Developments in broadband antennas

A survey is presented for the purpose of providing the nonspecialist with a basic understanding of the remarkable advances that have taken place over the past decade in the field of broadband antennas

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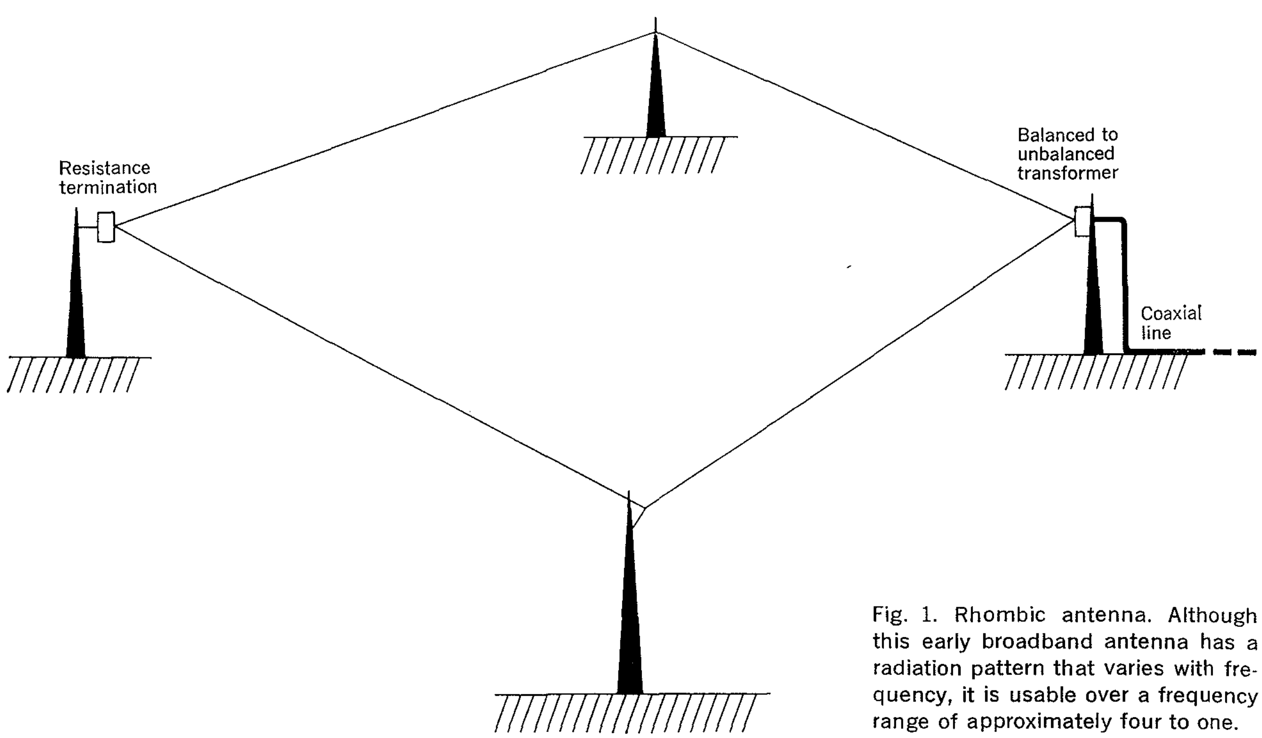


Fig. 1. Rhombic antenna. Although this early broadband antenna has a radiation pattern that varies with frequency, it is usable over a frequency range of approximately four to one.

A

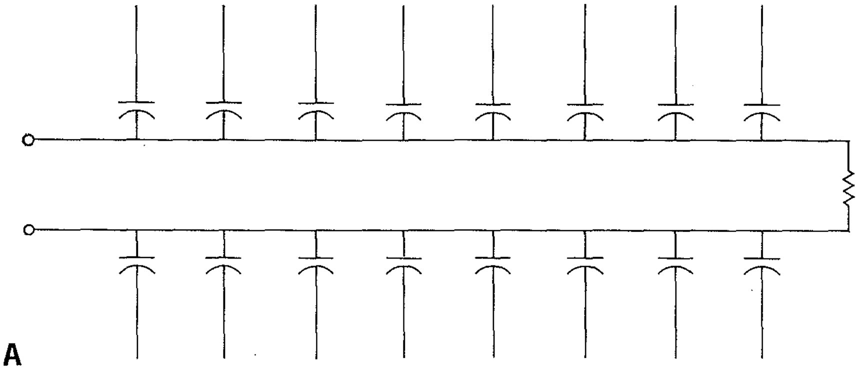
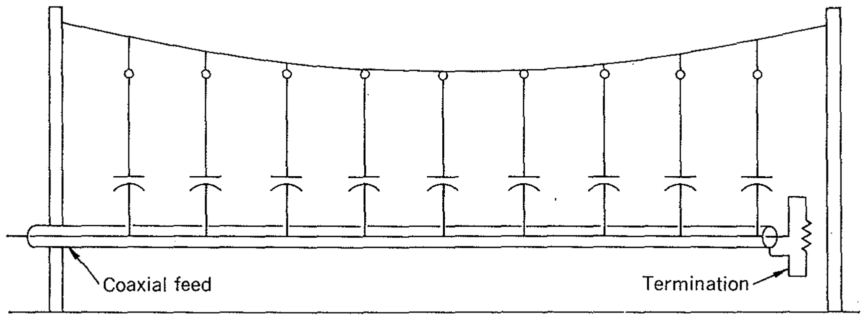


Fig. 2. The fishbone (A) and comb (B) antennas, which have a pattern and impedance that are relatively constant over a two-to-one frequency band.

B



Until a few years ago the ultimate limiting factor on the bandwidth of any communication system using radiated-wave propagation was most frequently the antenna. Since the antenna performs the dual functions of an impedance-matching device and a directional radiator, the characteristics of major importance are its impedance and its directional properties. Depending upon the application, one or the other (or both) of these characteristics may limit the useful bandwidth. The required or desired bandwidth also varies markedly with application; for example, the required bandwidth may vary from a few cycles per second for a single-channel VLF system up to about 6 Mc/s for a single television channel. For multichannel operation, the radio broadcast band covers a three-to-one bandwidth, the VHF television band covers a four-to-one bandwidth, and the high-frequency communication band covers a ten-to-one bandwidth, from 3 to 30 Mc/s. Finally, for countermeasures work it is usual to state that the desired frequency range extends from "dc to light."

Because of the wide variety of operational requirements that exist, there is no unique definition of antenna bandwidth. For our purposes, a broadband antenna will be considered to be one which retains certain desired or specified radiation pattern, polarization, or impedance characteristics over more than an octave (that is, a two-to-one frequency range).

Early broadband antennas

Some of the earliest broadband antennas were long-wire types designed to operate in the high-frequency (short-wave) band or in the low-frequency band. For the most part, they were broadband only in the sense that impedance remained relatively constant over the useful range; in general, no attempt was made to achieve a constant pattern. Among these antennas the well-known rhombic antenna has held a dominant place since the early days of radio. This antenna (Fig. 1) is essentially a resistance-terminated transmission line that has been opened out to form the four sides of a rhombus. Because of the traveling-wave current distribution along the terminated line, the main beam is in the forward direction (toward the termination) at an elevation angle that depends, in a complicated fashion, on the included angle of the rhombus and the lengths of the sides in wavelengths. Fortunately the beam is quite broad in the vertical plane and the angle above ground of the maximum increases as the frequency decreases. This change of angle with frequency is in the correct direction for transmission or reception of ionospherically reflected waves, so a rhombic antenna of fixed dimensions is usable over a wide frequency range (of the order of four to one) in the short-wave band.

The wave antenna, consisting of a long, elevated wire parallel to the ground and resistance-terminated at both ends, is another traveling-wave-type antenna. In contrast with most antennas, which operate best over a highly conducting ground, the wave antenna depends for its operation upon the finite conductivity of the earth beneath it. An incident radio wave traveling along the surface of a finitely conducting earth has a forward tilt and a horizontal component of electric field intensity. It is this horizontal component of electric field, produced by the finite earth conductivity, that induces a traveling wave of voltage in the horizontal wire and the resulting antenna action. Because the antenna has an impedance that is nearly independent of frequency, it is known as an aperiodic antenna. However, the radiation pattern does vary with the length of the wire in wavelengths, and hence with frequency. Wave antennas are used for long-wave or low-frequency reception.

The fishbone receiving antenna consists of a long resistance-terminated transmission line loosely coupled by capacitors to an array of closely spaced (less than $\lambda/4$), untuned, horizontal dipoles; see Fig. 2(A). For vertically polarized signals, one half of a fishbone antenna is erected vertically and fed against ground to form a comb antenna; see Fig. 2(B). Tapering the coupling capacitors to larger values towards the termination equalizes the antenna currents and reduces resonance effects. Because the capacitive coupling of the elements to the transmission line is tighter at the higher frequencies, fewer of the elements are strongly excited. Hence, the effective length of the array varies inversely with frequency in such a manner as to maintain a fairly constant pattern, gain, and impedance, over the useful bandwidth of more than two to one. The fishbone and comb antennas have been described chiefly for comparison with the log-periodic dipole and log-periodic monopole types to be described later.

In contrast to the terminated wire and loaded transmission line types just described, there is a class of antennas that owes its broadband properties to broad, specially shaped surfaces. It was recognized quite early that a fat dipole had a much lower antiresonant (full-wavelength) impedance than a thin one, and that in general, fat antennas had smaller impedance variations than thin ones. The importance of broad surfaces was emphasized by Schelkunoff in the treatment of the biconical antenna, and many broad-surfaced specially shaped antennas found early application in television transmitting antennas and countermeasures antennas. A very successful broadband antenna was the discone[1] (a cone fed against a disk), which maintained good impedance and pattern characteristics over a four-to-one bandwidth (Fig. 3). For countermeasures work many antennas having surfaces of various shapes were developed,[2] some of which had remarkably wide impedance bandwidths and usable pattern bandwidths of the order of five to one. It must be admitted, however, that most of these early designs were arrived at by an intuitive or cut-and-try approach.

Another group of antennas, some of which display fairly wide bandwidths, consists of various helical and spiral shapes. When the circumference of a helical antenna is of the order of a free-space wavelength, the antenna radiates in the axial mode—that is, with the maximum radiation along the axis of the helix. In this mode, the helical antenna has desirable impedance, pattern, and circular polarization properties over nearly an octave.[3] By expanding the diameter of the helix along its length to form a conical monofilar helix fed from the base end, Springer[4] showed that the bandwidth could be increased. His observation that there appeared to be an effective aperture that moved toward the smaller end of the cone as the operating wavelength decreased was perhaps the first indication of things to come. Later, Chatterjee[5,6] also considered monofilar helical antennas formed on a conical surface and fed against a ground plane. He demonstrated that they could be excited from either end, and obtained usable bandwidths of approximately four to one. At about the same time, Turner[7] proposed a balanced antenna constructed in the form of an Archimedes spiral. This planar antenna, constructed with narrow constant-width arms and radiating a broad lobe on each side of the structure, gave promise of being usable over the then remarkable bandwidths of between seven and eight to one.

Frequency-independent antennas

In 1954, Rumsey[8] put forth the idea that a structure entirely definable by angles, without any characteristic length dimension, should have properties that are independent of the frequency of operation. However, all such angle structures extend to infinity, so the key question was which of such structures retained these frequency-independent characteristics when truncated to a finite length. It should be noted that the well-known biconical structure is an angle structure that is *not* frequency independent when it is truncated to form a practical antenna. Both impedance and pattern vary with frequency for any finite length.

Rumsey proposed that an equiangular spiral structure, which satisfies the angle requirement, might have the desired properties, and Dyson[9–11] undertook a comprehensive experimental study of an antenna based on the equiangular spiral geometry shown in Fig. 4. The equiangular or logarithmic spiral* is defined by

$$\rho = e^{a(\phi - \delta)} \quad \text{or} \quad \phi - \delta = \frac{1}{a} \ln \rho$$

where ρ and ϕ are conventional polar coordinates, and a and δ are constants. In Fig. 4 the edges of the metallic arms are defined by

$$\rho_1 = ke^{a\phi} \quad \text{and} \quad \rho_2 = ke^{a(\phi - \delta)}$$

for one arm, and by

$$\rho_3 = ke^{a(\phi - \pi)} \quad \text{and} \quad \rho_4 = ke^{a(\phi - \pi - \delta)}$$

for the other arm, where the constants a, k, and δ determine the rate of spiral, size of the terminal region, and arm width, respectively. With this particular spiral the angle between the radius vector and the spiral remains the same for all points on the curve—hence the term "equiangular spiral." Experimental investigation established that this particular geometry did indeed retain its frequency-independent properties after truncation, and this design was the basis for a large class of successful frequency-independent antennas.

When this angular structure is excited in a balanced manner at the origin, the current flows outward with small attenuation along the spiral arms until a region of given size in wavelengths is reached. In this region (the active or radiating region) essentially all of the incident energy transmitted along the spiral arms is radiated, and somewhat beyond this region the presence or absence of the arms is of no consequence. Because the radiating region is of constant size in wavelengths, it moves toward the origin as the wavelength of operation decreases. The size of effective radiating aperture thus automatically adjusts or scales with frequency of operation in such a manner that the antenna behaves the

* The logarithmic spiral was first discussed by Descartes (1638) and later (1691–1693) studied by Jacques Bernouilli, who gave it its name. Bernouilli was so delighted by the property of the spiral reproducing itself under various transformations that he requested that the spiral be engraved on his tomb with the inscription "Eadem Mutata Resurgo."

same at all frequencies. Because of the spiraling of the arms, this scaling is accompanied by a rotation of the radiated field about the axis of the antenna.

It is now known that this automatic scaling of the radiating aperture is a condition for operation in a frequency-independent manner. It is interesting to note that Springer observed this phenomenon on the expanding helix, but unfortunately the methods of construction and excitation limited the bandwidth obtainable to something over an octave, so the importance of scaling of effective aperture with frequency was not fully recognized. Chatterjee's measurements also show evidence of scaling with frequency in the near-field amplitude plots from which he calculated radiation patterns; but again, possibly because of the physical configuration and method of feed, the full significance of this scaling does not appear to have been appreciated. In a similar manner, the radiating aperture of the Archimedes spiral antenna tends to scale with frequency; however, because the width and spacing of the spiral arms in the radiating region are not constant in wavelengths, as frequency is varied, the antenna characteristics change (albeit slowly) with frequency.

At this point it is necessary to define the term "frequency independent" when it is used with a practical finite-sized structure. If the antenna is excited by a voltage applied between the two arms at the origin, it has an impedance and radiation pattern that are essentially constant* (that is, independent of frequency) for all frequencies above that for which the outer diameter of the truncated structure is approximately half a wavelength up to the frequency at which the diameter of the feed region (as determined by the transmission line feed) is comparable with a half wavelength. Since these two dimensions can be specified independently, the design bandwidth can be made arbitrarily large; actually it is limited only by practical considerations of construction—that is, how large the outer diameter is made and how finely the geometry at the feed region can be modeled.

The equiangular spiral antenna, which is bidirectional, radiates a very broad, circularly polarized beam on both sides of its surface. This bidirectional characteristic severely restricts its utility in practice, but a modified version, to be described later, provides a highly practical, extremely broadband antenna.

Log-periodic antennas

In 1955, working with Rumsey on broadband antenna development, DuHamel[12] proposed that it should be possible to force radiation from otherwise "angle structures" by the use of appropriately located discontinuities. One of the first geometries chosen to investigate the validity of this concept was that shown in Fig. 5. Here two wedge-shaped metallic angle structures have teeth cut into them along circular arcs. The radii of the arcs which define the location of successive teeth are chosen to have a constant ratio $\tau = R_{n+1}/R_n$. This same ratio τ defines the lengths and the widths of suc-

* The pattern actually rotates with frequency about an axis perpendicular to the plane of the spiral. If the pattern-measuring coordinate system is allowed to rotate at the same rate, the measured pattern remains constant; otherwise there will be a (generally small) periodic variation of magnitude proportional to the rotational asymmetry of the pattern.

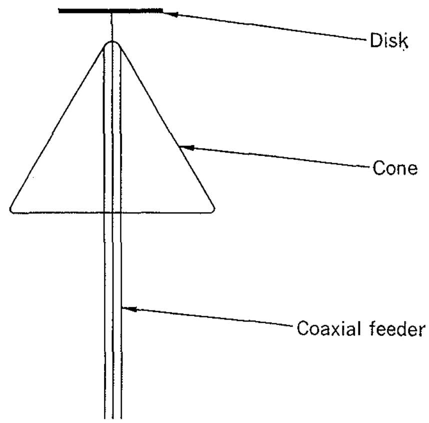


Fig. 3. The discone, a successful early broadband antenna having a useful bandwidth of approximately four to one.

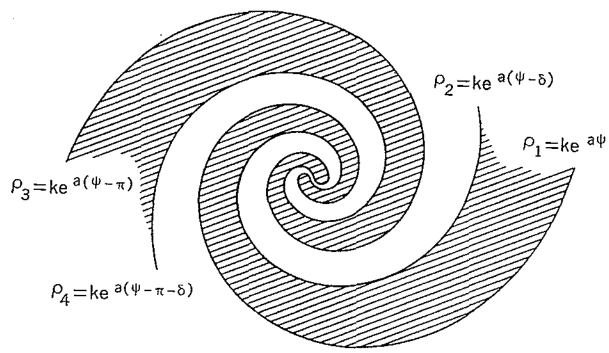
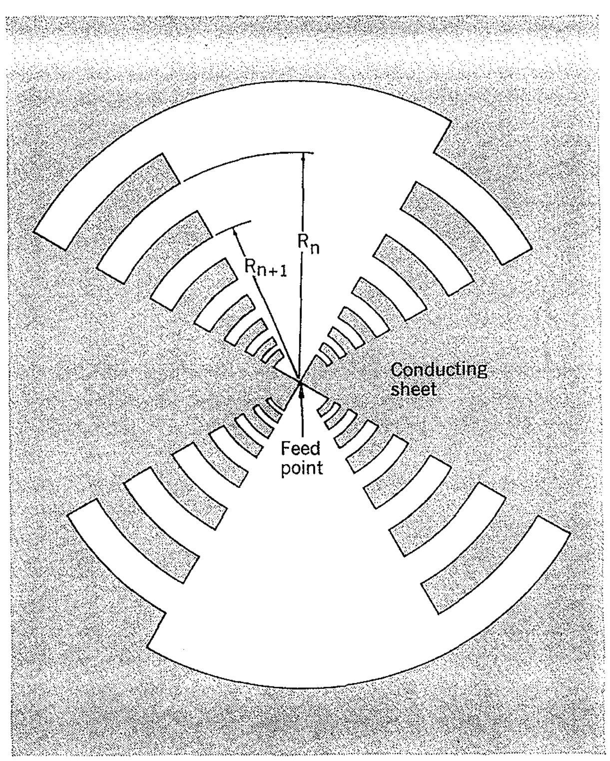


Fig. 4. Sketch showing the geometry of the equiangular (or logarithmic) spiral antenna with equations of the edges.

Fig. 5. A sheet-metal log-periodic antenna.



cessive teeth. From the principle of modeling it is evident for this structure, extending from zero to infinity and energized at the vertex, that whatever properties it may have at a frequency f will be repeated at all frequencies given by $\tau^n f$, where n is an integer. When plotted on a logarithmic scale, these frequencies are equally spaced with a period equal to the logarithm of τ; hence the name "log-periodic" structure. Log-periodicity guarantees only periodically repeating radiation pattern and impedance. However, for certain types of such structures and for values of τ not too far from unity, variation of characteristics over a period can be quite small, and an essentially frequency-independent structure results. It is important to note that only a relatively few of the nearly infinite variety of log-periodic structures will make successful broadband antennas in the sense that the impedance and pattern characteristics will remain constant when the structure is truncated to a finite length. It happens that the geometry of Fig. 5 did result in a successful log-periodic antenna.

The antenna of Fig. 5 was designed to have one other rather special property; namely, that the metal cut away from the plane sheet to form the antenna arms has identical shape with the metal that remains. In other words, the complementary slot antenna has the same size and shape as the metallic dipole antenna. Now by an extension of Babinet's principle it is known that complementary-dipole and slot antennas have impedances Z_d and Z_s, respectively, related by $Z_d Z_s = (60\pi)^2$. Because the slot antenna and dipole antenna are the *same* (for the geometries chosen) it follows that $Z_d = Z_s = 60\pi \approx 189$ ohms, a result that is independent of frequency. Hence this particular geometry assured constant impedance, although not constant radiation pattern, independently of the other consideration of log-periodic geometry. In view of this use of Babinet's principle in the design of these planar structures, the next step to be taken was a bigger one than might at first appear.

Unidirectional frequency-independent and log-periodic antennas. Both the equiangular spiral antenna (Fig. 4) and the log-periodic antenna (Fig. 5) radiate equally on both sides of the plane of the antennas, a result that severely limits their usefulness. A major step forward was made in extending the range of practical application when Isbell[13] bent the two arms of the planar log-periodic structure toward each other (out of the plane) to form the nonplanar V-shaped antenna of Fig. 6. Two rather surprising results were observed. As the angle between the two arms of the antenna was decreased from 180° the radiation pattern changed from bidirectional to undirectional, with the major radiation off the apex of the antenna—that is, in the backward direction. Moreover, although one of the necessary conditions for Babinet's principle (that of a plane surface) was now violated, the impedance continued to remain nearly constant with frequency, but at a different value, which depended upon the angle between the arms. This nonplanar version of the log-periodic structure, radiating a plane-polarized unidirectional beam, greatly increased the utility of the log-periodic structures.

The frequency-independent logarithmic spiral structure also found wider use when Dyson developed a unidirectional version by wrapping the balanced spiral arms on the surface of a cone, as shown in the antenna of Fig.

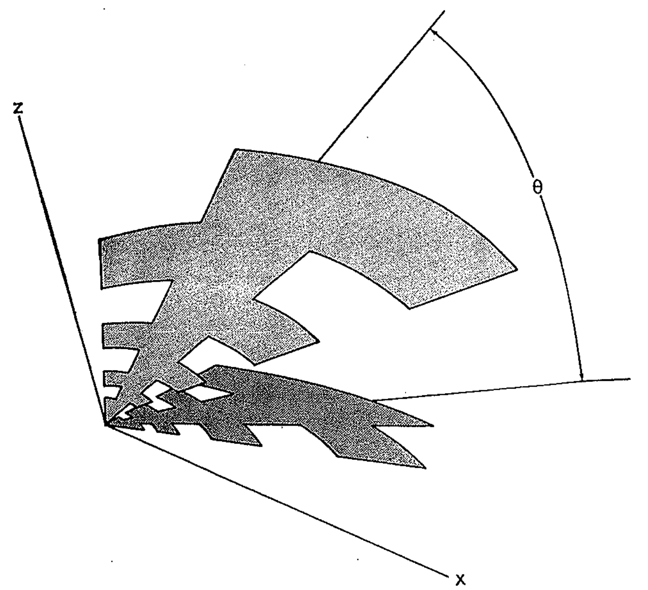
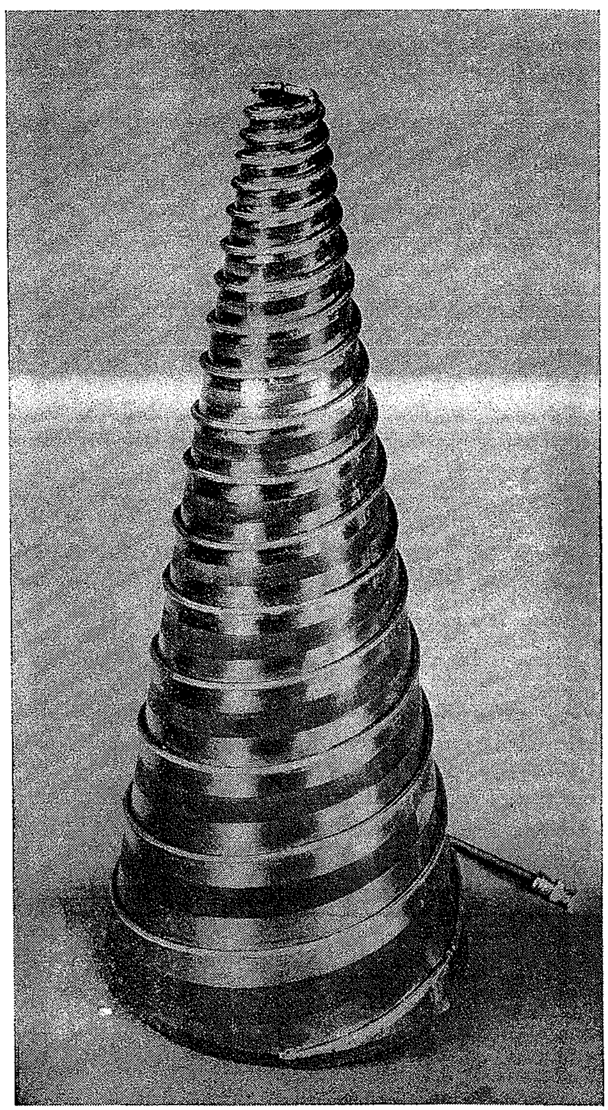


Fig. 6. Nonplanar, unidirectional log-periodic antenna.

Fig. 7. Unidirectional conical spiral antenna.



7. For appropriately chosen rates of spiral this modified version continued to yield essentially frequency-independent performance. For cone angles of less than about 45° the pattern became unidirectional with a broad-lobed beam, again in the backward direction off the apex of the cone.

The conical equiangular spiral antenna is a balanced structure, which may be fed (at the apex) by means of a balanced transmission carried up inside and along the axis of the cone. Alternatively, it may be fed as illustrated in Fig. 7 by a coaxial cable carried along and soldered in contact with one of the arms. Because the amplitude of antenna current on the arms, and also on the outside of the coaxial cable, falls off quite rapidly with distance from the apex, the ends of the arms where the cable enters is essentially a field-free region. This type of feed automatically provides a frequency-independent balun (balanced converter), permitting the balanced antenna to be fed by means of an unbalanced coaxial line. To maintain physical symmetry a dummy cable is usually soldered to the other arm. Conical equiangular or log-spiral antennas have been constructed to operate over bandwidths of higher than 40 to 1. The bandwidth obtained is at the discretion of the designer. The upper usable frequency is determined by the truncated region at the apex, which must remain small in terms of wavelengths, and the lowest usable frequency is set by the base diameter of the cone, which must be at least 3/8 wavelength at the lowest frequency of operation for spirals that are wrapped fairly tightly.

A further modification of the conical equiangular spiral results in a very practical, easily constructed antenna. If the width of the expanding arms is narrowed and they are allowed to degenerate to constant-width structures, the cables alone can form the arms. For fairly tightly spiraled antennas there is little change in the characteristics from those of an antenna with narrow expanding arms.

Other types of log-periodic antennas. The practical value of the log-periodic approach was enhanced even further when DuHamel[14] and co-workers demonstrated that successful log-periodic antennas could be made with wire structures as well as sheet structures. This development extended the range of application down from microwaves through the high-frequency band. A typical wire version of a log-periodic antenna is shown in Fig. 8. It was also demonstrated that for higher gain a frequency-independent array of log-periodic antennas could be constructed by arranging the antennas like the spokes of a wheel with the origins of the individual antennas at the hub.

Still another application of the log-periodic principle is the log-periodic dipole array[15] of Fig. 9. As with all log-periodic geometries, all dimensions are increased by a constant ratio in moving outward from the origin. Thus the lengths and spacings of adjacent elements must be related by a constant scale factor τ, as follows:

$$\frac{l_n}{l_{n-1}} = \frac{d_n}{d_{n-1}} = \tau$$

Although at first glance this antenna might appear similar to the early fishbone antenna with $\tau = 1$, there are several essential differences. For successful operation, the log-periodic dipole array must be fed with a transposition of the transmission line between adjacent dipole elements. The antenna is then caused to radiate in the backfire direction (that is, toward the source), a condition which appears to be necessary for successful unidirectional frequency-independent or log-periodic operation. The

Fig. 8. Log-periodic wire antenna for frequencies of 11 to 60 Mc/s. (Photo courtesy Collins Radio Company.)

