ANTENNAS FOR MEDIUM-FREQUENCY BROADCASTING

The beginnings of medium-frequency broadcast antennas (530 to 1700 kHz) can be traced to the early 1920s. The first antennas constructed were made of a pair of steel or wooden masts supporting an antenna structure consisting of a vertical wire or wire cage, sometimes accompanied by a horizontal section consisting of a wire or flat surface or cage of wires. The antennas with the horizontal members where referred to as T- or L-type antennas. Figure 1 illustrates the physical characteristics of the early medium-frequency antennas. Most of these antennas did not exceed a physical height of 50 to 70 electrical degrees. In 1924, Ballantine (1) showed that longer antennas would result in a substantial gain in the horizontal plane radiation. Heights were then increased to as much as 135 electrical degrees with the first commercial antennas being constructed in the early 1930s.

During the 1930s the present type of medium-frequency broadcast antenna, a self-supporting or guyed tower in which a base-insulated tower is utilized as the radiating element with an accompanying ground system, was developed. The classic paper by Chamberlain and Lodge (2) spearheaded this development and offered many advantages over the earlier antennas. The radiation efficiency of a nondirectional radiator was often more than double when this new type of design was put into service and the cost of such a structure decreased significantly as the number of required towers was cut in half. It was also found necessary to use breakup insulators to reduce guy wire current. Figure 2 shows a single tower radiator of this type. As the number of medium-frequency broadcast stations increased during the 1930s, it was necessary to develop directional antenna systems to minimize interference between stations. A directional antenna consists of multiple towers used in a phased array configuration and excited with various amplitude and phase relationships to form a pattern in the desired shape. The single excited tower radiating element made the directional antenna concept an economically feasible possibility. The first directional antenna system designed by Dr. Raymond M. Wilmotte, was constructed by WSUN in St. Petersburg, Florida and employed two towers to produce a radiation pattern null toward cochannel station WTMJ in Milwaukee, Wisconsin to resolve a nighttime interference controversy.

The 1940s brought further development to the design of medium-frequency broadcast antennas. Top-loading, sectionalizing, and improved ground systems were introduced to improve antenna efficiency as well as to control vertical radiating characteristics. The number of towers used in directional arrays was increased to as many as nine elements as the power dividing and phasing systems were improved.

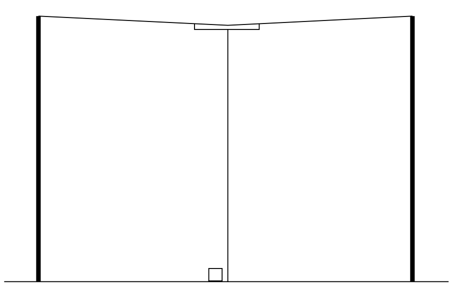

Figure 1. Early AM antenna utilizing two vertical masts supporting the radiating structure.

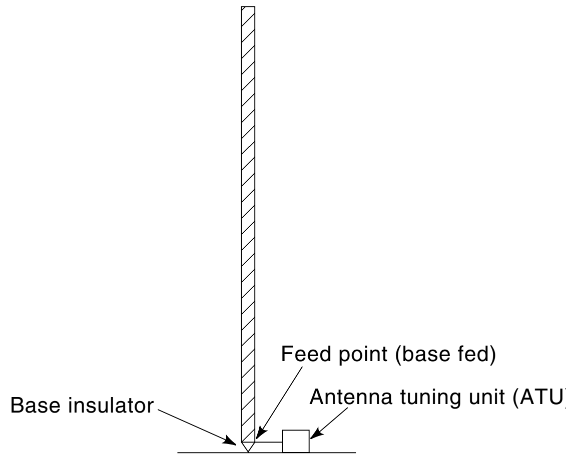

Figure 2. Present antenna configuration using the tower as the radiating element. Structure can be self-supporting or supported with guy wires.

Since the 1940s, there has not been significant development in the area of the antenna element itself as self-supporting and guyed radiating towers continue to be used much as they were 50 years ago. In the 1960s, three 12-tower arrays were constructed at WJBK in Detroit, Michigan, CFGM in Toronto, Canada and KLIF in Dallas, Texas. These were the largest arrays of driven elements ever constructed for medium-frequency broadcast use. New developments since then principally involve auxilliary equipment used to test and monitor the antenna system and the means by which the antenna is theoretically analyzed. Digital antenna monitors are used today to accurately monitor the relative current magnitudes and phase relationships of the towers in a directional array while portable solid-state field strength meters are used to measure radiation patterns. Advances have also been made with improved RF current meters, sampling transformers, and impedance measuring equipment. The design of medium-frequency broadcast antennas for optimized performance has also seen tremendous advances over the past 15 years with the introduction of numerical solutions to electromagnetic problems and nodal modeling of feeder systems made economically possible with personal computers vastly simplifying both design and implementation.

GENERAL ANTENNA CHARACTERISTICS

The medium-frequency range is generally defined from 300 to 3000 kHz. The portion of the band allocated to AM broadcasting is from 530 to 1700 kHz in North America. The channels for the individual broadcast stations are spaced 10 kHz apart. In contrast, within the medium-frequency broadcast band in other regions of the world, the stations are spaced 9 kHz apart.

Electrical

The typical antenna used for medium-frequency broadcasting is electrically equivalent to one or more base-excited monopoles over a finite perfectly conducting ground plane. In directional antenna systems, where more than one element is employed, the amplitude and phase of the current to each monopole are varied in relationship to one another to determine the pattern size and shape. The height of each monopole can vary in height from as little as 45 to as much as 225 electrical degrees. This translates to a range of physical heights of 25 to 200 m within the medium-frequency band. In a few rare instances taller towers are employed; they are center-fed and known as *Franklin Antennas.*

Mechanical

A self-supported or guyed steel tower is usually used as the radiating element. The tower can be triangular or square in cross-section and can have a face width ranging from a fraction of a meter to several meters. A ground system consisting of copper wires typically extends radially from the base of each tower a length of 90 electrical degrees from the tower base. Normal practice is to use 120 wires equally spaced (every three degrees) and equal in length except where they would overlap between adjacent towers for the ground system. Overlapping of wires is avoided by shortening them and bonding them to a transverse conductor (usually a copper strap).

Propagation

The electromagnetic field propagates from a medium-frequency antenna system in two modes. The first mode travels along the surface of the ground and is referred to as groundwave propagation. The second mode radiates directly into space and refracts from the ionosphere before reaching the target area and is referred to as skywave propagation. Effective skywave propagation is severely attenuated during daytime hours and is significant only at night. Groundwave propagation is dependent on the characteristics of the terrain over which the signal propagates. Propagation models for groundwave and skywave signals, as they affect spectrum management and individual station authorization, are prescribed by the government agency having jurisdiction within the country where a station is located. The agency with jurisdiction over stations operation within the United States is the Federal Communications Commission (FCC), while many foreign countries use propagation standards published by the ITU/CCIR.

STATION CLASSIFICATIONS

The FCC of the United States regulates all medium-wave radio broadcasting in the US. The FCC has classified all medium-frequency radio stations into categories defining their coverage areas and power levels (3). An *unlimited* time station can broadcast at all times during the daytime and nighttime, whereas a *limited* time station (usually daytime) can only broadcast at certain specified times. A *primary* service area is defined as the area within close proximity to the station and where groundwave propagation provides a high-quality signal. A *secondary* service area is more distant from the station and usually depend on skywave propagation during nighttime hours.

Under the international agreements governing medium-wave broadcasting in the western hemisphere (ITU Region III), there are three classes of frequency allotments. By bilateral agreements among the North American countries, certain channels or frequencies are reserved for use by stations providing various classes of service.

Clear Channel

A *clear channel* classification is assigned to stations covering wide service areas and is subdivided into three classes:

1. *Class A.* Unlimited stations assigned to primary and secondary service area. Power levels range between 10 and 50 kW.
2. *Class B.* Unlimited service assigned to primary service areas only. Power levels range between 0.25 and 50 kW.
3. *Class D.* Limited service daytime or unlimited service with no nighttime service or nighttime power less than 0.25 kW. Power levels range between 0.25 and 50 kW.

Regional Channel

A *regional channel* classification is assigned to stations serving a principal center of population and the surrounding rural areas and is subdivided into two classes:

1. *Class B.* Unlimited service assigned to a primary service area. Power levels range between 0.25 and 50 kW.
2. *Class D.* Limited service daytime or unlimited service with no nighttime service or nighttime power less than 0.25 kW. Power levels range between 0.25 and 50 kW.

Local Channel

A *local channel* classification is assigned to stations serving a community and the surrounding suburban and rural areas and consists of one class:

1. *Class C.* Unlimited service assigned to a primary service area. Power level range between 0.25 and 1 kW.

ALLOCATION STUDIES

Before a medium-frequency station can be licensed, an analysis is required to determine compliance with rules governing acceptable interference levels between stations. In the United States, the analysis is based on propagation models and conductivity maps as defined in the Code of Federal Regulations (3).

Propagation Models

Field strength algorithms exist that can be used to predict the coverage of a particular antenna system. The *equivalent-distance groundwave model* is a prediction method used for daytime groundwave field strength calculations when signals propagate over one or more conductivity regions. The *1992 FCC Skywave Model* (47 CFR 73.183) is a prediction method used for most nighttime field strength calculations within the United States. The *Region 2 Annex II Figure 4 Skywave Model* is a predication method used for nighttime skywave field strength calculations between the United States and Central and South America, and the Caribbean Islands. Also, the *US–Canada Bilateral Agreement, Annex II Figure 4A Sky-*

wave Model is a predication method used for nighttime skywave field strength calculations between the United States and Canada. Propagation models and techniques for determining allowable interference differ substantially from one part of the world to another. Propagation models for other regions of the world are specified in regional agreements administered by the International Telecommunications Union (ITU) in Geneva, Switzerland.

Ground Conductivity Maps

The calculation of groundwave field strength levels depend on predicted ground conductivity and dielectric constant values for the area of interest. It is an acceptable simplification for most engineering analysis to define ground conductivity with a fixed dielectric constant. Predicted conductivities are usually presented in the form of a map or computer model delineating boundaries between regions of different conductivities. The M3 map is included in the FCC rules and shows predicted ground conductivity for the continental United States. The Region 2 map covers a larger area and shows predicted ground conductivity for much of the Western Hemisphere. The M3 map is used for calculations between stations within the United States while the Region 2 map is used between stations in the United States and stations in Canada, Central America, and the Caribbean. Other countries utilize conductivity maps developed for their own regions.

