea of just how large an element must be to disturb the pattern of an array with correct feed-point currents, a two-element cardioid array of quarter wave vertical elements was modeled at 10 MHz. With thin, 0.1-inch diameter elements, the front/back ratio was 35 dB, the very small reverse lobe caused by slightly unequal element current distributions. Increasing the element diameter to 20 inches decreased the front/back ratio to 20 dB. Returning the front/back ratio of the array of 20-inch elements to >35 dB required changing the feed-point current ratio from the nominal value of 1.0 at an angle of 90° to 0.88 at 83°.

The same array was first modeled with 0.1 inch diameter elements, where it has a front/back ratio of 35 dB, then the elements were lengthened. The front/back ratio dropped to 20 dB at an element length of 36 feet, or about 0.37 λ. In that case, adjustment of the feed-point current ratio to about 0.9 at about 83° restored a good front/back ratio.

In the discussion and development which follow, the assumption is made that the fields will be very nearly proportional to the feed-point currents. If the elements are fat or long enough to make this assumption untrue, some adjustment of feed-point current ratio will be necessary to achieve the desired pattern, particularly nulls. Most feed systems can be designed for any current ratio. Modeling will reveal the ratio required for the desired pattern, and then the feed system can be designed accordingly.

COMMON PHASED-ARRAY FEED SYSTEMS

This section will first describe several popular approaches to feeding phased arrays that often don't produce the desired results. It will describe why they don't work as well as hoped. It also briefly discusses systems that could be used, but that often aren't appropriate or optimum for amateur arrays.

This will be followed in the next section by detailed descriptions of array feed systems that *do* produce the predicted element current ratios and array patterns.

The "Phasing-Line" Approach

For an array to produce the desired pattern, the element currents must have the required magnitude and the required phase relationship. As explained above, this can generally be achieved well enough by causing the feed point currents to have that same relationship.

On the surface, this sounds easy—just make sure that the difference in electrical lengths of the feed lines to the elements equals the desired phase angle. Unfortunately, this approach doesn't necessarily achieve the desired result. The first problem is that the phase shift through the line is not equal to its electrical length. The current (or, for that matter, voltage) delay in a transmission line is equal to its electrical length in only a few special cases—cases which don't exist in most amateur arrays! The impedance of an element in an array is frequently very different from the impedance of an isolated element and the impedances of all the elements in an array can be different from each other.

See the ***EZNEC* Example—Mutual Coupling** in Appendix A for a graphic illustration of the effect of mutual coupling on feed-point impedance. Also look at the Four-Square array example in the **Phased Array Design Examples** section. The array in that example has one element with a *negative* feed-point resistance, if ground loss is low. Without mutual coupling, the resistance of that same element would be about 36 Ω plus ground loss.

Because of mutual coupling, the elements seldom provide a matched load for the element feed lines. The effect of mismatch on phase shift can be seen in **Fig 14**. Observe what happens to the phase of the current and voltage on a line terminated by a purely resistive impedance that is lower than the characteristic impedance of the line (Fig 14A). At a point 45° from the load the current has advanced less than 45°, and the voltage more than 45°. At 90° from the load both are advanced 90°. At 135° the current has advanced more and the voltage less than 135°. This apparent slowing down and speeding up of the current and voltage waves is caused by interference between the forward and reflected waves. It occurs on any line that is not terminated with a pure resistance equal to its characteristic impedance. If the load resistance is greater than the characteristic impedance of the line, as shown in Fig 14B, the voltage and current exchange angles. Adding reactance to the load causes additional phase shift. The *only cases in which the current (or voltage) delay is equal to the electrical length of the line are*

1) When the line is flat; that is, terminated in a purely resistive load equal to its characteristic impedance;

2) When the line length is an integral number of half wavelengths;

3) When the line length is an odd number of quarter wavelengths and the load is purely resistive; and

4) When other specific lengths are used for specific load impedances.

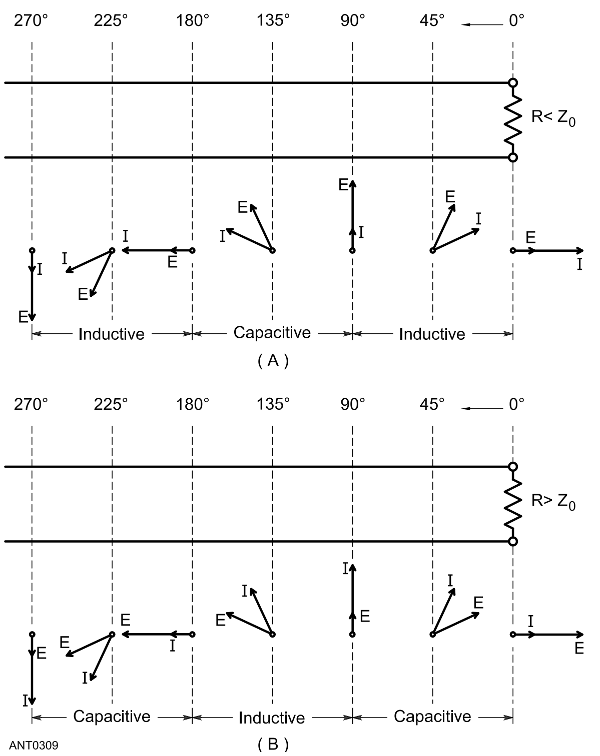


Fig 14—Resultant voltages and currents along a mismatched line. At A, R less than Z_0, and at B, R greater than Z_0.

Just how much phase error can be expected if two feed lines are simply hooked up to form an array? There is no simple answer. Some casually designed feed systems might deliver satisfactory results, but most will not. See the ***EZNEC* Example—"Phasing-Line" Feed** in Appendix A for the typical consequences of using this sort of feed system.

A second problem with simply connecting feed lines of different lengths to the elements is that the lines will change the *magnitudes* of the currents. The magnitude of the current (or voltage) out of a line does not equal the magnitude into that line, except in cases 1, 2 and 4 above. The feed systems presented later in this chapter assure currents that are correct in both magnitude and phase.

The elementary phasing-line approach *will* work in three very special but common situations. If the array consists of only two identical elements and those elements are fed in-phase, mutual coupling will modify the element impedances, but both will be modified exactly the same amount. Consequently, if the two elements are fed through equal-length transmission lines, the lines will transform and delay the currents by the same amount and result in equal, in-phase currents at the element feed points.

Similarly, an array of two identical elements fed 180º out-of-phase will have the same feed-point impedances and can be fed with two lines of any length so long as one line is an electrical half wavelength longer than the other. But this can't be extended to any two elements in a larger array, since mutual coupling to the other elements can result in different feed-point impedances. Methods will be described later which do assure a correct current ratio in this situation.

The third application in which the phasing-line approach works is in receiving arrays where the elements are very short in terms of wavelength and/or very lossy. In either of these cases, mutual coupling between elements is much less than an element's self-impedance. This allows the elements to be individually matched to the feed lines, with no significant change taking place when the elements are formed into an array. Under those conditions, the transmission lines can be matched and the lines used as simple delay lines with easily predictable phase shift and with no transformation of current or voltage magnitude other than cable loss. This is discussed in the later section on receiving antennas.

Many arrays *can* be correctly fed with a feed system consisting of only transmission lines, but the technique requires knowledge of the element feed-point impedances in a correctly fed array. Line lengths can then be computed that provide the correct ratio of currents into those particular load impedances. The line lengths generally differ by amount that's considerably different from the element phase angle difference, and appropriate line lengths can't always be found for all arrays. This technique is described more fully in the **"The Simplest Phased Array Feed System—That Works"** section later in the chapter and illustrated in the examples in **Phased Array Design Examples**.

The Wilkinson Divider

The *Wilkinson divider*, sometimes called the *Wilkinson power divider*, was once heavily promoted as a means to distribute power among the elements of a phased array. While it's a very useful device for other purposes, it won't produce the desired current ratios in antenna elements. In most phased arrays, element feed-point resistances are different and therefore require different amounts of power to achieve the desired equal magnitude currents. (See the section on mutual coupling above.) A Wilkinson divider is intended to deliver equal powers, not currents, to multiple loads. And it won't even do that when the load impedances are different.

The Hybrid Coupler

Hybrid couplers are promoted as solving the problem of achieving equal magnitude currents with a 90º phase difference between elements. Unfortunately, they provide equal magnitude, quadrature (90° phased) currents only when the load impedances are equal and correct. And this simply isn't true of arrays with quadrature-fed elements, except for arrays consisting of short and/or lossy elements, usually suitable only for receiving. In those arrays, the hybrid coupler can be useful for the same reasons as the phasing-line approach, discussed in an earlier section.

At the time of this writing, hybrid couplers are being used in a popular commercial product for phasing at least one type of array. Reports are that it works satisfactorily. However, this shouldn't be taken as proof that the hybrid coupler forces equal magnitude 90º phased currents in loads of arbitrary impedances. No passive network, including the hybrid coupler, is capable of doing that. See **The "Magic Bullet"** below for more information.

Large Array Feed Systems

The author once worked on a radar system where the transmit array consisted of over 5,000 separate dipole elements and the receive array over 4,000 pairs of crossed dipoles, all over a metal reflecting plane, which was the sloping side of a 140 foot high building. In such large arrays, each element is in essentially the same environment as every other element except near the array edges, so almost all elements have very nearly the same feed-point impedance. While producing the phase shifts and magnitude tapers is a considerable mathematical challenge, the problem of unequal element feed-point impedances can largely be ignored. Consequently, feed methods for these large arrays are generally not suitable for typical amateur arrays of a few elements.

The Broadcast Approach

Networks can be designed to transform the element base impedances from their values in an excited array to, say, 50 Ω resistive. Then another network can be inserted at the junction of the feed lines to properly divide the power among the elements (not necessarily equally!). And finally, additional networks must be added to correct for the phase shifts and magnitude transformations of the other networks. This general approach is used by the broadcast industry, in installations that are typically adjusted only once for a particular frequency and pattern.

Although this technique can be used to correctly feed any type of array, design is difficult and adjustment is tedious, since all adjustments interact. When the relative currents and phasings are adjusted, the feed-point impedances change, which in turn affect the element currents and phasings, and so on. A further disadvantage of using this method is that switching the array direction is generally impossible. Information on applying this technique to amateur arrays can be found in Paul Lee's book, listed in the Bibliography.

The "Magic Bullet"

For about 15 years prior to this writing, this *Antenna Book* section has contained a specification for a hypothetical passive circuit that would provide equal-magnitude, 90° phased currents into two loads without respect to the load impedances. This would be a circuit to which we could connect any two elements and guarantee that they'd have exactly the correct currents.

Along with the specifcation was a request that any person knowing of such a circuit would contact the author (Roy Lewallen, W7EL) or the book editors. During this time, only a single response was received, in 1996. It was from Kevin Schmidt, W9CF, who has formulated a mathematical proof that such a circuit—in fact, one resulting in any relative phase other than 0º or 180º—*cannot exist* if restricted to reciprocal elements. (That is, it can't exist unless directional components such as ferrite circulators are used.) This means that, in order to design a network to feed elements at any other phase angle other than 0 or 180º, we *must* know the impedance of at least one element, and correct feed system operation depends on that impedance. There's no way around this requirement. At the time of this writing, Schmidt's proof can be found at **fermi.la.asu.edu/w9cf/articles/magic/index.html**.

RECOMMENDED FEED METHODS FOR AMATEUR ARRAYS

The following feed methods are able to produce element feed-point currents having a desired magnitude and phase relationship, resulting in desirable and predictable patterns. Most methods require knowing the feed-point impedance of one or more array elements *when the array element currents are the correct values*. This isn't possible to measure directly, because if the element currents were correct, the feed system would already be working properly and no further design would be necessary.

By far the easiest way to get this information, if possible, is by computer modeling. Modeling programs such as *EZNEC-ARRL* (included on the CD in the back of this book) allow you to construct an ideal array with perfect element currents, then look at the resulting feed-point impedances. Because of its simplicity and versatility, this approach is highly recommended and it's the one used for the array design examples in this chapter.

Some feed systems allow adjustment, so even an approximate result provides an adequate starting point on which to base the feed-system design. There are several other alternatives to computer modeling. One is to first eliminate the effects of coupling of the element to be measured from all other elements, usually by open circuiting the feed points of the other elements. Then the feed-point impedance of the element is measured. Next, the impedance change due to mutual coupling from all other elements has to be calculated, based on the intended currents in the other elements, their lengths and their distances from the element being measured. Mutual impedance (which is not the same as the impedance *change* due to mutual coupling) between each pair of elements must be known for this calculation and it can be determined by measurement, calculation or from a graph.

The latter two methods are possible only for the simplest element types and measurement is very difficult to do accurately because it involves resolving very small differences between two relatively large values. Accuracy of a calculated result will be reduced if any elements are relatively fat (that is, they have a large diameter, because this impacts the current distribution) or they aren't perfectly straight and parallel.

So the only situations where you're likely to get good results from approaches other than modeling are the very easiest ones to model! And modeling allows determination of the feed-point impedances of many antennas that areimpossible to calculate by manual or graphical methods. Therefore, the manual approach isn't discussed or used here. Appendix B, on the CD, contains equations and manual techniques from previous editions of *The ARRL Antenna Book*, for those who are interested. You can also find a great deal of additional information in many of the texts listed in the Bibliography, particularly Jasik and Johnson.

Current Forcing with λ/4 Lines—Elements In-Phase or 180° Out-of-Phase

The feed method introduced here has been used in its simplest form to feed television receiving antennas and other arrays, as presented by Jasik, pages 2-12 and 24-10 or Johnson, on his page 2-14. However, until first presented in *The ARRL Antenna Book*, this feed method was not widely applied to amateur arrays.

The method takes advantage of an interesting property of λ/4 transmission lines. (All references to lengths of lines are electrical length and lines are assumed to have negligible loss.) See **Fig 15**. The magnitude of the *current out* of a λ/4 transmission line is equal to the *input voltage* divided by the characteristic impedance of the line. This is independent of the load impedance. In addition, the phase of the output current lags the phase of the input voltage by 90°, also independent of the load impedance. These properties can be used to advantage in feeding arrays with certain phase angles between elements.

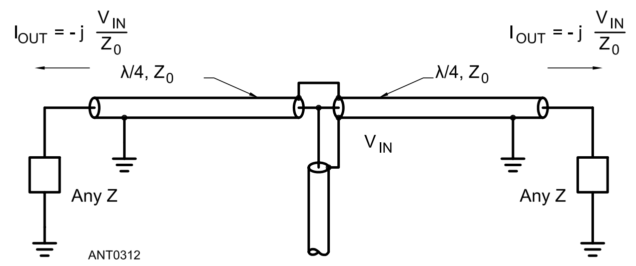


Fig 15—A useful property of λ/4 transmission lines; see text. This property is utilized in the "current-forcing" method of feeding an array of coupled elements.

If any number of loads are connected to a common driving point through λ/4 lines of equal impedance, the currents in the loads will be *forced* to be equal and in-phase, regardless of the load impedances. So any number of in-phase elements can be correctly fed using this method, regardless of how their impedances might have been changed by mutual coupling. Arrays that require unequal currents can be fed through λ/4 lines of unequal impedances to achieve other current ratios.

The properties of λ/2 lines also are useful. Since the current out of a λ/2 line equals the input current shifted 180°, regardless of the load impedance, any number of half wavelengths of line may be added to the basic λ/4, and the current and phase forcing property will be preserved. For example, if one element is fed through a λ/4 line and another element is fed from the same point through a 3λ/4 line of the same characteristic impedance, the currents in the two elements will be forced to be equal in magnitude and 180° out-of-phase, regardless of the feed-point impedances of the elements.

If an array of two, and only two, identical elements is fed in-phase or 180° out-of-phase with equal magnitude currents, both elements have the same feed-point impedance. The reason is that each element sees exactly the same thing when looking at the other. In an in-phase array, each sees another element with an identical current; in an out-of-phase array, each sees another element with an equal magnitude current that's 180° out-of-phase, the same distance away in both cases. This isn't true in something like a 90° fed array, where one element sees another with a current leading its current by 90°, while the other sees another element with a lagging current.

With arrays fed in-phase or 180° out-of-phase, feeding the elements through equal lengths of feed line (in-phase) or lengths differing by 180° (out-of-phase) will lead to the correct current magnitude ratio and phase difference, regardless of the line length and regardless of how much the element feed-point impedances depart from the lines' Z_0.

Unless the feed-point impedances equal the line Z_0 or the lines are an integral number of half wavelengths long, the magnitudes of the currents out of the lines will not be equal to the input magnitudes, and the phase will not be shifted an amount equal to the electrical lengths of the lines. But both lines will produce the same transformation and phase shift because their load impedances are equal, resulting in a properly fed array. In practice, however, feed-point impedances of elements frequently are different even in these arrays, because of such things as different ground systems (for ground mounted vertical elements), proximity to buildings or other antennas, or different heights above ground (for horizontal or elevated vertical elements).

In many larger arrays, two or more elements must be fed either in-phase or out-of-phase with equal currents, but coupling to other elements can cause their impedances to change unequally—sometimes extremely so. Using the current-forcing method allows the feed system designer to ignore all these effects, while guaranteeing equal and correctly phased currents in any combination and number of 0° and 180° fed elements.

This method is used to develop feed systems for the Four Square and 4-element rectangular arrays in the **Practical Array Design** section. The front and rear elements of a Four-Square antenna provide a good example of elements having very different feed-point impedances that are forced to have equal out-of-phase currents.

"The Simplest Phased Array Feed System—That Works"

This is the title of an article in *The ARRL Antenna Compendium*, *Vol 2*, which describes how arrays can be fed with a feed system consisting of only transmission lines. (The article is available for viewing at **eznec.com/Amateur/Articles/Simpfeed.pdf** and is also on the CD included in the back of this book, along with the program *Arrayfeed1*, which solves the equations presented in the article.)

As explained earlier in the Phasing Line section, this method requires knowing what the element feed-point impedances will be in a correctly fed array. Feed-line lengths can then be computed, for most but not all arrays. These lengths will produce the desired current ratio in array elements that do present those feed-point impedances. If you know the load impedances connected to transmission lines whose inputs are connected to a common source, it's simple to calculate the resulting load currents for any transmission line lengths. However, the reverse problem is much more difficult; that is, given the load impedances and desired currents to calculate the required cable lengths.

One way to solve the problem is to choose some feed-line lengths, solve for the currents, examine the answer, adjust the feed-line lengths, and try again until the desired currents are obtained. The author used this iterative approach, using first a programmable calculator and later a computer, for some time before developing a direct way of solving for the transmission-line lengths. The direct solution method is described briefly in the *Compendium* article.

Fig 16 shows the basic so-called "simplest" system applied to a two element array. Although it resembles an elementary phasing-line system as described earlier, the critical difference is that the lengths of Lines 1 and 2 are calculated to provide the correct current relative magnitude transformation and phase shift when terminated with the actual feed-point impedances.

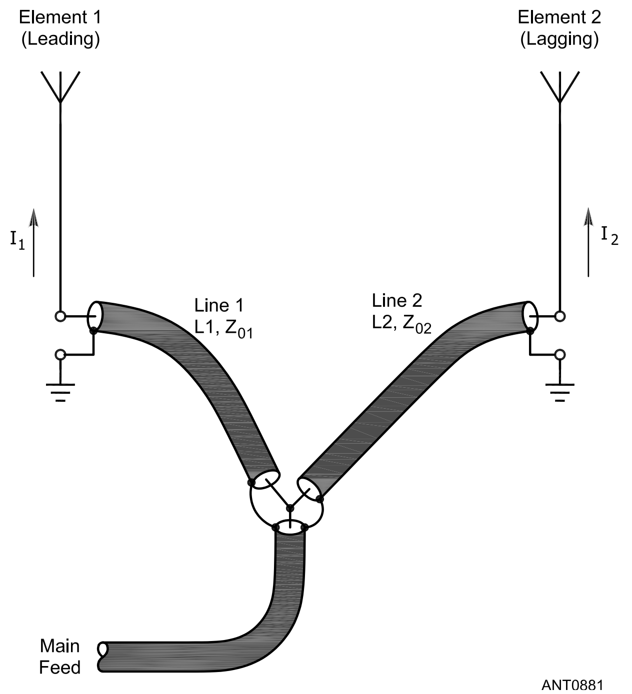


Fig 16—"Simplest" feed system for 2-element array. No matching or phasing network is used here, only transmission lines.

The advantage to using this "simplest" feed system is indeed its simplicity. It's no more complicated than the elementary phasing-line approach but actually works as planned. The disadvantage over some other methods is that there's no convenient adjustment to compensate for environment factors, array imperfections or inaccurately known feed-point impedances.

Also, while unusual, it's possible that no suitable feedline lengths can be found for some arrays, or at least none with practical feed-line characteristic impedances. The difference in electrical feed-line lengths almost never equals the difference in phase angles between element currents. This is because of the different line delays resulting from different feed-point impedances.

Program *Arrayfeed1*, included on the CD included with this book, can do the calculations for any two elements (alone or in a larger array), a Four-Square array or a rectangular array in which two in-phase elements are driven at any current magnitude and phase relative to the other two in-phase elements. These possibilities cover a large number of common arrays.

Arrayfeed1 can also be applied to other types of arrays using the method described in the **Feeding Larger Arrays** section in Appendix B (on the CD). The required knowledge of element feed-point impedances in a correctly fed array can be obtained using *EZNEC-ARRL*, also included on the CD. Examples of the design of a "simplest" feed system for several different arrays using *EZNEC-ARRL* and *Arrayfeed1* can be found in the **Phased Array Design Examples** section.

When a solution is possible for a given choice of line characteristic impedances, a second solution with different lengths is always available. See the comments in the introductory part of the **Phased Arrray Design Examples** section about choosing the solution to use.

An Adjustable L-Network Feed System

Adjustment of the current ratio of any two elements requires varying two independent quantities; for example the magnitude and phase of the current ratio. Two degrees of freedom—adjustments that are at least partially independent—are required. The "simplest" all-transmission line feed system described earlier adjusts the lengths of the two transmission lines to achieve the correct ratio.

But if the antenna characteristics aren't well known—for example, if the ground resistance isn't known even approximately—then the initial "simplest" design won't be optimum and adjustment can be difficult and tedious. The current-forcing method produces correct currents independently of the element characteristics, so it doesn't require adjustment as long as the elements are identical. But it's suitable only for feeding elements in-phase or 180° out-of-phase and a few fixed current-magnitude ratios.

The addition of a simple network as shown in **Fig 17** allows you to easily adjust feeding of element pairs at other relative phase angles and/or magnitude ratios. Any desired current ratio (magnitude and phase) can be obtained with two elements fed with any lengths of wire, equal or unequal, by adding a network.

However, calculations for the general case are complex. The problem becomes much simpler if the transmission lines are restricted to lengths of odd multiples of λ/4, forming a modified "forcing" system that includes an added network. There are at least three additional advantages of this scheme. One is that a λ/4 line is easy to measure, even if the velocity factor isn't known. This is described in the **Practical Aspects of Phased Array Design** section.

Second is that the feed system becomes completely insensitive to the feed-point impedance of one of the two elements. And the third is that the transmission lines of "forcing" systems feeding groups of elements in larger arrays can be used in place of the normal λ/4 lines. This greatly simplifies both the design of feed systems for larger arrays and the feed systems themselves. Note that both lines can be changed to 3λ/4 if necessary to span the physical distance between elements, but *both* lines must be the same 3λ/4 length.

This basic feed method can be used for any pair of elements, or for two groups of elements having forced equal currents. (See **Feeding Four Element and Larger Arrays** below.) Many networks can accomplish the desired function, but a simple L network is adequate for most feed systems. The network can be designed to produce a phase lead or phase lag. The basic two-element L-network feed system is shown in Fig 17. Many variations of this general method can be used, but the equations, program, and method to be discussed here apply only to the feed system shown.

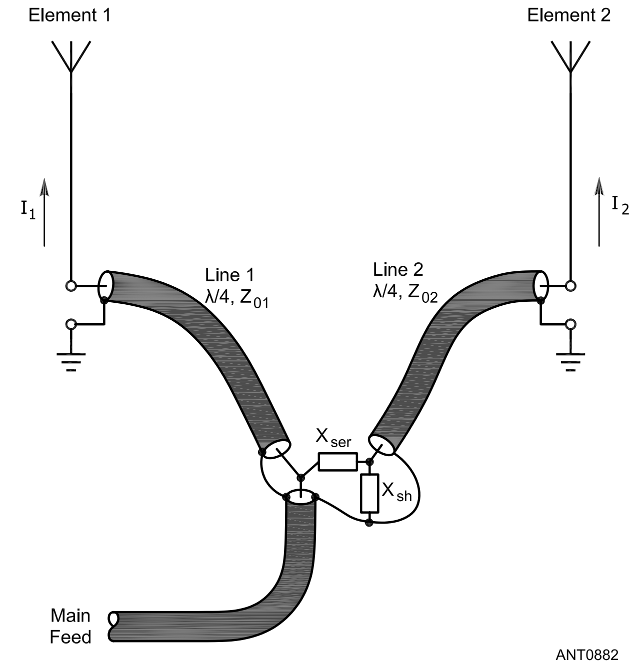


Fig 17—The addition of a simple L-network to Fig 16 allows you to easily adjust feeding of element pairs at other relative phase angles and/or magnitude ratios.

If the phase angle of I_2/I_1 is negative (element 2 is lagging element 1), the L network will usually resemble a low-pass network (X_{ser} is an inductor and X_{sh} is a capacitor). But if the phase angle is positive (element 2 lagging element 1), the L network will resemble a high-pass network (X_{ser} is a capacitor and X_{sh} is an inductor). However, some current ratios and feed-point impedances could result in both components being inductors or both being capacitors.

If it's desired to maintain symmetry in the feed system, X_{ser} can be divided into two components, each being inserted in series with a transmission line conductor. If X_{ser} is an inductor, the new components will each have half the value of the original X_{ser}, as shown in **Fig 18**. If X_{ser} is a capacitor, each of the new components will be twice the original value of X_{ser}.

