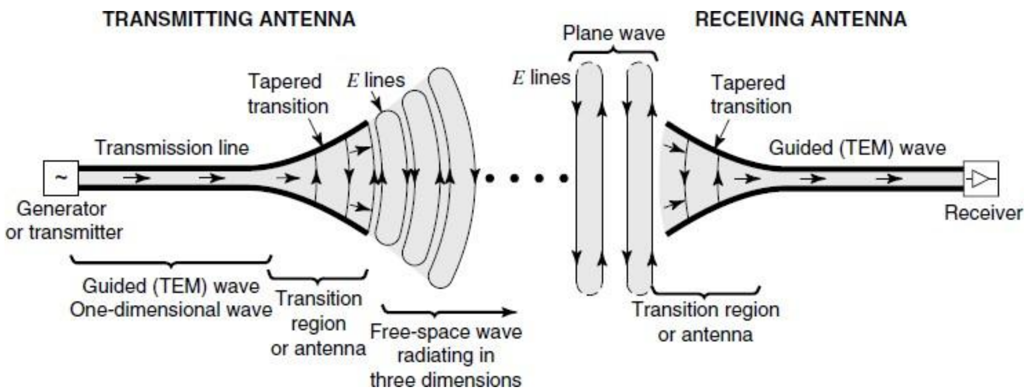
UNIT 1 FUNDAMENTALS OF ANTENNA

INTRODUCTION

An antenna is defined by Webster‘s Dictionary as ―a usually metallic device (as a rod or wire) for radiating or receiving radio waves.‖ The IEEE Standard Definitions of Terms for Antennas (IEEE Std 145–1983) defines the antenna or aerial as ―a means for radiating or receiving radio waves.‖ In other words the antenna is the transitional structure between free-space and a guiding device. The guiding device or transmission line may take the form of a coaxial line or a hollow pipe (waveguide), and it is used to transport electromagnetic energy from the transmitting source to the antenna or from the antenna to the receiver. In the former case, we have a transmitting antenna and in the latter a receiving antenna.



An antenna is basically a transducer. It converts radio frequency (RF) signal into an electromagnetic (EM) wave of the same frequency. It forms a part of transmitter as well as the receiver circuits. Its equivalent circuit is characterized by the presence of resistance, inductance, and capacitance. The current produces a magnetic field and a charge produces an electrostatic field. These two in turn create an induction field.

Definition of antenna

An antenna can be defined in the following different ways:

1. An antenna may be a piece of conducting material in the form of a wire, rod or any other shape with excitation.
2. An antenna is a source or radiator of electromagnetic waves.
3. An antenna is a sensor of electromagnetic waves.
4. An antenna is a transducer.
5. An antenna is an impedance matching device.
6. An antenna is a coupler between a generator and space or vice-versa.

Radiation Mechanism

The radiation from the antenna takes place when the Electromagnetic field generated by the source is transmitted to the antenna system through the Transmission line and separated from the Antenna into free space.

Radiation from a Single Wire

Conducting wires are characterized by the motion of electric charges and the creation of current flow. Assume that an electric volume charge density, q_v (coulombs/m^3), is distributed uniformly in a circular wire of cross-sectional area A and volume V.

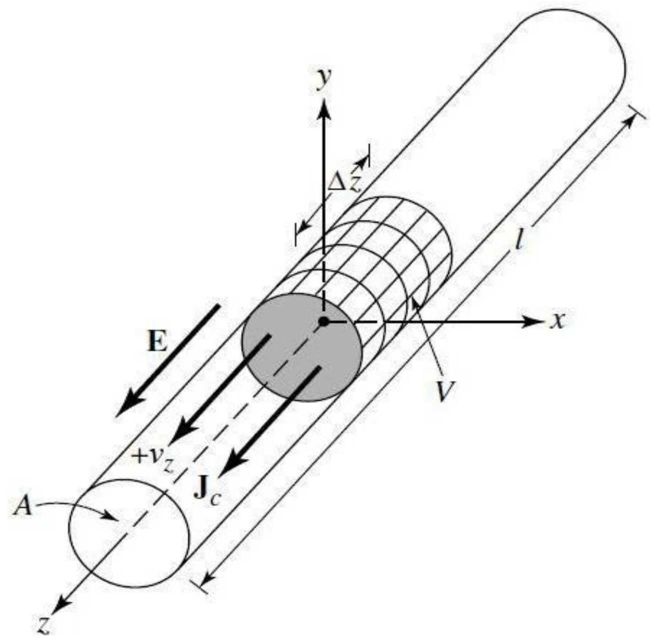


Figure: Charge uniformly distributed in a circular cross section cylinder wire.

Current density in a volume with volume charge density q_v (C/m^3)

$$J_z = q_v v_z \ (A/m^2) \qquad (1)$$

Surface current density in a section with a surface charge density qs (C/m^2)

$$J_s = q_s v_z \ (A/m) \qquad (2)$$

Current in a thin wire with a linear charge density q_l (C/m):

$$I_z = q_l v_z \ (A) \qquad (3)$$

To accelerate/decelerate charges, one needs sources of electromotive force and/or discontinuities of the medium in which the charges move. Such discontinuities can be bends or open ends of wires, change in the electrical properties of the region, etc.

In summary:

It is a fundamental single wire antenna. From the principle of radiation there must be some time varying current. For a single wire antenna,

1. If a charge is not moving, current is not created and there is no radiation.
2. If charge is moving with a uniform velocity:

a. There is no radiation if the wire is straight, and infinite in extent.

b. There is radiation if the wire is curved, bent, discontinuous, terminated, or truncated, as shown in Figure.

3. If charge is oscillating in a time-motion, it radiates even if the wire is straight.

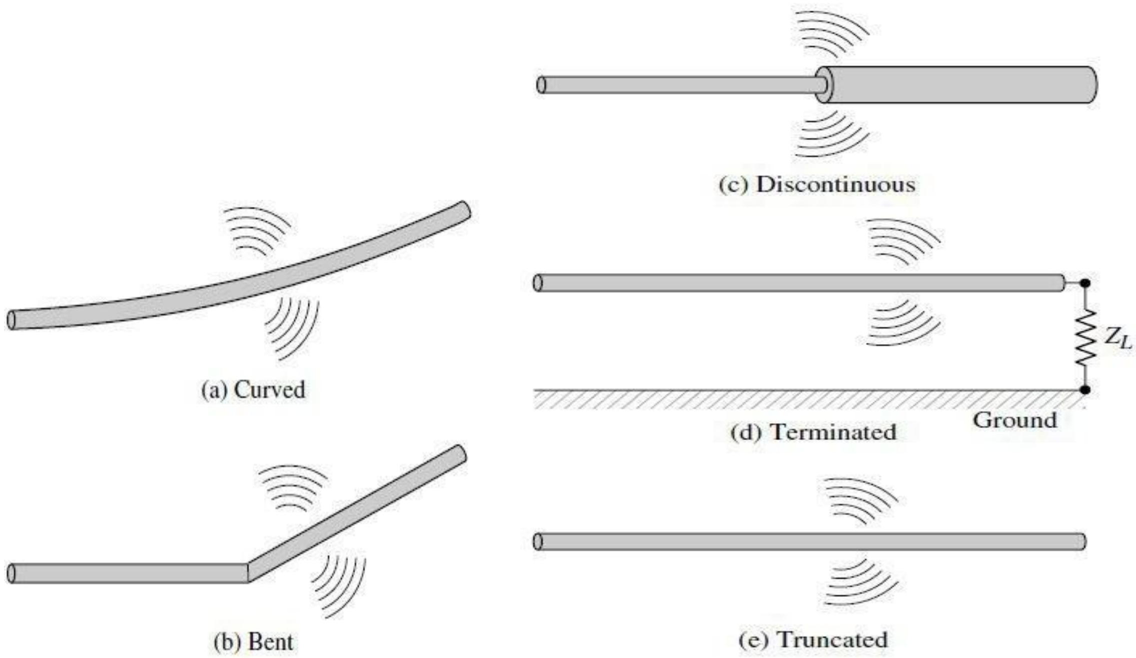


Figure : Wire Configurations for Radiation

<u>Radiation from a Two Wire</u>

Let us consider a voltage source connected to a two-conductor transmission line which is connected to an antenna. This is shown in Figure (a). Applying a voltage across the two conductor transmission line creates an electric field between the conductors. The electric field

has associated with it electric lines of force which are tangent to the electric field at each point and their strength is proportional to the electric field intensity. The electric lines of force have a tendency to act on the free electrons (easily detachable from the atoms) associated with each conductor and force them to be displaced. The movement of the charges creates a current that in turn creates magnetic field intensity. Associated with the magnetic field intensity are magnetic lines of force which are tangent to the magnetic field. We have accepted that electric field lines start on positive charges and end on negative charges. They also can start on a positive charge and end at infinity, start at infinity and end on a negative charge, or form closed loops neither starting or ending on any charge. Magnetic field lines always form closed loops encircling current-carrying conductors because physically there are no magnetic charges. In some mathematical formulations, it is often convenient to introduce equivalent magnetic charges and magnetic currents to draw a parallel between solutions involving electric and magnetic sources.

The electric field lines drawn between the two conductors help to exhibit the Distribution of charge. If we assume that the voltage source is sinusoidal, we expect the electric field between the conductors to also be sinusoidal with a period equal to that of the applied source. The relative magnitude of the electric field intensity is indicated by the density (bunching) of the lines of force with the arrows showing the relative direction (positive or negative). The creation of time-varying electric and magnetic fields between the conductors forms electromagnetic waves which travel along the transmission line, as shown in Figure 1.11(a). The electromagnetic waves enter the antenna and have associated with them electric charges and corresponding currents. If we remove part of b the antenna structure, as shown in Figure (b), free-space waves can be formed by ―connecting‖ the open ends of the electric lines (shown dashed). The free-space waves are also periodic but a constant phase point P0 moves outwardly with the speed of light and travels a distance of λ/2 (to P1) in the time of one-half of a period. It has been shown that close to the antenna the constant phase point P0 moves faster than the speed of light but approaches the speed of light at points far away from the antenna (analogous to phase velocity inside a rectangular waveguide).

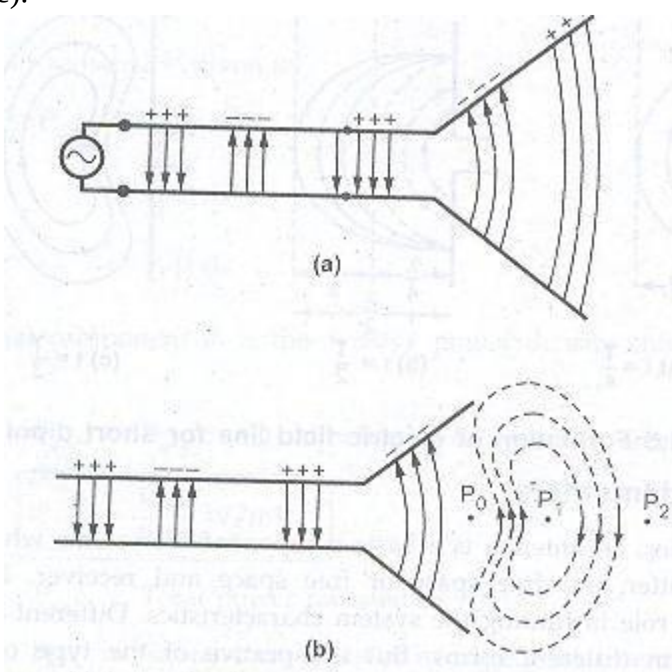


Fig. Radiation from two wire antennas

Radiation from a Dipole

Now let us attempt to explain the mechanism by which the electric lines of force are detached from the antenna to form the free-space waves. This will again be illustrated by an example of a small dipole antenna where the time of travel is negligible. This is only necessary to give a better physical interpretation of the detachment of the lines of force. Although a somewhat simplified mechanism, it does allow one to visualize the creation of the free-space waves. Figure(a) displays the lines of force created between the arms of a small center-fed dipole in the first quarter of the period during which time the charge has reached its maximum value (assuming a sinusoidal time variation) and the lines have traveled outwardly a radial distance λ/4. For this example, let us assume that the number of lines formed is three. During the next quarter of the period, the original three lines travel an additional λ/4 (a total of λ/2 from the initial point) and the charge density on the conductors begins to diminish. This can be thought of as being accomplished by introducing opposite charges which at the end of the first half of the period have neutralized the charges on the conductors. The lines of force created by the opposite charges are three and travel a distance λ/4 during the second quarter of the first half, and they are shown dashed in Figure (b). The end result is that there are three lines of force pointed upward in the first λ/4 distance and the same number of lines directed downward in the second λ/4. Since there is no net charge on the antenna, then the lines of force must have been forced to detach themselves from the conductors and to unite together to form closed loops. This is shown in Figure(c). In the remaining second half of the period, the same procedure is followed but in the opposite direction. After that, the process is repeated and continues indefinitely and electric field patterns are formed.

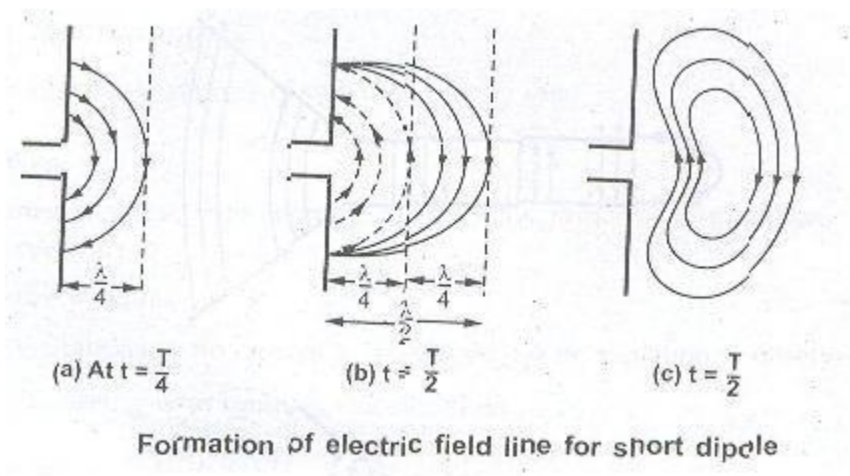
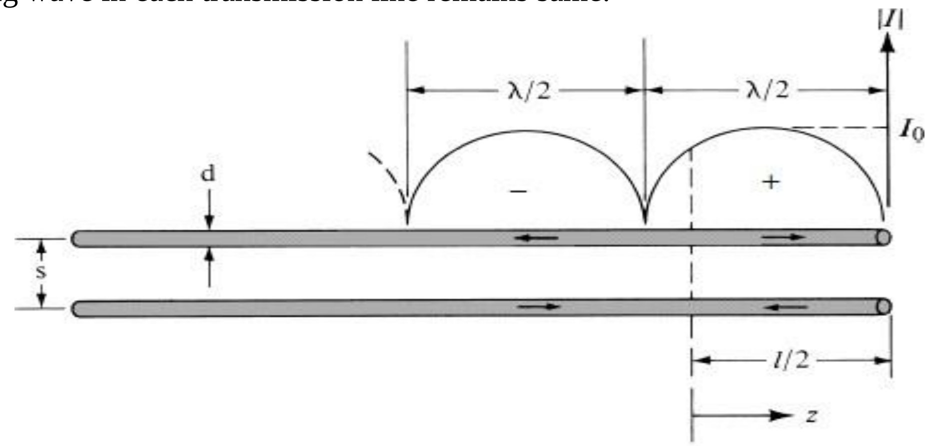


Fig. Formation of electric field line for short dipole

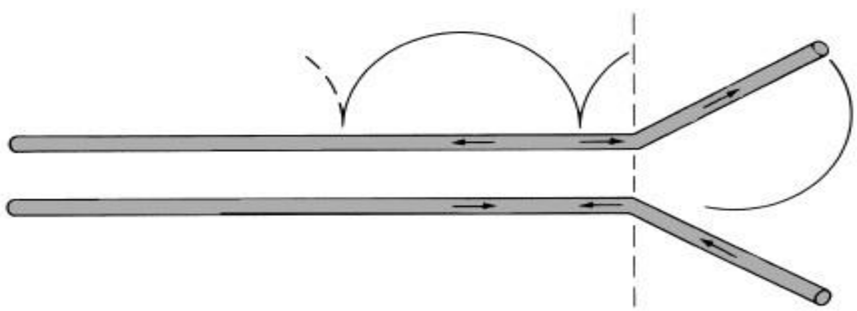
Current distribution on a thin wire antenna

Let us consider a lossless two wire transmission line in which the movement of charges creates a current having value I with each wire. This current at the end of the transmission line is reflected back, when the transmission line has parallel end points resulting in formation of standing waves in combination with incident wave.

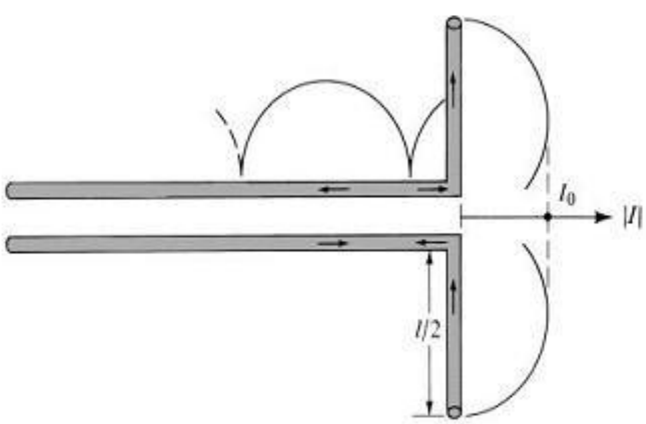
When the transmission line is flared out at 90^0 forming geometry of dipole antenna (linear wire antenna), the current distribution remains unaltered and the radiated fields not getting cancelled resulting in net radiation from the dipole. If the length of the dipole $l< \lambda/2$, the phase of current of the standing wave in each transmission line remains same.



(a) Two–wire transmission line



(b) Flared transmission line

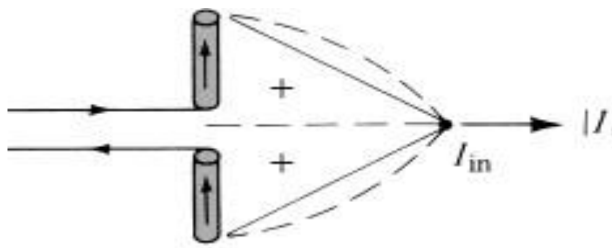


(c) Linear dipole

Fig. Current distribution on a lossless two-wire transmission line, flared transmission line, and linear dipole.

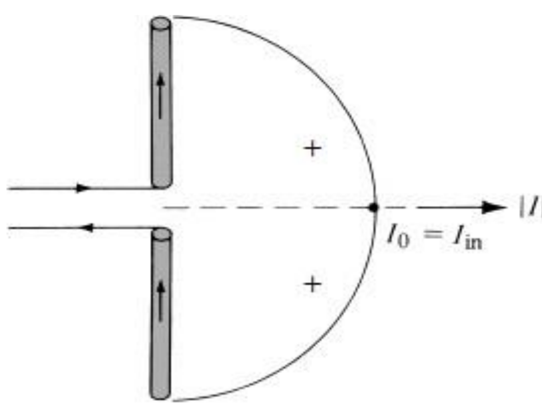
If diameter of each line is small d<< λ/2, the current distribution along the lines will be sinusoidal with null at end but overall distribution depends on the length of the dipole (flared out portion of the transmission line).

The current distribution for dipole of length *l* << λ



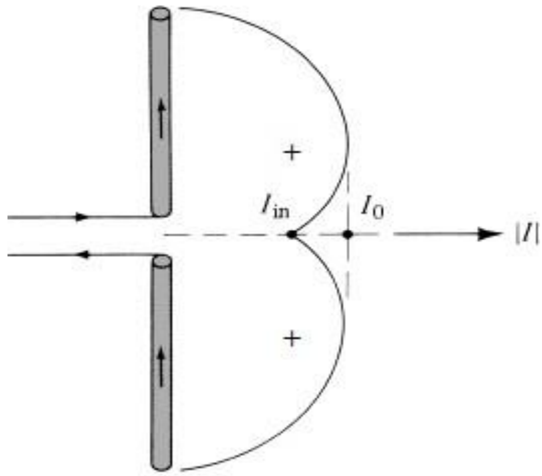
(a) $l << \lambda$

For *l*= λ /2



(b) $l = \lambda/2$

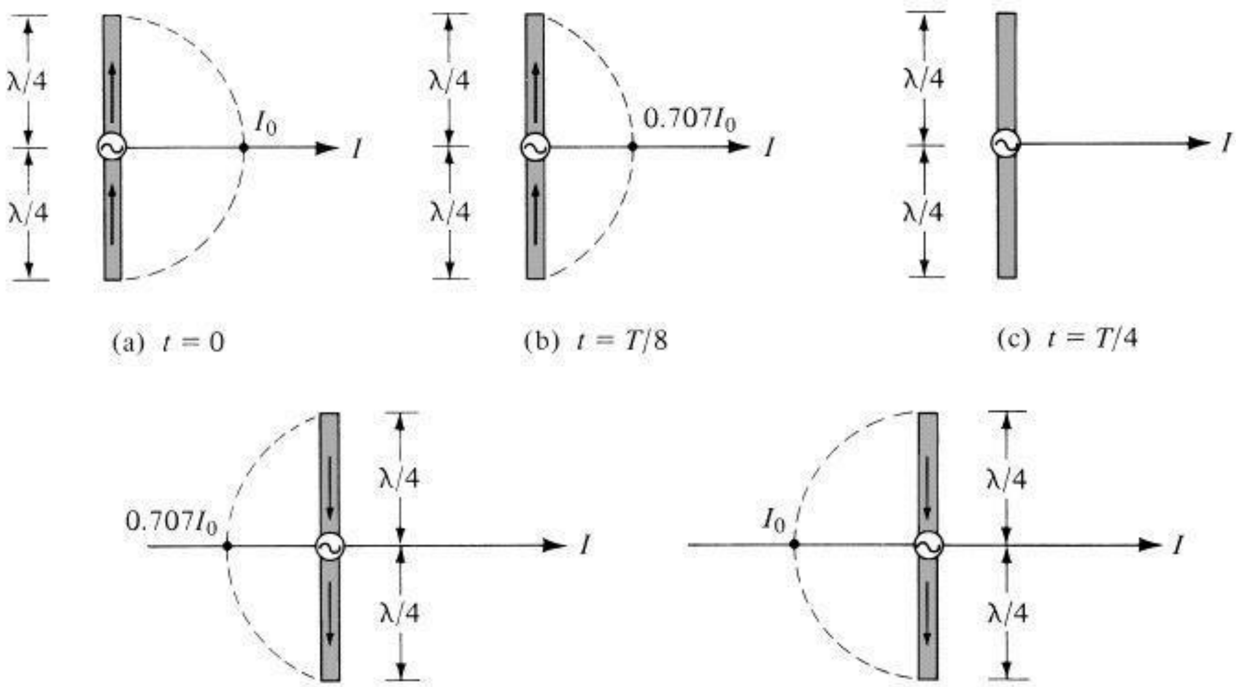
For λ /2<l< λ



(c) $\lambda/2 < l < \lambda$

When $l > \lambda$, the current goes phase reversal between adjoining half-cycles. Hence, current is not having same phase along all parts of transmission line. This will result into interference and canceling effects in the total radiation pattern.

The current distributions we have seen represent the maximum current excitation for any time. The current varies as a function of time as well.



(a) $t = 0$ (b) $t = T/8$ (c) $t = T/4$

(d) $t = 3T/8$ (e) $t = T/2$

ANTENNA PARAMETERS

INTRODUCTION:

To describe the performance of an antenna, definitions of various parameters are necessary. Some of the parameters are interrelated and not all of them need be specified for complete description of the antenna performance.

RADIATION PATTERN

An antenna radiation pattern or antenna pattern is defined as ―a mathematical function or a graphical representation of the radiation properties of the antenna as a function of space coordinates. In most cases, the radiation pattern is determined in the far field region and is represented as a function of the directional coordinates. Radiation properties include power flux density, radiation intensity, field strength, directivity, phase or polarization.‖ The radiation property of most concern is the two- or three dimensional spatial distribution of radiated energy as a function of the observer‘s position along a path or surface of constant radius. A convenient set of coordinates is shown in Figure 2.1. A trace of the received electric (magnetic) field at a constant radius is called the amplitude field pattern. On the other hand, a graph of the spatial variation of the power density along a constant radius is called an amplitude power pattern.

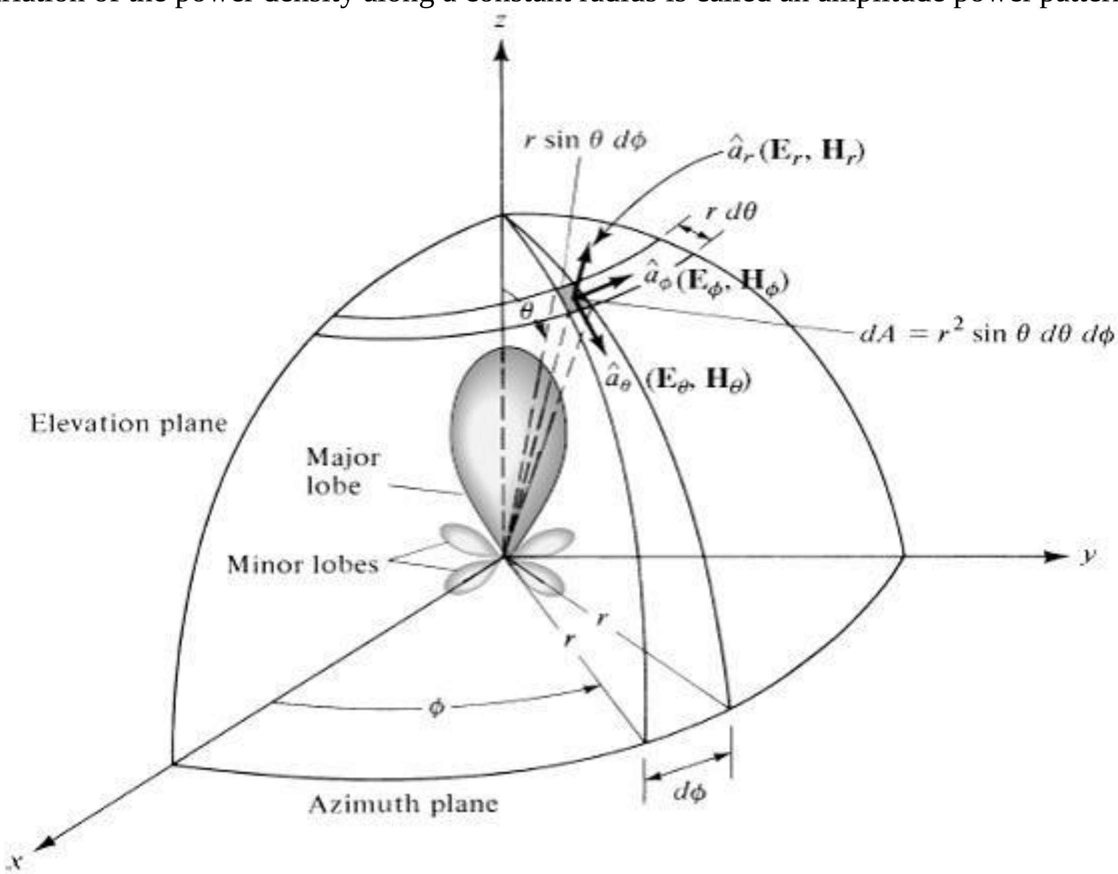


Fig. Coordinate system for antenna analysis

Often the field and power patterns are normalized with respect to their maximum value, yielding normalized field and power patterns. Also, the power pattern is usually plotted on a logarithmic scale or more commonly in decibels (dB). This scale is usually desirable because a logarithmic scale can accentuate in more details those parts of the pattern that have very low values, which later we will refer to as minor lobes.