Field Strength Contours

When analyzing the coverage from a given antenna system, it is useful to calculate field strength contours at various levels to determine if the station is providing adequate coverage to the target area and if interference exists between stations on cochannel or adjacent channel frequencies. A field strength level of 50–25 mV/m is considered necessary to provide premium service to heavily built-up urban and industrialized areas, whereas a field strength of 5 mV/m is often considered to be satisfactory for the less heavily built-up surrounding areas. A field strength of 2.0 mV/m provides service to residential areas and 0.5 mV/m is the minimum signal level for service to rural areas in non-tropical regions of the world.

Daytime Allocation Study

A daytime allocation study involves the calculation of groundwave field strength contours to determine if interference exists between stations on cochannel or adjacent channel frequencies. The required protections as specified by the FCC are given in Fig. 3. The most stringent protection is afforded to cochannel stations with decreasing protection levels to the first, second, and third adjacent channels chosen to eliminate *splatter* between the signals of nearby stations.

Nighttime Allocation Study

A nighttime allocation study involves the calculation of nighttime skywave field strength levels to determine if interference exists between stations on cochannel or adjacent frequency channels. In contrast to the daytime study in which field strength contours are determined, the nighttime study involves point-to-point calculations. In the United States, the method for determining protection between stations requires calculating the received interfering field strengths from all cochannel and first-adjacent-channel stations. The square root of the sum of the squares (RSS) is calculated using all the interfering signals in descending order and determines the overall interference level. The levels of interference which are defined by the FCC are the 50 and 25% RSS levels. The 50% RSS level includes only the stations that contribute a signal level of at least 50% to the running RSS total, whereas the 25% RSS level includes all stations that contribute a signal level of at least 25% to the running RSS total. When a station is newly licensed or undergoes a major change (increase in power or modified pattern), it cannot increase interference to existing stations above the 25% RSS level. If a station presently causes interference at a level between 25 and 50% of another station, its interference contribution cannot be increased at all, and, if it presently caused interference above the 50% level, its interfering signal must be decreased at the affected station by at least 10% under the present FCC rules. The service area of a station during nighttime hours is considered to be the area which is defined by the 50% RSS boundary.

Frequency separation (kHz)	Contour of proposed station (classes B, C and D) (mV/m)	Contour of any other station (mV/m)
0	0.005	0.100 (Class A)
	0.025	0.500 (Other classes)
	0.500	0.025 (All classes)
10	0.250	0.500 (All classes)
	0.500	0.250 (All classes)
20	5.0	5.0 (All classes)
30	25.0	25.0 (All classes)

Figure 3. Daytime protection limits as specified by the Federal Communications Commission. Protection limits are instituted to reduce co-channel and adjacent channel interference.

ANTENNA DESIGN

Antenna design involves the selection of physical and electrical parameters that meet all design requirements as determined by the allocation studies while simultaneously providing a satisfactory level of interference-free coverage to the proposed coverage area, including the "community of license," from the selected transmitter site.

Mechanical

Radiating Elements. Two types of radiating elements are typically used in medium-frequency broadcast antenna system today: self-supporting or guyed towers. A self-supporting tower consists of a free-standing tapered steel structure. A guyed tower is usually of uniform cross section and is supported by insulated steel guy cables or nonconductive cables attached at multiple levels. Either type of tower can be top-loaded (with a horizontal steel circular cap or a portion of the guy wires connected directly to the tower to achieve greater electrical height with a physically shorter structure) or sectionalized (when the tower is broken into sections and a series inductance is inserted between them to reduce the reactance of the upper sections). In some cases, where towers on the order of a wavelength tall are employed, they are center fed. Such center-fed towers are known as Franklin Antennas.

Feed Point. The feed point is the location at which the radiating element is fed power from the transmission line. A series feed system feeds the power across a tower's base insulator, whereas a tower may be shunt-fed either with a slant wire attached part way up its structure or with a wire skirt at its base. A major advantage of grounded tower radiators with skirt wires is the elimination of isolation components for lighting circuits. This is especially true for very high-power operation of stations outside the US. A Franklin Antenna may be fed across either an insulator or gap in a wire skirt at approximately one half of its height. Some type of balun (usually a quarter-wave line section) must be employed to isolate the circuit across the ground level insulator of a Franklin Antenna.

Ground System. The ground system is a conductive screen or grid of wires imbedded in the earth around the base of each radiating element to allow ground currents to return directly to its base. A typical ground system consists of 120 buried copper wires, equally spaced, extending radially outward from each radiating element base to a minimum distance of 90 electrical degrees. An exposed copper mesh may also be used around the base when high voltages are expected.

When multiple elements are used to form an array, the ground radial wires from adjacent towers will often overlap. When this occurs, a copper transverse strap or cable is employed for bonding the radials together so that they do not extend into the area where they would overlap.

Lighting System. A system of beacons or continuously illuminated lights mounted on each tower at various heights is often required for towers above a certain height. Series excited towers must have some means of coupling the ac power to the lighting circuit on the towers while the wiring may proceed directly from ground level up a shunt-fed grounded tower. Lighting chokes, which provide a high impedance at the RF frequency while conducting ac current are often used for connecting the lighting circuits on towers across their base insulators. Another method employed uses a *ring* transformer that is constructed so that the primary and secondary have sufficient spacing between them to withstand typical base voltages while adding only a slight amount of capacitance across the base as far as the RF energy is concerned.

Lightning Protection. Because of the relatively tall and conductive nature of a medium-frequency antenna systems, they are very susceptible to lightning strikes. Additionally, high transient voltages can be induced at their bases due to distant lightning strikes and they are subject to high static buildup under certain environmental conditions. Therefore it is necessary to include a system to protect the radiating elements and associated tuning components from being damaged by lightning strikes. A tower protection system usually consists of a pointed vertical rod or rods at its top, extending above the tower lighting beacon if one is employed, a conductive circuit such as an RF choke across the tower base to provide a low impedance path to ground (for series-excited towers) and a set of arc-gaps directly across its feedpoint.

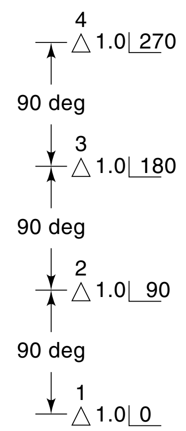

Figure 4. Orientation of antenna array for a given set of field parameters. The current magnitude and phase is given relative to tower no. 1.

Electrical

Electrical parameters are chosen such that size and shape of an antenna pattern meet the radiation limits identified in the allocation studies. Antenna patterns fall into two broad categories, directional and nondirectional. Where a power level to provide satisfactory coverage can be had without any interference to other stations, a new facility may employ a nondirectional antenna consisting of only one radiating element. Otherwise, a multielement array (directional antenna) must be employed to meet the protection requirements. The parameters used to design a directional antenna pattern include the field ratios and phase relationships between elements, the number of elements, the height of each element, and the physical orientation of each element. These are the factors that determine the size and shape of the pattern such that the amount of energy radiated is controlled in any given direction.

Electrical Parameters. The following example (Table 1) shows how the electrical design parameters are typically specified for a four element array. Note that Tower No. 1 is used as reference with phase, spacing, and bearing set to zero.

Table 1. Electrical Design Parameters in a Four Element Array

Tower No.	Field Ratio	Phase (degrees)	Spacing (degrees)	Bearing (degrees)	Height (degrees)
1	1.000	0.0	0.0	0.0	90.0
2	1.000	+90.0	90.0	0.0	90.0
3	1.000	+180.0	180.0	0.0	90.0
4	1.000	+270.0	270.0	0.0	90.0

The field ratio gives the relative magnitude of the radiated field from each element. The spacing and bearing of each element is given with respect to the reference element. The relative electrical phase relationships between the elements are also specified. The bearing of each element (physical orientation) is given in true degrees azimuth. The spacing and height of each element are given in electrical degrees. A plan of the preceding example is shown in Fig. 4 while the horizontal pattern is shown in Fig. 5.

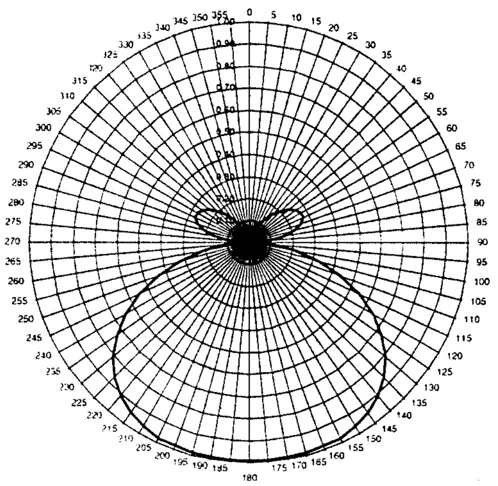

Figure 5. Horizontal pattern of field parameters given in Fig. 4. Use of multiple towers produces pattern directivity.

Pattern Shape. The shape of a radiation pattern is controlled by varying the electrical parameters and the geometry of the individual radiating elements (usually towers). The most elementary directional antenna radiation patterns are developed using two-tower arrays of elements. More towers are added as necessary to meet more complicated radiation pattern requirements.

Theoretical Pattern. A theoretical radiation pattern can be calculated using the following formulation, which represents the inverse distance field at 1 km for a given azimuth and elevation angle.

$$E(\phi,\theta)_{\text{th}} = \left| K \sum_{i=1}^{n} F_i f_i(\theta) \angle \overline{S_i \cos\theta \cos(\phi_{\text{i}} - \phi) + \varphi_i} \right| \tag{1}$$

where

- K = multiplying constant which determines the basic pattern size
- n = number of elements in the directional array
- i = the ith element in the array
- F_i = field ratio of the ith element in the array at $\theta = 0$
- θ = vertical elevation angle measured from the horizontal plane
- $f_i(\theta)$ = vertical plane radiation characteristic of the ith element in the array
- S_i = electrical spacing of the ith element from the reference point
- ϕ_i = orientation (with respect to true north) of the ith element in the array
- ϕ = azimuth (with respect to true north)
- φ_i = electrical phase angle of the current in the ith element in the array

Figure 6 shows the reference coordinate system.