Because of the current-forcing properties of λ/4 lines, we need to make the ratio of *voltages* at the inputs of the lines equal to the desired ratio of *currents* at the output ends of the lines; that is, at the element feed points. The job of the L network is to provide the desired voltage transformation. If the output-to-input voltage ratio of the network is, say, 2.0 at an angle of –60°, then the ratio of element currents (I_2/I_1)

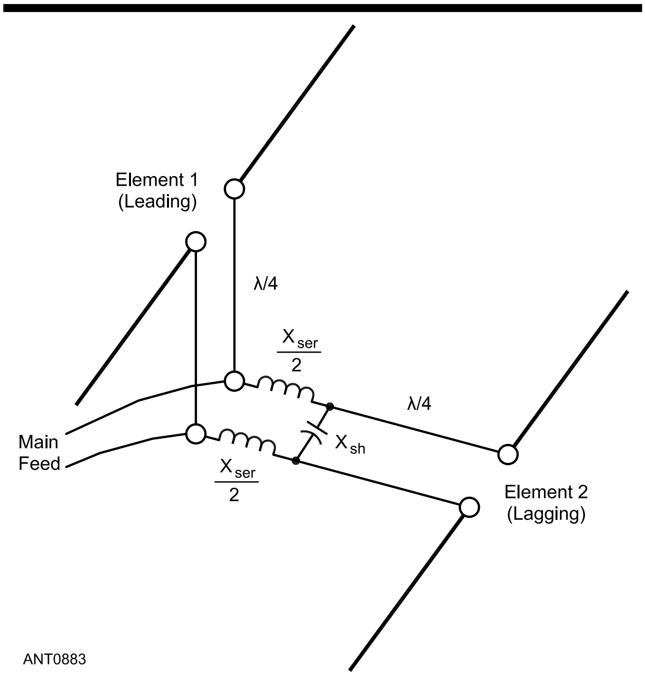


Fig 18—Symmetrical feed system similar to Fig 17, where feed network is split into two symmetrical parts.

will be 2.0 at an angle of –60°. The voltage transformation of the network is affected by the impedance of element 2, but not by the impedance of element 1. So only the impedance of element 2 must be known to design the feed system.

Equations for designing the L network are given in Appendix B, but the program *Arrayfeed1* is included on the CD to make it unnecessary to solve them. The feed-point impedance of the lagging element or group of elements must be known in order to design the network. This can best be determined by modeling the array with *EZNEC-ARRL*. The impedance can be manually calculated for some simple element and array types by using the equations in Appendix B, but those same types of element and array are simple to model.

Examples of the design of L-network feed systems for several different arrays using *EZNEC-ARRL* and *Arrayfeed1* can be found in the **Phased Array Design Examples** section. A similar application of this feed system and a spreadsheet program for calculation was developed by Robye Lahlum, W1MK, and described in *Low-Band DXing* (see the Bibliography). *Arrayfeed1* can be used for the applications of the feed system described in that book if desired.

Additional Considerations

Feeding 4-Element and Larger Arrays

Both the simplest and L-network feed systems described above can be extended to feeding larger arrays having two groups of elements in which all the elements in a group are in-phase or 180° out-of-phase with each other—basically, any group that can be fed with the current-forcing method. The elements in each group are connected to a common point with λ/4 or 3λ/4 lines to force the currents to be in the correct ratio within the group. Then the "simplest" or L-network feed system can be used to produce the correct phasing between the two groups, just as it does between two individual elements.

Two common arrays that fit this description are the Four Square and the 4-element rectangular array. But more elaborate arrays could be constructed and fed using this method, such as a pair of binomial arrays. (A single binomial array is described in the **Phased Array Design Example** section below.) The *Arrayfeed1* program incorporates additional calculations necessary for designing Four Square and 4-element rectangular arrays. The general procedure for adapting the feed methods to other larger arrays can be found in Appendix B.

What If the Elements Aren't Identical?

Getting the desired pattern requires getting the correct relative magnitude and phase of the *fields* from the elements. If the elements are identical, which we've generally assumed up to this point, then producing currents of the desired magnitude and phase will create the desired fields (neglecting mutual coupling current distribution effects, discussed elsewhere).

But what if the elements aren't identical? Fortunately, the feed systems described here can still be used for any 2-element array and some more complex arrays, provided that the system can be accurately modeled. But a slightly different approach is required than for identical elements.

The first step is to model the array with a current source at the feed point of each element. Next, the magnitudes and phases of the model source currents are varied until the desired pattern is achieved. Then the ratio of feed-point (source) currents is calculated and this value, along with the feed-point impedances reported by the model, are used for the feed system design. The feed system will produce the same ratio of currents as the model, resulting in the same pattern.

In general, this approach won't work with shunt fed towers or gamma-fed elements because of the difficulty of accurately modeling those structures. See **Shunt and Gamma Fed Towers and Elements** for more information.

Shunt and Gamma Fed Towers and Elements

In a shunt, gamma, or similarly fed tower or element, the feed-point current isn't the same as the main current flowing in the element. The ratio between the feed-point current and element current isn't a constant, but depends on a number of factors. The ratio of currents in shunt or gamma fed elements are typically different—often vastly different—from the currents at the feed points. This complicates the design of feed systems for arrays of these elements.

An even more limiting problem is that the feed-point impedances are difficult to determine. The feed-point impedances of one or more elements in a properly fed array must be known in order to design a feed system for anything but

2-element in-phase or 180° out-of-phase arrays.

The only practical way to get this information for a shunt or gamma fed array is by modeling an array having the desired element currents. But Cebik has pointed out ("Two Limitations of NEC-4"—see the Bibliography) that many common antenna analysis programs, including *EZNEC-ARRL*, have difficulty accurately modeling folded dipoles with unequal diameter wires. The same problem applies to shunt and gamma-fed elements when the element diameter is significantly different from the diameter of the shunt or gamma feed wire. Without accurate feed-point impedances, feed systems can't be designed to work without adjustment. It might be possible to get reasonably accurate results from a *MININEC*-based modeling program, but there are a number of issues which must be given great care when using one. (See Lewallen, "*MININEC*—The Other Edge of the Sword," listed in the Bibliography.)

If such a *MININEC* program is available, you would have to model the complete array including feed system, with sources at the normal feed points in the shunt or gamma wires. Next, you would have to adjust the magnitudes and phases of the sources to produce the desired pattern. The reported source impedances and currents would be the ones you would use to design the feed system. It's likely that some adjustment would be necessary, so an adjustable system such as the L-network feed system described later would be best.

Loading, Matching and Other Networks

Adding a component such as a loading inductor in series with an element or element feed point won't change the ratio of element current to feed-point current. As a result, a feed system designed to produce a particular ratio of element currents will still function properly if the elements contain series components. The extra feed-point impedance introduced by the loading component(s) must be considered when designing the feed system, however. Similarly, end or top loading won't alter the relationship between feed-point and element current, provided that the current distribution in the elements is essentially the same. (See **Feed-point vs Element Current** in a previous section.)

However, insertion of any *shunt* component, or a network containing a shunt component, *will* alter the relationship between feed-point and element current because it will divert part of the feed-line current that would otherwise flow into the antenna. As a result, a feed system designed to deliver correct currents at the feed points will produce incorrect element currents and therefore an incorrect pattern. Therefore, any components or networks other than a series loading component should be avoided at any place in the feed system on the antenna side of the point at which the feed system splits to go to the various elements.

There are a few exceptions to this rule. If the feed-point impedances of the elements when in the excited array are equal, then identical networks with or without shunt components can be put at the feed points of the elements and the proper element current ratio maintained—so long as the feed system is designed to deliver the proper feed-point current ratio with the networks in place. Equal element impedances occur in arrays having only two identical elements fed in-phase or 180º out-of-phase, or arrays of any number of elements where the elements are electrically short and/or very lossy.

Baluns in Phased Arrays

For purposes of achieving the correct array pattern, baluns aren't usually required when feeding grounded vertical elements with coaxial cable feed lines. However, a balun might be desirable if current induced onto the outside of the feed line by mutual coupling to the elements is causing RF in the shack. And with arrays of dipole or other elevated elements, baluns can be important to achieve the proper element current ratio, as explained below.

First, however, the general rules for using baluns in phased arrays will be stated. Here, "main feed line" means the feedline going from the transmitter or receiver to the common point where the system splits to feed the various elements. "Phasing-system lines" means any transmission lines between that common point and any element. The rules:

- **Rule 1:** A balun or baluns (more specifically, a current, sometimes called a choke balun) should be used as necessary to suppress unbalanced current on the main feed line. This usually isn't required when feeding grounded elements with coaxial feed line from an unbalanced rig or tuner. Unbalanced current can occur on either coax or parallel-conductor line.

"Baluns: What They Do and How They Do It", listed in the Bibliography, describes conducted-imbalance (common-mode) currents. Imbalance can also be caused by mutual coupling to the array elements. Common-mode currents have at least two undesirable effects on array performance. First, the imbalance current can flow from the main feed line to the phasing system lines, not necessarily splitting in the right proportion to maintain the correct element current ratio. This can affect the array pattern. In practice, however, this effect is likely to be small unless the common-mode current is unusually large. Even a small common-mode current, however, results in main feed-line radiation, and even a small amount of radiation can significantly degrade array pattern nulls. Any type of current balun can be used on the main feed line, at any place along the line, without any effect on the array pattern except to the extent that it reduces common mode current.

- **Rule 2:** No balun or any other component or network should be inserted in any phasing system line that will alter the line length or characteristic impedance. This means that baluns in phasing system lines must be of a type made from the phasing line itself. Options are the W2DU type balun, consisting of ferrite cores placed along the outside of the feed line; an air-core balun made by winding part of the line into an approximately self-resonant or otherwise high-impedance coil; or winding part of the line onto a ferrite core or rod to make a several-turn winding. When coaxial cable is used, the feed system characteristics are dictated by the inside of the cable. Any cores or winding of the outside prevents common-mode current on the outside, but otherwise have no effect on the phasing performance. This rule applies equally to parallel wire line, where the balun affects only common-mode current (equivalent to current on the outside of coax) while the phasing performance depends on differential mode current (equivalent to the current on the inside of coax).

Baluns are important when feeding dipole or other elevated arrays, unless a fully balanced tuner is used. This is because common-mode current represents a diversion of some of the current that should be going to the array elements. The presence of common-mode current means that the element currents are being altered from the desired ratio and therefore the pattern won't as intended. A balun should be placed wherever a path for current exists other than along a parallel-line conductor or on the inside of a coaxial line. Such a path exists, for example, where a coaxial cable connects to a dipole, as shown in Fig 1 of the balun article referenced above. Or a path can exist where a parallel-conductor transmission line connects to an unbalanced tuner or to a coaxial line, as shown in Fig 2 of that article. In both those cases, a path exists for a common-mode current to flow on the outside of the coax cable. A balun creates a high impedance to this current, thereby reducing its magnitude. But remember that all baluns must conform to the rules above.

Fig19 shows recommended balun locations for a coax-fed dipole array using an L-network feed system.

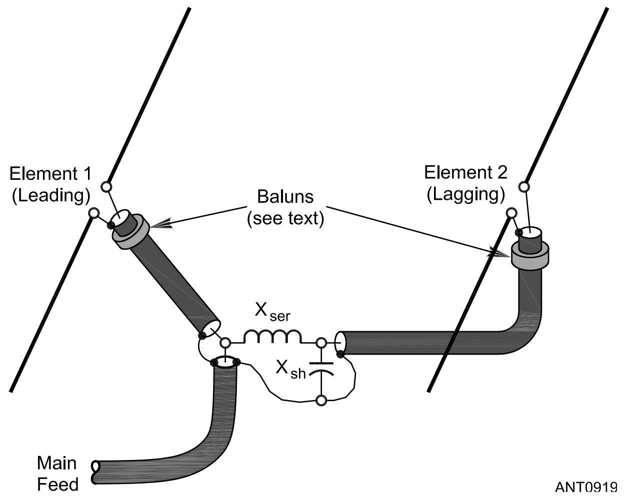


Fig 19—Adding choke baluns to a two-dipole feed system to get rid of common-mode currents radiated onto the coax shields.

Receiving Arrays

While it might not be entirely intuitive, an array designed for a particular gain and pattern for transmitting that considers mutual coupling, element currents, field reinforcement and cancellation, and so forth, will perform exactly the same when receiving. So a receiving array can be designed by approaching the problem as though the array were to be used for transmitting.

However, at HF and below, the system requirements for transmitting and receiving antennas are different, so receiving-only arrays can be designed that aren't suitable for transmitting but are perfectly adequate for receiving in that frequency range. The reason, described in more detail in Chapter 13 of this book, is that at HF and below atmospheric noise is typically much greater than a receiver's internally generated noise. Lowering a receiving antenna's gain and efficiency reduces the signal and atmospheric noise both by the same factor. Because the overall noise is for practical purposes all atmospheric noise, the signal/noise ratio isn't affected by antenna efficiency.

Of course, a point can be reached where the atmospheric noise is so reduced by inefficiency that the receiver itself becomes the dominant source of noise, but this typically doesn't happen until the antenna is extremely inefficient. When transmitting, reduced efficiency lowers the transmitted signal, but it has no effect on the receiving station's noise. So reduced efficiency of a transmitting antenna results in a reduced signal/noise ratio at the receiving end, and consequently should be avoided.

Mutual coupling effects can be minimized by increasing the loss (and therefore reducing the efficiency) of the elements, or by reducing the element sizes to a small fraction of a wavelength. Doing the second without the first isn't usually a good idea because the feed-point impedance tends to change rapidly with frequency for very small elements, making an antenna that works well over only a narrow bandwidth. But increasing loss broadens the bandwidth, even for small elements, as well as reducing mutual coupling effects. So this approach is often taken for designing a receiving-only array. With mutual coupling effects minimized because of loss, feed-system design becomes relatively simple, provided a few simple rules are followed. See **Loss Resistance, Mutual Coupling, and Antenna Gain**, above.

Phased Array Design Examples

This section, also written by Roy Lewallen, W7EL, presents examples of feed-system design for several kinds of array using the design principles given in previous sections. All but the last example array are assumed to be made of λ/4 vertical elements. The last example is for a halfwave-dipole array, which illustrates that exactly the same method can be used for arrays of any shape of elements, including dipole, square (quad) and triangular. Likewise, the methods shown here apply equally well to VHF and UHF arrays. The first example includes more detail than the remaining ones, so you should read it before the others.

General Array Design Considerations

If either the "simplest" feed system (Fig 16) or L-network feed system (Fig 17) is used, the feed-point impedance of one or more elements—*when the array elements all have the correct currents*—must be known. By far the best way to determine this is by modeling. If accurate modeling isn't practical for some reason, an estimation should be made from an approximate model, and you should expect to have to adjust the feed system after building and installing it.

Manual calculation methods for some simple configurations are given in Appendix B (on the CD), but calculation is tedious and, as stated earlier, the configurations for which this method works are the very ones which are easiest to model. *EZNEC-ARRL* (also included on the CD) is used in the following examples to determine feed-point impedance. Space doesn't permit detailed instructions here on creating the models, so they are included in complete form. They should provide a convenient starting point for any variations you might like to try. See the *EZNEC-ARRL* manual (accessed by clicking **Help/Contents** in the main *EZNEC-ARRL* window) for help in using this program.

In the following examples, vertical elements are close to λ/4 high and dipole elements close to λ/2, and their lengths have been adjusted for resonance when all other elements are absent or open circuited. There's actually no need in practice to make the elements self-resonant—it's simply used as a convenient reference point for these examples. You'll also find it interesting to see how much reactance is present at the feed points of the elements when in the arrays, knowing that it's very nearly zero when only one element is present.

In any real grounded vertical array, there is ground loss associated with each element. The amount of loss depends on the length and number of ground radials, and on the type and wetness of the ground under and around the antenna. This resistance becomes part of the feed-point resistance, so it must be included in the model used to determine feed-point impedance. The 90° Fed, 90° Spaced Array example below discusses how this is done. **Fig 20** gives resistance values for typical ground systems, based on measurements by Sevick (July 1971 and March 1973 *QST*). The values of feed system components based on Fig 20 will be reasonably close to correct, even if the ground characteristics are somewhat different than Sevick's.

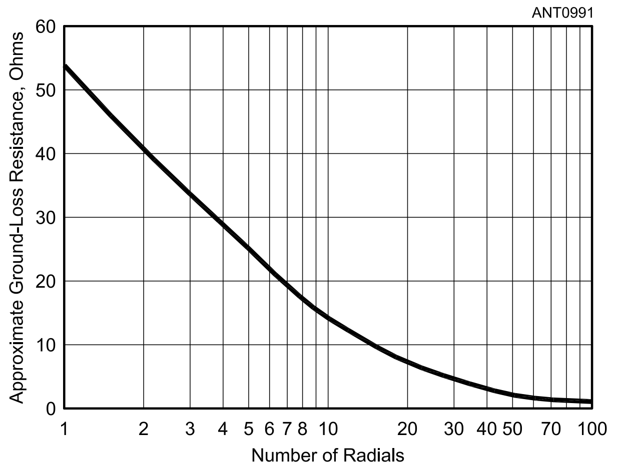


Fig 20—Approximate ground system loss resistance of a resonant λ/4 ground-mounted vertical element versus the number of radials, based on measurements by Jerry Sevick, W2FMI. Moderate length radials (0.2 to 0.4 λ) were used for the measurements. The exact resistance, especially for only a few radials, will depend on the nature of the soil under the antenna. Add 36 Ω for the approximate feed-point resistance of a thin resonant λ/4 vertical.

Feed systems for the design example arrays to follow are based on the resistance values given below.

Number of Radials	*Loss Resistance, Ω*
4	29
8	18
16	9
Infinite	0

Elevated radial systems also have some ground loss, although it can be considerably less than a system with the same number of buried radials. This loss will be automatically included in the feed-point impedance of a model which includes the elevated radials, so no further estimation is required. Be sure to use Perfect, High-Accuracy ground type when modeling an elevated radial system with *EZNEC-ARRL*. In other *NEC-2* based programs, this might be referred to as Sommerfeld type ground. More information can be found in the *EZNEC-ARRL* manual.

The matter of matching the array for the best SWR on the feed line to the station is not dealt with here, since it's a separate problem from that of the main topic, which is designing feed systems to produce a desired pattern. Some of the simpler arrays provide a match that is close to 50 or 75 Ω, so no further matching is required. However, as shown by program *Arrayfeed1*, many larger arrays present a less favorable impedance for direct connection and will require matching if a low SWR on the main feedline is required. If matching is necessary, the appropriate network should be placed in the single feed line running to the station. Attempts to improve the match by adjustment of the phasing L network, individual element lengths, matching at the element feed points or individual element feeder lengths will usually ruin the current balance of the array. Program *TLW*, included on the CD, can be used for designing an appropriate matching network. Additional information on impedance matching may be found in Chapters 25 and 26 of this book.

Choosing Arrayfeed1 Solutions

When designing a feed system for a two element array, *Arrayfeed1* program allows you to choose the characteristic impedances of the two transmission lines going to the elements, which don't have to be equal, so you have your choice of more than one solution. However, directional array switching is much more difficult if the lines have different impedances, so in general you should use the same characteristic impedances.

For larger arrays, *Arrayfeed1* requires the feed lines to all elements to have the same impedance. In choosing the transmission line impedance values, usually you can simply use convenient impedances. But in general, you should avoid solutions where component reactance (X) values are vastly different (say, more than three times or less than one third as large) as the line characteristic impedances. Such networks will become more critical to adjust, and both the impedance and pattern will change more rapidly with changes in frequency. You can usually avoid this situation by choosing feed-line impedances that are in the same ballpark as the element feed-point impedances. The last example in the Practical Array Design section illustrates this problem and its solution.

When designing a "simplest" feed system, the most broadbanded and least critical system is usually one where the difference in electrical feed-line lengths is closest to the relative element phase angle. Here, "broadbanded" means that the pattern changes less with frequency, not necessarily that the SWR changes less. However, an array that's broadbanded in the pattern sense is usually also relatively broadbanded with respect to SWR.

Arrayfeed1 reports the impedance seen at the main array feed point. While it might be tempting to choose the solution producing the lowest SWR on the main feedline, you'll end up with a less critical and more broadbanded system if you base your choice on the criteria given above, and provide separate impedance matching at the array's main feed point when necessary.

90° Fed, 90° Spaced Vertical Array

This example illustrates the design of both "simplest" and L-network feed systems for a 2-element, 90° spaced and fed vertical array. The first task when using either feed system is to determine the feed-point impedances of the elements when placed in an array having the desired element currents. The "simplest" feed system method requires knowledge of both element impedances, while the L-network system requires you to know only one. Actually, it's equally easy to determine both as it is to find just one, using *EZNEC-ARRL*. (Appendix B contains equations for those interested in manual methods or for more insight as to how the impedances come about.) The first step is to specify the antenna we want. For this example, we'll specify:

- Frequency: 7.15 MHz.
- Two identical, one inch (2.54 cm) diameter, 33 feet (10.06 meter) long elements spaced 90 electrical degrees, with element currents equal in magnitude and 90° apart in-phase.
- 8 buried radial wires, 0.3 λ long, under each element.

A model of this antenna has been created and furnished with *EZNEC-ARRL*. So the next step is to start *EZNEC-ARRL*, click the **Open** button, enter **ARRL_Cardioid_Example** in the text box (or double-click it in the file list) to open example file **ARRL_Cardioid_Example.EZ**.

This *EZNEC* example model uses a *MININEC*-type ground, which is the same as perfect ground when calculating antenna currents and impedances. A real antenna would have some additional resistive loss due to the finite conductivity of the ground system. The only way to model a buried radial ground system with an *NEC-2* based program like *EZNEC* is to create radial wires just above the ground (using the Real, High-Accuracy ground type), because *NEC-2* can't handle buried conductors.

This provides only a moderate approximation of a buried system. Another way to estimate ground-system resistance is to measure the feed-point impedance of a single element, then subtract from that the resistance reported for a model of that element over perfect (or *MININEC*-type) ground. For most uses, however, an adequate approximation can be made by simply referring to the graph of Fig 20. As stated previously, the feed system design depends on the feed-point impedances of the elements, which in turn depend on the ground system resistance. So the ground system resistance must be known, approximately anyway, before designing the feed system. At the end of this example we'll investigate the effect of changes in the ground system or errors in estimating the resistance on the pattern.

For 8 radials, Fig 20 shows the ground system resistance to be about 18 Ω. This is included in the example model as a simple resistive load at the feed point of each element. Click the **Src Dat** button to see the feed-point impedances of the two elements. In this model, Source 1 is at the base of Wire 1 (element 1), and Source 2 is at the base of Wire 2 (element 2). Notice in the **Source Data** display that the Source 1 current has been specified at 1 amp at 0°, and Source 2 is 1 amp at –90°. So the Source 2 element is the lagging element. You should see impedances of $37.53 - j19.1$ Ω for element 1 and $68.97 + j18.5$ Ω for element 2. These are the feed-point impedances resulting when the array is ideally fed, with equal magnitude and 90° phased currents. Record these values for use in *Arrayfeed1*.

Click the **FF Plot** button to generate a plot of the azimuth pattern at an elevation angle of 10°. In the 2D Plot Window, open the **File** menu and select **Save Trace As**. Enter **Cardioid_Ideal Feed** in the **File Name** box, then click **Save**. This saves the cardioid pattern plot so you can compare it later to the pattern you get with the transmission line feed system.

Now it's time to design the feed system. Refer to the appropriate subheading below for the design of each of the two kinds of feed systems. Both systems use program *Arrayfeed1* program.

"Simplest" (Transmission Line Only) Feed System

Start *Arrayfeed1*. In the **Array Type** frame, select **Two Element**. In Feed System Type, select **"Simplest."** In the **Inputs** frame, enter the following values:

Frequency MHz = 7.15; Feed-point impedances – Leading Element: R ohms = 37.53, X ohms = –19.1; Lagging Element: R ohms = 68.97, X ohms = 18.5 (these are the element R and X values from *EZNEC-ARRL*). We'll be discussing the array input impedance, so check the **Calc Zin** box near the lower left corner of the main window if it's not already checked.

We're free to choose any transmission-line characteristic impedances we want, so long as we can get cables with those impedances. And the two cables don't have to have the same characteristic impedances. Each choice will lead to a different set of solutions. But sometimes a solution isn't possible, which then requires choosing different line impedances. Let's try 50 Ω for both lines. Enter **50** in both **Z0** boxes.

Finally, enter **1** for the lagging:leading **I Mag**, and **–90** for the **Phase**. Click **Find Solutions**. The result is no solution! So enter **75** into both the line impedance boxes and click **Find Solutions** again. You should now see two sets of results in the **Solutions** frame, electrical lengths of 68.80° and 156.03° for the first solution and 131.69° and 185.00° for the second. (Notice that the difference in length between the two lines isn't 90° for either solution, although the first solution is quite close. It's normal for the feed-line length difference to be different than the phase difference, due to the unequal element feed-point impedances caused by mutual coupling.)

The solution with a line length difference closest to the element phase difference is usually preferable. Also, all else being equal, the solution with shortest lines is better providing that the lines will physically reach the elements. This is because the current magnitude and phase will change less with frequency than for a longer-length solution. However, there might be some cases where the change with frequency luckily compensates for the changing electrical distance between elements, so it's not a bad idea to model both solutions unless you plan on using the antenna over only a narrow frequency range.

In this case, the first solution looks best in all respects. The sum of the two lines in the first solution is about 225 electrical degrees. Assuming the lines have a velocity factor of 0.66, the total length of the lines will be more than 148 physical degrees. Since our two elements are spaced 90 physical degrees apart, the lines will comfortably reach. If they didn't, we could either use the second solution's lengths, use cable with a higher velocity factor or add a half wavelength to both the line lengths in the first solution.

The impedance **Zin** shown by *Arrayfeed1* is the impedance at the input to the feed system, so it's the impedance that will be seen by the main feed line. The second solution provides nearly a perfect match for a 50-Ω transmission line. But the first solution is good for nearly all applications. Also a 50-Ω line connected to the first solution's feed system would have an SWR of only 1.65:1, which wouldn't require any matching under most circumstances. Normal line loss would reduce the SWR even more at the transmitter end of the feed coax.

To find the required physical line lengths, enter the cable velocity factor and make your choice of units in the **Physical Lengths** frame. The design is now complete; all you have to do is cut two lines to the specified lengths and connect one from a common feed point to each element as shown in Fig 16, or the screen capture from *Arrayfeed1* shown in **Fig 21**.

Next, we'll design an L-network feed system for the same array.