For an antenna, the

a. field pattern(in linear scale) typically represents a plot of the magnitude of the electric or magnetic field as a function of the angular space.

b. power pattern(in linear scale) typically represents a plot of the square of the magnitude of the electric or magnetic field as a function of the angular space.

c. power pattern(in dB) represents the magnitude of the electric or magnetic field, in decibels, as a function of the angular space.

Below Figures a,b are principal plane field and power patterns in polar coordinates. The same pattern is presented in Fig.c in rectangular coordinates on a logarithmic, or decibel, scale which gives the minor lobe levels in more detail.

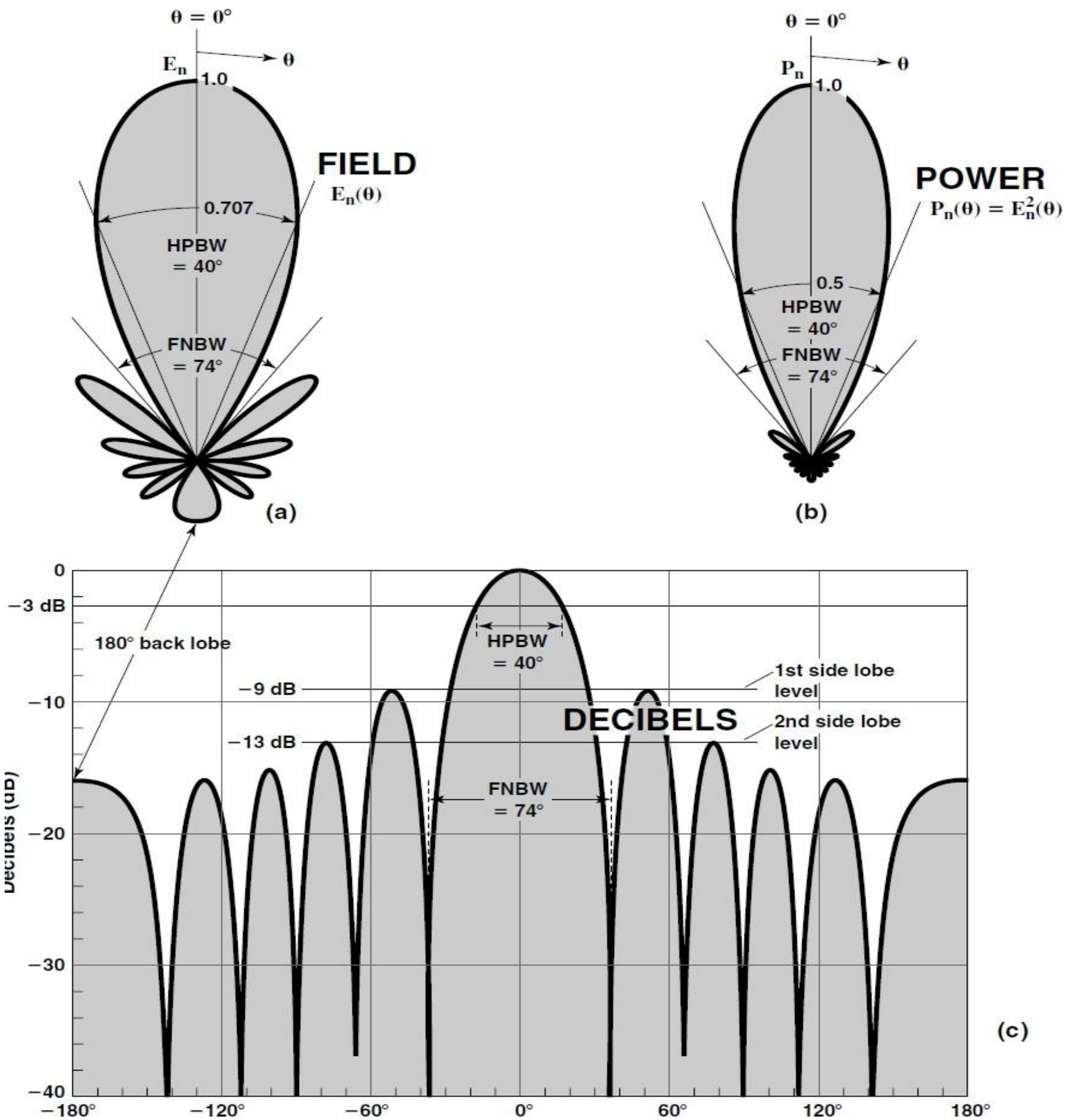
The angular beamwidth at the half-power level or half-power beamwidth (HPBW) (or −3-dB beamwidth) and the beamwidth between first nulls (FNBW) as shown in Fig. ,are important pattern parameters.

Dividing a field component by its maximum value, we obtain a *normalized or relative field pattern* which is a dimensionless number with maximum value of unity

$$\textit{Normalized field pattern} = E_\theta(\theta,\phi)_n = \frac{E_\theta(\theta,\phi)}{E_\theta(\theta,\phi)_{\max}} \qquad \text{(dimensionless)}$$

The half-power level occurs at those angles θ and φ for which $E_\theta(\theta, \varphi)_n = 1/\sqrt{2} = 0.707$.

At distances that are large compared to the size of the antenna and large compared to the wavelength, the shape of the field pattern is independent of distance. Usually the patterns of interest are for this far-field condition. Patterns may also be expressed in terms of the power per unit area [or Poynting vector S(θ, φ)]. Normalizing this power with respect to its maximum value yields a normalized power pattern as a function of angle which is a dimensionless number with a maximum value of unity.



Isotropic, Directional, and Omni directional Patterns:

An isotropic radiator is defined as ―a hypothetical lossless antenna having equal radiation in all directions.‖ Although it is ideal and not physically realizable, it is often taken as a reference for expressing the directive properties of actual antennas. A directional antenna is one ―having the property of radiating or receiving electromagnetic waves more effectively in some directions than in others. This term is usually applied to an antenna whose maximum directivity is significantly greater than that of a half-wave dipole.‖ Examples of antennas with directional radiation patterns are shown in Figures 2.5 and 2.6. It is seen that the pattern in Figure 2.6 is non directional in the azimuth plane [f (ϕ), $\theta = \pi/2$] and directional in the elevation plane [g(θ),

ϕ = constant]. This type of a pattern is designated as Omni directional, and it is defined as one ―having an essentially non directional pattern in a given plane (in this case in azimuth) and a directional pattern in any orthogonal plane (in this case in elevation).‖ An Omni directional pattern is the special type of a directional pattern.

Principal Patterns

For a linearly polarized antenna, performance is often described in terms of its principal E- and H-plane patterns. The E-plane is defined as ―the plane containing the electric field vector and the direction of maximum radiation,‖ and the H-plane as ―the plane containing the magnetic-field vector and the direction of maximum radiation.‖ Although it is very difficult to illustrate the principal patterns without considering a specific example, it is the usual practice to orient most antennas so that at least one of the principal plane patterns coincide with one of the geometrical principal planes. An illustration is shown in Figure 2.5. For this example, the x-z plane (elevation plane; $\varphi = 0$) is the principal E-plane and the x-y plane (azimuthal plane; $\theta = \pi/2$) is the principal H-plane. Other coordinate orientations can be selected. The omni directional pattern of Figure 2.6 has an infinite number of principal E-planes (elevation plan es; $\varphi = \varphi c$) and one principal H-plane (azimuthal plane; $\theta = 90^\circ$).

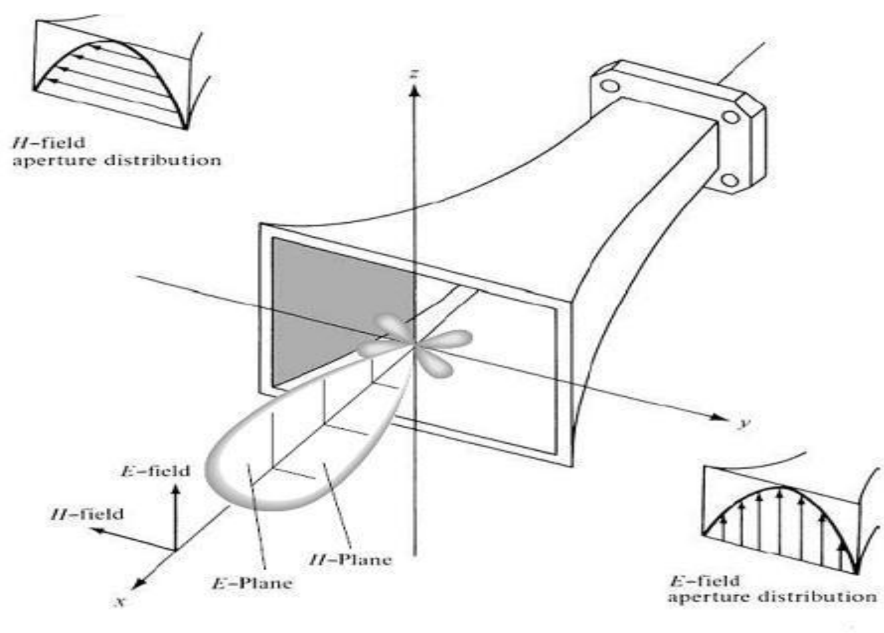


Fig. Principal E- and H-plane patterns for a pyramidal horn antenna

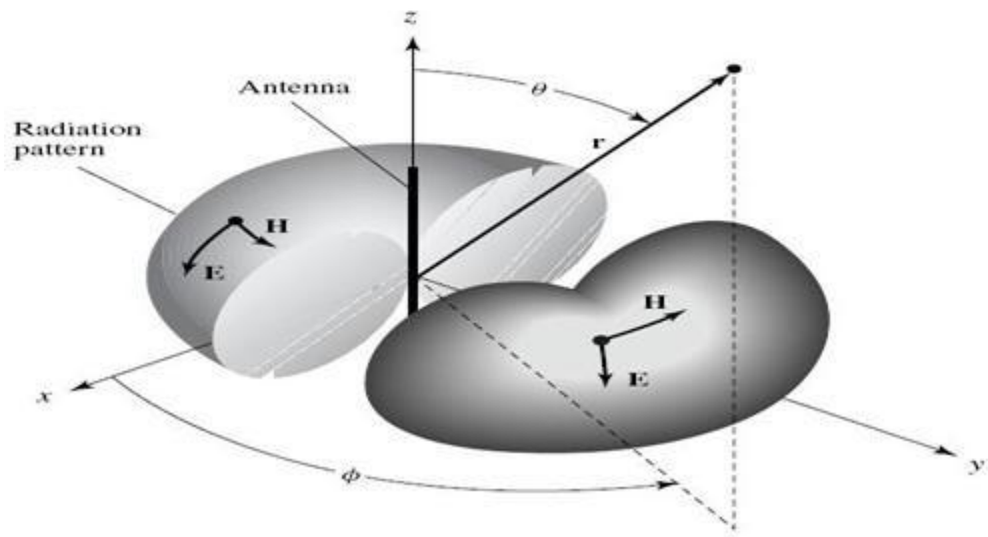


Fig. Omnidirectional antenna pattern

Radian and Steradian

The measure of a plane angle is a radian. One radian is defined as the plane angle with its vertex at the center of a circle of radius r that is subtended by an arc whose length is r. A graphical illustration is shown in Figure (a). Since the circumference of a circle of radius r is $C = 2\pi r$, there are 2π rad ($2\pi r/r$) in a full circle.

The measure of a solid angle is a steradian. One steradian is defined as the solid angle with its vertex at the center of a sphere of radius r that is subtended by a spherical surface area equal to that of a square with each side of length r. A graphical illustration is shown in Figure (b). Since the area of a sphere of radius r is $A = 4\pi r^2$, there are 4π sr ($4\pi r^2/r^2$) in a closed sphere.

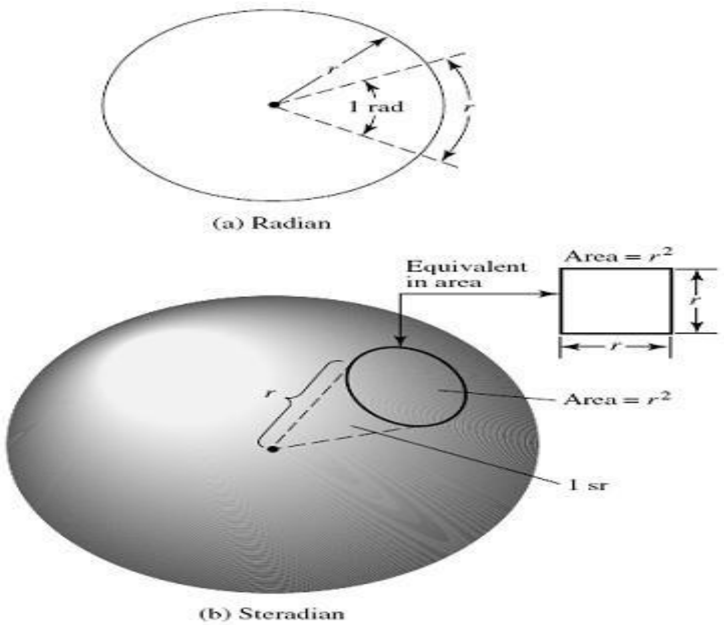


Fig. Geometrical arrangements for defining a radian and a steradian

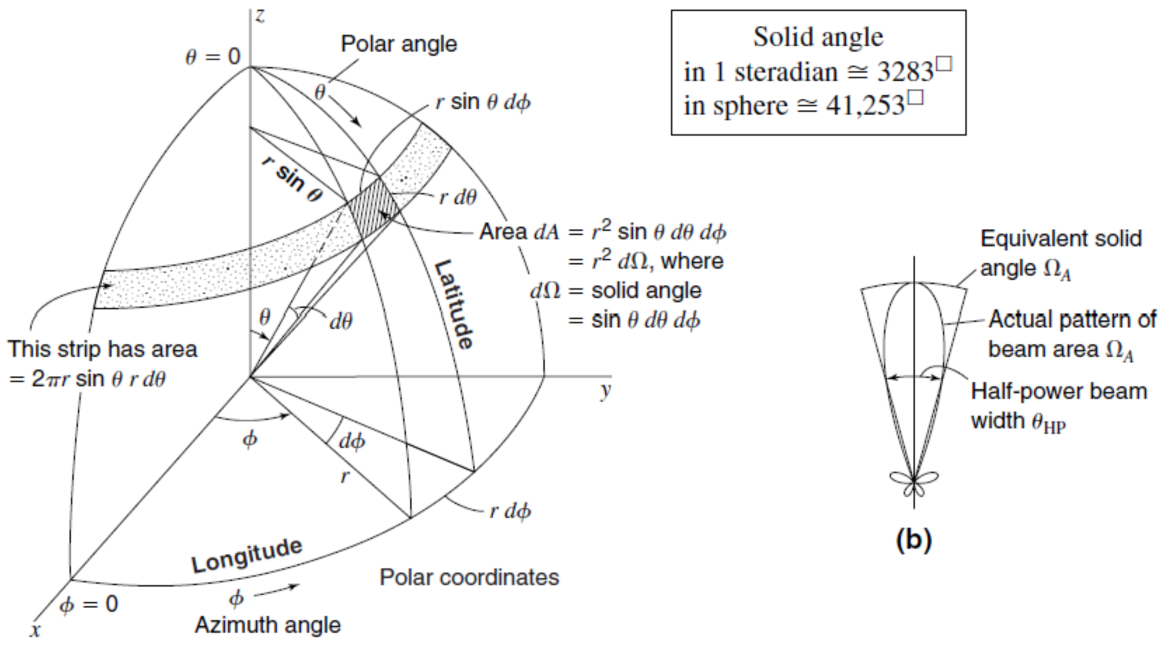
Although the radiation pattern characteristics of an antenna involve three-dimensional vector fields for a full representation, several simple single-valued scalar quantities can provide the information required for many engineering applications.
These are:
Half-power beamwidth, HPBW
Beam area, Ω_A
Beam efficiency, ε_M
Directivity *D* or gain *G*
Effective aperture *Ae*

<u>Beam Area (or beam solid angle):</u>

In polar two-dimensional coordinates an incremental area *dA* on the surface of a sphere is the product of the length $r\,d\theta$ in the θ direction (latitude) and $r \sin\theta\, d\varphi$ in the φ direction (longitude), as shown in Fig.
Thus,

$$dA = (r\,d\theta)(r \sin\theta\, d\varphi) = r^2\, d\Omega \qquad (1)$$

Where
$d\Omega$ = solid angle expressed in steradians (sr) or square degrees (□)
$d\Omega$ = solid angle subtended by the area dA



Polar coordinates showing incremental solid angle $dA = r^2\, d\Omega$ on the surface of a sphere of radius r where $d\Omega$ = solid angle subtended by the area dA. (b) Antenna power pattern and its equivalent solid angle or beam area Ω_A.

The area of the strip of width $r\,d\theta$ extending around the sphere at a constant angle θ is given by $(2\pi r \sin\theta)(r\,d\theta)$. Integrating this for θ values from 0 to π yields the area of the sphere. Thus,

$$\text{Area of sphere} = 2\pi r^2 \int_0^{\pi} \sin\theta\, d\theta = 2\pi r^2[-\cos\theta]_0^{\pi} = 4\pi r^2 \qquad (2)$$

where 4π = solid angle subtended by a sphere, sr

Thus,

1 steradian = 1 sr = (solid angle of sphere)/(4π)

= 1 rad² = $(180/\pi)^2(\text{deg}^2)$ = 3282.8064 square degrees (3)

therefore

4π steradians = 3282.8064 × 4π = 41,252.96 ~=41,253 square degrees = 41,253□ (4)

= solid angle in a sphere

The beam area or *beam solid angle* or Ω_A of an antenna (Fig. 2–5*b*) is given by the integral of the normalized power pattern over a sphere (4π sr)

$$\Omega_A = \int_{\phi=0}^{\phi=2\pi}\int_{\theta=0}^{\theta=\pi} P_n(\theta,\phi)\sin\theta\, d\theta\, d\phi \qquad (5a)$$

And

$$\Omega_A = \iint_{4\pi} P_n(\theta,\phi)\, d\Omega \quad (\text{sr}) \quad \textbf{\textit{Beam area}} \qquad (5b)$$

where d_ = sin θ $d\theta$ $d\varphi$, sr.
The beam area _*A* is the solid angle through which all of the power radiated by the antenna would stream if $P(\theta, \varphi)$ maintained its maximum value over Ω_A and was zero elsewhere. Thus the power radiated = $P(\theta, \varphi)\ \Omega_A$ watts.

The *beam area* of an antenna can often be described *approximately* in terms of the angles subtended by the *half-power points* of the main lobe in the two principal planes.
Thus,

$$\boxed{\textit{Beam area} \cong \Omega_A \cong \theta_{HP}\phi_{HP} \qquad (\text{sr})} \tag{6}$$

where θ_{HP} and φ_{HP} are the *half-power beamwidths* (HPBW) in the two principal planes, minor lobes being neglected.

RADIATION INTENSITY

The power radiated from an antenna per unit solid angle is called the *radiation intensity U* (watts per steradian or per square degree). The normalized power pattern of the previous section can also be expressed in terms of this parameter as the ratio of the radiation intensity $U(\theta, \varphi)$, as a function of angle, to its maximum value. Thus,

$$P_n(\theta, \phi) = \frac{U(\theta, \phi)}{U(\theta, \phi)_{\max}} = \frac{S(\theta, \phi)}{S(\theta, \phi)_{\max}}$$

Whereas the Poynting vector *S* depends on the distance from the antenna (varying inversely as the square of the distance), the radiation intensity *U* is independent of the distance, assuming in both cases that we are in the far field of the antenna.

BEAM EFFICIENCY

The (total) *beam area* Ω_A (or *beam solid angle*) consists of the main beam area (or solid angle) Ω_M plus the minor-lobe area (or solid angle) Ω_m. Thus,

$$\Omega_A = \Omega_M + \Omega_m$$

The ratio of the main beam area to the (total) beam area is called the (main) *beam efficiency* ε_M. Thus,

$$\boxed{\textit{Beam efficiency} = \varepsilon_M = \frac{\Omega_M}{\Omega_A} \qquad (\text{dimensionless})}$$

The ratio of the minor-lobe area (Ω_m) to the (total) beam area *is* called the *stray factor*.
Thus,

$$\varepsilon_m = \frac{\Omega_m}{\Omega_A} = \text{stray factor}$$

It follows that

$$\varepsilon_M + \varepsilon_m = 1$$

DIRECTIVITY *D* AND GAIN *G*

The directivity D and the gain G are probably the most important parameters of an antenna. The *directivity* of an antenna is equal to the ratio of the maximum power density $P(\theta, \varphi)_{max}$ (watts/m^2) to its average value over a sphere as observed in the far field of an antenna. Thus,

$$D = \frac{P(\theta,\phi)_{max}}{P(\theta,\phi)_{av}} \qquad \textit{Directivity from pattern} \tag{1}$$

The directivity is a dimensionless ratio ≥1.

The average power density over a sphere is given by

$$\begin{aligned} P(\theta,\phi)_{av} &= \frac{1}{4\pi}\int_{\phi=0}^{\phi=2\pi}\int_{\theta=0}^{\theta=\pi} P(\theta,\phi)\sin\theta\, d\theta\, d\phi \\ &= \frac{1}{4\pi}\iint_{4\pi} P(\theta,\phi)\, d\Omega \qquad (\text{W sr}^{-1}) \end{aligned} \tag{2}$$

Therefore, the directivity

$$D = \frac{P(\theta,\phi)_{max}}{(1/4\pi)\iint_{4\pi} P(\theta,\phi)\, d\Omega} = \frac{1}{(1/4\pi)\iint_{4\pi}[P(\theta,\phi)/P(\theta,\phi)_{max}]\, d\Omega} \tag{3}$$

And

$$D = \frac{4\pi}{\iint_{4\pi} P_n(\theta,\phi)\, d\Omega} = \frac{4\pi}{\Omega_A} \qquad \textit{Directivity from beam area } \Omega_A \tag{4}$$

where $P_n(\theta, \varphi)\, d\Omega = P(\theta, \varphi)/P(\theta, \varphi)_{max}$ = normalized power pattern

Thus, the directivity is the ratio of the area of a sphere (4π sr) to the beam area Ω_A of the antenna

The smaller the beam area, the larger the directivity D. For an antenna that radiates over only half a sphere the beam area $\Omega A = 2\pi$ sr in fig and the directivity is

$D = 4\pi/2\pi = 2$ (= 3.01 dBi) (5)

where dBi = decibels over isotropic

Note that the idealized *isotropic antenna* (_$A = 4\pi$ sr) has the lowest possible directivity $D = 1$. All actual antennas have directivities greater than 1 *(*$D > 1$*)*. The simple short dipole has a beam area _$A = 2.67\pi$ sr and a directivity $D = 1.5$ *(*= 1.76 dBi*)*.

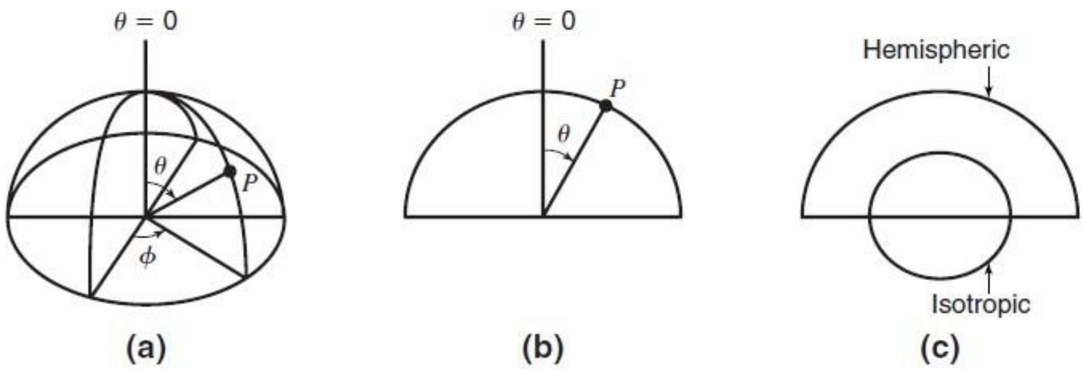


Figure 2–6
Hemispheric power patterns, (*a*) and (*b*), and comparison with isotropic pattern (*c*).

The *gain G* of an antenna is an actual or realized quantity which is less than the directivity *D* due to ohmic losses in the antenna or its radome (if it is enclosed). In transmitting, these losses involve power fed to the antenna which is not radiated but heats the antenna structure. A mismatch in feeding the antenna can also reduce the gain. The ratio of the gain to the directivity is the *antenna efficiency factor*. Thus,

$$G = kD \tag{6}$$

where k = efficiency factor ($0 < k < 1$), dimensionless.

In many well-designed antennas, k may be close to unity. In practice, G is always less than D, with D its maximum idealized value.

If the half-power beamwidths of an antenna are known, its directivity

$$D = \frac{41{,}253^{\square}}{\theta^{\circ}_{HP}\phi^{\circ}_{HP}} \tag{7}$$

\

where

$41{,}253^{\square}$ = number of square degrees in sphere = $4\pi(180/n)^2$ square degrees ($^{\square}$)
θ°_{HP} = half-power beamwidth in one principal plane
ϕ°_{HP} = half-power beamwidth in other principal plane

Since (7) neglects minor lobes, a better approximation is a

$$D = \frac{40{,}000^{\square}}{\theta^{\circ}_{HP}\phi^{\circ}_{HP}} \qquad \textit{Approximate directivity} \tag{8}$$

If the antenna has a main half-power beamwidth (HPBW) = 20° in both principal planes, its directivity

$$D = \frac{40{,}000^{\square}}{400^{\square}} = 100 \text{ or } 20\,\text{dBi} \qquad (9)$$

which means that the antenna radiates 100 times the power in the direction of the main beam as a non-directional, isotropic antenna.
If an antenna has a main lobe with both half-power beamwidths (HPBWs) = 20◦, its directivity from (7) is *approximately*

$$D = \frac{4\pi\,(\text{sr})}{\Omega_A\,(\text{sr})} \cong \frac{41{,}253(\text{deg}^2)}{\theta^\circ_{\text{HP}}\phi^\circ_{\text{HP}}} = \frac{41{,}253(\text{deg}^2)}{20^\circ \times 20^\circ}$$

$$\cong 103 \cong 20 \text{ dBi (dB above isotropic)}$$

which means that the antenna radiates a power in the direction of the main-lobe maximum which is about 100 times as much as would be radiated by a non-directional (isotropic) antenna for the same power input.

DIRECTIVITY AND RESOLUTION

The resolution of an antenna may be defined as equal to half the beam width between first nulls (FNBW)/2, for example, an antenna whose pattern FNBW = 2° has a resolution of 1° and, accordingly, should be able to distinguish between transmitters on two adjacent satellites in the Clarke geostationary orbit separated by 1°. Thus, when the antenna beam maximum is aligned with one satellite, the first null coincides with the adjacent satellite. Half the beamwidth between first nulls is approximately equal to the half-power beamwidth (HPBW) or

$$\frac{\text{FNBW}}{2} \cong \text{HPBW}$$

The product of the FNBW/2 in the two principal planes of the antenna pattern is a measure of the antenna beam area. Thus,

$$\Omega_A = \left(\frac{\text{FNBW}}{2}\right)_\theta \left(\frac{\text{FNBW}}{2}\right)_\phi$$

It then follows that the number N of radio transmitters or point sources of radiation distributed uniformly over the sky which an antenna can resolve is given approximately by

$$N = \frac{4\pi}{\Omega_A}$$

However

$$D = \frac{4\pi}{\Omega_A}$$

and we may conclude that *ideally* the number of point sources an antenna can resolve is numerically equal to the directivity of the antenna or

$$\boxed{D = N}$$

the directivity is equal to the number of point sources in the sky that the antenna can resolve under the assumed ideal conditions of a uniform source distribution.