***K* Factor.** The multiplying constant, K, can be obtained by numerically integrating the effective field intensity as calculated at each vertical angle in half space. Calculation at 5° or 10° intervals is satisfactory for results that are acceptable from an engineering standpoint.

$$K = \frac{E_{\text{s}}\sqrt{P}}{e_{\text{h}}} \tag{2}$$

where E_s is the horizontal radiation from a standard isotropic radiator in half space at 1 km distance and 1 kW power level, P is the antenna input power, and e_h is the root-mean-square effective field strength in half space.

$$e_{\text{h}} = \left\{ \frac{\pi\Delta}{180} \left[\frac{e_{\text{a}}^2(0)}{2} + \sum_{m=1}^{N} e_{\text{a}}^2(m\Delta)\cos(m\Delta) \right] \right\}^{1/2} \tag{3}$$

where Δ is the interval between vertical elevation angles, $N = (90/\Delta) - 1$ (number of intervals minus one), and $e_a(m\Delta)$ is the root-mean-square field strength at angle $m\Delta$.

$$e_{\text{a}}(m\Delta) = \left\{ \sum_{i=1}^{n}\sum_{j=1}^{n} F_i f_i(m\Delta) F_j f_j(m\Delta) \cos\psi_{ij} J_0[S_{ij}\cos(m\Delta)] \right\}^{1/2} \tag{4}$$

where

- j = jth element
- $f_i(m\Delta)$ = vertical radiation characteristic of the ith element
- F_j = field ratio of the jth element
- $f_j(m\Delta)$ = vertical radiation characteristic of the jth element
- ψ_{ij} = difference in phase angles of the currents in the ith and jth elements
- S_{ij} = spacing between ith and jth elements
- $J_0(x)$ = Bessel function of the first kind and zero order

Vertical Plane Radiation Characteristic. The vertical plane radiation characteristics show the relative field being radiated at a given vertical angle (θ), with respect to the horizontal plane. The general form is

$$f(\theta) = \frac{E(\theta)}{E(0)} \tag{5}$$

where $E(\theta)$ is the radiation from an element at angle θ and $E(0)$ is the radiation from an element in the horizontal plane.

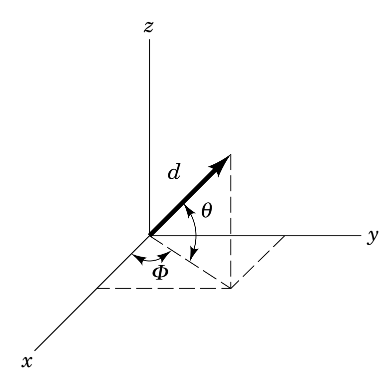

Figure 6. Definition of reference coordinate system. $\theta = 0$ in the x–y plane.

Assuming sinusoidal current distribution for a typical element that is not top-loaded or sectionalized, the vertical radiation is

$$f(\theta) = \frac{\cos(G\sin\theta) - \cos G}{(1-\cos G)\cos\theta} \tag{6}$$

where G is the electrical height of the element. For a top-loaded element, the vertical radiation is

$$f(\theta) = \frac{\cos B\cos(A\sin\theta) - \sin\theta\sin B\sin(A\sin\theta) - \cos(A+B)}{\cos\theta[\cos B - \cos(A+B)]} \tag{7}$$

where A is the physical height of the element in electrical degrees, B is the difference between the apparent electrical height and the actual physical height in electrical degrees, and G is the apparent electrical height, $A + B$.

For a sectionalized element the vertical radiation is

$$f(\theta) = \frac{\begin{aligned}&\sin J[\cos B\cos(A\sin\theta) - \cos G]\\&+\sin B[\cos D\cos(C\sin\theta) - \sin\theta\sin D\sin(C\sin\theta)\\&-\cos J\cos(A\sin\theta)]\end{aligned}}{\cos\theta[\sin J(\cos B - \cos G) + \sin B(\cos D - \cos J)]} \tag{8}$$

where A is the physical height of the lower section of the element in electrical degrees, B is the difference between the apparent electrical height of the lower section of the element and the physical height of the lower section of the element, C is the physical height of the entire element in electrical degrees, D is the difference between the apparent height of the element and the physical height of the entire element (D will be zero if the sectionalized tower is not top loaded), and

$$G = A + B$$
$$H = C + D$$
$$J = H - A$$

Standard Pattern. The FCC has also defined a standard pattern that is an envelope around the theoretical pattern and is intended to provide a tolerance within which the actual operating pattern can be maintained. All designs must be based on the standard pattern and are calculated as follows:

$$E_{\text{std}} = 1.05\sqrt{E^2 + Qg^2(\theta)} \tag{9}$$

$Q = 10\sqrt{P}$ or $0.025E_{\text{rss}}$ (whichever is greater)
P = power (kW)
$E_{\text{rss}} = E_1\sqrt{\sum_i^n F_i^2}$
E_1 = reference field
F_i = field ratio of the ith element
$g(\theta) = f_s(\theta)$, if the shortest element is shorter than $\lambda/2$
$= \sqrt{f_s^2(\theta) + 0.0625/1.030776}$, otherwise
$f_s(\theta)$ = vertical radiation characteristic of the shortest element

Pattern Augmentation. The FCC Rules include a provision to augment the standard pattern to take into account actual operating conditions when radiation is greater than the standard pattern in certain directions. Radiation is augmented over a specified azimuthal span and is calculated as follows:

$$E_{\text{aug}}(\theta) = \sqrt{E^2_{\text{std}}(\theta) + A\left[g(\theta)\cos\left(\frac{180D_{\text{a}}}{S}\right)\right]^2} \tag{10}$$

E_{std} = standard radiation pattern
$A = E^2_{\text{aug}}(\theta) - E^2_{\text{std}}(\theta)$ at central azimuth of augmentation
S = azimuthal span of augmentation centered on central azimuth of augmentation
D_{a} = absolute difference between azimuth of calculation and central azimuth of augmentation; note, D_{a} cannot exceed $S/2$ for augmentation within a particular span

Pattern Size. The size of an antenna pattern is the magnitude of the radiation and is determined by the term K as introduced in the previous section. The term K is evaluated by calculating the total power radiated from the antenna. The total power radiating from any antenna structure can be determined by integrating the power flowing outward from a closed surface completely enclosing the antenna. The Poynting vector expresses the rate of power flow in watts/meter^2 at a given point in space and is expressed as

$$P = \boldsymbol{E} \times \boldsymbol{H} \tag{11}$$

where P is the power flow (W/m^2), $\boldsymbol{E}$ is the electric field intensity vector (V/m), and $\boldsymbol{H}$ is the magnetic field intensity vector (A/m).

In the far field, the two field vectors $\boldsymbol{E}$ and $\boldsymbol{H}$ are orthogonal and related in free space by the permeability and permittivity of air and is expressed as

$$H = \sqrt{\epsilon_0/\mu_0}\,E \tag{12}$$

where $\mu_0 = 4\pi \times 10^{-7}$ permeability (H/m), $\epsilon_0 = 1/\mu_0 c^2$ permittivity (f/m), and $c = 299.776 \times 10^6$ velocity of light (m/s).

The intrinsic impedance of free space is defined as

$$Z_c = \boldsymbol{E}/\boldsymbol{H} = \sqrt{\epsilon_0/\mu_0} = 376.71\,\Omega \tag{13}$$

The total amount of power flowing in free space from a given source is

$$p = E^2/Z_c \tag{14}$$

The total power radiated is calculated by integrating over a closed surface enclosing the source and can be expressed as

$$P = \int p\,dS = (1/Z_c)\int E^2\,dS \tag{15}$$

where P is the total power radiated (watts), Z_c is the intrinsic impedance (ohms), E is the total field at the closed surface (V/m), and dS is the incremental area on the closed surface (m^2).

If a sphere is chosen as the closed surface then the integration becomes

$$dS = d\cos\theta\,d\theta\,d\phi \tag{16}$$

Substituting Eq. (16) into Eq. (15) yields the total power radiating from a given source,

$$P = (1/Z_c)\int_0^{2\pi}\int_{-\pi/2}^{+\pi/2} E^2 d^2 \cos\theta\, d\theta\, d\phi \tag{17}$$

Isotropic Antenna in Free Space. If the radiating source is isotropic, the power radiates equally in all directions and Eq. (17) becomes

$$P = (1/Z_c)E_0^2 d^2 \int_0^{2\pi}\int_{-\pi/2}^{+\pi/2} \cos\theta\, d\theta\, d\phi \tag{18}$$

$$P = 4\pi(1/Z_c)E_0^2 d^2 \tag{19}$$

Solving Eq. (19) for E_0, the root-mean-square (rms) field intensity of an isotropic radiator is

$$E_0 = \sqrt{\frac{PZ_c}{4\pi d^2}} \tag{20}$$

For 1 kW of power at a distance of 1 km, Eq. (20) yields

$$E_0 = 173.14 \text{ mV/m}$$

Isotropic Antenna in Half-Space. Upon placing the isotropic radiator in a half space over a perfectly conducting plane, the limits of integration change and Eq. (18) becomes

$$P = (1/Z_c)E_0^2 d^2 \int_0^{2\pi}\int_0^{+\pi/2} \cos\theta\, d\theta\, d\phi \tag{21}$$

Solving Eq. (21) for E_0, the maximum rms field intensity of an isotropic radiator half-space is

$$E_0 = \sqrt{\frac{PZ_c}{2\pi d^2}} \tag{22}$$

For 1 kW of power at a distance of 1 km, Eq. (19) yields

$$E_0 = 244.86 \text{ mV/m}$$

Current Element in Free Space. If we replace the isotropic radiating source with an infinitesimally small vertical current element, the field intensity term is

$$E = E_0 \cos\theta \tag{23}$$

and Eq. (17) becomes

$$P = (1/Z_c)\int_0^{2\pi}\int_{-\pi/2}^{+\pi/2} (E_0 d\,\cos\theta)^2\, d\theta\, d\phi \tag{24}$$

Solving Eq. (24) for E_0, the maximum rms field intensity at $\theta = 0$ is

$$E_0 = \sqrt{\frac{3PZ_c}{8\pi d^2}} \tag{25}$$

For 1 kW of power at a distance of 1 km, Eq. (25) yields

$$E_0 = 212.05 \text{ mV/m}$$

Current Element in Half-Space. Upon placing the current element just above a perfectly conducting plane, the limits of integration change and Eq. (24) becomes

$$P = (1/Z_c)\int_0^{2\pi}\int_0^{+\pi/2} (E_0 d\,\cos\theta)^2\, d\theta\, d\phi \tag{26}$$