L Network Feed System

In *Arrayfeed1*, select **L Network** in the **Feed System Type** frame. The program doesn't need to know the leading element impedance to calculate the L-network values, but it does need it to calculate the array input impedance. If you want to know the impedance, check the **Zin** box at the lower left corner of the main window, otherwise you can uncheck it and the input box for the leading element Z will disappear. The values from the "Simplest" analysis should still be present in the appropriate boxes; if not, refer to the "Simplest" feed system design above and re-enter the values. Again, we'll use 75 Ω for the line impedances, since that gave us a

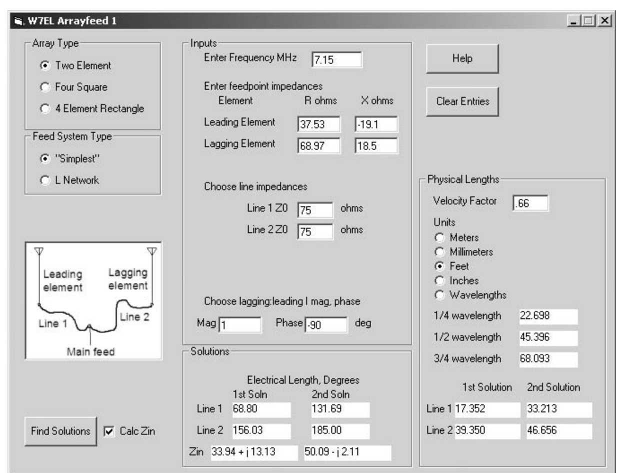


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Fig 21—Screen capture from *Arrayfeed1* program for "Simplest" 2-element phased array shown in Fig 16 and whose feed-point impedances are modeled by *EZNEC-ARRL*.

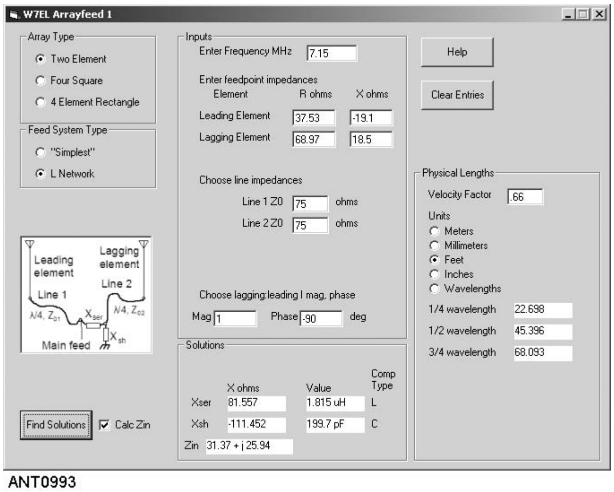

Fig 22—Screen capture from *Arrayfeed1* program for L-network feed system using "current-forcing" properties of λ/4 feed lines.

solution for the "Simplest" feed system. This feed system is more versatile, though, so we could use 50-Ω lines with this feed system if desired.

Click **Find Solution** and see the results in the **Solution** frame. See screen capture in **Fig 22**. With 75-Ω lines, the L network consists of a series inductor of 1.815 µH and a shunt capacitor of 199.7 pF, connected as shown in the diagram in the left part of the program window. To find the physical length of the λ/4 lines, enter the velocity factor and choice of units in the **Physical Lengths** frame.

The main feed-point impedance of $31.37 + j25.94$ Ω would result in about a 2.2:1 SWR on a 50-Ω feed line, which would be acceptable for many applications. It could easily be reduced to 1.6:1 by the simple addition of a series capacitor of 25.94 Ω reactance (858 pF) at the main feed point or, of course, reduced to 1:1 with a simple L network or other matching system designed with the *TLW* program.

Pattern Verification and Effect of Loss Resistance—"Simplest" System

EZNEC-ARRL doesn't have the capability to model an L-network, so *EZNEC-ARRL* verification of the pattern and the effect of various modifications can be done only for the "simplest" feed system.

EZNEC model **ARRL_Cardioid_TL_Example.EZ** has been created to model the "simplest" feed system just designed. Open it with *EZNEC-ARRL*. In the View Antenna Display, you can see the transmission lines connecting to the source midway between the antennas. In *EZNEC*, the physical locations of the ends of transmission line models don't have to be the same as the physical locations, so the view isn't a precise representation of what the actual setup would look like. (You can find more about this in **ARRL_Cardioid_TL_Example.txt**, the Antenna Notes file that accompanies example file **ARRL_Cardioid_TL_Example.EZ.**)

Click **FF Plot** to generate a 2D pattern of the antenna. In the 2D Plot Window, open the **File** menu and select **Add Trace**. Select **Cardioid—Ideal Feed** (which you saved earlier) and click **Open**. The added plot overlays perfectly, indicating that the pattern using this feed system is identical to the pattern we got with perfect current sources at each feed point.

To check the feed-point currents, click the **Currents** button. In the resulting table, you can see that **Wire 1 Segment 1** current is 0.56467 A at a phase of –56.73° and **Wire 2 Segment 1** current is 0.56467 A at –146.7°. (If you get the correct phase angles but wrong magnitudes, open the main window **Options** menu, select **Power level**, and make sure the **Absolute V, I sources** box is checked.) The ratio is 1.0000 at an angle of –89.97°, which is within normal error bounds for the desired 1 at –90°.

As a check on *Arrayfeed1*, click the **Src Dat** button to find the impedance seen by the source. This would be the impedance at the main feedline connection in the real array. *EZNEC-ARRL* reports $33.96 + j13.11$ Ω, very close to the $33.94 + j13.13$ Ω given by *Arrayfeed1* in Fig 21. Small differences of this order are normal and to be expected. This provides a further check that the *EZNEC-ARRL* model is correctly analyzing the *Arrayfeed1* feed system.

This *EZNEC-ARRL* model uses lossless transmission lines of a fixed physical length rather than a fixed electrical length (number of degrees), so they'll behave like real lines as the frequency is changed. By changing the *EZNEC* frequency and re-running the 2D plot, you can see that the front-to-back ratio degrades at 7.0 and 7.3 MHz. A slight adjustment of one or more line lengths, or a new *Arrayfeed1* solution at a slightly different frequency might produce a better compromise for some uses.

Other things you can try are to evaluate the second *Arrayfeed1* solution, or to try using different line impedances. (Keep the two line impedances equal if you anticipate doing array direction switching.) The effect of varying ground system resistance can also be evaluated by clicking the **Loads** line in the main window and changing the load resistance values. For example, if the ground system resistance were 9 Ω instead of the 18 Ω we have assumed, the front/back ratio would drop from about 32 to about 20 dB. Note that changing the *EZNEC* ground conductivity in this model has no effect on the feed-point current ratio. With a *MININEC* type ground, it's used only for pattern calculation—the ground is assumed perfect during impedance and current calculations, and the only ground system loss resistance in the model is what we've specifically put in as loads.

Not surprisingly, the forward gain is affected very little by changes in frequency or ground system loss. To find the gain relative to a single element, compare the reported dBi gain of **ARRL_Cardioid_Example** with the same model with one of the elements deleted. You'll find it's very close to 3.0 dB. The 90° fed, 90° spaced array is a special case of array where the effects of mutual coupling on the two elements are opposite and cancel, resulting in the same gain as if mutual coupling didn't exist. But mutual coupling most certainly does exist!

The second solution presented a more favorable main feed-point impedance, so it would be tempting to use that one instead of the first solution. Replacing the feed-line lengths with the second solution lengths to model the second solution shows that the front/back ratio deteriorates more at the band edges when the second solution is used. This might be tolerable if restricted frequency use is anticipated. But it does illustrate that the solution with shorter lines is generally more broadbanded and that the choice of solution shouldn't in general be based on the one giving the most favorable impedance.

A Three-Element Binomial Broadside Array

An array of three in-line elements spaced λ/2 apart and fed in-phase gives a pattern that is generally bidirectional. If the element currents are equal, the resulting pattern has a forward gain of 5.7 dB (for lossless elements) compared to a single element, but it has substantial side lobes. If the currents are tapered in a binomial coefficient 1:2:1 ratio (twice the current in the center element as in the two end elements), the gain drops slightly to just under 5.3 dB, the main lobes widen and the side lobes disappear.

The array is shown in **Fig 23**, and an *EZNEC-ARRL* model of the antenna over perfect ground to show the ideal pattern is provided as **ARRL_Binomial_Example.EZ**. To obtain a 1:2:1 current ratio in the elements, each end element is fed through a 3λ/4 line of impedance Z_0. Line lengths of 3λ/4 are chosen because λ/4 lines will not physically reach. The center element is fed from the same point through two parallel 3λ/4 lines of the same characteristic impedance. This is equivalent to feeding it through a line of impedance $Z_0/2$. The currents are thus forced to be in-phase and to have the correct ratio. **ARRL_Binomial_TL_Example.EZ** is an *EZNEC-ARRL* model that shows this feed system with lossless transmission lines. The reader is encouraged to experiment with this model to see the effect of changes in frequency, the addition of loss resistance (as resistive loads at the element feed points) and other alterations on the array pattern and gain. You should also replace the perfect ground with *MININEC* type of ground to show how radiation patterns over real ground differ from the theoretical perfect-ground pattern.

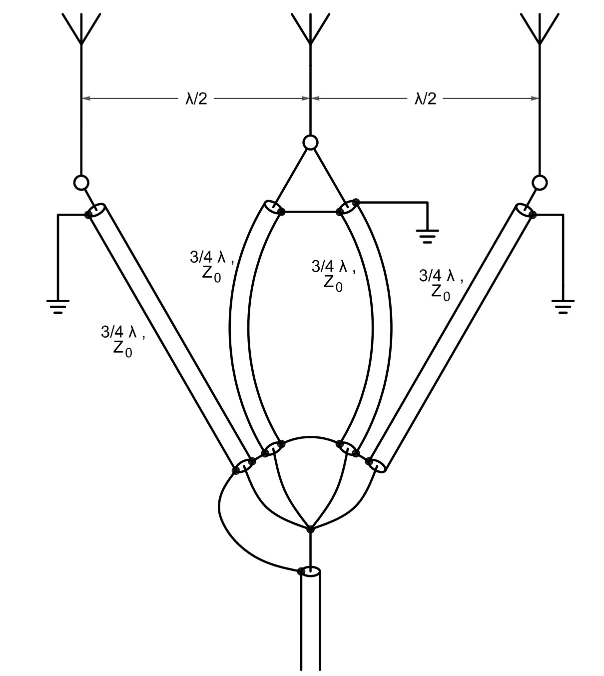

Fig 23—Feed system for the three element 1:2:1 binomial array. All feed lines are ¾ electrical wavelength long and have the same characteristic impedance.

A "Four Square" Array

Several types of feed system are used for feeding this popular array, and most share a common problem—they don't provide the correct element current ratio—although a number of them produce a workable approximation. The feed systems described here are capable of producing exactly the correct current ratio. The only significant variable is the element feed-point impedances, so the quality of the result depends on your ability to model the feed-point impedances of a correctly fed array. As in the examples above, *EZNEC-ARRL* will be used for that purpose and *Arrayfeed1* for the design of the feed system itself.

In this array (see **Fig 24**), four elements are placed in a square with λ/4 sides. (A variation of the Four Square uses wider spacing.) The rear and front elements (1 and 4) are 180° out-of-phase with each other. The side elements (2 and 3) are in phase with each other and 90° delayed from the front ele-

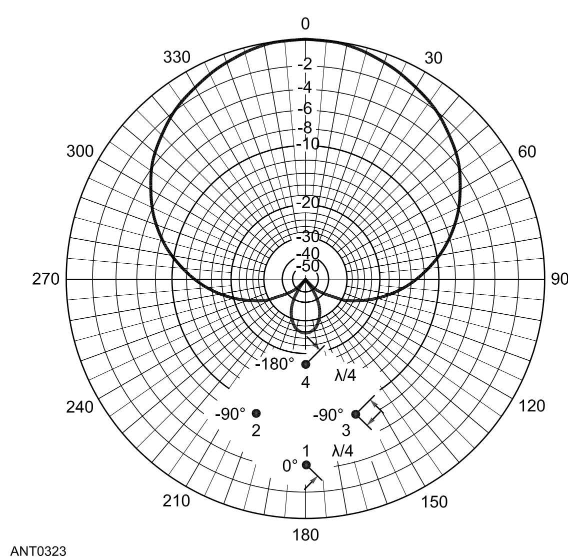

Fig 24—Pattern and layout of the four-element Four-Square array. Gain is referenced to a single similar element; add 5.5 dB to the scale values shown.

ment. The magnitudes of the currents in all four elements are equal. The front and rear elements can be forced to be 180° out-of-phase and to have equal currents by using the current-forcing method described earlier. One element is connected to a line that is either λ/4 or 3λ/4 long, the other to a line that is λ/2 longer, and the two lines to a common point.

Likewise, the two side element currents are forced to be equal by connecting them to a common point via λ/4 or 3λ/4 lines. **Fig 25** shows the basic current-forcing system.

If the pattern is to be electrically rotated, it is necessary to bring lines from all four elements to a common location. If solid-polyethylene dielectric coaxial cable, which has a dielectric constant of 0.66, is used, λ/4 lines won't reach the center of the array. So 3λ/4 lines must be used. Alternatively, you can use λ/4 lines with foam or other dielectric having a velocity factor of more than about 0.71 (plus a little extra margin). These will reach to the center. Whatever your choice, three of the lines must be the same length and the fourth must be λ/2 longer.

In this array, the side elements (2 and 3) have equal impedances, but the rear and front (1 and 4) are different from each other, and both are different from the side elements. We have to know the feed-point impedances of the front, rear and side elements in order to design the "simplest" feed system, but only the side element impedances are needed to design the L-network system. Knowledge of all feed-point impedances is necessary if the array main feedpoint impedance **Zin** is to be calculated. *EZNEC-ARRL* model **4Square_Example.EZ** shows a 40-meter Four Square array with 18 Ω of loss resistance at each element, to approximate an 8-radial per element ground system. (See the cardioid array example above for more information about modeling ground system loss.) Opening the file in *EZNEC-ARRL* and clicking the **Src Dat** button gives the following impedances:

Source 1: 16.4 – *j*15.85 Ω
Sources 2 and 3: 57.47 – *j*19.44 Ω
Source 4: 77.81 + *j*54.8 Ω

It's interesting to note that the resistive part of source 1 is less than the 18 Ω of loss resistance we intentionally added to simulate ground system loss. That means that the element 1 feed-point resistance would be negative if the ground resistance were less than about an ohm and a half. This isn't uncommon in phased arrays and simply means that the element is feeding power *into* the feed system. This power is coming via mutual coupling from the other elements.

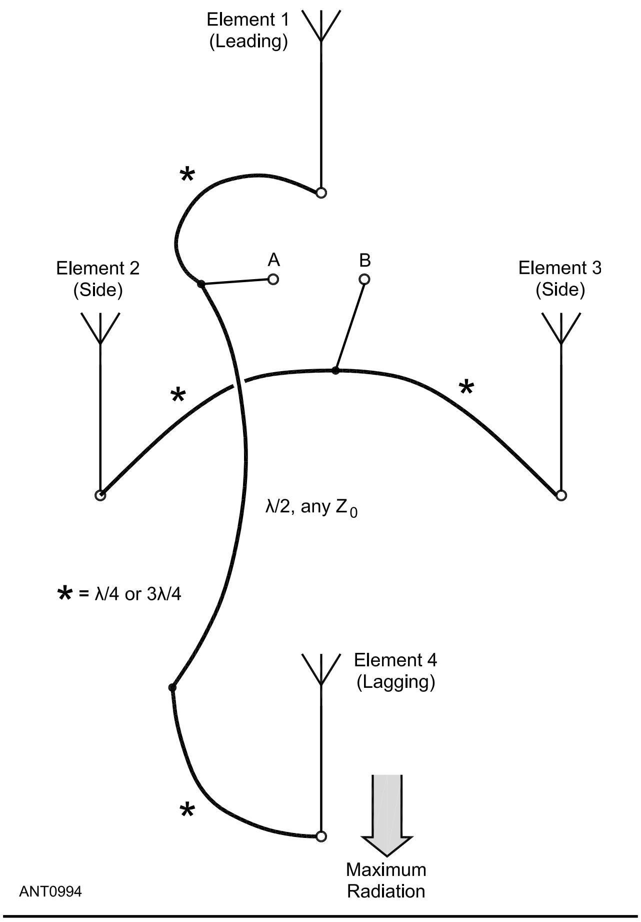

Fig 25—"Simplest" feed system for the Four-Square array in Fig 24. Grounds and cable shields have been omitted for clarity.

"Simplest" (Transmission Line Only) Feed System

To design a "simplest" feed system, start *Arrayfeed1*. In the **Array Type** frame, select **4 Square**, and select **"Simplest"** in the **Feed System Type** frame. In the **Inputs** frame, enter the frequency and the impedances from *EZNEC-ARRL*:

Frequency = 7.15 MHz
Leading Element: R = 16.4, X = –15.85
Side elements: R = 57.47, X = –19.44
Lagging Element: R = 77.81, X = 54.8

We'll try using 50 Ω for all lines, so enter **50** into the next three boxes.

Enter **1** for the lagging:leading I magnitude and **–90** for the phase.

Click **Find Solutions**.

The result is shown in the **Solutions** frame, shown in **Fig 26**. As always when any solution exists, there are two to choose from. The one with the shortest lines is generally preferable, so we'll choose it. For this example, we'll use λ/4 lines with velocity factor of 0.82. So enter **0.82** in the **Velocity Factor** box in the **Physical Lengths** frame, and read the physical lengths from the bottom of that frame. The λ/4 lines (marked in the *Arrayfeed1* diagram with an asterisk) are 28.2 feet, line 1 is 7.483 feet and line 2 is 51.668 feet. The "simplest" feed system is shown in Fig 26, and the complete feed system consists of this connected to the array of Fig 25.

EZNEC-ARRL model **ARRL_4Square_TL_Example.EZ** is simulates the array fed with this system. Comparison of the pattern plot to one from ideal-current model **ARRL_4Square_Example.EZ** and examination of

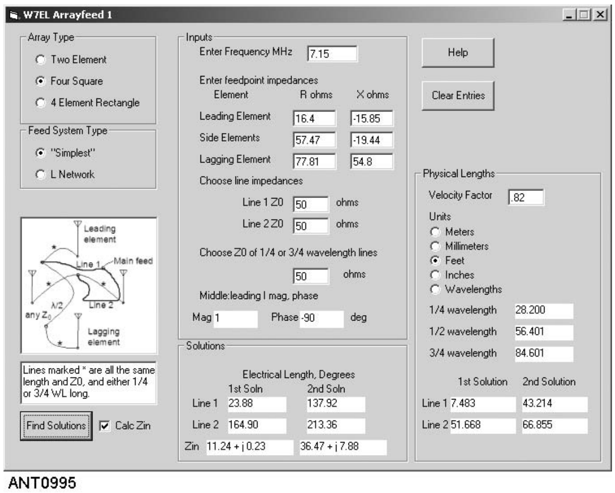

Fig 26—Screen capture from *Arrayfeed1* for "Simplest" feed system for Four Square feed system shown in Fig 25.

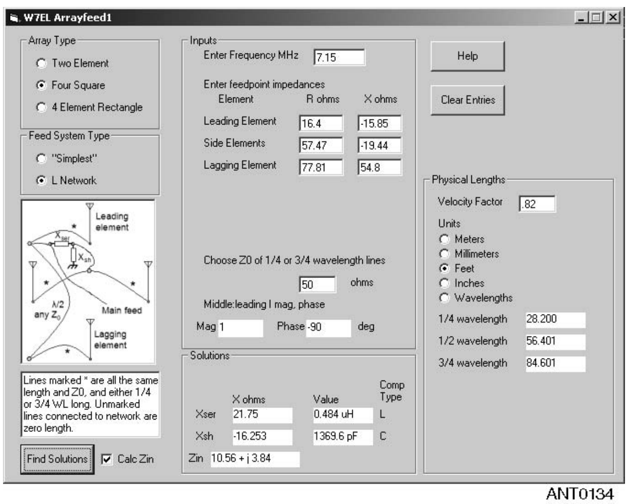

Fig 27—L-network setup for Four-Square array in Fig 25, fed with λ/4 (or 3λ/4) current-forcing feed system.

the element currents verify that the feed system is producing the desired pattern and element currents. You can use **ARRL_4Square_Example.EZ** to investigate the effect of frequency change, ground loss and other changes on the array gain and pattern.

L-Network Feed System

To design the L-network feed system, simply change the **Feed System Type** to **L Network** and click **Find Solutions**. The results you should see are a 0.484 µH inductor for the series component X_{ser}, and a 1369.6 pF capacitor for the shunt component X_{sh}. The L-network feed system is shown in **Fig 27**, and the complete feed system consists of this L network connected to the array of Fig 25.

EZNEC-ARRL doesn't have the ability to directly model

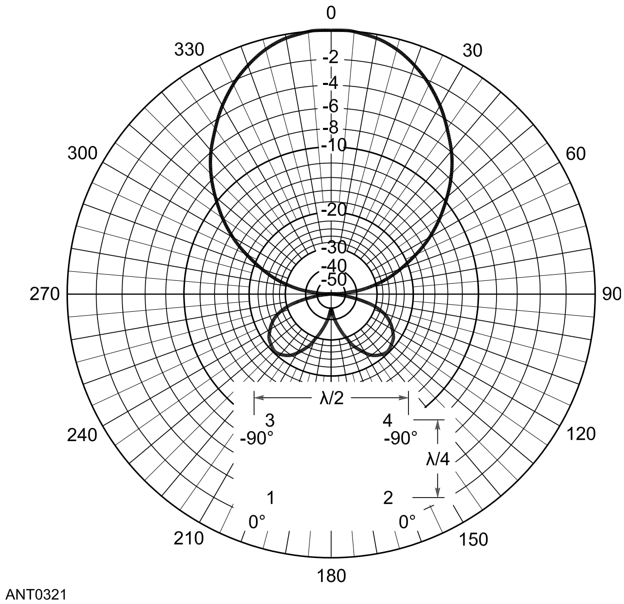

Fig 28—Pattern and layout of the four-element rectangular array. Gain is referenced to a single similar element; add 6.8 dB to the scale values shown.

the L network, so it's unable to model the complete system. However, the system has been modeled using the network capability of *EZNEC v.5* and found to work as designed. Arrays have also been built using this feed system and the element currents measured, with exactly the expected results.

This array is more sensitive to adjustment than the 2-element 90° fed, 90° spaced array. Adjustment procedures and a method of remotely switching the direction of this array are described in the **Practical Aspects of Phased Array Design** section that follows.

A 4-Element Rectangular Array

The 4-element rectangule array shown with its pattern in **Fig 28** has appeared numerous times in amateur publications. However, many of the accompanying feed systems fail to deliver currents in the proper amounts and phases to the various elements. The array can be correctly fed using the principles discussed in this chapter and the design methods that follow.

Elements 1 and 2 can be forced to be in-phase and to have equal currents by feeding them through 3λ/4 lines. (As in the binomial and Four Square array examples, 3λ/4 lines are chosen because λ/4 lines won't physically reach.) The currents in elements 3 and 4 can similarly be forced to be equal and in-phase. **Fig 29** shows the "current-forcing" feed system. Elements 3 and 4 are made to have currents of equal magnitude but of 90° phase difference from elements 1 and 2 by use of either a "simplest" all-transmission line feed system or an L-network feed system. Both will be designed in this example.

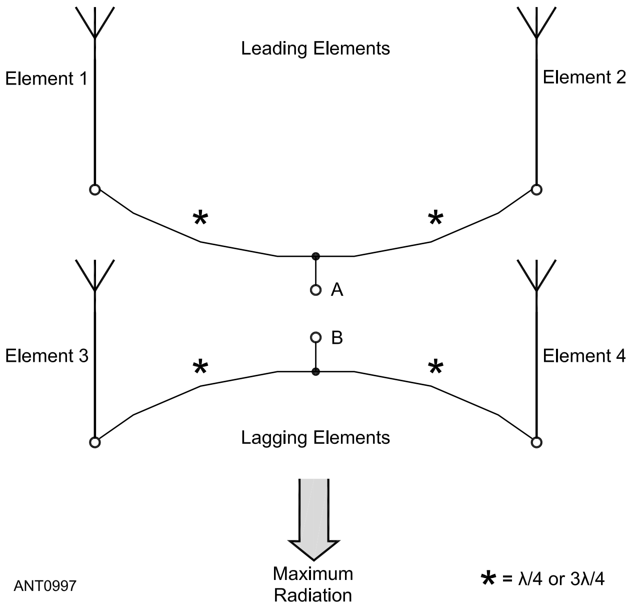

Fig 29—"Simplest" feed system for four-element rectangular array, using four equal-length λ/4 (or 3λ/4) cables.

For this array, we have to know the feed-point impedances of two elements (one of each pair) in order to design either type of feed system. *EZNEC-ARRL* model **Rectangular_Example.EZ** shows a 20-meter rectangular array with 18 Ω of loss resistance at each element, again to approximate an 8-radial per element ground system. (See the cardioid array example above for more information about modeling ground system loss.) Open the file in *EZNEC-ARRL* and click the **Src Dat** button to find the following feed-point impedances:

Sources 1 and 2: 21.44 – *j*21.29 Ω
Sources 3 and 4: 70.81 – *j*5.232 Ω

"Simplest" (Transmission- Line Only) Feed System

To design a "simplest" feed system, start program *Arrayfeed1*. In the **Array Type** frame, select **4 Element Rectangle**, and select **"Simplest"** in the **Feed System Type** frame. In the **Inputs** frame, enter the frequency and the impedances from *EZNEC-ARRL*:

Frequency = 14.15 MHz
Leading Elements R = 21.44, X = –21.29
Lagging Elements R = 70.81, X = –5.232

We'll use 50 Ω for all lines, so enter **50** into the next three boxes.

Enter **1** for the lagging:leading I magnitude and **–90** for the phase.

Click Find Solutions.

The result is "No Solution"—This combination of line impedances can't be used. Several other combinations also produce this result, but making lines 1 and 2 each 75 Ω and the 3λ/4 lines 50 Ω does produce a solution. Enter **75** into the **Line 1 Z0** and **Line 2 Z0** boxes, and leave **50** in the **Choose Z0 of ¼ or ¾ wavelength lines** box, then click the Find Solutions button. There won't be any problem making lines 1 and 2 reach, so we'll choose the first solution because the lines are shorter. The physical lengths of all the lines are shown in the **Physical Lengths** frame when the velocity factor is entered in the appropriate box. Assuming that we use coax with a velocity factor of 0.66 (and the example frequency of 14.15 MHz), the lengths are:

Line 1: 4.982 feet
Line 2: 20.153 feet
3λ/4 lines (marked with an asterisk in the *Arrayfeed1* diagram): 34.408 feet

The lines are connected following the diagram in the upper left part of the *Arrayfeed1* window. This completes the "simplest" feed system design. *EZNEC-ARRL* model **Rectangular_TL_Example.EZ** simulates an array fed with this system.

Comparison of the pattern plot to one from ideal-current **Rectangular_Example.EZ**, and examination of the element currents verify that the feed system is producing the desired pattern and element currents.

L-Network Feed System

To design the L-network feed system using *Arrayfeed1*, change the **Feed System Type** to **L Network** and click **Find Solutions**. The resulting L-network values are a 0.484 µH inductor for the series component X_{ser} and a 1369.6 pF capacitor for the shunt component X_{sh}.