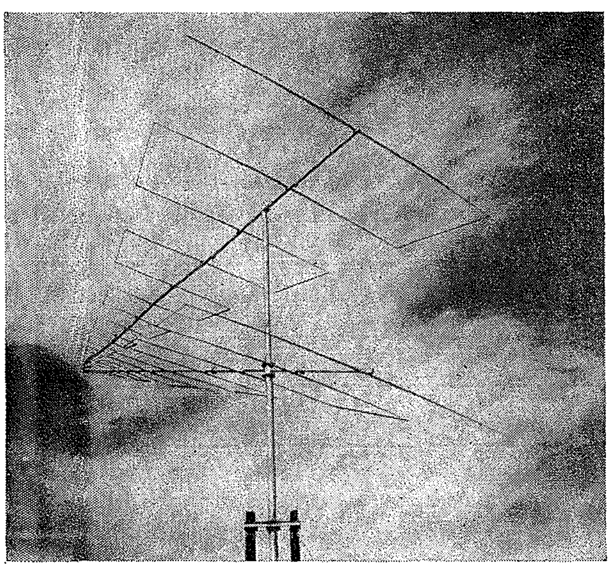
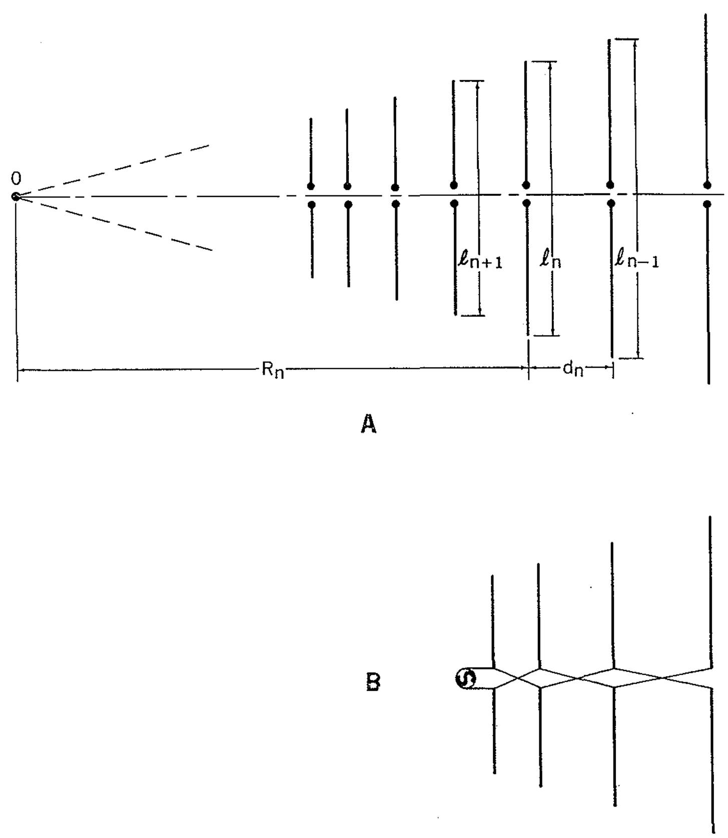


Fig. 9. Log-periodic dipole antenna array. A—Lengths and spacings of elements. B—Method of feeding.



active portion of the array from which most radiation occurs is centered around those elements near resonance (for which l_n is somewhat less than $\lambda/2$). As the frequency is changed the active region moves back or forth along the array. Because practically all of the input power is absorbed in and radiated by the active portion, the larger elements to the right of the active region are not excited. Moreover, because the beam is directed toward the feed point at the left, these larger elements are in an essentially field-free region, and so do not adversely affect the operation. The shorter elements to the left of the active region are in the beam but, because of their short lengths, close spacings, and alternate phasings, have small influence on the pattern.

Basic principles of operation of log-periodic and frequency-independent antennas. Of the almost unlimited variety of log-periodic structures that can be devised, only a small fraction will produce successful antennas when truncated. It is interesting to search out the essential requirements for successful design. The operation of the log-periodic dipole array of Fig. 9, being simple and easily understood, will be analyzed in some detail. From the understanding so gained it should be possible to extend the analysis to less familiar geometries, and then to frequency-independent antennas in general.

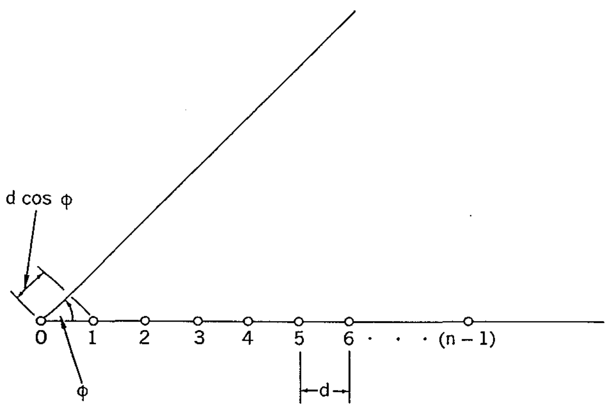


Fig. 10. Array of equispaced isotropic radiators.

At this point it will be advantageous to recall some of the basic notions of antenna array theory. Consider an n-element array of equispaced isotropic radiators (Fig. 10) having equal current amplitudes and a spacing d less than one-half wavelength. (An isotropic radiator is one that radiates uniformly in all directions; a simple dipole antenna is an isotropic radiator in the H plane perpendicular to its axis.) At a distant point the electric fields from these radiators will add with a phase angle between them which is dependent upon the relative phasings of the radiator currents and the relative phase delays produced by the difference in path lengths to the distant point. For the array shown the phase difference due to path length difference between adjacent elements is $(2\pi/\lambda)d \cos \phi$ radians. If the elements of the array are fed with a progressive phasing of currents equal to α, where α represents the angle by which the current in a given element *leads* the current in the preceding element, then at the distant receiving point the phase difference of the fields produced by adjacent elements will be

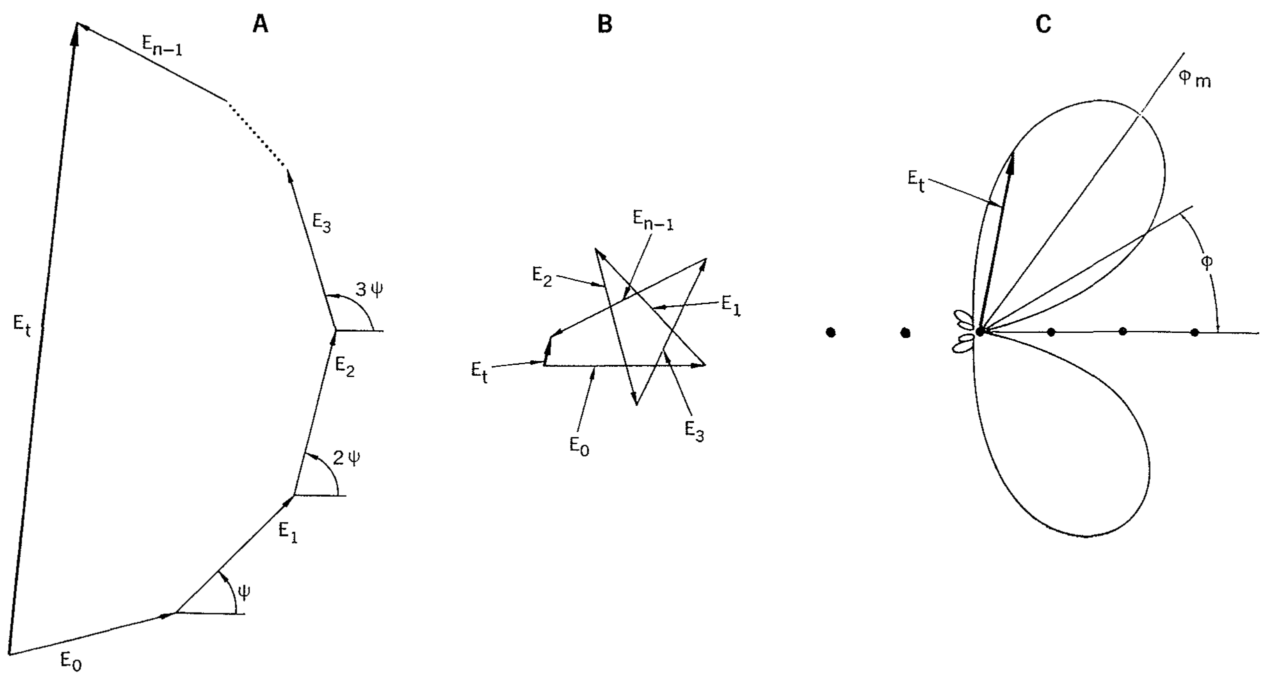
$$\psi = \alpha + \frac{2\pi}{\lambda} d \cos \phi = \alpha + kd \cos \phi \qquad (1)$$

where $k = 2\pi/\lambda$ is the free-space phase-shift constant. The total electric field at any distant point will be given by the phasor sum

$$E_t = E_0[1 + e^{j\psi} + e^{j2\psi} + \ldots + e^{j(n-1)\psi}] \qquad (2)$$

where E_0 is the field intensity at the reception point produced by current I_0. E_t can be obtained graphically

Fig. 11. A and B—Graphical construction for E_t. C—Calculated radiation pattern, E_t vs. ϕ.



from the construction of Fig. 11(A). Using the particular value of α, and computing ψ from (1) for various values of ϕ, the construction of Fig. 11(A) can be used to determine a radiation pattern of the array; see Fig. 11(C). It is evident that the total field intensity will be maximum when $\psi = 0$, so that all fields add in phase.

Therefore, for a maximum, $\psi = \alpha + kd \cos \phi = 0$. The angle ϕ_m for maximum radiation is given by

$$\cos \phi_m = -\frac{\alpha}{kd} \quad \text{or} \quad \phi_m = \cos^{-1} \frac{-\alpha}{kd} \tag{3}$$

If the elements are fed in phase, $\alpha = 0$, and $\phi_m = 90°$, so the maximum radiation is broadside. If successive elements are fed with a lagging phase of value, $\alpha = -kd$, than $\phi_m = 0$, so the maximum radiation is endfire in the forward direction. If successive elements are fed with a leading phase of value, $\alpha = +kd$, than ϕ_m will equal 180°, and the maximum radiation will be endfire in the backward direction. For values of α between $-kd$ and $+kd$, the angle of maximum radiation is at an angle between 0 and 180° as given by Eq. (3). By symmetry about the axis of the array, there is another maximum at an angle between 0 and $-180°$, which is also given by (3). When $|\alpha| > kd$, Eq. (3) cannot be satisfied for any real value of ϕ; that is, there is no value of ϕ in the "visible" range between 0 and 180° (hence, also between 180° and 360°) that will produce a maximum—in the sense that all the radiations add in phase. However, if $|\alpha|$ is only slightly greater than kd, so that ψ is not much larger than zero, the total field can still be quite strong in the forward direction ($\phi = 0$) for negative α, or in the backfire direction ($\phi = 180°$) for positive α. This case is illustrated by the sketch of Fig. 11(A). On the other hand, if $|\alpha|$ is considerably greater than kd (that is, the phase shift between elements is large), the phase diagram might be as illustrated in Fig. 11(B), with a resulting small total E_t for all values of ϕ.

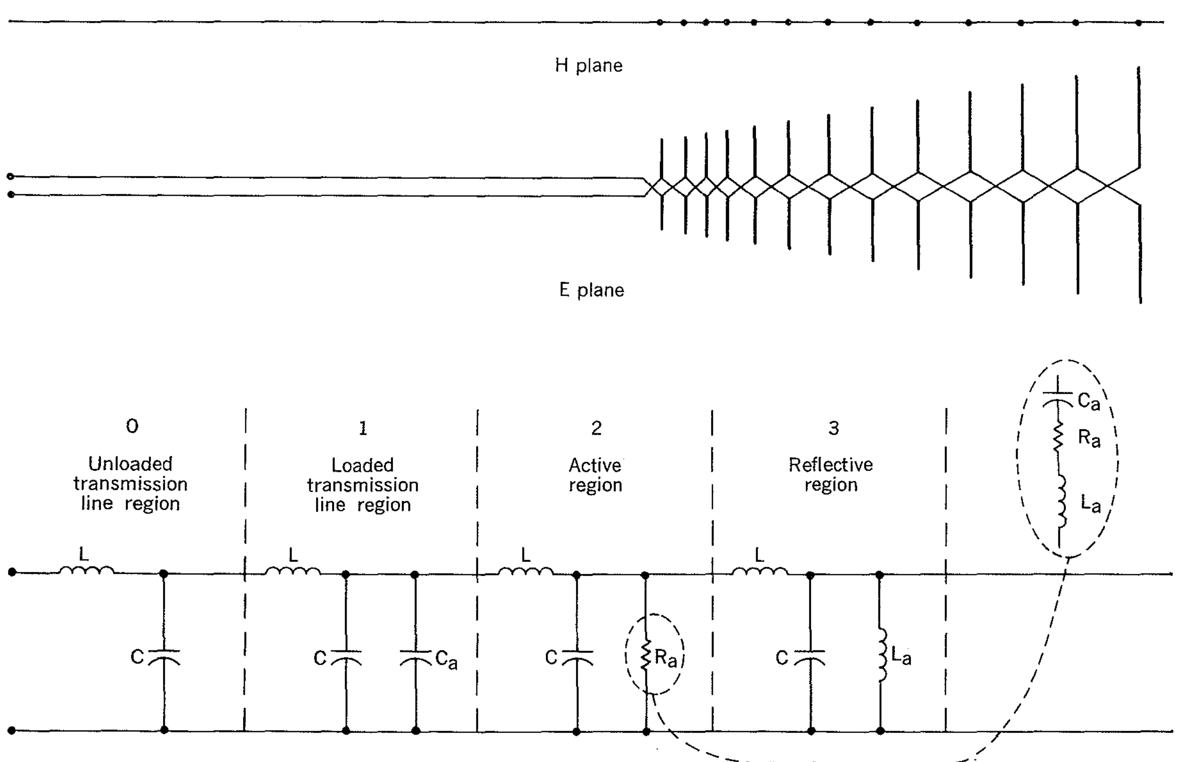
For these cases of large phase shift, as shown in Fig. 11(B), there is no major lobe anywhere, and the array radiates only feebly, scattering its small radiated energy in various directions.

The elementary notions just discussed can be applied with some slight modification to an analysis of the log-periodic dipole array sketched in Fig. 12. For this purpose, it is helpful to consider separately three main regions of the array.

1. *Transmission-line region.* The antenna elements in the transmission-line region are short compared with the resonant length (that is, $l \ll \lambda/2$), so the element presents a relatively high capacitive impedance. The element current is small and leads the base voltage supplied by the transmission line by approximately 90°. The element spacing is small in wavelengths and the phase reversal introduced by transposition of the transmission line means that adjacent elements are nearly 180° out of phase. More precisely, each element current leads the preceding element current approximately by $\alpha = \pi - \beta d$, where d is the element separation and $\beta = 2\pi/\lambda = \omega/v$ is the phase-shift constant along the line. In general β, λ, and v will differ from their free-space values owing to the loading effect of the elements on the transmission lines. Because of the phasing and close spacing of the elements, radiation from this region will be very small and in the backfire direction.

2. *Active region.* In the active region the element lengths approach the resonant length (l slightly less than

Fig. 12. Transmission-line representation of log-periodic dipole array.



λ/2), so the element impedance has an appreciable resistive component. The element current is large and more nearly in phase with the base voltage; the current is slightly leading just below resonance and slightly lagging just above resonance. The element spacing is now sufficiently large to allow the phase of current in a given element to lead that in the preceding element by an angle $\alpha = \pi - \beta d$, which may approximate $\pi/2$ radians. This combination of conditions will produce a strong radiation in the backfire direction.

3. *Reflection region.* The element lengths in the reflection region are greater than the resonant length ($l \geq \lambda/2$), so the element impedance becomes inductive and the element current lags the base voltage. The base voltage provided by the transmission line is now quite small, because in a properly designed array nearly all of the energy transmitted down the line has been abstracted and radiated by the active region. The element spacing may now be larger than λ/4. However, as will be shown later, the phase shift per unit length along the line in this region is small, so the resulting phasing between elements (including the phase reversal introduced by the transposition) is such that any small amount of radiation is still in the backfire direction. In addition, it will be demonstrated later that the characteristic impedance of the transmission-line becomes reactive in this region. Thus, any small amount of incident energy transmitted through the active region is not accepted in the reflection region but is reflected back toward the source.

The array as a loaded transmission line. Some of the remarkable properties of log-periodic and frequency-independent antennas are attributable to the propagation characteristics of the equivalent loaded transmission line that conveys energy from the source to the radiating portion of the antenna. These effects are particularly easy to see in the case of the log-periodic dipole array, shown in Fig. 12. On the feed line to the antenna, region 0, the series inductance and shunt capacitance per unit length are shown as L and C, respectively. In the transmission region of the antenna, region 1, the transmission line is loaded by a capacitance per unit length C_a that represents the loading effect of the short dipoles, which have a capacitance reactance. It is noted that to the first approximation C_a is nearly constant throughout this region because at the beginning of the region the capacitance per element is small, but the elements are closely spaced, whereas near the end of the region the capacitance per element is larger, but so is the spacing. The effect of the augmented shunt capacitance of the line ($C + C_a$) is to increase the phase delay per unit length, and since $\beta = 2\pi/\lambda = \omega/v$, this means a decrease of wavelength λ and a decrease of phase velocity v along the line below the free-space values. This is said to be a "slow wave" region of the transmission line. Note, however, that because of the transposition of the feed line between elements, successive elements are fed with a leading phase shift of $\pi - \alpha$ per section. This rapid phase shift in the reverse direction corresponds to a slow wave in the backward direction along the antenna elements.

In region 2, the element lengths approach the resonant length and the transmission line loading becomes resistive, designated by the shunt resistance R_a in series with the antenna capacitance C_a and antenna inductance L_a. The phase shift per unit length, the wavelength, and the phase velocity all approach their free-space values. Because of the transposition between elements, and accounting for the fact that the element current leads the base voltage by lesser amounts in successive elements as the resonant length is approached, it turns out that phasing of currents in the elements corresponds to a backward traveling wave having a velocity v somewhat less than c, the velocity of light.

In region 3 the element lengths become longer than the resonant length, the antenna inductive reactance pre-

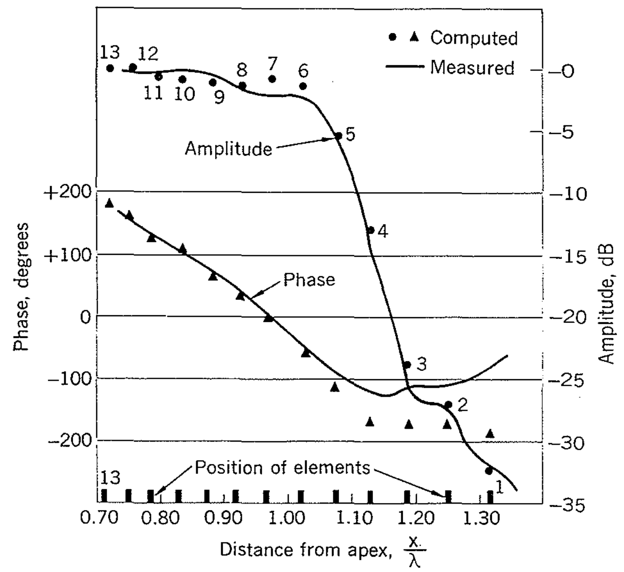
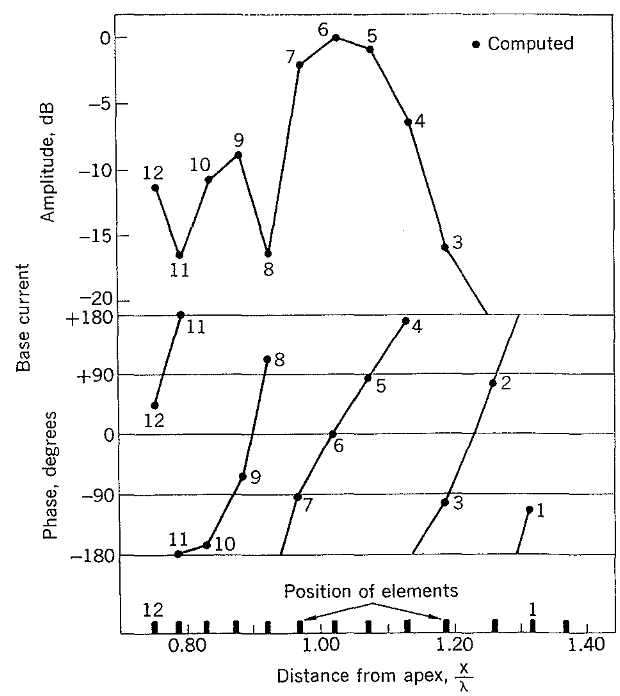


Fig. 13. Base voltages along a typical 13-element log-periodic dipole array at a frequency for which element 4 is half of a wavelength.

Fig. 14. Element currents corresponding to Fig. 13.



dominates, and the loading effect on the line is represented by the shunt inductance L_a. If the parallel combination of L_a and C is inductive, we have the equivalent of the attenuation region of a filter. The phase shift per unit length is then zero (for the lossless case) and the phase velocity is infinite; that is, there is no wave motion. The incident energy propagating down the line is no longer accepted but is reflected back toward the source. These results are strictly true only in the case of a lossless filter, but they form the first approximation in the case of a lossy filter.

The general features outlined in the foregoing discussion will be illustrated for a particular log-periodic dipole array, which has been analyzed in considerable detail.[16] Fig. 13 shows the amplitude and phase of the transmission line voltage along a particular 13-element log-periodic dipole array. Distance is shown measured from the apex of the array, and the elements are numbered starting with the largest element as number 1. This set of data is for a frequency f for which element number 4 is $\lambda/2$ long. Several interesting aspects of the data are immediately apparent: In the transmission region (elements 13 to 7), the amplitude of voltage along the line is approximately constant and the phase shift between element positions increases gradually from about 20° to 30°. (Because of the transposition between elements, this means that adjacent elements are fed with a progressive phase lead 160° to 150°.) In the active region (elements 7 to 4) the amplitude drops sharply because of power absorbed by the strongly radiating elements, and the phase shift averages about 90° between adjacent elements. Finally, in the unexcited or reflection region (elements 3 to 1), the amplitude drops to very low values and the phase shift between element positions is nearly zero (corresponding to the zero phase shift or infinite phase velocity in the attenuation region of a low-pass filter).

The resulting element currents for the log-periodic dipole array of Fig. 13 are shown in Fig. 14, both in amplitude and phase. From the current amplitudes (noting that small contributions from elements 12 through 8 tend to cancel one another because of the nearly 180° phase shift between them), it is evident that the only elements that will contribute appreciably to the radiation are elements 7, 6, 5, and 4. For these elements, the phase difference between adjacent members is approximately 90° leading, so a backfire radiation will be expected. The phasor diagrams for $\phi = 0°, 90°$, and 180° are shown in Fig. 15 and the resulting radiation patterns are shown in Fig. 16. (The E-plane pattern is the H-plane pattern modified by the directivity of the individual elements in this plane.)

As operating frequency is decreased or increased the active region moves up or down the array, but radiation pattern and input impedance remain almost constant.

General properties of log-periodic and frequency-independent antennas. The manner of operation of the log-periodic dipole array has been described in some detail because of the insight it gives into what are believed to be general requirements for successful frequency-independent operation. These appear to be as follows:

1. An excitation of the antenna or array from the high-frequency or small end of the antenna.

2. A backfire radiation (in the case of unidirectional radiators), so that the antenna fires through the small

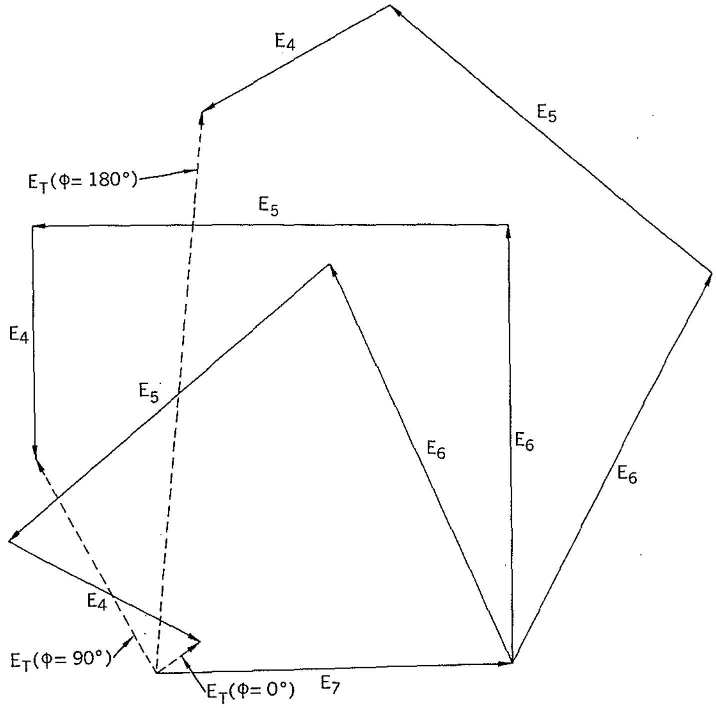
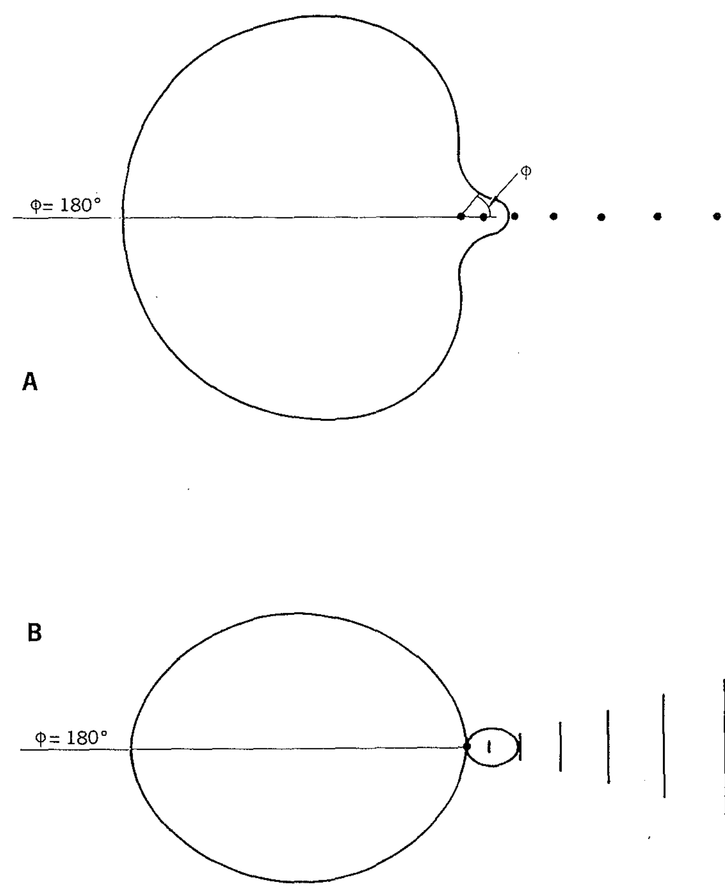


Fig. 15. Phasor sum of radiated fields from currents shown in Fig. 14, for ϕ values of 0°, 90°, and 180°.

Fig. 16. Radiation patterns resulting from fields shown in Fig. 15. A—H-plane pattern. B—E-plane pattern.



part of the antenna, with the radiation in the forward direction being zero or at least very small. For bidirectional antennas the backfire requirement is replaced by a requirement for broadside radiation. In any case, the radiation in the forward direction along the surface of the antenna (which theoretically extends to infinity) must be zero or very small.

3. A transmission region formed by the inactive portion of the antenna between the feed point and the active region. This transmission line region should have the proper characteristic impedance and negligible radiation.

4. An active region from which the antenna radiates strongly because of a proper combination of current magnitudes and phasings. The position and phasing of these radiating currents are such as to produce a very small radiation field along the surface of the antenna or array in the forward direction, and a maximum radiation field in the backward direction (broadside for bidirectional antennas). For successful backfire antennas these requirements are frequently met with separations less than a quarter wavelength and phasings near 90° leading, for adjacent elements in the active region. For broadside radiation the phasings must, of course, be zero.

5. An inactive or reflection region beyond the active region. All successful frequency-independent antennas must exhibit a rapid decay of current within and beyond the active region, so that operation will not be affected by truncation of the structure. A major cause of the rapid current decay is, of course, the large radiation of energy from the active region. An additional cause, in at least some types of frequency-independent and log-periodic antennas, is the attenuation resulting from the rejection of incident energy by the reflection region (the filter stop-band effect mentioned previously). The prevalence and importance of this latter filter action are still uncertain.

Finally, two other observations may be made. Although we have tended to think of the structures of Figs. 4, 5, 7, and 8 as single antennas and the structure of Fig. 9 as an antenna array, it appears that most frequency-independent and log-periodic antennas may be thought of as antenna arrays, with the array factor playing an important role in the formation of a proper endfire or broadside pattern. The localization of the individual radiating elements may be easier to see for the case of the log-periodic dipole array of Fig. 9, but the array action can also be observed in the other cases; it is particularly evident in the case of the fairly tightly wrapped conical log-spiral.

The second observation relates to the similarity between antennas derived from the angle concept and log-periodic concept.[17] Both lead to a solution of the unlimited-bandwidth problem and for this reason both have come to be known as frequency independent.

An example of the similarity between these two antenna types can be demonstrated in the case of the log-periodic wire antenna of Fig. 8, which produces a linearly polarized beam off the apex with the electric vector parallel to the transverse elements. If two such antennas are arranged in space quadrature along a common axis, and with a common origin but with one structure scaled a quarter period from the other, the resultant combination produces a circularly polarized beam with a pattern that rotates about the axis with frequency, exactly as in the case of the conical equiangular spiral antenna. Conversely, of course, if the pattern of a conical equiangular spiral is probed with a linear receiving antenna of fixed plane of polarization, the measured pattern will vary log-periodically with frequency, as does the pattern of the antenna of Fig. 8.

In addition, it is pertinent to note that if a narrow-armed conical equiangular spiral (an angle structure) is flattened sideways (along the axis), it becomes a log-periodic zigzag antenna.