ANTENNA APERTURES

The concept of aperture is most simply introduced by considering a receiving antenna. Suppose that the receiving antenna is a rectangular electromagnetic horn immersed in the field of a uniform plane wave as suggested in Fig. Let the Poynting vector, or power density, of the plane wave be *S* watts per square meter and the area, or physical aperture of the horn, be A_p square meters. If the horn extracts all the power from the wave over its entire physical aperture, then the total power *P* absorbed from the wave is

$$P = \frac{E^2}{Z} A_p = S A_p \qquad \text{(W)}$$

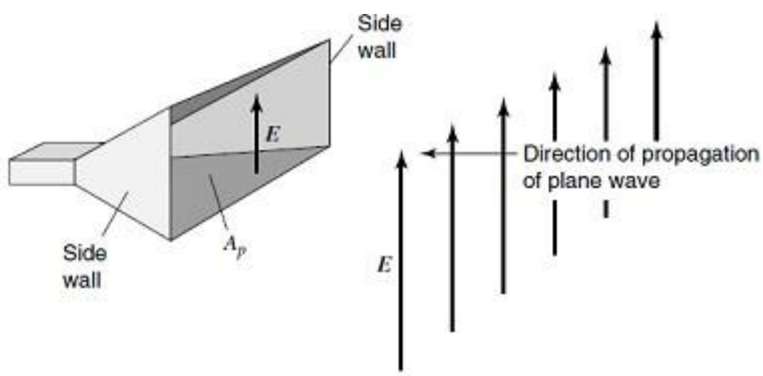


Figure
Plane wave incident on electromagnetic horn of physical aperture A_p.

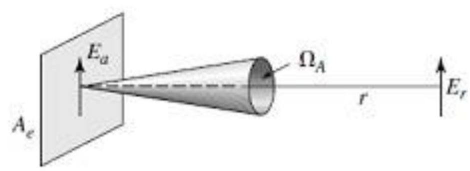


Figure
Radiation over beam area Ω_A from aperture A_e.

Thus, the electromagnetic horn may be regarded as having an aperture, the total power it extracts from a passing wave being proportional to the aperture or area of its mouth. But the field response of the horn is NOT uniform across the aperture A because E at the sidewalls must equal zero. Thus, the effective aperture A_e of the horn is less than the physical aperture A_p as given by

$$\varepsilon_{ap} = \frac{A_e}{A_p} \qquad \text{(dimensionless)} \qquad \textit{Aperture efficiency}$$

where ε_{ap} =aperture efficiency.

Consider now an antenna with an effective aperture A_e, which radiates all of its power in a conical pattern of beam area Ω_A, as suggested in above Fig. b. Assuming a uniform field Ea over the aperture, the power radiated is

$$P = \frac{E_a^2}{Z_0} A_e \qquad \text{(W)}$$

where Z_0 =intrinsic impedance of medium (377Ω for air or vacuum).
Assuming a uniform field E_r in the far field at a distance r, the power radiated is also given by

$$P = \frac{E_r^2}{Z_0} r^2 \Omega_A \qquad \text{(W)}$$

where Ω_A =beam area (sr).

$$\lambda^2 = A_e \Omega_A \qquad (\text{m}^2) \qquad \textit{Aperture–beam-area relation}$$

Thus, if A_e is known, we can determine Ω_A (or vice versa) at a given wavelength

$$D = 4\pi \frac{A_e}{\lambda^2} \qquad \textit{Directivity from aperture}$$

All antennas have an effective aperture which can be calculated or measured. Even the hypothetical, idealized isotropic antenna, for which $D = 1$, has an effective aperture

$$A_e = \frac{D\lambda^2}{4\pi} = \frac{\lambda^2}{4\pi} = 0.0796\lambda^2$$

All lossless antennas must have an effective aperture equal to or greater than this. By reciprocity the effective aperture of an antenna is the same for receiving and transmitting.

Three expressions have now been given for the directivity D. They are

$$D = \frac{P(\theta, \phi)_{\text{max}}}{P(\theta, \phi)_{\text{av}}} \qquad \text{(dimensionless)} \qquad \textit{Directivity from pattern}$$

$$D = \frac{4\pi}{\Omega_A} \qquad \text{(dimensionless)} \qquad \textit{Directivity from pattern}$$

$$D = 4\pi \frac{A_e}{\lambda^2} \qquad \text{(dimensionless)} \qquad \textit{Directivity from aperture}$$

for the case of the dipole antenna the load power
$P_{load} = SA_e$ (W)
where
S = power density at receiving antenna, W/m^2
A_e = effective aperture of antenna, m^2
a reradiated power
*P*rerad = Power reradiated/4π sr= SA_r (W)
where A_r =reradiating aperture = A_e, m^2 and
***P*rerad = *P*load**
The above discussion is applicable to a single dipole (λ/2 or shorter). However, it does not apply to all antennas. In addition to the reradiated power, an antenna may scatter power that does not enter the antenna-load circuit. Thus, the reradiated plus scattered power may exceed the power delivered to the load.

ANTENNA EFFICIENCY

The total antenna efficiency e_0 is used to take into account losses at the input terminals and within the structure of the antenna. Such losses may be due

1. reflections because of the mismatch between the transmission line and the antenna

2. I^2R losses (conduction and dielectric)

In general, the overall efficiency can be written as

$$e_0 = e_r e_c e_d$$

where

e_0 = total efficiency (dimensionless)
e_r = reflection(mismatch) efficiency = $(1 - |\Gamma|^2)$ (dimensionless)
e_c = conduction efficiency (dimensionless)
e_d = dielectric efficiency (dimensionless)
Γ = voltage reflection coefficient at the input terminals of the antenna
$\Gamma = (Z_{in} - Z_0)/(Z_{in} + Z_0)$ where Z_{in} = antenna input impedance,
Z_0 = characteristic impedance of the transmission line]

VSWR = voltage standing wave ratio = $\mathbf{(1 + |\Gamma|)/(1 - |\Gamma|)}$
Usually e_c and e_d are very difficult to compute, but they can be determined experimentally. Even by measurements they cannot be separated.

$$e_0 = e_r e_{cd} = e_{cd}(1 - |\Gamma|^2)$$

Where $e_{cd} = e_{ced}$ = antenna radiation efficiency, which is used to relate the gain and directivity.

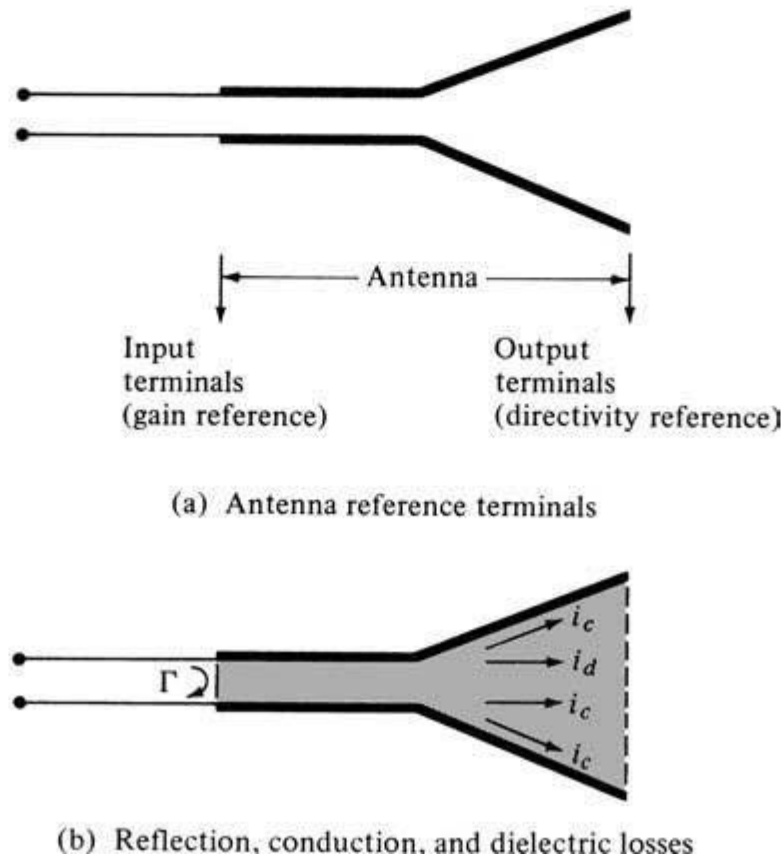


Fig. Reference terminals and losses of an antenna

EFFECTIVE HEIGHT

The effective height h (meters) of an antenna is another parameter related to the aperture. multiplying the effective height by the incident field E (volts per meter) of the same polarization gives the voltage V induced. Thus,

$$V = hE \qquad (1)$$

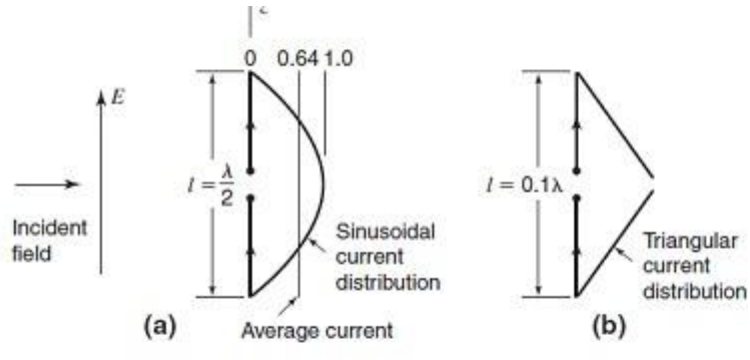
Accordingly, the effective height may be defined as the ratio of the induced voltage to the incident field or

$$h = V/E \text{ (m)} \qquad (2)$$

Consider, for example, a vertical dipole of length $l = \lambda/2$ immersed in an incident field E, as in below Fig.

If the current distribution of the dipole were uniform, its effective height would be l. The actual current distribution, however, is nearly sinusoidal with an average value $2/\pi = 0.64$ (of the maximum) so that its effective height $h = 0.64\ l$. It is assumed that the antenna is oriented for maximum response.

Figure
(*a*) Dipole of length $l = \lambda/2$ with sinusoidal current distribution.
(*b*) Dipole of length $l = 0.1\lambda$ with triangular current distribution.



If the same dipole is used at a longer wavelength so that it is only 0.1λ long, the current tapers almost linearly from the central feed point to zero at the ends in a triangular distribution, as in Fig.(b). The average current is 1⁄2 of the maximum so that the effective height is $0.5l$. Thus, another way of defining effective height is to consider the transmitting case and equate the effective height to the physical height (or length l) multiplied by the (normalized) average current or

$$h_e = \frac{1}{I_0}\int_0^{h_p} I(z)\,dz = \frac{I_{av}}{I_0}h_p \qquad \text{(m)}$$

where

h_e = effective height, m
h_p = physical height, m
I_{av} = average current, A

It is apparent that *effective height* is a useful parameter for transmitting tower-type antennas. It also has an application for small antennas. The parameter *effective aperture* has more general application to all types of antennas. The two have a simple relation, as will be shown.

For an antenna of radiation resistance R_r matched to its load, the power delivered to the load is equal to

$$P = \frac{1}{4}\frac{V^2}{R_r} = \frac{h^2E^2}{4R_r} \qquad \text{(W)}$$

In terms of the effective aperture the same power is given by

$$P = SA_e = \frac{E^2A_e}{Z_0} \qquad \text{(W)}$$

where Z_0 =intrinsic impedance of space (= 377 Ω)

$$h_e = 2\sqrt{\frac{R_r A_e}{Z_0}} \quad \text{(m)} \qquad \text{and} \qquad A_e = \frac{h_e^2 Z_0}{4R_r} \quad (\text{m}^2)$$

Thus, effective height and effective aperture are related via radiation resistance and the intrinsic impedance of space.

To summarize, we have discussed the space parameters of an antenna, namely, field and power patterns, beam area, directivity, gain, and various apertures.

Antenna Polarization

Polarization of an antenna in a given direction is defined as ―the polarization of the wave transmitted (radiated) by the antenna. *Note:* When the direction is not stated, the polarization is taken to be the polarization in the direction of maximum gain.‖ In practice, polarization of the radiated energy varies with the direction from the center of the antenna, so that different parts of the pattern may have different polarizations.
Polarization of a radiated wave is defined as ―that property of an electromagnetic wave describing the time-varying direction and relative magnitude of the electric-field vector; specifically, the figure traced as a function of time by the extremity of the vector at a fixed location in space, and the sense in which it is traced, *as observed along the direction of propagation*.‖ Polarization then is the curve traced by the end point of the arrow (vector) representing the instantaneous electric field. The field must be observed along the direction of propagation.
Polarization may be classified as linear, circular, or elliptical. If the vector that describes the electric field at a point in space as a function of time is always directed along a line, the field is said to be linearly polarized. In general, however, the electric field traces is an ellipse, and the field is said to be elliptically polarized.
Linear and circular polarizations are special cases of elliptical, and they can be obtained when the ellipse becomes a straight line or a circle, respectively. The electric field is traced in a clockwise (CW) or counterclockwise (CCW) sense. Clockwise rotation of the electric-field vector is also designated as right-hand polarization and counterclockwise as left-hand polarization.

In general, the polarization characteristics of an antenna can be represented by its polarization pattern whose one definition is ―the spatial distribution of the polarizations of a field vector excited (radiated) by an antenna taken over its radiation sphere. When describing the polarizations over the radiation sphere, or portion of it, reference lines shall be specified over the sphere, in order to measure the tilt angles (see tilt angle) of the polarization ellipses and the direction of polarization for linear polarizations. An obvious choice, though by no means the only one, is a family of lines tangent at each point on the sphere to either the θ or φ coordinate line associated with a spherical coordinate system of the radiation sphere. At each point on the radiation sphere the polarization is usually resolved into a pair of orthogonal polarizations, the co-polarization and cross polarization.

To accomplish this, the co-polarization must be specified at each point on the radiation sphere.‖ –Co-polarization represents the polarization the antenna is intended to radiate (receive) while cross-polarization represents the polarization orthogonal to a specified polarization, which is usually the co-polarization.‖

–For certain linearly polarized antennas, it is common practice to define the co polarization in the following manner: First specify the orientation of the co-polar electric-field vector at a pole of the radiation sphere. Then, for all other directions of interest (points on the radiation sphere), require that the angle that the co-polar electric-field vector makes with each great circle line through the pole remain constant over that circle, the angle being that at the pole.‖

–In practice, the axis of the antenna‘s main beam should be directed along the polar axis of the radiation sphere. The antenna is then appropriately oriented about this axis to align the direction of its polarization with that of the defined co-polarization at the pole.‖ –This manner of defining co-polarization can be extended to the case of elliptical polarization by defining the constant angles using the major axes of the polarization ellipses rather than the co-polar electric-field vector. The sense of polarization (rotation) must also be specified.‖

Linear, Circular, and Elliptical Polarizations

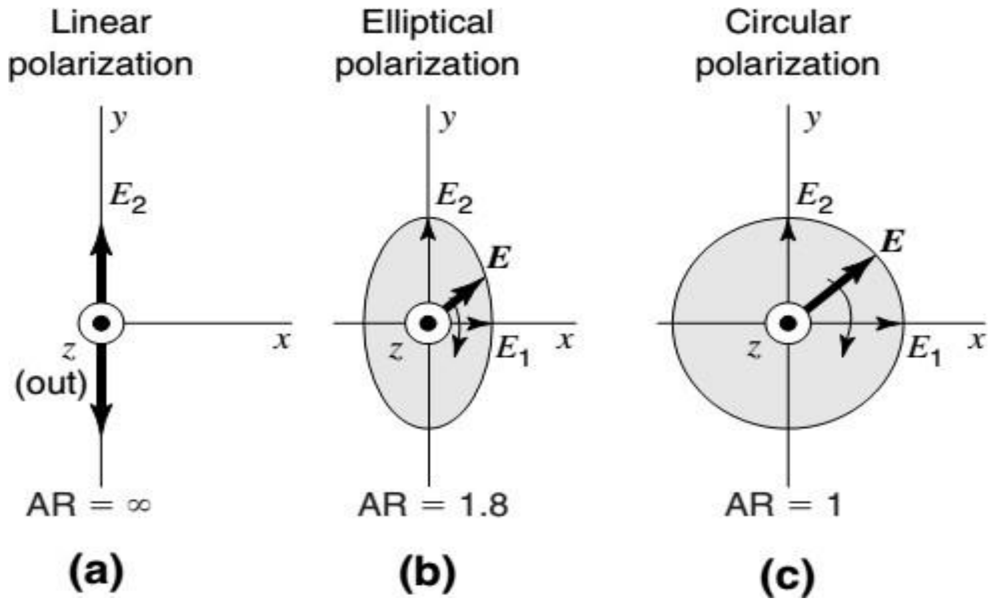
The instantaneous field of a plane wave, traveling in the negative *z* direction, can be written as

$$\mathscr{E}(z;t) = \hat{\mathbf{a}}_x \mathscr{E}_x(z;t) + \hat{\mathbf{a}}_y \mathscr{E}_y(z;t)$$

The instantaneous components are related to their complex counterparts by

$$\begin{aligned} \mathscr{E}_x(z;t) &= \mathrm{Re}[E_x^- e^{j(\omega t+kz)}] = \mathrm{Re}[E_{xo} e^{j(\omega t+kz+\phi_x)}] \\ &= E_{xo}\cos(\omega t + kz + \phi_x) \\ \mathscr{E}_y(z;t) &= \mathrm{Re}[E_y^- e^{j(\omega t+kz)}] = \mathrm{Re}[E_{yo} e^{j(\omega t+kz+\phi_y)}] \\ &= E_{yo}\cos(\omega t + kz + \phi_y) \end{aligned}$$

Where E_{xo} and E_{yo} are, respectively, the maximum magnitudes of the *x* and *y* components.



Linear Polarization

For the wave to have linear polarization, the time-phase difference between the two components must be

$$\Delta\phi = \phi_y - \phi_x = n\pi, \quad n = 0, 1, 2, 3, \ldots$$

Circular Polarization

Circular polarization can be achieved *only* when the magnitudes of the two components are the same *and* the time-phase difference between them is odd multiples of $\pi/2$.

$$|\mathscr{E}_x| = |\mathscr{E}_y| \Rightarrow E_{xo} = E_{yo}$$

$$\Delta\phi = \phi_y - \phi_x = \begin{cases} +(\frac{1}{2} + 2n)\pi, n = 0, 1, 2, \ldots & \text{for CW} \\ -(\frac{1}{2} + 2n)\pi, n = 0, 1, 2, \ldots & \text{for CCW} \end{cases}$$

If the direction of wave propagation is reversed (i.e., +z direction), the phases in for CW and CCW rotation must be interchanged.

Elliptical Polarization

Elliptical polarization can be attained only when the time-phase difference between the two components is odd multiples of π/2 and their magnitudes are not the same or when the time-phase difference between the two components is not equal to multiples of π/2 (irrespective of their magnitudes). That is,

$$|\mathscr{E}_x| \neq |\mathscr{E}_y| \Rightarrow E_{xo} \neq E_{yo}$$

$$\text{when } \Delta\phi = \phi_y - \phi_x = \begin{cases} +(\frac{1}{2} + 2n)\pi & \text{for CW} \\ -(\frac{1}{2} + 2n)\pi & \text{for CCW} \end{cases} \quad n = 0, 1, 2, \ldots$$

$$\Delta\phi = \phi_y - \phi_x \neq \pm\frac{n}{2}\pi = \begin{cases} > 0 & \text{for CW} \\ < 0 & \text{for CCW} \end{cases} \quad n = 0, 1, 2, 3, \ldots$$

Linear Polarization:

A time-harmonic wave is linearly polarized at a given point in space if the electric-field (or magnetic-field) vector at that point is always oriented along the same straight line at every instant of time. This is accomplished if the field vector (electric or magnetic) possesses:

a. Only one component, or

b. Two orthogonal linear components that are in time phase or 180◦ (or multiples of 180◦) out-of-phase.

Circular Polarization:

A time-harmonic wave is circularly polarized at a given point in space if the electric (or magnetic) field vector at that point traces a circle as a function of time.

The necessary and sufficient conditions to accomplish this are if the field vector (electric or magnetic) possesses *all* of the following:

a. The field must have two orthogonal linear components, and

b. The two components must have the same magnitude, and

c. The two components must have a time-phase difference of odd multiples of 90˚.

The sense of rotation is always determined by rotating the phase-leading component toward the phase-lagging component and observing the field rotation as the wave is viewed as it travels away from the observer. If the rotation is clockwise, the wave is right-hand (or clockwise) circularly polarized; if the rotation is counterclockwise, the wave is left-hand (or counterclockwise) circularly polarized. The rotation of the phase-leading component toward the phase-lagging component should be done along the angular separation between the two components that is less than 180˚. Phases equal to or greater than 0˚ and less than 180˚ should be considered leading whereas those equal to or greater than 180˚ and less than 360˚ should be considered lagging.

Elliptical Polarization A time-harmonic wave is elliptically polarized if the tip of the field vector (electric or magnetic) traces an elliptical locus in space. At various instants of time the field vector changes continuously with time at such a manner as to describe an elliptical locus. It is right-hand (clockwise) elliptically polarized if the field vector rotates clockwise, and it is left-hand (counterclockwise) elliptically polarized if the field vector of the ellipse rotates counter clockwise.

The sense of rotation is determined using the same rules as for the circular polarization. In addition to the sense of rotation, elliptically polarized waves are also specified by their axial ratio whose magnitude is the ratio of the major to the minor axis. A wave is elliptically polarized if it is not linearly or circularly polarized. Although linear and circular polarizations are special cases of elliptical, usually in practice elliptical polarization refers to other than linear or circular. The *necessary and sufficient* conditions to accomplish this are if the field vector (electric or magnetic) possesses *all* of the following:

a. The field must have two orthogonal linear components, and
b. The two components can be of the same or different magnitude.
c. (1) If the two components are not of the same magnitude, the time-phase difference between the two components must not be 0° or multiples of 180° (because it will then be linear).
(2) If the two components are of the same magnitude, the time-phase difference between the two components must not be odd multiples of 90° (because it will then be circular).

If the wave is elliptically polarized with two components not of the same magnitude but with odd multiples of 90° time-phase difference, the polarization ellipse will not be tilted but it will be aligned with the principal axes of the field components. The major axis of the ellipse will align with the axis of the field component which is larger of the two, while the minor axis of the ellipse will align with the axis of the field component which is smaller of the two.

ANTENNA FIELD ZONES

The fields around an antenna may be divided into two principal regions, one near the antenna called the near field or Fresnel zone and one at a large distance called the far field or Fraunhofer zone.

The boundary between the two may be arbitrarily taken to be at a radius

$$R = \frac{2L^2}{\lambda} \qquad \text{(m)}$$

where
L= Maximum dimension of the antenna in meters
λ=wavelength, meters

In the far or Fraunhofer region, the measurable field components are transverse to the radial direction from the antenna and all power flow is directed radially outward.

In the far field the shape of the field pattern is independent of the distance. In the near or Fresnel region, the longitudinal component of the electric field may be significant and power flow is not entirely radial. In the near field, the shape of the field pattern depends, in general, on the distance.

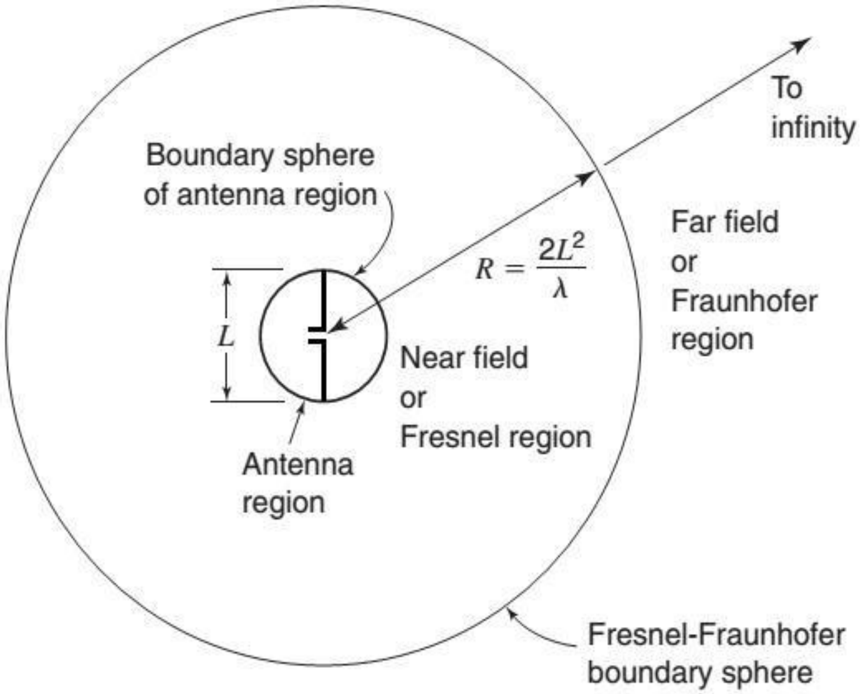


Figure: Antenna region, Fresnel region and Fraunhofer region.

FRIIS TRANSMISSION FORMULA

$$\frac{P_r}{P_t} = \frac{A_{er} A_{et}}{r^2 \lambda^2} \quad \text{(dimensionless)} \qquad \textit{Friis transmission formula}$$

where

P_r = received power, W
P_t = transmitted power, W
A_{et} = effective aperture of transmitting antenna, m^2
A_{er} = effective aperture of receiving antenna, m^2
r = distance between antennas, m
λ = wavelength, m

Radiation from Alternating current Element

If calculated outside the current distribution, then J = 0. Hence E is expressed in terms of a vector potential A.

To calculate the electromagnetic field radiated in the space by a short dipole, the retarded potential is used. A short dipole is an alternating current element. It is also called an oscillating current element. An alternating current element is considered as the basic source of radiation. It can be used as a building block for antenna analysis. For the calculation of electromagnetic field of the current element, the concept of retarded vector potential which is discussed earlier is most useful.

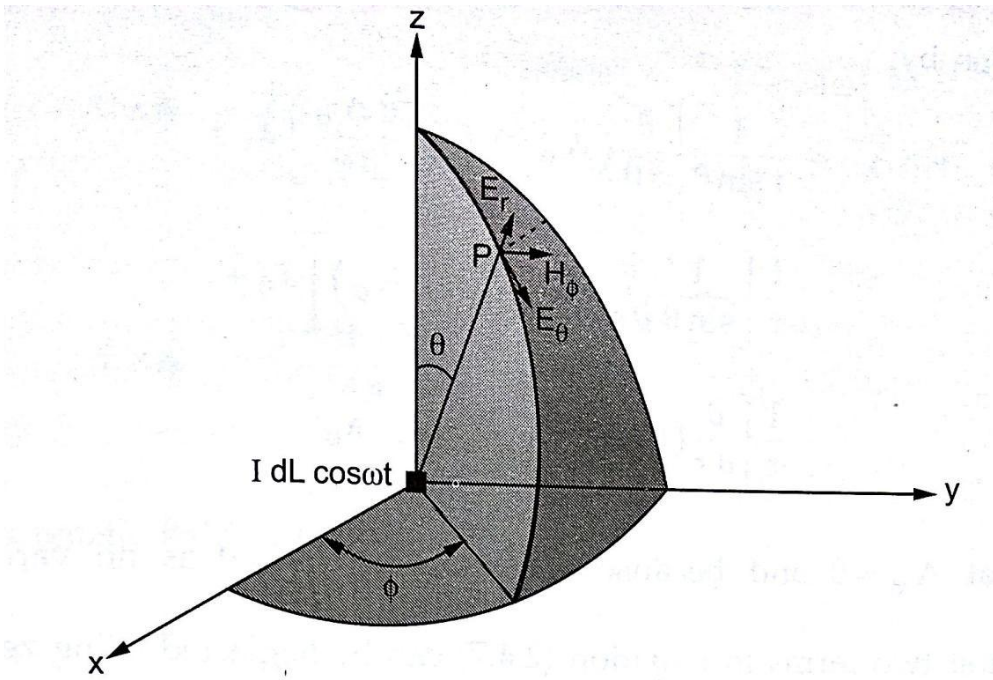
In general, a current element IdL is nothing but an element of length dL carrying filamentary current I. This length of a thin wire is assumed to be very short, so that the filamentary current can be considered as constant along the length of an element. The important usage of this approximation is observed in case of current carrying antenna. In such cases, an antenna can be considered as made up of large numbers of such elements connected end to end. Hence if the electromagnetic field of such small element is known, then the electromagnetic field of any long antenna can be easily calculated.