120° Fed, 60° Spaced Dipole Array

This example shows the design of "simplest" and L network feed systems for a 2-element 20- meter dipole array, rather than a vertical array. No special accommodation is required for the array made from dipoles rather than vertical elements—the same methods can be used regardless of element shape. This example also shows that both the "simplest" and L-network feed systems can readily be applied to elements that use phase angles other than 90°.

Any 2-element array made with identical elements spaced λ/2 or closer and having equal magnitude currents with a relative phase angle of 180° minus the spacing will produce a unidirectional pattern with a good null to the rear. In practice, very close spacings lead to very low feed-point resistances, with consequent losses and very narrowband characteristics. But this 60° spaced array is well within the range of practical realization. File **ARRL_Dipole_Array_Example.EZ** is a model created for this array, with ideal element currents. Open this file in *EZNEC-ARRL* and click **FF Plot** to show the pattern at an elevation angle of 10°. You can save this pattern for later comparison to the pattern with a "simplest" feed system by opening the **File** menu in the **2D Plot** window, selecting **Save Trace As**, entering a name for the trace file and clicking **Save**.

Following the same procedure as in the previous examples, we begin the array design by finding the element feed-point impedances in the ideally fed array using *EZNEC-ARRL* numbers. Having already opened **ARRL_Dipole_Array_Example.EZ**, all that's needed is to click **Src Dat.** The results are:

Leading element (source 1) : 36.16 – *j*46.05 Ω
Lagging element (source 2) : 49.56 + *j*51.47 Ω

"Simplest" (Transmission Line Only) Feed System

Select **Two Element** for the **Array Type** in *Arrayfeed1* and **"Simplest"** for the **Feed System Type**. Enter the frequency of 14.15 MHz and enter the element feed-point impedances from *EZNEC-ARRL* into the appropriate boxes in the **Inputs** frame. For line impedances, the section describing the "simplest" feed system recommends against choosing one which is very different from the element feed-point impedances, but for fun let's try 300 Ω for the two lines and see what happens. Enter **300** in the **Line 1 Z0** and **Line 2 Z0** boxes. Finally, enter the **lagging:leading I mag, phase** of **1** for **Mag** and **–120** for **Phase**.

Click **Find Solutions**. For this example we'll assume that TV type twinlead with a velocity factor of 0.8 is being used. So enter **0.8** for the **Velocity Factor** and read the physical line lengths in the **Physical Lengths** frame. A model of the array using the first solution has been created as **ARRL_Dipole_Array_TL_Example.EZ**. Open this file in *EZNEC-ARRL* and click **FF Tab**. You should see that the plot is virtually identical to the one saved earlier from the ideal-current model. Note the gain and front/back ratio or 8.79 dBi and 31.01 dB respectively reported in the data box below the 2D plot.

Don't subtract 2.15 dB to find the gain relative to a single element! This isn't a free-space model, and the gain of a single dipole over ground is much greater than 2.15 dBi. Instead, delete one of the elements in **ARRL_Dipole_Array_Example.EZ** to find the gain of a single element and subtract that value from the array gain. You can use the undo feature or re-open the file to restore the array.

Now, go back to the model with the "simplest" feed system in *EZNEC-ARRL* and change the Frequency to **14.0** MHz. Click **FF Tab** again. The gain has decreased a little, to 8.54 dBi and the front/back ratio has also decreased, to 21.8 dB. At 14.3 MHz, the gain is slightly higher, 9.04 dBi, but the front/back is again worse, down to 18.64 dB. But this isn't bad overall.

Let's take a look at the second solution. Click the **Trans Lines** line in the main *EZNEC-ARRL* window to open the Transmission Lines Window. Change the length of the first line to **26.856** feet, the second to **28.356** feet, and press the **Enter** key to complete the change. Change the **Frequency** back to **14.15** MHz and click **FF Tab**. You should see exactly the same pattern as for both the first solution and for the ideal current model. But now change the **Frequency** to **14.0** MHz, click FF Tab, and look at the pattern.

What happened? The gain has dropped to 5.95 dBi and the front/back to only 3.1 dB. The array is now nearly bi-directional! It's almost as bad at 14.3 MHz. So we've created a terribly touchy system. The chance of its working correctly even at the design frequency is slim, because there are inevitably some differences between the model and real antenna.

We did have a clue this might happen. As stated in the section describing the "simplest" feed system, the best choices for line Z_0 and for the resulting solution give a difference in electrical line lengths about equal to the desired phase delay of the current. The difference in electrical line lengths for the first solution was about 152°—not as close to the 120° current phase difference as we'd like, but much better than the mere 9.7° difference of the lines for the second solution. While the 300-Ω line Z_0 is quite different from the element feed-point impedances, the first solution result is quite good. If desired, you can try other line impedance values into *Arrayfeed1* and evaluate the results with *EZNEC-ARRL*.

Please see the information about baluns in the **Baluns in Phased Arrays** section. Baluns are placed the same as in Fig 19, which shows the L-network feed system.

L-Network Feed System

To design an L network feed system, change the *Arrayfeed1* **Feed System Type** to **L Network** and click **Find Solutions**. The results aren't good ones to use. The component reactance magnitudes of about 1573 and 2619 Ω are more than five times the 300-Ω Z_0 of the feed lines. As explained in the section describing the L-network feed system, it's undesirable to have such a large ratio of component reactance to line Z_0. Among other problems, the inductor and capacitor values are quite extreme and capacitor stray inductance and inductor capacitance would have a significant impact on performance.

The problem occurs because the feed-line impedance we chose is much larger than the element feed-point impedances, so the λ/4 lines transform the feed-point impedances to much higher values at the L network and main feed point. This feed system would be extremely critical, narrowbanded and difficult to adjust. We can do better by choosing feed-line impedances that aren't too drastically different than the element feed-point impedances. In this case, 50 or 75 Ω would be a much better choice than 300. Let's try 75.

In *Arrayfeed1*, change the **Line 1 Z0** and **Line 2 Z0** impedances from **300** to **75** and click **Find Solutions**. L-network component reactance magnitudes are now about 98 and 164 Ω, much better than before. This will be a relatively uncritical and broadbanded feed system.

Again, be sure to read the information about baluns in the **Baluns in Phased Arrays** section. Fig 19 shows the completed feed system including baluns.

PRACTICAL ASPECTS OF PHASED ARRAY DESIGN

With almost any type of antenna system, there is much that can be learned from experimenting with, testing and using various array configurations. In this section, Roy Lewallen, W7EL, shares the benefit of years of his experience from actually building, adjusting and using phased arrays. There is much more work to be done in most of the areas covered here, and Roy encourages the reader to build on this work.

Adjusting Phased Array Feed Systems

If a phased array is constructed only to achieve forward gain, adjusting it is seldom worthwhile. This is because the forward gain of most arrays is quite insensitive to either the magnitude or phase of the relative currents flowing in the elements. If, however, good rejection of unwanted signals is desired, adjustment may be required. And achieving very deep nulls will almost surely require some adjustment.

The in-phase and 180° out-of-phase current-forcing method supplies very well-balanced and well-phased currents to elements without adjustment. If the pattern of an array fed using this method is unsatisfactory, it's generally the result of environmental differences—where the elements, even though furnished with correct currents, aren't generating the correct fields. Such an array can be optimized in a single direction, but a more general approach than the current-forcing method must be taken. Some possibilities are described by Paul Lee and Forrest Gehrke (see Bibliography).

Unlike the current-forcing method, the "simplest" and L-network feed systems described earlier in this chapter are dependent on the self and mutual impedance of one or more elements. The required transmission-line lengths or L-network component values can be computed to a high level of precision, but the results are only as good as the knowledge of the relevant feed-point impedances.

While the simplest feed system doesn't readily lend itself to adjustment, the components of an L network can easily be made adjustable or can be experimentally changed in increments. A practical approach is to model the array as accurately as possible, design and build the feed system based on the model results and then adjust the network for the best performance.

Simple arrays such as the two-element 90° fed and spaced array can be adjusted as follows. Place a low-power signal source at a distance from the array (preferably several wavelengths), in the direction a null should be. While listening to the signal on a receiver connected to the array, alternately adjust the two L-network components for the best rejection of the signal.

This has proved to be a very good way to adjust 2-element arrays. However, variable results were obtained when a Four-Square array was adjusted using this technique. The probable reason is that more than one combination of current balance and phasing can produce a null in a given direction but each produces a different overall pattern. So a different method must be used for adjusting more complex arrays. This involves actually measuring the element currents in some way, and adjusting the network until the currents are correct. After adjusting the currents, small adjustments can be made to deepen the null(s) if desired.

Measuring Element Currents

You can measure the element currents two ways. One way is to measure them directly at the element feed points, as shown in **Fig 30**. A dual-channel oscilloscope is required to monitor the currents. This method is the most accurate and it provides a direct indication of the actual relative magnitudes and phases of the element currents. The current probe is shown in **Fig 31**.

Instead of measuring the element currents directly, you could measure them indirectly by measuring the voltages on the feed lines an electrical $\lambda/4$ or $3\lambda/4$ distance from the array. The voltages at these points are directly proportional to the element currents. This introduces additional variables that can reduce the accuracy of the result, but the method generally produces adequate performance. The 2-element arrays fed with the L-network system and all the four element arrays presented earlier have $\lambda/4$ or $3\lambda/4$ lines from all elements to a common location, making this second measurement method convenient. The voltages can be observed with a dual-channel oscilloscope, or, to adjust for equal-magnitude currents and 90° phasing, you can use the test circuit shown in **Fig 32**.

The test circuit is connected to the feed lines of two elements that are to be adjusted for 90° phasing (such as elements 1 and 2, or 2 and 4 of the Four-Square array of Figs 24 and 25). Adjust the L-network components alternately until both meters read zero. Proper operation of the test circuit can be verified by disconnecting one of the inputs. The *phase* output

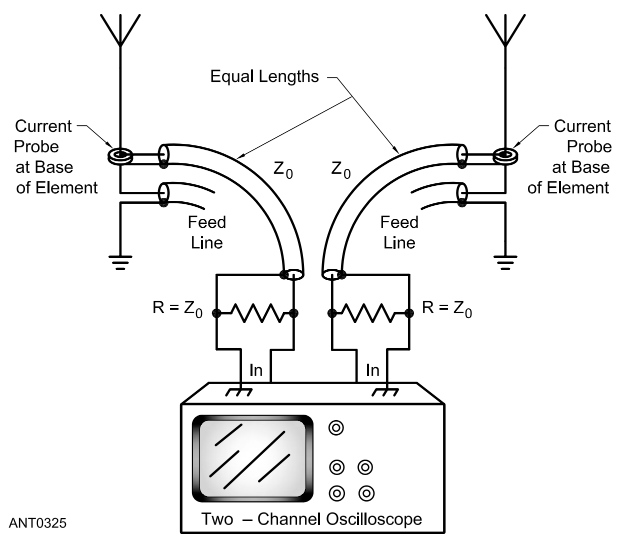

Fig 30—One method of measuring element currents in a phased array. Details of the current probe are given in Fig 31. Caution: Do not run high power to the antenna system for this measurement, or damage to the test equipment may result.

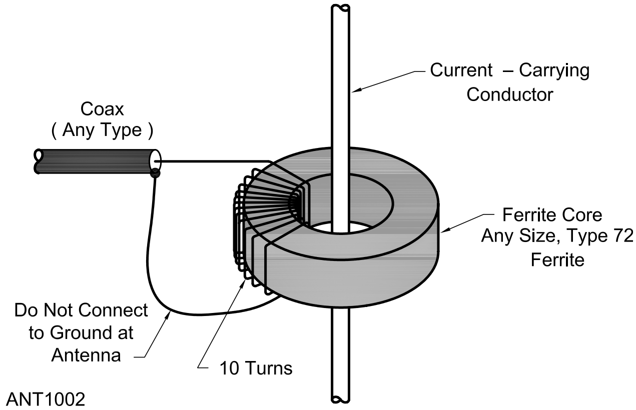

Fig 31—The current probe for use in the test setup of Fig 30. The ferrite core is of type 72 material, and may be any size. The coax line must be terminated at the opposite end with a resistor equal to its characteristic impedance. You should build this probe in a plastic or metal box to provide mechanical ruggedness.

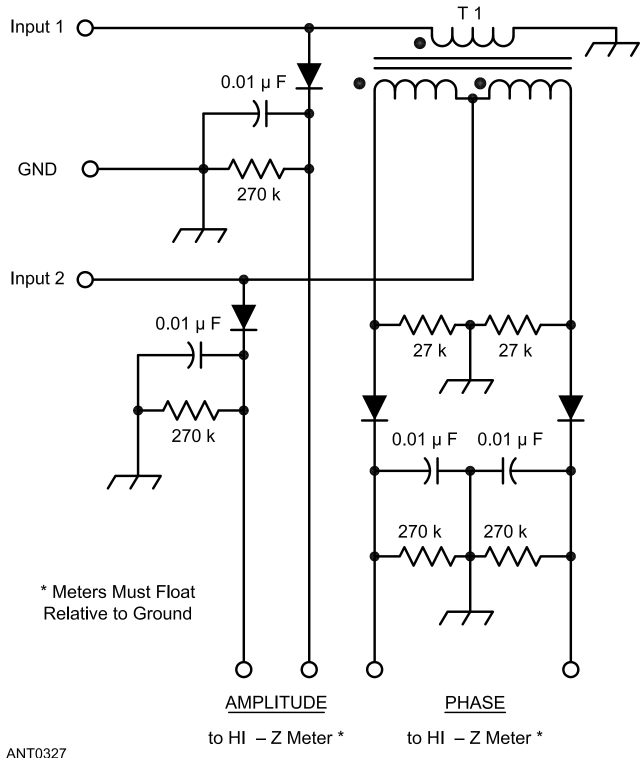

Fig 32—Quadrature test circuit. All diodes are germanium, such as 1N34A, 1N270, or equiv. Hot carrier or silicon diodes can be used at higher power levels. All resistors are ¼ or ½ W, 5% tolerance. Capacitors are ceramic. Alligator clips are convenient for making the input and ground connections to the array.

T1—7 trifilar turns on an Amidon FT-37-43, -75, -77, or equivalent ferrite toroid core.

should remain close to zero. If not, there is an undesirable imbalance in the circuit, which must be corrected. Another means of verification is to first adjust the L network so the tester indicates correct phasing (zero volts at the *phase* output). Then reverse the tester input connections to the elements. The *phase* output should remain close to zero.

Directional Switching of Arrays

One ideal directional-switching method would take the entire feed system, including the lines to the elements and physically rotate it. The smallest possible increment of rotation would depend on the symmetry of the array—the feed system would need to rotate until the array again *looks* the same to it. For example, any 2-element array can be rotated 180° (although that wouldn't accomplish anything if the array is bidirectional to begin with). The 4-element rectangular array of Figs 28 and 29 can also be reversed, and the Four-Square array of Figs 24 and 25 can be switched in 90° increments.

Smaller switching increments can be accomplished only by reconfiguring the feed system, including any network if used, effectively creating a different kind of array. Switching in smaller increments than dictated by symmetry will create a different pattern in some directions than in others, and must be thoughtfully done to maintain equal and properly phased element currents. The methods illustrated here will deal only with switching in increments related to the array symmetry, except for one: a 2-element broadside/end-fire array.

In all arrays, the success of directional switching depends on the elements and ground systems being identical so that equal element currents result in equal fields. It's even more important in arrays fed with any method other than current forcing, because the effectiveness of those methods depends on the element feed-point impedances. Few of us can afford the luxury of having an array many wavelengths away from all other conductors, so an array will nearly always perform somewhat differently in each direction. The array should be adjusted when steered in the direction requiring the most signal rejection in the nulls. Forward gain will, for all practical purposes, be equal in all the switched directions, since gain is much more tolerant of error than are nulls.

Basic Switching Methods

Following is a discussion of basic switching methods, how to power relays through the main feed line and other practical considerations. In diagrams, grounds are frequently omitted to aid clarity, but connections of the ground conductors must be carefully made. In fact, it is recommended that the ground conductors be switched just as the center conductors are, as explained in more detail in **Improving Array Switching Systems** below. In all cases, interconnecting lines must be very short.

A pair of elements spaced λ/2 apart can readily be switched between broadside and end-fire bidirectional pat-

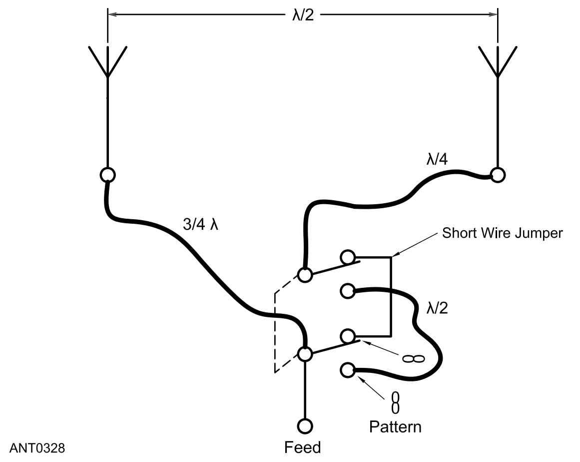

Fig 33—Two-element broadside/end-fire switching. All lines must have the same characteristic impedance. Grounds and cable shields have been omitted for clarity.

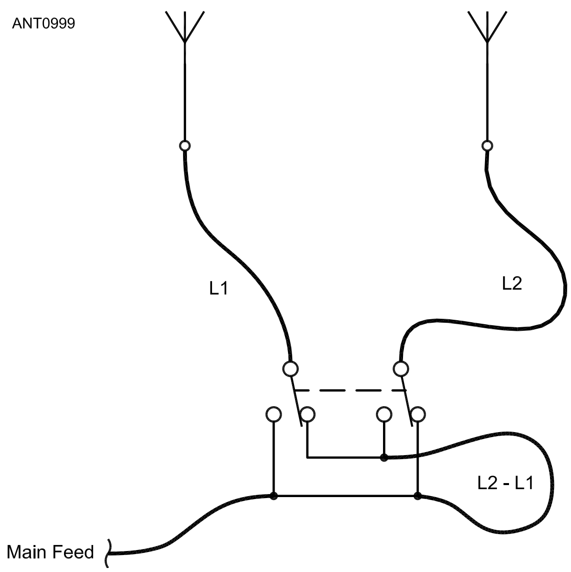

Fig 34—Directional switching for 90°, 90° spaced 2-element array fed with a "simplest" feed system.

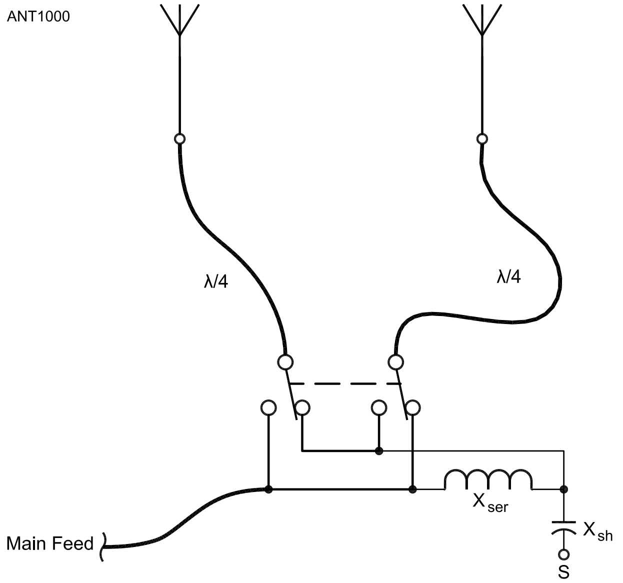

Fig 35—Directional switching for 90°, 90° spaced 2-element array fed with an L-network, current-forcing feed system.

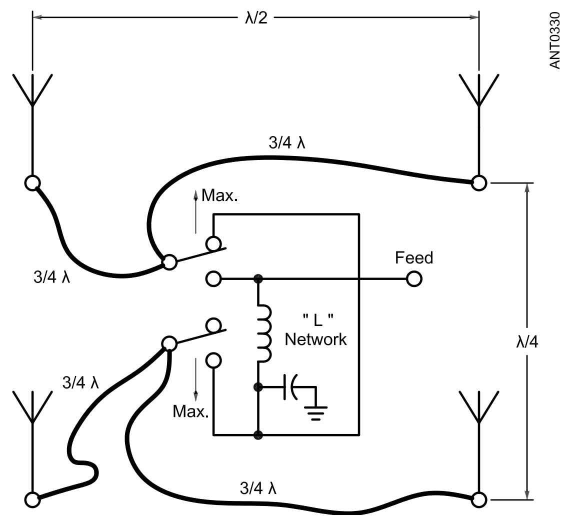

Fig 36—Directional switching of a four-element rectangular array. All interconnections must be very short. As usual, grounds and cable shields have been omitted for clarity.

terns, using the current-forcing properties of λ/4 lines. The method is shown in **Fig 33**. The switching device can be a relay powered via a separate cable or by dc sent along the main feed line.

Fig 34 shows directional switching of a 90° fed, 90° spaced array fed with a "simplest" feed system, where L1 and L2 are the required lengths of the two feed lines. **Fig 35** shows how to switch the same array when fed with an L-network, current-forcing system.

The rectangular array of Fig 28 can be switched in a similar manner, as shown in **Fig 36**. To switch a "simplest" fed rectangular array, use the switching circuit of Fig 34, but connect the two equal length lines to points A and B of Fig 29 in place of the two elements shown in Fig 34.

Switching the direction of an array in increments of 90°, when permitted by symmetry, requires at least two relays. A method of 90° switching of the Four-Square array with L-network feed is shown in **Fig 37**.

Powering Relays Through Feed Lines

All of the above switching methods can be implemented without additional wires to the switch box. A single-relay system is shown in **Fig 38A**, and a two-relay system in

Fig 38B. Small 12 or 24-V dc power relays can be used in either system at power levels up to at least a few hundred watts. Do not attempt to change directions while transmitting, however. Blocking capacitors C1 and C2 should be good quality ceramic or transmitting mica units of 0.01 to 0.1 µF. No problems have been encountered using 0.1 µF, 300-V monolithic ceramic units at RF output levels up to 300 W. C2 may be omitted if the antenna system is an open circuit at dc. C3 and C4 should be ceramic, 0.001 µF or larger.

In Fig 38B, capacitors C5 through C8 should be selected with the ratings of their counterparts in Fig 38A, as given above. Electrolytic capacitors across the relay coils, C9 and C10 in Fig 38B, should be large enough to prevent the relays from buzzing, but not so large as to make relay operation too slow. Final values for most relays will be in the range from 10 to 100 µF. They should have a voltage rating of at least double the relay coil voltage. Some relays do not require this capacitor. All diodes are 1N4001 or similar. A rotary switch may be used in place of the two toggle switches in the two-relay system to switch the relays in the desired sequence.

Improving Array-Switching Systems

The extra circuitry involved in switching arrays can degrade array performance by altering the relative currents fed to each element. One common cause is current sharing in common ground conductors, even when connections are kept very short. The author has seen a 30° phase shift in voltage along a 4-inch piece of #12 wire in a 40-meter array feed system.

When the two conductors of a feedline are physically separated from each other, the impedance increases. This is especially true when the main lines are coaxial cables. If currents from two elements share the ground conductor of a split line, a relatively large voltage drop results. Voltage changes λ/4 from the elements translate to current changes at the elements. Although keeping all leads extremely short is sometimes adequate, the best way to reduce current sharing problems is to keep the two conductors of each transmission line as close together as possible, and switch both conductors of each line rather than just a single or "hot" conductor.

An example of a carefully designed switching system is shown in **Fig 39**. It avoids the problem of shared ground conductor currents, as well as another common problem, namely that effective line lengths are often different along different switching paths. Notice how the path from the main feed point travels through a single line to each element with no common ground connections to other lines except at the

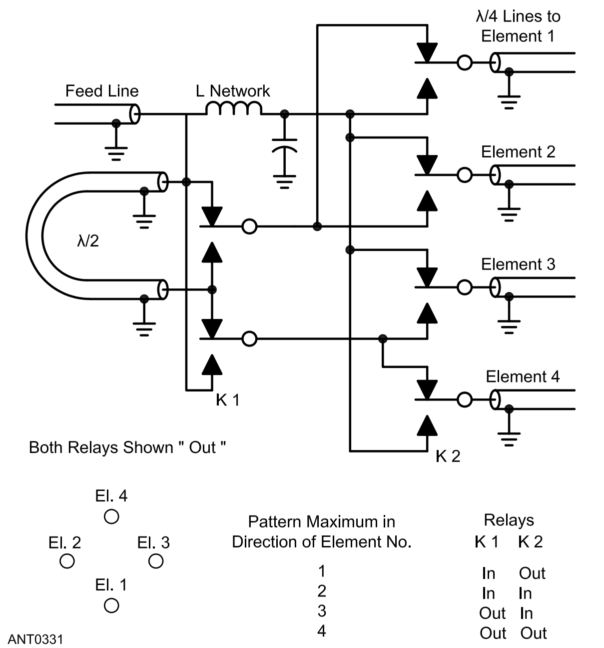

Fig 37—Directional switching of the four-square array. All interconnections must be very short.

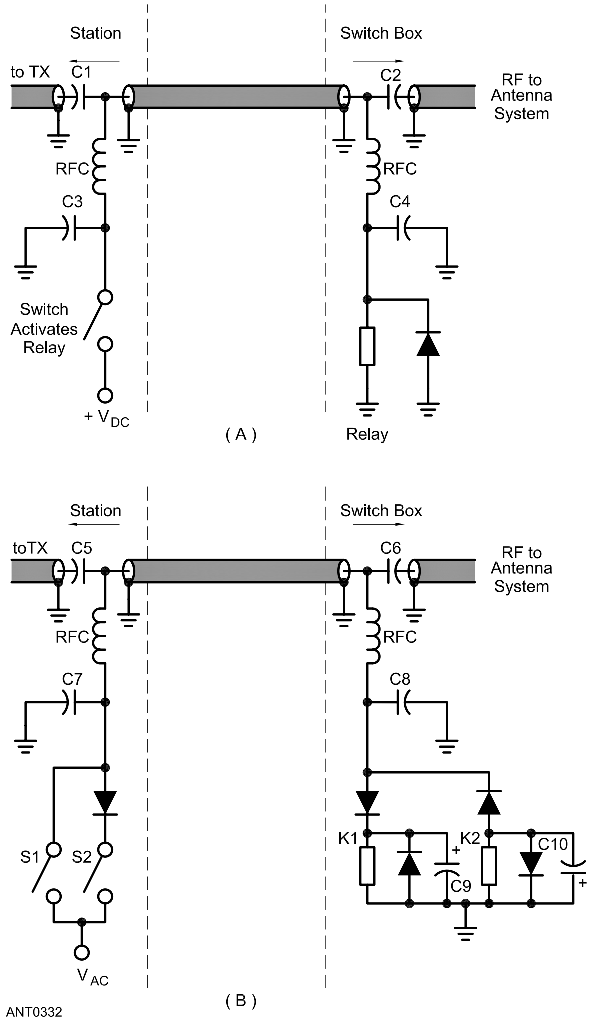

Fig 38—Remote switching of relays. See text for component information. A one-relay system is shown at A, and a two-relay system at B. In B, S1 activates K1, and S2 activates K2.

main feed point. Notice also that the distance doesn't change as the direction is switched. The λ/4 lines going to the two elements must be shortened by the length ℓ of the lines on the feed side of the relays, so that the total line length from the main feed point to each element is λ/4 (or 3λ/4).