Recent developments

The log-periodic and angle concepts have been used to generate many highly useful antennas of large bandwidth. Fig. 17 shows a very practical two-element array of log-periodic dipole arrays capable of maintaining a nearly constant radiation pattern and a 50-ohm input impedance over the frequency range from 450 to 2000 Mc/s. The 50-ohm input impedance results from feeding two 100-ohm arrays in parallel. Although the dipole array is a balanced structure, it can be fed as shown with a coaxial cable running up the inside of one of the hollow transmission lines, utilizing the frequency-independent balun effect previously noted in connection with the conical log-spiral antenna.

Fig. 17. Two-element array of log-periodic dipole arrays.

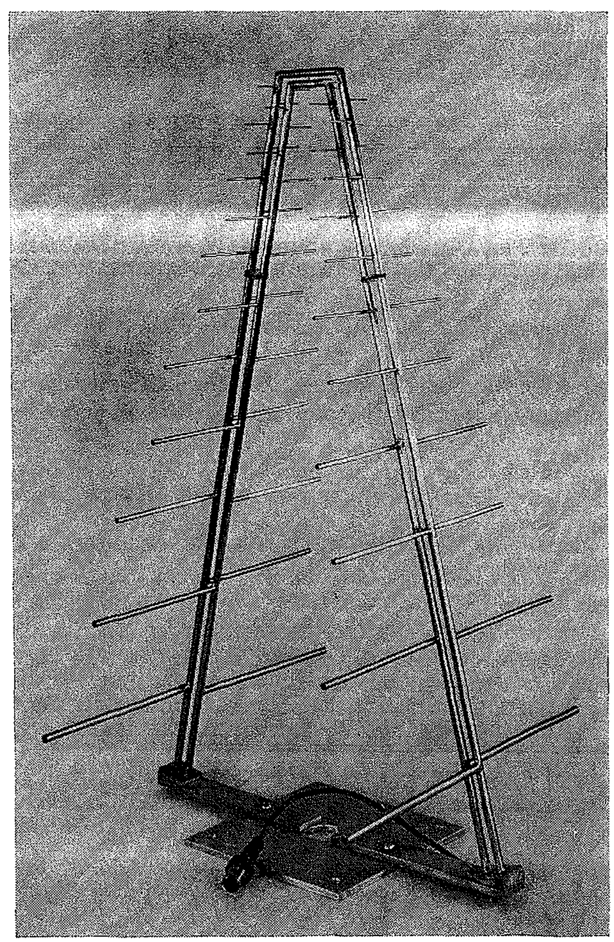


Fig. 18 shows one version of the log-periodic resonant-V developed by Mayes and Carrel.[18,19] This antenna was designed to overcome one of the major shortcomings of the ordinary log-periodic dipole array—namely, the long physical length of array required to cover a very wide band of frequencies. The antenna of Fig. 18 is designed to operate in several modes. In the lowest order $\lambda/2$ mode, the operation is similar to that of the log-periodic dipole array because the forward tilt of the elements has small effect for this mode. However, as the frequency of operation is increased beyond that at which the shortest elements are resonant—that is, when the active region runs off the front end of the array—the largest elements at the rear become active in the $3\lambda/2$ resonance mode. In this mode the forward tilting of the elements ensures a good unidirectional pattern of high directivity. As the frequency is further increased, the active region moves forward through the array in the $3\lambda/2$ mode until once again it runs off the front end, to return to the rear in the $5\lambda/2$ mode. This scheme makes it possible to obtain large bandwidths of the order of 20 to 1 with a relatively compact array. The pattern and impedance characteristics remain good over the entire frequency spectrum except for intervals about the mode-transition frequencies. Based on these principles, arrays have been designed to cover all of the television channels from 2 through 83, corresponding to a frequency range from 54 to 890 Mc/s.

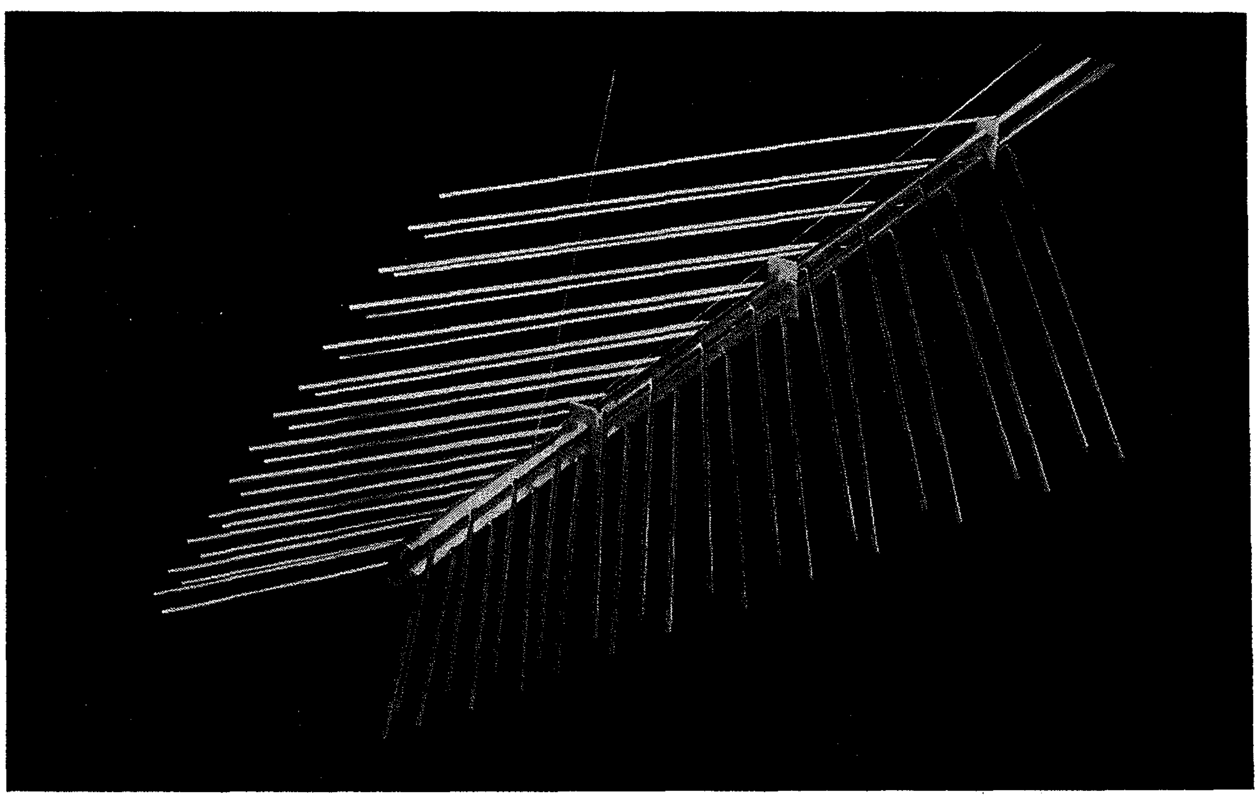
Another interesting development is that of a log-periodic folded-dipole array. At first thought it would appear that such an array could not work because the short elements at the front of the array present a very low impedance, thus short-circuiting the transmission region leading to the active region. This difficulty is circumvented[20] by connecting the folded dipoles in series with the transmission line, rather than in shunt, and recognizing that the active region will occur near first resonance, that is near the element length ($\lambda/4 < l < \lambda/2$) where the capacitive reactance of the short antenna resonates with the inductive reactance of the folded dipole viewed as a short-circuited transmission line. This unusual operating mode for the folded dipole results in a shorter element length for resonance, and consequently a narrow width for the resulting folded-dipole array.

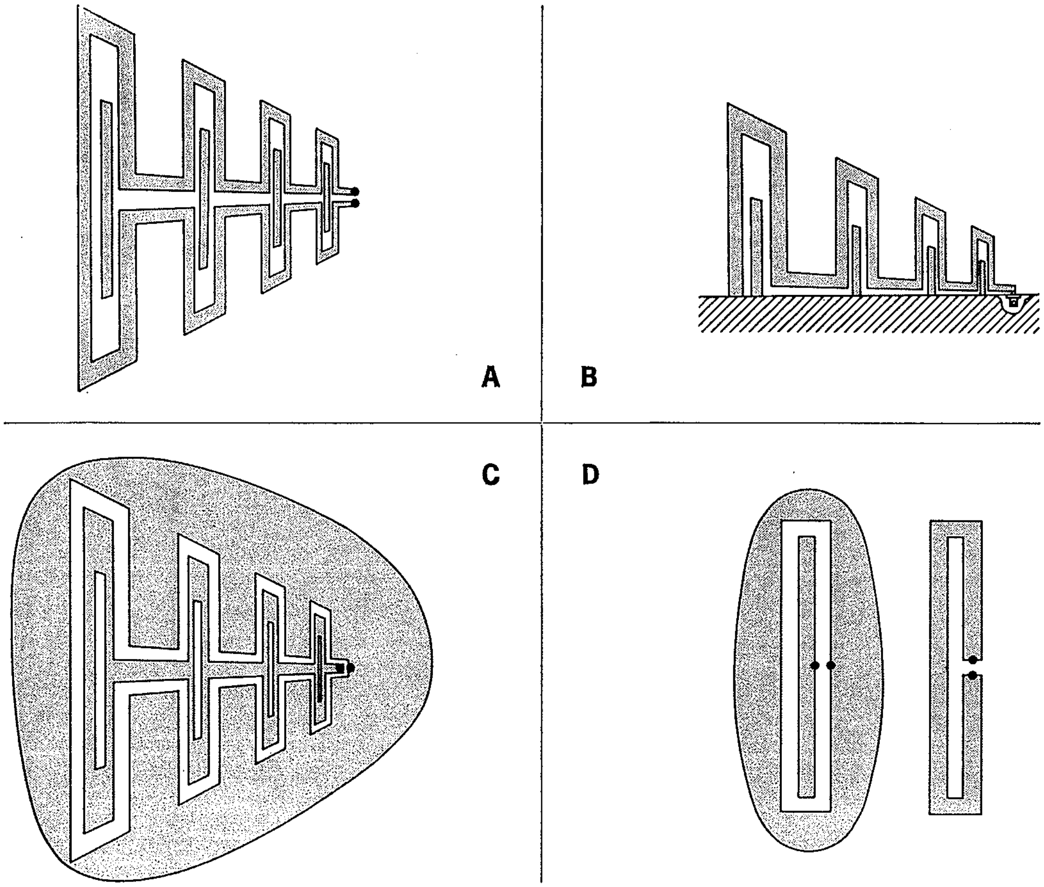
A major problem with log-periodic structures has been the design of an antenna that will operate successfully when fed against a ground plane to produce vertical polarization. One half of the antenna of Fig. 8 can be operated over ground to produce horizontal polarization, as can an inclined horizontal log-periodic dipole array. For vertical polarization, particularly in the high-frequency band (3–30 Mc/s), it is desirable to use the equivalent of a log-periodic monopole array that has a height of only approximately $\lambda/4$ at the lowest operating frequency, rather than $\lambda/2$. Because of the necessity for introducing a transposition between elements (or otherwise producing the required phase difference between elements) it is not possible simply to use one half of a log-periodic dipole array fed against ground.

Several solutions to this problem, having varying degrees of success for different applications, have been developed by a number of workers in the field.[21–23] A quite recent development[24] using folded monopoles with added phasing elements promises to be very useful.

Three versions of this antenna are shown in Fig. 19.

Fig. 18. Log-periodic resonant-V array, for operation in several modes.




A
B
C
D

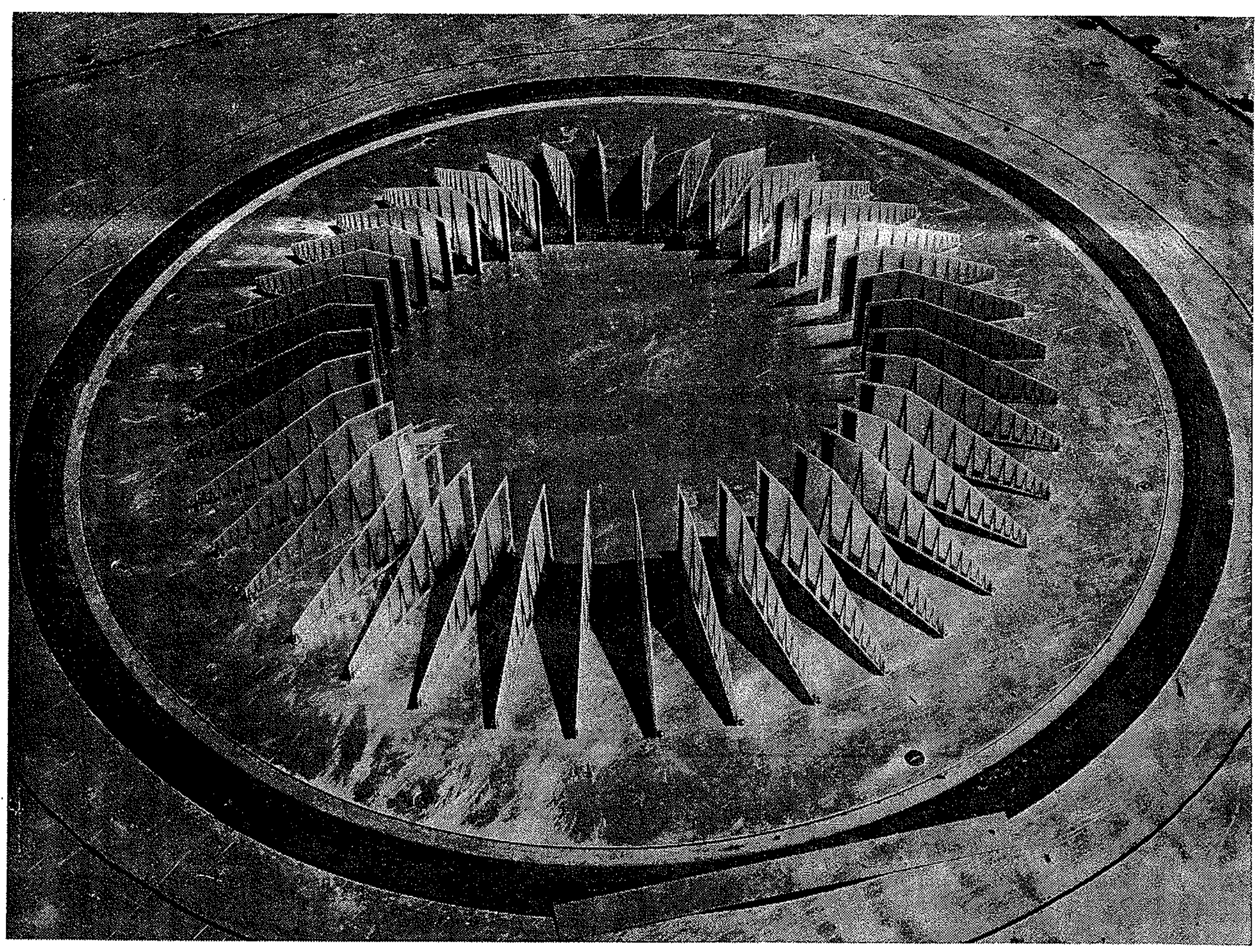


Fig. 19. Log-periodic arrays of folded elements.
A—Log-periodic folded-dipole array. B—Log-periodic folded-monopole array. C—Log-periodic folded-slot array. D—Duals: folded slot and folded dipole.

Fig. 20. Model of wide-aperture log-periodic array for high-frequency radio direction finding (3–30 Mc/s).

The log-periodic folded-slot array (A) was conceived first, but by duality, the log-periodic folded-dipole array (B) is obtained automatically. Because this array possesses the proper image symmetry about the horizontal axis (horizontal currents in opposite directions, vertical currents in the same direction), one half of the array can be fed against a ground plane to produce the folded monopole array of (C). The duals, folded slot and folded dipole, are illustrated in (D). The dimensions of the phasing slots in (A), or phasing strips in (B) and (C), are adjusted experimentally to provide the required phasing between successive dipoles or monopoles to produce a good backfire beam.

For greater directivity than can be achieved with a single frequency-independent antenna (or array) it is possible to use the frequency-independent structure as the broadband feed of a large paraboloid. Although the resultant combination is no longer frequency independent, high-gain antennas having a usable bandwidth as high as ten to one have been built by use of this approach.

Some of the high-gain paraboloid tracking antennas for the Atlantic Missile Range have been modified to use two conical log-spiral antennas as a circularly polarized broadband feed in a conical scan system. This application covers a frequency band of 215 to 1000 Mc/s, but the feed elements themselves are capable of operating continuously to 2300 Mc/s.

An alternative approach to the high-gain broadband problem is illustrated in the model of a broadband (3–30 Mc/s) wide-aperture radio-direction-finding array shown in Fig. 20. For frequency-independent arraying, the individual elements should lie along radials and be arranged to fire inward toward the common origin (toward the hub of the wheel). Unfortunately this arrangement requires opposite elements to fire through each other, and severe pattern deterioration results. In the array of Fig. 20 the log-periodic antennas fire outward. A 100° sector of elements is connected together through an appropriate phasing network and rotating switch or goniometer to form a narrow beam, which rotates with the goniometer as the latter connects in elements on one side of the sector and disconnects them on the other side. Again, this arrangement is far from being frequency independent, but the use of broad-band log-periodic structures as array elements is an improvement over the earlier use of frequency-sensitive elements.

This last example indicates that although truly remarkable progress has been made in the past decade in achieving broadband antenna operation there still remain some challenging problems for the future. Among these challenges are the design of broadband antennas having very high gain, and the design of frequency-independent antennas to produce specified radiation patterns.

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Down-to-earth Army antenna

Built low and compact for field radio operations, a vertical loop is easy to transport and set up. Early square-shaped models got high efficiency ratings, and an octagonal version is even better

By Kenneth H. Patterson

Army Limited War Laboratory, Aberdeen Proving Ground, Md.

Fight fans say a good big man will beat a good little man every time. But Army signalmen setting up radio sites in the jungles of Vietnam would much rather use a small radio antenna with adequate efficiency than a larger, more efficient one that has to be strung on high masts. Mobility and concealment are high-priority features in a guerrilla war.

For communications at medium and high frequencies, the Army normally uses various forms of long wires, dipoles, and rhombics. To be reasonably effective, the long wire should be a minimum of about 100 feet (plus a good ground), the dipole should be at least twice as long, and the better rhombics can exceed 600 feet. In addition, all of these antennas should be elevated to a minimum height of about 40 feet, with 80 feet being better. Such space and height requirements are invariably so difficult to meet that compromises are usually made in the field.

But a solution has been developed at the Army Limited War Laboratory at Aberdeen Proving Ground, Md. A vertical loop antenna (in the shape of an equilateral octagon, 5 feet to a side) has been built that doesn't have to be raised above the ground. Despite the inherently low radiation resistance of a loop, the new design usually does as good a job as a full-length dipole 40 feet above the ground, even though the length of the dipole is about 234 feet at 2 Mhz and 94 feet at 5 Mhz.

One can get an idea of the difference between the low-height performance of vertical loops and horizontal dipoles by visualizing their basic characteristics. The most significant consideration is the effect on the signal of reflection from the ground.

The theory of image antennas tells us that vertically polarized signals are reflected by the earth without a phase shift. Horizontally polarized signals, on the other hand, are phase inverted upon being reflected. The exact amount of inversion depends on orientation, but in most instances it can be expected to approximate the full 180°. It then follows that when the dipole antenna is used at heights lower than about $.12\lambda$, the reflected wave has a cancelling effect on the incident wave. In fact, zero height and a perfectly conducting earth are conditions that cause total cancellation and zero radiation.

Hit the dirt

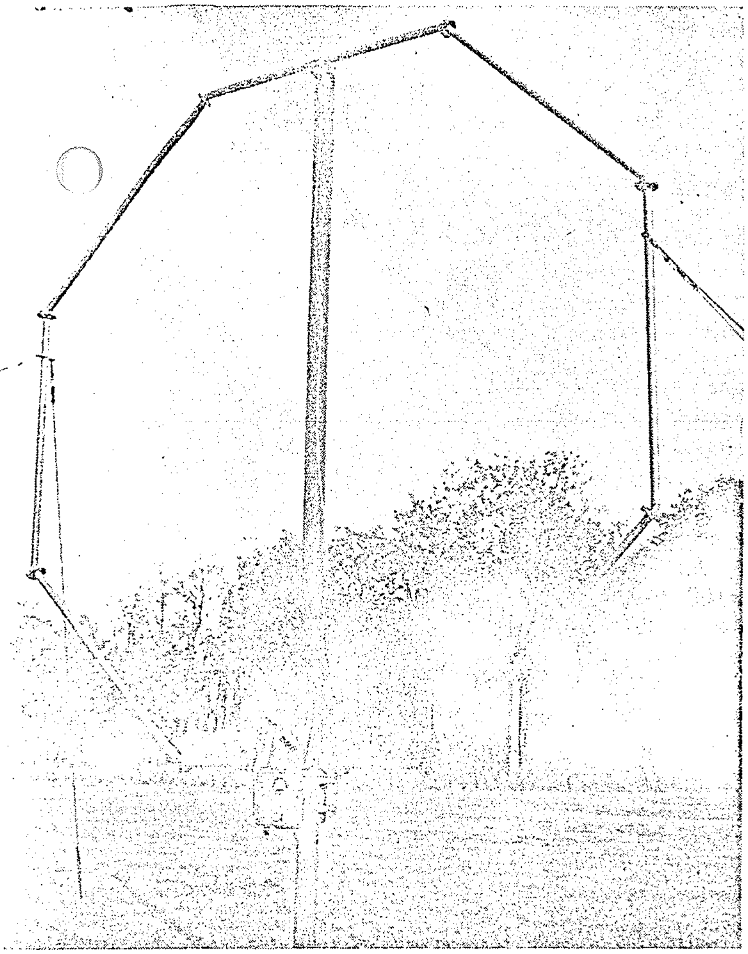
Under the same circumstances, however, the reflected wave from a vertical loop combines with the incident wave under what is predominantly an in-phase condition. This quite naturally results in an increase in the amount of energy radiated in useful directions. With greater proximity to the earth, therefore, the radiation from the loop may actually improve. It cannot, of course, exceed a limit of twice its radiation in free space.

A close examination of the expressions for the radiation resistance values of the two antenna types, including the height-modifying terms, shows the loop approaching twice its free-space value as a function of decreasing height, while the horizontal antenna goes to zero.

If we wish to operate our antenna at a really low height, therefore, we must use vertical polarization. And if we cannot accept a null in our overhead

The author

Kenneth H. Patterson, who has been developing electronic communications equipment since 1923, designed the loop antenna described in his article. Before joining the Army Limited War Laboratory, he was chief of radio frequency development at the Ballistic Research Laboratories at Aberdeen. There he supervised work on a large portion of the instrumentation used to measure missile performance at the White Sands, N.M., range.



Will travel. Octagonal loop antenna with 5-foot sides is easily set up, dismantled, and transported. Efficiency near the earth matches that of conventional elevated antennas as long as several hundred feet.

pattern, we are restricted to a vertical loop. The problem is to improve the efficiency of the loop antenna.

Besides solving spatial and height problems, the antenna developed at Aberdeen:

- is operational from below 2.5 Mhz to above 5.5 Mhz—although tuning ranges as great as 10:1 are possible;
- offers a pattern factor well suited to both short- and long-range ionospheric propagation with no overhead nulls;
- has a predicted efficiency level of from 20% to 80% throughout the operational band. (Future models will do much better.);
- is self-supporting in use;
- can be transported by a small vehicle when packed;
- can be set up or dismantled by a crew of no more than three men in less than 30 minutes;
- doesn't require an artificial ground plane;
- has sufficient strength to withstand normal wind and rain storms.

The power rating, determined by the voltage breakdown and current ratings of the capacitors used in the matching network, is arbitrary. A 1-kilowatt version has been built.

The loop isn't being suggested as the best design for every antenna application. It won't, for instance, outperform the large rhombic in specific low-angle unidirectional tasks. But the loop can do the job at installations where real estate is limited and a complex of high antenna masts is impractical.

Field tested

Loop antennas, some shaped as squares and others as octagons, have undergone extensive field trials in the U.S. and Vietnam. The results in all cases have been excellent.

Before the loop was selected, other configurations were considered and rejected. Grounded verticals, or whips, were ruled out because of their height and their inherently restricted radiation pattern.

The horizontal series-fed antennas, including dipoles, long wires, V's, and rhombics, were also dismissed because of their height requirements. As noted before, the radiation resistance of these antennas approaches zero as height is reduced. (Radiation resistance is obtained by dividing the total radiated power of an antenna by the square of the effective antenna current measured at the point where power is supplied.)

The vertical loop, being compact and vertically polarized, and having no overhead null in its pattern, got the nod almost by default. The superior low-height performance of a loop antenna was suggested to the Proving Ground by David Sunstein of the General Atronics Corp., Philadelphia.

To improve the loop's efficiency, two problems had to be solved: radiation resistance (R_R) had to be increased, and the sum of all other losses (R_L) had to be reduced. To assess the problems and the expected efficiency (E), the standard efficiency equation was used—efficiency equals radiation resistance divided by the sum of radiation resistance and R_L.

Cutting losses

The two principal sources of loss in a loop antenna are the inherent resistance in the conductor used to form the loop and the resistance in the antenna-matching network.

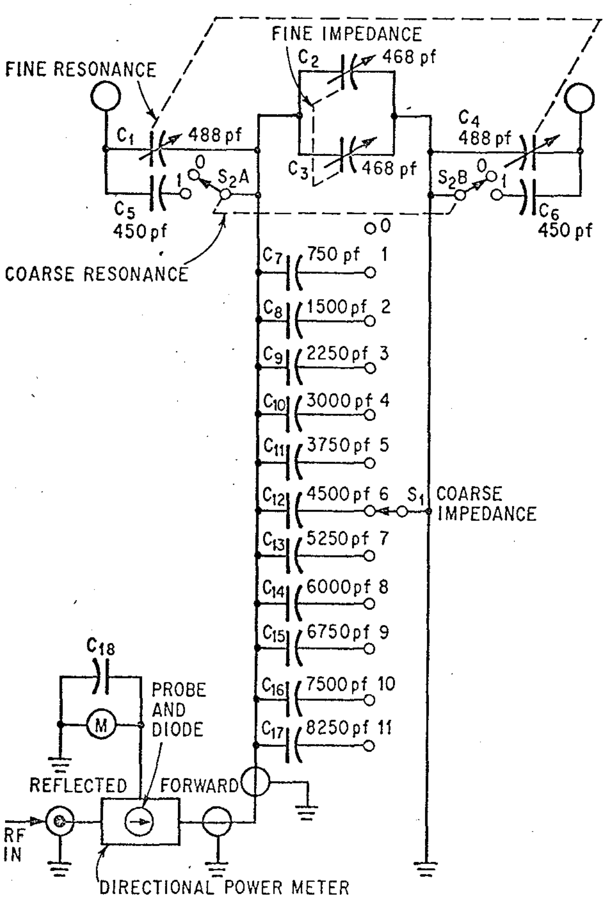
In considering the losses in the matching network, it was decided the antenna would be driven from a 50-ohm or 70-ohm coaxial line, thus requiring the input portion of the matching network to handle the comparatively high currents characteristic of low impedance levels. Particular care was taken in the design of the input part of the network to reduce the normal resistive power losses to a practical minimum.

To do this, the customary taps or links, with their inherent resistive losses, were discarded in favor of variable air-dielectric and fixed mica capacitors, both of which have low loss characteristics. The schematic diagram of this matching network on page 113 is conventional except that the low-value, high-impedance capacitor is divided into two equal sections to provide a close—though not precise—balance with respect to ground, and to permit higher voltages. For low-power applications where the emphasis is on portability, the second tuning capacitor is omitted.

In either balanced or unbalanced form, the capacitative matching network is practical, versatile, and efficient. Matching losses can be cut to the point where they can be ignored in many instances. If a loop with a high ratio of inductance to capacitance (L/C ratio) is attempted, however, a significant portion of the circulating current flows through the stray (shunt) capacitance and a poor match and an inefficient coupling condition results. This could occur, for example, if the operating frequency is raised to a level closely approaching the self-resonant frequency formed by the loop inductor and the stray shunting capacitance.

In using the matching network, the transmitter is turned on. With the arrow on the pick-up probe in the "REFL" direction the meter reads the reverse power level. The controls are adjusted to the deepest null until the meter reads zero, indicating a perfect match. The transmitter power can then be advanced to a maximum level and the pick-up probe rotated to "FOR" to read the forward power level.

The method selected to reduce loss in the loop



Indicator. The arrow on the front of the probe assembly indicates whether forward or reflected power is being provided by the pickup link. The link applies this energy to an r-f rectifier whose d-c output drives the power meter movement.

itself is a straightforward brute-force technique using conductors with an exceptionally large surface area. In the majority of the models built to date, the loop conductors are composed of 1½-inch tubing. With such large conductors, the r-f losses due primarily to skin effect are sharply pared.

Radiation resistance

The next step in improving efficiency is to obtain the greatest practical radiation resistance. This involves another classic equation:

$R_R = 3.12 \times 10^4\ [NA^2/\lambda^2]^2$

where R_R = the approximate radiation resistance of a small loop.

N = the number of turns in the loop.