Solving Eq. (26) for E_0, the maximum rms field intensity at $\theta = 0$ is

$$E_0 = \sqrt{\frac{3PZ_c}{4\pi d^2}} \tag{27}$$

For 1 kW of power at a distance of 1 km, Eq. (27) yields

$$E_0 = 299.89 \text{ mV/m}$$

Center-Fed Conductor in Free Space. Now replace the radiating source with a center-fed conductor in free space of length $2G$ having a sinusoidal current distribution, the field intensity term is

$$E = E_0\left[\frac{\cos(G\sin\theta) - \cos G}{(1-\cos G)\,\cos\theta}\right] \tag{28}$$

and Eq. (14) becomes:

$$P = (1/Z_c)\int_0^{2\pi}\int_{-\pi/2}^{+\pi/2}\left\{E_0\left[\frac{\cos(G\sin\theta - \cos G)}{(1-\cos G)\,\cos\theta}\right]\right\}^2 d^2\cos\theta\, d\theta\, d\phi \tag{29}$$

Equation (29) has been solved by Ramo and Whinnery (3), the maximum rms field intensity of a center-fed conductor in free space:

$$E_0 = \frac{\left[\dfrac{\cos(G\sin\theta) - \cos G}{\cos\theta}\right]\sqrt{\dfrac{PZ_c}{2\pi d^2}}}{\{\gamma + \ln(2G) - Ci(2G) + 0.5[Si(4G) - 2\sin(2G)]\sin(2G) + 0.5[\gamma + \ln G - 2Ci(2G) + Ci(4G)]\cos(2G)\}^{1/2}} \tag{30}$$

where γ is the Euler's Constant = 0.57721566, Ci is the cosine integral function, Si is the sine integral function, and $2G$ is the length of the conductor.

For 1 kW of power at a distance of 1 km and defining the angle of radiation as well as the electrical length of the radiating element,

$$\theta = 0^\circ \quad \text{and} \quad G = 90^\circ$$

Equation (30) yields a maximum field intensity of

$$E_0 = 221.78 \text{ mV/m}$$

Vertical Conductor in Half-Space. The final step in this process is to determine the maximum field intensity of a vertical conductor in half-space (a monopole over perfecting conducting ground plane of infinite extent). Again, the limits of integration change such that Eq. (29) becomes

$$P = (1/Z_c)\int_0^{2\pi}\int_0^{+\pi/2}\left\{E_0\left[\frac{\cos(G\sin\theta - \cos G)}{(1-\cos G)\,\cos\theta}\right]\right\}^2 d^2\cos\theta\, d\theta\, d\phi \tag{31}$$

It follows from the previous analysis that solving for Eq. (31) yields

$$E_0 = \frac{\left(\dfrac{\cos(G\sin\theta) - \cos G}{\cos\theta}\right)\sqrt{\dfrac{PZ_c}{2\pi d^2}}}{\{\gamma + \ln(2G) - Ci(2G) + 0.5[Si(4G) - 2\sin(2G)]\sin(2G) + 0.5[\gamma + \ln G - 2Ci(2G) + Ci(4G)]\cos(2G)\}^{1/2}} \quad (32)$$

For 1 kW of power at a distance of 1 km, and

$$\theta = 0^\circ \quad \text{and} \quad G = 90^\circ$$

Equation (32) yields a maximum field intensity of

$$E_0 = 313.66 \text{ mV/m}$$

Pattern Synthesis. The antenna designer is required to fit a pattern within a given set of radiation limits as defined by the allocation studies. As this defines the general shape and size of the pattern, a set of field parameters must be chosen with regard to the number, height, and physical orientation of the towers. Useful techniques have been developed to synthesize the design of antenna patterns. The general expression given in Eq. (1) for calculating the pattern shape is simplified for a two-element array, which is used as the basic building block for pattern synthesis.

Using element 1 as the reference, the field ratio F_2 can be defined as

$$F_2 = E_1/E_2 \quad (33)$$

The difference in phase angle, α_2 has two components:

$$\alpha_2 = S\cos\phi\cos\theta + \Psi_2 \quad (34)$$

The first term of Eq. (34) relates to the space phase difference, and the second term relates to the time phase difference between E_1 and E_2.

If the towers are of equal height, it can be shown that the total field is

$$E = E_1 f(\theta)\sqrt{2F_2}\left[\frac{1+F_2^2}{2F_2} + \cos(S\cos\phi\cos\theta + \Psi_2)\right]^{1/2} \quad (35)$$

where

$$f(\theta) = f_1(\theta) = f_2(\theta)$$

If $F_2 = 1$ and $f(0) = 1$, Eq. (35) further reduces to

$$E = 2E_1 \cos\left(\frac{S}{2}\cos\phi + \frac{\Psi_2}{2}\right) \quad (36)$$

It follows from Eq. (36) that nulls occur in the pattern when

$$S\cos\phi + \Psi_2 = \pm 180^\circ \quad (37)$$

Any number of horizontal plane patterns from Eq. (36) can be generated by varying spacing and phase relationships of one element to the other. These patterns are then used as basic building blocks when computing patterns for multielement arrays. The most common technique for computing these patterns is known as pattern pair multiplication.

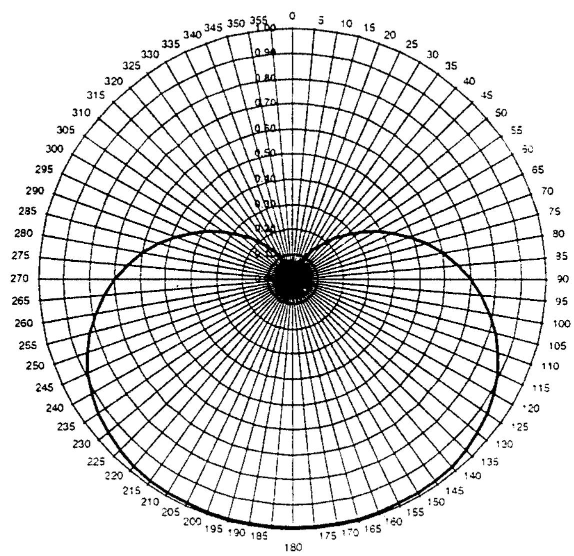

Figure 7. Horizontal pattern of a two-element array ($S = 90^\circ$ and $\Psi_2 = 90^\circ$). Pattern is broad with no radiation at 0°.

Pattern Pair Multiplication. Multielement arrays can be designed by multiplying the pattern of individual elements with the pattern of an array of vertical radiators having the same locations, relative amplitudes, and phases as the individual elements. As an example, take the two array patterns as shown in Figs. 7 and 8. Using Eq. (36) the array patterns can be expressed as follows:

$$E_a = 2E_{1a}\cos\left(\frac{\pi}{4}\cos\phi + \frac{\pi}{4}\right) \quad (38)$$

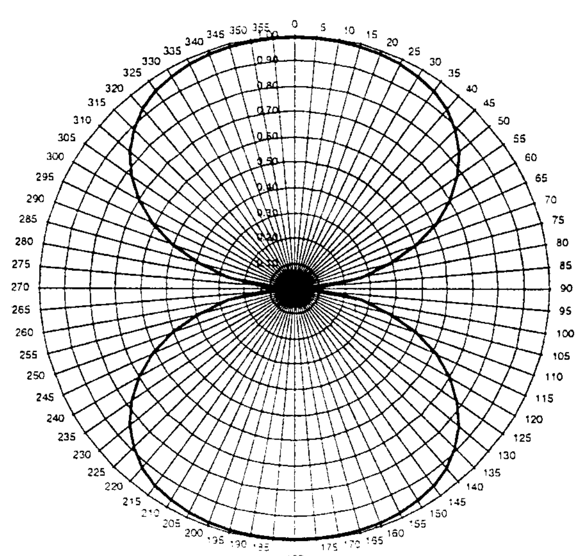

Figure 8. Horizontal pattern of a two-element array ($S = 180^\circ$ and $\Psi_2 = 180^\circ$). Pattern is symmetric with equal radiation at 0° and 180°.

when $S = \pi/2$ and $\Psi_2 = \pi/2$

$$E_b = 2E_{1b} \cos\left(\frac{\pi}{2}\cos\phi + \frac{\pi}{2}\right) \tag{39}$$

when $S = \pi$ and $\Psi_2 = \pi$

The resulting equation for the combined pattern is simply the product of individual arrays

$$E_a = 2E_{1a} \cos\left(\frac{\pi}{4}\cos\phi + \frac{\pi}{4}\right) 2E_{1b} \cos\left(\frac{\pi}{2}\cos\phi + \frac{\pi}{2}\right) \tag{40}$$

and Fig. 5 shows the combined pattern.

Array Simplification. The four-tower array as illustrated in the previous section can be simplified if equal spacing is maintained between the towers, which is frequently the case for medium-frequency in-line arrays used in broadcasting. Using Eqs. (38) and (39) the following field relationships can be defined as

$$F_a = 1.0 \quad \underline{/+90}$$
$$F_b = 1.0 \quad \underline{/+180}$$

The four-tower array can be reduced to a three-tower array using the following relations:

$$\text{Tower 3} = F_a \times F_b = 1.00 \quad \underline{/+270}$$
$$\text{Tower 2} = F_a \times F_b = 1.41 \quad \underline{/+135}$$
$$\text{Tower 1} = \text{Reference} = 1.00 \quad \underline{/+0}$$

The horizontal pattern for this set of parameters is shown in Fig. 9. Comparing Fig. 5 with Fig. 9 reveals little difference in pattern shape with the economical advantage of saving the cost of one tower.