You can see that in either relay position, there's an open ended stub of length ℓ connected at the main feed point and another at the output end of the L network. These will add capacitance at those points. Extra C at the main feed point will alter the overall impedance seen by a transmitter, but won't otherwise have any effect on the array or its performance. The one at the output of the L network will, however, change the transformation and phase shift properties of the network. But it's easy to compensate—the value of the shunt capacitor element is simply reduced by the amount of the C added by the stub. The amount of C for any kind of transmission line can be calculated from:

$$C(\text{pf/ft}) = \frac{1017}{Z_0 VF}$$

or

$$C(\text{pf/m}) = \frac{3336}{Z_0 VF}$$

where Z_0 = the characteristic impedance of the line and VF = the velocity factor. This works out to 31 pF/foot or 101 pF/meter for 50-Ω solid polyethylene insulated coax, which has a velocity factor of 0.66.

The general principles illustrated in Fig 39 can be extended to other switching systems. If switching the ground conductors as described above isn't practical, use of a metal box for the switching circuitry is recommended, so that the relatively large surface area of the box can be used for the common ground conductors, minimizing their inductance. Always keep leads extremely short.

Measuring the Electrical Length of Feed Lines

When using the feed methods described earlier the feed lines must be very close to the correct length. For best results, they should be correct within 1% or so. This means that a line that is intended to be, say, λ/4 at 7 MHz, should actually be λ/4 at some frequency within 70 kHz of 7 MHz. A simple but accurate method to determine at what frequency a line is λ/4 or λ/2 is shown in **Fig 40A**. The far end of the line is short circuited with a very short connection. A signal is applied to the input and the frequency is swept until the impedance at the input is a minimum. This is the frequency at which the

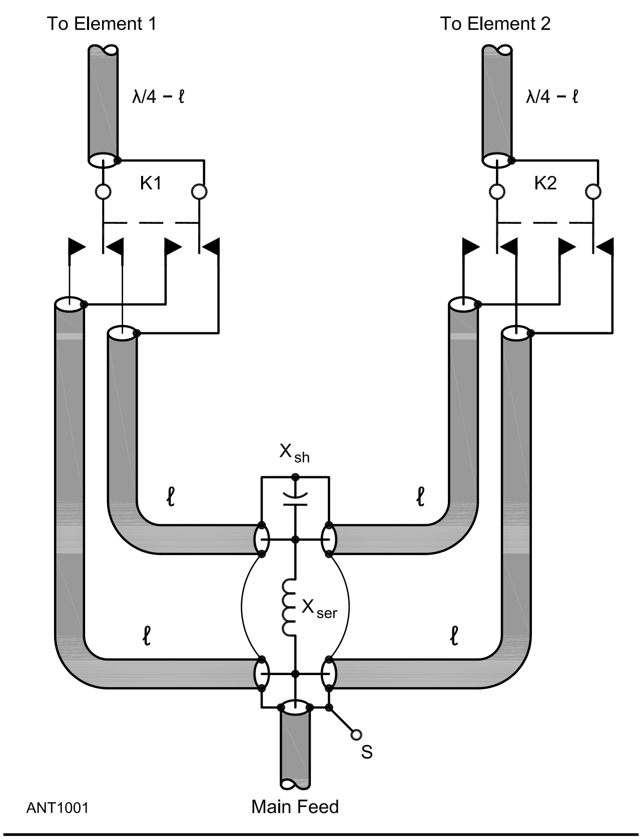

Fig 39—A carefully designed L-network, current-forcing switching system that switches both hot and shield conductors in feed coaxes.

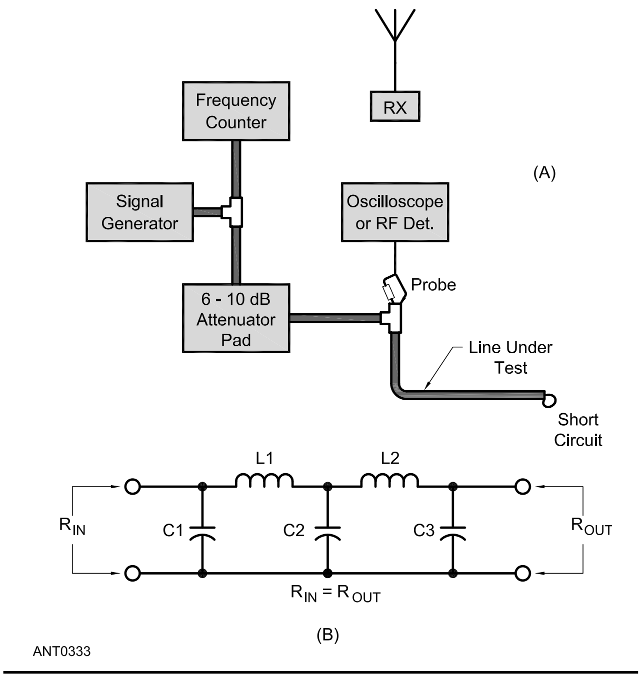

Fig 40—At A, the setup for measurement of the electrical length of a transmission line. The receiver may be used in place of the frequency counter to determine the frequency of the signal generator. The signal generator output must be free of harmonics; the half-wave harmonic filter at B may be used outboard if there is any doubt. It must be constructed for the frequency band of operation. Connect the filter between the signal generator and the attenuator pad.

C1, C3—Value to have a capacitive reactance = R_{IN}.

C2—Value to have a capacitive reactance = ½ R_{IN}.

L1, L2—Value to have an inductive reactance = R_{IN}.

line is λ/2. Either the frequency counter or the receiver may be used to determine this frequency. The line is, of course, λ/4 at one half the measured frequency.

The detector can be a simple diode detector or an oscilloscope may be used if available. A 6 to 10 dB attenuator pad is included to prevent the signal generator from looking into a short circuit at the measurement frequency. The signal generator output must be free of harmonics. If there is any doubt, an outboard low-pass filter, such as a half-wave harmonic filter, should be used. The half-wave filter circuit is shown in Fig 40B, and must be constructed for the frequency band of operation.

Another satisfactory method is to use a noise or resistance bridge or antenna analyzer at the input of the line, again looking for a low impedance at the input while the output is short circuited. Simple resistance bridges are described in Chapter 27.

Dip oscillators have been found to be unsatisfactory. The required coupling loop has too great an effect on measurements.

Measuring Element Self and Mutual Impedances

The need for measuring element self and mutual impedances has been made largely unnecessary with the ready availability of modeling software. Few amateurs appreciate the considerable difficulty of making accurate impedance measurements and accurate mutual impedance measurements are very difficult even with professional test equipment and skills. Despite the limitations of computer modeling, results very often are better than measured values because of the multiple factors affecting measurement accuracy.

Those who are interested in measuring self and mutual impedances can find more detailed information about doing so in Appendix B. The information there is from earlier editions of *The ARRL Antenna Book*.

Broadside Arrays

Broadside arrays can be made up of collinear or parallel elements or combinations of the two. This section was contributed by Rudy Severns, N6LF.

COLLINEAR ARRAYS

Collinear arrays are always operated with the elements in-phase. (If alternate elements in such an array are out-of-phase, the system simply becomes a harmonic type of antenna.) A collinear array is a broadside radiator, the direction of maximum radiation being at right angles to the line of the antenna.

POWER GAIN

Because of the nature of the mutual impedance between collinear elements, the feed-point resistance (compared to a single element, which is ≈73 Ω) is increased as shown earlier in this chapter (Fig 9). For this reason the power gain does not increase in direct proportion to the number of elements. The gain with two elements, as the spacing between them is varied, is shown by **Fig 41**. Although the gain is greatest when the end-to-end spacing is in the region of 0.4 to 0.6 λ, the use of spacings of this order is inconvenient constructionally and introduces problems in feeding the two elements. As a result, collinear elements are almost always operated with their ends

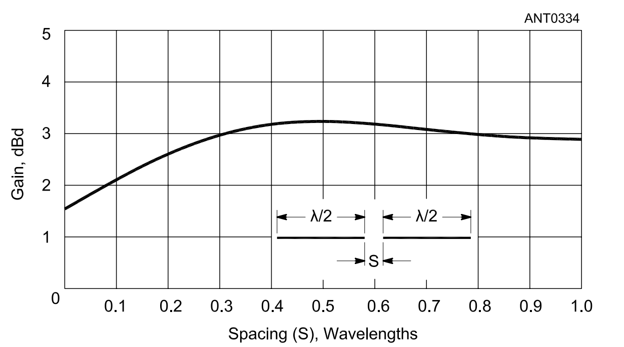

Fig 41—Gain of two collinear λ/2 elements as a function of spacing between the adjacent ends.

quite close together—in wire antennas, usually with just a strain insulator between.

With very small spacing between the ends of adjacent elements the theoretical power gain of collinear arrays, assuming the use of #12 copper wire, is approximately as follows over a dipole in free space:

2 collinear elements—1.6 dB
3 collinear elements—3.1 dB
4 collinear elements—3.9 dB

More than four elements are rarely used.

DIRECTIVITY

The directivity of a collinear array, in a plane containing the axis of the array, increases with its length. Small secondary lobes appear in the pattern when more than two elements are used, but the amplitudes of these lobes are low enough so that they are usually not important. In a plane at right angles to the array the directive diagram is a circle, no matter what the number of elements. Collinear operation, therefore, affects only E-plane directivity, the plane containing the antenna.

When a collinear array is mounted with the elements vertical, the antenna radiates equally well in all geographical directions. An array of such *stacked* collinear elements tends to confine the radiation to low vertical angles.

If a collinear array is mounted horizontally, the directive pattern in the vertical plane at right angles to the array is the same as the vertical pattern of a simple λ/2 antenna at the same height (Chapter 3).

TWO-ELEMENT ARRAYS

The simplest and most popular collinear array is one using two elements, as shown in **Fig 42**. This system is commonly known as *two half-waves in phase*. The directive pattern in a plane containing the wire axis is shown in **Fig 43**, which shows superimposed patterns for a dipole and 2, 3 and 4-element collinear arrays. Depending on the conductor size, height, and similar factors, the impedance at the feed point can be expected to be in the range of 4 to 6 kΩ, for wire antennas. If the elements are made of tubing having a low λ/dia (wavelength to diameter) ratio, values as low as 1 kΩ are representative. The system can be fed through an open-wire tuned line with negligible loss for ordinary line lengths, or a matching section may be used if desired.

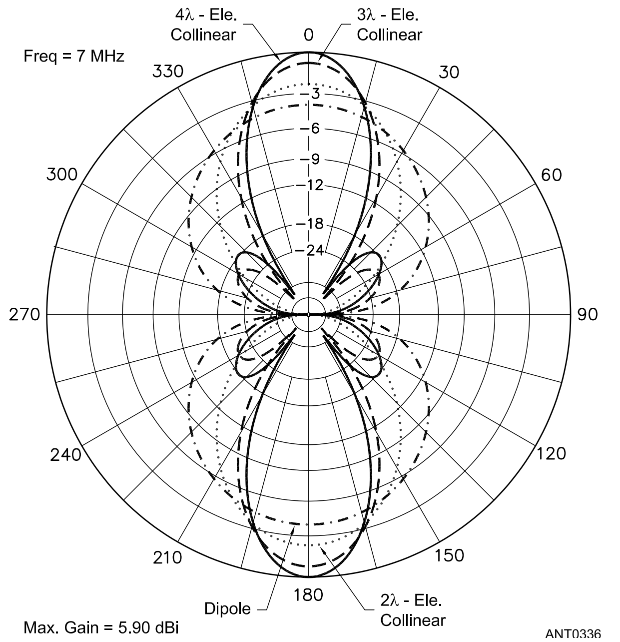

Fig 43—Free-space E-plane directive diagram for dipole, 2, 3 and 4-element collinear arrays. The solid line is a 4-element collinear; the dashed line is for a 3-element collinear; the dotted line is for a 2-element collinear and the dashed-dotted line is for a λ/2 dipole.

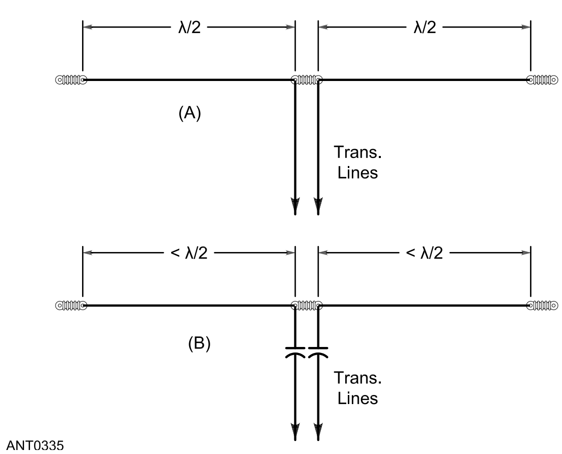

Fig 42—At A, two-element collinear array (two half-waves in phase). The transmission line shown would operate as a tuned line. A matching section can be substituted and a nonresonant line used if desired, as shown at B, where the matching section is two series capacitors.

A number of arrangements for matching the feed line to this antenna are described in Chapter 26. If elements somewhat shorter than λ/2 are used, then additional matching schemes can be employed at the expense of a slight reduction in gain. When the elements are shortened two things happen—the impedance at the feed-point drops and the impedance has inductive reactance that can be tuned out with simple series capacitors, as shown in Fig 42B.

Note that these capacitors must be suitable for the power level. Small *doorknob* capacitors such as those frequently used in power amplifiers, are suitable. By way of an example, if each side of a 40-meter 2-element array is shortened from 67 to 58 feet, the feed-point impedance drops from nearly 6000 Ω to about 1012 Ω with an inductive reactance of 1800 Ω. The reactance can be tuned out by inserting 25 pF capacitors at the feed-point. The 1012 Ω resistance can be transformed to 200 Ω using a λ/4 matching section made of 450-Ω ladder line and then transformed to 50 Ω with a 4:1 balun. Shortening the array as suggested reduces the gain by about 0.5 dB.

Another scheme that preserves the gain is to use a 450-Ω

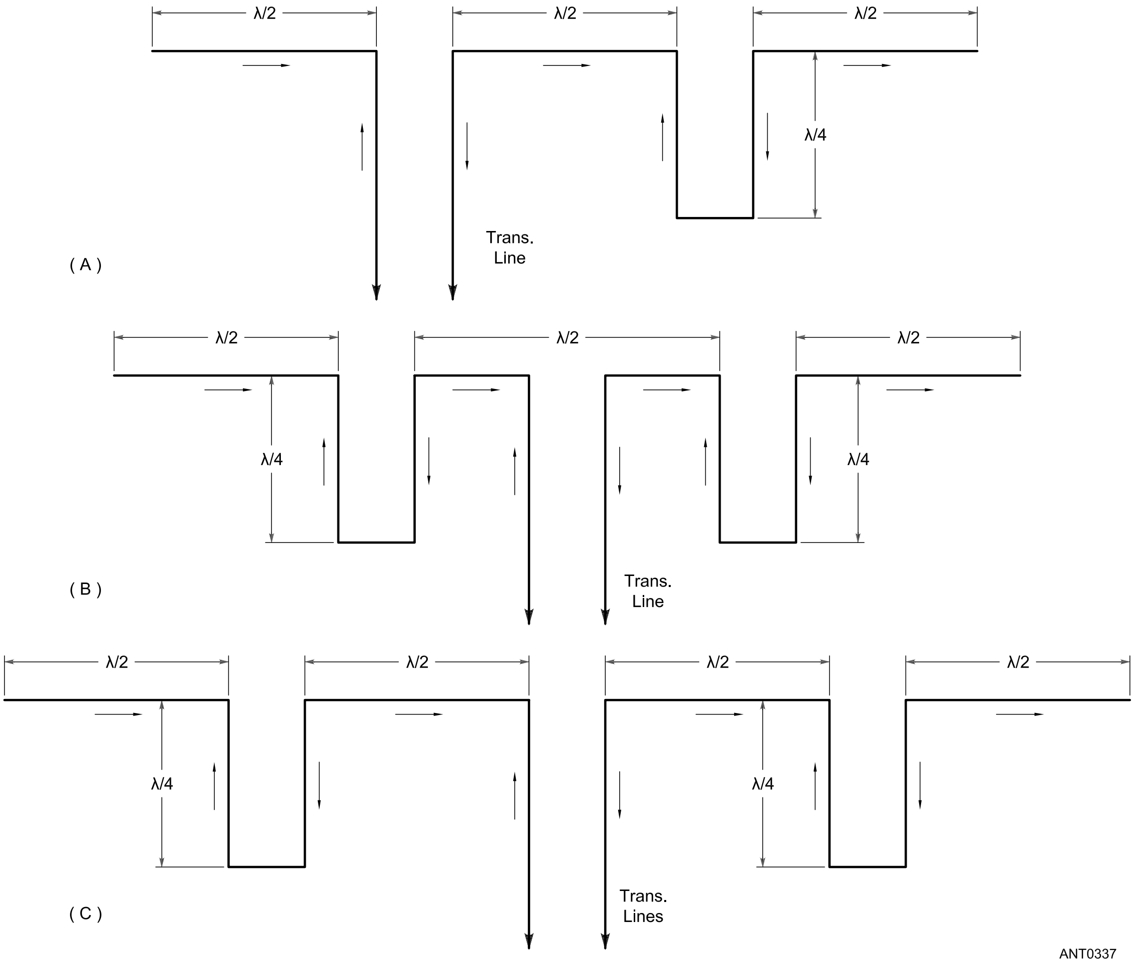

Fig 44—Layouts for 3- and 4-element collinear arrays. Alternative methods of feeding a 3-element array are shown at A and B. These drawings also show the current distribution on the antenna elements and phasing stubs. A matched transmission line can be substituted for the tuned line by using a suitable matching section.

λ/4 matching section and shorten the antenna only slightly to have a resistance of 4 kΩ. The impedance at the input of the matching section is then near 50 Ω and a simple 1:1 balun can be used. Many other schemes are possible. The free-space E-plane response for a 2-element collinear array is shown in Fig 43, compared with the responses for more elaborate collinear arrays described below.

THREE- AND FOUR-ELEMENT ARRAYS

In a long wire the direction of current flow reverses in each λ/2 section. Consequently, collinear elements cannot simply be connected end to end; there must be some means for making the current flow in the same direction in all elements. When more than two collinear elements are used it is necessary to connect *phasing* stubs between adjacent elements in order to bring the currents in all elements in-phase. In **Fig 44A** the direction of current flow is correct in the two left-hand elements because the shorted λ/4 transmission line (*stub*) is connected between them. This stub may be looked upon simply as the alternate λ/2 section of a long-wire antenna folded back on itself to cancel its radiation. In Fig 44A the part to the right of the transmission line has a total length of three half wavelengths, the center half wave being folded back to form a λ/4 phase-reversing stub. No data are available on the impedance at the feed point in this arrangement, but various considerations indicate that it should be over 1 kΩ.

An alternative method of feeding three collinear elements is shown in Fig 44B. In this case power is applied at the center of the middle element and phase-reversing stubs are used between this element and both of the outer elements. The impedance at the feed point in this case is somewhat over 300 Ω and provides a close match to 300 Ω line. The SWR will be less than 2:1 when 600-Ω line is used. Center feed of this type is somewhat preferable to the arrangement in Fig 44A because the system as a whole is balanced. This assures more uniform power distribution among the elements. In Fig 44A, the right-hand element is likely to receive somewhat less power than the other two because a portion of the input power is radiated by the middle element before it can reach the element located at the extreme right.

A four-element array is shown in Fig 44C. The system is symmetrical when fed between the two center elements as shown. As in the three-element case, no data are available on the impedance at the feed point. However, the SWR with a 600 Ω line should not be much over 2:1.

Fig 43 compares the directive patterns of 2, 3 and 4-element arrays. Collinear arrays can be extended to more than four elements. However, the simple 2-element collinear array is the type most frequently used, as it lends itself well to multi-band operation. More than two collinear elements are seldom used because more gain can be obtained from other types of arrays.

ADJUSTMENT

In any of the collinear systems described, the lengths of the radiating elements in feet can be found from the formula $468/f_{MHz}$. The lengths of the phasing stubs can be found from the equations given in Chapter 26 for the type of line used. If the stub is open-wire line (500 to 600 Ω impedance) you may assume a velocity factor of 0.975 in the formula for a λ/4 line. On-site adjustment is, in general, an unnecessary refinement. If desired, however, the following procedure may be used when the system has more than two elements.

Disconnect all stubs and all elements except those directly connected to the transmission line (in the case of feed such as is shown in Fig 44B leave only the center element connected to the line). Adjust the elements to resonance, using the still-connected element. When the proper length is determined, cut all other elements to the same length. Make the phasing stubs slightly long and use a shorting bar to adjust their length. Connect the elements to the stubs and adjust the stubs to resonance, as indicated by maximum current in the shorting bars or by the SWR on the transmission line. If more than three or four elements are used it is best to add elements two at a time (one at each end of the array), resonating the system each time before a new pair is added.

THE EXTENDED DOUBLE ZEPP

One method to obtain higher gain that goes with wider spacing in a simple system of two collinear elements is to make the elements somewhat longer than λ/2. As shown in **Fig 45**, this increases the spacing between the two in-phase λ/2 sections at the ends of the wires. The section in the center carries a current of opposite phase, but if this section is short the current will be small; it represents only the outer ends of a λ/2 antenna section. Because of the small current and short length, the radiation from the center is small. The optimum length for each element is 0.64 λ. At greater lengths the system tends to act as a long-wire antenna, and the gain decreases.

This system is known as the *extended double Zepp*. The gain over a λ/2 dipole is approximately 3 dB, as compared with about 1.6 dB for two collinear λ/2 dipoles. The directional pattern in the plane containing the axis of the antenna

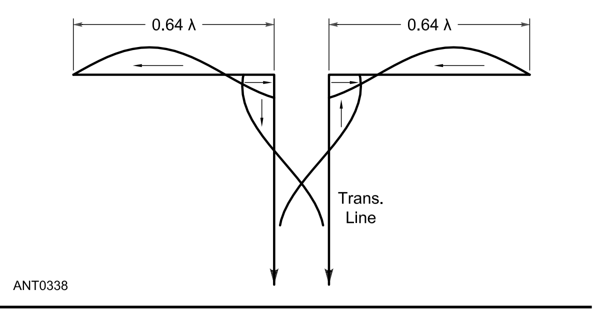

Fig 45—The extended double Zepp. This system gives somewhat more gain than two λ-sized collinear elements.

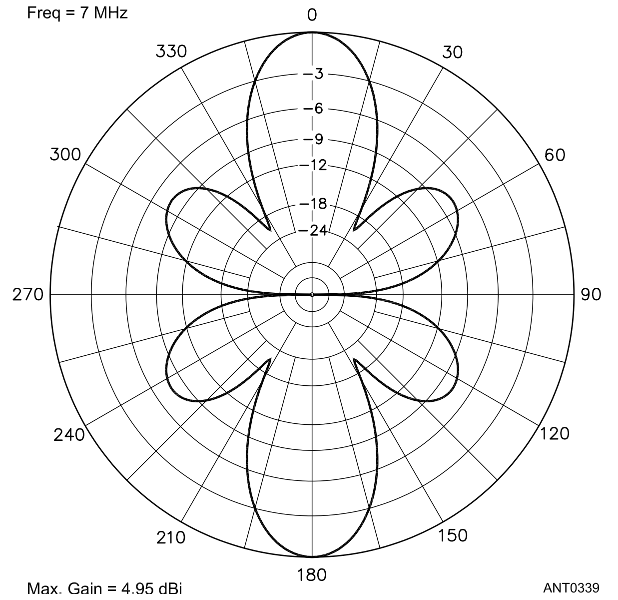

Fig 46—E-plane pattern for the extended double Zepp of Fig 45. This is also the horizontal directional pattern when the elements are horizontal. The axis of the elements lies along the 90°-270° line. The free-space array gain is approximately 4.95 dBi.

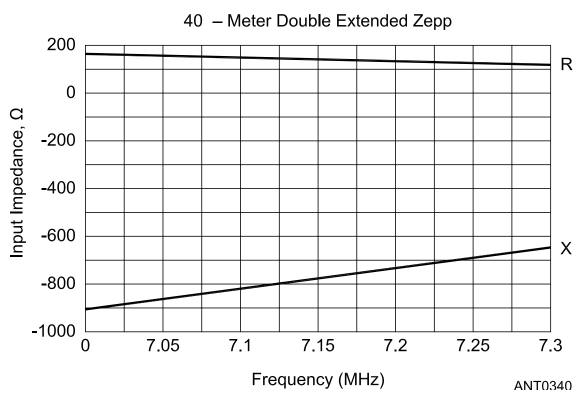

Fig 47—Resistive and reactive feed-point impedance of a 40-meter extended double Zepp in free space.

is shown in **Fig 46**. As in the case of all other collinear arrays, the free-space pattern in the plane at right angles to the antenna elements is the same as that of a λ/2 antenna—circular.

This antenna is not resonant at the operating frequency so that the feed-point impedance is complex (R ± *j*X). A typical example of the variation of the feed-point impedance over the band for a 40-meter double-extended Zepp is shown in **Fig 47**. This antenna is normally fed with open-wire transmission line to an antenna tuner. Other matching arrangements are, of course, possible. A method for transforming the feed-point impedance to 450 Ω and eliminating the minor lobes is given in Chapter 6.

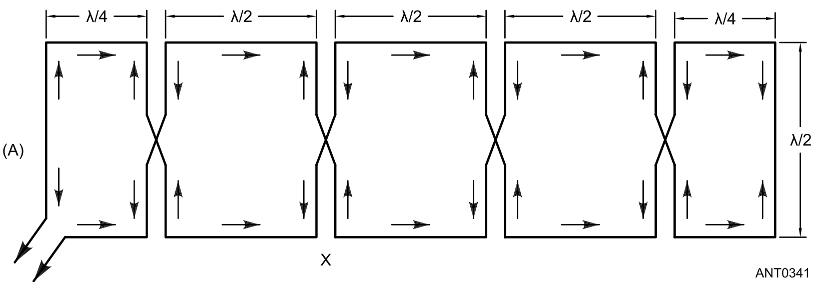

Fig 48—Typical Sterba array, an 8-element version.