A = the area enclosed.

λ = the wavelength of the frequency being used.

The dimensional terms are expressed in the same units, squared or lineal.

The equation shows that two parameters, N and A, may be controlled to possible advantage. In the case of N, imagine a small loop antenna composed of a single turn shaped to form a square 3 feet on a side. Further assume that in all cases a very short drive cable is used, a low-to-moderate L/C ratio is employed, and losses in the capacitative matching network are negligible. In other words, essentially all the losses in this antenna stem from resistance in the wire conductor. Under these circumstances, the losses (R_L) are, to a great extent, directly proportional to the total length of the conductor.

As one would expect, the efficiency of this imaginary reference antenna is extremely low. In fact, if it were made of 18-gauge copper wire, the computed free-space efficiency at 3 Mhz would be a mere 0.0457%.

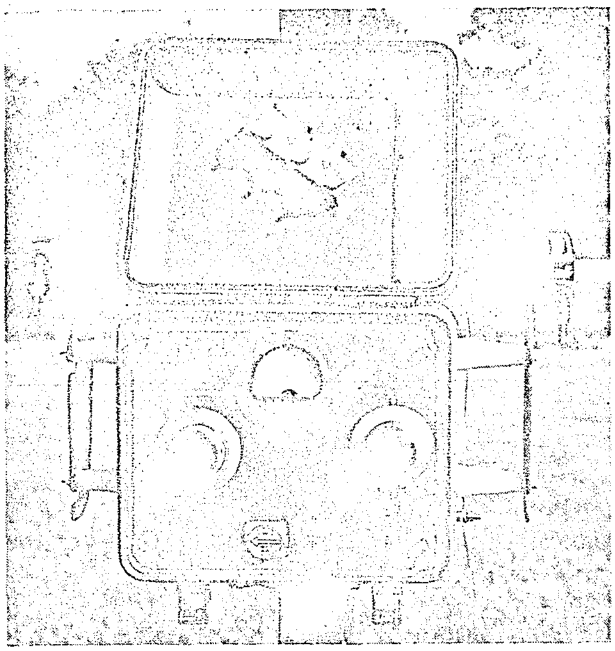
If the number of turns in a 3-foot-square loop were increased to three, the radiation resistance would be increased by a factor of nine. Unfortunately, the extra turns in this example also increase the loss resistance by a factor of three. Consequently, the efficiency improvement is reduced to a net factor of three; this is a reasonably significant improvement, but efficiency is still intolerably low.

The same conductor used for the three-turn loop can be reshaped to form a single-turn square loop with 9-foot sides, increasing its area by a factor of nine. Since R_R is proportional to A^2, the radiation resistance is now greater by a factor of 81. Here again, however, the losses are three times greater, resulting in a net efficiency improvement factor of 27. Using 18-gauge copper wire, the free-space calculated efficiency of this antenna is 1.23% at 3 Mhz and 13.1% at 6 Mhz.

Unless extremely severe restrictions are imposed on space, therefore, the largest practical single-turn loop is preferable to a smaller multiturn unit.

Shaping up

The final consideration is form. Efficiency has been shown to be highest when a fixed-length peri-



Precise tuning. Reflected power meter permits fine tuning and exact matching, and monitors forward output power level. The coarse adjustments make discrete, step-by-step changes, while the fine controls provide continuous variations overlapping the coarse steps.

meter encloses the greatest possible area. A circular configuration is best from a performance standpoint, but practical factors, such as the types of commercial tubing available, must be considered. This is especially true of developmental models, which are most economically built with straight tubing and "els."

Comparisons were made between a square, an equilateral octagon, and a circle, all of the same total perimeter. The octagon encloses 20% more space than the square, and the circle 29% more. Since radiation resistance is proportional to the square of the area, the radiation efficiency of the small octagon loop exceeds that of the square by about 45%, while the circular loop outperforms its square counterpart by about 65%. But because straight tubing packs more compactly, is more readily available, and is much cheaper than the circular kind, the octagon shape was chosen. The size—5 feet is the maximum length for any part—was held down so the antenna could be easily disassembled and carried in a small motor vehicle.

With the loss reduction achieved, the loop antennas that have been built provide a calculated free-space efficiency of about 22% at 2.5 Mhz and about 77% at 5 Mhz.

In the final analysis, however, an antenna's efficiency, or, perhaps more important, its effectiveness, must be proven in field tests.

As noted before, preliminary tests in the U.S. were followed by several hundred more conducted by the Army in Vietnam. Numerous frequencies and different ranges were used at various times of day in the later trials. Also, the tests were at all times confined to relatively low power levels, usually on the order of a few watts.

Nevertheless, all contacts attempted in Vietnam were not only 100% successful, but, with few exceptions, were rated "perfect." These few exceptions were never ranked lower than "good."

Competitive trials

The earlier U.S. tests, however, were perhaps more informative in providing direct comparisons between the loop and other antennas. As a matter of fact, the octagonal loop made of 1½-inch tubing is actually a second-generation model. For convenient construction, the earlier developmental models were constructed in a square shape and formed from ¾-inch tubing. Later tests have revealed a very significant 8-decibel superiority for the octagon version over the square.

Despite this, however, the square models showed themselves to be effective radiators in the early trials. In numerous ionospheric field tests conducted between Aberdeen Proving Ground and two widely separated mountain valleys in the Alleghenies, the square loop proved, in almost all instances, as effective as a full-sized dipole, and actually surpassed the dipole in most matchups. Several frequencies between 2.5 Mhz and 5.5 Mhz were used, and in each instance the dipole was cut to the optimum length. Furthermore, the dipole was strung in a cleared area at a height of 40 feet, while the bottom of the loop was only 4 feet above the ground.

In one field test to determine ground-wave propagation over a distance of two miles, a dipole supported at a height of 6 feet delivered a measured field strength of about 68 microvolts on a transmitter power of 6 watts. The square loop gave about 600 microvolts on the same power, and the octagon produced approximately 1,500 microvolts. The frequency used was 3.275 Mhz—not the best for the small loops.

It's true that the dipole's propagation in this particular test suffered from a less favorable polarization. But polarization characteristics are basically inherent and have to be considered in choosing antennas for operation at low heights.

Modifications

Though the octagon loop is a second-generation version, further improvements are being made. With the fiberglass mast used in some recent models, a single soldier can set up the antenna. For some applications, the diameter of the tubing can perhaps be advantageously increased from the present 1½ inches to 3, 4, or even to 6 inches. For other applications, the usual output tank circuit and matching network can be eliminated if the final amplifier is located close to the loop. The loop then can serve as both the radiation element and the final tank circuit, a technique especially useful where power is low and portability a prime requisite. Models of such long-range, highly portable antennas have been successfully tested at the Limited War Laboratory.

Basics for Beginners

Antennas and Feeders

Part I — Resonance in Linear Circuits

BY GEORGE GRAMMER,* W1DF

Some years ago a widely-used textbook on radio engineering began a chapter on antennas with this statement: "An understanding of the mechanism by which energy is radiated . . . involves conceptions which are unfamiliar to the ordinary engineer." Obviously radiation must be a stiff subject. So we simply ask you to accept the well-known fact that energy *is* radiated in the form of electromagnetic waves. We won't attempt to explain why.

Everyone who has studied for an amateur license has been introduced to wavelength and frequency. The formula is:

$$\textit{Wavelength in meters} = \frac{300}{\textit{Frequency in Mc.}}$$

In Fig. 1 the transmitter is generating a radio-frequency voltage, indicated by the sine wave in the upper drawing. When the voltage is applied to an antenna, energy is radiated into space and travels away with the speed of light. As shown by the lower drawing, it covers a certain distance — one **wavelength** — in the time the voltage takes to go through one cycle.

Current in a Wire

This relationship between wavelength and frequency has a very practical use. Suppose we connect an r.f. ammeter in the center of a wire having a length L, as in Fig. 2. Further, suppose that by some means we introduce r.f. energy of adjustable frequency into the wire. If the frequency is gradually raised, it will be found that the current indicated by the ammeter will also rise, at first. But after reaching a maximum at some frequency, f, the current will start to go down again if we continue to raise the frequency. This is the sort of thing we found to happen in an LC circuit, as discussed in an earlier article.[1] The wire, in fact, acts like a resonant circuit. It is tuned to the frequency, f, for which its length is equal to one-half wavelength. If the wire is 40 meters long, for example, 40 meters would be one-half wavelength, and the full wavelength would be 80 meters. From the formula above, this would correspond to a resonant frequency of 300/80, or 3.75 Mc.

A wire such as this is called a **dipole,** when its length is of the order of a half wavelength, or less. One exactly a half-wavelength long is called **a half-wave dipole.** Very often, the simple term "dipole" is used when a half-wave dipole actually is meant.

Two Practical Points

Before going farther, it is well to translate this into a more familiar unit of length, the foot. Converting units and changing to a *half* wavelength gives us

$$\tfrac{1}{2}\ \textit{wavelength in feet} = \frac{492}{\textit{Frequency in Mc.}}$$

or

$$\textit{Resonant frequency in Mc.} = \frac{492}{\textit{Length in feet}}$$

The second point is this: these formulas are not quite accurate for an actual wire. They apply only to a wave traveling in space. In a practical half-wave antenna the difference amounts to about 5 per cent, on the average. Thus an *average* formula for resonant length would be

$$\tfrac{1}{2}\ \textit{wavelength in feet} = \frac{468}{\textit{Frequency in Mc.}}$$

* Technical Director, ARRL.

[1] "A.C. in Radio Circuits," Part II, *QST*, April, 1963.

The earlier group of articles in this "Basics" series concerned itself with a class of circuits using concentrated values of inductance, capacitance and resistance, generally called "lumped constants." The present article opens a discussion of circuits having "distributed constants" — "linear" circuits. A linear circuit is simply a wire, or group of conductors, running in a more-or-less straight line.

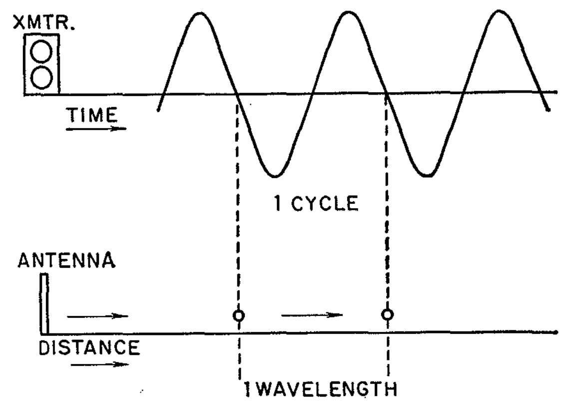


Fig. 1—One wavelength s the distance that radiated energy will cover, traveling at the speed of light, during one cycle of the radiated frequency.

Remember that this is only an average. In a particular case the actual resonant length might differ by a few per cent from the length given by this formula. The difference is usually small enough to have little practical effect.

Electrical Length

Electrically, the length given by the last formula is a half wavelength because it is a *resonant* length, even though it is physically short of being a half wavelength in space. We can account for the difference in length by the fact that energy does not travel quite as fast along a wire as it does in **free space.**

In some cases, as you will see later when we get to transmission lines, there can be quite a marked difference between electrical wavelength and free-space wavelength. When you see the length of an antenna or line expressed in terms of wavelength you can safely assume that an electrical measure is being used, unless it is made plain that free-space measure is meant.

Enter Time

In the preceding series of articles[2] we dealt with circuits that offered a complete path around which electrical energy could move. A wire such as is shown in Fig. 2 doesn't offer any such path. How is it that current can flow in it?

In the "closed" circuits considered earlier it was assumed, without our having said it in so many words, that electrical energy traveled around the circuit so rapidly that its action could be taken to be instantaneous. In the circuits we use in transmitters and receivers for frequencies up to 30 Mc., at least, this is a satisfactory assumption. As long as the circuit is small compared with the wavelength (the wavelength corresponding to the frequency we happen to be using) the action is instantaneous, for all practical purposes.

But an antenna such as the wire in Fig. 2 is *not* small compared with the wavelength. If the length L is one-half wavelength, a length of time equal to one-half cycle of the applied frequency is needed for energy to go from one end of the wire to the other. Imagine a voltage applied to the left-hand end of the wire at an instant when the voltage is at the positive peak of the cycle. A voltage impulse will go along the wire to the right, reaching the end one-half cycle later. But at this instant the *applied voltage* has moved on to its *negative* peak.

Standing Waves

In other words, when the left-hand end of the wire is negative the right-hand end is positive,

[2] "A.C. in Radio Circuits," Parts I–V, incl., *QST*, March–July, 1963.

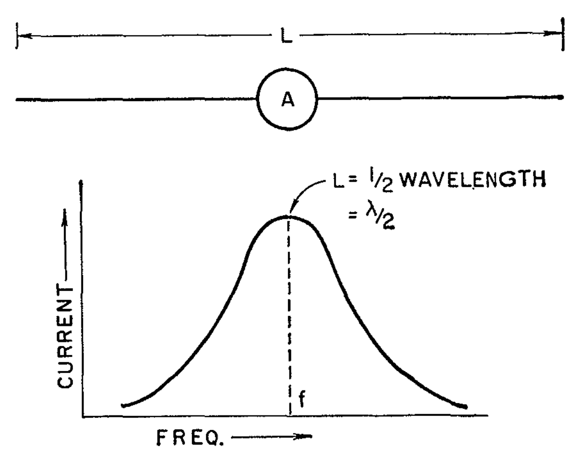


Fig. 2—The r.f. current at the center of a wire is highest when the wire length is equal to one-half wavelength.

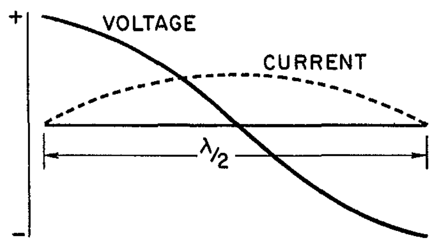


Fig. 3—The current and voltage along a half-wave wire have different values at all points along the wire. When plotted as shown above, the graphs are wave-like in shape, and since their positions are fixed with respect to the wire they are called standing waves.

and vice versa. Also, when the voltage reaches the end of the wire it comes to the end of the track. There is no place for it to go except back over the same path. The energy is **reflected** from the end. In going back, it combines with energy — from a later part of the cycle — that is going out.

Fig. 3 shows what happens in a wire one-half wavelength long. All the components of voltage, those traveling out and those reflected back, add up to make a **standing wave** of voltage. If we could go along the wire with a meter for measuring r.f. voltage we should find that the voltage is highest at the ends of the wire and is practically zero at the center. Between the ends and the center it gradually decreases. When plotted against length, as in Fig. 3, it is like part of a sine wave.

Polarity

The plus and minus signs on the scale at the left can be somewhat misleading. One end of the antenna isn't always positive, and the other end isn't always negative. In fact, both ends alternate between positive and negative each r.f. cycle. What the picture tries to show is that *when* the left-hand half of the wire is positive the right-hand half is negative. The reverse is also true. The voltages in the two halves of the antenna always have opposite polarity.

On the other hand, if we went along the wire with a meter for measuring the r.f. current, we should find that the current is zero at the ends. This you might expect, since current can't flow off the wire into space. The current gets larger as we move toward the center, and is largest right in the middle of the wire. In the drawing, the current is shown entirely on the plus side of the scale. Again this shouldn't be taken literally; it actually goes from positive to negative and back again each cycle. The picture means that the current is always flowing in the *same* direction, at any given instant, throughout the entire length of a half-wave wire.

Loops and Nodes

The point where the amplitude of a standing wave passes through zero is called a **node.** Thus in Fig. 3 the standing wave of voltage has a node at the center of the antenna. The standing wave of current has two nodes in this figure, one at each end of the wire.

A point of maximum amplitude is called an **antinode** or, sometimes, a **loop.** (Properly, the term loop refers to the entire segment of the standing wave between two nodes.) The standing wave of voltage has antinodes at the ends of the wire in Fig. 3, while the current antinode is at the center.

Note that where there is a current antinode there is a voltage node, and where there is a current node there is a voltage antinode. Also, an antinode of current is one-quarter wavelength away from a current node; similarly with the voltage. These two statements are true, in general, of all standing waves along wires.

Longer Wires

Tuned circuits using coils and capacitors are resonant at just one frequency, that for which the inductive and capacitive reactances are equal. An antenna isn't quite so simple. If the things shown in Fig. 3 happen when a wire is a half wavelength long because of the *time* it takes energy to surge back and forth, it seems reasonable to expect that another half wavelength of wire added to the first will see a repetition of these same events. And so it is. There will be a repetition each time a half wavelength is added.

Fig. 4 shows the **current** and **voltage distribution** when the wire is two half-waves (or one wavelength) long and three half-waves (1½ wavelengths) long. At the ends of each half-wave section the voltage is high and the current is zero. In the middle of each such section the current is high and the voltage is zero. But there

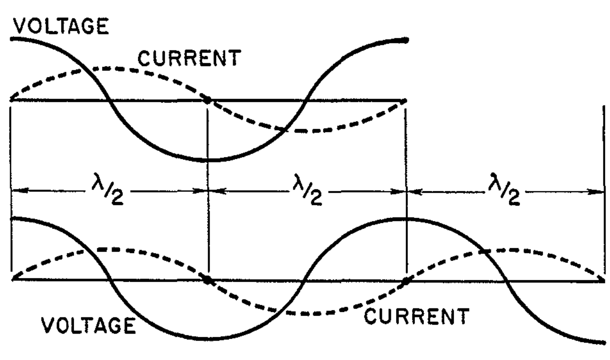


Fig. 4—Harmonic resonance. The upper drawing shows the standing waves on a wire one wavelength long; the lower shows them on a wire 1½ wavelengths long.

is a difference between two adjacent half-wave sections. You can see that if the voltage at the left end of the first section is positive, as shown, the voltage at the left end of the second section is negative. It has to be the same as the voltage at the right-hand end of the first section, of course, since the two sections are connected together. Also, if the current in the first section is positive, as shown, the current in the next section will be negative. That is, the currents in adjacent half-wave sections flow in opposite directions. This is called a **phase reversal.**

Phase

In the third section, shown in the bottom drawing of Fig. 4, there is again a phase reversal. This brings the phase relationships in this section back to exactly what they are in the first section. In other words, *alternate* half-wave sections have identical standing waves of current and voltage on them. They are said to be **in phase.** *Adjacent* sections are **out of phase.** This goes on no matter how many half-wave sections are added to the wire.

Harmonic Resonance

Each of these sections is just as much resonant to the applied frequency as another. In effect, we have two resonant antennas end-to-end in the upper drawing of Fig. 4, and three in the lower drawing. These are called **harmonic resonances,** since they occur at the same frequencies as the harmonics of a fundamental frequency. That is, they are integral (whole-number) multiples of the fundamental.

In the case of an antenna, the fundamental frequency is the one for which the entire wire length is equal to one-half wavelength. For example, an antenna that is a half wavelength long at 7150 kc. will be two half-waves long at 14,300 kc. (second harmonic), three half-waves long at 21,450 kc. (third harmonic), and so on up the scale. The actual multiples are approximate, not exact, integers. The resonant frequencies will differ slightly from exact harmonics. The reasons are the same as given earlier, in the discussion of the length of a practical antenna.

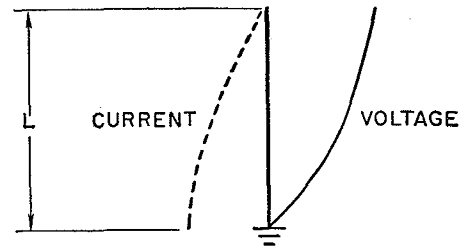


Fig. 5—Grounding one end of the antenna chops off one-half of the standing wave—that is, the length *L* need be only a quarter wavelength for the antenna to be resonant.

Grounded Antennas

A half wavelength is the shortest length of wire that will be resonant to a given frequency, if the wire is simply considered by itself. However, if we connect one end of the wire to earth the grounded end is no longer free. We can't raise the potential of the earth itself, so the voltage at the grounded end has to be zero. On the other hand, we *can* make current flow into the earth. Thus the earth can be made to act as a substitute for one half of the half-wave antenna.

Fig. 5 shows this. The current is large at the earth connection, and decreases to zero at the open end of the antenna. The voltage is zero at the bottom and has its greatest value at the top. But the length *L* for this antenna is only a *quarter* wavelength, at resonance. So a grounded antenna need be only half as long as a dipole antenna to be resonant at the same frequency. QST

(*Part II of this series will appear in an early issue.* — EDITOR.)

Strays

Outstanding achievement in any aspect of amateur radio will be recognized yearly by the St. Louis Amateur Radio Club, with an Amateur-of-the-Year Award. Nominations are solicited from all amateurs of the St. Louis area, and these should be sent to Lane Jackson, KØKJX, 645 Marshall Ave., Webster Groves, Mo., before Nov. 1. The award will be presented at a meeting to be held in the Mosely Auditorium Nov. 15, with ARRL Midwest Division Director Denniston as principal speaker.

W9IOP's famous "Second OP" has been entirely revised. The dial-a-prefix DX-operating accessory now includes the new African republics and most of the recent prefix changes; a total of more than 300 countries, each cross-referenced with its great circle bearings, time differential, postage rates, continent, and zone. QSL bureaus of the world are listed on one side and operating instructions on the other. The W9IOP Second Op is available from Electro-Voice, Inc., Buchanan, Michigan, for a dollar.

The QSO Club of Pasadena (Calif.) City College has announced their bi-annual (school radio clubs only) Field Day. The next session will be October 25–26, *local standard* time; and all school and college clubs are invited to participate. Write the QSO Club, c/o Ken Johnson, W6VEB, Department of Engineering and Technology, Pasadena City College, Pasadena 4, California.

OUR COVER

Our cover this month shows a transmitter and its schematic — products of W9YRV and WA9DNF. An exciting new balanced-modulator circuit is featured in this compact six-meter beauty. Read all about it in "The Single Sideband Sixer," beginning on page eleven.

Basics for Beginners

Antennas and Feeders

Part II — Antenna Impedance; Directivity

BY GEORGE GRAMMER,* W1DF

The way in which current and voltage are distributed along a wire, discussed in Part I,[1] may be in itself an interesting electrical phenomenon, but it is not merely that. It has important practical effects, too, in the workings of antennas and transmission lines. In antennas, the distribution exerts control over the radiation — taken up later in this Part — and establishes the conditions that must be met when r.f. power is fed to the antenna. The latter point can be summed up by saying that, as a result of the r.f. current and voltage distribution, the antenna has an *impedance* that must be considered when power is applied. In this respect the antenna is just like any other load in which energy is to be used up. Until we know the impedance of the load we don't know where to start in settling on the right way to feed it.

Antenna Impedance

Impedance, as it was defined in the earlier series,[2] is equal to voltage divided by current. When the current and voltage both change as we move along the antenna, as they do in Fig. 3, Part I, the impedance also is different everywhere along the antenna. Therefore, if we want to talk about antenna impedance we have to specify the point at which it is measured.

The customary place to measure the impedance of a simple antenna is at the center of the wire. In Fig. 1 an r.f. generator, *G*, is inserted in series with the antenna at its center. The voltage from the generator will cause a current, *I*, to flow; this current has the same value on both sides of the terminals. The antenna behaves like a circuit having resistance, inductance and capacitance in series. At the resonant frequency of such a circuit the inductive and capacitive reactances cancel each other,[2] leaving only the resistance. This is also true of the antenna. Thus at its resonant frequency the antenna "looks like" a simple resistance, and it is at this frequency that the current is largest. A half-wave antenna has a resistive impedance, measured at this point, in the neighborhood of 70 ohms. It is rarely exactly 70 ohms in any practical case, because the actual resistance depends on the same factors that affect the resonant frequency.

If the frequency is moved off resonance the impedance rises, just as it does in a series *LC* circuit. It also becomes complex — there is reactance, now, along with the resistance.

Now suppose the r.f. generator to be connected to one end of the antenna, as in Fig. 1B, with one ammeter at the end and the other at the center. As the frequency is varied, the current I_2 will reach its highest value at resonance, where the antenna is a half wavelength long. But the current I_1 at the terminal where the generator is connected will be *smallest* at this frequency. As seen by the generator, the antenna is just like a parallel *LC* circuit. That is, at resonance its impedance is maximum, and is a simple resistance. As the frequency is moved away from resonance the current I_1 increases; the impedance becomes smaller and is again complex, containing both reactance and resistance.

Impedance Values

Although it is possible to feed at any point along the wire, antennas are usually fed with r.f. power either at the center or the end. Thus the two cases illustrated by Fig. 1 have some practical importance. The resonant impedance at the end is much more dependent on the thickness of the antenna conductor and other such factors than is the impedance at the center. Values can range from a few hundred to several thousand ohms. The thicker the conductor the lower the resistance as viewed from the end. At the center, the

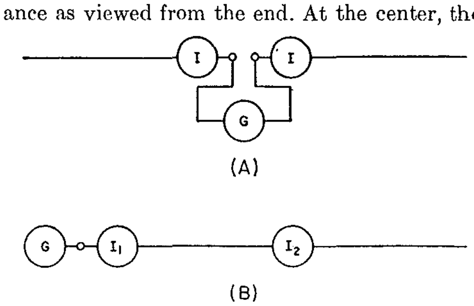


Fig. 1—A half-wave wire driven at the center behaves like a series-resonant circuit. One driven at the end acts like a parallel-resonant circuit.

* Technical Director, ARRL.
[1] "Antennas and Feeders," Part I, *QST*, October, 1963.
[2] "A.C. in Radio Circuits," Part II, *QST*, April, 1963.

effect of conductor thickness on the resistance, at resonance, is relatively minor.

The impedance of a grounded antenna usually is measured between the earth and the bottom of the antenna. Like the center-fed antenna with free ends, the grounded antenna acts like a circuit having I, C and R in series. As the antenna is only half as long, for the same resonant frequency, the resistance is only half as great. That is, it is in the neighborhood of 35 ohms, for an antenna a quarter wavelength long. This assumes a "perfect" ground — one that has extremely low losses at the operating frequency. Ordinary ground is far from perfect, and the earth connection usually adds quite a considerable amount of resistance to the system — often as much as 25 ohms. The ground resistance can be reduced by burying a large number of wires, having a length of about a half wavelength, going out from the base of the antenna like the spokes of a wheel. To be effective, though, a really large number of them — several dozen — has to be used.

The Nature of Antenna Resistance

Resistance, defined in broad terms, is something in which power is used up — usefully or otherwise. The resistance of an antenna divides into two parts, one useful and one not. The useful part is called **radiation resistance.** The power used up in this resistance is the power actually radiated into space from the antenna. The non-useful part of the resistance is represented by losses, partly in the conductor (because of its ordinary resistance at the operating frequency), partly in insulation associated with the wire, and partly in conductors and dielectrics close enough to the antenna to be in a strong electromagnetic field. These are lumped together and often called the **ohmic** resistance. Power dissipated in ohmic resistance is turned into heat.

Since only the power used up in the radiation resistance is useful, we want the radiation resistance to be much larger than the ohmic resistance. It is the *ratio* of the former to the latter, rather than the actual values in ohms, that is of interest. We may measure different values of total resistance at different points along a given antenna, but the ratio of the two components of the resistance does not change. In other words, it does not matter where power is introduced into the antenna; the same proportion will be radiated, and the same fraction lost, in every case.

Why Impedance is Important

Since it is only the *ratio* of radiation resistance to ohmic resistance that counts, you would be justified in concluding that the actual value of resistance is unimportant. This is so in the antenna itself. But another factor must be taken into account. Somehow, r.f. power must be put into the antenna before there can be any radiation. In feeding power to the antenna the actual antenna resistance — or impedance — *is* important.

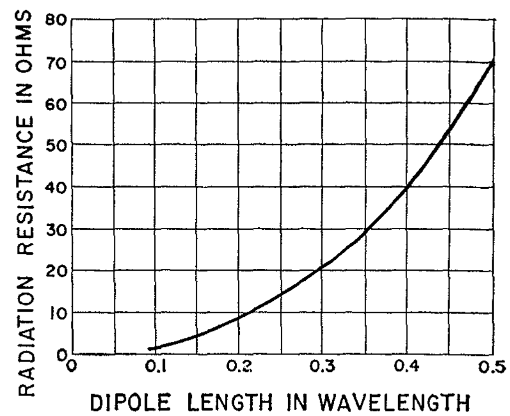


Fig. 2—Radiation resistance measured at the center of an antenna as the length of the wire is varied. Lengths here are in terms of free-space wavelength.

R.f. circuits using practical components work at best efficiency when the impedance level is between perhaps 25 and 2000 ohms. These are not exact limits by any means, but do indicate the general range. If the impedance is only an ohm or two, or is many thousands of ohms, the losses in the circuits themselves may be far greater than the power that can be delivered through them to a load. And between the plate of the transmitter's final-amplifier tube and the antenna itself there must be circuits — often several of them. Each exacts its toll of power.

The resistance of a half-wave antenna is about 70 ohms, as we have mentioned. This value is well within the optimum range for minimizing the losses in any circuits we may use to match the antenna to the final amplifier. Furthermore, it is nearly all radiation resistance. Ohmic resistance amounts to only a few per cent of the total if the antenna is mounted in a clear spot. However, the radiation resistance decreases if the antenna is shortened. For example, if a dipole is a quarter wavelength long its radiation resistance as measured at the center is only about 14 ohms, as shown in Fig. 2. If the length is shortened to one-eighth wavelength the resistance drops to around 4 ohms.