Let us study how to calculate the electromagnetic field due to an alternating current element. Consider spherical co-ordinate system. Consider that an alternating current element IdL cos o) t is located at the centre as shown in the figure. The aim is to calculate electromagnetic field at a point P placed at a distance R from the origin. The current element IdL cos ω t is placed along the z-axis.

Let us write the expression for vector potential $\bar{A}$ at point P, using previous knowledge. The vector potential is given by,

$$\bar{A}(r) = \frac{\mu}{4\pi} \int_v \frac{\bar{J}\left(t - \frac{r}{v}\right)}{R} dv'$$

$$A_z = \frac{\mu}{4\pi} \int_v \frac{\bar{J}\left(t - \frac{r}{v}\right)}{R} dv'$$



Electromagnetic field at point P when an alternating current element IdL cosωt placed at origin

$$\int_v \bar{J}\left(t-\frac{r}{v}\right) dv' = I\ dL\ \cos\omega\left(t-\frac{r}{v}\right)$$

The Hertzian dipole – Radiation between a current element and Electric dipole

Hertzian dipole is nothing but an infinitesimal current element IdL. Actually such a current element does not exist in the real life, but it serves as block in electric field of the alternating current element contains the terms of building calculating the field of a practical antenna using integration. It that the field of an. electric dipole which correspond to observe.

A Hertzian dipole consisting two equal and opposite charges at the end of the current element separated by a short distance dL is as shown in the Figure.

The wire between the two spheres where charges can accumulate is very thin as compared to the radius of -spheres. Thus the current I is uniform through the wires. Also the distance dL is greater as compared to the radii of the spheres.

$$i = I \cos w t$$

Then the charge accumulated at the ends of the element and current flowing through the wire are related to each other by the expression,

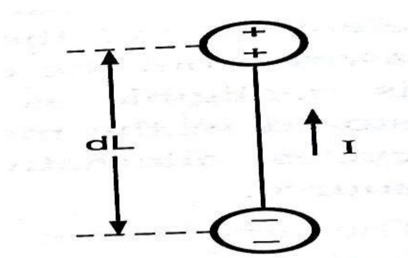
$$dq = I \cos\omega t \, dt$$

Substituting the value of q in terms of current I we will get

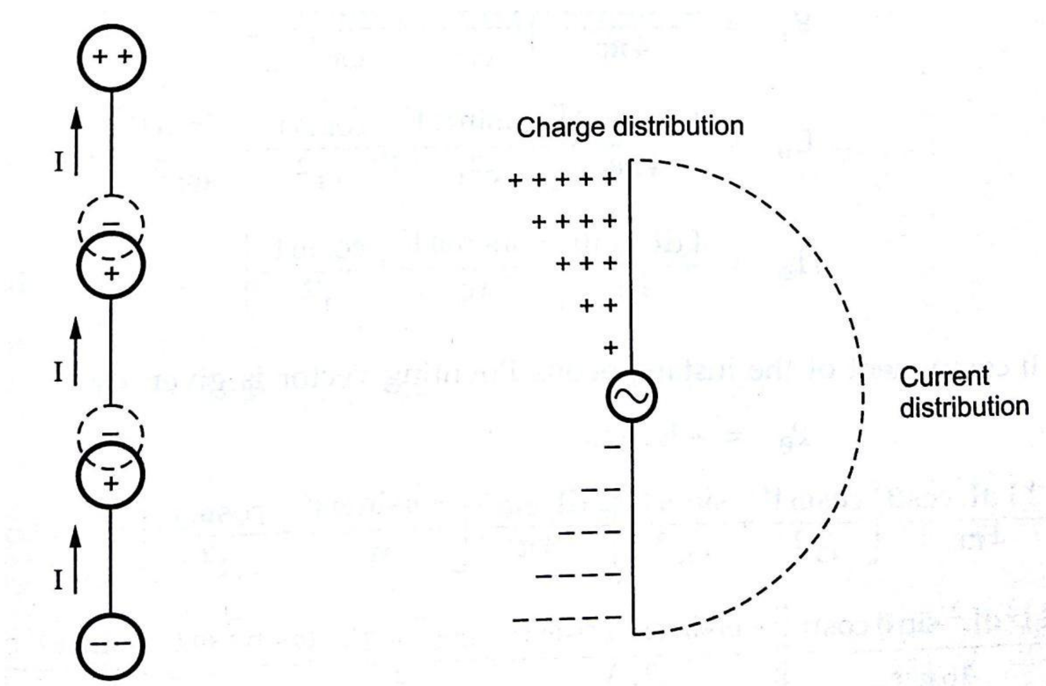
$$E_\theta = \frac{I\, dL \sin\theta \sin\omega t'}{4\pi\varepsilon\omega r^3}$$

Hertzian dipole – Radiation between current element and Electric dipole

Hertzian dipole is nothing but an infinitesimal current element. Actually such a current element does not exist in the real life, but it serves as block in electric field of the alternating current in terms of building calculating the field of a practical antenna using integration.



Hertzian dipole



Chain of Hertzian dipoles and charge and current distributions on linear antenna

When such Hertzian dipoles are connected end to end forming a practical antenna, it is observed that the positive charge at one end of the dipole gets cancelled by the equal and opposite charge at lower end of the next dipole. Hence when the current is uniform along the antenna, then there is no charge accumulation at the ends of the dipole which indicates that $1/r^3$ term is absent and only induction and radiation fields are present. The chain of Hertzian dipole forming part of antenna is as shown in figure

But if the current through antenna is not uniform throughout then - there is a accumulation of charge as shown in the Figure These charges causes stronger electric field component normal to the surface of the wire.

$$P_r = \frac{\eta_0}{2}\left(\frac{\omega I \, dL \sin\theta}{4\pi r c}\right)^2$$

Power radiated by a current element

Consider a current element placed at a center of a spherical coordinate system. Then the power radiated per unit area at point p can be calculated using pointing theorem.

The radial power is

$$P_r = \frac{\eta_0}{2}\left(\frac{\omega I \, dL \sin\theta}{4\pi r c}\right)^2$$

Short Linear Antennas

The current element that we have considered previously is not a practical, but it is hypothetical. It is useful in the theoretical calculations such as the components of the fields, radiation of power etc. The practical example of the centre-fed antenna is an elementary dipole.

The length of such centre-fed antenna is very short in wavelength. The current amplitude on such antenna is maximum at the center and it decreases uniformly to zero at the ends. The current distribution of short dipole is as shown in the Figure.

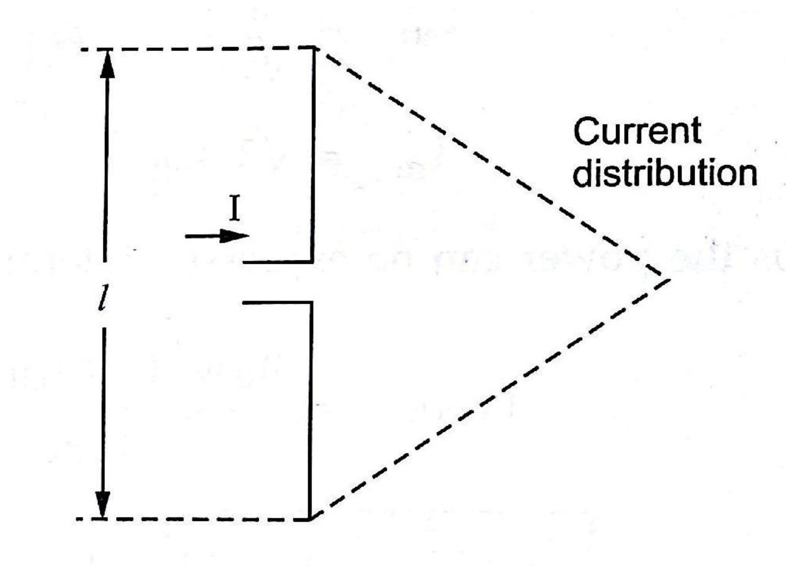
If we consider same current I flowing through the hypothetical current element and the practical short dipole, both of same length, then the practical short dipole radiate only one-quarter of the power that is radiated by the current element. This is because the field strengths at every point on the short dipole reduce to half of the values for the current element and hence the power density reduces to one quarter. So obviously for same current, the radiation resistance for the short dipole is ¼ times. Hence the radiation resistance is given by

Another practical example of an antenna is a monopole or short vertical antenna mounted on a reflecting plane as shown in the Figure.

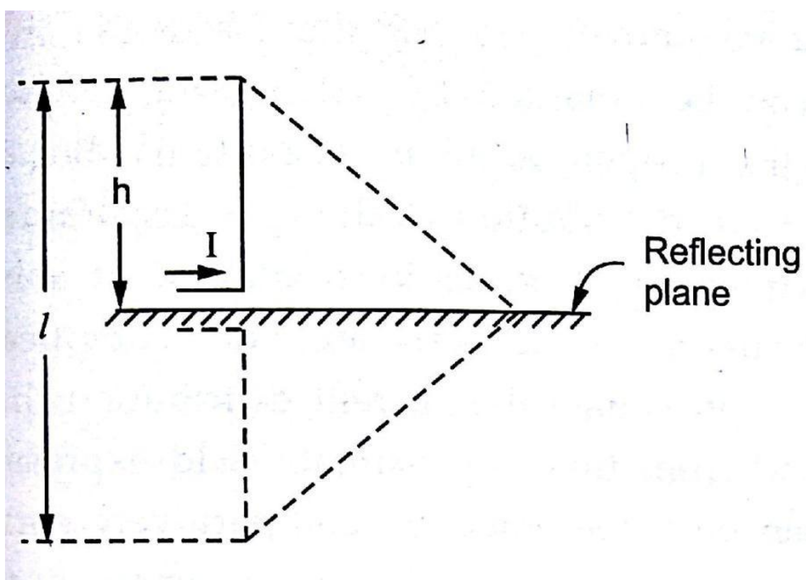
Let the monopole is of length h. Again if we consider same current I flows through a monopole of length h and a short dipole of length / = 2h then the field strength produced by both the antennas is same above the reflecting plane. But the monopole radiates only through the hemispherical surface above the plane. So the radiated power of a monopole is half of that radiated by a short dipole. Hence the radiation resistance of a monopole is half of the radiation resistance of the short dipole.

$$R_{\text{rad (short dipole)}} = 200(L/\lambda)^2$$

$$R_{rad\ (monopole)} = 400(h/\lambda)^2$$



Current distribution of short dipole



Current distribution of monopole

The half wave dipole and monopole

In order to calculate the radiated electromagnetic field of longer antenna, the the discussion in the current distribution along the antenna must be known. As boundary solving the Maxwell's previous sections, the current distribution can be obtained by equations for the time varying fields with the proper boundary conditions. But it is observed that the actual calculation of the current distribution of the cylindrical antenna is very difficult and complicated task. The mathematical expressions obtained by solving the Maxwell's equations with appropriate boundary conditions are very complicated. Hence, in general it is a common practice to approximate the current distribution that is more or less same as the real distribution and from that approximate field expressions are calculated. Such field expressions can then be represented by comparatively simpler expressions. Obviously the accuracy of the fields calculated with approximate current distribution assumption depends on the fact that how good an assumption is made for the current distribution. The centre fed antenna as an open circuited transmission line that is opened out, with a current distribution of sinusoidal type with current nodes at ends is studied in the last section. This assumption is the outcome of Abraham's work on the thin ellipsoids and it is observed that this assumption holds good for the thin antennas only.

A very commonly used antenna is the half wave dipole with a length one half of the free space wavelength of the radiated wave. It is found the linear current distribution is not suitable for this antenna. But when such antenna is fed at its centre with the help of a transmission line, it gives a current distribution which is approximately sinusoidal, with maximum at the centre and zero at the ends. The UHF and VHF regions, the dimensions of the half wave dipole make it most suitable as an antenna or as an antenna system element.

The half wave clippie can be considered as a chain of Hertzian dipoles. For the uniform current distribution, the positive charges at the end of one Hertzian dipole gets cancelled with an equal negative charge at the opposite end of the adjacent dipole. But when the current distribution is not constant (i.e. sinusoidal as assumed here), the successive dipoles of the chain have slightly different current amplitudes, where adjacent charges are not cancelled completely.

Power Radiated by the Half Wave Dipole and the Monopole

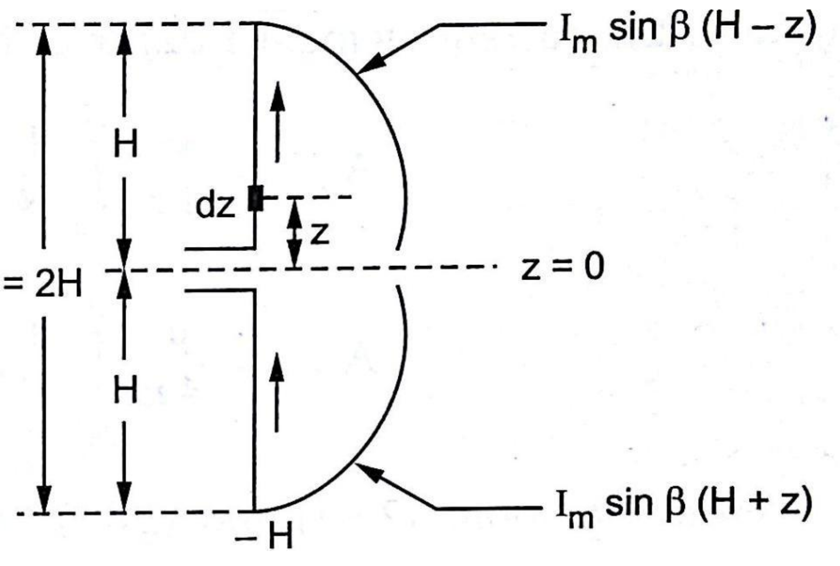
A dipole antenna is a vertical radiator fed in the centre. It produces maximum is the overall length.

The vertical antenna of height H =L/2 produces the radiation characteristics above the plane which is similar to that produced by the dipole antenna of length L = 2H. The vertical antenna is referred as a monopole.

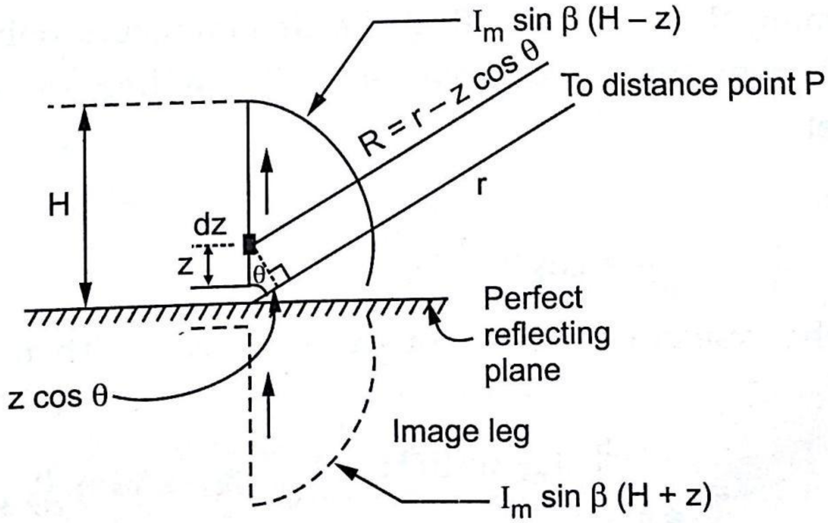
In general antenna requires large current to radiate large amount of power. To generate such a large current at radio frequency it is practically impossible. In case of Hertzian dipole the expressions for E and H are derived assuming uniform current throughout the length. But we have studied that at the ends of the antenna current is zero. In other words the current is not uniform throughout the length as it is maximum at centre and zero at the ends. Hence practically Hertzian dipole is not used. The practically used antennas are half wave dipole (λ / 2) and quarter wave monopole (λ / 4).

The half wave dipole consists two legs each of length L/2. The physical length of the half wave dipole at the frequency of operation is λ/2 in free space.

The quarter wave mono pole consists of single vertical leg erected on the perfect ground i.e. on the perfect conductor. The length of the leg of the quarter wave monopole is λ/4.



Assumed sinusoidal current distribution in half wave dipole



Assumed sinusoidal current distribution in quarter wave monopole

$$H_\phi = \frac{j I_m e^{-j\beta r}}{2\pi r}\left[\frac{\cos\left(\frac{\pi}{2}\cos\theta\right)}{\sin\theta}\right]$$

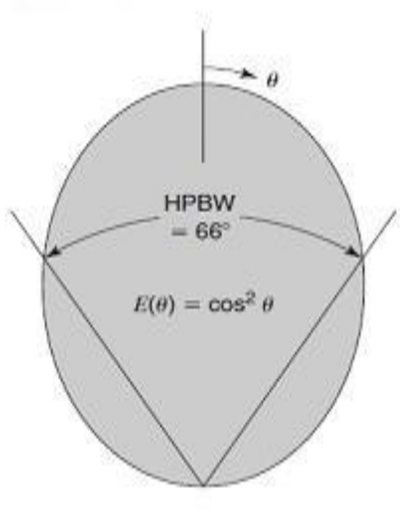
Problems

Problem 1

An antenna has a field pattern given by

$$E(\theta) = \cos^2\theta \quad \text{for } 0^\circ \leq \theta \leq 90^\circ$$

Find the half-power beamwidth (HPBW).



Solution

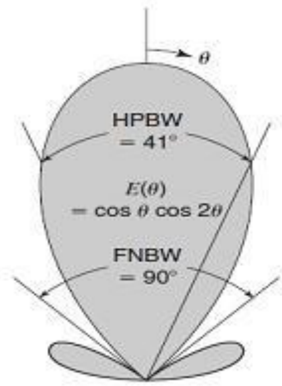
$E(\theta)$ at half power $= 0.707$. Thus $0.707 = \cos^2\theta$ so $\cos\theta = \sqrt{0.707}$ and $\theta = 33^\circ$

HPBW $= 2\theta = 66^\circ$ *Ans.*

Problem 2

Half-Power Beamwidth and First Null Beamwidth

An antenna has a field pattern given by $E(\theta) = \cos\theta \cos 2\theta$ for $0^\circ \leq \theta \leq 90^\circ$. Find (*a*) the half-power beamwidth (HPBW) and (*b*) the beamwidth between first nulls (FNBW).



Solution

(*a*) $E(\theta)$ at half power $= 0.707$. Thus $0.707 = \cos\theta \cos 2\theta = 1/\sqrt{2}$.

$$\cos 2\theta = \frac{1}{\sqrt{2}\cos\theta} \qquad 2\theta = \cos^{-1}\left(\frac{1}{\sqrt{2}\cos\theta}\right) \qquad \text{and}$$

$$\theta = \frac{1}{2}\cos^{-1}\left(\frac{1}{\sqrt{2}\cos\theta'}\right)$$

Iterating with $\theta' = 0$ as a first guess, $\theta = 22.5^\circ$. Setting $\theta' = 22.5^\circ$, $\theta = 20.03^\circ$, etc., until after next iteration $\theta = \theta' = 20.47^\circ \cong 20.5^\circ$ and

$$\text{HPBW} = 2\theta = 41^\circ \qquad \textit{Ans. (a)}$$

(*b*) $0 = \cos\theta \cos 2\theta$, so $\theta = 45^\circ$ and

$$\text{FNBW} = 2\theta = 90^\circ \qquad \textit{Ans. (b)}$$

Problem 3

The normalized radiation intensity of an antenna is represented by

$$U(\theta) = \cos^2(\theta)\cos^2(3\theta), \quad (0 \le \theta \le 90^\circ, \quad 0^\circ \le \phi \le 360^\circ)$$

The three- and two-dimensional plots of this, plotted in a linear scale, are shown in Figure 2.11. Find the

a. half-power beamwidth HPBW (*in radians and degrees*)
b. first-null beamwidth FNBW (*in radians and degrees*)

Solution:

a. Since the $U(\theta)$ represents the *power* pattern, to find the half-power beamwidth you set the function equal to half of its maximum, or

$$U(\theta)|_{\theta=\theta_h} = \cos^2(\theta)\cos^2(3\theta)|_{\theta=\theta_h} = 0.5 \Rightarrow \cos\theta_h \cos 3\theta_h = 0.707$$
$$\theta_h = \cos^{-1}\left(\frac{0.707}{\cos 3\theta_h}\right)$$

Since this is an equation with transcendental functions, it can be solved iteratively. After a few iterations, it is found that

$$\theta_h \approx 0.25 \; radians = 14.325^\circ$$

Since the function $U(\theta)$ is symmetrical about the maximum at $\theta = 0$, then the HPBW is

$$\text{HPBW} = 2\theta_h \approx 0.50 \; radians = 28.65^\circ$$

b. To find the first-null beamwidth (FNBW), you set the $U(\theta)$ equal to zero, or

$$U(\theta)|_{\theta=\theta_n} = \cos^2(\theta)\cos^2(3\theta)|_{\theta=\theta_n} = 0$$

This leads to two solutions for θ_n.

$$\cos\theta_n = 0 \Rightarrow \theta_n = \cos^{-1}(0) = \frac{\pi}{2}\ radians = 90^\circ$$
$$\cos 3\theta_n = 0 \Rightarrow \theta_n = \frac{1}{3}\cos^{-1}(0) = \frac{\pi}{6}\ radians = 30^\circ$$

The one with the smallest value leads to the FNBW. Again, because of the symmetry of the pattern, the FNBW is

$$\text{FNBW} = 2\theta_n = \frac{\pi}{3}\ radians = 60^\circ$$

Problem4

A lossless resonant half-wavelength dipole antenna, with input impedance of 73 ohms, is connected to a transmission line whose characteristic impedance is 50 ohms. Assuming that the pattern of the antenna is given approximately by

$$U = B_0 \sin^3\theta$$

find the maximum absolute gain of this antenna.

Solution: Let us first compute the maximum directivity of the antenna. For this

$$U|_{\max} = U_{\max} = B_0$$
$$P_{\text{rad}} = \int_0^{2\pi}\int_0^{\pi} U(\theta,\phi)\sin\theta\, d\theta\, d\phi = 2\pi B_0 \int_0^{\pi} \sin^4\theta\, d\theta = B_0\left(\frac{3\pi^2}{4}\right)$$
$$D_0 = 4\pi\frac{U_{\max}}{P_{\text{rad}}} = \frac{16}{3\pi} = 1.697$$

Since the antenna was stated to be lossless, then the radiation efficiency $e_{cd} = 1$.

Thus, the total maximum gain is equal to

$$G_0 = e_{cd} D_0 = 1(1.697) = 1.697$$
$$G_0(\text{dB}) = 10\log_{10}(1.697) = 2.297$$

which is identical to the directivity because the antenna is lossless.

There is another loss factor which is not taken into account in the gain. That is the loss due to reflection or mismatch losses between the antenna (load) and the transmission line. This loss is accounted for by the reflection efficiency of (2-44) or (2-45), and it is equal to

$$e_r = (1 - |\Gamma|^2) = \left(1 - \left|\frac{73-50}{73+50}\right|^2\right) = 0.965$$
$$e_r(\text{dB}) = 10\log_{10}(0.965) = -0.155$$

Therefore the overall efficiency is

$$e_0 = e_r e_{cd} = 0.965$$
$$e_0(\text{dB}) = -0.155$$

Thus, the overall losses are equal to 0.155 dB. The absolute gain is equal to

$$G_{0abs} = e_0 D_0 = 0.965(1.697) = 1.6376$$
$$G_{0abs}(\text{dB}) = 10\log_{10}(1.6376) = 2.142$$

The gain in dB can also be obtained by converting the directivity and radiation efficiency in dB and then adding them. Thus,

$$e_{cd}(\text{dB}) = 10\log_{10}(1.0) = 0$$
$$D_0(\text{dB}) = 10\log_{10}(1.697) = 2.297$$
$$G_0(\text{dB}) = e_{cd}(\text{dB}) + D_0(\text{dB}) = 2.297$$

which is the same as obtained previously. The same procedure can be used for the absolute gain.

PART A

1. Define an antenna.
Antenna is a transition device or a transducer between a guided wave and a free space wave or vice versa. Antenna is also said to be an impedance transforming device.

2. What is meant by radiation pattern?
Radiation pattern is the relative distribution of radiated power as a function of distance in space. It is a graph which shows the variation in actual field strength of the EM wave at all points which are at equal distance from the antenna. The energy radiated in a particular direction by an antenna is measured in terms of field strength. (E Volts/m)

3. Define Radiation intensity?
The power radiated from an antenna per unit solid angle is called the radiation intensity U (watts per steradian or per square degree). The radiation intensity is independent of distance.

4. Define Beam efficiency?
The total beam area (Ω_A) consists of the main beam area (Ω_M) plus the minor lobe area (Ω_m) . Thus $\Omega_A = \Omega_M + \Omega_m$
The ratio of the main beam area to the total beam area is called beam efficiency.
Beam efficiency = $\Sigma_M = \Omega_M / \Omega_A$.

5. Define Directivity?
The directivity of an antenna is equal to the ratio of the maximum power density $P(\theta,\varphi)_{max}$ to its average value over a sphere as observed in the far field of an antenna.
$D = P(\theta,\varphi)_{max} / P(\theta,\varphi)_{av}$. Directivity from Pattern.

$D = 4\pi / \Omega_A$. Directivity from beam area(Ω_A).

6. What is meant by Polarization.?

The temporal behavior of the tip of the E-field vector is called as polarization.
The polarization are three types. They are
Elliptical polarization, Circular polarization and Linear polarization.

7. Define different types of aperture.?

Effective Aperture(A_e). It is the area over which the power is extracted from the incident wave and delivered to the load is called effective aperture.

Scattering Aperture(As.) It is the ratio of the reradiated power to the power density of the incident wave.

Loss Aperture. (A_e).
It is the area of the antenna which dissipates power as heat.

Collecting aperture. (A_e).
It is the addition of above three apertures. Physical aperture. (Ap). This aperture is a measure of the physical size of the antenna.

8. Define Aperture efficiency?

The ratio of the effective aperture to the physical aperture is the aperture efficiency. i.e
Aperture efficiency = η_{ap} = Ae / Ap (dimensionless).

9. What is meant by effective height?

The effective height h of an antenna is the parameter related to the aperture. It may be defined as the ratio of the induced voltage to the incident field that is H= V / E.

10. What are the field zone?

The fields around an antenna ay be divided into two principal regions.
i. Near field zone (Fresnel zone)
ii. Far field zone (Fraunhofer zone)

11. Define a Hertzian dipole?

Oscillating dipole or Hertzian dipole is a current carrying conductor in which the charges at both the ends starts at oscillate. Its length is very small compared to λ.

12. What is radiation resistance of a half wave dipole?

($Rr = 80\,\pi^2 (dl/\lambda)^2$ ohms. Where Rr = Radiation resistance Dl = length of the current element

λ = Wavelength.

13. List some applications of monopole antenna.

It is used in compact communications system like, Hand phones Remote control etc.,

14. What is radiation resistance?

Radiation resistance is the amount of opposition offered by an antenna to radiate the energy to free space. It is the ratio between power radiated by an antenna to the square of rms current flow in that antenna.