Antenna Impedance. The base impedance defines the relationship of the voltage to the current in both magnitude and phase at the base of each radiating element. This is a complex quantity and is typically given in the following form:

$$Z_b = R_b + jX_b \tag{41}$$

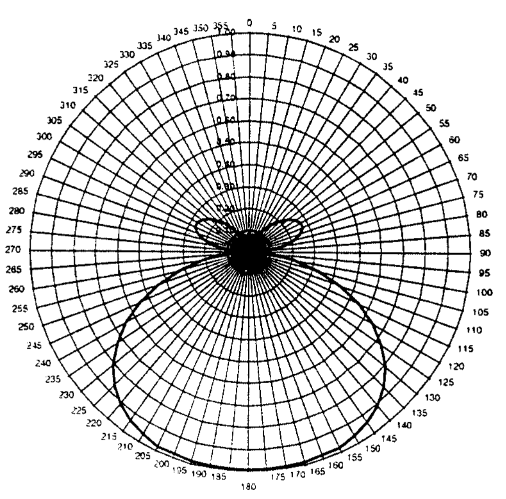

Figure 9. Horizontal pattern of a three-element array. Combined pattern produces greater directivity with no radiation at 0°.

where R_b is the base resistance (ohms) and X_b is the base reactance (Ω).

The base resistance, R_b, has two components

$$R_b = R_r + R_l \tag{42}$$

where R_r is the radiation resistance (Ω) and R_l is the loss resistance (Ω).

The radiation resistance determines the total power radiated from the antenna while the loss resistance takes into account all dissipative losses associated with the antenna and the ground system. It is the ratio of these two quantities that determines the efficiency of an antenna system. As the height of the radiating element decreases below 90 electrical degrees, the loss resistance becomes an appreciable percentage of the radiation resistance thus decreasing the overall efficiency of the antenna system.

It is important to take into account all series and shunt reactance found between the base of the antenna and the point at which the matching networks are connected to the antenna. Knowing an accurate impedance at this point of the antenna is very important when designing the feeder system for a multielement antenna system.

It follows then that the power radiated from a given antenna element is given as

$$P_r = I_b^2 R_r \tag{43}$$

where I_b is the base current (A).

Self Impedance Using Traditional Methods. The traditional method of determining the self-impedance (the impedance of a single radiating element apart from the influence of other radiating elements in close proximity) of a vertical radiating element uses the theory of nonuniform transmission lines as introduced by Schelkunoff (4). This method assumes a single radiator of uniform cross section over an infinite perfectly conducting ground plane. The first order approximation is given as

$$Z_b = Z_0 \left[\frac{A\sin G + j(B-C)\sin G - j(2Z_0 - D)\cos G}{(2Z_0 + D)\sin G + (B+C)\cos G - j(A\cos G)}\right] \tag{44}$$

where

$Z_b = R_b + jX_b$ base self-impedance (Ω)
$Z_0 = 60[\ln(2G/a) - 1]$ average characteristic impedance (Ω)
G = antenna height (degrees)
a = antenna radius (degrees)
$A = 60[\gamma + \ln(2G) - Ci(2G)] + 30[\gamma + \ln G - 2Ci(2G) + Ci(4G)]\cos(2G) + 30[Si(4G - 2Si(2G)]\sin(2G)$
$B = 60Si(2G) + 30[Ci(4G) - \ln G - \gamma]\sin(2G) - 30Si(4G)\cos(2G)$
$C = 60[Si(2G) - \sin(2G)]$
$D = 60[\ln(2G) - Ci(2G) + \gamma - 1 + \cos(2G)]$
γ = 0.5772 Euler's constant
Ci = cosine integral function
Si = sine integral function

Impedance of the Elements in Directional Array Using Traditional Methods. The impedance for an individual element of a directional array is not only dependent on its own current but also the current induced in it due to mutual coupling from other elements in the array. The relationship between the voltages and currents of the individual elements are given in terms of mutual impedance. The matrix of equations for a three-element array will be

$$V_1 = I_1Z_{11} + I_2Z_{21} + I_3Z_{31}$$
$$V_2 = I_1Z_{12} + I_2Z_{22} + I_3Z_{32}$$
$$V_3 = I_1Z_{13} + I_2Z_{23} + I_3Z_{33}$$

where V_1 and I_1 are the base voltage and current for element 1, Z_{11} is the self-impedance of element 1, and Z_{21} is the mutual impedance between element 1 and element 2

The values for mutual impedance as a function of element separation and height have been solved by Brown (5) and Cox (6). The equations are as follows:

$$Z_{21} = R_{21} + jX_{21} \tag{45}$$

where

$$\begin{aligned} R_{21} = \frac{15}{\sin\beta l_1 \sin\beta l_2}\{&\cos\beta\Delta[Ci(u_1) - Ci(u_0) + Ci(v_1) - Ci(v_0) \\ &+ 2Ci(y_0) - Ci(y_1) - Ci(s_1)] \\ &+ \sin\beta\Delta[Si(u_1) - Si(u_0) + Si(v_0) - Si(v_1) \\ &- Si(y_1) + Si(s_1)] \\ &+ \cos\beta L[Ci(w_1) - Ci(v_0) + Ci(x_1) - Ci(u_0) + 2Ci(y_0) \\ &- Ci(y_1) - Ci(s_1)] \\ &+ \sin\beta L[Si(w_1) - si(v_0) + Si(u_0) - Si(x_1) \\ &- Si(y_1) + Si(s_1)]\} \end{aligned} \tag{46}$$

$$\begin{aligned} X_{21} = \frac{15}{\sin\beta l_1 \sin\beta l_2}\{&\cos\beta\Delta[Si(u_0) - Si(u_1) + Si(v_0) - Si(v_1) \\ &+ Si(y_1) - 2Si(y_0) + Si(s_1)] \\ &+ \sin\beta\Delta[Ci(u_1) - Ci(u_0) + Ci(v_0) - Ci(v_1) \\ &- Ci(y_1) + Ci(s_1)] \\ &+ \cos\beta L[Si(v_0) - Si(x_1) + Si(u_0) - Si(x_1) \\ &- 2Si(y_0) + Si(y_1) + Si(s_1)] \\ &+ \sin\beta L[Ci(w_1) - Ci(v_0) + Ci(u_0) - Ci(x_1) \\ &- Ci(y_1) + Ci(s_1)]\} \end{aligned} \tag{47}$$

where l_1 and l_2 are the heights of elements 1 and 2, respectively, d is the distance between elements, and

$$\begin{aligned} L &= l_1 + l_2 \\ \Delta &= l_2 - l_1 \\ w_0 &= \beta[\sqrt{d^2 + l_1^2} + l_1] = v_0 \\ w_1 &= \beta[\sqrt{d^2 + L^2} + L] \\ v_1 &= \beta[\sqrt{d^2 + \Delta^2} - \Delta] \\ x_0 &= \beta[\sqrt{d^2 + l_1^2} - l_1] = u_0 \\ x_1 &= \beta[\sqrt{d^2 + L^2} - L] \\ u_1 &= \beta[\sqrt{d^2 + \Delta^2} + \Delta] \\ y_0 &= \beta d = s_0 \\ y_1 &= \beta[\sqrt{d^2 + l_2^2} + l_2] \\ s_1 &= \beta[\sqrt{d^2 + l_2^2} - l_2] \end{aligned}$$

Once the mutual impedances are known, the above matrix of equations is solved for the ratio of the voltage to the current in each element that defines the impedance of that element.

Operating impedances may be calculated with fair accuracy using the traditional method of calculating self- and mutual impedances as presented herein for tower heights up to approximately 120 electrical degrees. For taller towers, it has been standard practice to design matching units very conservatively and with components to provide a wide adjustment range. As presented in the next section, modern moment method analysis gives excellent results for all tower heights with properly chosen assumptions.

Impedance Determination Using Moment Methods. Moment method antenna modeling has proven to be a very useful tool in overcoming the limitations of traditional antenna theory. The moment method technique divides each radiator into a large number of individual segments for which corresponding current values can be calculated. In order for this technique to be useful in the design of a medium-frequency antenna system, it is necessary to relate the fields as produced by the antennas system to the drive point conditions of the antenna (voltage, current, and impedance).

A convenient method for specifying a medium-frequency directional antenna system uses field parameters that easily allow the designer to determine the radiation characteristics of any given antenna configuration. The field parameters for each tower in a directional antenna array are the ratios of the magnitudes and phases, relative to an arbitrary reference, of the electric field component of the radiation that results from integrating the current over the length of that particular tower or element of the directional antenna. Because field parameters are the standard method of specifying directional antennas, most notably with the FCC, it is necessary to relate these parameters to the driving point conditions (base voltage and current) in order to utilize modern moment method techniques to design these antennas. Once the driving point conditions are determined, the antenna feed system can be designed to provide the necessary power division and phase relationship between the elements in the array.

The means of exciting the antenna model with numerical electromagnetics code (NEC) (18) and MININEC (19) involves voltage sources. A problem involving a monopole over perfectly conducting ground plane excited with 1 + j0 volts at the base would yield the current distribution on the monopole and the fields, both electric and magnetic, produced by the monopole.

Field Parameters Versus Voltage Drives (21). The field parameters are calculated by ratioing the electric fields as produced by each element in the directional array to an arbitrary reference. The electric field produced by a finite current element over a perfectly conducting ground plane is proportional to the current in the wire and can be expressed as follows:

$$\overline{E} \propto \int_0^l \overline{I}\, dz \tag{48}$$

where $\overline{I}$ is the current distribution of the current element, dz is the incremental distance along the current element, and l is the length of the current element.

A close approximation for the solution of Eq. (48) is found using moment method techniques by summing the current moments of each element. The mathematical representation is

$$\overline{E} \propto \sum_{i=0}^{n} \overline{I}_i l_i \quad (49)$$

where $\overline{I}_i$ is the current in the ith segment and l_i is the length of the ith segment.

Four terminal network theory can now be used to relate the field parameters to the driving voltages for each tower in a directional array. Using a two-tower array as an example, the following set of equations is formulated:

$$\overline{E}_1 = \overline{T}_{11}\overline{V}_1 + \overline{T}_{12}\overline{V}_2$$
$$\overline{E}_2 = \overline{T}_{21}\overline{V}_1 + \overline{T}_{22}\overline{V}_2$$

where $\overline{E}_1$ is the field radiated from tower 1, $\overline{V}_1$ is the voltage drive of tower 1, $\overline{T}_{11}$ is the current moment summation of tower 1, and $\overline{T}_{12}$ is the current moment summation of tower 2 (as induced by the current in tower 1).