Coupling Losses

If the same power can be put into all these values of resistance, all of the power will be

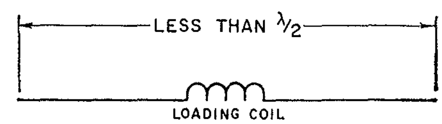
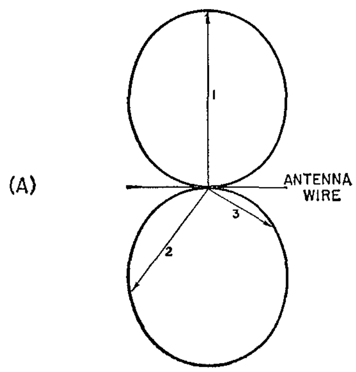


Fig. 3—Inductive "loading" of a short antenna to make it resonant. The shorter the antenna the greater the inductance required. The term loading, as used in this connection, dates from early radio times, and refers to tuning a circuit —usually by adding inductance—to a lower frequency than the one to which it is naturally resonant. The natural resonance in this case would be that of the wire without the coil.



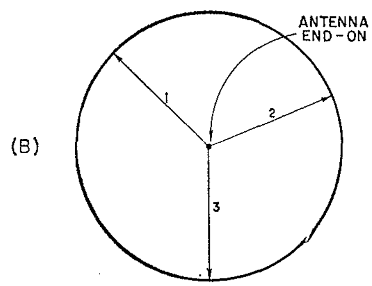


Fig. 4—Cross-sections of directional pattern of a half-wave antenna. A—in the plane in which the wire lies; B—in a plane cutting through the center of the wire at right angles to it.

radiated. However, the "if" is a big one. The half-wave antenna is resonant, and so needs no tuning. The shorter antennas are not resonant; their impedances have large amounts of reactance along with resistance. In order to put power into a short antenna the reactance has to be "tuned out," by adding the same value of reactance, but of the opposite kind, at the antenna terminals. A short antenna has capacitive reactance, so inductive reactance has to be added to cancel it, as in Fig. 3. But coils inherently have resistance, and a coil of the size needed for tuning a ⅛-wave antenna, for instance, will have more resistance than the radiation resistance of the antenna itself. As a result, more power is used up in heating the coil than is radiated by the antenna.

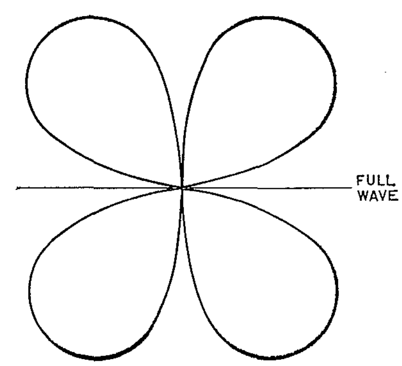
Aside from considerations such as these, there is nothing sacred about the resonant length. The antenna will radiate just as well whether or not it is resonant. However, it will not *get* all the power output of the transmitter if it is so far off resonance that the tuning apparatus uses up an appreciable portion of the power.

Beginners often take antenna resonance far more seriously than it warrants. A small departure from the resonant length is of little consequence. The resistance and reactance change rather slowly around the resonant point, so there is no observable increase in loss if the antenna isn't exactly resonant. As a matter of fact, an antenna can't be resonant at more than one single frequency. Yet it isn't by any means necessary to use different antennas for each frequency within an amateur band.

Directivity

Offhand you might think that the strength of the signal radiated from an antenna would be the same in all directions—up, down, and to all sides. It isn't. The radiation is stronger in some directions than in others. This comes about because the ends of the antenna always have opposite polarity, and because the antenna is not just a point but has a length that isn't small compared with the wavelength.

You can think of it as a case of timing, or phase. The electromagnetic field from one part of the antenna doesn't reach a distant point at the same time as the field from another part. In an extreme case, the fields reaching such a distant point may even get there with the same amplitude but *opposite* polarity. Then they add up to zero; there is no radiation in that direction. Or, in another direction, the fields may reach the distant point with the same amplitude and the *same* polarity. Being "in phase," they add together to give the strongest field the antenna is capable of producing. In still other directions, neither of these conditions is met completely, so



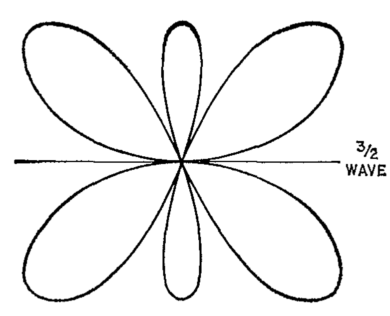


Fig. 5—Cross-sections of directional patterns of (A) a full-wave antenna and (B) one having a length of 1½ wavelengths. The cross-sections correspond to the one in Fig. 4A, in relationship to the antenna wire.

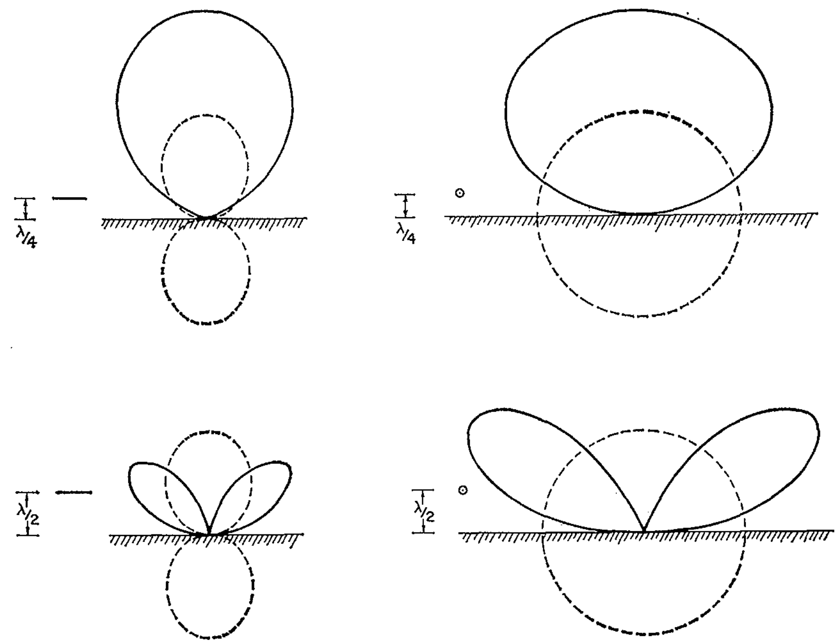


Fig. 6—Effect of the ground on the radiation from a horizontal half-wave antenna, for heights of one-fourth and one-half wavelength. Dashed lines show what the pattern would be if there were no reflection from the ground. Orientation of the antenna is shown to the left of each pattern.

the strength of the signal has an intermediate value.

Directive Patterns

This rather complex operation is summed up in what is called the **directive pattern** of the antenna. The pattern is a graph showing the relative strength of the radiation in all directions. We can't show a pattern completely on a sheet of paper, since the paper has only two dimensions, while the antenna actually radiates into all the space surrounding it. Antenna patterns usually are a "slice" or cross section of the full pattern.

Fig. 4A shows typical cross-sectional patterns for a half-wave dipole. The arrows marked 1, 2 and 3 show, by their length and direction, the relative strength of the radiated field. Don't forget that this drawing is a slice; in order to visualize the complete pattern you would have to imagine that the pattern rotates around the antenna wire, in and out of the paper, to form a doughnut with a point, not a hole, in the middle. Then when you turn the antenna on end, as in B, a slice at right angles would give you just a circle, as shown.

Taking these two patterns together, you can see that a *horizontal* half-wave antenna will radiate best directly upward and downward (if you are looking at the antenna from the side) and won't radiate at all directly off the ends. If you imagine yourself *over* the antenna in A, it radiates best at right angles to the direction of which the wire runs. On the other hand, if you are looking directly down on a *vertical* antenna, as in B, the antenna is radiating equally well in all directions. These last directions, of course, are along the ground, going around the compass.

If the antenna is shorter than a half wavelength the patterns will still have much the same shape. However, if the length is two or more half wavelengths there are rather drastic changes. Figs. 5A and 5B show, respectively, the patterns for the "full-wave" and "three half-wave" antennas whose current and voltage distribution are shown in Fig. 4, Part I. The maximum radiation is no longer broadside to the wire but goes off at an angle, as you can see by comparing these drawings with Fig. 4A. These, too, are cross-sections of a solid pattern that you can visualize by imagining the cross-section drawing to be rotating around the antenna.

The Earth's Part

Since the antenna radiates in all directions, some of the energy must go toward the ground. The earth acts more-or-less like a huge reflector for radio waves. The rays hitting it bounce off much like light rays from a mirror. These reflected rays combine with the direct rays from the antenna at a distance. The result is that the directive pattern of the antenna is modified by the presence of the earth "mirror." Just what the mirror does depends on the height of the antenna above it, and whether the antenna is horizontal or vertical.

Fig. 6 shows a couple of typical cases for a half-wave antenna. The patterns at the left show the relative radiation when you view the antenna from the side; those at the right show the radiation pattern you would "see" when you

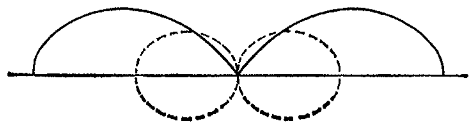


Fig. 7—Effect of the ground on radiation from a half-wave vertical antenna. In the absence of the ground, the pattern would be like the dashed line.

look at the end of the antenna. Changing the height from one-fourth to one-half wavelength makes quite a difference in the upward radiation — that is, the radiation at high angles. The **radiation angle** is measured from the ground up.

Fig. 7 shows what happens to the pattern of a vertical half-wave antenna sitting on the ground. Here the maximum radiation is along the ground.

Lest you take these pictures too seriously, we have to warn you that the ground isn't like the mirror on your wall. It's pretty foggy, as a matter of fact. In other words, it isn't by any means the perfect reflector that these pictures assume it to be. The fogginess is principally the result of energy losses; a fairly husky proportion of the wave energy striking the ground is used up in the ground resistance. The principal effect of this is that you don't get the radiation at very low angles that Fig. 7 would lead you to expect. Practically, there isn't a great deal of difference between horizontal and vertical antennas in this respect, if the horizontal is a half wavelength or more above the earth. QST

Strays

Most fields have their historians and collectors. There are antique car fans, early railroad "buffs" and now amateur radio historians. Such a group met recently for the first time, sponsored by the Antique Wireless Association, at Holcomb, New York. Representatives of many pioneer organizations attended. Among them were R. B. Bourne, W1ANA, Curator of the ARRL Museum; Frank Davis, Curator of the Ford Science Museum; James Jones of VWOA; W2ZI, Historian of the OOTC; W4ZM of the QCWA; W1NTE, Curator of the New England Wireless Museum; W2ONE, Historian, the Morse Telegraph Club; and Radio Club of America President Ralph Batcher, who delivered the meeting's keynote address. Our picture shows W1ANA addressing the group. The museum room is seen to the rear.

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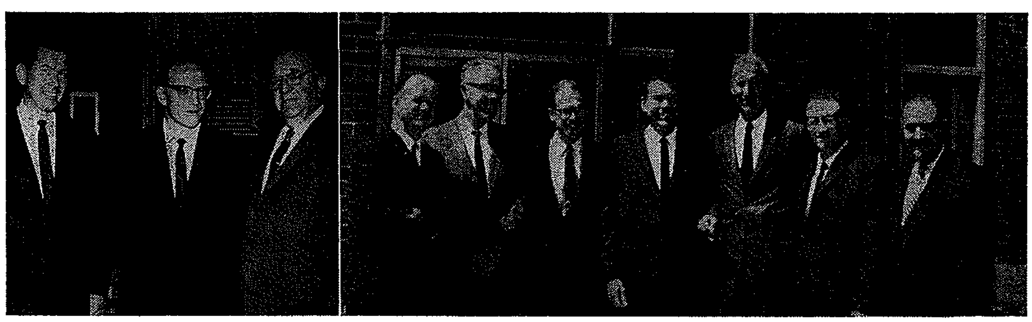
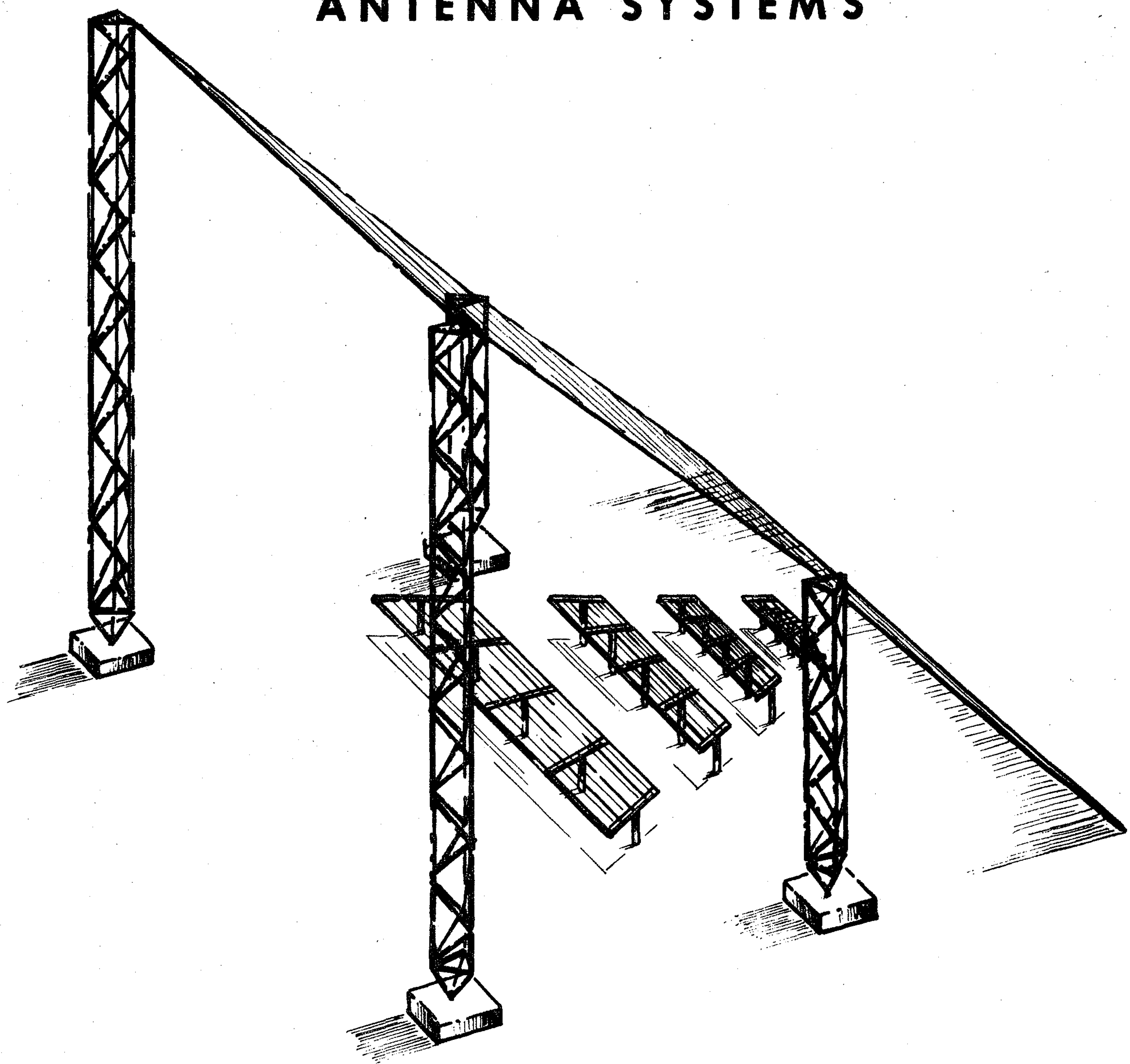


Domenico Petti, HV1CN, recently made a whirlwind week-long jaunt through the eastern U. S. He made stops at Chicago, where he was feted at the Hamfesters Radio Club picnic and the Hallicrafters labs and factory; in Detroit, where he visited the Ford Museum and an automobile assembly line; in New York; and at ARRL headquarters. Among the firsts that Dom tallied during his visit were subways, hot dogs, and mobile hamming (mobiling is not permitted in Italy). An engineer with Vatican Radio, Dom was guided through radio and TV studios of WGN in Chicago and NBC, New York.

The picture on the left shows HV1CN (center) and his Hallicrafters hosts K9EBE (left) and W9AC. The other picture was taken at ARRL's front door. Left to right are W1WPO of DXCC; W1BDI, ARRL Communications Manager; W1LVQ, ARRL General Manager; K9EBE; W1VG, QST Advertising Manager; HV1CN; and K1JMN, Dom's traveling companion and interpreter.

A COMPARISON OF HIGH FREQUENCY ANTENNA SYSTEMS



The Log Sequential Array

Continental Electronics MFG. CO.

DALLAS. TEXAS

HIGH FREQUENCY ANTENNAS and THE LOG SEQUENTIAL ARRAY

When considering antennas for high frequency broadcasting and for communications systems there are a few basic types which have been in use for many years. Continental Electronics has recently added a new type called a Log Sequential Antenna which provides considerable improvement over those now in use. A comparison of the capability and limitations of present day high frequency antennas is included as a guide to selecting the most suitable types for a proposed system.

A directional antenna is usually desired since the target area often includes a single language speaking country or the circuit is a point-to-point communications path. The directional antenna conserves energy and concentrates the total transmitter power in the desired direction or areas. The width of the beam and the resulting power gain are closely correlated in the following way:

Beam width	90	60	30	10	Degrees
Power gain	30	45	90	275	Ratio-to-one
Gain	15	17	20	24	dBI

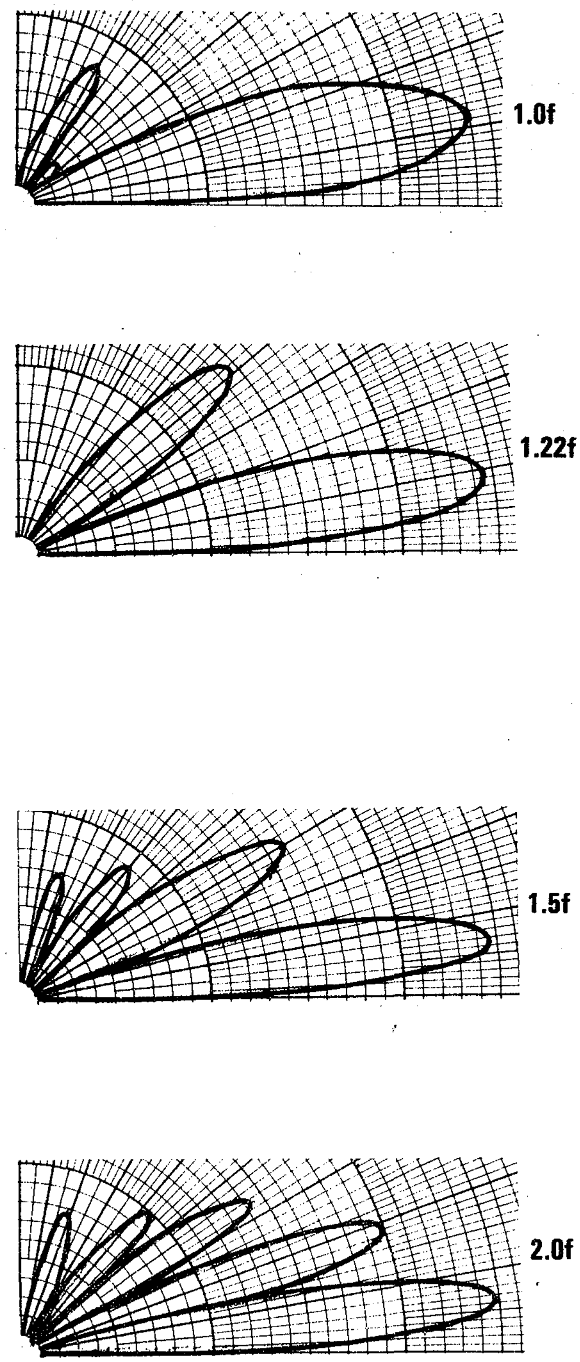
These are nominal values and vary somewhat with the vertical beam width and magnitude of side lobes.

CURTAIN ANTENNAS

The most used antenna for International Broadcasting is the 16 element curtain antenna stacked 4 high and 4 wide using half wave dipoles in front of a reflector screen. It is a good antenna to use as a standard for comparison of all other types. The principle concern is how many antennas are required to cover the frequency range from say 5 to 30 MHz, and consequently the total cost of the system.

The bandwidth of the curtain antenna is limited both by its impedance bandwidth and its radiation pattern bandwidth. Classic curtain antennas such as the Voice of America Type IIC are used over two adjacent international bands or a frequency ratio of 1.3:1. With improved dipole design, the ratio can be increased to 1.5:1. There is no point in further improvement to the impedance bandwidth as the pattern bandwidth is restricted to this range as shown in Figure 1.

As shown in Figure 1, the beam is set to a good usable low angle of ten or twelve degrees and then as the frequency is increased, the pattern becomes multilobed and its performance in gain and multipath deteriorates rapidly, beyond a frequency range of 1.5:1. Nothing further can be done to improve the pattern performance since it is a function of the physical height of the



f = lowest design frequency

Figure 1 Typical Vertical Plane Patterns For Curtains, Rhombics, and Horizontal Log Periodic Antennas array above ground. These facts are well documented in classical antenna literature. This 1.5:1. range requires a minimum of four curtains to cover from 5 to 25 MHz and preferrably five for good performance.

RHOMBIC ANTENNAS

The impedance bandwidth of a terminated Rhombic is unlimited. Its pattern bandwidth however has the same limitation as the curtain and for the same reasons. Very long Rhombics tend to control the over hanging lobes a little better than curtains and have been used over a 2:1 frequency ratio for economy. If the leg lengths are in excess of 5 wavelengths, one need not terminate the far end with a load resistor since the reflection coefficient is largely radiated before it returns to the input terminals and the impedance excursions are consequently small.

Four large Rhombics can be used to cover the frequency range with good performance but require considerable real estate.

LOG PERIODIC ANTENNAS

The log periodic antenna like the Rhombic can be designed for low SWR over an unlimited frequency range and thus the impedance bandwidth is unlimited. When we consider pattern bandwidth, several types of log periodic antennas must be designated.

A horizontal log periodic antenna has exactly the same pattern bandwidth limitation as the Curtain and Rhombic antennas and for the same reasons. It is possible to design a steeply sloping horizontal log periodic where each of its elements has constant electrical height above ground. This will produce a constant pattern over any frequency range but the pattern has one high angle lobe good for short distance circuits only. It cannot be used for long range work.

Vertically polarized log periodic antennas are being offered and are in use to overcome the pattern deficiency of horizontal log periodics. However, while the pattern bandwidth is unlimited it is not necessarily what is desired unless a very extensive ground system is used for thousands of feet out in front of the array.

Low angle summation of the incident and ground reflected waves as in-phase vectors is based upon the assumption of a perfectly conductive earth. This is approximately true at medium and low frequencies, but at five Megahertz and above, it is far from true as shown in Figure 2.

As can be seen in Figure 2, the reflection coefficient is 180 degrees out of phase at low angles for average soil conductivity from 5 to 30 MHz. Only above 20 degrees does it become reasonably close to being in-phase. However at these higher angles,

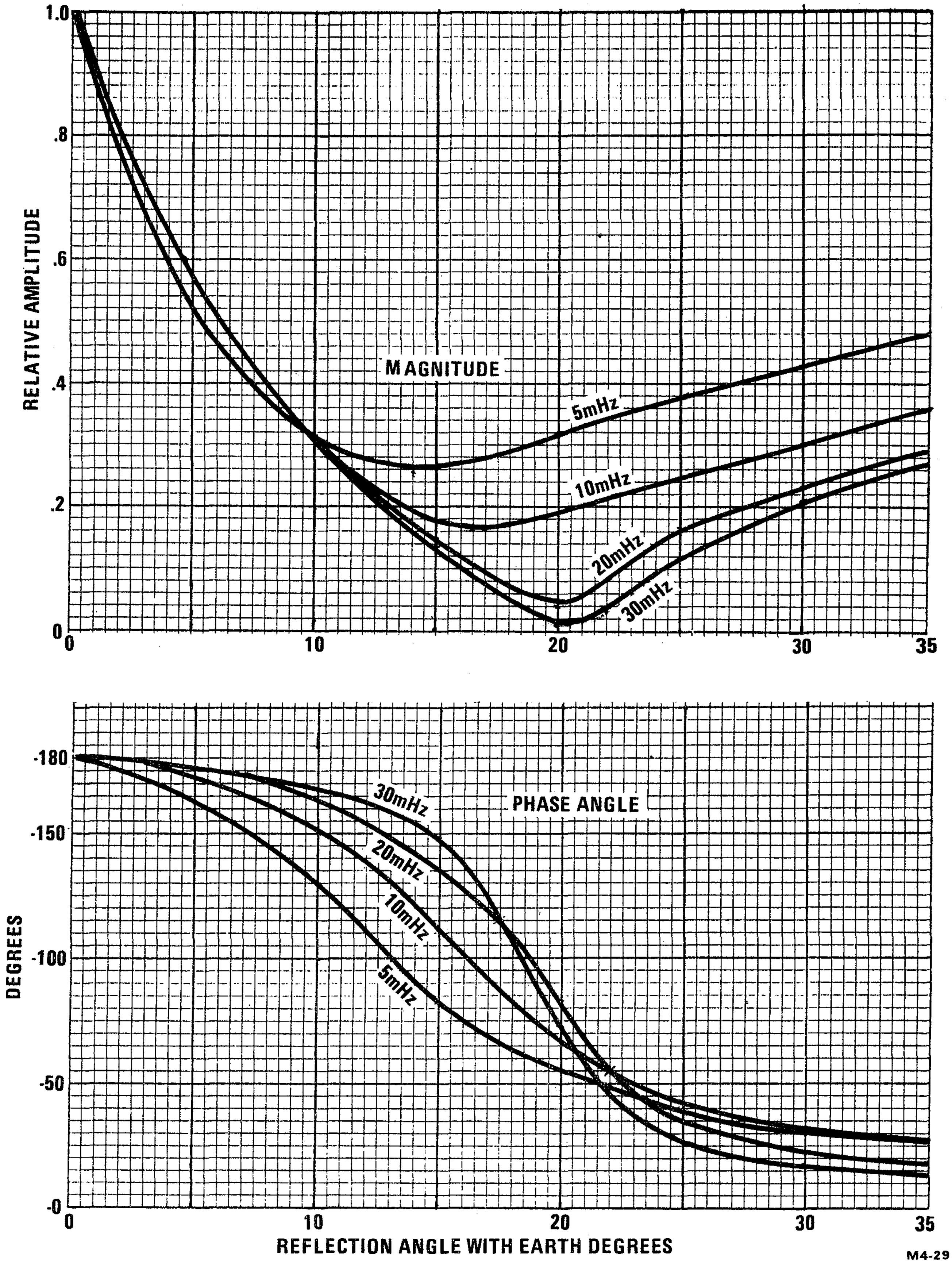


Figure 2 Reflection Coefficient Vertical Polarization Conductivity 4mmho - Dielectric Constant 8
High Frequencies

its magnitude has decreased to a small fraction of its value before the reflection due to dissipation in the soil. So then there is little or no radiation at low angles and at high angles there is up to 3 dB loss in the system, which means that one half the power is wasted in the earth.

The only solution is to provide a ground screen out through the first fresnel zone which for low angles is quite extensive. This requires a lot of copper, labor, and real estate.

LOG SEQUENTIAL ANTENNA

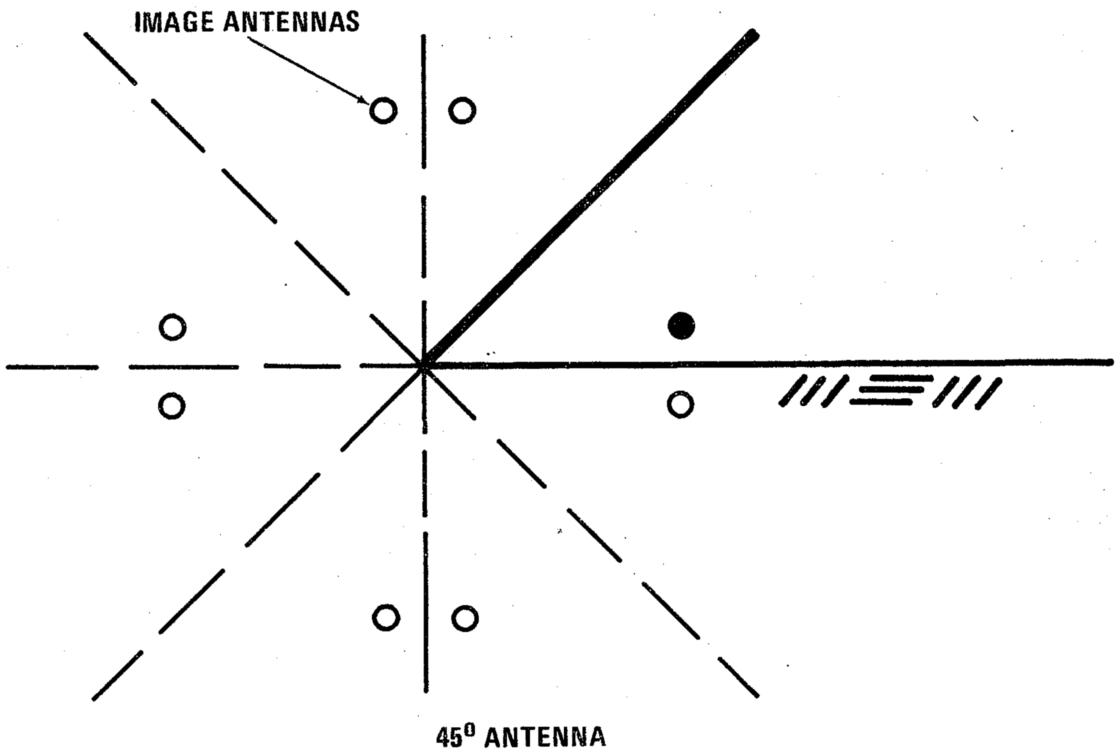
The Log Sequential Antenna has the same fundamental limitations on its subsystem parts as does a curtain antenna, but is a unique configuration which allows a large number of curtains to utilize a single reflector screen no bigger than that required for the lowest frequency of operation. By slanting the screen and using a primary ground reflection to excite it, the log sequential antenna requires only a single colinear line of feed dipoles.