15. Define Hertz antenna.

It is a symmetrical dipole antenna in which the two ends are at equal potential relative to mid point whose length is equal to the half of the wavelength.

16. Define self- impedance

Self -impedance of an antenna is defined as its input impedance with all other antennas are completely removed i.e away from it.

17. What is point source?

It is the waves originate at a fictitious volumelessemitter source at the center of the observation circle.

18. What is mean by loop antenna?

An antenna is a radio antenna consisting of a loop (or loops) of wire, tubing, or other electrical conductor with its ends connected to a balanced transmission line.

19. Define half wave dipole antenna

A dipole antenna is the simplest type of radio antenna, consisting of a conductive wire rod thatis half the length of the maximum wavelength the antenna is to generate. This wire rod is split in the middle, and the two sections are separated by an insulator.

20. What is meant by isotropic radiator?

A isotropic radiator is a fictitious radiator and is defined as a radiator which radiates fields uniformly in all directions. It is also called as isotropic source or omni directional radiator or simply unipole.

UNIT II ANTENNA ARRAYS

INTRODUCTION

The field radiated by a small linear antenna is not distributed uniformly in the case of a short dipole, the direction perpendicular to the axis of the antenna. As in maximum radiation takes place in the direction right angles to the axis of the dipole. But it decreases to minimum when the polar angle decreases. So, these non-uniform radiation characteristics may be used for many broadcast services. But such a characteristics are not preferred in point to point communication. In the point to point communication, it is desired to have most of the energy radiated in one particular direction. That means it is desired to have greater directivity in a desired direction particularly which is not possible with single dipole antenna.

In general, antenna array is the radiating system in which several antennas are spaced properly so as to get greater field strength at a far distance from the radiating system by combining radiations at point from all the antennas in the system. In general, the total field produced by the antenna array at a far distance is the vector sum of the fields produced by the individual antennas of the array. The individual element is generally called element of an antenna array.

The antenna array is said to linear if the elements of the antenna array are equally spaced along a straight line. The linear antenna array is said to be uniform linear array if all the elements are fed with a current of equal magnitude with progressive uniform phase shift along the line.

In general, the element in the antenna array is a -f dipole. The length of half wavelength dipole may not be equal to the electrical wavelength. If the variation of the electrical length from is within 5 % then it is assumed that the radiation properties of individual elements are not affected.

As the antennas may be used in various configurations such as straight line, circle, rectangle etc., many configurations of antenna arrays are possible. But practically limited number of configurations is used extensively.

Hence antenna array is a radiating system in contribute to obtain maximum field strength in the

individual field strength in all other directions desired direction.

Various Forms of Antenna Arrays

Practically various forms of the antenna array are used as radiating systems. Some of the practically used forms are as follows.

1. Broadside Array 2. End fire Array

3. Collinear Array 4. Parasitic Array

This form of the antenna array is one of the most important practical forms used in practice. The broadside array is the array of antennas in which all the elements are placed parallel to each other and the direction of maximum radiation is always perpendicular to the plane consisting elements.

A typical arrangement of a Broadside array is as shown in the Figure.

A broadside array consist number of identical antennas placed parallel to each other along a straight line. This straight line is perpendicular to the axis of individual antenna. It is known as axis of antenna array. Thus each element is perpendicular to the axis of antenna array. All the individual antennas are spaced equally along the axis of the antenna array. The spacing between any two elements is denoted by_d'. All the elements are fed with currents with equal magnitude and same phase. As the maximum point sources with equal amplitude and phase radiation is directed in broadside direction i.e. perpendicular to the line of axis of array, the radiation pattern for the broadside array is bidirectional. Thus we can define broadside array as the arrangement of antennas in which maximum radiation is in the direction perpendicular to the axis of array and plane containing the elements of array.

Now consider two isotropic point sources spaced equally with respect to the origin of the co-ordinate system as shown in the Fig. 4.2.2 Assume that the two point sources are with equal amplitude and phase.

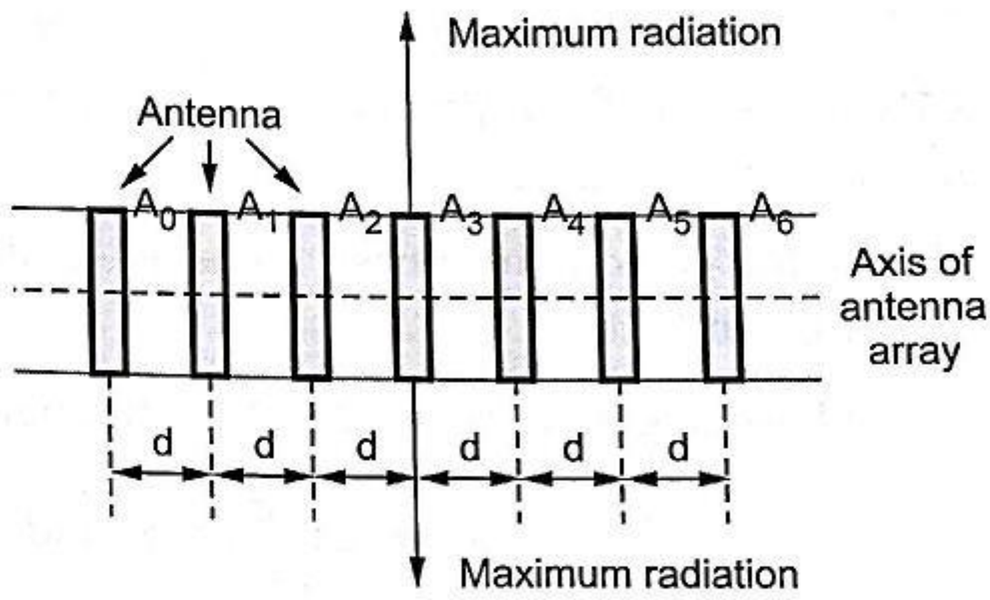
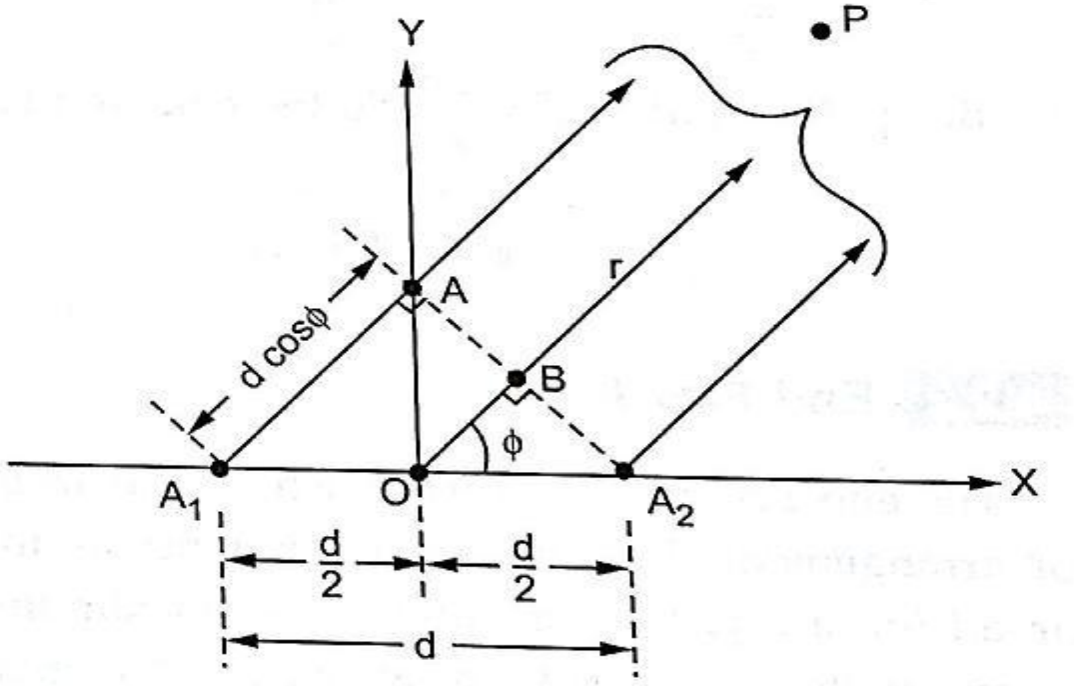


Figure - Broadside array of antennas



Broadside array with two isotropic point sources with equal amplitude and phase

Consider that point P is far away from the origin. Let the distance of point P from origin be r. The wave radiated by radiator A_2 will reach point P as compared to that radiated by radiator A_1.

This is due to the path difference that the wave radiated by radiator A_1 has to travel extra distance. Hence the path difference is given by,

$$\text{Path difference} = d \cos \phi$$

This path difference can be expressed in terms of wave length as

$$\text{Path difference} = \frac{d}{\lambda} \cos \phi$$

From the optics the phase angle is 2π times the path difference. Hence the phase angle is given by

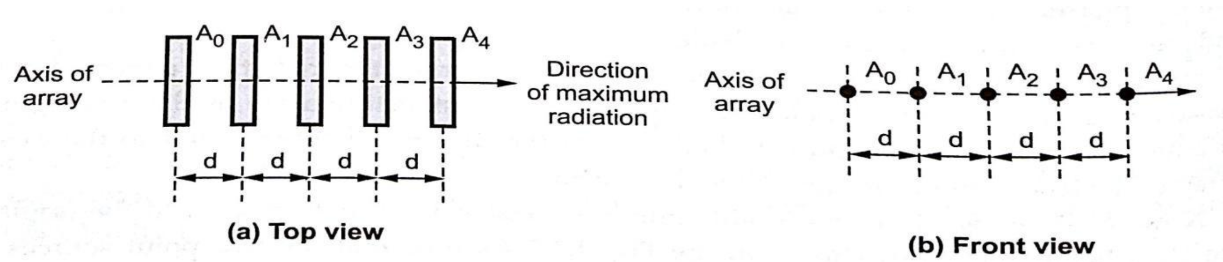
$$\text{Phase angle} = \psi = 2\pi \text{ (Path difference)}$$

$$\Psi = 2\pi \left(\frac{d}{\lambda} \cos \phi\right)$$

$\Psi = \left(\frac{2\pi}{\lambda}\right) d \cos\phi$ $\Psi = \beta d\cos\phi$

End Fire Array

The end fire array is very much similar to the broadside array from the point of view of arrangement. But the main difference is in the direction of maximum radiation. In broadside array, the direction of the maximum radiation is perpendicular to the axis of array; while in the end fire array, the direction of the maximum radiation is along the axis of array.



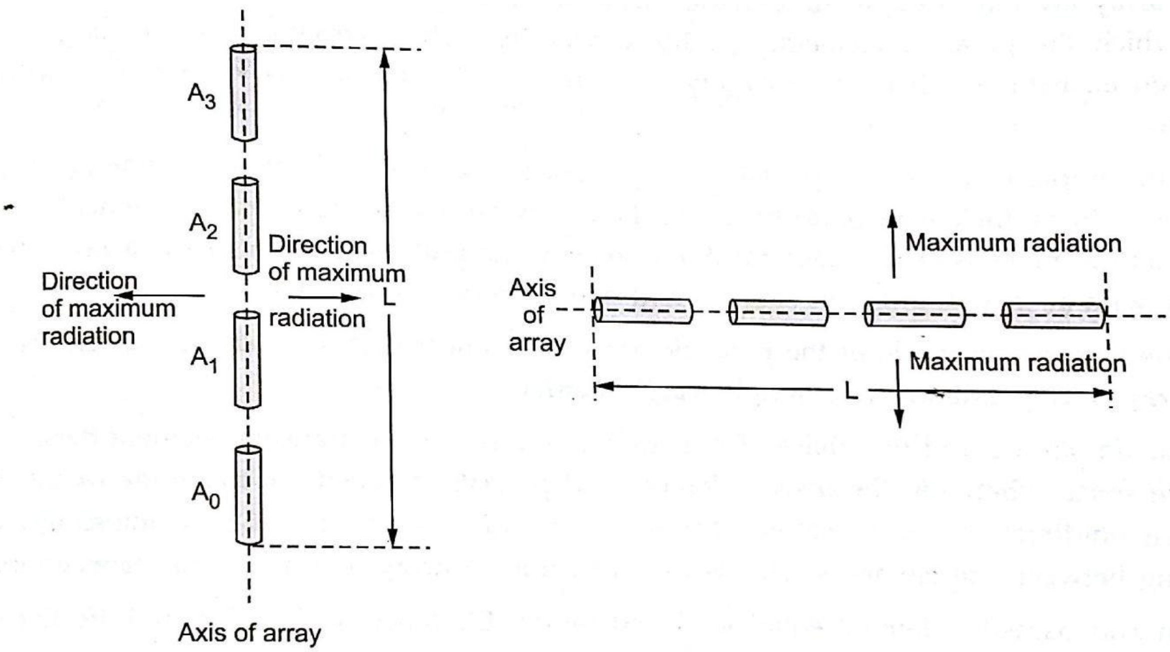
End Fire Array

Thus in the end fire array number of identical antennas are spaced equally along a line. All the antennas are fed individually with currents of equal magnitudes but their phases vary progressively along the line to get entire arrangement unidirectional finally. i.e. maximum radiation along the axis of array.

Thus end fire array can be defined as an array with direction of maximum radiation coincides with the direction of the axis of array to get unidirectional radiation.

Collinear Array

As the name indicates, in the collinear array, the antennas are arranged co-axially i.e. the antennas are arranged end to end along a single line as shown in the Fig. 4.2.4 (a) and (b).



(a) Vertical **(b) Horizontal**

Different Types of Collinear Array

The individual elements in the collinear array are fed with currents equal in magnitude and phase. This condition is similar to the broadside array. In collinear array the direction of maximum radiation is perpendicular to the axis of array. So the radiation pattern of the collinear array and the broadside array is very much similar but the radiation pattern of the collinear array has circular symmetry with main lobe perpendicular everywhere to the principle axis. Thus the collinear array is also called omnidirectional array or broadcast array.

The gain of the collinear array is maximum if the spacing between the elements is of the order of 0.3 λ to 0.5 λ. But this small spacing introduces constructional and feeding

same.

To derive different expressions following conditions can be applied to the antenna array

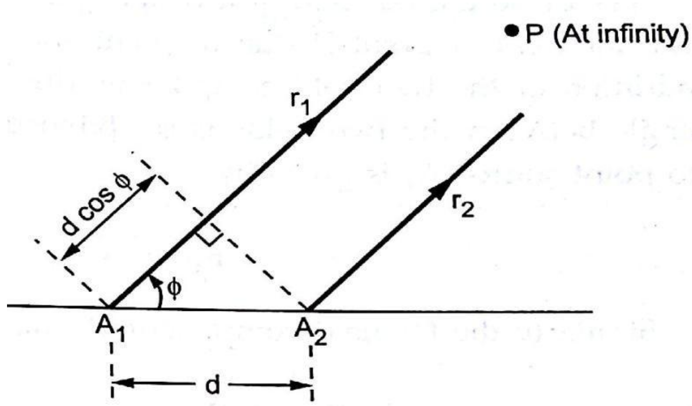
- Two point sources with currents of equal magnitudes and with same phase.
- Two point sources with currents of equal magnitude but with opposite phase.
- Two point sources with currents of unequal magnitudes and with opposite phase.

Two Point Sources with Currents Equal in Magnitude and Phase

Consider two point sources A_1 and A_2 separated by distance d as shown in the Figure of two element array. Consider that both the point sources are supplied with currents equal in magnitude and phase.

Consider point P far away from the array. Let the distance between point P and point sources A_1 and A_2 be r_1 and r_2 respectively. As these radial distances are extremely large as compared with the distance of separation between two point sources i.e. d, we can assume,

$r_1 = r_2 = r$



Two Element Array

The radiation from the point source A_2 will reach earlier at point P than that from point source A_1 because of the path difference. The extra distance is travelled by the radiated wave from point source A_1 than that by the wave radiated from point source A_2.

Hence path difference is given by,

$$\text{Path difference} = d\cos\theta$$

This path difference can be expressed in terms of wave length as

$$\text{Path difference} = \frac{d}{\lambda}\cos\theta$$

From the optics the phase angle is 2π times the path difference. Hence the phase angle is given by

$$\text{Phase angle} = \psi = 2\pi\,(\text{Path difference})$$

$$\Psi = 2\pi\left(\frac{d}{\lambda}\cos\phi\right)$$

$$\Psi = \left(\frac{2\pi}{\lambda}\right)d\cos\phi$$

$$\Psi = \beta d\cos\phi$$

$$E_T = 2E_0\cos\left(\frac{\beta d\cos\phi}{2}\right)$$

Above equation represents total field intensity at point P, due to two point sources having currents of same amplitude and phase. The total amplitude of the field at point P is $2E_0$ while the phase shift is $\beta d\frac{\cos\phi}{2}$

The array factor is the ratio of the magnitude of the resultant field to the magnitude of the maximum field.

$$\text{Therefore A.F.} = \frac{|E_T|}{2|E_{max}|}$$

$$\text{But maximum field is } E_{max} = 2E_0$$

$$\text{A.F.} = \frac{|E_T|}{|2E_0|} = \cos\left(\pi\frac{d}{\lambda}\cos\phi\right)$$

The array factor represents the relative value of the field as a function of ϕ. It defines the radiation pattern in a plane containing the line of the array.

Maxima Direction

From above equation , the total field is maximum when cos $\left(\frac{\beta d \cos\emptyset}{2}\right)$ is maximum.

As we know, the variation of cosine of a angle is ± 1. Hence the condition for maxima is given by,

$$\left(\frac{\beta d \cos\emptyset}{2}\right) = \pm 1$$

Let spacing between the two point sources be λ/2. Then we can write

$$\cos\left[\frac{\beta\frac{\lambda}{2}\cos\phi}{2}\right] = \pm 1$$

then we can say $\emptyset_{max} = 90^0 \ or 270^0$

Minima direction

Again from equation (4.4.9), total field strength is minimum when cos $\left(\frac{\beta d \cos\emptyset}{2}\right)$ is minimum that is 0 as cosine angle has minimum value 0. Hence the condition for minima is given by,

$$\cos\left(\frac{\beta d \cos\emptyset}{2}\right) = 0$$

then we can say that $\emptyset_{min} = 0^0 \ or \ 180^0$

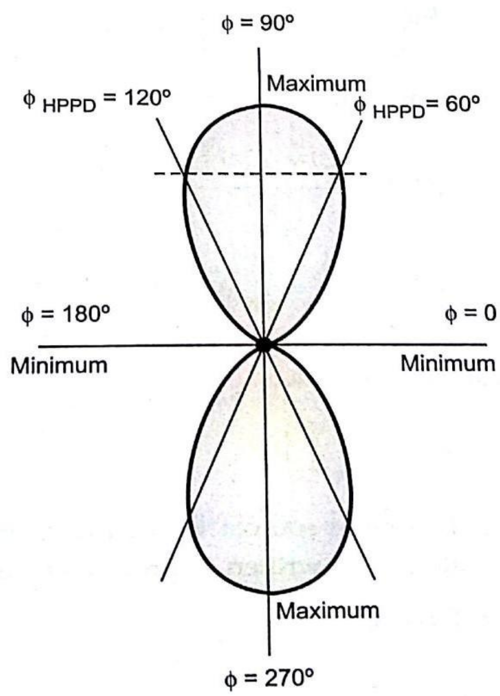
Half power point directions

When the power is half, the voltage or current is $\frac{1}{\sqrt{2}}$ times the maximum value. Hence the condition for half power point is given by,

$$\cos\left(\frac{\beta d \cos\emptyset}{2}\right) = \pm\frac{1}{\sqrt{2}}$$

Then by simplifying the above expression we will get $\varnothing_{HPPD} = 60^0\ or\ 120^0$

The field pattern drawn with E_T against ϕ for d = $\frac{\lambda}{2}$ then the pattern is bidirectional as shown in the figure. The field pattern obtained is bidirectional and it is a figure of eight (8). If this pattern is rotated by 360° about axis, it will represent three dimensional doughnut shaped space pattern. This is the simplest type of broadside array of two point sources and it is called Broadside couplet as two radiations of point sources are in phase.



Field pattern for two point source with spacing d= d = $\frac{\lambda}{2}$ and fed with currents equal in magnitude and phase

Two Point Sources with Currents Equal in Magnitudes but Opposite in Phase

Consider two point sources separated by distance d and supplied with currents equal

magnitude but opposite in phase. For the above figure all the conditions are exactly same except the phase of the currents is opposite i.e. 180°. With this condition, the total field at far point P is given by,

$$E_T = (-E_1) + (E_2)$$

Assuming equal magnitudes of currents, the fields at point P due to the point sources A_1 and A_2 can be written as,

$E_1 = E_0\, e^{-j\frac{\psi}{2}}$

$E_2 = E_0\, e^{j\frac{\psi}{2}}$

And substituting the values of E_1 and E_2 in the above equation we will get

$E_T = E_0\, e^{-j\frac{\psi}{2}} + E_0\, e^{-j\frac{\psi}{2}}$ Finally we will get

$E_T = j2E_0 \sin(\left(\frac{\psi}{2}\right)$

Now as the condition for two point sources with currents in phase and out of phase is exactly same, the phase angle can be written as previous case

Phase angle = $\psi = \beta d \cos\phi$

$E_T = j2E_0 \sin\left(\frac{\beta d \cos\emptyset}{2}\right)$

Substituting value of phase angle in equation we get,

$E_T = j2E_0 \sin\left(\frac{\beta d \cos\emptyset}{2}\right)$

Maxima direction

From the above equation, the total field is maximum when sin $\left(\frac{\beta d \cos\emptyset}{2}\right)$ is maximum that is ± 1, Hence the condition for maxima is sin $\left(\frac{\beta d \cos\emptyset}{2}\right) = \pm 1$

By taking the spacing between two isotropic point sources be equal to $\frac{\lambda}{2}$ that is d = $\frac{\lambda}{2}$ and β = $\frac{2\pi}{\lambda}$ in the above equation and simplifying we will get

$\Phi_{max} = 0^0 \; or \; 180^0$

Minima direction

Again from above equation total field strength is minimum when sin $\left(\frac{\beta d cos\emptyset}{2}\right)$ is minimum that is zero.

Hence the condition is given by

sin $\left(\frac{\beta d cos\emptyset}{2}\right) = 0$

By taking the spacing between two isotropic point sources be equal to $\frac{\lambda}{2}$ that is d = $\frac{\lambda}{2}$ and β = $\frac{2\pi}{\lambda}$ in the above equation and simplifying we will get

$\Phi_{min} = -90^0 \; or +90^0$

Half Power Point Direction

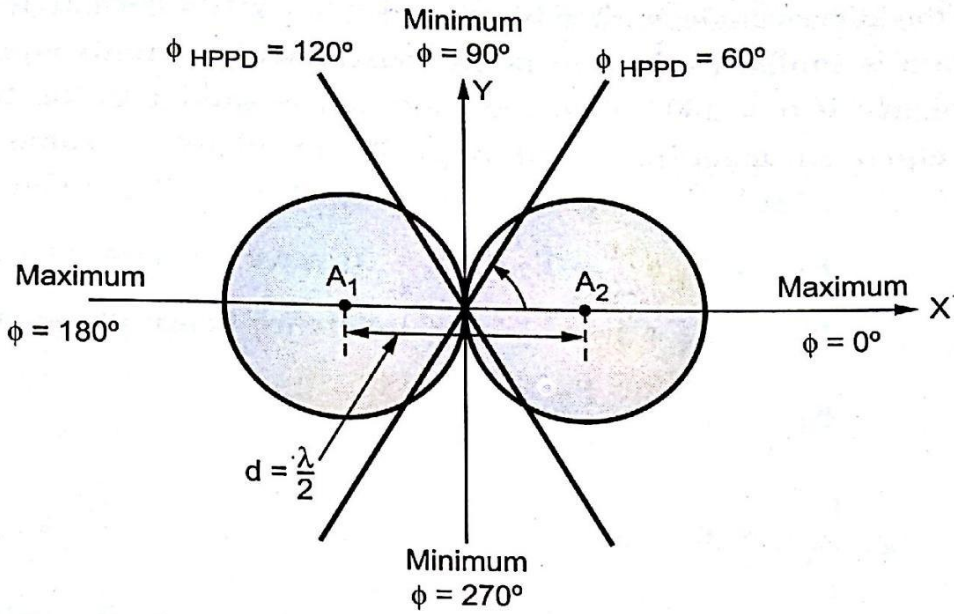
When the power is half, the voltage or current is $\frac{1}{\sqrt{2}}$ times the maximum value. Hence the condition for half power point is given by,

$$sin\left(\frac{\beta d cos\emptyset}{2}\right) = \pm\frac{1}{\sqrt{2}}$$

By taking the spacing between two isotropic point sources be equal to $\frac{\lambda}{2}$ that is d = $\frac{\lambda}{2}$ and β = $\frac{2\pi}{\lambda}$ in the above equation and simplifying we will get

Then by simplifying the above expression we will get $\emptyset_{HPPD} = 60^0 \; or \; 120^0$

As compared with the field pattern for two point sources with in-phase currents, the maxima have shifted by 90° along X-axis in case of out-phase currents in two point source array. Thus the maxima is along the axis of the array or along the line joining two point sources. In first case, we have obtained vertical figure of 8. Now in above case we have obtained horizontal figure of 8. AS the maximum field is along the line joining the two point sources, this is simple type of end fire array.



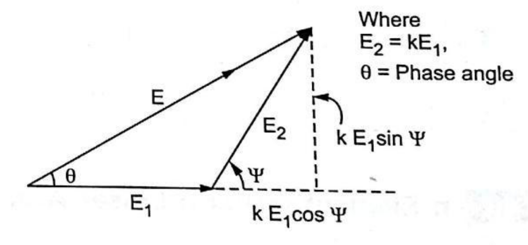
Field pattern for two point sources with spacing d = $\frac{\lambda}{2}$ and fed with currents equal in magnitude but out of phase by 180^0

Two Point Sources with Currents Unequal in Magnitudes and with any Phase

If the two point sources are separated by distance d and supplied with currents which are different in magnitudes and with any phase difference say α. Consider that source 1 is assumed to be reference for phase and amplitude of the fields E_1 and E_2, which are due to source 1 and source 2 respectively at the distant point P. Let us assume that E_1 is greater than E_2 in magnitude as in diagram

Now the total phase difference between the radiations by the two point sources at any far point is given by

$$\Psi = \frac{2\pi}{\lambda}\cos\theta+\alpha$$



Vector Diagram of fields E_1 and E_2

where α is the phase angle with which current I_2 leads current I_1. Now if a then the condition is similar to the two point sources with currents equal in magnitude and phase. Similarly if α = 180°, then the condition is similar to the two point source with currents equal in magnitude but opposite in phase. Assume value of phase difference a as 0 < α < 180°. Then the resultant field at point P is given by,

$$E_T = E_1\, e^{-j0} + E_2\, e^{j\psi}$$

$$E_T = E_1 + E_2\, e^{j\psi}$$

$$E_T = E_1\ (E_1 + \frac{E_2}{E_1}\, e^{j\psi})$$

Let $\frac{E_2}{E_1}$= k note that $E_2 > E_1$, the value of k is less than unity. Moreover the value of k is given by $0 \leq k \leq 1$

Then E_T=E_1[1+k(cosψ + jsinψ)]

The magnitude of the resultant field at point P is given by

$$|E_T| = E_1\sqrt{(1 + kcos\psi)\text{^}2\ ) + (ksin\psi)^2}$$

The phase difference between two fields at the far point P is given by

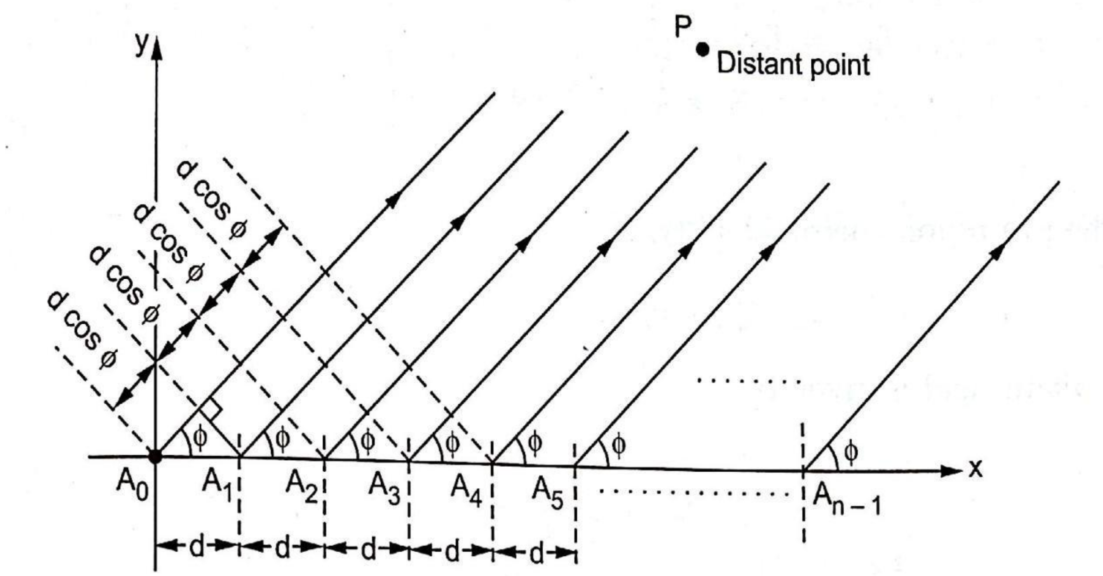
$$\theta = \tan^{-1}\frac{ksin\psi}{1 + kcos\psi}$$

n Element Uniform Linear Arrays

At higher frequencies, for point to point communications it is necessary to have a pattern with single beam radiation. Such highly directive single beam pattern can be obtained by increasing the point sources in the arrow from 2 to n say.