THE STERBA ARRAY

Two collinear arrays can be combined to form the *Sterba array*, often called the *Sterba curtain*. An 8-element example of a Sterba array is shown in **Fig 48**. The four λ/4 elements joined on the ends are equivalent to two λ/2 elements. The two collinear arrays are spaced λ/2 and the λ/4 phasing lines connected together to provide λ/2 phasing lines. This arrangement has the advantage of increasing the gain for a given length and also increasing the E-plane directivity, which is no longer circular. An additional advantage of this array is that the wire forms a closed loop. For installations where icing is a problem a low voltage dc or low frequency (50 or 60 Hz) ac current can be passed through the wire to heat it for deicing. The heating current is isolated from RF by decoupling chokes. This is standard practice in commercial installations.

The number of sections in a Sterba array can be extended as far as desired but more than four or five are rarely used because of the slow increase in gain with extra elements, the narrow H-plane directivity and the appearance of multiple sidelobes. When fed at the point indicated the impedance is about 600 Ω. The antenna can also be fed at the point marked X. The impedance at this point will be about 1 kΩ. The gain of the 8-element array in Fig 48 will be between 7 to 8 dB over a single element.

Parallel Broadside Arrays

To obtain broadside directivity with parallel elements the currents in the elements must all be in-phase. At a distant point lying on a line perpendicular to the axis of the array and also perpendicular to the plane containing the elements, the fields from all elements add up in phase. The situation is like that pictured in Fig 1 in this chapter, where four parallel λ/2 dipoles were fed together a broadside array.

Broadside arrays of this type theoretically can have any number of elements. However, practical limitations of construction and available space usually limit the number of broadside parallel elements.

POWER GAIN

The power gain of a parallel-element broadside array depends on the spacing between elements as well as on the number of elements. The way in which the gain of a two-element array varies with spacing is shown in **Fig 49**. The greatest gain is obtained when the spacing is in the vicinity of 0.67 λ.

The theoretical gains of broadside arrays having more than two elements are approximately as follows:

No. of Parallel Elements	*dB Gain with λ/2 Spacing*	*dB Gain with 3λ/4 Spacing*
3	5.7	7.2
4	7.1	8.5
5	8.1	9.4
6	8.9	10.4

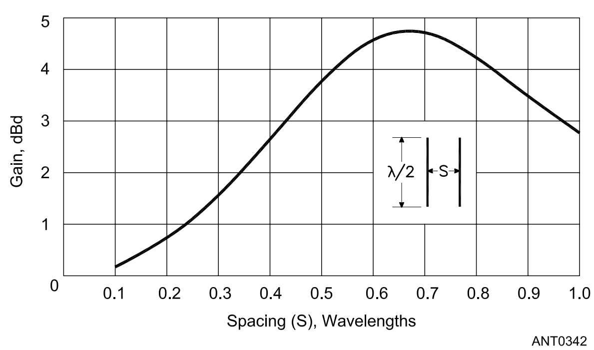

Fig 49—Gain as a function of the spacing between two parallel elements operated in-phase (broadside).

The elements must, of course, all lie in the same plane and all must be fed in-phase.

DIRECTIVITY

The sharpness of the directive pattern depends on spacing between elements and number of elements. Larger element spacing will sharpen the main lobe, for a given number of elements, up to a point as was shown in Fig 41. The two-element array has no minor lobes when the spacing is λ/2, but small minor lobes appear at greater spacings. When three or more elements are used the pattern always has minor lobes.

Other Forms Of Broadside Arrays

For those who have the available room, multi-element arrays based on the broadside concept have something to offer. The antennas are large but of simple design and non-critical dimensions; they are also very economical in terms of gain per unit of cost.

Large arrays can often be fed at several different points. However, the pattern symmetry may be sensitive to the choice of feed point within the array. Non-symmetrical feed points will result in small asymmetries in the pattern but these are not usually of great concern.

Arrays of three and four elements are shown in **Fig 50**. In the 3-element array with λ/2 spacing at A, the array is fed at the center. This is the most desirable point in that it tends to keep the power distribution among the elements uniform. However, the transmission line could alternatively be connected at either point B or C of Fig 50A, with only slight skewing of the radiation pattern.

When the spacing is greater than λ/2, the phasing lines must be 1 λ long and are not transposed between elements. This is shown Fig 50B. With this arrangement, any element spacing up to 1 λ can be used, if the phasing lines can be folded as suggested in the drawing.

The 2-element array at C is fed at the center of the system to make the power distribution among elements as uniform as possible. However, the transmission line could be connected at either point B, C, D or E. In this case the section of phasing line between B and D must be transposed to make the currents flow in the same direction in all elements. The 4-element array at C and the 3-element array at B have approximately the same gain when the element spacing in the array at B is 3λ/4.

An alternative feeding method is shown in Fig 50D. This system can also be applied to the 3-element arrays, and will result in better symmetry in any case. It is necessary only to move the phasing line to the center of each element, making connection to both sides of the line instead of one only.

The free-space pattern for a 4-element array with λ/2 spacing is shown in **Fig 51**. This is also approximately the pattern for a 3-element array with 3λ/4 spacing.

Larger arrays can be designed and constructed by following the phasing principles shown in the drawings. No accurate figures are available for the impedances at the various feed points indicated in Fig 50. You can estimate it to be in the vicinity of 1 kΩ when the feed point is at a junction between the phasing line and a λ/2 element, becoming smaller as the number of elements in the array is increased. When the feed point is midway between end-fed elements as in Fig 50C, the feed-point impedance of a 4-element array is in the vicinity of 200 to 300 Ω, with 600-Ω open-wire phasing lines. The impedance at the feed point with the antenna shown at D should be about 1.5 kΩ.

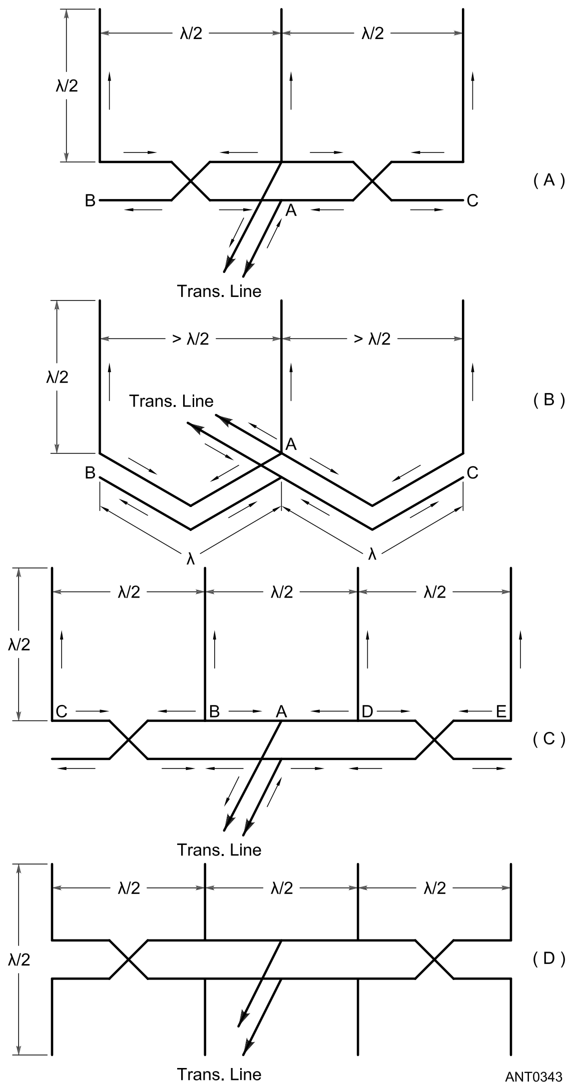

Fig 50—Methods of feeding 3- and 4-element broadside arrays with parallel elements.

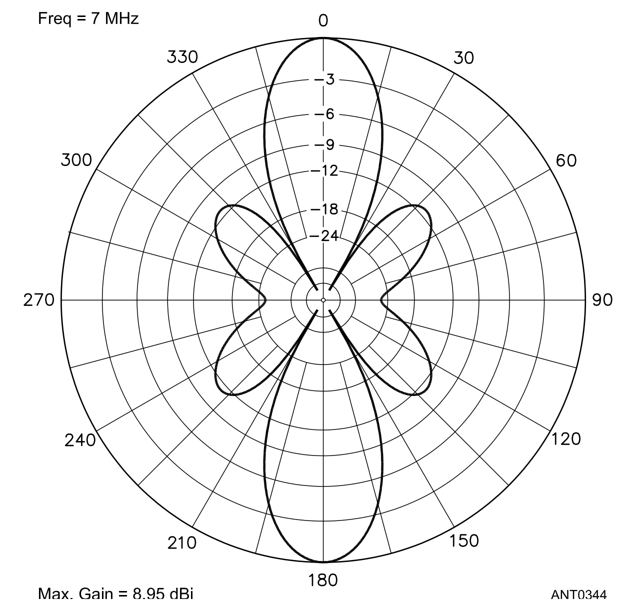

Fig 51—Free-space E-plane pattern of a 4-element broadside array using parallel elements (Fig 50). This corresponds to the horizontal directive pattern at low wave angles for a vertically polarized array over ground. The axis of the elements lies along the 90°-270° line.

NON-UNIFORM ELEMENT CURRENTS

The pattern for a 4-element broadside array shown in Fig 51 has substantial side lobes. This is typical for arrays more than λ/2 wide when equal currents flow in each element. Sidelobe amplitude can be reduced by using non-uniform current distribution among the elements. Many possible current amplitude distributions have been suggested. All of them have reduced current in the outer elements and greater current in the inner elements. This reduces the gain somewhat but can produce a more desirable pattern. One of the common current distributions is called *binomial current grading*. In this scheme the ratio of element currents is set equal to the coefficients of a polynomial. For example:

$$
\begin{aligned}
&1\,x+1, \quad \Rightarrow \; 1,1 \\
&(x+1)^2 = 1x^2+2x+1, \quad \Rightarrow \; 1, 2, 1 \\
&(x+1)^3 = 1x^3+3x^2+3x+1, \quad \Rightarrow \; 1, 3, 3, 1 \\
&(x+1)^4 = 1x^4+4x^3+6x^2+6x+1, \quad \Rightarrow \; 1, 4, 6, 4, 1
\end{aligned}
\qquad \text{(Eq 7)}
$$

In a 2-element array the currents are equal, in a 3-element array the current in the center element is twice that in the outer elements, and so on.

HALF-SQUARE ANTENNA

On the low-frequency bands (40, 80 and 160 meters) it becomes increasingly difficult to use λ/2 elements because of their size. The half-square antenna is a 2-element broadside array with λ/4-high vertical elements and λ/2 horizontal spacing. See **Fig 52**. The free-space H-plane pattern for this array is shown in **Fig 53**. The antenna gives modest (4.2 dBi) but useful gain and has the advantage of only λ/4 height. Like all vertically polarized antennas, real-world performance depends directly on the characteristics of the ground surrounding it.

The half-square can be fed either at the point indicated or at the bottom end of one of the vertical elements using a voltage-feed scheme, such as that shown in **Fig 54** for the bobtail curtain. The feed-point impedance is in the region of 50 Ω when fed at a corner as shown in Fig 52. A typical SWR plot is shown in **Fig 55**. Chapter 6 has a detailed discussion of the half-square antenna with several variations, together with practical considerations.

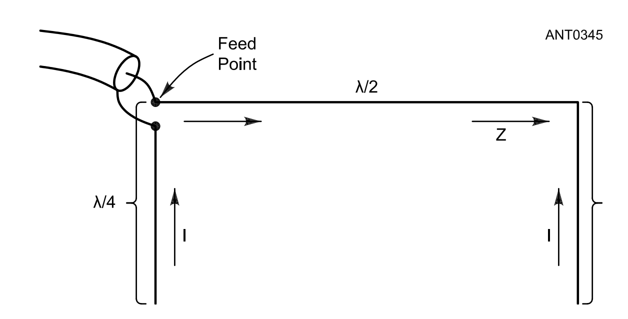

Fig 52—Layout for the half-square antenna.

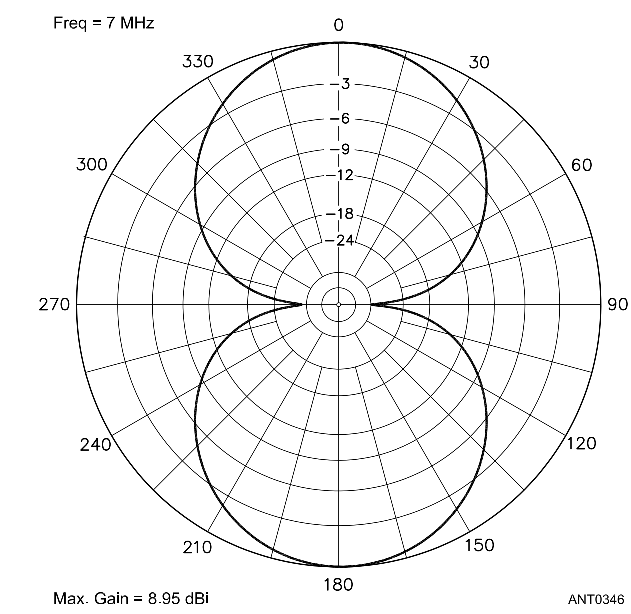

Fig 53—Free-space E-plane directive pattern for the half-square antenna.

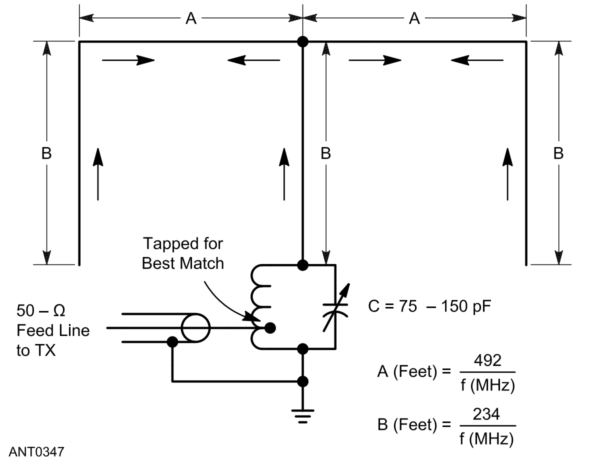

Fig 54—The bobtail curtain is an excellent low-angle radiator having broadside bidirectional characteristics. Current distribution is represented by the arrows. Dimensions A and B (in feet, for wire antennas) can be determined from the equations.

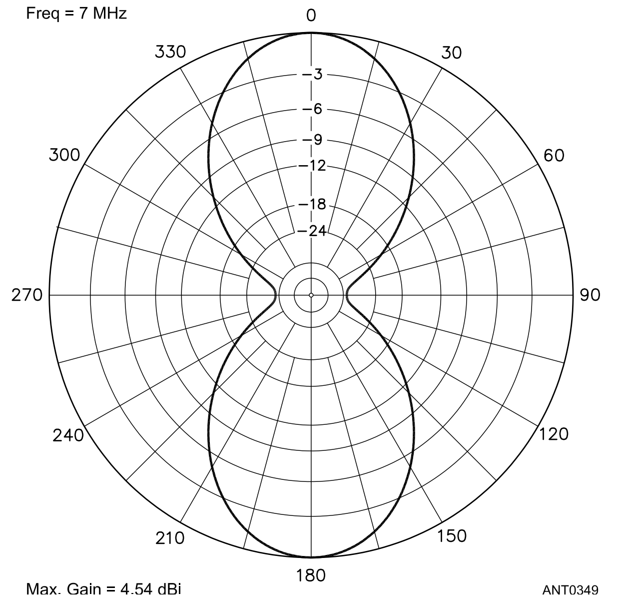

Fig 56—Calculated free-space E-plane directive diagram of the bobtail curtain shown in Fig 54. The array lies along the 90°-270° axis.

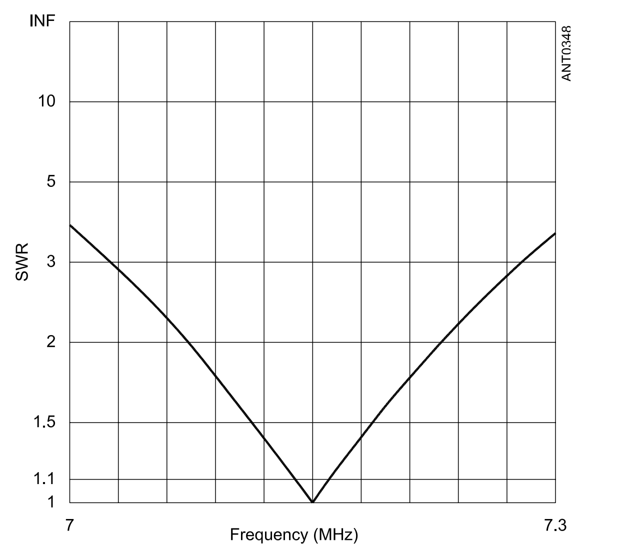

Fig 55—Typical SWR plot for a 40-meter half-square antenna fed at one corner. Antenna in free space.

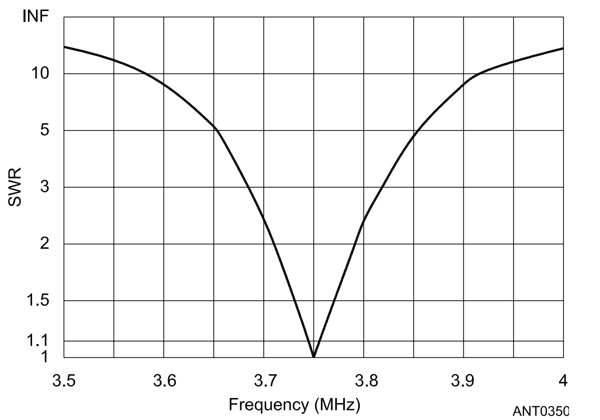

Fig 57—Typical SWR plot for an 80-meter bobtail curtain in free space. This is a narrow-band antenna.

BOBTAIL CURTAIN

The antenna system in Fig 54 uses the principles of co-phased verticals to produce a broadside, bidirectional pattern providing approximately 5.1 dB of gain over a single λ/4 element. The antenna performs as three in-phase, top-fed vertical radiators approximately λ/4 in height and spaced approximately λ/2. It is most effective for low-angle signals and makes an excellent long-distance antenna for 1.8, 3.5 or 7 MHz.

The three vertical sections are the actual radiating components, but only the center element is fed directly. The two horizontal parts, A, act as phasing lines and contribute very little to the radiation pattern. Because the current in the center element must be divided between the end sections, the current distribution approaches a binomial 1:2:1 ratio. The radiation pattern is shown in **Fig 56**.

The vertical elements should be as vertical as possible. The height for the horizontal portion should be slightly greater than B, as shown in Fig 54. The tuning network is resonant at the operating frequency. The L/C ratio should be fairly low to provide good loading characteristics. As a starting point, a maximum capacitor value of 75 to 150 pF is recommended, and the inductor value is determined by C and the operating frequency. The network is first tuned to resonance and then the tap point is adjusted for the best match. A slight readjustment of C may be necessary. A link coil consisting of a few turns can also be used to feed the antenna.

A feeling for the matching bandwidth of this antenna can be obtained by looking at a feed point located at the top end of the center element. The impedance at this point will be approximately 32 Ω. An SWR plot (for $Z_0 = 32\ \Omega$) for an 80-meter bobtail curtain at this feed-point is shown in **Fig 57**. However, it is not advisable to actually connect a feed line at this point since it would detune the array and alter the pattern. This antenna is relatively narrow band. When fed at the bottom of the center element as shown in Fig 54, the SWR can be adjusted to be 1:1 at one frequency but the operating bandwidth for SWR < 2:1 may be even narrower than Fig 57 shows. For 80-meters, where operation is often desired in the CW DX window (3.510 MHz) and in the phone DX window (3.790 MHz), it will be necessary to retune the matching network as you change frequency. This can be done by switching a capacitor in or out, manually or remotely with a relay.

While the match bandwidth is quite narrow, the radiation pattern changes more slowly with frequency. **Fig 58** shows the variation in the pattern over the entire band (3.5 to 4.0 MHz). As would be expected, the gain increases with frequency because the antenna is larger in terms of wavelengths. The general shape of the pattern, however, is quite stable.

THE BRUCE ARRAY

Four variations of the Bruce array are shown in **Fig 59**. The Bruce is simply a wire folded so that the vertical sections carry large in-phase currents, while the horizontal sections carry small currents flowing in opposite directions with respect to the center of a section (indicated by dots). The radiation is vertically polarized. The gain is proportional to the length of the array but is somewhat smaller than you can obtain from a broadside array of λ/2 elements of the same length. This is because the radiating portion of the elements is only λ/4.

The Bruce array has a number of advantages:

1) The array is only λ/4 high. This is especially helpful on 80 and 160 meters, where the height of λ/2 supports becomes impractical for most amateurs.

2) The array is very simple. It is just a single piece of wire folded to form the array.

3) The dimensions of the array are very flexible. Depending on the available distance between supports, any number of elements can be used. The longer the array, the greater the gain.

4) The shape of the array does not have to be exactly 1.05 λ/4 squares. If the available height is short but the array can be made longer, then shorter vertical sections and longer horizontal sections can be used to maintain gain and resonance. Conversely, if more height is available but width is restricted then longer vertical sections can be used with shorter horizontal sections.

5) The array can be fed at other points more convenient for a particular installation.

6) The antenna is relatively low Q, so that the feed-point impedance changes slowly with frequency. This is very helpful on 80 meters, for example, where the antenna can be relatively broadband.

7) The radiation pattern and gain is stable over the width of an amateur band.

Note that the nominal dimensions of the array in Fig 59 call for section lengths = 1.05 λ/4. The need to use slightly longer elements to achieve resonance is common in large wire arrays. A quad loop behaves in the same manner. This is quite different from wire dipoles, which are typically *shortened* by 2-5% to achieve resonance.

Fig 60 shows the variations in gain and pattern for 2 to 5-element 80-meter Bruce arrays. **Table 2** lists the gain over a vertical λ/2 dipole, a 4-radial ground-plane vertical and the size of the array. The gain and impedance parameters listed are for free space. Over real ground the patterns and gain will depend on the height above ground and the ground characteristics. Copper loss using #12 conductors in included.

Worthwhile gain can be obtained from these arrays, especially on 80 and 160 meters, where any gain is hard to come by. The feed-point impedance is for the center of a vertical section. From the patterns in Fig 60 you can see that sidelobes start to appear as the length of the array is increased beyond 3λ/4. This is typical for arrays using equal currents in the elements.

It is interesting to compare the bobtail curtain (Fig 54) with a 4-element Bruce array. **Fig 61** compares the radiation patterns for these two antennas. Even though the Bruce is shorter (3λ/4) than the bobtail (1 λ), it has slightly more

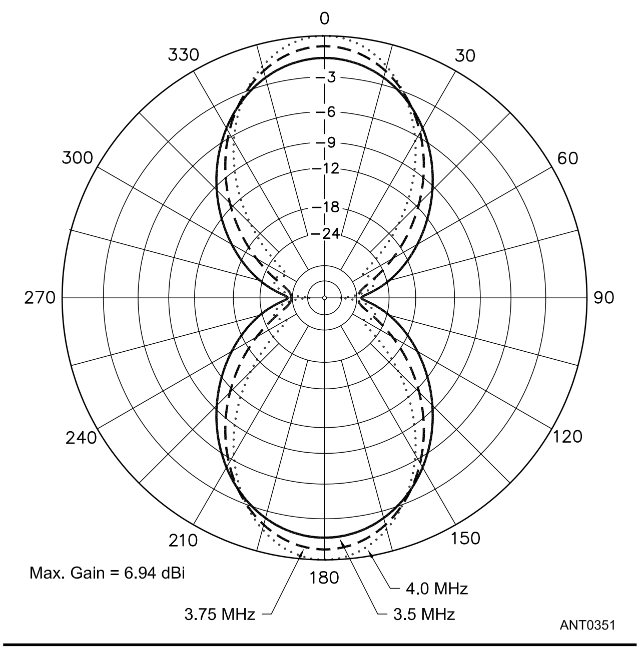

Fig 58—80-meter bobtail curtain's free-space E-plane pattern variation over the 80-meter band.

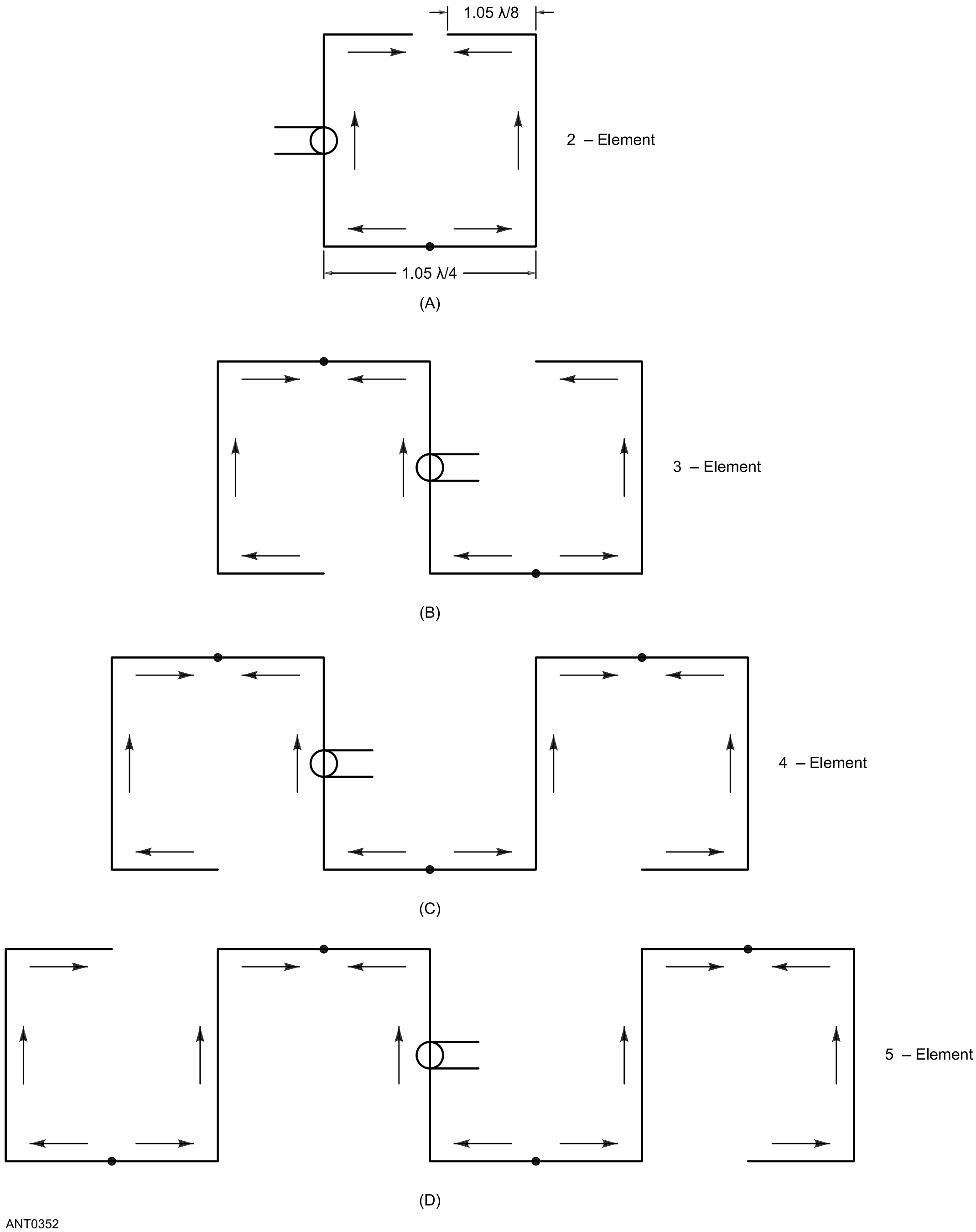

Fig 59—Various Bruce arrays: 2, 3, 4 and 5-element versions.