As can be seen from the above equations, to determine the T elements of the matrix, it is necessary to calculate the current summations by individually exciting each tower in the array. For example, to determine the elements, T_{11} and T_{21} it is necessary to excite tower 1 with voltage V_1 while grounding tower 2. The current moment summations are calculated for each tower. The same procedure is used to determine the elements, T_{21} and T_{22} by exciting tower 2 with voltage V_2 while grounding tower 1. Using matrix algebra, the drive point voltages can be determined from the field parameters by inverting the T matrix and multiplying by the field parameters.

$$[\overline{V}] = [\overline{F}][\overline{T}]^{-1} \quad (50)$$

where $[\overline{F}]$ is the set of field parameters as determined from the calculated electric fields and $[\overline{T}]^{-1}$ is the inverted current summation matrix.

The drive voltages for a given set of field parameters can now be determined. With these drive voltages the drive point currents and impedances are calculated, which determines the power division and phase relationships of the directional array elements.

It is possible to adjust an antenna system using moment method modeling with little, if any, experimentation, if the conditions at the site approach the ideal in terms of flat terrain and an absence of nearby reradiating structures. Even where conditions are not ideal, moment method modeling is a very useful tool in relating current drives to field parameters and reducing the amount of trial-and-error work necessary to achieve the required radiation pattern.

DIRECTIONAL ANTENNA FEEDER SYSTEMS

Once the base impedances of the individual elements in the antenna system have been calculated, it is possible to design the feeder system to provide the required current amplitude and phase for each tower in the array. Figure 10 shows a block diagram of the basic components that comprise a directional antenna feeder system.

Computer modeling techniques have been developed to analyze the feeder systems making it possible to obtain exact theoretical solutions for bandwidth analysis. The advance of computational capabilities has allowed the development of new approaches for the design of power dividing, phasing, and matching networks. The two areas of concern when designing the feeder system are the impedance and the pattern bandwidth which directly impact the quality of the audio within the entire coverage area.

Nodal Analysis

The technique of nodal analysis is well known in the field of electrical engineering. This technique works very well when predicting the bandwidth performance of directional antenna phasing and coupling equipment, since admittance values can be given for each component and the tower bases can be modeled as nodes with self- and mutual admittance values determined using moment method analysis. An exact solution for carrier and sideband currents and impedances can be found for every branch in a system. This solves the problem with simpler techniques that assume a set of base current parameters to determine operating impedances, which, when presented to the system of networks, yield a different set of base current parameters and render the starting assumptions invalid.

Power Dividing Circuits

Prior to the 1970s, two traditional circuits were used for the purpose of dividing the power between the towers of a directional antenna system in virtually all cases. The circuits are shown on Fig. 11. The first circuit is a *series* or *tank* type of power divider which goes back to the earliest days of radio and the *parallel* or *Ohms's law* design, which became popular during the 1950s.

Both circuits of Fig. 11 function primarily as power dividers, with separate networks necessary for phase adjustments.

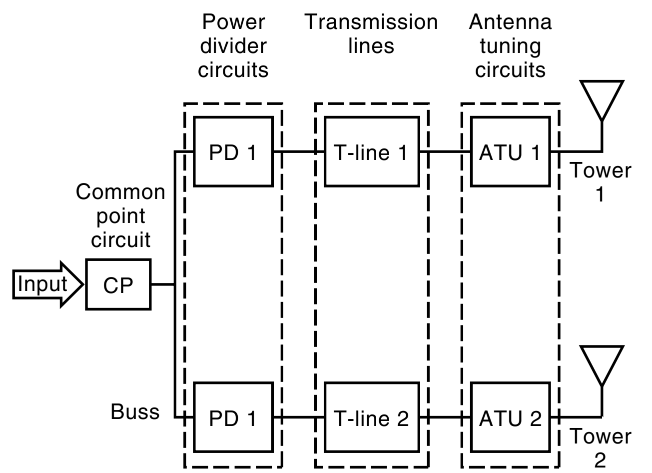

Figure 10. Basic components of a two-tower directional antenna feeder system. Additional towers can be added to the buss using similar networks.

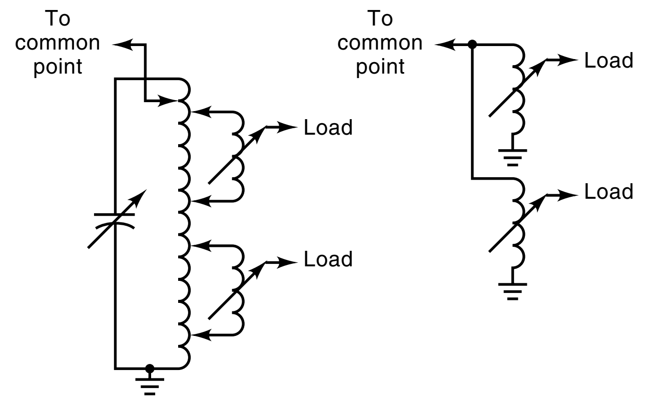

Figure 11. Traditional power divider circuits principally used in early medium-wave antenna designs.

Both can introduce high system Q, thus possibly restricting bandwidth. The series circuit circulates all of the power fed into the system through a parallel tuned antiresonant circuit, and the parallel circuit can result in relatively high circulating current due to the low resistance presented when several tower feeds are tied together across a common buss. The high Q of such circuits could serve to counteract bandwidth problems inherent in an array design. This would require careful system modeling to be effective and was not practical when such systems were built.

The general principle for all power divider circuits is illustrated in Fig. 12. If the common feed for all power dividing circuits is considered to be a voltage buss, the power delivered to each tower is determined by the conductance value presented to the buss by that tower's power dividing circuit. The voltage for the desired buss impedance can be determined and then the circuits necessary to present the required conductances, when terminated in the transmission lines, can be designed.

It is usually desirable to design for a buss impedance of 50 Ω when 50 Ω transmission lines are used, unless another factor suggests otherwise. Such an alternative situation would arise where one tower in a system needs much higher power than any of the others and could be fed directly off the buss without adjustment capability and satisfy the requirements

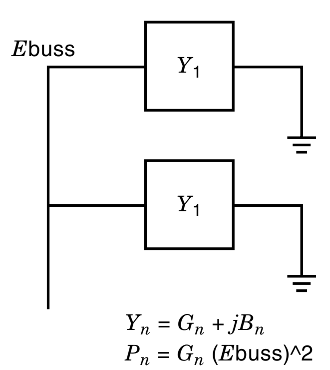

Figure 12. General power divider principle. The buss voltage is determined by the parallel combination of admittances produced by each tower and the input power.

for optimum overall phase shift. For example, a 25 Ω buss would feed half of its power directly to a 50 Ω transmission line.

Modern Power Divider Circuits. Any network that can adjust the conductance presented across the buss for a tower feed can be used as a power divider circuit. It is not necessary to have the same type of power divider network for every tower in an array. From the standpoint of adjustability and bandwidth, it is often desirable to have different types of intermixed networks in a given system.

Figures 13(a) through 13(f) show several power divider circuits. Each one shown is capable of serving for control of both power and phase, making separate phase adjustment networks unnecessary. If properly applied, the circuits of Fig. 13 can generally lead to lower power dividing and phasing network Q than attainable with either power divider from Fig. 11. Most of the circuits of Fig. 13 do not offer separate controls for both power and phase. This is not a great disadvantage, because the circuits that do only offer totally independent control when connected to load impedances that remain constant. This is not the case for any power divider that is feeding elements in an array because effects of mutual coupling between the elements make each tower's impedance change as the current flowing in the other towers change.

The low Q circuits of Fig. 13 are popular modern alternatives to the traditional power dividers of Fig. 11. For some directional arrays with highly volatile power division, however, the traditional power dividers, with their higher Q, may be desirable if easy adjustibility is important. Proper system modeling could be used to minimize the high Q effects or actually use them to improve overall system bandwidth.

Figures 14 through 17 show how the basic power divider circuits of Fig. 13 can be applied in phasing system design. Figure 14 offers good control and a 50-Ω buss, but can be simplified to Fig. 15 if the proper value is chosen for the power divider coil of the lowest power tower so that the capacitor necessary to antiresonate it is of the same reactance magnitude as the top tower's fixed L network shunt coil. This would be possible in a case where the lowest power tower would not change power flow direction. In the process, the power divider Q is lowered by the elimination of a parallel antiresonant circuit across the buss.

Figure 16 shows how, if the phase shift requirements allow it, the high-power tower feed can be connected directly to the buss, eliminating the three components of the L network. The circuit of Fig. 17 is identical to the circuit of Fig. 16, except that the buss has been divided with the series L–C slope network. In the case shown, the high-power towers need to have the phase shift of their feed tailored to track the lower-power tower in order to preserve pattern bandwidth. This is the purpose of the L–C slope network as shown.

As can be seen from the circuit of Fig. 17, high Q circuits can be inserted at appropriate locations in phasing equipment to effectuate broadbanding. Such processes require modeling of total system performance, such as with nodal analysis, in order to be effective. In many cases, it may be necessary to improve pattern bandwidth with high Q circuits added after the common buss, with an additional network to improve im-