An array of n elements is said to be linear array if all the individual elements are spaced equally along a line. An array is said to be uniform array if the elements in the array are fed with currents with equal magnitudes and with uniform progressive phase shift along the line.

Consider a general n element linear and uniform array with all the individual elements spaced equally at distance d from each other and all elements are fed with currents equal in magnitude and uniform progressive phase shift along line as shown in figure.



Uniform, linear array of n elements

The total resultant field at the distant point P is obtained by adding the fields due to n individual sources vectorically. Hence we can write,

$$E_T = E_0\, e^{j0} + E_0\, e^{j\psi} + E0\, e^{2j\psi} + \ldots\ldots + E_0\, e^{j(n-1)\psi}$$

$$E_T = E_0[\, e^{j0} + e^{j\psi} + + \ldots\ldots + e^{j(n-1)\psi}]$$

Note that $\psi = (\beta d\cos(\theta) + \alpha)$ indicates the total phase difference of the fields from

adjacent sources calculated at point P. Similarly α is the progressive phase shift between two adjacent point sources. The value of a may lie between 0° and 180°. If α = 0°, we get n element uniform li near broadside array. If α = 180°, we get n element uniform linear-end-fire-array.

Multiplying above equation by $e^{j\psi}$, we get,

$$E_T e^{j\psi} = E_0[e^{j\psi} + e^{j2\psi} + +.......+ e^{jn\psi}]$$

Subtracting and the above two equations and simplifying we will get

$$E_T = E_0 \left[\frac{1-e^{jn\psi}}{1-e^{j\psi}}\right]$$

Simplifying we will get

$$E_T = E_0 \left[\frac{\sin n\frac{\psi}{2}}{\sin\frac{\psi}{2}}\right] e^{j\left(\frac{n-1}{2}\right)\psi}$$

Then the magnitude of the resultant field is given by

$$E_T = E_0 \left[\frac{\sin n\frac{\psi}{2}}{\sin\frac{\psi}{2}}\right]$$

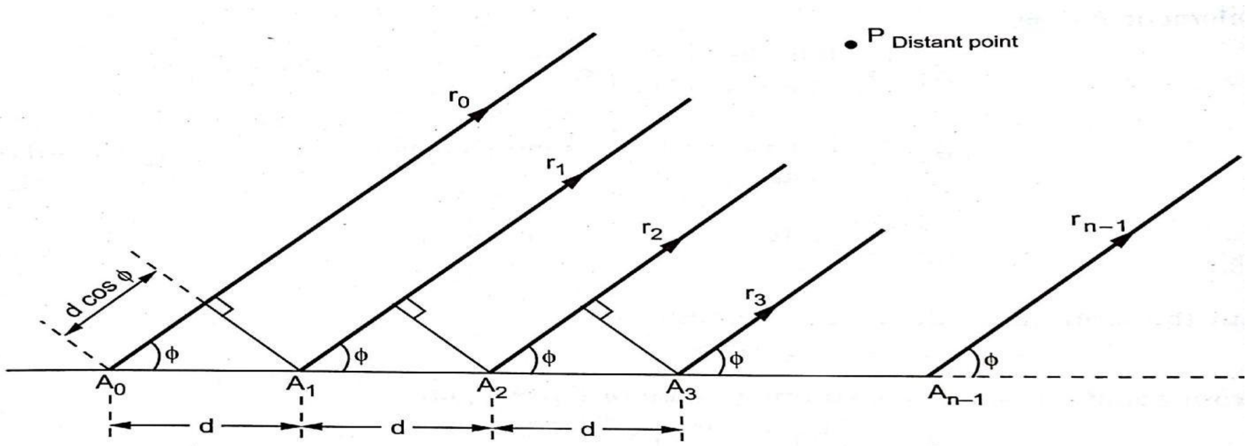
The phase angle θ of the resultant field at point P is given by,

$$\theta = \left(\frac{n-1}{2}\right)\psi = \beta d\cos\phi + \alpha$$

Array of n Elements with Equal Spacing and Currents Equal in Magnitude and Phase - Broadside Array

Consider the 'n' number of identical radiators carry currents which are equal magnitude and in phase. The identical radiators are equispaced. Hence the maxim radiation occurs in the directions normal to the line of array. Hence such an array known as Uniform broadside array.

Consider a broadside array with n identical radiators as shown in figure.



The electric field produced at point P due to an element A_0 is given by,

$$E_0 = \frac{IdL\, sin\theta}{4\pi\omega\varepsilon_0}\left[j\frac{\beta^2}{r_0}\right] e^{-j\beta r_0}$$

As the distance of separation d between any two array elements is very sai compared to the radial distances of point P from A_0, A_1, ...A_{n-1}, we can assume r_1, r_2, $r_{n-1}$are approximately same.

Now the electric field produced at point P due to an element A_1 will differ in as r_0 and r_1 are not actually same. Hence the electric field due to A_1 is given by,

$$E_1 = \frac{IdL\, sin\theta}{4\pi\omega\varepsilon_0}\left[j\frac{\beta^2}{r_1}\right] e^{-j\beta r_1}$$

But

$$r_1 = r_0 - dcos\phi$$

$$E_1 = \frac{IdL\, sin\theta}{4\pi\omega\varepsilon_0}\left[j\frac{\beta^2}{r_0}\right] e^{-j\beta(r_0 - dcos\phi)}$$

$$E_1 = E_0 \; e^{j\beta d cos\emptyset}$$

Exactly on the similar lines we can write the electric fieto the ld produced at point P due to an element A_2 is

$$E_2 = \frac{IdL\, sin\theta}{4\pi\omega\varepsilon_0}\left[j\frac{\beta^2}{r_2}\right] e^{-j\beta r_2}$$

$$E_2 = \frac{IdL\, sin\theta}{4\pi\omega\varepsilon_0}\left[j\frac{\beta^2}{r_1}\right] e^{-j\beta(r_1 - dcos\phi)}$$

$$E_2 = E_1 \; e^{j\beta d cos\emptyset}$$

By substituting E_1 the equation becomes

$$E_2 = E_1 \; e^{j2\beta d cos\emptyset}$$

Similarly

$$E_{n-1} = E_0 \; e^{j(n-1)\beta d cos\emptyset}$$

Then the total electric field at point P becomes

$$E_T = E_0 + E_1 + \ldots\ldots\ldots\ldots + E_{n-1}$$

$$E_T = E_0 + E_0 e^{j\beta d cos\emptyset} + E_0 e^{j2\beta d cos\emptyset} + \cdots + E_0 e^{j(n-1)\beta d cos\emptyset}$$

By writing the $\beta dcos\phi = \psi$ then the above equation becomes

$$E_T = E_0 + E_0 e^{j\psi} + E_0 e^{j2\psi} + \cdots + E_0 e^{j(n-1)\psi}$$

Considering the series by $s = 1 + r + r^2 + r^3 + \ldots\ldots\ldots r^{n-1}$

Where $r = e^{j\psi}$

Multiplying the above equation on the both sides by r and simplifying we will get

$$s = \frac{1 - r^n}{1 - r}$$

By this series the ET becomes

$$E_T =$$

$$E_0 \left[\frac{1-e^{jn\psi}}{1-e^{j\psi}}\right] \quad E_T = E_0 \frac{e^{jn\frac{\psi}{2}}}{e^{j\frac{\psi}{2}}} \left[\frac{e^{-jn\frac{\psi}{2}} - e^{jn\frac{\psi}{2}}}{e^{-j\frac{\psi}{2}} - e^{j\frac{\psi}{2}}}\right]$$

By simplifying we will get

$$\frac{E_T}{E_0} = e^{j(n-1)\frac{\psi}{2}} \left[\frac{\sin\left(\frac{n\psi}{2}\right)}{\sin\left(\frac{\psi}{2}\right)}\right]$$

The exponential term in above equation represents the phase shift. Now considering magnitudes of the electric fields, we can write

$$\left|\frac{E_T}{E_0}\right| = \frac{\sin\left(\frac{n\psi}{2}\right)}{\sin\left(\frac{\psi}{2}\right)}$$

Properties of Broadside Array

1. Major lobe

In case of broadside array, the field is maximum in the direction normal to the axis of the array. Thus the condition for the maximum field at point P is given by $\psi = 0$ i.e. $\beta d\cos\phi = 0$

i.e. $\cos\phi = 0$

i.e. $\phi = 90^0$ or 270^0

Thus $\phi = 90°$ and $\phi = 270°$ are called directions of principle maxima.

2. Magnitude of major lobe

The maximum radiation occurs when $\psi = 0$. Hence we can write,

$$|\text{Major lobe}| = \lim_{\psi \to 0} \left\{ \frac{\frac{d}{d\psi}\left(sinn\frac{\psi}{2}\right)}{\frac{d}{d\psi}\left(sin\frac{\psi}{2}\right)} \right\}$$

$$|\text{Major lobe}| = n$$

where, n is the number of elements in the array.

Thus from equation, it is clear that, all the field components add up together to give total field which is 'n' times the individual field when ϕ = 90° or ϕ= 270°.

3. Nulls

The ratio of total electric field to an individual electric field is given by

$$\left|\frac{E_T}{E_0}\right| = \frac{\sin\left(\frac{n\psi}{2}\right)}{\sin\left(\frac{\psi}{2}\right)}$$

By making above equation to zero we can find the minima, but the above eq uation becomes zero then $\frac{n\psi}{2} = \pm m\pi$

Now ψ = βdcosϕ

$$\text{Therefore } \frac{n}{2}\left(\frac{2\pi}{\lambda}d\right)cos\phi_{min} = \pm m\pi$$

$$\phi_{min} = \cos^{-1}\left(\pm\frac{m\lambda}{nd}\right)$$

N= number of elements in array

d= Spacing between elements in meter

λ = Wavelength in meter

m= constant = 1,2,3……..

4. Subsidary maxima (or side lobes)

The directions of the subsidiary maxima or side lobes can be obtained if in above equation

$\sin\left(\frac{n\psi}{2}\right) = \pm 1$

$n\frac{\psi}{2} = \pm\frac{3\pi}{2} \pm \frac{5\pi}{2}, \ldots \ldots \ldots$

Hence sin($n\frac{\psi}{2}) = \pm 1$ is not considered because if $n\frac{\psi}{2} = \frac{\pi}{2}$ then sin ($n\frac{\psi}{2}$) =1 whch is the direction of principle maxima

Hence we can skip $n\frac{\psi}{2} = \pm\frac{\pi}{2}$ value

Thus, we can get

$\psi = \pm\frac{3\pi}{n} \pm \frac{5\pi}{n}, \ldots \ldots \ldots$

Now

$$\psi = \beta d\cos\theta$$

By simplifying we can get

$\phi = \cos^{-1}\left[\pm\frac{\lambda(2m+1)}{2nd}\right]$

The above equation represents the directions where certain radiation which is not maximum. Hence it represents directions of subsidary maxima or side lobes.

5. Beamwidth of major lobe

The beamwidth is defined as the angle between first nulls. Alternatively beamwidth is the angle equal to twice the angle between first null and the major lobe maximum direction.

Hence the beamwidth between first nulls is given by, BWFN = 2*γ, where γ = 90-ϕ

$$\phi_{min} = \cos^{-1}\left[\pm\frac{m\lambda}{nd}\right]$$, where m= 1, 2, 3………

And $$90-\gamma = \cos^{-1}\left[\pm\frac{m\lambda}{nd}\right]$$

Taking the cosine angle on both sides

$$\cos(90-\gamma) = \cos\left(\cos^{-1}\left[\pm\frac{m\lambda}{nd}\right]\right)$$

If γ is very small, sinγ=γ, substituting in the above equation, we can get

$$\gamma = \pm\frac{m\lambda}{nd}$$

But for the first null m=1

$$\gamma = +\frac{\lambda}{nd}$$

$$BWFN = 2\gamma = +\frac{2\lambda}{nd}$$

But nd ≈ (n-1) d if n is very large. This nd indicates total length of aray in meter. This is denoted by L.

$$BWFN = \frac{2\lambda}{L} = \frac{2}{\left(\frac{L}{\lambda}\right)}$$

$$BWFN = \frac{114.6}{L} = \frac{114.6}{\left(\frac{L}{\lambda}\right)} \text{ degrees}$$

The Half Power beam width (HPBW) is given by

$$HPBW = \frac{BWFN}{2} = \frac{1}{\left(\frac{L}{\lambda}\right)}$$

$$HPBW = \frac{57.3}{\left(\frac{L}{\lambda}\right)} \text{ degrees}$$

6. Directivity

The directivity is defined as

$$G_{Dmax} = \frac{Maximum\ radiation\ intensity}{Average\ radiation\ Intensity} = \frac{U_{max}}{U_{avg}} = \frac{U_{max}}{U_0}$$

Where the U_0 is the average radiation intensity and is given by

$$U_0 = \frac{1}{4\pi}\int_{\emptyset=0}^{2\pi}\int_{\theta=0}^{\pi}|E(\theta,\emptyset|^2 sin\theta d\theta d\emptyset$$

From the expression of ratio of magnitudes we can write,

$$\left|\frac{E_T}{E_0}\right| = n$$

For the normalized condition

$$|E_T| = n$$

Thus field from array is maximum in any direction 8 when w = 0. Hence normalized field pattern is given by,

$$E_{Normalized} = \left|\frac{E_T}{E_{T\,max}}\right| = \frac{1}{n}\left|\frac{E_0}{E_0}\right| = \frac{1}{n}$$

Hence the field is given by,

$$E_{Normalized} = \frac{\sin\left(\frac{n\psi}{2}\right)}{n\sin\left(\frac{\psi}{2}\right)}$$

Where $\psi = \beta d\cos\theta$

The equationindicates array factor, hence we can write, the electric field due to n arrays as

$$E = \frac{1}{n}\left[\frac{sin\frac{n\beta d cos\emptyset}{2}}{sin\frac{\beta d cos\emptyset}{2}}\right]$$

Assuming d very small as compared to length of array, we can approximate $sin\frac{n\beta dcos\emptyset}{2}=\frac{n\beta dcos\emptyset}{2}$

Substituting the value of E in equation we get

$$U_0=\frac{1}{4\pi}\int_{\emptyset=0}^{2\pi}\int_{\theta=0}^{\pi}\left[\frac{sin\frac{n\beta dcos\emptyset}{2}}{sin\frac{\beta dcos\emptyset}{2}}\right]^2 sin\theta d\theta d\emptyset$$

Let $z=\frac{n\beta dcos\emptyset}{2}$ and then simplifying we will get

$$U_0=-\frac{1}{n\beta d}\int_{\frac{n\beta d}{2}}^{-\frac{n\beta d}{2}}\left[\frac{sinz}{z}\right]^2 dz$$

For large array, n is large hence nf3d is also very large (assuming tending to ∞). Hence of threwriting above equation.

$$U_0=-\frac{1}{n\beta d}\int_{\infty}^{-\infty}\left[\frac{sinz}{z}\right]^2 dz$$

But $\left[\frac{sinz}{z}\right]^2 dz=\pi.$ so the equation becomes

$$U_0=\frac{1}{n\beta d}\pi$$

the directivity is given by

$$G_{Dmax}=\frac{Umax}{U_0}$$

But U_{max} = 1 at θ =90^0 and directivity we will get is

$$G_{Dmax}=\frac{2nd}{\lambda}$$

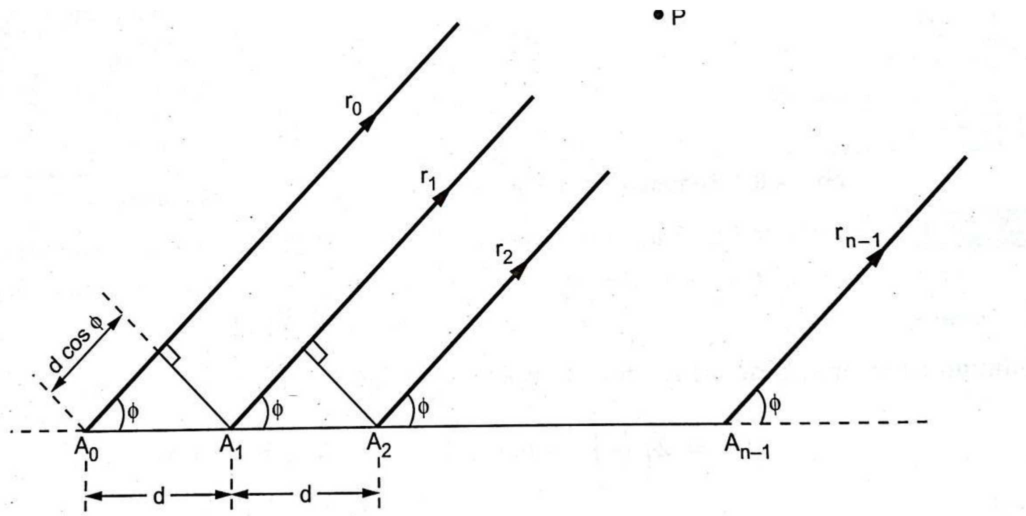
L= (n-1)d, d≈nd if n is very large

Then the directivity in terms of the total length array as

$$G_{Dmax} = 2\left(\frac{L}{\lambda}\right)$$

Array of n Elements with Equal Spacing and Currents Equal in Magnitude but with Progressive Phase Shift - End Fire Array

Consider n number of identical radiators supplied with equal current which are not in phase as shown in the figure. Assume that there is progressive phase lag of 13d radians in each radiator.



End fire Array

Consider that the current supplied to first element A_0 be I_0. Then the current supplied to A_1 is given by,

$$I_1 = I_0 e^{-j\beta d}$$

Similarly the current supplied to A 2 is given by,

$$I_2 = I_1 . e^{-j\beta d} = I_0 e^{-j2\beta d}$$

Thus the current supplied to the last element is given by,

$$I_{n-1} = I_0 e^{-j(n-1)\beta d}$$

The electric field produced at point P, due to A_0 is given by,

$$E_0 = \frac{IdL\,sin\theta}{4\pi\omega\varepsilon_0}\left[j\frac{\beta^2}{r_0}\right]e^{-j\beta r_0}$$

The electric field produces at point P due to A_1 is given by

$$E_1 = \frac{IdL\,sin\theta}{4\pi\omega\varepsilon_0}\left[j\frac{\beta^2}{r_1}\right]e^{-j\beta r_1}\,e^{-j\beta d}$$

But

$$r_1 = r_0 - dcos\phi$$

$$E_1 = \frac{IdL\,sin\theta}{4\pi\omega\varepsilon_0}\left[j\frac{\beta^2}{r_0}\right]e^{-j\beta(r_0 - dcos\phi)}e^{-j\beta d}$$

$$E_1 = E_0\; e^{j\beta d(cos\emptyset - 1)}$$

Let $E_1 = E_0\, e^{j\psi}$

The electric field at point P due to A2 is given by

$$E_2 = E_0\;\; e^{j2\psi}$$

Similarly

$$E_{n-1} = E_0\;\; e^{j(n-1)\psi}$$

Then the total electric field at point P becomes

$$E_T = E_0 + E_1 + \ldots\ldots\ldots\ldots.. + E_{n-1}$$

$$E_T = E_0 + E_0 e^{j\psi} + E_0 e^{j2\psi} + \cdots + E_0 e^{j(n-1)\psi}$$

Considering the series by $s = 1 + r + r^2 + r^3 + \ldots\ldots\ldots r^{n-1}$

Where $r = e^{j\psi}$

Multiplying the above equation on the both sides by r and simplifying we will get

$$s = \frac{1 - r^n}{1 - r}$$

By this series the E_T becomes

$$E_T = E_0 \left[\frac{1 - e^{jn\psi}}{1 - e^{j\psi}}\right]$$

$$E_T = E_0 \frac{e^{jn\frac{\psi}{2}}}{e^{j\frac{\psi}{2}}} \left[\frac{e^{jn\frac{\psi}{2}} - e^{jn\frac{\psi}{2}}}{e^{-j\frac{\psi}{2}} - e^{j\frac{\psi}{2}}}\right]$$

By simplifying we will get the magnitude as

$$\left|\frac{E_T}{E_0}\right| = \left[\frac{\sin\left(\frac{n\psi}{2}\right)}{\sin\left(\frac{\psi}{2}\right)}\right]$$

Properties of End Fire Array

1. Major lobe

For the end fire array where currents supplied to the antennas are but the phase changes progressively through array, the phase angle is

$\psi = \beta d(\cos(\theta - 1))$

i.e. $\cos\theta = 1$

$\theta = 0^0$

Thus $\theta = 0^0$ indicates the direction of principle maxima. Also it indicates that maximum radiation is along the axis of array or line of array.

2. Magnitude of major lobe

The maximum radiation occurs when $\psi = 0$. Hence we can write,

$$|\text{Major lobe}| = \lim_{\psi \to 0} \left\{ \frac{\frac{d}{d\psi}\left(sinn\frac{\psi}{2}\right)}{\frac{d}{d\psi}\left(sin\frac{\psi}{2}\right)} \right\}$$

$$|\text{Major lobe}| = n$$

where, n is the number of elements in the array.

Thus from equation, it is clear that, all the field components add up together to give total field which is 'n' times the individual field when $\theta = 0^0$

3. Nulls

The ratio of total electric field to an individual electric field is given by

$$\left|\frac{E_T}{E_0}\right| = \frac{\sin\left(\frac{n\psi}{2}\right)}{\sin\left(\frac{\psi}{2}\right)}$$

By making above equation to zero we can find the minima, but the above equation becomes zero then $\frac{n\psi}{2} = \pm m\pi$

$$n\beta d(\cos\theta-1)/2 = \pm m\pi$$

By simplifying we will get

$$\frac{nd}{\lambda}(cos\phi - 1) = \pm m$$

$$\phi_{min} = \cos^{-1}\left(1 - \frac{m\lambda}{nd}\right)$$

N= number of elements in array

d= Spacing between elements in meter

λ = Wavelength in meter

m= constant = 1,2,3……..

Then $\frac{\emptyset_{min}}{2} = \sin^{-1}\left[\pm\sqrt{\frac{m\lambda}{2nd}}\right]$

4. Subsidary maxima (or side lobes)

The directions of the subsidiary maxima or side lobes can be obtained if in above equation

$$\sin\left(\frac{n\psi}{2}\right) = \pm 1$$

$$n\frac{\psi}{2} = \pm\frac{3\pi}{2} \pm \frac{5\pi}{2}, \dots \dots \dots$$

Hence sin($n\frac{\psi}{2}$) $= \pm 1$ is not considered because if $n\frac{\psi}{2} = \frac{\pi}{2}$then sin ($n\frac{\psi}{2}$) =1 whch is the direction of principle maxima

Hence we can skip $n\frac{\psi}{2} = \pm\frac{\pi}{2}$value

Thus, we can get

$n\frac{\psi}{2} = \pm(2m+1)\frac{\pi}{2}$, where m=1, 2, 3…………

$$\frac{n\beta d(cos\emptyset - 1)}{2} = \pm(2m+1)\frac{\pi}{2}$$

By simplifying we can get

$$\phi = \cos^{-1}\left[1 - \frac{\lambda(2m+1)}{2nd}\right]$$

The above equation represents the directions where certain radiation which is not maximum. Hence it represents directions of subsidary maxima or side lobes.

5. Beamwidth of major lobe

The Beamwidth of the end fire array is greater than broad side array.

Beamwidth = 2*Angle between first nulls and maximum of the major lobe i.e. θ_{min}

$$\frac{\emptyset_{min}}{2} = \sin^{-1}\left[\pm\sqrt{\frac{m\lambda}{2nd}}\right]$$

If $\frac{\emptyset_{min}}{2}$ is very low, then we can write $\sin\frac{\emptyset_{min}}{2} \approx \frac{\emptyset_{min}}{2}$.Using this property in above equation we will get

$$\frac{\emptyset_{min}}{2} = \pm\sqrt{\frac{m\lambda}{2nd}}$$

$$\phi_{min} = \pm\sqrt{\frac{4m\lambda}{2nd}}$$

But n=L i.e. length of the antenna array, so the equation can be written as

$$\phi_{min} = \pm\sqrt{\frac{2m\lambda}{L}}$$

$$BWFN = 2\,\phi_{min} = \pm 2\sqrt{\frac{2m}{\frac{L}{\lambda}}}$$

$$BWFN = 2\,\phi_{min} = \pm 2\sqrt{\frac{2m}{\frac{L}{\lambda}}} * 57.3$$

$$BWFN = \pm 114.6\sqrt{\frac{2m}{\frac{L}{\lambda}}} \text{ degree}$$

6. Directivity

The directivity is defined as

$$G_{Dmax} = \frac{Maximum\ radiation\ intensity}{Average\ radiation\ Intensity} = \frac{U_{max}}{U_{avg}} = \frac{U_{max}}{U_0}$$

$$G_{Dmax} = 4\left(\frac{nd}{\lambda}\right)$$

But nd = L = Length of the array then

$$G_{Dmax} = 4\left(\frac{L}{\lambda}\right)$$

Array of n Elements with Equal Spacing and Currents with Equal Amplitude and Progressive Phase Shift-End Fire Array with Increased Directivity

The maximum radiation can be obtained along the axis of the uniform end fire array if the progressive phase shift a between the elements is given by,

$$\alpha = \pm\beta d = -\beta d \text{ for maximum in } \theta = 0^0 \text{ direction}$$

$$= +\beta d \text{ for maximum in } \theta = 180^0 \text{ direction}$$

It is found that the field produced in the direction θ = 0° is maximum; but the directivity is not maximum. In many applications it is necessary to have the maximum possible directivity of the linear array.