Table 2
Bruce array length, impedance and gain as a function of number of elements

Number Elements	*Gain Over λ/2 Vertical Dipole*	*Gain over λ/4 Ground-Plane*	*Array Length Wavelengths*	*Approx. Feed Z, Ω*
2	1.2 dB	1.9 dB	¼	130
3	2.8 dB	3.6 dB	½	200
4	4.3 dB	5.1 dB	¾	250
5	5.3 dB	6.1 dB	1	300

gain. The matching bandwidth is illustrated by the SWR curve in **Fig 62**. The 4-element Bruce has over twice the match bandwidth (200 kHz) than does the bobtail (75 kHz in Fig 57). Part of the gain difference is due to the binomial current distribution—the center element has twice the current as the outer elements in the bobtail. This reduces the gain slightly so that the 4-element Bruce becomes competitive. This is a good example of using more than the minimum number of elements to improve performance or to reduce size. On 160 meters the 4-element Bruce will be 140 feet shorter than the bobtail, a significant reduction. If additional space is available for the bobtail (1 λ) then a 5-element Bruce could

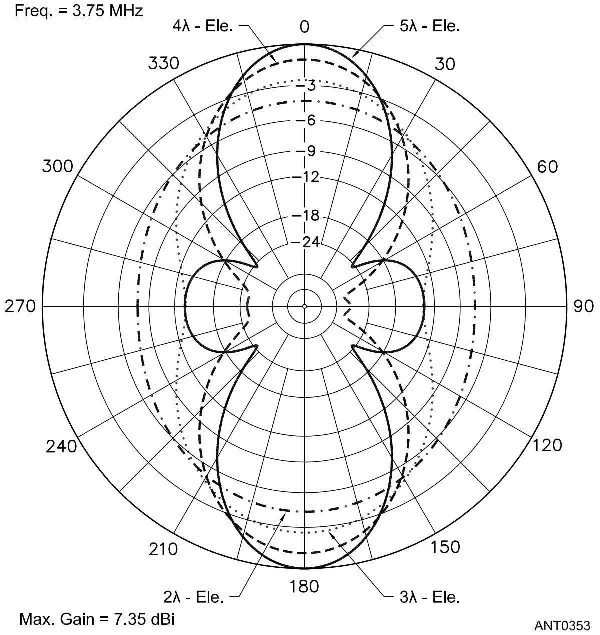

Fig 60—80-meter free-space E-plane directive patterns for the Bruce arrays shown in Fig 59. The 5-element's pattern is a solid line; the 4-element is a dashed line; the 3-element is a dotted line, and the 2-element version is a dashed-dotted line.

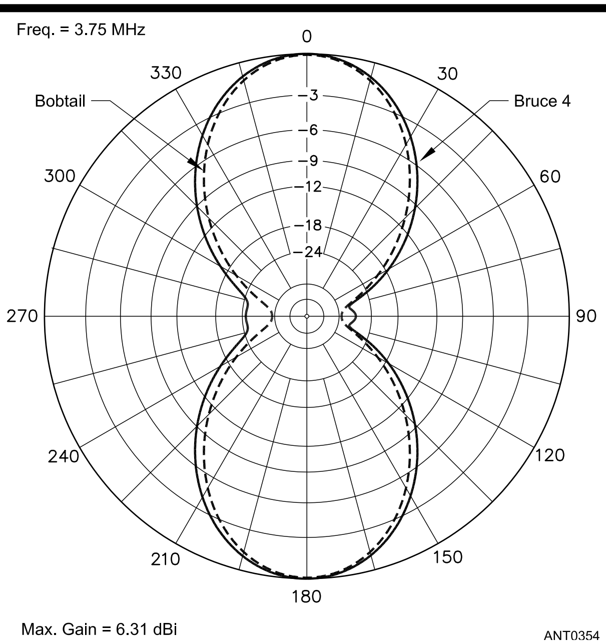

Fig 61—Comparison of free space patterns of a 4-element Bruce array (solid line) and a 3-element bobtail curtain (dashed line).

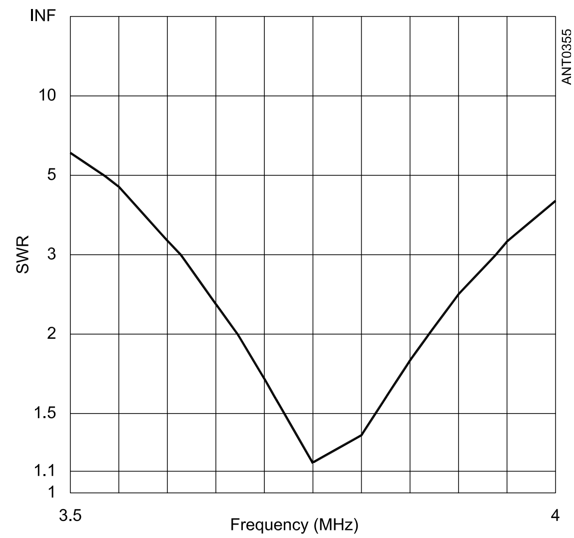

Fig 62—Typical SWR curve for a 4-element 80-meter Bruce array.

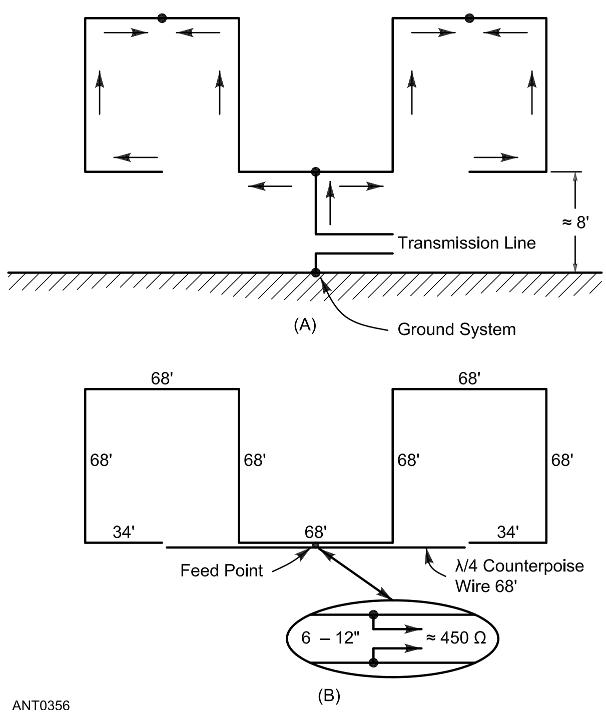

Fig 63—Alternate feed arrangements for the Bruce array. At A, the antenna is driven against a ground system and at B, it uses a two-wire counterpoise.

be used, with a small increase in gain but also introducing some sidelobes.

The 2-element Bruce and the half-square antennas are both 2-element arrays. However, since the spacing between radiators is greater in the half-square (λ/2) the gain of the half-square is about 1 dB greater. If space is available, the half-square would be a better choice. If there is not room for a half-square then the Bruce, which is only half as long (λ/4), may be a good alternative. The 3-element Bruce, which has the same length (λ/2) as the half-square, has about 0.6 dB more gain than the half-square and will have a wider match bandwidth.

The Bruce antenna can be fed at many different points and in different ways. In addition to the feed points indicated in Fig 59, you may connect the feed line at the center of any of the vertical sections. In longer Bruce arrays, feeding at one end will result in some current imbalance among the elements but the resulting pattern distortion is small. Actually, the feed-point can be anywhere along a vertical section. One very convenient point is at an outside corner. The feed-point impedance will be higher (about 600 Ω). A good match for 450-Ω ladder-line can usually be found somewhere on the vertical section. It is important to recognize that feeding the antenna at a voltage node (dots in Fig 59) by breaking the wire and inserting an insulator, completely changes the current distribution. This will be discussed in the section on endfire arrays.

A Bruce can be fed unbalanced against ground or against a counterpoise as shown in **Fig 63**. Because it is a vertically polarized antenna, the better the ground system, the better the performance. As few as two elevated radials can be used as shown in Fig 63B, but more radials can also be used to improve the performance, depending on local ground constants. The original development of the Bruce array in the late 1920s used this feed arrangement.

FOUR-ELEMENT BROADSIDE ARRAY

The 4-element array shown in **Fig 64** is commonly known as the *lazy H*. It consists of a set of two collinear elements and a set of two parallel elements, all operated in-phase to give broadside directivity. The gain and directivity will depend on the spacing, as in the case of a simple parallel-element broadside array. The spacing may be chosen between the limits shown on the drawing, but spacings below 3λ/8 are not worthwhile because the gain is small. Estimated gains compared to a single element are:

3λ/8 spacing—4.2 dB
λ/2 spacing—5.8 dB
5λ/8 spacing—6.7 dB
3λ/4 spacing—6.3 dB

Half-wave spacing is generally used. Directive patterns for this spacing are given in **Figs 65** and **66**. With λ/2 spacing between parallel elements, the impedance at the junction of the phasing line and transmission line is resistive and in the vicinity of 100 Ω. With larger or smaller spacing the impedance at this junction will be reactive as well as resistive. Matching stubs are recommended in cases where a non-resonant line is to be used. They may be calculated and adjusted as described in Chapter 26.

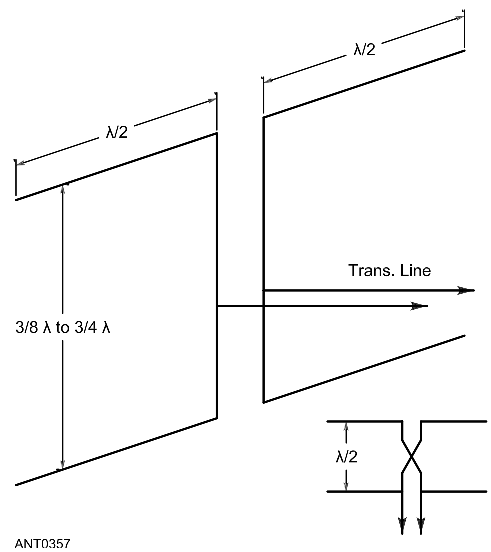

Fig 64—Four-element broadside array ("lazy H") using collinear and parallel elements.

The system shown in Fig 64 may be used on two bands having a 2-to-1 frequency relationship. It should be designed for the higher of the two frequencies, using 3λ/4 spacing between parallel elements. It will then operate on the lower frequency as a simple broadside array with 3λ/8 spacing.

An alternative method of feeding is shown in the small diagram in Fig 64. In this case the elements and the phasing line must be adjusted exactly to an electrical half wavelength. The impedance at the feed point will be resistive and on the order of 2 kΩ.

THE BI-SQUARE ANTENNA

A development of the lazy H, known as the *bi-square antenna*, is shown in **Fig 67**. The gain of the bi-square is somewhat less than that of the lazy-H, but this array is attractive because it can be supported from a single pole. It has a circumference of 2 λ at the operating frequency, and is horizontally polarized.

The bi-square antenna consists of two 1 λ radiators, fed 180° out-of-phase at the bottom of the array. The radiation resistance is 300 Ω, so it can be fed with either 300- or 600-Ω line. The free space gain of the antenna is about 5.8 dBi, which is 3.7 dB more than a single dipole element. Gain may be increased by adding a parasitic reflector or director. Two bi-square arrays can be mounted at right angles and switched to provide omnidirectional coverage. In this way, the antenna wires may be used as part of the guying system for the pole.

Although it resembles a loop antenna, the bi-square is not a true loop because the ends opposite the feed point

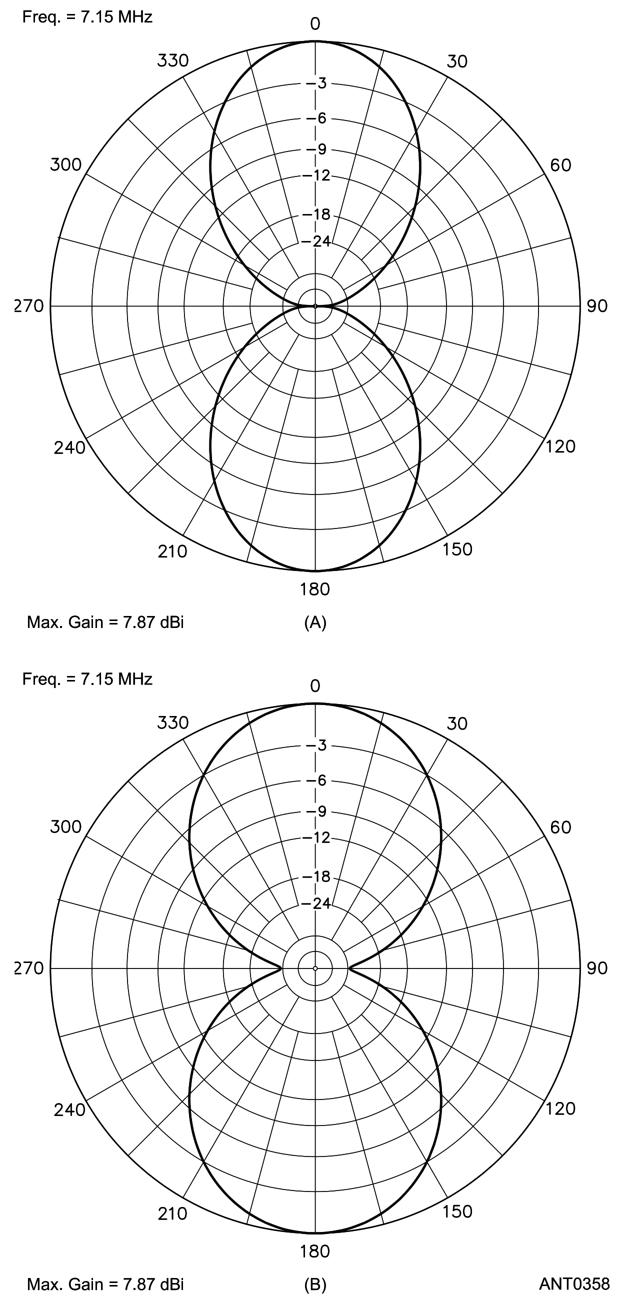

Fig 65—Free-space directive diagrams of the 4-element antenna shown in Fig 64. At A is the E-plane pattern. The axis of the elements lies along the 90°-270° line. At B is the free-space H-plane pattern, viewed as if one set of elements is above the other from the ends of the elements.

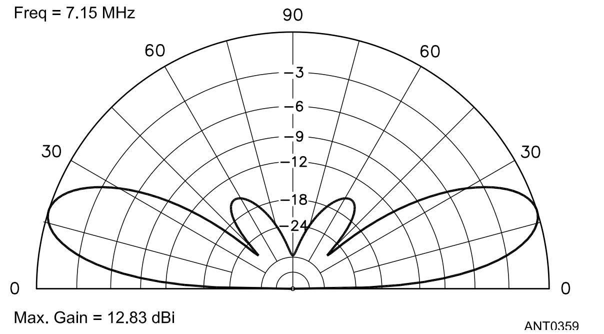

Fig 66—Vertical pattern of the 4-element broadside antenna of Fig 64, when mounted with the elements horizontal and the lower set λ/4 above flat ground. Stacked arrays of this type give best results when the lowest elements are at least λ/2 high. The gain is reduced and the wave angle raised if the lowest elements are too close to ground.

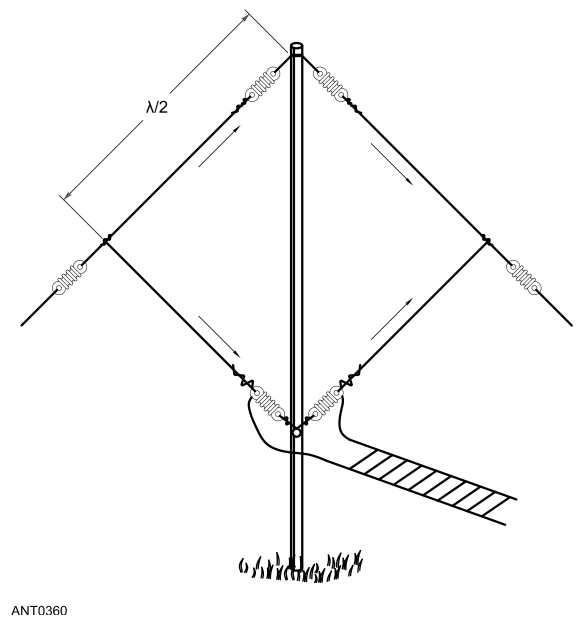

Fig 67—The bi-square array. It has the appearance of a loop, but is not a true loop because the conductor is open at the top. The length of each side, in feet, is 480/f (MHz).

are open. However, identical construction techniques can be used for the two antenna types. Indeed, with a means of remotely closing the connection at the top for lower frequency operation, the antenna can be operated on two harmonically related bands. As an example, an array with 17 feet per side can be operated as a bi-square at 28 MHz and as a full-wave loop at 14 MHz. For two-band operation in this manner, the side length should favor the higher frequency. The length of a closed loop is not as critical.

End-Fire Arrays

The term *end-fire* covers a number of different methods of operation, all having in common the fact that the maximum radiation takes place along the array axis, and that the array consists of a number of parallel elements in one plane. End-fire arrays can be either bidirectional or unidirectional. In the bidirectional type commonly used by amateurs there are only two elements, and these are operated with currents 180° out-of-phase. Even though adjustment tends to be complicated, unidirectional end-fire driven arrays have also seen amateur use, primarily as a pair of phased, ground-mounted λ/4 vertical elements. Extensive discussion of this array is contained in earlier sections of this chapter.

Horizontally polarized unidirectional end-fire arrays see little amateur use except in log-periodic arrays (described in Chapter 10). Instead, horizontally polarized unidirectional arrays usually have parasitic elements (described in Chapter 11) and are called Yagis.

TWO-ELEMENT END-FIRE ARRAY

In a 2-element array with equal currents out-of-phase, the gain varies with the spacing between elements as shown in **Fig 68**. The maximum gain occurs in the neighborhood of 0.1 λ spacing. Below that the gain drops rapidly due to conductor loss resistance.

The feed-point resistance for either element is very low at the spacings giving greatest gain, as shown in Fig 8 earlier in this chapter. The spacings most frequently used are λ/8 and λ/4, at which the resistances of center-fed λ/2 elements are about 9 and 32 Ω, respectively.

The effect of conductor resistance on gain for various spacings is shown in **Fig 69**. Because current along the element is not constant (it is approximately sinusoidal), the resistance shown is the equivalent resistance (R_{eq}) inserted at the center of the element to account for the loss distributed along the element.

The equivalent resistance of a λ/2 element is one half the ac resistance (R_{ac}) of the complete element. R_{ac} is usually >> R_{dc} due to skin effect. For example, a 1.84 MHz dipole using #12 copper wire will have the following R_{eq}:

Wire length = 267 feet
Rdc = 0.00159 [/foot] × 267 [feet] = 0.42
Fr = Rac/Rdc = 10.8
Req =(Rdc/2) × Fr = 2.29

For a 3.75 MHz dipole made with #12 wire, Req = 1.59 Ω. In Fig 69, it is clear that end-fire antennas made with #12 or smaller wire will limit the attainable gain because of losses. There is no point in using spacings much less than λ/4 if you use wire elements. If instead you use elements made of aluminum tubing then smaller spacings can be used to increase gain. However, as the spacing is reduced below λ/4 the increase in gain is quite small even with good conductors. Closer spacings give little gain increase but can drastically reduce the operating bandwidth due to the rapidly increasing Q of the array.

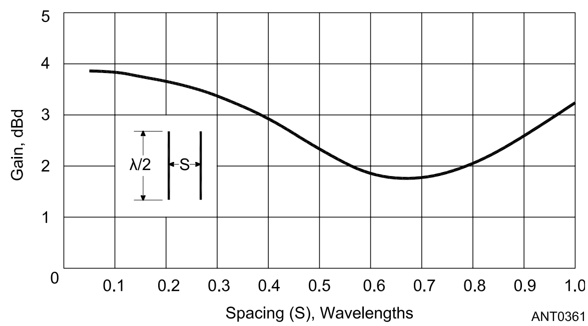

Fig 68—Gain of an end-fire array consisting of two elements fed 180° out-of-phase, as a function of the spacing between elements. Maximum radiation is in the plane of the elements and at right angles to them at spacings up to λ/2, but the direction changes at greater spacings.

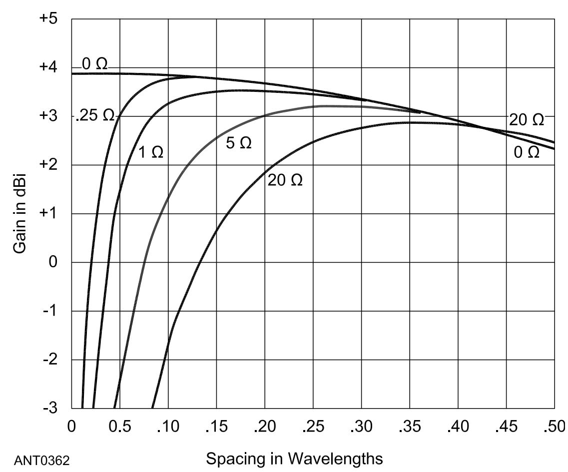

Fig 69—Gain over a single element of two out-of-phase elements in free space as a function of spacing for various loss resistances.

Unidirectional End-Fire Arrays

Two parallel elements spaced λ/4 apart and fed equal currents 90° out-of-phase will have a directional pattern in the plane at right angles to the plane of the array. See **Fig 70**. The maximum radiation is in the direction of the element in which the current lags. In the opposite direction the fields from the two elements cancel.

When the currents in the elements are neither in-phase nor 180° out-of-phase, the feed-point resistances of the elements are not equal. This complicates the problem of feeding equal currents to the elements, as discussed in earlier sections.

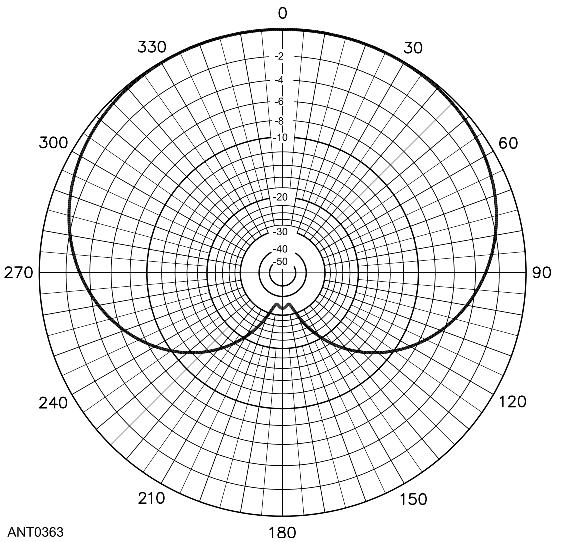

Fig 70—Representative H-plane pattern for a 2-element end-fire array with 90° spacing and phasing. The elements lie along the vertical axis, with the uppermost element the one of lagging phase. Dissimilar current distributions are taken into account. (Pattern computed with *ELNEC*.)

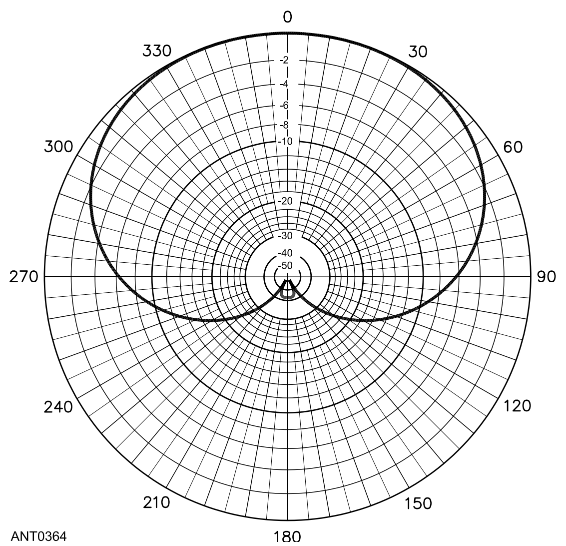

Fig 71—H-plane pattern for a 3-element end-fire array with binomial current distribution (the current in the center element is twice that in each end element). The elements are spaced λ/4 apart along the 0°-180° axis. The center element lags the lower element by 90°, while the upper element lags the lower element by 180° in phase. Dissimilar current distributions are taken into account. (Pattern computed with *ELNEC*.)

More than two elements can be used in a unidirectional end-fire array. The requirement for unidirectivity is that there must be a progressive phase shift in the element currents equal to the spacing, in electrical degrees, between the elements. The amplitudes of the currents in the various elements also must be properly related. This requires binomial current distribution. In the case of three elements, this requires that the current in the center element be twice that in the two outside elements, for 90° (λ/4) spacing and element current phasing. This antenna has an overall length of λ/2. The directive diagram is shown in **Fig 71**. The pattern is similar to that of Fig 70, but the 3-element binomial array has greater directivity, evidenced by the narrower half-power beamwidth (146° versus 176°). Its gain is 1.0 dB greater.