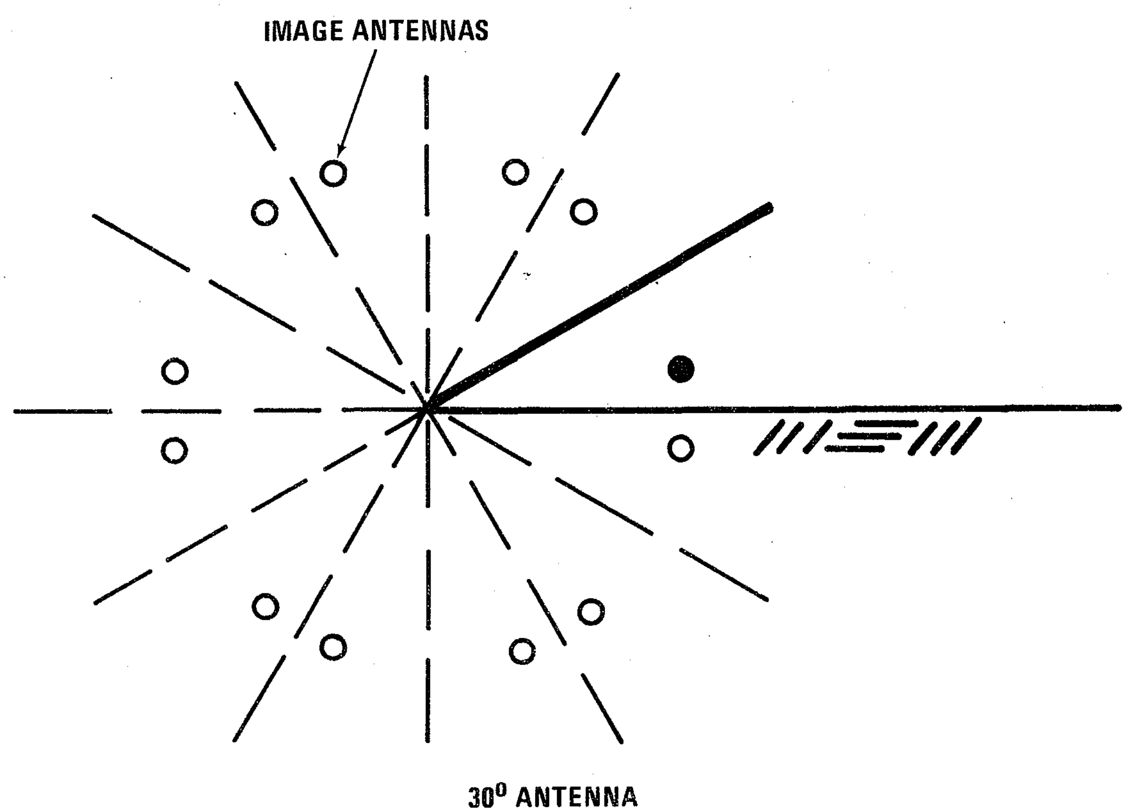
This allows a separate feed system for each band of operation within a single array without physical overlap. It is in essence the missing link between the curtain antenna and the log periodic antenna and overcomes the limitations of both.

The geometry of the Log Sequential Antenna is described completely by angles as is the case with Log Periodic arrays. This gives unlimited impedance bandwidth and it is switched sequentially for each band of operation to hold the radiation patterns constant with frequency change. It is the only known antenna which successfully covers the total frequency range in a single unit the size of an equivalent curtain antenna.

The single colinear feed is feasible because the system utilizes multiple images or reflections in the slant screen as shown in Figure 3. This eliminates over-hanging or grating lobes in the vertical pattern.

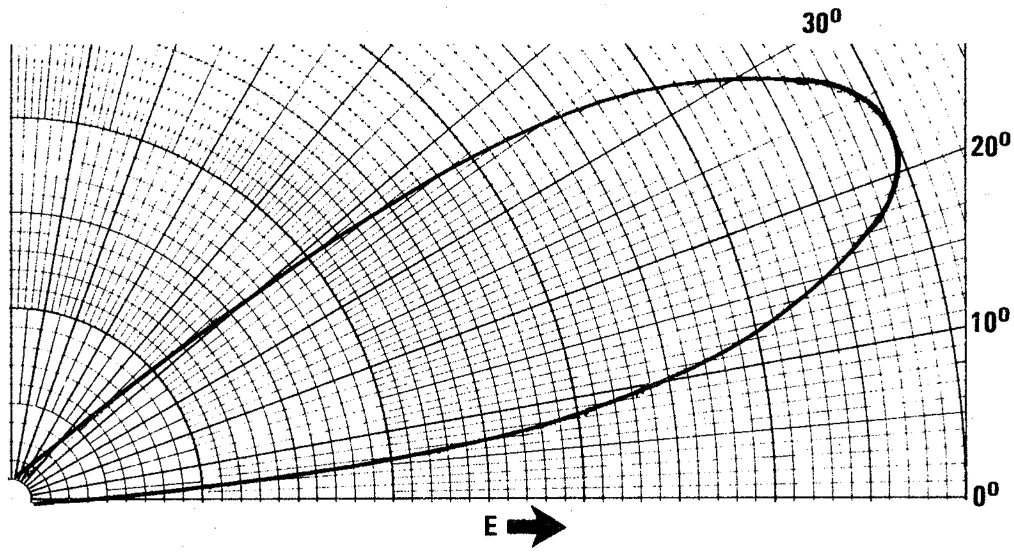
A considerable advantage is that there are no active feed elements in the superstructure. All feed components are at a height of 0.2 wavelengths above ground. In addition, the vertical angle of radiation can be steered by selecting the distance of the feed system from the apex of the screen and ground. Vertical angles from 8 to 22 degrees may be selected for each or all frequencies throughout the operating range. Typical vertical and horizontal plane patterns as shown in Figures 4, 5 and 6 can be held constant from the lowest frequency of design to any selected upper frequency limit.

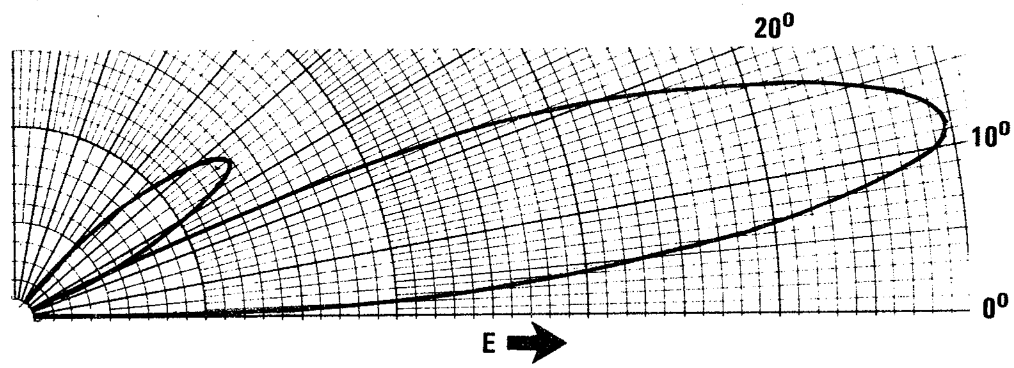




M4-30

Figure 3 Log Sequential Antennas





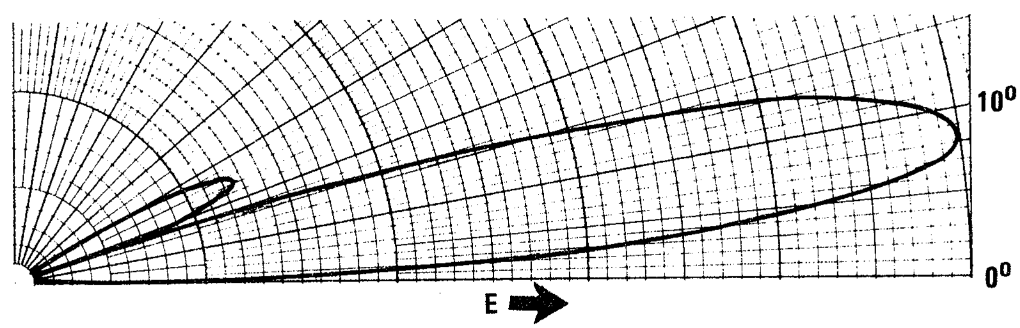
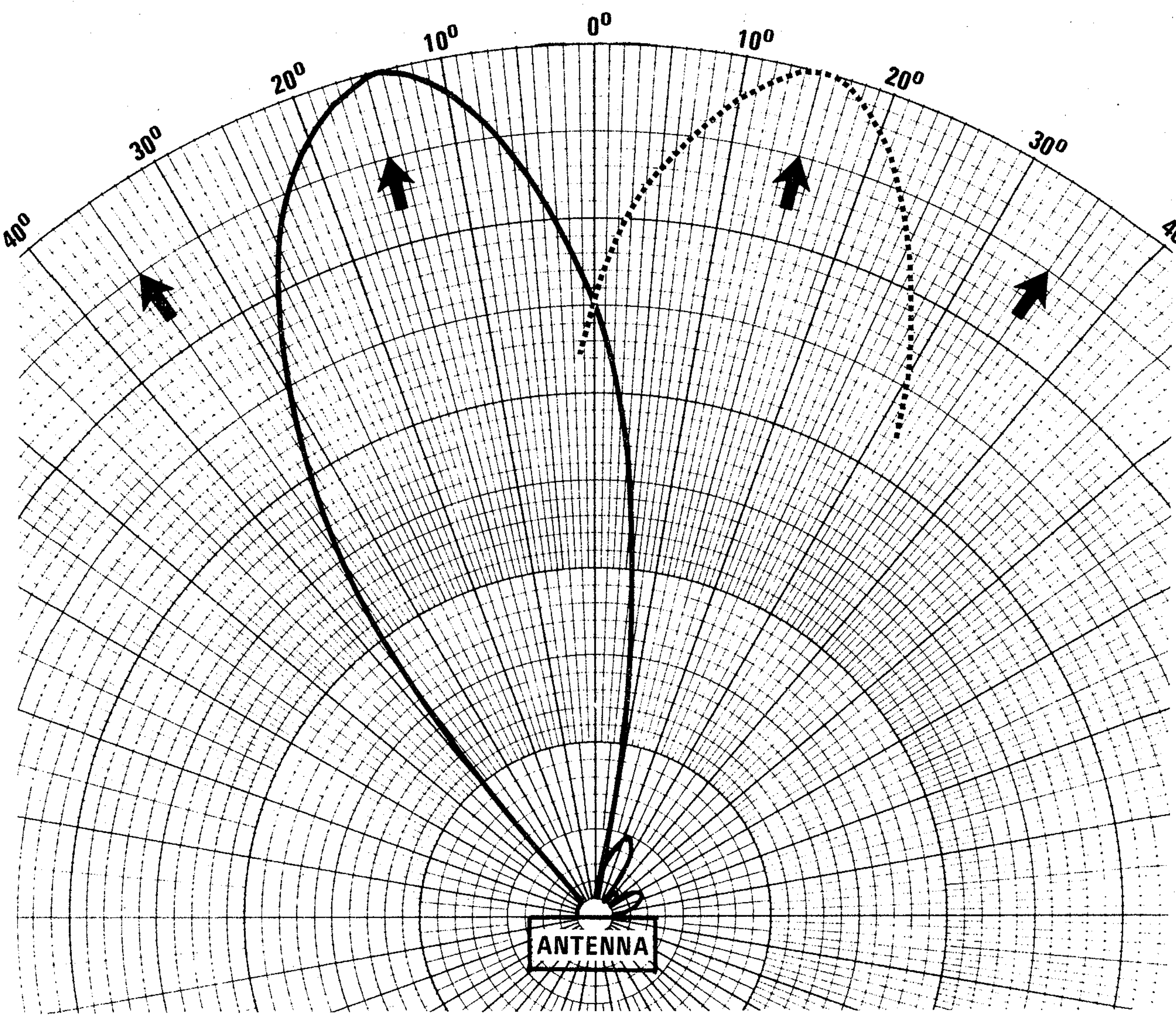


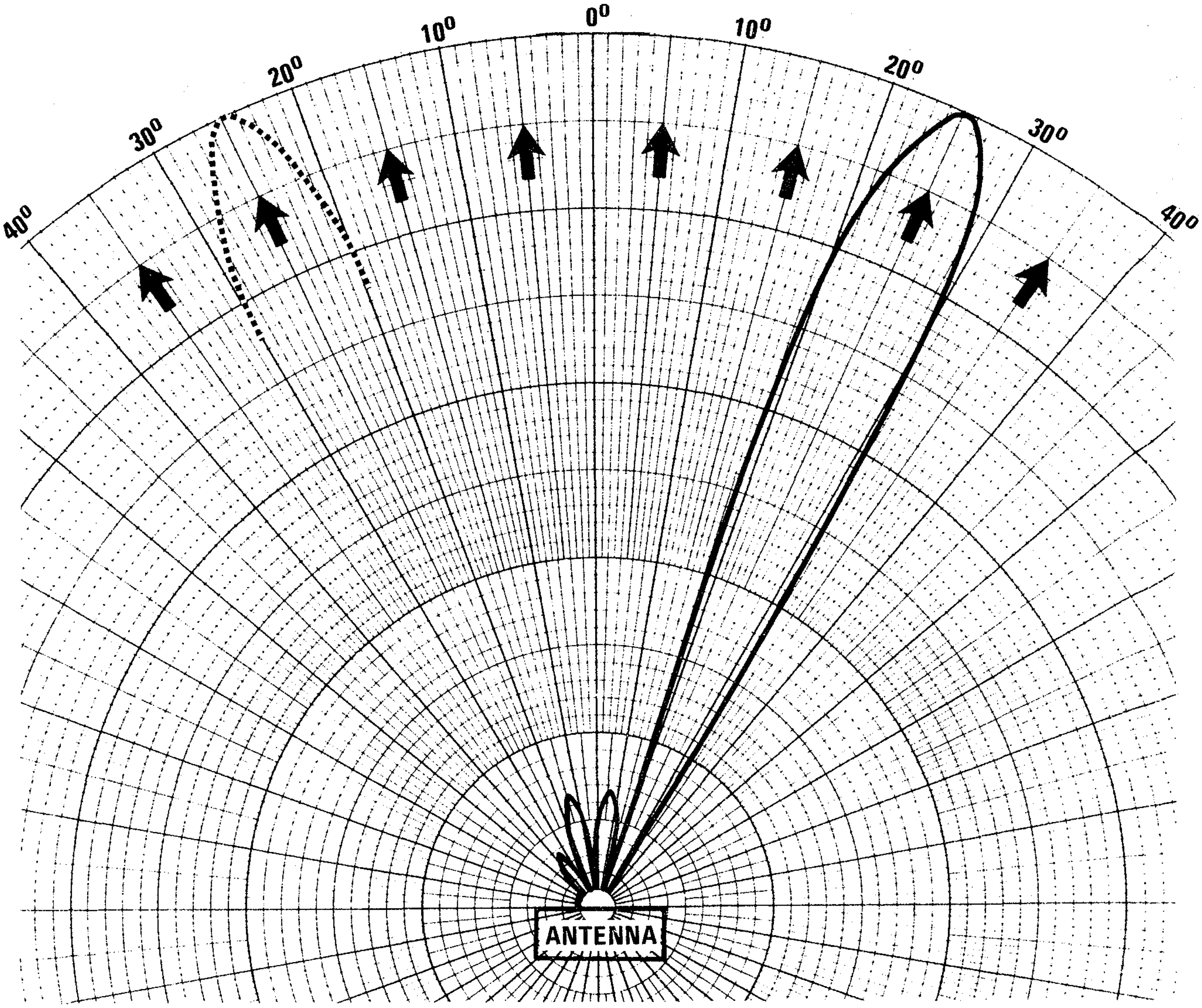
Figure 4 Available Vertical Plane Patterns Log Sequential Antenna

Continental Electronics



M5-2

Figure 5 Horizontal Plane Steerable Patterns Medium Gain Log Sequential Antenna



M5-3

Figure 6 Horizontal Plane Steerable Patterns High Gain Log Sequential Antenna

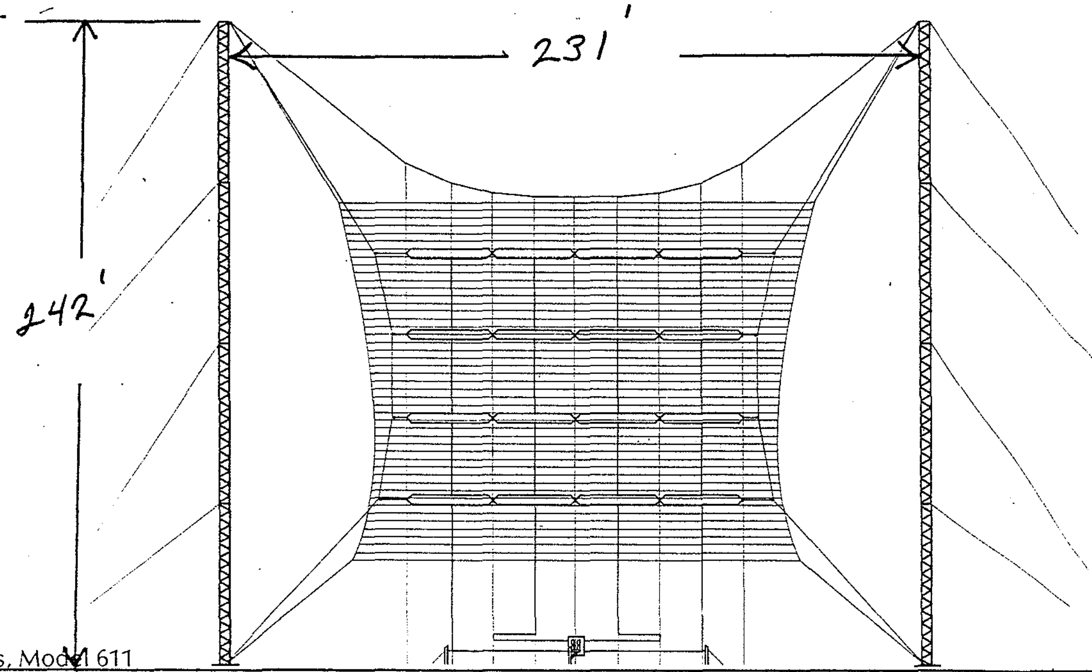
Steering the horizontal plane pattern can be achieved by phasing the feed system for curtains, log sequential and multiple element log periodic arrays. With uniform power in all elements, the curtain and log sequential antennas have equivalent steering. With tapered illumination they both can be steered to wide angles. Log periodic arrays have their sub-arrays converging inward to a point in front of the system. The sub-arrays are therefore not aligned on boresight and steering is restricted to smaller angles if side lobes are to be kept within reasonable bounds.

The log sequential antenna therefore is the only antenna with nominally constant impedance and pattern performance over the total frequency range. Its feed system is easily added-to or modified in the event of changes in systems performance requirements and without effecting the superstructure. As stated before, it collects the best features of existing antennas and is the missing link between log periodics and curtains.

ANTENNAS H.F

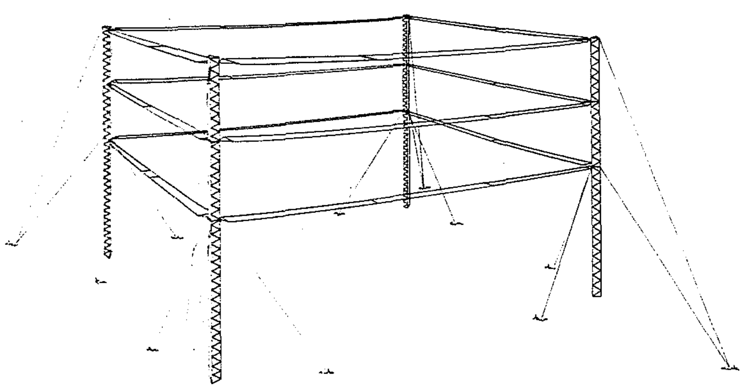
TECHNOLOGY FOR COMMUNICATIONS INTERNATIONAL **TCI**

Shortwave Broadcast Antennas



Wide Band Dipole Curtain Antennas, Model 611

TCI - 611 4x4x1 (9/11/15/17 mhz)
250KW carrier power.



Omnidirectional Quadrant Antennas, Model 605

Dipole Curtain Antennas, Model 611

Dipole curtain antennas have long been favored for high-power short-wave broadcasting. Historically, these antennas have had several undesirable traits: narrow bandwidth, extensive fabrication necessary in the field, complicated and time-consuming installation, and materials being subject to corrosion.

Wide-band techniques developed at TCI make it possible to supply dipole curtain antennas with frequency bands in excess of 2:1. In most cases, this bandwidth encompasses 4 adjacent broadcast bands. One TCI four-band antenna replaces four single frequency antennas with the associated savings in land, antenna cost, transmission lines, switching, and installation cost.

As with all TCI antennas, complete fabrication and pre-assembly are accomplished in the factory. No measuring, cutting, swaging, welding, or other manufacture is required in the field. Installation consists of only the tower erection and hoisting the preassembled curtains. The few connections required are accomplished with nuts and bolts. Installation cost and time can be reduced by at least half that of the more conventional arrays. While each dipole curtain is usually designed for a specific application, TCI manufacturing technique maximizes the commonality of parts and allows fabrication using mass production.

Like other standard TCI antennas, the TCI broadcast antennas consist of high-quality, exhaustively tested components and materials. All radiators, feedlines, and supporting catenaries are of Alumoweld, a wire composed of a high strength steel

- **High Power**
- **High-Power Gain**
- **Wide Band**
- **Slewable**
- **Rugged Construction**
- **Factory Preassembled for easy and quick installation**

core and a highly conductive, corrosion-resistant welded coating of aluminum. All insulators are made of high-strength glazed alumina or high quality porcelain. No organic or fiberglass material is used anywhere in the antenna. Dissimilar metal contacts, long a troublesome cause of corrosion, have been eliminated.

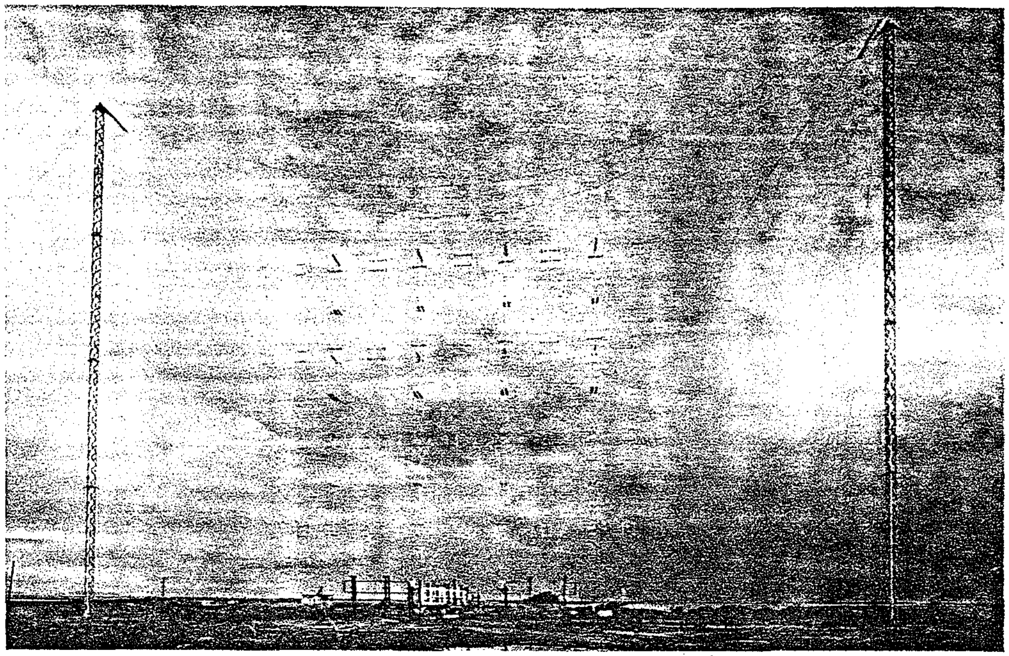
TCI has vast experience in high voltage technology, having antennas operating in excess of 1 megawatt. All components are thoroughly tested for voltage flash-over to assure safety factors well in excess of the actual voltage stress experienced.

The take-off angle is determined by the height of the lowest radiator above ground and the number of elements stacked vertically. Figure 1 provides the take-off angle for various conditions.

The gain at beam maximum is determined by the number of elements wide and high and the height above ground. Figure 2 provides power antenna gain for various antenna configurations.

The Model 611 can be supplied in any configuration dependent on the particular application. Model 611 antennas, which are reversible and slewable, are also available.