In 1938, Hansen and Woodyard proposed certain conditions for the end fire case which are helpful in enhancing the directivity without altering other characteristics of the end fire array. These conditions are popularly known as Hansen-Woodyard conditions for End Fire Radiation. According To Hansen-Woodyard conditions, the phase-shift between closely spaced radiators of a very long array should be

$$\alpha = (\beta d + 2.94/n) \approx -\left(\beta d + \frac{\pi}{n}\right) \text{ for maximum in } \theta = 0^0 \text{ direction}$$

$$\text{and } \alpha = (\beta d + 2.94/n) \approx +\left(\beta d + \frac{\pi}{n}\right) \text{ for maximum in } \theta = 180^0 \text{ direction}$$

Note that with above conditions also maximum possible directivity cannot be achieved. That means the maximum may not even occur at ϕ=0° and ϕ=180°, its magnitude maximum may not be unit and even side lobe level may not be -13.46 dB. Basically the maxima level and side lobe level, both depend on 'n' i.e. number of elements in the array.

The enhanced directivity due to Hansen-Woodyard conditions can be realized by using above equation along with assumptions for $|\psi|$ values given as below.

1) For maximum radiation along $\theta = 0^0$

$$|\psi| = |\beta d cos\emptyset + \alpha|_{\emptyset=0^0} = \frac{\pi}{n}$$

and

$$|\psi| = |\beta d cos\emptyset + \alpha|_{\emptyset=180^0} = \pi$$

ii) For maximum radiation along $\theta = 180°$

$$|\psi| = |\beta d cos\emptyset + \alpha|_{\emptyset=1800^0} = \frac{\pi}{n}$$

and

$$|\psi| = |\beta d cos\emptyset + \alpha|_{\emptyset=0^0} = \pi$$

Even though above equations represent conditions obtained from equation of first set only, the precaution must be taken to fulfill the condition $|\psi| = \pi$ for each array. In general, for an array of n elements, the condition $|\psi| = \pi$ It can be satisfied by using equation of first set for $\theta = 0°$ and $\theta = 180°$ by selecting the spacing between two elements as,

$$d = \left(\frac{n-1}{n}\right)\frac{\lambda}{4}$$

If the number of elements is considerably large, then we can write,

$$d = \frac{\lambda}{4}$$

Hence for large uniform array, the Hansen-Woodyard conditons illustrate enhanced directivity if the spacing between the two adjacent elements is approximately λ/ 4

Consider n element array. The array factor of the n-element array is given by,

$$(AF)_n = \frac{1}{n}\left[\frac{\sin\frac{n\psi}{2}}{\sin\frac{\psi}{2}}\right]$$

But ψ = βdcosθ +α. Putting in above equation ,we can get

$$(AF)_n = \frac{1}{n}\left[\frac{\sin\frac{n}{2}(\beta d\cos\emptyset+\alpha)}{\sin\frac{1}{2}(\beta d\cos\emptyset+\alpha)}\right]n$$

For smaller values of ψ the above expression becomes

$$(AF)_n = \frac{1}{n}\left[\frac{\sin\frac{n}{2}(\beta d\cos\emptyset+\alpha)}{\frac{1}{2}(\beta d\cos\emptyset+\alpha)}\right]n$$

$$(AF)_n = \left[\frac{\sin\frac{n}{2}(\beta d\cos\emptyset+\alpha)}{\sin\frac{1}{2}(\beta d\cos\emptyset+\alpha)}\right]$$

Let the progressive phase shift be α = -pd, where p is constant. Then above equation becomes

$$(AF)_n = \left[\frac{\sin\frac{n}{2}(\beta d\cos\emptyset-pd)}{\frac{n}{2}(\beta d\cos\emptyset-pd)}\right]$$

$$(AF)_n = \left[\frac{\sin\frac{nd}{2}(\beta d\cos\emptyset-p)}{\frac{nd}{2}(\beta d\cos\emptyset-p)}\right]$$

Let $\frac{nd}{2} = q$. Hence the above equation becomes

$$(AF)_n = \left[\frac{\sin q(\beta d\cos\emptyset - p)}{q(\beta d\cos\emptyset - p)}\right]$$

Let $q(\beta d\cos\emptyset - p) = z$ then the above equation becomes

$$(AF)_n = \left[\frac{\sin z}{z}\right]$$

The radiation intensity $U(\phi) = |(AF)n|^2 = \left[\frac{\sin z}{z}\right]^2$

At $\theta=0^0$, the radiation intensity is given by

$$U(\phi=0^0) = \left[\frac{\sin z}{z}\right]^2 = \left[\frac{\sin q(\beta - p)}{q(\beta - p)}\right]^2$$

Dividing above two equations and let z=q(β-p)

$U(\theta)_n = 1$

The directivity of the array factor is given by,

$$D_0 = \frac{4\pi U_{max}}{P_{rad}} = \frac{U_{max}}{\frac{P_{rad}}{4\pi}} = \frac{U_{max}}{U_0}$$

The average radiation intensity is given by

$$U_0 = \frac{P_{rad}}{4\pi} = \frac{1}{4\pi}\int_{\emptyset=0}^{2\pi}\int_{\theta=0}^{\pi} U(\theta)\sin\theta\, d\theta\, d\emptyset$$

$$U_0 = \frac{1}{2}\left[\frac{z}{\sin z}\right]^2 \frac{1}{4\pi}\int_{\emptyset=0}^{\pi}\frac{\sin z}{z}\sin\theta\ d\theta$$

By simplifying we can get

$$U_0 = \frac{1}{2\beta q}[g(v)] \text{ where } v = q(\beta - p)$$

$$g(v) = \left[\frac{v}{sinv}\right]^2 \left[\frac{\pi}{2} + \frac{\cos(2v) - 1}{2v} + \sin(2v)\right]$$

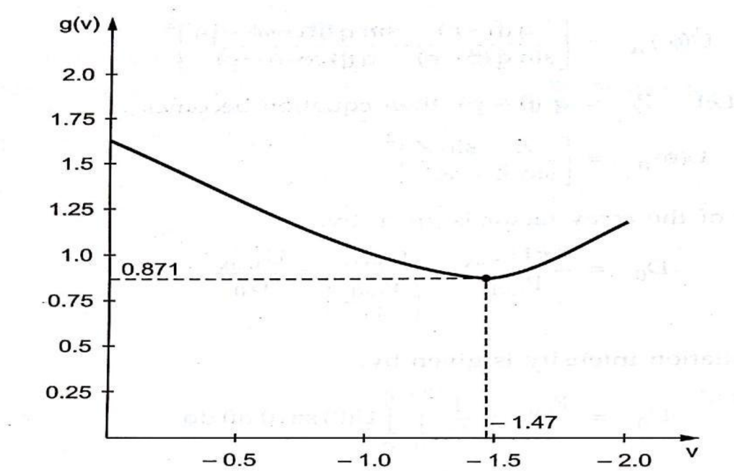
When the g(v) is plotted against v, its minimum value appears when

$$v = q(\beta\text{-}p) = \frac{nd}{2}(\beta - p) = -1.47$$

$$\alpha = -pd = -\left(\beta d + \frac{2.94}{n}\right)$$

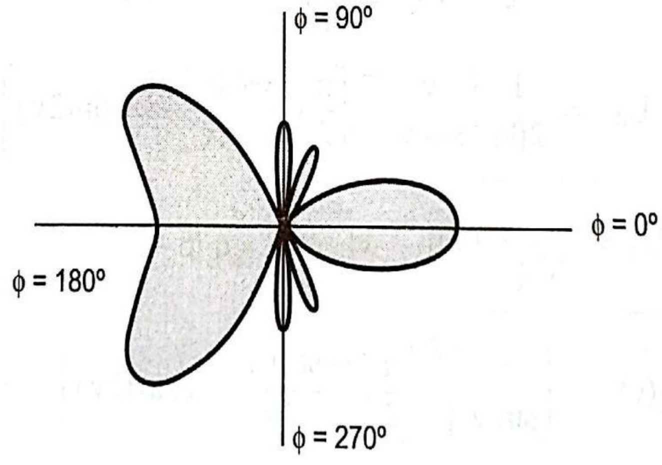
The above equation gives the condition for the end fire array with enhanced directivity based on Hansen woodyard conditions

The variation g(v) as a function of v is as figure below.



variation g(v) as a function of v

The field pattern for 4 element end fire array with equal amplitude and $\frac{\lambda}{2}$ spacing for increased directivity is as drawn below



field pattern for 4 element end fire array with equal amplitude and $\frac{\lambda}{2}$ spacing for increased directivity

Directivity of end fire array with increased directivity

For end fire array with increased directivity and maximum radiation in ϕ=00 direction, the radiation on intensity for small spacing between elements (d<<λ) is given

$$U_0 = \frac{1}{n\beta d}\left(\frac{\pi}{2}\right)^2\left[\frac{\pi}{2}+\frac{2}{\pi}-1.8515\right]$$

By simplifying we can get

$$U_0 = 0.559\left(\frac{\pi}{2n\beta d}\right)$$

And the directivity becomes

D =1.789(4(L/λ)) where L= (n-1)d = nd

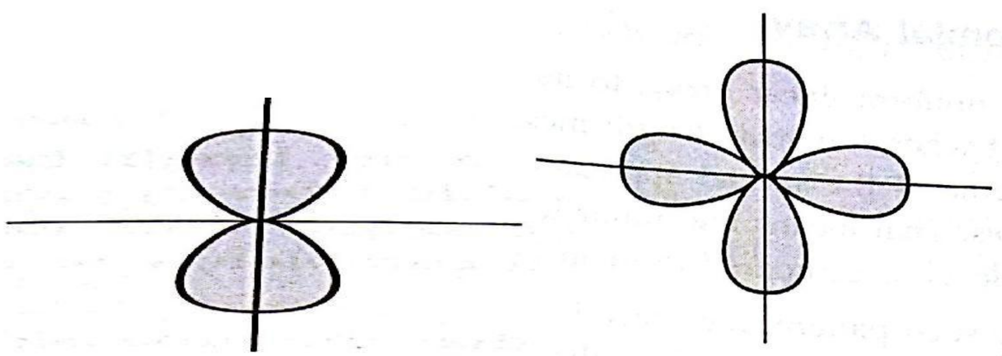
Pattern Multiplication Method

The simple method of obtaining the patterns of the arrays. This method is known as pattern multiplication method. This method is a very useful in the design of arrays because it makes possible to draw the patterns of complicated arrays rapidly, almost by inspection. To illustrate this method, consider 4 element array of equispaced identical antennas as shown in the figure. Let the spacing between two units be d = λ/2. Also assume that all the elements are supplied with equal magnitude currents which are in phase.

As the point P at which the resultant field has to be obtained is far away, we can assume the radiation from the antenna in the form of parallel lines.

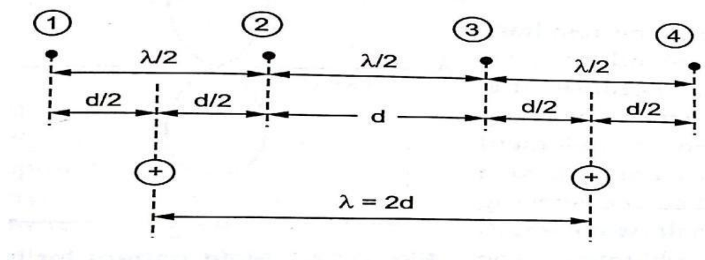
The radiation pattern of the antennas (1) and (2) treated to be operating as a single unit is as shown in the Figure (a). Similarly the radiation pattern of the antennas (3) and (4), spaced d = λ/2 distance apart and fed with equal current in phase, treated to be operated as single unit is again as shown in the Figures(a).

Now instead of considering two separate elements (1) and (2), we can replace it by a single antenna located at a point midway between them $\frac{d}{2}$ as shown in the figure(c). Now Similarly replacing antennas (3) and (4) by single antenna having same pattern as shown in the Figure (c). Now both the antennas have bi direction pattern i.e. figure eight pattern spaced distance λ apart from each other, fed with equal currents in phase is as shown in the Figure (b). Now the resultant radiation pattern of four element array can be obtained as the multiplication of pattern as shown in the Figure (d). Note that this multiplication is polar graphical multiplication for different values of ϕ.

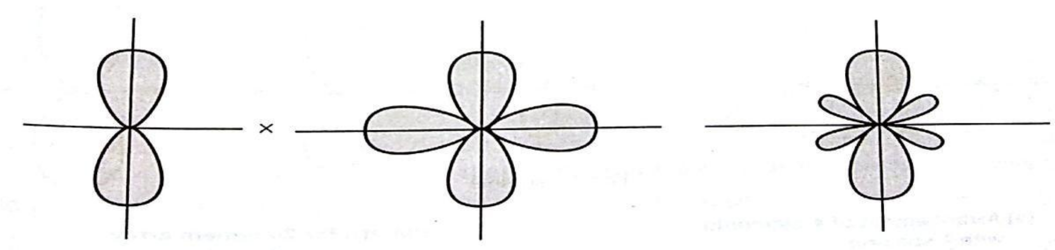


(a) Radiation Pattern of two antennas spaced at distance $\frac{\lambda}{2}$ and fed with equal currents in phase

(b) Radiation Pattern of two antennas spaced at distance λ and fed with equal currents in phase



(c) Array of 4 identical elements. Replacement of array by two single antennas placed at distance λ



(d) Multiplication of pattern

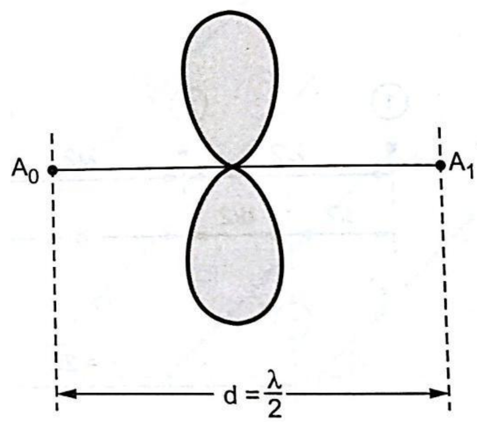
Binomial Array

In case of uniform linear array, to increase the directivity, the array length has to be increased. But when the array length increases, the secondary or side lobes appear in the pattern. In some of the special applications, it is desired to have single main lobe with no minor lobes. That means the minor lobes should be eliminated completely or reduced to minimum level as compared to main lobe.

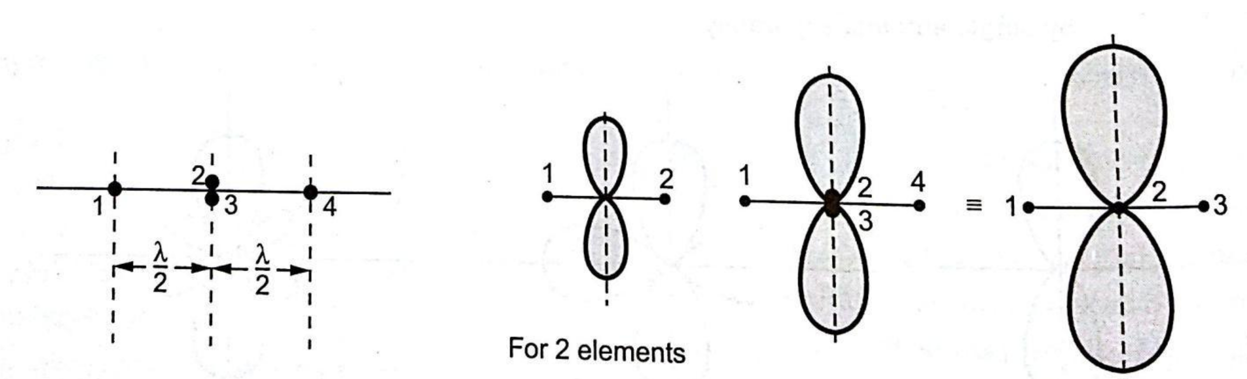
To achieve such pattern, the array is arranged in such a way that the broadside array radiate more strongly at the centre than that from edges. Let us consider array of the two identical in-phase point sources spaced λ/2 apart. Then the far-field pattern is given by

$$E = \cos\left(\frac{\pi}{2}\cos\theta\right)$$

In case of uniform 4-element array, the resultant pattern shows four side lobes. The secondary lobes appear in the resultant pattern, because the elements producing the group pattern have a spacing greater than one-half eave length. So 4 – element array, the elements producing pattern are spaced a full wave length apart. So if we reduce the spacing between two elements to one half wavelength then only the primary lobes are obtained.



Field pattern for two point sources with equal amplitude in-phase current



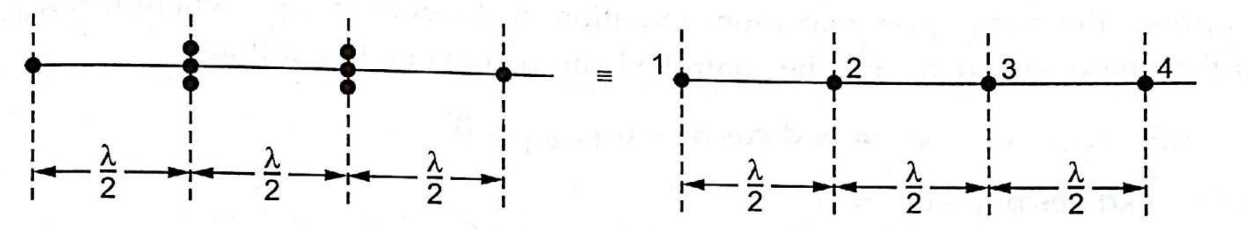
(a) Arrangement of 4-elements with λ/2 spacing

(b) Pattern for 2-element array and 4-element array

The two element arrays are spaced —2 distance apart from each other. Such array produces increased radiation pattern with no secondary lobes.

Here antenna 2 and 3 coincide at the centre as shown in the Figure (a). Hence it can be replaced by a single element carrying double current compared with other elements. Thus as shown in the Figure (b), the resultant array consists three elements with current ratio 1 : 2 : 1.

The same concept can be extended further by considering three element array as a unit and with a second similar three element array spaced half-wave length from it. This results in 4-element array as shown in the Fig. 4.12.3. In this array, the current ratio is given by 1 : 3 : 3 : 1.



Four element Array

If we continue this process, we can obtain the pattern with arbitrarily large directivity without minor lobes.. In general the pattern for the binomial array is given by

$$E = \cos^{n-1}[\pi/2\cos\theta]$$

n= number of sources in the array

In order to increase the directivity of an array its total length need to be increased. In this approach, number of minor lobes appears which are undesired for narrow beam applications. In has been found that number of minor lobes in the resultant pattern increases whenever spacing between elements is greater than λ/2. As per the demand of modern communication where narrow beam (no minor lobes) is preferred, it is the greatest need to design an array of only main lobes. The ratio of power density of main lobe to power density of the longest minor lobe is termed side lobe ratio. A particular technique used to reduce side lobe level is called tapering.

Since currents/amplitude in the sources of a linear array is non-uniform, it is found that minor lobes can be eliminated if the centre element radiates more strongly than the other sources. Therefore tapering need to be done from centre to end radiators of same specifications. The principle of tapering are primarily intended to broadside array but it is also applicable to end-fire array. Binomial array is a common example of tapering scheme and it is an array of n-isotropic sources of non-equal amplitudes. Using principle of pattern multiplication, John Stone first proposed the binomial array in 1929, where amplitude of the radiating sources arc arranged according to the binomial expansion. That is. if minor lobes appearing in the array need to be eliminated, the radiating sources must have current amplitudes proportional to the coefficient of binomial series, i.e. proportional o the coefficient of binomial series, i.e.

$$(1+x)^n = 1 + (n-1)x + \frac{(n-1)(n-2)}{!2}x^2 + \frac{(n-1)(n-2)(n-3)}{!3}x^3 \pm \cdots$$

where n is the number of radiating sources in the array.

For an array of total length nλ/2, the relative current in the nth element from the one end is given by

$$= \frac{n!}{r!(n-r)!}$$

where r = 0, 1, 2, 3, and the above relation is equivalent to what is known as Pascal's triangle. For example, the relative amplitudes for the array of 1 to 10 radiating sources are as follows:

No. of sources	*Pascal's triangle*
$n = 1$	1
$n = 2$	1 1
$n = 3$	1 2 1
$n = 4$	1 3 3 1
$n = 5$	1 4 6 4 1
$n = 6$	1 5 10 10 5 1
$n = 7$	1 6 15 20 15 6 1
$n = 8$	1 7 21 35 35 21 7 1
$n = 9$	1 8 28 56 70 56 28 8 1
$n = 10$	1 9 36 84 126 126 84 36 9 1

Since in binomial array the elements spacing is less than or equal to the half-wave length, the HPBW of the array is given by

$$\text{HPBW} = \frac{10.6}{\sqrt{n-1}} = \frac{1.06}{\sqrt{\frac{2L}{\lambda}}} = \frac{0.75}{\sqrt{L_\lambda}}$$

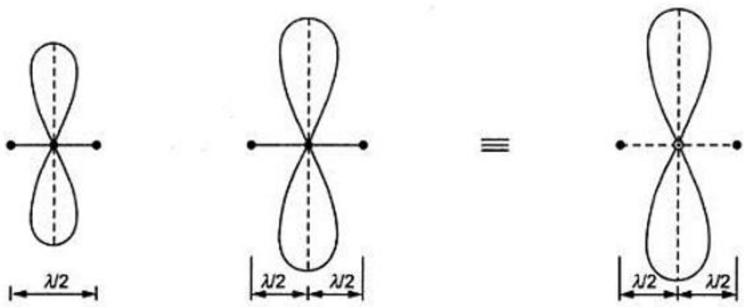
And dirrectivity

$$D_0 = 1.77\sqrt{n} = 1.77\sqrt{1 + 2L_\lambda}$$

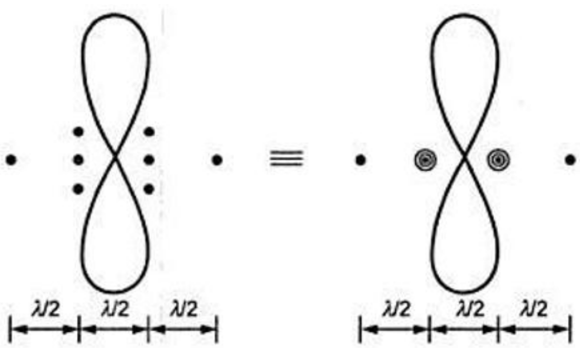
Using principle of multiplication, the resultant radiation pattern of an n-source binomial array is given by

In particular, if identical array of two point sources is superimposed one above other, then three effective sources with amplitude ratio 1:2:1 results. Similarly, in case three such elements are superimposed in same fashion, then an array of four sources is obtained whose current amplitudes are in the ratio of 1:3:3:1.

The far-field pattern can be found by substituting n = 3 and 4 in the above expression and they take shape as shown in below Fig.



Radiation pattern of 2-element array with amplitude ratio 1:2:1.



Radiation pattern of 3-element array with amplitude ratio 1:3:3:1.

It has also been noticed that binomial array offers single beam radiation at the cost of directivity, the directivity of binomial array is greater than that of uniform array for the same length of the array. In other words, in uniform array secondary lobes appear, but principle lobes are narrower than that of the binomial array.

Disadvantages of Binomial Array

(a) The side lobes are eliminated but the directivity of array reduced.

(b) As the length of array increases, larger current amplitude ratios are required.

PART – A

1. Define beam width? 2. Define broadside array. 3. What is end fire array. 4. Give the directivity expression for broadside array. 5. Give the directivity expression for end fire array. 6. Define pattern multiplication? 7. Define binomial array with necessary diagram.

PART – B

1. Derive the expression for broadside array and draw the radiation pattern for the same. 2. Derive the expression for end fire array and draw the radiation pattern for the same. 3. Derive the beam width and draw the radiation pattern for two point sources with equal amplitude and same phase. 4. Derive the beam width and draw the radiation pattern for two point sources with equal amplitude and opposite phase.
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4. With neat diagram explain the following(a) binomial array (b) pattern multiplication. 6. Derive the general expression for linear array of point sources.

Problems

Problem 1:For two element array consisting identical radiators carrying equal currents in phase, obtain positions of maxima and minima of the radiation pattern if the distance of separation d = λ.

Solution : i) Maxima : If $\phi = \pm \frac{\pi}{2}$, we get the maxima.

Also, another condition for maxima is given by,

$$\pi \frac{d}{\lambda} \cos\phi = 0, \pm \pi$$

But d = λ, thus the condition becomes,

$$\therefore \quad \pi \cos\phi = 0, \pm \pi$$

i.e. $\cos\phi = 0$ or $\cos\phi = \pm 1$ ... if $\pi \neq 0$

$$\therefore \quad \phi = \frac{\pi}{2} \text{ or } \phi = 0 \text{ or } \phi = \pi$$

Thus the positions of maxima are

$$\phi = 0, +\frac{\pi}{2}, -\frac{\pi}{2}, \pi$$

ii) Minima : The null condition is obtained if,

$$\pi\frac{d}{\lambda}\cos\phi = \pm\frac{\pi}{2}, \pm\frac{3\pi}{2}$$

But $d = \lambda$.

$$\therefore \quad \pi\cos\phi = \pm\frac{\pi}{2}, \pm\frac{3\pi}{2}$$

$$\therefore \quad \cos\phi = \pm\frac{1}{2} \text{ or } \pm\frac{3}{2}$$

Thus the positions of minima are,

$$\therefore \quad \phi = \pm 60^\circ \text{ or } \pm 120^\circ$$

Problem 2

Sketch the radiation pattern of a two element array having identical radiators spaced $\lambda/4$ apart and current in one radiator lags behind other by 90°.

Solution : For the two element array with $\lambda / 4$ separated radiators fed with currents of equal magnitude but phase difference of 90°, we can write,

$$\left|\frac{E_T}{E_0}\right| = 2\cos\left[\pi\left(\frac{\lambda/4}{\lambda}\right)\cos\phi - \frac{\pi/2}{2}\right] \qquad \ldots\ \alpha = 90^\circ = \frac{\pi}{2}\text{ rad}$$

$$\therefore \quad \left|\frac{E_T}{E_0}\right| = 2\cos\left[\frac{\pi}{4}\cos\phi - \frac{\pi}{4}\right] \quad \ldots (1)$$

Maxima : The maximum radiation is possible if

$$\cos\left[\frac{\pi}{4}\cos\phi - \frac{\pi}{4}\right] = 1$$

i.e. $\left[\frac{\pi}{4}\cos\phi - \frac{\pi}{4}\right] = 0, \ \pm\pi, \ \ldots..$

i.e. $\cos\phi = 1$ or $\cos\phi = 5$ or $\cos\phi = -3$

But $\cos\phi \ngtr 1$, selecting appropriate value of $\cos\phi$, we get,

$$\cos\phi = 1$$

$$\therefore \quad \phi = 0 \quad \ldots \text{ Only location of maximum radiation.}$$

Minima : The minimum radiation is possible if

$$\cos\left[\frac{\pi}{4}\cos\phi - \frac{\pi}{4}\right] = 0$$

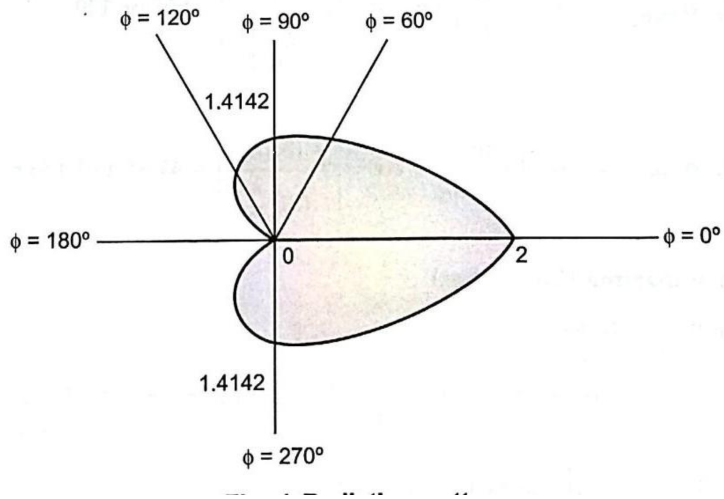
i.e. $\left[\frac{\pi}{4}\cos\phi - \frac{\pi}{4}\right] = \pm\frac{\pi}{2}, \pm\frac{3\pi}{2}, \ldots$

$$\frac{\pi}{4}\cos\phi = \frac{3\pi}{4}, -\frac{\pi}{4}, \ldots$$

$$\cos\phi = 3, -1, \ldots.$$

$$\cos\phi = -1 \quad \ldots \text{ Neglecting other values as } \cos\phi \ngtr 1.$$

$$\phi = 180^\circ = \pi^c \quad \ldots \text{ Only location of null radiation.}$$



Problem 3 : A broadside array of identical antennas consists 8 isotropic radiators separated by distance X/2. Find radiation field in a plane containing the line of array showing directions of maxima and null.