THE W8JK ARRAY

As pointed out earlier, John Kraus, W8JK, described his bidirectional flat-top W8JK beam antenna in 1940. See **Fig 72**. Two λ/2 elements are spaced λ/8 to λ/4 and driven 180° out-of-phase. The free-space radiation pattern for this antenna, using #12 copper wire, is given in **Fig 73**. The pattern is representative of spacings between λ/8 and λ/4 where the gain varies less than 0.5 dB. The gain over a dipole is about 3.3 dB (5.4 dBi referenced to an isotropic radiator), a worthwhile improvement. The feed-point impedance (including wire resistance) of *each* element is about 11 Ω for λ/8 spacing and 33 Ω for λ/4 spacing. The feed-point impedance at the center connection will depend on the length and Z_0 of the connecting transmission line.

Kraus gave a number of other variations for end-fire arrays, some of which are shown in **Fig 74**. The ones fed at the center (A, C and E) are usually horizontally polarized flat-top beams. The end-fed versions (B, D & F) are usually vertically polarized, where the feed point can be conveniently near ground.

A practical variation of Fig 74B is given in **Fig 75**. In this example, the height is limited to λ/4 so the ends can be bent over as shown, producing a 2-element Bruce array. This reduces the gain somewhat but allows much shorter supports, an important consideration on the low bands. If additional height is available, then you can achieve some additional gain. The upper ends can be bent over to fit the available height. The feed-point impedance will greater than 1 kΩ.

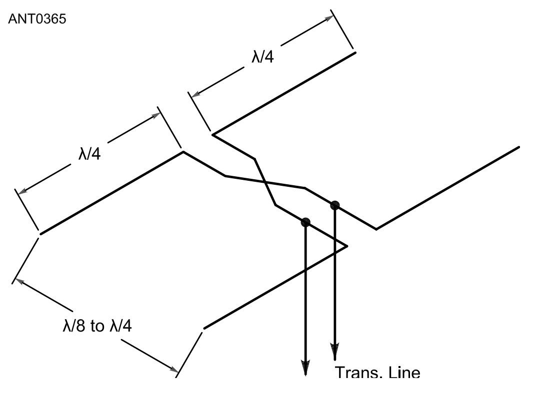

Fig 72—A 2-element W8JK array.

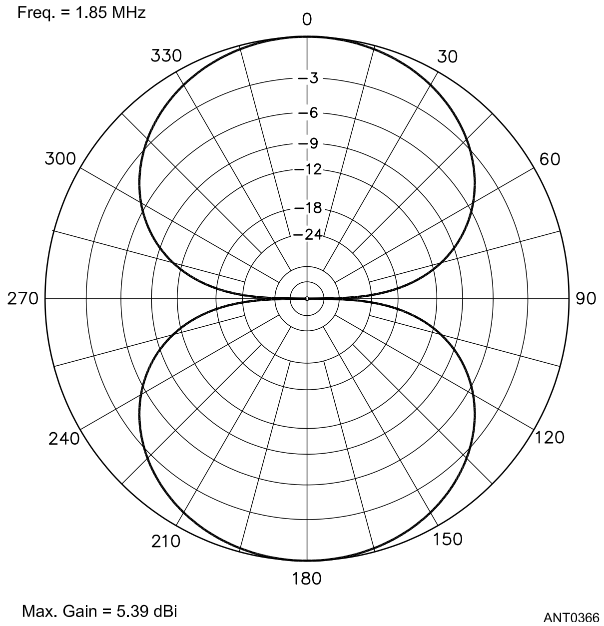

Fig 73—Free-space E-plane pattern for the 2-element W8JK array

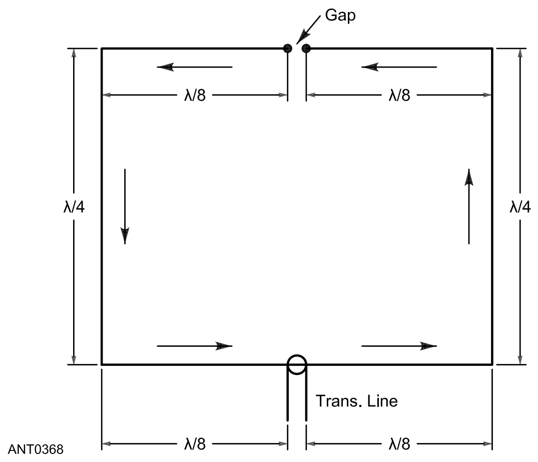

Fig 75—A 2-element end-fire array with reduced height.

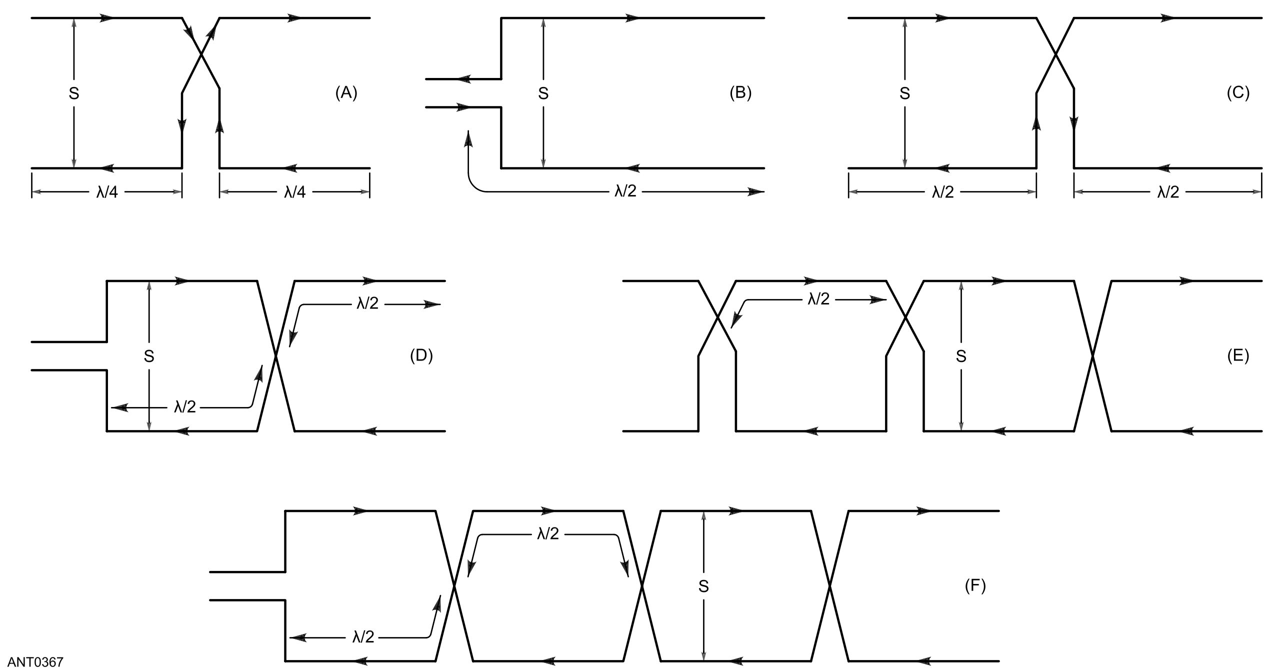

Fig 74—Six other variations of W8JK "flat-top beam" antennas.

FOUR-ELEMENT END-FIRE AND COLLINEAR ARRAYS

The array shown in **Fig 76** combines collinear in-phase elements with parallel out-of-phase elements to give both broadside and end-fire directivity. It is a *two-section W8JK*. The approximate free-space gain using #12 copper wire is 4.9 dBi with λ/8 spacing and 5.4 dBi with λ/4 spacing. Directive patterns are given in **Figs 77** for free space, and in **Fig 78** for heights of 1 λ and λ/2 above flat ground.

The impedance between elements at the point where the phasing line is connected is of the order of several thousand ohms. The SWR with an unmatched line consequently is quite high, and this system should be constructed with open-wire line (500 or 600 Ω) if the line is to be resonant. With λ/4 element spacing the SWR on a 600 Ω line is estimated to be in the vicinity of 3 or 4:1.

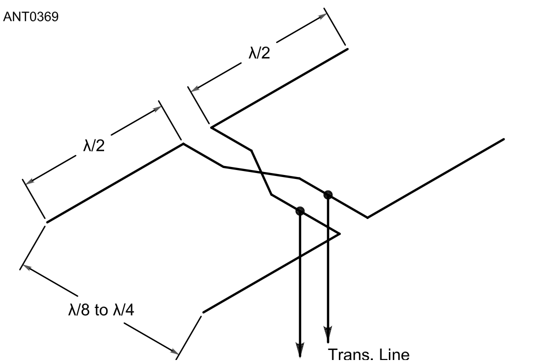

Fig 76—A 4-element array combining collinear broadside elements and parallel end-fire elements, popularly known as a two-section W8JK array.

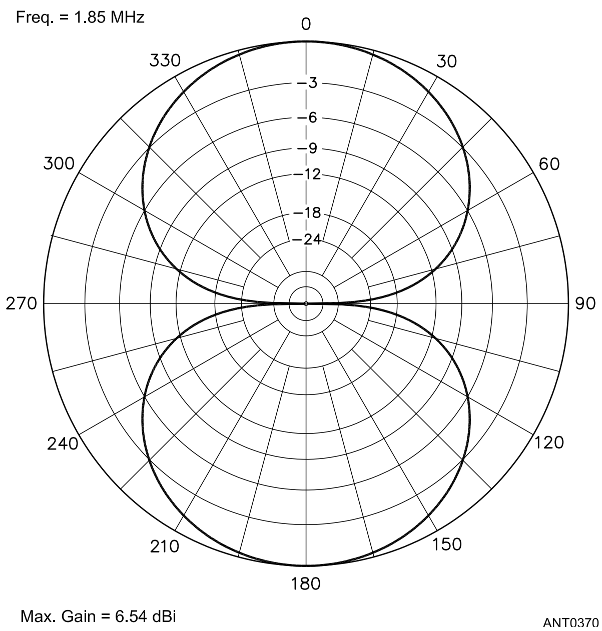

Fig 77—Free-space E-plane pattern for the antenna shown in Fig 76, with λ/8 spacing. The elements are parallel to the 90°-270° line in this diagram. Less than a 1° change in half-power beamwidth results when the spacing is changed from λ/8 to λ/4.

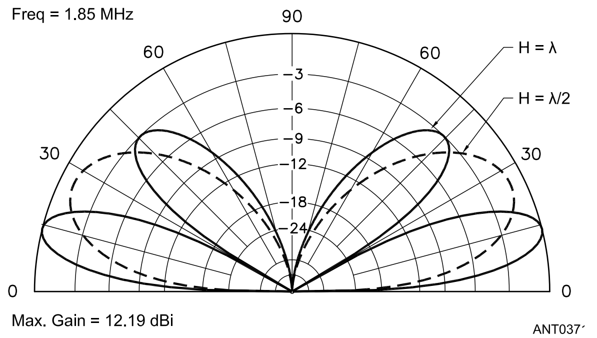

Fig 78—Elevation-plane pattern for the 4-element antenna of Fig 76 when mounted horizontally at two heights over flat ground. Solid line = 1 λ high; dashed line = λ/2 high.

To use a matched line, you could connect a closed stub 3λ/16 long at the transmission-line junction shown in Fig 76. The transmission line itself can then be tapped on this matching section at the point resulting in the lowest line SWR. This point can be determined by trial.

This type of antenna can be operated on two bands having a frequency ratio of 2 to 1, if a resonant feed line is used. For example, if you design for 28 MHz with λ/4 spacing between elements, you can also operate on 14 MHz as a simple 2-element end-fire array having λ/8 spacing.

Combination Driven Arrays

You can readily combine broadside, end-fire and collinear elements to increase gain and directivity, and this is in fact usually done when more than two elements are used in an array. Combinations of this type give more gain, in a given amount of space, than plain arrays of the types just described. Since the combinations that can be worked out are almost endless, this section describes only a few of the simpler types.

The accurate calculation of the power gain of a multi-element array requires a knowledge of the mutual impedances between all elements, as discussed in earlier sections. For approximate purposes it is sufficient to assume that each set (collinear, broadside, end-fire) will have the gains as given earlier, and then simply add up the gains for the combination. This neglects the effects of cross-coupling between sets of elements. However, the array configurations are such that the mutual impedances from cross-coupling should be relatively small, particularly when the spacings are λ/4 or more, so the estimated gain should be reasonably close to the actual gain. Alternatively, an antenna modeling program, such as *EZNEC*, can give good estimates of all parameters for a real-world antenna, providing that you take care to model all applicable parameters.

FOUR-ELEMENT DRIVEN ARRAYS

The array shown in **Fig 79** combines parallel elements with broadside and end-fire directivity. The smallest array (physically)—3λ/8 spacing between broadside and λ/8 spacing between end-fire elements—has an estimated gain of 6.5 dBi and the largest—3λ/4 and λ/4 spacing, respectively—about 8.4 dBi. Typical directive patterns for a λ/4 × λ/2 array are given in **Figs 80** and **81**.

The impedance at the feed point will not be purely resistive unless the element lengths are correct and the phasing lines are exactly λ/2 long. (This requires somewhat less than λ/2 spacing between broadside elements.) In this case the impedance at the junction is estimated to be over 10 kΩ. With other element spacings the impedance at the junction will be reactive as well as resistive, but in any event the SWR will be quite large. An open-wire line can be used as a resonant line, or a matching section may be used for non-resonant operation.

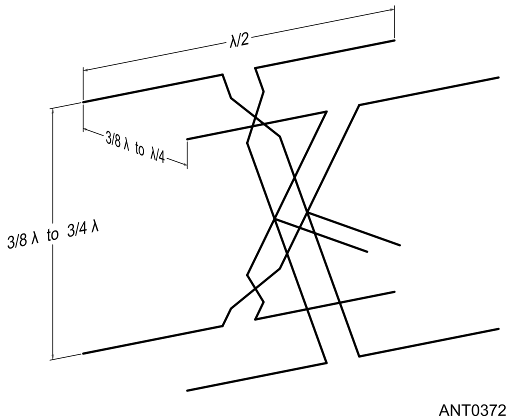

Fig 79—Four-element array combining both broadside and end-fire elements.

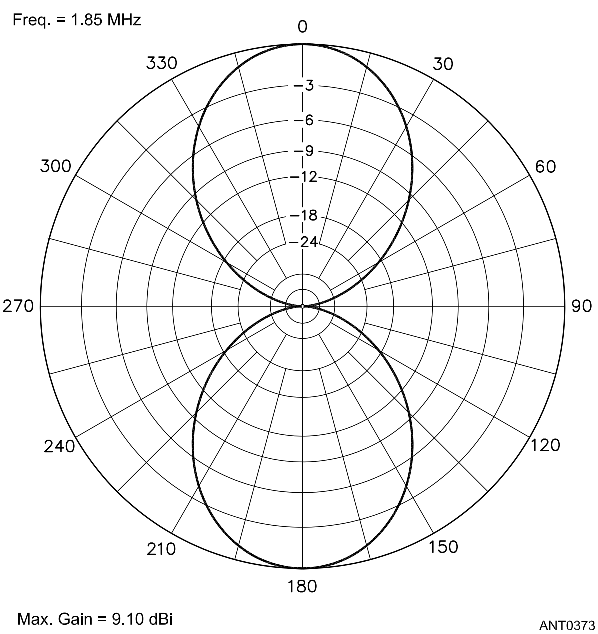

Fig 80—Free-space H-plane pattern of the 4-element antenna shown in Fig 79.

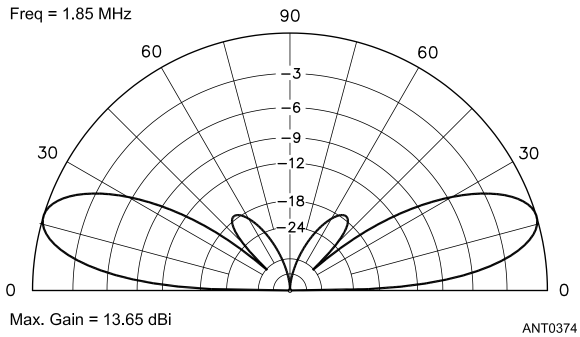

Fig 81—Vertical pattern of the antenna shown in Fig 79 at a mean height of 3λ/4 (lowest elements λ/2 above flat ground) when the antenna is horizontally polarized. For optimum gain and low wave angle the mean height should

EIGHT-ELEMENT DRIVEN ARRAYS

The array shown in **Fig 82** is a combination of collinear and parallel elements in broadside and end-fire directivity. Common practice in a wire antenna is to use λ/2 spacing for the parallel broadside elements and λ/4 spacing for the end-fire elements. This gives a free-space gain of about 9.1 dBi. Directive patterns for an array using these spacings are similar to those of Figs 80 and 81, but are somewhat sharper.

The SWR with this arrangement will be high. Matching stubs are recommended for making the lines non-resonant. Their position and length can be determined as described in Chapter 26.

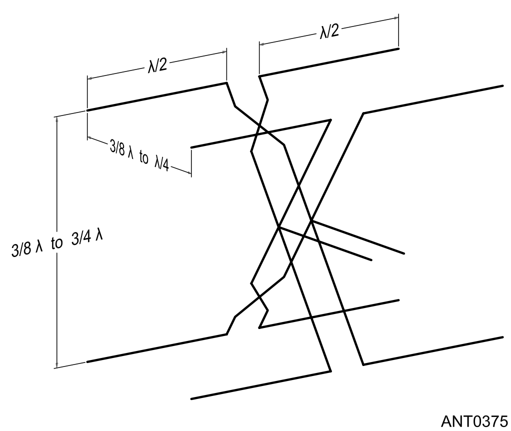

Fig 82—Eight-element driven array combining collinear and parallel elements for broadside and end-fire directivity.

This system can be used on two bands related in frequency by a 2-to-1 ratio, providing it is designed for the higher of the two, with 3λ/4 spacing between the parallel broadside elements and λ/4 spacing between the end-fire elements. On the lower frequency it will then operate as a 4-element antenna of the type shown in Fig 79, with 3λ/8 broadside spacing and λ/8 end-fire spacing. For two-band operation a resonant transmission line must be used.

PHASING ARROWS IN ARRAY ELEMENTS

In the antenna diagrams of preceding sections, the relative direction of current flow in the various antenna elements and connecting lines was shown by arrows. In laying out any antenna system it is necessary to know that the phasing lines are properly connected; otherwise the antenna may have entirely different characteristics than anticipated. The phasing may be checked either on the basis of current direction or polarity of voltages. There are two rules to remember:

1) In every λ/2 section of wire, starting from an open end, the current directions reverse. In terms of voltage, the polarity reverses at each λ/2 point, starting from an open end.
2) Currents in transmission lines always must flow in opposite directions in adjacent wires. In terms of voltage, polarities always must be opposite.

Examples of the use of current direction and voltage polarity are given at A and B, respectively, in **Fig 83**. The λ/2 points in the system are marked by small circles. When current in one section flows toward a circle, the current in the next section must also flow toward it, and vice versa. In the 4-element antenna shown at A, the current in the upper right-hand element cannot flow toward the transmission line because then the current in the right-hand section of the phasing line would have to flow upward and thus would be flowing in the same direction as the current in the left-hand wire. The phasing line would simply act like two wires in parallel in such a case. Of course, all arrows in the drawing could be reversed, and the net effect would be unchanged.

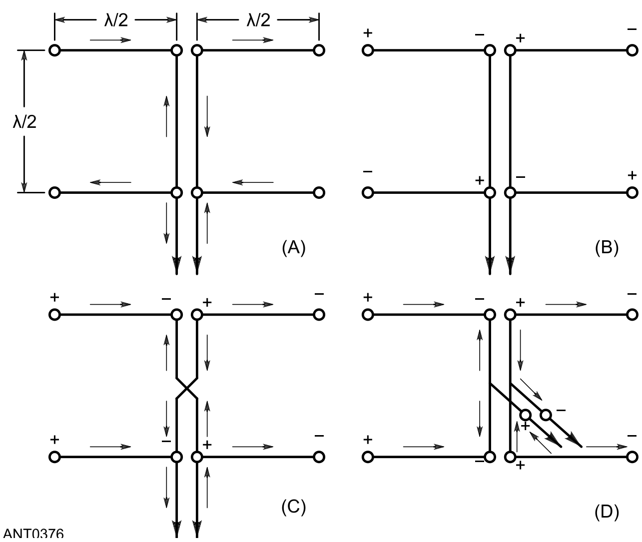

Fig 83—Methods of checking the phase of currents in elements and phasing lines.

C shows the effect of transposing the phasing line. This transposition reverses the direction of current flow in the lower pair of elements, as compared with A, and thus changes the array from a combination collinear and end-fire arrangement into a collinear-broadside array.

The drawing at D shows what happens when the transmission line is connected at the center of a section of phasing line. Viewed from the main transmission line, the two parts of the phasing line are simply in parallel, so the half wavelength is measured from the antenna element along the upper section of phasing line and thence along the transmission line. The distance from the lower elements is measured in the same way. Obviously, the two sections of phasing line should be the same length. If they are not, the current distribution becomes quite complicated; the element currents are neither in-phase nor 180° out-of-phase, and the elements at opposite ends of the lines do not receive the same current. To change the element current phasing at D into the phasing at A, simply transpose the wires in one section of the phasing line. This reverses the direction of current flow in the antenna elements connected to that section of phasing line.

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EZNEC is an antenna-modeling computer program for Microsoft Windows. See **eznec.com** for full information.

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Appendix A—*EZNEC-ARRL* Examples

This appendix contains step-by-step procedures using *EZNEC-ARRL* (included on the *Antenna Book* CD) to illustrate various topics discussed in the main chapter. A standard *EZNEC* program type of v. 4.0 or later may also be used. Different versions, program types and calculating engines may give results that are slightly different from those shown in the examples. However, any differences should be insignificantly small.

EZNEC Example—Mutual Coupling

Illustrates the effect of mutual coupling on feed-point impedance. Open the **ARRL_Cardioid.EZ** file, which is mounted over "perfect" ground. Click the **View Ant** button to see a diagram of the antenna, a 2-element array of vertical elements. Click on the **Wires** line in the main window to open the Wires Window. Click the button at the left of the Wire 2 line, and then press the **Delete** key on your keyboard to delete wire #2. After clicking **Ok**, note that one of the verticals has disappeared from the View Antenna display, leaving a single element. Click **Src Dat** and note that the feed-point impedance of this single vertical is about $37 + j\,1\ \Omega$--it's very nearly resonant.

Next, in the Wires Window, open the **Edit** menu at the top and click **Undo delete wire(s)** to restore the second element. Click **Src Dat** again and notice that the feed-point impedance of wire #1 is now about $21 - j\,19\ \Omega$. The feed-point impedance of the second element, which is identical to the first, is about $52 + j\,21\ \Omega$. This difference, and the change from the self-impedance of $37 + j\,1\ \Omega$, is due to mutual coupling. As you see, it's not at all a minor effect.

As an additional exercise, change the magnitude or phase angle of the source at the base of wire #2 (click **Sources** in the main window), and see how this changes the feed-point impedances of both elements. You should be able to confirm each of the four points enumerated in the **MUTUAL COUPLING** section.

EZNEC Example—Nulls

Illustrates the effect of current magnitude on nulls and gain. Again, open the **ARRL_Cardioid.EZ** file. Click the **FF Plot** button to generate the azimuth pattern of an ideal array. Save the plot for future reference as follows: In the plot window, open the **File** menu and select **Save Trace As**. Enter the name **Cardioid** and click **Save**. Now, in the main window click on the **Sources** line to open the Sources Window. Change the magnitude of source 1 from 1 to 1.1, and of source 2 from 1 to 0.9 and press **Enter** on your keyboard so that *EZNEC-ARRL* will accept the last change.

Click **FF Plot** to generate a pattern with the new currents. In the plot window, open the File menu and select **Add**

Trace. Enter the name **Cardioid** and click Open. You should now see the original plot and new plot overlaid. Notice that the null is much less deep with the altered currents, but the forward patterns are nearly identical. By clicking on the names of the traces, **Primary** and **Cardioid**, you can see in turn the gain and front-to-back ratio of each of the traces. The original, **Cardioid**, has a front-to-back ratio of about 32 dB, while Primary, the new plot, has a ratio of about 22.5 dB. The forward gain, however, differs by only 0.02 dB, a completely insignificant amount.

EZNEC Example—"Phasing-Line" Feed

Illustrates the effect of using a "phasing-line" feed. Open the **ARRL_CardTL.EZ** file. This is a model of an array fed with transmission lines whose lengths were designed using the *Arrayfeed1* program to take into account the actual load impedances of elements in a phased array. This model is mounted over "perfect" ground.

Click the **View Ant** button to show the array. Note that the lengths of the lines from the source (circle) to the elements don't represent the actual physical lengths of the lines. In the main window, click on the **Trans Lines** line to open the Transmission Lines window. In it you can see that the lengths of the feed lines, both of which are connected to the same source, are about 81° and 155°, a difference of 74° rather than 90°.

In the main window, click the **Currents** button and take a look at the current shown for segment 1 of wires 1 and 2. These are the currents at the element feed points. The ratio of the magnitude of currents is 4.577/4.561 = 1.003, and the phase difference is –56.3°–(–147.5°) = 91.2°. (A more accurate determination of feed-line lengths with program *Arrayfeed1* gives lengths of 80.61° and 153.70°, resulting in a current ratio of 1.000 at a phase of 90.02°. But the resulting pattern is very nearly the same.) But let's see what happens when we make the lines exactly 90° different in length.

First, click the **FF Plot** button to generate the azimuth pattern of the original model. Save the plot for future reference as follows: In the plot window, open the **File** menu and select **Save Trace As**. Enter the name **CardTL** and click **Save**. Now in the Transmission Lines Window, change the length of line number 1 from 80.56° to 90°. *Important:* In the line 1 Length box, enter **90d** to make the line 90° long. If you omit the **d**, it will become 90 meters long! Similarly, change the length of line 2 to 180° by entering **180d** in the line 2 Length box, then press **Enter** on your keyboard so that *EZNEC* will accept the last change.

Click **FF Plot** to generate a pattern with the new line lengths. In the plot window, open the **File** menu and select **Add Trace**. Enter the name **CardTL** and click **Open**. You should now see the original plot and new plot overlaid together. Notice that the gain of the modified model is about 1 dB greater than the original, but the front-to-back ratio has deteriorated to about 10 dB.

Experiment with different combinations of line lengths that differ by 90°--for example, 45° and 135° (don't forget the '**d**'!), or change the impedance of one or both lines and you'll see that you can get a wide variety of patterns. None, however, are likely to be as close to the ideal cardioid pattern as the original.