Dipole curtains for any application are available from TCI and are described by Standard Broadcast Nomenclature as follows:



H R R S 4 / 4 / 1 – Height of lowest radiator above ground in wavelengths
- Number dipoles stacked vertically (one-half wavelength spacing)
- Number dipoles (one-half wavelength) wide
- Array may be slewed
- Radiation direction is reversible
- Reflectors present
- Horizontal Polarization

System Design Tables

Figure 1. Take-Off Angle — Dipole Curtain Array with reflecting screen

Number of elements in vertical stack (m) (½ wavelength spacing)	Height above ground of lowest element in wavelengths (h)			
	0.25	0.5	0.75	1.0
1	**48°***	**30°**	**20°**	**15° & 48°****
2	**23°**	**18°**	**14°**	**11°**
3	**15°**	**12°**	**10°**	**9°**
4	**11°**	**10°**	**8°**	**7°**
5	**9°**	**8°**	**7°**	**6°**
6	**7°**	**7°**	**6°**	**5°**

*90° w/o Reflector
**Two Lobes Present

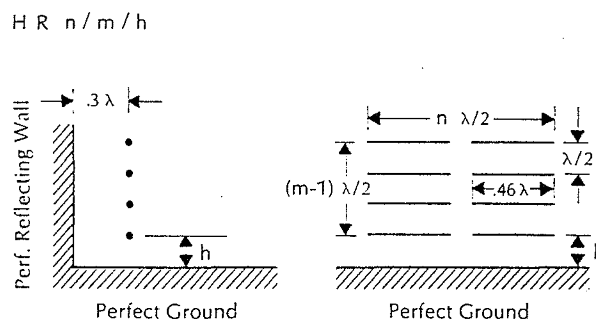


Figure 2. Gain in dBi of Dipole Curtain Array with perfect reflecting screen spaced 0.3 wavelengths from elements

Number of elements in vertical stack (m) (½ wavelength spacing)	Number of half wave elements wide (n)															
	1				2				3				4			
	Height above ground of lowest element in wavelengths (h)															
	0.25	0.50	0.75	1.0	0.25	0.50	0.75	1.0	0.25	0.50	0.75	1.0	0.25	0.50	0.75	1.0
1	**12.6**	**12.7**	**13.3**	**12.7**	**13.8**	**13.9**	**14.9**	**14.3**	**14.4**	**15.1**	**15.6**	**15.0**	**15.6**	**16.3**	**16.8**	**16.2**
2	**13.3**	**14.5**	**14.9**	**15.0**	**15.1**	**16.4**	**16.8**	**16.9**	**16.0**	**17.3**	**17.7**	**17.8**	**17.2**	**18.5**	**18.9**	**19.0**
3	**14.9**	**15.8**	**16.4**	**16.5**	**16.8**	**17.7**	**18.4**	**18.4**	**17.7**	**18.7**	**19.4**	**19.4**	**18.9**	**19.9**	**20.6**	**20.6**
4	**15.8**	**16.7**	**17.2**	**17.5**	**17.7**	**18.7**	**19.2**	**19.5**	**18.6**	**19.7**	**20.2**	**20.5**	**19.8**	**20.9**	**21.4**	**21.7**

Figure 3. Azimuth Beamwidth*

Number of elements in vertical stack (m) (½ wavelength spacing)	Number of half wave elements wide (n)		
	1	2	4
1	**85°**	**54°**	**28°**
2	**85°**	**50°**	**27°**
3	**80°**	**46°**	**26°**
4	**78°**	**44°**	**24°**

*between half power points

SPECIFICATIONS

TCI Model 611 Dipole Curtain

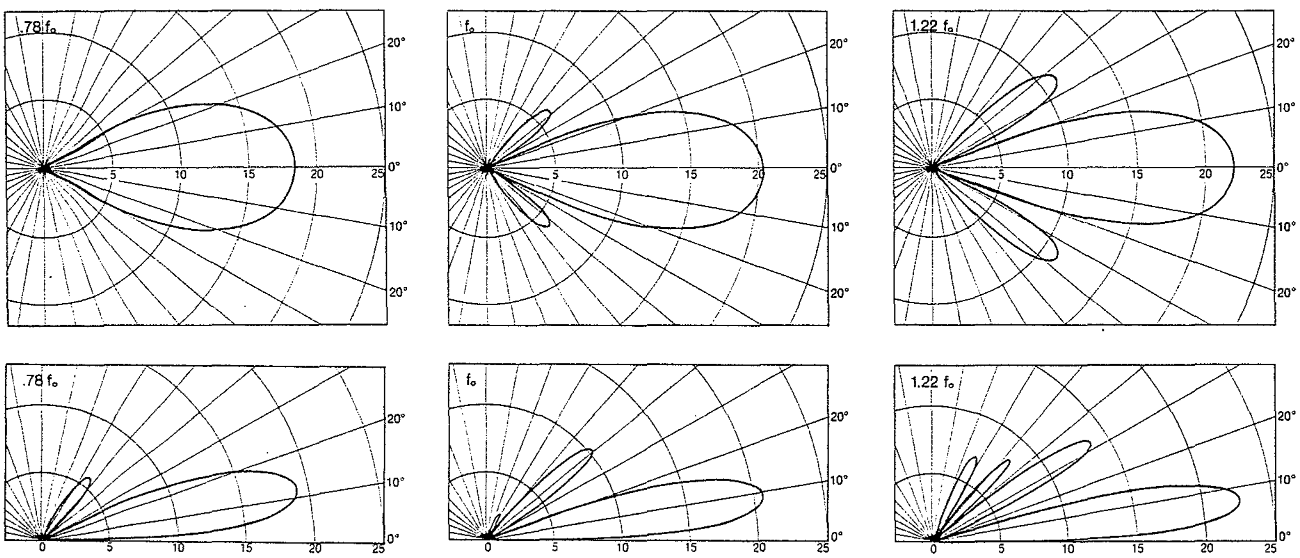
Polarization	**Horizontal**
Frequency	**6-26 MHz, one, two, three and four band antennas available**
Input Impedance	**300 ohms balanced (others available on request)**
VSWR	**1.5:1 Nominal (HRS 4/4)**
Power	**Up to 500 kw carrier with 100% modulation**
Slew	**±30°**
Other Options	**Reversing switches available**
Environmental Performance	**100 MPH wind (higher environmental capability available upon request)**
Gain	**See Fig. 2 for gain figures of various options**
Radiation Patterns	**Take-off angle is shown in Fig. 1. The radiation patterns for an HR 4/4/ three-band array at frequencies .78 f_O, f_O, and 1.22 f_O are shown below. f_O is the center frequency.**

Application Engineering

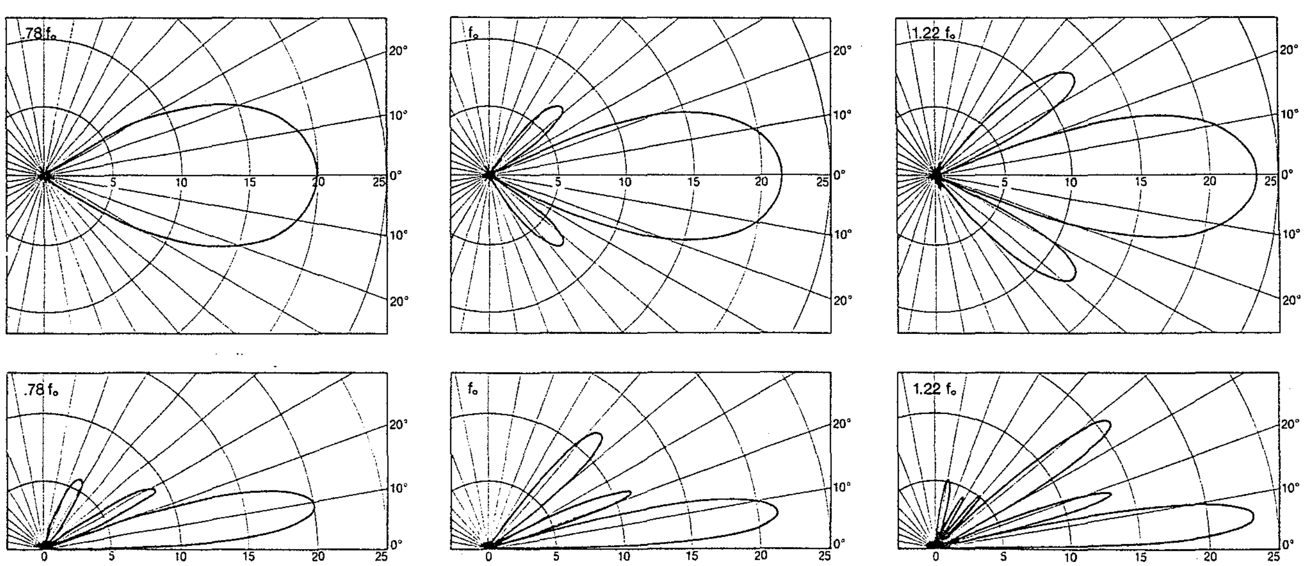
TCI has developed and demonstrated expertise in high-power short-wave broadcast antennas. The 515, 516 and 528 log-periodic antennas (see data sheet) have wide application for short-wave broadcasting and can be used in many projects. In those instances where the particular application requires a special solution, TCI can offer related antennas such as our curtain arrays or linear arrays and accessories such as compensated high-power open-wire transmission line assemblies. For assistance with your specific application, please contact your nearest TCI Engineering Office.

Model 611 ELEVATION AND AZIMUTH PATTERNS (over perfect earth) gain in dBi

H R 4/4/0.5



H R 4/4/1.0



Omnidirectional Quadrant Antennas, Model 605

The TCI Model 605 Quadrant Antenna is well suited for medium distance, omnidirectional broadcast applications. The quadrant antenna is two orthogonal radiators whose apex and feed is at the supporting mast. Four individual antennas tuned to different frequency bands can be supported by four towers. A separate 600 ohm (300 ohm if required) input is provided for each antenna. The take-off angle is a function of the height above ground and is consistent with the data in Fig. 1.

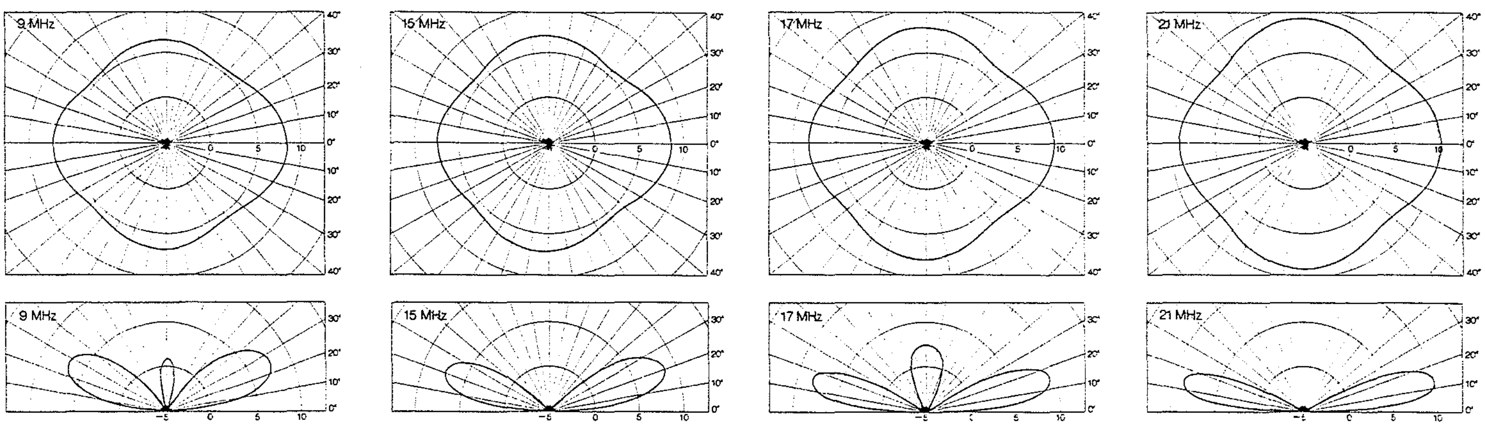
Additional power gain may be achieved by stacking the quadrant antennas. The 605 antenna illustrated is a 4-band antenna tuned for 9, 15, 17, and 21 MHz with the upper three bands consisting of two stacked arrays. The specifications for this particular antenna are:

Freq.	TOA	Gain at Beam Max.
9 MHz	27°	8 dBi
15 MHz	22°	8.5 dBi
17 MHz	18°	9 dBi
21 MHz	15°	10 dBi

SPECIFICATIONS

Frequency	**Any four separate broadcast frequency bands**
Input Impedance	**600 ohms balanced (300 ohm, 50 ohm and others available on request)**
Polarization	**Horizontal**
Radiation Pattern	
Azimuth	**Essentially omnidirectional**
Elevation	**Consistent with Fig. 1**
Gain at Beam Maximum	**8-10 dBi depending on the number of stacked elements and height above ground.**
Power Handling Capability	**Up to 500 kw carrier with 100% modulation**
VSWR	**1.4:1 maximum**
Environmental Performance	**140 MPH wind (higher environmental capability available upon request)**

ELEVATION AND AZIMUTH PATTERNS (over perfect earth) gain in dBi



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Design Data for

By H. W. HASENBECK

Radio Section, Engineering Laboratory
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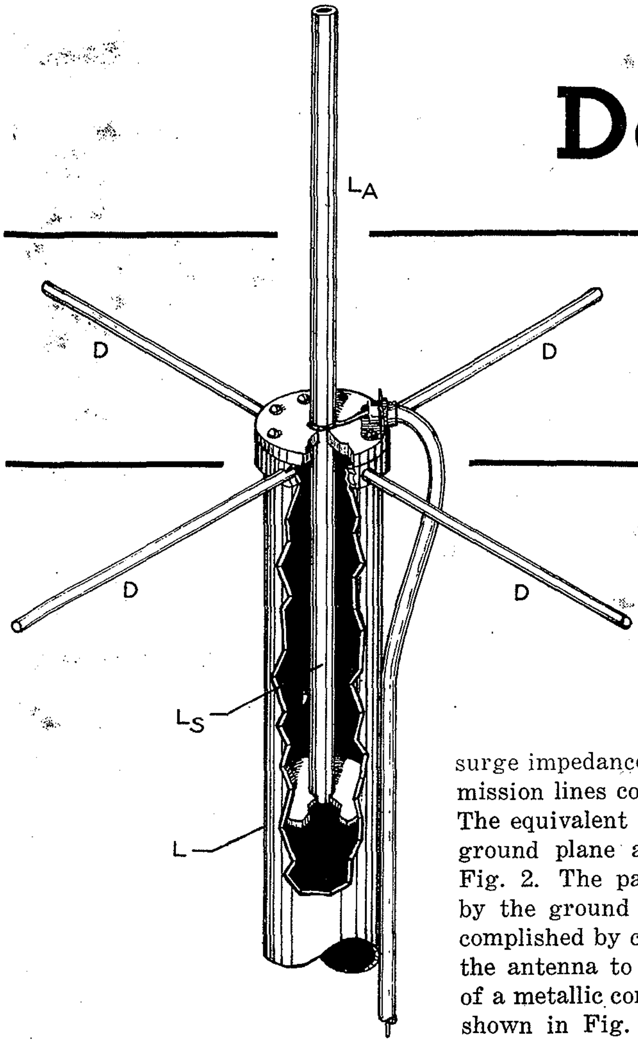

FIG. 1 — Constructional details of ground plane antenna. *L* may be any convenient length provided it is at least as long as L_s

A VERTICAL antenna located several wavelengths above ground does not display the low angle of radiation expected unless certain features are included as an integral part of the installation.

One type of high-frequency antenna designed to produce a low vertical angle of radiation is the so-called "ground plane." Its design was conceived by Brown and Epstein,[1] and is unique in that its features not only produce the desired low angle of radiation, but also allow for proper termination of the transmission line and place the antenna at ground potential. Figure 1 illustrates the mechanical construction.

Basically, the antenna assembly is so designed that a parallel circuit is formed, whose impedance can be made to match the characteristic surge impedance of concentric transmission lines commercially available. The equivalent electrical circuit of a ground plane antenna is shown in Fig. 2. The parallel circuit formed by the ground plane antenna is accomplished by connecting the base of the antenna to the center conductor of a metallic concentric base support, shown in Fig. 1 as *L*. This center conductor, shown as L_s in Fig. 1, acts as an inductance, and is shunted at the top by the antenna base impedance, composed of capacitive reactance and radiation resistance. If the inductance, capacitance and resistance are of the proper values, the parallel circuit thus formed will act as a pure resistance at the operating frequency.

As illustrated in Fig. 1, the outer conductor of the coaxial transmission line is connected to the outer conductor of the antenna base support. Since the inner conductor is common to the outer conductor, due to a metallic shorting disc at the base of L_s, the antenna assembly may be placed at ground potential, providing the antenna supporting structure or the outer sheath of the coaxial line is grounded.

Design Data

Consider the circuit shown in Fig. 2. If the values of reactance and resistance are properly proportioned, the circuit can be made to act as a pure resistance, practically equal to the surge impedance of coaxial transmission lines in common use. Proof of this fact is demonstrated by

$$\frac{j53.3\,(25 - j36)}{25 + j53.3 - j36} = 77 \text{ ohms} \qquad (1)$$

The term $j53.3$ represents the reactance of L_s in Fig. 2, and the term $(25 - j36)$ represents the resistance and reactance of L_a. The combined branches of the circuit equal 77 ohms and form the termination of the coaxial transmission line.

The four arms extending out from the top end of the outer sleeve of the concentric base support, labelled *D* are ground radials. These radials are necessary to lower the radiation resistance and decrease the effect of high-angle interference radiation originating on the supporting structure or coaxial feed line.

The terms found on the left side of Eq. (1) are entirely possible to obtain in actual practice, and the resulting parallel impedance should be satisfactory for terminating coaxial lines having a characteristic surge impedance between 60 and 90 ohms.

It can be shown mathematically that the base impedance of a quarter-wave antenna, having four radi-

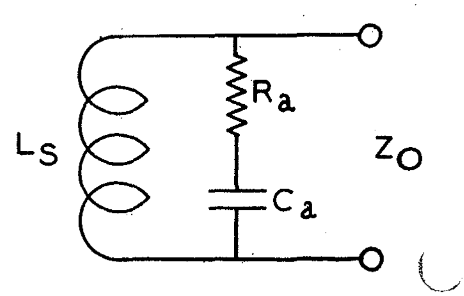

FIG. 2—Equivalent electrical circuit of ground plane antenna

Ground Plane Antennas

The addition of a turnstile element to a high-frequency vertical antenna lowers the angle of radiation. This paper develops the basic formulas for designing such a ground plane antenna. An example is worked out for a 78-ohm termination at 33.78 Mc

als extending from its base, is approximately 30 ohms. Since Eq. (1) indicates that the antenna must offer a capacitive reactance at its base equal to $-j36$ ohms, the antenna must be made slightly shorter than a quarter-wave. Obviously, if the antenna is shortened, the radiation resistance will decrease. Measurements indicate that when the antenna offers $-j36$ at the base, the radiation resistance decreases to 25 ohms as required by Eq. (1).

The $j53.3$ term of Eq. (1) is easily obtainable by properly proportioning the length of the concentric base supporting structure. Once the surge impedance of this section has been determined, the reactance of the center conductor can be calculated quite accurately by

$$jX = Z_o \tan Y \qquad (2)$$

Since jX must equal 53.3,

$$\tan Y = 53.3/Z_o \qquad (3)$$

Z_o may be calculated once the material has been chosen from which the antenna base support is to be constructed.

$$Z_o = 138.15 \log b/a \qquad (4)$$

where b is the inside diameter of the outer conductor and a is the outside diameter of the inner conductor.

The number of electrical degrees which the tangent value represents, as calculated by Eq. (3), may be obtained from a table of trigonometric functions. The length in inches which represents one electrical degree at the operating frequency f in Mc is

$$L = 32.8/f \qquad (5)$$

The total length of the concentric antenna base supporting section L_s (as measured from the top of the shorting disc to the top of the outer conductor) is

$$L_s = LY \qquad (6)$$

where L is the length of one electrical degree in inches, and Y is the number of electrical degrees necessary to form 53.3 ohms reactance.

Each of the ground radials should be one quarter-wave in length, and may be calculated by

$$d = 85.5\, L \qquad (7)$$

For those who do not have the necessary equipment to measure the characteristics at the base of the antenna so as to determine when the antenna length is properly adjusted to fulfill the term (25—$j36$) given by Eq. (1), the following equations may be of interest:

$$\pm jX = Z_{oa} \tan \left\{ (2\pi h/\lambda) + (1.5\pi) + 0.01 \left[\sqrt{2.86\,(\log \lambda/a) - 4.6} + 3.4\right] \right\} \qquad (8)$$

where

$\pm jX$ is reactance of antenna
Z_{oa} is characteristic impedance of antenna
h is length of antenna
λ is electrical length of one cycle at the operating frequency
a is radius of antenna element

$$Z_{oa} = K_1 + K_2$$
$$K_1 = 138.15 \log h/a$$
$$K_2 = -(60 + 69 \log 2h/\lambda)$$

The reactance of the antenna can be calculated quite accurately by the use of Eq. (8). Since the length of h must be assumed in solving for Z_{oa}, several trial values may be necessary before the exact length is ascertained. Previous calculations indicate that the length of h is usually between 87 percent and 90 percent of the true 90 deg. electrical length. Several values of h should be chosen between these limits and the reactance calculated. The values should then be plotted as a function of h to obtain the exact length for a reactance of $-j36$ ohms at its base.

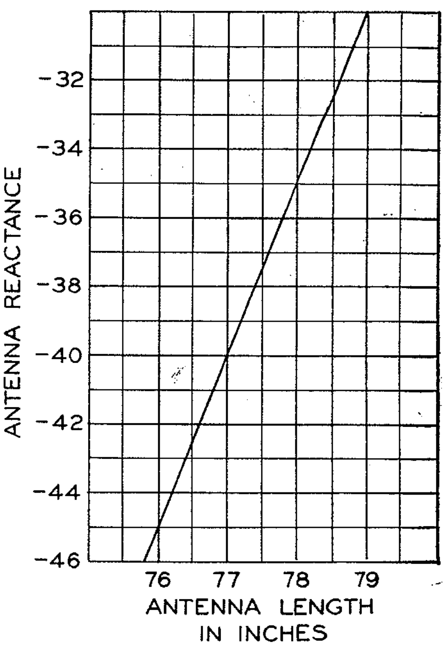

FIG. 3—Reactance as a function of length h for the hypothetical antenna design given in the text

Example of Calculations

A practical example may help in making the included data more easily followed. Assume it is desired to design a ground plane antenna to operate on a frequency of 33.780 Mc. The characteristic impedance of the transmission line is 78 ohms. The material available for the antenna base support is a section of copper tubing with an inside diameter of 2 inches and a center conductor having an outside diameter of 0.625 inches. The antenna section is to be a continuation of the center conductor of the antenna base support.

If the various sections of the an-

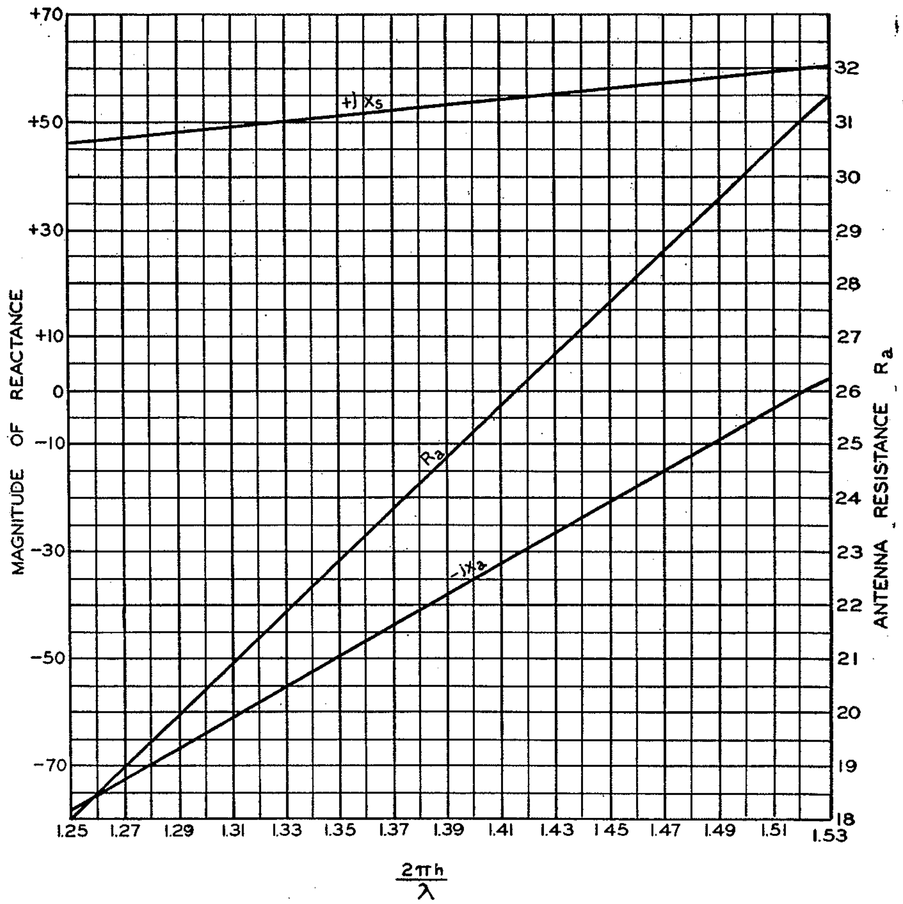

FIG. 4—Antenna reactance, antenna resistance and antenna base reactance as a function of $2\pi h/\lambda$

NOTE: The frequency range covered by the curves in Fig. 4 and 5 is approximately 30 to 37 Mc, with h being the correct value of 77.8 inches for 33.78 Mc

tenna system satisfy the terms given on the left side of Eq. (1), the 78-ohm transmission line should be satisfactorily terminated.

Since the antenna base support L_s is to offer the required 53.3 ohms reactance to the base of the antenna, the length of this section must be calculated. First, it is necessary to calculate the Z_o of this section by the use of Eq. (4). $Z_o = 138.15 \log 2/0.625 = 70$ ohms.

Substituting known values in Eq. (3), we get $53.3 = 70 \tan Y$, or $\tan Y = 53.3/70 = 0.76143$. Then $\tan^{-1} 0.76143 = Y = 37° 17'$.

From Eq. (5) the length of one electrical degree may be calculated: $L = 32.8/33.780 = 0.973$ inches. From Eq. (6), $L_s = 37.2833° \times 0.973 = 36.28$ in.

The next operation solves for the length of the ground radials. The length is measured from the outer wall of the outer conductor of the concentric antenna base support to the tip of the radial: $d = 85.5 \times 0.973 = 93$ in.

The antenna must offer $-j36$ at its base. The proper length may be calculated by the use of Eq. (8) and (9). As mentioned, the length usually will be between 87 percent and 90 percent of the true 90 deg. electrical length. The 90 deg. length is $90°L = 90° \times 0.973 = 87.5$ in. The correct length of h should be between 0.87×87.5 in. and 0.9×87.5 in. or between 76 in. and 78.8 in.

If the reactance of the antenna is calculated for the lengths 76 and 78.8 in. and these reactive values plotted as a function of h, the proper value may be ascertained.

The characteristic impedance of the 76 in. element is $Z_{oa} = K_1 + K_2$.

$$K_1 = 138.15 \log 76/0.3125 = 330$$
$$K_2 = -(60 + 69 \log 152/350) = -35.7$$
$$Z_{oa} = 330 - 35.7 = 294.3 \text{ ohms.}$$

The characteristic impedance of the 78.8 in. element is

$$K_1 = 138.15 \log 78.8/0.3125 = 331$$
$$K_2 = -(60 + 69 \log 157.6/350) = -36$$
$$Z_{oa} = 331 - 36 = 295 \text{ ohms}$$

From Eq. (8), the reactance of the 76 in. element is

$$\pm jX = 294.3 \tan \{6.28 \times 76/350 + 4.71 + 0.01 [\sqrt{2.86} (\log 350/0.3125) - 4.6 + 3.4]\}$$
$$\pm jX = 294.3 \tan 6.073 + 0.0555$$
$$= 294.3 \tan 6.1285$$
$$\pm jX = 294.3 \tan 351.3 \text{ deg.}$$

Since 351.3 deg. is in the fourth quadrant, the tangent value is negative and the reactive component of the antenna will be $-jX$.

$$-jX = 294.3 \tan 8.7 \text{ deg.}$$
$$= 294.3 \times 0.153 = 45 \text{ ohms}$$

The reactance of the 78.8 in. element is

$$\pm jX = 294.3 \tan \{6.28 \times 78.8/350 + 4.71 + 0.01 [\sqrt{2.86} (\log 350/0.3125) - 4.6 + 3.4]\}$$
$$\pm jX = 295 \tan 6.1785$$
$$\pm jX = 295 \tan 354 \text{ deg.}$$

Since 354 deg. is also in the fourth quadrant the tangent value is negative and the reactive component is $-jX$.

$$-jX = 295 \tan 6 \text{ deg.} = 295 \times .1051$$
$$= 31 \text{ ohms}$$

When these two values are plotted as a function of h as shown in Fig. 3, the correct value of h to produce $-j36$ ohms is 77.8 in.

Thus, to construct a ground plane antenna to operate on a frequency of 33.78 Mc, which will properly terminate a 78-ohm transmission line, the antenna must be ⅝ in. diameter and 77.8 in. long. The antenna base support must have a surge impedance of 70 ohms and a length of 36.28 in. The lengths of the four ground radials, extending out at right angles from the top of the antenna base support, must be 83 in. each.

The curves in Fig. 4 and 5 illus-

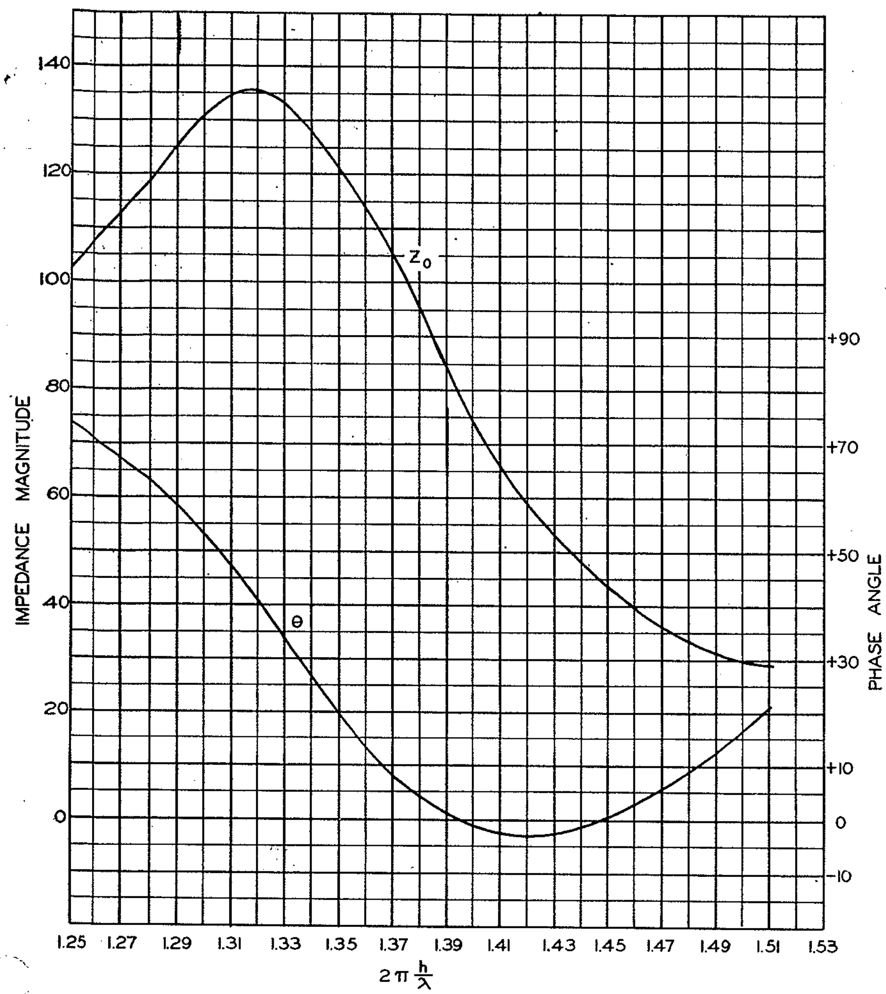

FIG. 5—Antenna terminating impedance and its phase angle as a function of $2\pi h/\lambda$

trate the variation of reactive and resistive component parts of the antenna when the operating frequency is varied from the proper operating frequency of 33.78 Mc. The frequency range represented by both curves is approximately 30 to 37 Mc. Figure 5 combines the reactive and resistive components to illustrate the variation in magnitude of the antenna terminating impedance and its phase angle.