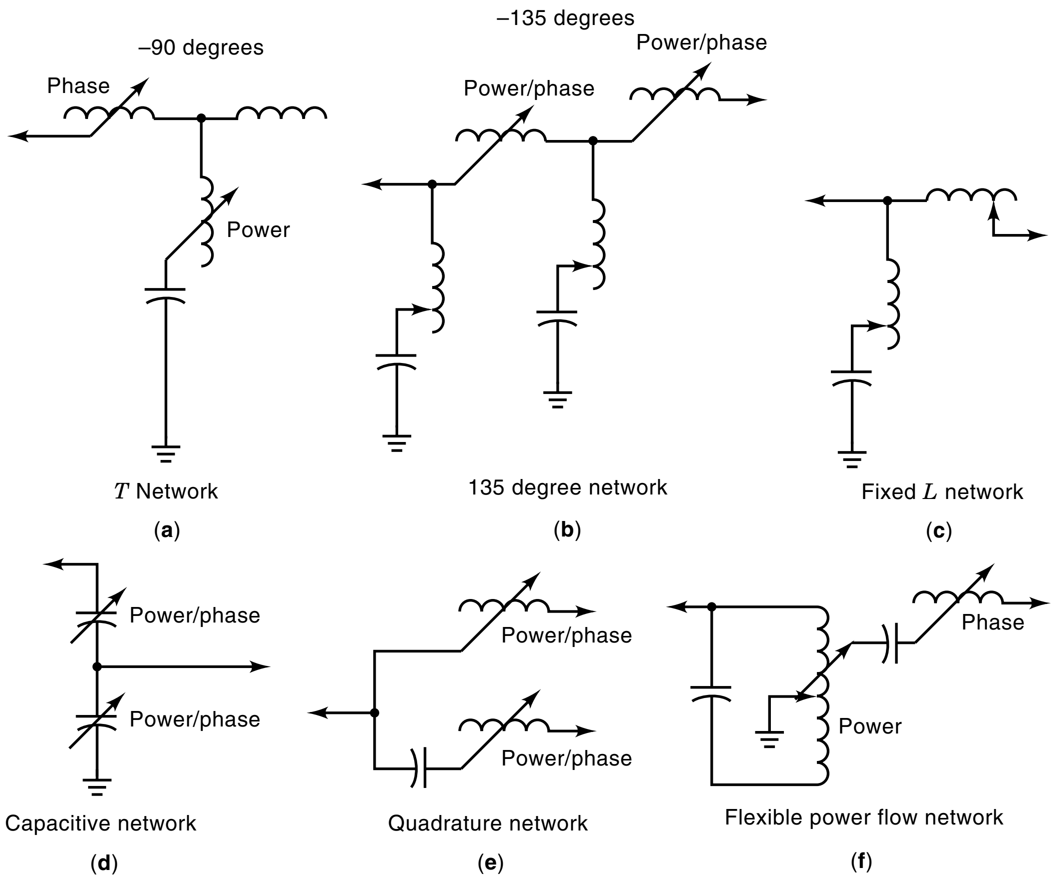

Figure 13. Modern power divider circuits. The selection of a particular circuit is dependent on overall system and load characteristics.

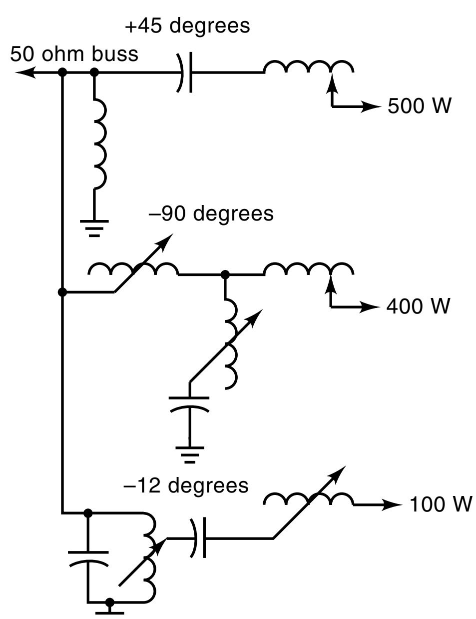

Figure 14. Mixed use of power divider circuits. Power division and system phase shifts determine the best combination of power divider circuits.

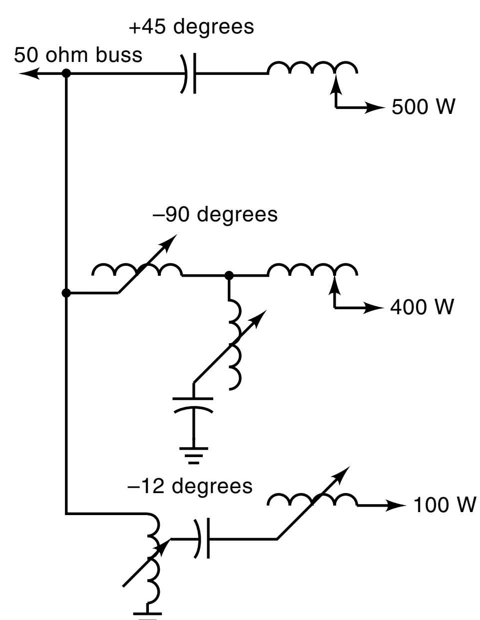

Figure 15. Simplification by elimination of parallel components as compared to Fig. 14.

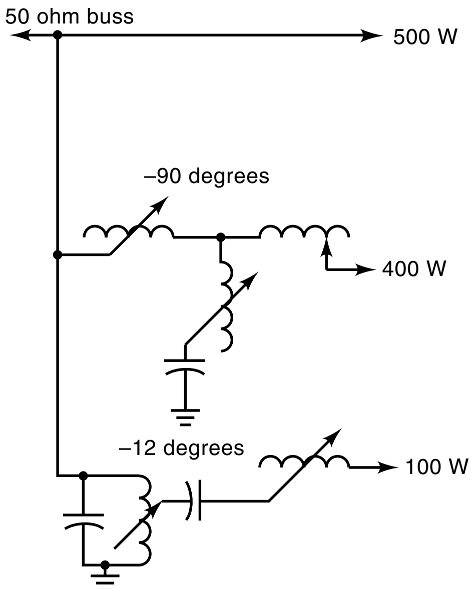

Figure 16. Direct feed to highest power tower. Another simplification which reduces the number of components.

pedance bandwidth included in the common point matching circuit.

Phasing and Matching Circuits

The conventional T network is the basic building block for antenna matching and phase shifting functions. Figure 18 shows the circuit for this network type. If the series input and output branches exhibit overall inductive reactance with the shunt branch having overall capacitive reactance, the network will produce negative phase shift and is said to be lagging. Conversely, if the series branches are capacitive and the shunt branch is inductive, the network will produce a positive phase shift and is said to be leading. The values chosen for each component determine the impedance transformation, the phase shift as well as the bandwidth performance.

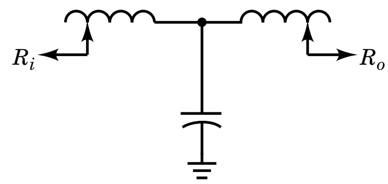

Figure 18. T network: basic circuit for impedance matching and phase shift. Circuit shown is for a phase-lagging network.

Referring to Fig. 18, the following equations can be used to determine the impedance transformation and phase shift of a T network.

$$X_1 = \frac{\sqrt{R_i R_o}}{\sin\beta} - \frac{R_i}{\tan\beta} \tag{51}$$

$$X_2 = \frac{\sqrt{R_i R_o}}{\sin\beta} - \frac{R_o}{\tan\beta} \tag{52}$$

$$X_3 = -\frac{\sqrt{R_i R_o}}{\sin\beta} \tag{53}$$

where R_i is the input resistance (ohms), R_o is the output resistance (Ω), and β is the phase angle.

Although conventional thinking would suggest that optimum bandwidth performance results with the phase shift of a T network adjusted to 90°, the family of curves on Fig. 19 indicate otherwise. There is an optimum T network phase shift for each transformation ratio and these values are generally lower than 90°.

Figure 19 also shows that the VSWR bandwidth worsens as the transformation ratio increases. The negative impact of transforming an impedance to one that is very much higher (or lower) can be lessened by cascading networks together. Figure 20 shows how two networks can be cascaded to achieve a gradual step-up of resistance that requires that only one additional shunt branch be added alongside the normal T network configuration. Figure 21 shows the bandwidth perfor-

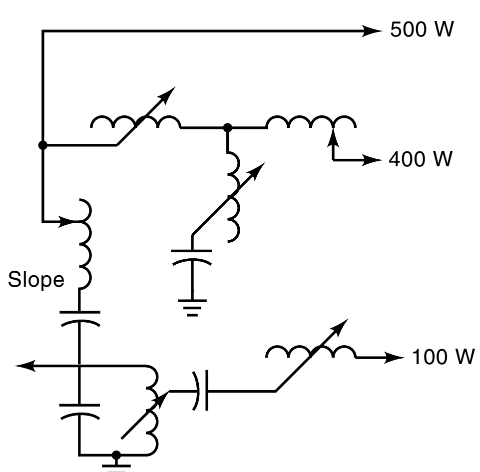

Figure 17. Split buss with pattern bandwidth improvement. Improved pattern bandwdith is most noticeable in the minima regions of radiation.

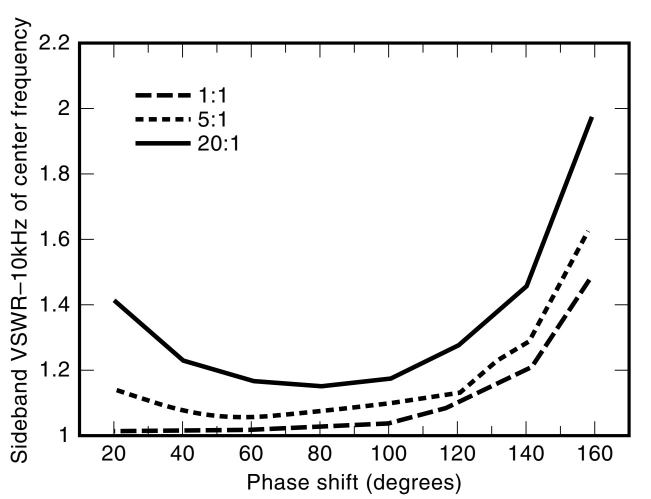

Figure 19. T network sideband VSWR versus phase shift for various transformation ratios. Smaller transformation ratios produce better sideband VSWR.

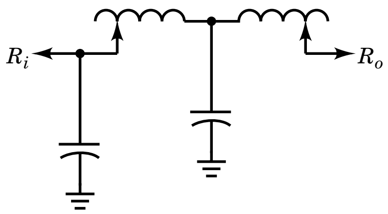

Figure 20. Cascaded T and L networks for optimizing phase shift and transformation ratio. Such a configuration results in better impedance bandwidth with fewer components.

mance of such a circuit designed with two cascaded networks. For the cost of an additional network branch, there is an approximate three to one improvement in sideband VSWR.

PROCUREMENT AND INSTALLATION

Once authorization has been received to construct a new or modify an existing medium-wave antenna system, it is necessary to prepare a request for quotation to procure the necessary equipment. The document necessary to receive a competitive quotation includes a detailed description of the required equipment usually in the form of a specification as well as a detailed statement of work that outlines all additional labor required outside of the manufacturing of the equipment. This usually includes any labor associated with installation and adjustments in the field. It is also important to specify the terms of a warranty if it is not already expressed in an off-the-shelf product. A line item must also be included for shipping.

It is recommended to request quotations from a number of reputable manufacturers to encourage a competitive bid. The evaluation of the bids should not only consider the price but also the quality of the product being proposed as well as how long it will take to deliver the product. It may be necessary to visit the prospective proposers' plants to ascertain their interest and commitment in providing you a quality product in a timely manner. Recommendations from others in the field of broadcasting are also invaluable when making a decision between quotations.

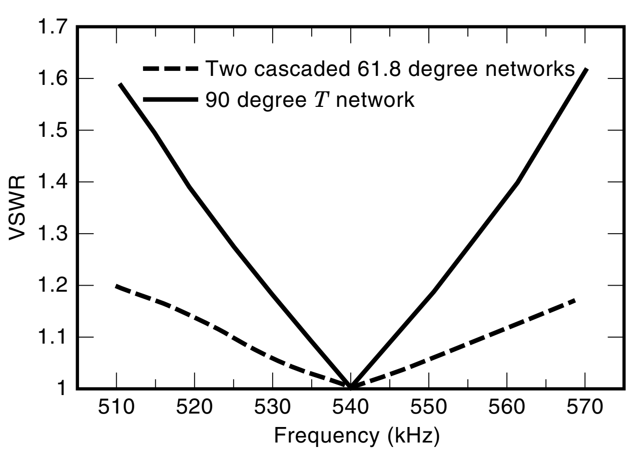

Figure 21. VSWR introduced by network for transformation ratio of 20 : 1. Cascaded networks produce significant improvement with sideband VSWR for high transformation ratios.

When overseeing the installation of a medium-wave antenna system there are a number of areas that one should pay special attention. For each tower, the connections of the ground system must be made and checked very carefully, since they will ultimately be below ground level and not visible. For directional antennas, the spacings and orientations of the elements must be carefully determined with reference to true north with a careful survey using celestial reference data.

The governing authorities of the country in which the new or modified medium-frequency antenna system is being constructed will usually require the system be tested to confirm compliance with the radiation characteristics as outlined in the construction permit. Once the system has been adjusted to theoretical parameters, a number of measurements are necessary to confirm compliance.

ANTENNA ADJUSTMENT

After the equipment has been installed and properly connected, the branches of the individual networks must be set to their design values. This involves using an impedance bridge and a frequency generator or a network analyzer to set the required reactance value at the operating frequency. Care must be taken when making these measurement to take into account all stray reactances that are inherent in an antenna system with long conductors between components (series inductive reactance) with close proximity to a grounded surface (shunt capacitive reactance). High-power systems, where physical dimensions are large, have very large stray impedances.

The electrical lengths of the sampling transmission lines must be measured to determine the values to which the antenna monitor will be adjusted. This is a critical step in the adjustment of the antenna system as errors interjected at this juncture will make it very difficult if not impossible to bring the antenna system into adjustment. A procedure has been developed for making such a measurement using an impedance bridge and a frequency generator by measuring the adjacent resonant frequencies of the line when it is in a short-circuited condition. The equation for calculating the length based on the two measurements is

$$\Psi_{\mathrm{L}} = \frac{180}{[(f_{\mathrm{H}}/f_{\mathrm{L}} - 1]} \tag{54}$$

where Ψ_{L} is the line length at lower frequency (degrees), f_{H} is the higher frequency (kHz), and f_{L} is the lower frequency (kHz).

Once the network branches have been set to their theoretical levels, low power can be applied to the system and the common point impedance adjusted to match the system to the transmitter. The phasor is then adjusted to bring the antenna monitor to the theoretical values as previously calculated. During this process it is oftentimes found that the adjustment of the ratio or phase to one tower will have an affect on the ratios and phases of other towers. The amount of interdependance between towers is determined by how closely the towers are coupled as well the component layout design within the phasor and ATU cabinets. While making these adjustments, the common point impedance must continually be readjusted to insure a proper load to the transmitter.

When the antenna monitor has been adjusted to theoretical parameters, a field strength meter is used to measure the radiated field levels at critical radials. The critical radials are usually located at the places where the pattern shape has inflections, that is, the pattern minima and minor lobe maxima. In the United States, the FCC requires that a number of measurements be made on each radial, ratioed to nondirectional reference measurements, averaged, and then multiplied by the measured unattenuated nondirectional measured field to determine the predicted level of radiation in a given direction. The inherent accuracy of a field intensity meter is largely dependent on the local environment in which a measurement is taken. Power lines and other reradiating structures necessitate that a number of measurements be taken to achieve reasonable results. The FCC requires at least twenty measurements be made on each radial between the distances of 2 and 20 miles. A number of close-in measurements, less than 2 miles, are also required for the nondirectional analysis.

At this point, there may be a one or more radials at which the measured radiation exceeds the maximum level specified in the construction permit. It may be necessary to adjust the antenna parameters away from their initial values to bring the pattern into compliance. A common approach used to bring a radiation pattern into compliance involves placing a number of field intensity monitors at locations on the critical radials and adjusting parameters until the pattern is in. While one person adjusts the parameters, the monitors at locations on the critical radials report variations in signal strength after each adjustment. Once the pattern appears to come into adjustment based on individual measurements at each point, the entire radial must be remeasured to confirm the adjustment. If the pattern is still out of adjustment, the procedure must be repeated. This method requires that the points monitored all represent their associated radials, a condition that often is not obtained, particularly with directional antenna patterns with deep radiation nulls. This method is generally a useful technique for patterns with minima that are not extremely deep, and where conductivity near the antenna is uniform and few reradiation sources affect the measurements.

Complex-Plane Mapping

Complex-plane mapping is an alternate approach which has been developed to adjust an antenna pattern. This technique is based on the knowledge that the field found at any point as produced by the antenna system is a vector quantity, having both magnitude and phase. In theory, the resultant vector field at any point in the far field can be calculated by adding the individual vectors as contributed from each radiating element in the array. It is theoretically straightforward to determine the change in the resultant vector at a point of interest when one or more of the ratios or phases of the radiating element is varied. In the real world, however, it is often impossible to accurately correlate the field strength measurements taken with a field intensity meter that measures only the magnitude of a signal with a theoretically determined resultant due to reradiation and variations in the ground characteristics.

The problem of how to determine the magnitude and phase of the resultant vector for each radial of interest is solved by making a series of trial measurements. First, a reference measurement is made by taking a sample of field strength readings at each radial of interest. These measurements are ratioed with the nondirectional measurements and for each critical measurement radial. Next, only one of the parameters of antenna is changed. Usually, the tower having the least interaction with the others in the array is chosen so that the adjustment is simplified. The magnitude is changed by an appropriate amount with all other parameters remaining the same and the field strength measurements are taken at the same points. Again the readings are ratioed with the nondirectional measurements and averaged. Finally, the magnitude of the tower that has changed is returned to the original value and the phase of that tower is changed by an appropriate amount. The measurements are retaken and analyzed as before. The results of all three trials (A, B, and C) are plotted on polar paper as circles with radii corresponding to the calculated averages for each measurement radial. The theoretical vectors for each radiating element are also plotted. Knowing that the field of the radiating element was changed both in magnitude (Trial B) and phase (Trial C), the delta (difference) vectors can be plotted. The magnitudes and phases of the delta vector gives the necessary information to determine the positions of the resultant vectors for the various radials for the beginning (Trial A) operating parameters.

Once the actual resultant vectors have been determined for each radial of interest, one is able to predict the impact of parameter variations to the individual radials themselves. In extremely difficult cases of signal scatter along measurement radials corresponding to deep radiation pattern nulls, it may be necessary to apply the complex plane mapping technique to individual measurement points rather than to entire radials in order to avoid analysis ambiguity. The pattern is brought into adjustment by making changes that will simultaneously change the field strengths at each radial in accordance with the previously calculated limits. Once the required adjustment has been determined and made, measurements are taken at all radials to access compliance. If compliance is confirmed, a full set of measurements is made to be filed with the proof-of-performance to be submitted to the proper governing authorities.

DETUNING TO CONTROL RERADIATION

Sometimes there are objects, usually other radio antenna towers or high-tension power lines, capable of scattering sufficient radiofrequency (RF) energy to distort a medium-wave radio station's antenna radiation pattern located near its transmitter site. Most such objects can be made transparent to the medium-wave RF energy if properly treated.

A tower located near a transmitter site may be detuned by installing the necessary apparatus to control its current distribution to minimize reradiation. For a short tower (shorter than one quarter wavelength), it is often sufficient to insulate its base or produce an impedance pole at its base with an arrangement of skirt wires and a detuning network.

For a taller tower or any critically located shorter tower, a null in tower current at a height somewhat above its base may be necessary for proper detuning. The correct treatment to produce the detuned condition and the corresponding current distribution for verification are best determined using the moment method directional antenna analysis procedures (with the field of the tower to be detuned set to zero) described herein.

Control over the tower current may be achieved by either placing a reactance across its base or tuning the open end of a wire skirt mounted on it. In general, it will be necessary to place the null in tower current at approximately one third of its height for towers up to approximately one half wavelength tall. For structures taller than one half wavelength, it may be necessary to produce more than one current distribution null. Two or more wire skirts may be required for this purpose.

Sampling loops may be mounted at the appropriate height or heights on a tower to verify the placement of current nulls corresponding to the detuned condition. They may be connected to detectors near the tower's base to facilitate observation during adjustment efforts and to verify continued detuning. For such sampling loops to be useful, they must not be mounted within a wire skirt span.

It is often sufficient to eliminate objectionable reradiation from a high tension power line by insulating the ground conductor running along the tops of adjacent support towers, thus breaking up the loops of current flowing in them. To maintain the integrity of the power line's protection system, insulators with arc gaps to conduct transient energy from the ground conductor to the support towers may be employed. When ground wire insulation is not sufficient to reduce high-tension power line reradiation to an acceptable level, individual towers may be detuned utilizing wire skirts and detuning networks to control their current distribution.

MULTIPLE FREQUENCY DIPLEXING

It is possible for a medium-wave antenna to radiate more than one frequency. Filters are employed to keep RF energy at each frequency out of a transmitter at the other frequency where a spurious signal might be generated.

For a nondirectional antenna, a series network providing an impedance zero at the desired frequency and an impedance pole at the undesired frequency and a shunt network providing an impedance pole at the desired frequency and an impedance zero at the undesired frequency to ground are normally placed at the feedpoint. For diplexed directional antennas, it is typical to have series filters at the tower bases but shunt filters only across the system input terminals (the common point).

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RONALD RACKLEY
MATTHEW FOLKERT
du Treil, Lundin, & Rackley

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ANTENNAS, LOADED. See DIELECTRIC-LOADED ANTENNAS.

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