The author wishes to acknowledge the cooperation of the technical staff of the Radio Section of the Electrical Division of the City of San Diego in aiding with construction and measurements of the antenna described in this paper.

APPENDIX

The constant 32.8 in Eq. (5) was derived from

$$\lambda_{meters} = 3 \times 10^8/f_{cycles} = 3 \times 10^2/f_{Mc}$$
$$\lambda_{inches} = 39.37 \times 3 \times 10^2/f_{Mc}$$
$$L = 39.37 \times 3 \times 10^2/360 f_{Mc} = 32.8/f_{Mc}$$

Measurements indicate that the length of the ground radials is not critical providing they are equal to 90 deg. electrical length or longer. The constant 85.5 when multiplied by L gives the approximate 90 deg. electrical length, after allowance has been made for "end-effect". (85.5 is 95 percent of 90 deg.)

In computing the resistance and reactance at the base of the antenna, the antenna was considered as an open-circuited transmission line. The methods used in obtaining the characteristic impedance Z_{oa} of the antenna were developed by Morrison and Smith.[2] (For a complete derivation, refer to p. 693 of their paper.[2]) The accuracy of Morrison and Smith's methods of calculation seems to be substantiated by the correlation between the reactance values calculated by the use of Eq. (8) and the results obtained by the more rigorous formula given by King and Blake.[3]

By including the 1.5 π term in Eq. (8), a rotation of 270 deg. is produced, making it possible to multiply the Z_{oa} term by the tangent value without regard to the antenna length, up to $\lambda/2$.

The square root term in Eq. (8) corrects for "end effect" which is influenced by the relationship between the antenna radius and operating wave length.

Over the range of $2\pi h/\lambda$ under consideration for this investigation (1.25 to 1.53), the reactance values calculated by Eq. (8) are in complete agreement with those calculated by the formula given by King and Blake.[3]

The values of resistance used in formulating curve R_a in Fig. 4 and calculation to obtain the impedance and phase angle curves of Fig. 5 were derived from the data given by King and Blake.

Several methods of calculating the antenna resistance were investigated, and the results indicate that the values obtained are not in perfect agreement. The formula cited by Brown and Epstein gives a resistance of 21.159 ohms at $\lambda/4$, while the transmission line formula given by Morrison and Smith yields a resistance of approximately 36 ohms. The rigorous method of King and Blake gives a value of approximately 32.5 ohms.

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(2) Morrison, J. F., and Smith, P. H., The Shunt-Excited Antenna, *Proc. I.R.E.*, **25**, p. 691–695, June, 1937.

(3) King, R., and Blake, Jr., F. G., The Self-Impedance of a Symmetrical Antenna, *Proc. I.R.E.*, **30**, p. 335–349, 1942.

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Hula-Hoop Antennas:

When antenna height is reduced by loading, efficiency deteriorates rapidly. Here is a system of antenna height-reduction where circumferential aperture is substituted for antenna portions lost

TODAY, when the military needs reliable world-wide communications, long waves have again assumed great importance. This trend has revived interest in the electrically short antenna.

While important for long-wavelength applications, the short-height antenna is also valuable in mobile and portable communications systems in the high-frequency bands where the vertical extent of a resonant quarter-wave antenna is mechanically impractical.

SHORT ANTENNA—A naturally resonant, vertical antenna such as the grounded quarterwave radiator is a colinear aperture. Its properties are ideal for general communications, providing an omnidirectional radiation pattern with most signals delivered at low angles. At full height, its radiation resistance is far greater than any electrical-loss resistance in the conductors. Even when operated over soil, the vertical's characteristics permit excellent radiation efficiency, even with simple wire radial ground networks.

When the height of this classical antenna is sharply reduced and electrical resonance restored by a conjugate reactor, its performance deteriorates severely. Fortunately, excellent theoretical contributions[1] on reduced electrical size antennas and modern supergain theory[2] show the reason for this effect. Reduction of physical height removes a portion of the colinear aperture of the vertical antenna. Loss of colinear aperture means loss of radiation resistance, resulting in less input power coupled to space. Loss in wire ground planes is severe in low-frequency radiator use.

When dealing purely with a colinear aperture, loss of electrical height means less efficiency.

DDRR ANTENNA—In the directional-discontinuity ring-radiator (DDRR) antenna,[3] circumferential aperture is substituted for the colinear portion lost in height reduction. Using normalized dimensions and, assuming that an antenna is desired whose height is 2.5 electrical degrees at the operating wavelength, it is specified that no lumped inductive elements be used to achieve electrical resonance to reduce electrical loss. Radiation efficiency must be within 2 to 3 decibels of a full quarter-wave vertical antenna erected over the same ground plane.

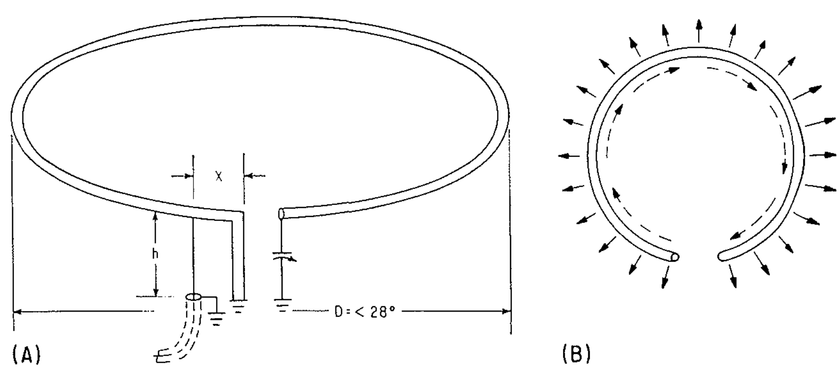
A circular array with a circumference of one full electrical quarter wavelength results in an antenna diameter of 28 electrical degrees. To attain an in-phase electric distribution over this circumferential aperture means using a center-fed radial waveguide like the flat top antenna. Study of a 28-degree diameter electric array by techniques such as those developed by Chireix,[4] shows that even with a 90-degree shift in phase around the aperture, an omnidirectional radiation pattern in the horizon plane will result. To establish a circular array a conductor one-quarter wavelength in length is conductively joined to the

HOOP WITH A PURPOSE

Low-frequency antennas require either long wires supported by tall masts, or vertical tower radiators 60 to 300 feet. Here is an antenna that looks like a child's hula-hoop and has performance characteristics closely approaching those of a full quarter-wave vertical.

The DDRR antenna (this week's cover) offers a height reduction of up to thirty-to-one over verticals now in use and can range in size from 6-inches to 5,000 feet in diameter and 2-inches to 300 feet in vertical height.

A model of this new antenna only 2-feet high, recently equaled the performance of a 60-foot vertical radiator



RING TRANSMISSION LINE *showing feed and tuning, if required (A) and current in ring element (B)—Fig. 1*

A Coming Trend?

By J. M. BOYER
Chief, Radio Physics Group,
Northrop Ventura, Van Nuys, Calif.

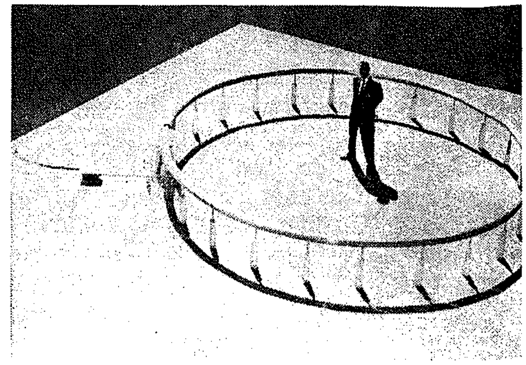
top of a 2.5-degree high vertical element and bent around in the horizon plane at this height to form a circle as shown in Fig. 1A. If a generator is connected across the slot formed by the circular conductor and the ground plane, an energy wave can be launched in this curved boundary region.

When a change of direction occurs in any electromagnetic waveguide system, higher order modes are established.[6] If this discontinuity takes place in a completely closed system such as rectangular waveguide, the equivalent circuit of the discontinuity will contain only reactive components because no power is lost in radiation. If, however, the fields are not confined, but extend beyond the guide boundaries, the discontinuity equivalent circuit contains resistive and reactive components. Field line fringing or extension effect is present in dielectric waveguides and open-wire transmission lines.

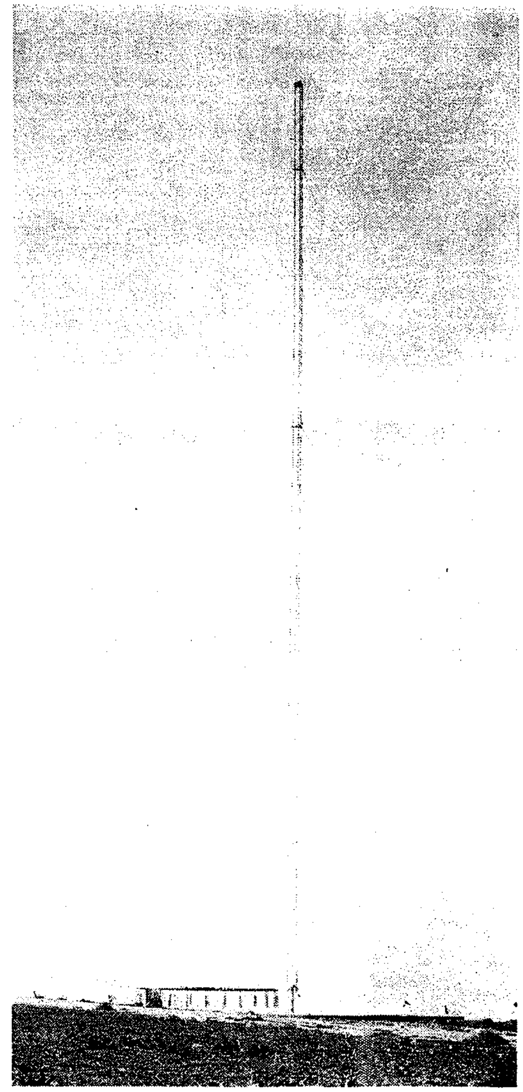
The constant-height ring-conductor just described forms a single-wire transmission line with the ground-plane surface. It runs a straight path rather than in a curve, this close-spaced line produces little radiation as shown by King.[6] This condition is true, to a first order, because the TEM mode exists over the length of this line, with discontinuities occurring only at the ends.

If the same single wire line is bent into a curve, however, a wave launched at its input terminals encounters a different set of conditions. The bent waveguide is a constant series of directional discontinuities. Thus, the launched wave radiates continuously in a direction transverse to the line-axis throughout its entire length. Radiation occurs from two sources; a horizontally polarized wave is launched from the current-flow in the ring element itself, but is cancelled because of the antisymmetric current relation in the image plane. At the same time, vertically polarized radiation takes place from the higher-order modes established by the direction discontinuity. The launched wave radiates as it moves around the ring until it meets the far end. The energy still remaining at this point reflects, radiating on its path back to the generator. Thus, the DDRR antenna might well be called a leaky-waveguide radiator, with radiation being integrated over its entire circumferential boundary or aperture.

The DDRR antenna produces typical dipole doughnut radiation pattern when its diameter is not large in wavelength. Chireix analysis quickly explains the omnidirectional azimuth plane pattern. In the elevation plane, the zenith null must occur due to the out-of-phase relationship of parallel current components on the ring element and the typical antisymmetric distribution of electric lines of force around

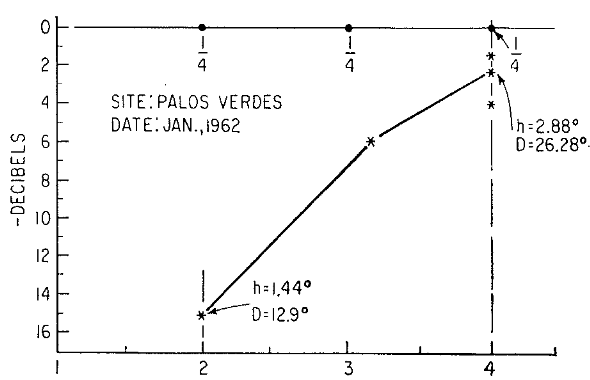


(A)



(B)

MODEL *of the DDRR antenna operating between 2 and 4 Mc (A) and the same antenna compared to a quarter-wave vertical (B)—Fig. 3*



AVERAGE GAIN *versus frequency of a DDRR antenna compared to a full, quarter-wave vertical radiator with same counterpoise—Fig. 2* the circular aperture as shown in Fig. 1B.

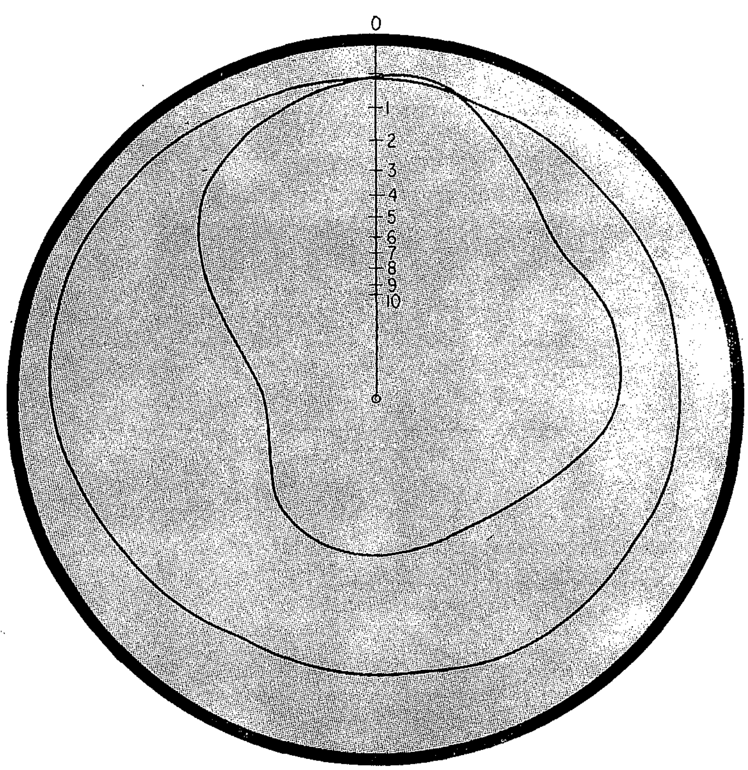
Although the ring-type-DDRR antenna design simplifies low-height radiator structure by eliminating the large current sheet found in the flat top model, it has other advantages. Unlike the flat top, the DDRR design is naturally resonant when the diameter of the aperture is approximately 28 electrical degrees. Resonance is relatively unaffected by the height h above the ground plane if kept well below 90 degrees. When size limitations restrict the diameter to less than 28 degrees, electrical resonance can be restored with a low-loss air or vacuum capacitor connected from the open end of the ring to the ground plane.

The DDRR design permits direct connection of transmission lines across the aperture and ground plane. Any line from 36 to 500 ohms may be used if dimension X is varied to suit its characteristic impedance. Thus, the impedance-matching network required with other short height designs is eliminated, together with attendant electrical loss and additional restriction of bandwidth. When point X has been determined to conform with the transmission line, a DDRR antenna may be capacitively tuned over at least a 2:1 frequency range without input vswr exceeding the 2:1 limit.

PERFORMANCE—No single antenna design is a panacea for the problems of communications, but the performance and flexibility of the DDRR antenna, represents a significant improvement.

Figure 2 shows typical gain variation of a simple DDRR antenna over a 2:1 frequency band. The magnitudes are the average for many field-strength comparisons to full quarter-wave vertical antennas taken at steadily increasing range over a combination land and sea path, where distances ranged from 1 to 132 miles in both a North and South path. The model used was small electrically, being only 12.9 degrees in diameter and 1.44 degrees high at the 2-Mc limit and 26.28 degrees in diameter and 2.88 degrees high at 4 Mc. This antenna appears in Fig. 3A. At the same frequency limits, the comparison test antenna was a one-foot, triangular cross-section tower, 110 feet tall, at 2 Mc and 68 feet high at 4 Mc. as shown in Fig. 3B. Operation was over rocky soil, using 90 radials each one-half wave long. Tests were made under the special call KM2XOP by authority of the FCC. Tuning of the test antenna over the entire frequency range was completely remote, using a servo actuated Jennings vacuum variable capacitor adjusted from the transmitter console 500 feet away; vswr was under 2:1 in 50-ohm coaxial cable at all frequencies.

MOBILE USE—At higher frequencies, the DDRR has proven a convenient and efficient device for mobile communications. A mobile model designed for use between 26.5 and 31 Mc is 27 inches in diameter and projects only 3½ inches above the vehicle roof. The horizontal radiation pattern of the DDRR and that of a full quarter-wave whip antenna on the same vehicle are shown in Fig. 4. Departure from a perfect omnidirectional pattern for both antennas is due to the nonsymmetrical ground plane geometry provided by the metal skin of the car. In addition to increased efficiency, the antenna possesses other important advantages for the mobile service. At high road speeds, there is no signal-flutter effect from the DDRR antenna. This wind effect is severe in vertical antennas cut for the same frequency range. Also, the simple DDRR design used in the mobile service acts as a sharp band-pass filter centered on the operating channel. Thus, better isolation from adjacent-channel interference is achieved during reception than with conventional designs. Being directly grounded to the car frame, an automatic static drain to ground is provided for static charge induced by fog, dust and precipitation, affording improved receiver signal-to-noise ratio.



POLAR PLOT *of a vehicular quarter-wave whip (inner) and that of a DDRR antenna (outer)—Fig. 4*

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Broadband Log-Periodic Antennas

These antennas can operate over a 10:1 frequency bandwidth while a range of antenna configurations confer important directional characteristics. Representative types of antennas are presented in this short survey

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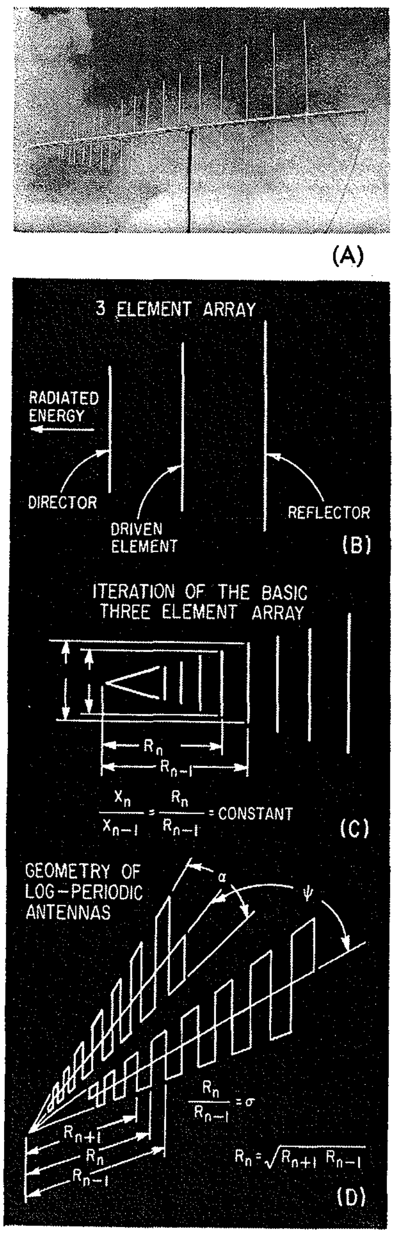


FIG. 1—Versatile log-periodic antenna is capable of horizontal, vertical and circular polarization (A), basic three-element array (B), iteration of the basic three element array (C), geometry of the log-periodic antenna (D)

THE LOG-PERIODIC antenna concept, originated by DuHamel, Isbell and their colleagues at the University of Illinois, is a remarkably flexible design approach for good antenna performance over wide frequency bands. Omnidirectional, bidirectional and unidirectional configurations are available that have good patterns and vswr characteristics, and constant performance over frequency ranges up to 10:1. Some designs permit remote selection of polarization. Circularly polarized strutcures are also practical. Applications include monitoring and signal interception, direction finding, satellite tracking, radio astronomy and h-f communications.

Logarithmically periodic antennas achieve wideband properties by geometric iteration. Consider the simple combination of driven dipole, parasitic reflector and direction of Fig. 1B. This combination radiates a unidirectional beam in the direction of the arrow, with the field polarized in the plane of the elements. With a proper choice of element lengths and spacing, the vswr and gain can be made reasonably constant over a frequency band of ±10 percent. When the antenna is driven at frequencies which are well outside these limits, however, the resonant character of the self and mutual impedances, and the space phasing which accounts for the directivity, result in rapid deterioration of performance. The vswr increases, the gain decreases, and undesirable secondary lobes appear. If additional elements are added, as in Fig. 1C, and the feed arrangement is appropriate, much wider band performance is obtainable. DuHamel and Isbell arranged these additional elements so that the significant dimensions increased in a logarithmic fashion from one end of the array to the other, and such that the ratio of element length to spacing was constant. It is apparent that this amounts to iterating the basic 3-element array, so that the final array can be regarded as a superposition of 3-element arrays, differing from one another only by a constant geometric scaling factor.

The usual log-periodic factors can be expressed in terms of the parameters defined in Fig. 1.

$$X_n = (X_{n+1}\, X_{n-1})^{1/2} = X_{n+1}/\sigma = \sigma\, X_{n-1} \quad (1)$$

$$R_n = (R_{n+1}\, R_{n-1})^{1/2} = R_{n+1}/\sigma = \sigma\, R_{n-1} \quad (2)$$

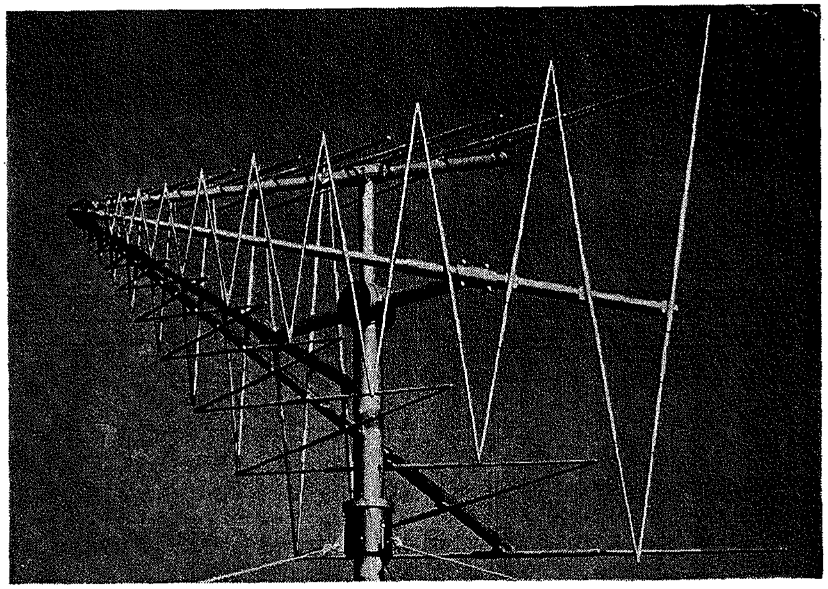
where σ is a constant that is less than unity.

Research on antennas structures of this type has shown that it is possible to design for any of a wide range of performance parameters while retaining characteristic wideband features. The most significant design parameters are σ, as defined above, d, the taper angle, and, in arrays of log-periodic elements, Ψ, the angle between the axis of the elements in the array.

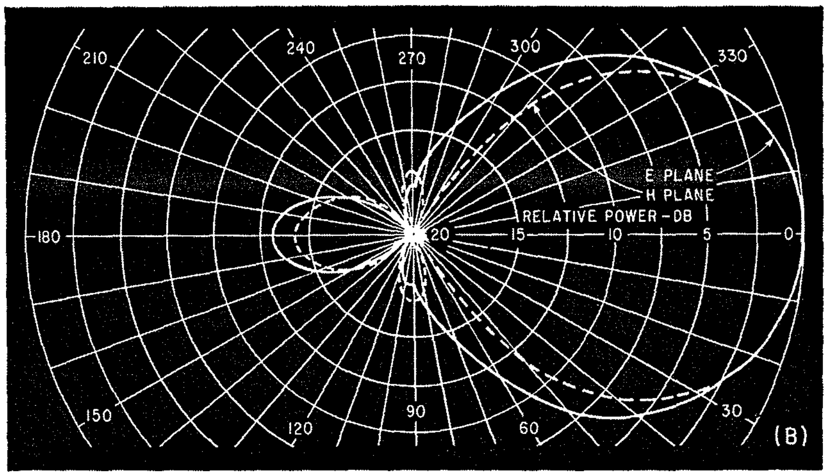
The log-periodic structure has a characteristic phase behavior that permits its application to circularly polarized and omnidirectional rerequirements. This frequency-independent phase shift capability is directly related to the behavior of the structure when it is regarded as a delay line supporting a transmission line mode.

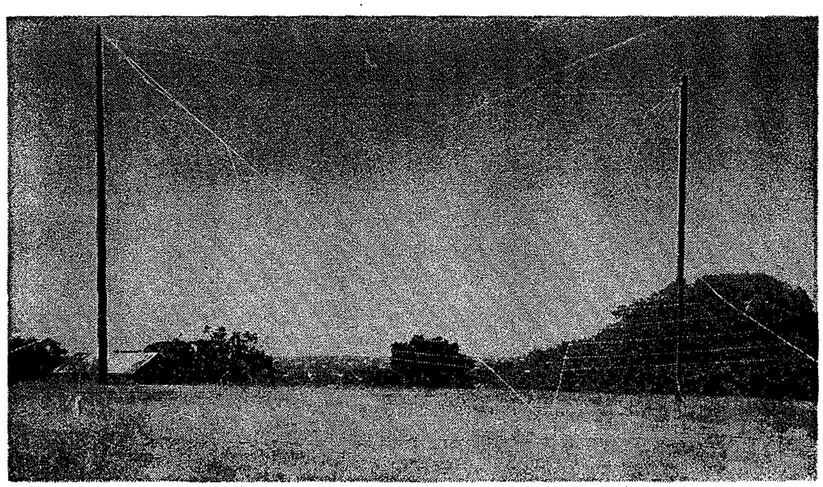
From the wide band performance and simplicity of phasing, gains up

FIG. 2—Antenna can be vertically or horizontally polarized by remote switching (A); polar diagram of top antenna (B). The h-f antenna radiates a low-elevation-angle beam (C); polar diagram of h-f antenna (D)

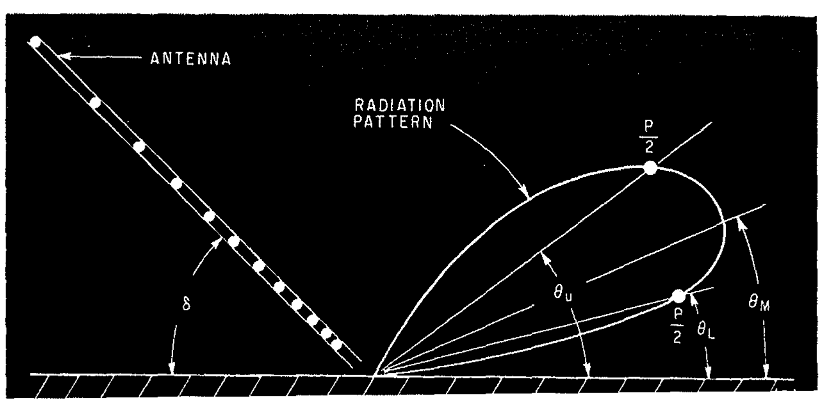


(A)





(C)



to 14-db covering a wide bandwidth may be had without using several large antennas and awkward switching arrangements. In intercept receiving systems this means smaller cheaper arrangements and greater intercept probability. In transmitting systems, the wide antenna bandwidth greatly simplifies antenna configurations, eliminating interference and switching problems.

While most applications of l-f arrays have been military, there are obvious civilian uses, including amateur radio, TV reception and aircraft control.

Figure 2A shows an antenna that is capable of either vertical or horizontal polarization by remote switching. The structure consists of two linearly polarized antennas mounted orthogonally. The radiators are of triangular construction and the structure is operational from 100 to 1,000 Mc. Typical E- and H-plane patterns for either polarization are also shown in Fig. 2B. This structure has an average impedance of 90 ohms with a maximum vswr of less than 2.0 to 1 referenced to this value.

A horizontally-polarized antenna arrayed against the ground for h-f application is shown in Fig. 2C. The radiators of this structure are formed from No. 10 copper wire and are held parallel to the ground by dielectric catenaries suspended from the vertical poles. The structure imaged in this manner radiates a low elevation-angle beam that has approximately 50 degrees between half-power points in the azimuth plane. Other versions of this type antenna permit even wider ranges in the elevation angles of the main beam.