Solution: Given : n = 8, d = λ/2.

1) Major lobe

For broadside array, the direction of maxima is along the direction normal to axis the array. Hence the direction of the major lobe is given by,

(1) =90° and (I) =270'

2) Magnitude of major lobe

The magnitude of the major lobe is given by,

I Major lobe (= n = 8

3) Nulls

The directions of nulls are given by,

$$\phi_{min} = \cos^{-1}\left[\pm\frac{m\lambda}{nd}\right], \quad \text{where } m = 1, 2, 3, \ldots\ldots\ldots$$

$$\text{For } m = 1,\ \phi_{min_1} = \cos^{-1}\left[\pm\frac{(1)\lambda}{\left(\frac{\lambda}{2}\right)(8)}\right] = \cos^{-1}\left[\pm\frac{1}{4}\right] = \mathbf{75.52^\circ} \text{ and } \mathbf{104.47^\circ}$$

$$\text{For } m = 2,\ \phi_{min_2} = \cos^{-1}\left[\pm\frac{m\lambda}{n\,d}\right] = \cos^{-1}\left[\pm\frac{(2)(\lambda)}{(8)\left(\frac{\lambda}{2}\right)}\right] = \mathbf{60^\circ} \text{ or } \mathbf{120^\circ}$$

$$\text{For } m = 3,\ \phi_{min_3} = \cos^{-1}\left[\pm\frac{m\lambda}{n\,d}\right] = \cos^{-1}\left[\pm\frac{(3)(\lambda)}{(8)\left(\frac{\lambda}{2}\right)}\right] = \mathbf{41.4^\circ} \text{ and } \mathbf{138.6^\circ}$$

4) The subsidary lobes

The direction of side lobes is given by

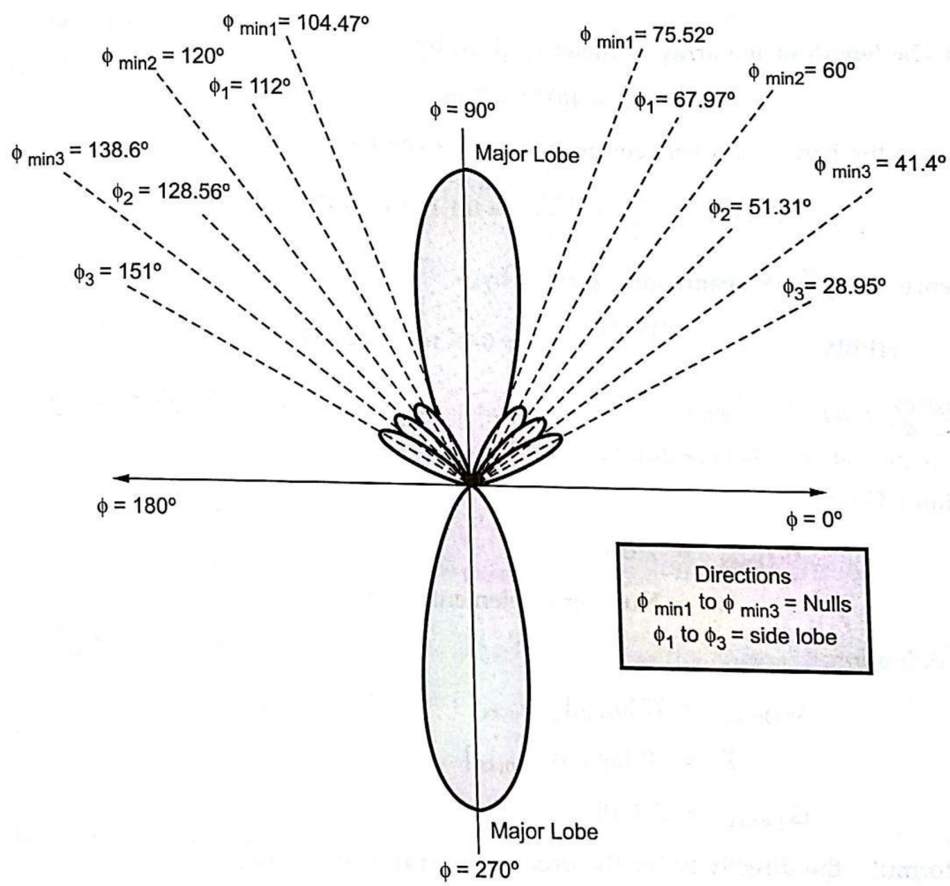
$$\phi = \cos^{-1}\left[\pm\frac{\lambda(2m+1)}{2nd}\right], \text{ where } m = 1, 2, 3, \ldots\ldots\ldots$$

$$\text{For } m = 1,\ \phi_1 = \cos^{-1}\left[\pm\frac{\lambda(2+1)}{2\times 8\times\frac{\lambda}{2}}\right] = \cos^{-1}\left[\pm\frac{3}{8}\right] = \mathbf{67.97^\circ} \text{ and } \mathbf{112^\circ}$$

The radiation pattern for the broadside array of 8 identical isotropic radiators is as shown in the figure.

$$\text{For } m = 2,\ \phi_2 = \cos^{-1}\left[\pm\frac{\lambda(4+1)}{2\times 8\times\frac{\lambda}{2}}\right] = \cos^{-1}\left[\pm\frac{5}{8}\right] = \mathbf{51.31^\circ} \text{ and } \mathbf{128.68^\circ}$$

$$\text{For } m = 3,\ \phi_3 = \cos^{-1}\left[\pm\frac{\lambda(6+1)}{2\times 8\times\frac{\lambda}{2}}\right] = \cos^{-1}\left[\pm\frac{7}{8}\right] = \mathbf{28.95^\circ} \text{ and } \mathbf{151^\circ}$$



UNIT III SPECIAL PURPOSE ANTENNAS

3.1 Loop Antennas

An RF current carrying coil is given a single turn into a loop, can be used as an antenna called as loop antenna. The currents through this loop antenna will be in phase. The magnetic field will be perpendicular to the whole loop carrying the current.

Frequency Range

The frequency range of operation of loop antenna is around 300MHz to 3GHz. This antenna works in UHF range.

Construction & Working of Loop Antennas

A loop antenna is a coil carrying radio frequency current. It may be in any shape such as circular, rectangular, triangular, square or hexagonal according to the designer's convenience.

Loop antennas are of two types.

- Large loop antennas
- Small loop antennas

Large loop antennas

Large loop antennas are also called as resonant antennas. They have high radiation efficiency. These antennas have length nearly equal to the intended wavelength.

$$L=\lambda$$

Where,

- L is the length of the antenna
- λ is the wavelength

The main parameter of this antenna is its perimeter length, which is about a wavelength and should be an enclosed loop. It is not a good idea to meander the loop so as to reduce the size, as that increases capacitive effects and results in low efficiency.

Small loop antennas

Small loop antennas are also called as magnetic loop antennas. These are less resonant. These are mostly used as receivers.

These antennas are of the size of one-tenth of the wavelength.

$$L=\lambda 10$$

Where,

- L is the length of the antenna
- λ is the wavelength

The features of small loop antennas are −

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The features of small loop antennas are −

- A small loop antenna has low radiation resistance. If multi-turn ferrite core constructions are used, then high radiation resistance can be achieved.
- It has low radiation efficiency due to high losses.
- Its construction is simple with small size and weight.

Due to its high reactance, its impedance is difficult to match with the transmitter. If loop antenna have to act as transmitting antenna, then this impedance mis-match would definitely be a problem. Hence, these loop antennas are better operated as receiver antennas.

Frequently Used Loops

Small loop antennas are mainly of two types −

- Circular loop antennas
- Square loop antennas

These two types of loop antennas are mostly widely used. Other types (rectangular, delta, elliptical etc.) are also made according to the designer specifications.

Fig 1: Circular loop antenna

Fig 2: Square loop antenna

The above images show circular and square loop antennas. These types of antennas are mostly used as AM receivers because of high Signal-to-noise ratio. They are also easily tunable at the Q-tank circuit in radio receivers.

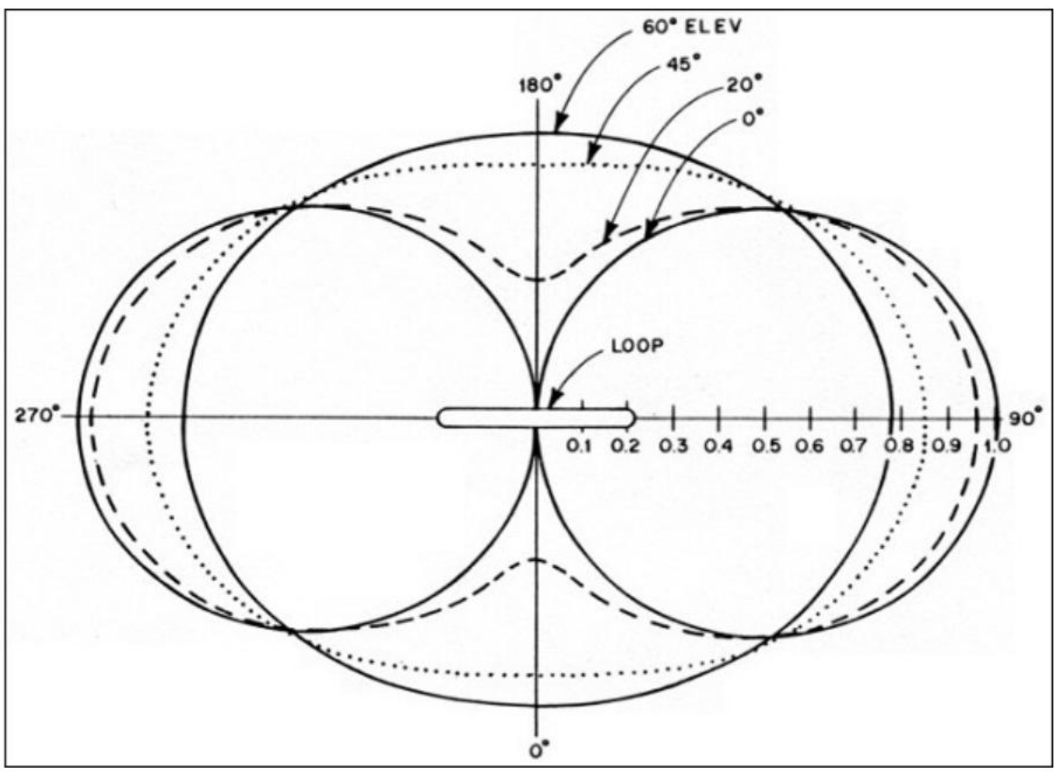
Polarization of Loop

The polarization of the loop antenna will be vertically or horizontally polarized depending upon the feed position. The vertical polarization is given at the center of the vertical side while the horizontal polarization is given at the center of the horizontal side, depending upon the shape of the loop antenna.

The small loop antenna is generally a linearly polarized one. When such a small loop antenna is mounted on top of a portable receiver, whose output is connected to a meter, it becomes a great direction finder.

Radiation Pattern

The radiation pattern of these antennas will be same as that of short horizontal dipole antenna.



Advantages

The following are the advantages of Loop antenna −

- Compact in size

- High directivity

Disadvantages

The following are the disadvantages of Loop antenna −

- Impedance matching may not be always good
- Has very high resonance quality factor

Applications

The following are the applications of Loop antenna −

- Used in RFID devices
- Used in MF, HF and Short wave receivers
- Used in Aircraft receivers for direction finding
- Used in UHF transmitters

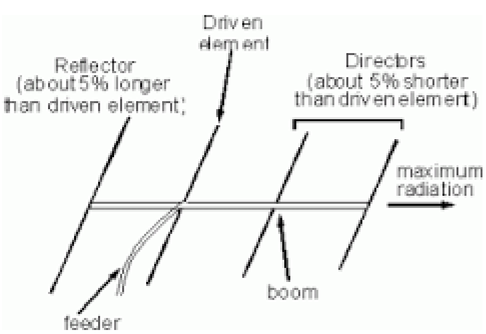
3.2 Yagi-Uda Array

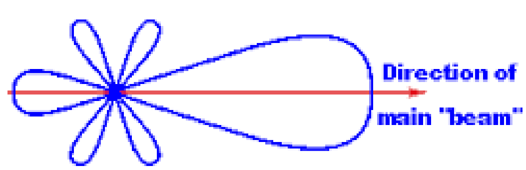
A Yagi-Uda array is an example of a parasitic array. Any element inan array which is not connected to the source (in the case of a transmitting antenna) or thereceiver (in the case of a receiving antenna) is defined as a parasitic element. A parasitic array isany array which employs parasitic elements. The general form of the N-element Yagi-Uda array isshown below.

Driven element - usually a resonant dipole or folded dipole.), folded dipoles are employed as driven elements to increase the array input impedance

Reflector - slightly longer than the driven element so that it isinductive (its current lags that of the driven element).Approximately 5 to 10 % longer than the driven element.

Director - slightly shorter than the driven element so that it iscapacitive (its current leads that of the driven element).Approximately 10 to 20 % shorter than the driven element), not necessarily uniform.

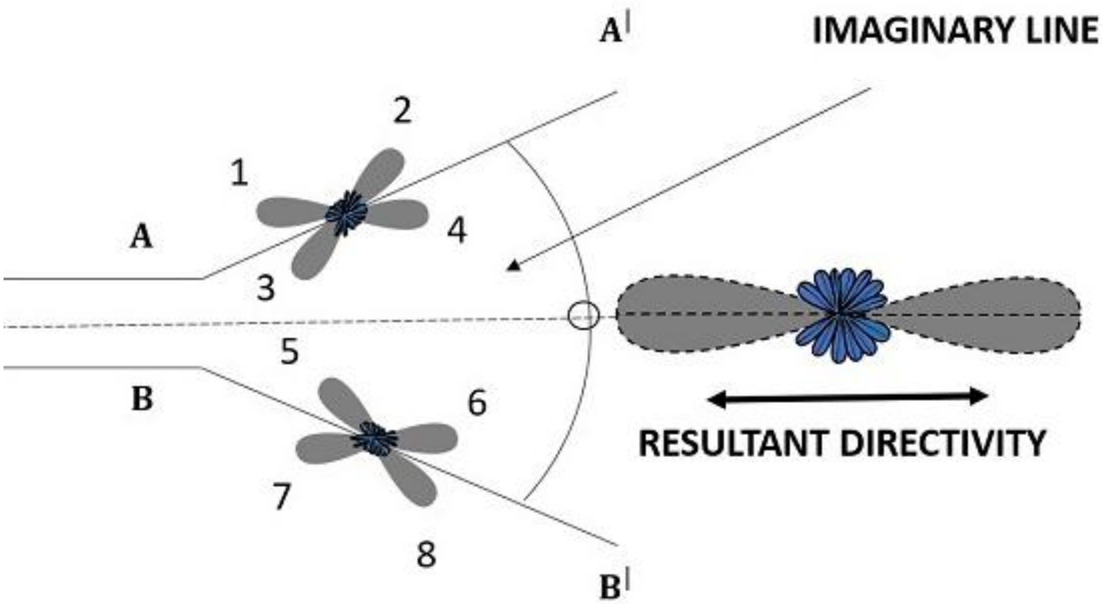




Advantages
1. Lightweight, Low cost
2.Simple construction
3.Unidirectional beam (front-to-back ratio)
4.Increased directivity over other simple wire antennas
5.Practical for use at HF (3-30 MHz), VHF (30-300 MHz), andUHF (300 MHz - 3 GHz)
Reflector spacing 0.1 to 0.258

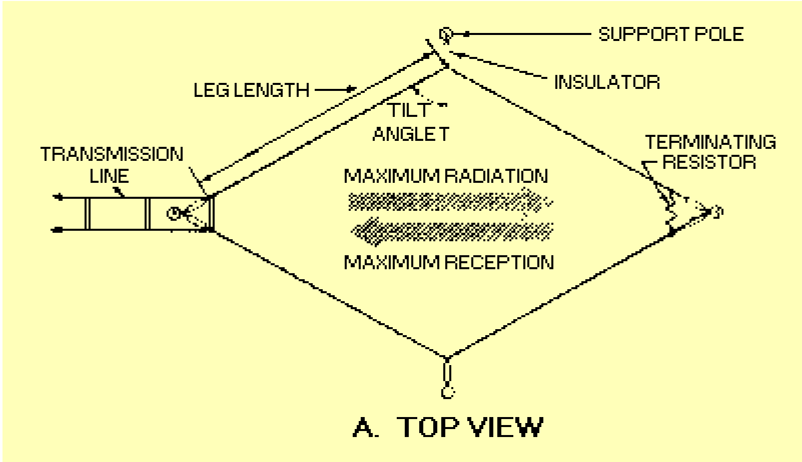
3.3 Vee Traveling Wave Antenna

The main beam of single electrically long wire guiding waves in one direction (traveling wave segment) was found to be inclined at an angle relative to the axis of the wire. Traveling wave antennas are typically formed by multiple traveling wave segments. These traveling wave segments can be oriented such that the main beams of the component wires combine to enhance the directivity of the overall antenna. A vee traveling wave antenna is formed by connecting two matched traveling wave segments to the end of a transmission line feed at an angle of 22degrees relative



3.4 Rhombic Antenna

The highest development of the long-wire antenna is the RHOMBIC ANTENNA . It consists of four conductors joined to form a rhombus, or diamond shape. The antenna is placed end to end and terminated by a noninductive resistor to produce a uni-directional pattern. A rhombic antenna can be made of two obtuse-angle V antennas that are placed side by side, erected in a horizontal plane, and terminated so the antenna is non resonant and unidirectional.



The rhombic antenna is widely used for long-distance, high-frequency transmission and reception. It is one of the most popular fixed-station antennas because it is very useful in point-to-point communication.

Radiation Patterns

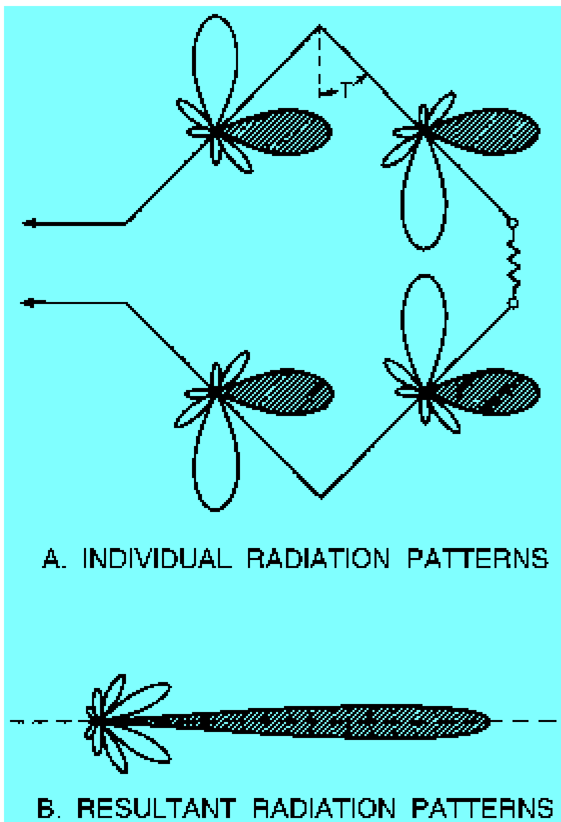


Figure a. shows the individual radiation patterns produced by the four legs of the rhombic antenna and the resultant radiation pattern. The principle of operation is the same as for the V and the half-rhombic antennas. Figure b. Formation of a rhombic antenna beam.

Advantages

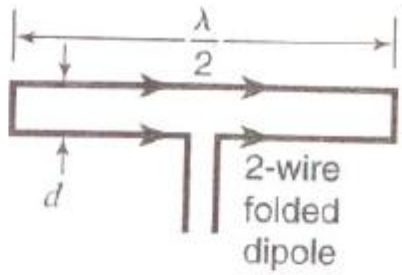
- The input impedance and radiation pattern of rhombic antenna do not change rapidly over a considerable frequency range.
- It is highly directional broad band antenna with greatest radiated or received power along the main axis.
- Simple and cheap to erect
- Low weight

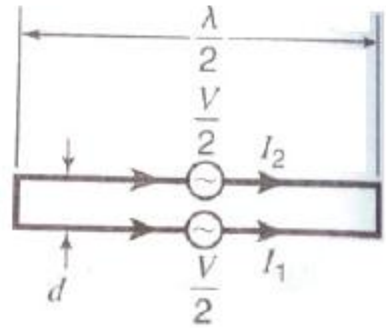
Disadvantages

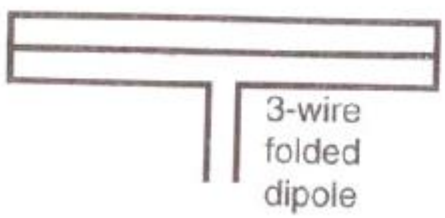
- It needs a larger space for installation
- Due to minor lobes, transmission efficiency is low

3.5 Folded Dipole:

A folded dipole is a dipole antenna with the ends folded back around and connected to each other, forming a loop as shown in Figure. It turns out the impedance of the folded dipole antenna will be a function of the impedance of a transmission line of length L/2. Also, because the folded dipole is "folded" back on itself, the currents can reinforce each other instead of cancelling each other out, so the input impedance will also depend on the impedance of a dipole antenna of length L.







The input impedance for a dipole is 73 Ω. Hence for a folded dipole with 2 arms the radiation resitance is 2* 73Ω= 292Ω. If 3 arms are used the resistance will be 3^2 * 73Ω = 657Ω

Advantages

- High input impedance
- Wide band in frequency
- Acts as built in reactance compensation network

Uses:

Folded dipole is used in conjunction with parasitic elements in wide band operation such as television. In this application, in the yagi antenna, the driven element is folded dipole and remaining are reflector and director

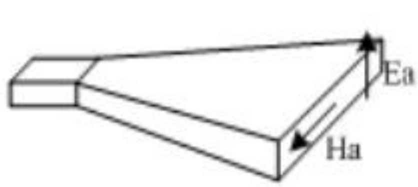
3.6 Horn Antennas

Horn antennas are popular in the microwave band (above 1 GHz). Horns provide high gain, low VSWR (with waveguide feeds), relatively widebandwidth, and they are not difficult to make. The horns can be also flared exponentially. This provides better matching ina broad frequency band, but is technologically more difficult and expensive.The rectangular horns are ideally suited for rectangular waveguide feeders.The horn acts as a gradual transition from a waveguide mode to a free-spacemode of the EM wave. When the feeder is a cylindrical waveguide, the antenna is usually a conical horn.

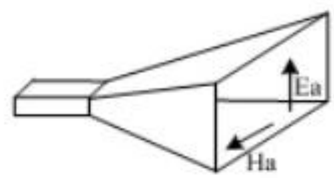
Types of the horn antennas as - Plane Sectoral Horn - Plane Sectoral Horn - Pyramidal and Conical Horn These horns are fed by a rectangular waveguide oriented its broad wall horizontal.

If flaring is done only in one direction, then it is called sectoral horn.Flaring in the direction of E and H, the sectoral E-plane and sectoral H plane are obtained respectively.If flaring is done along both the walls (E&H),then pyramidal horn is obtained.

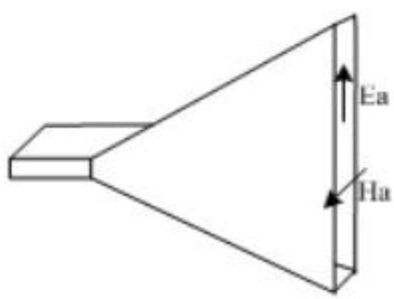
Horn antenna emphasizes traveling waves leads to wide bandwidth and low VSWR. Because of longer path length from connecting waveguide to horn edge, phase delay across aperture causes phase error. Dielectric or metallic plate lens in the aperture are used to correct phase error. Those with metallic ridges increase the bandwidth. Horns are also used for a feed of reflector antennas.



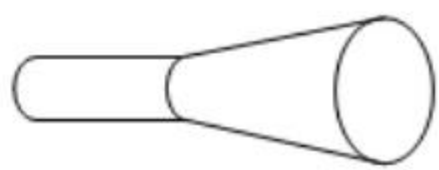
H-plane sectoral horn



Pyramidal horn



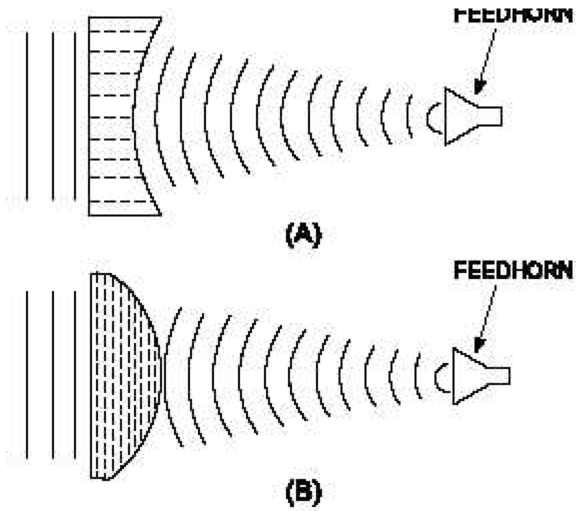
E-plane sectoral horn


Conical Horn Antenna

3.7 LENS ANTENNA:

Another antenna that can change spherical waves into flat plane waves is the lens antenna. This antenna uses a microwave lens, which is similar to an optical lens to straighten the spherical wavefronts. Since this type of antenna uses a lens to straighten the wavefronts, its design is based on the laws of refraction, rather than reflection.

Two types of lenses have been developed to provide a plane-wavefront narrow beam for tracking radars, while avoiding the problems associated with the feedhorn shadow. These are the conducting (acceleration) type and the dielectric (delay) type.



The lens of an antenna is substantially transparent to microwave energy that passes through it. It will, however, cause the waves of energy to be either converged or diverged as they exit the lens. This type of lens consists of flat metal strips placed parallel to the electric field of the wave and spaced slightly in excess of one-half of a wavelength. To the wave these strips look like parallel waveguides. The velocity of phase propagation of a wave is greater in a waveguide than in air. Thus, since the lens is concave, the outer portions of the transmitted spherical waves are accelerated for a longer interval of time than the inner

Advantages :

1. The lens antenna, feed and feed support do not block the aperture as the rays are transmitted away from the feed

2. It has greater design tolerance

3.It can be used to feed the optical axis and hence useful in applications where a beam is required to be moved angularly with respect to the axis.

3.8 Parobolic Reflector Antenna

A parabolic antenna is an antenna that uses a parabolic reflector, a curved surface with the cross-sectional shape of aparabola, to direct the radio waves. The most common form is shaped like a dish and is popularly called a dish antennaor parabolic dish. The main advantage of a parabolic antenna is that it has high directivity. It functions similarly to a search light or flashlight reflector to direct the radio waves in a narrow beam, or receive radio waves from one particular direction only. Parabolic antennas have some of the highest gains, that is, they can produce the narrowest beamwidths, of any antenna type. In order to achieve narrow beam widths, the parabolic reflector must be much larger than thewavelength of the radio waves used,so parabolic antennas are used in the high frequency part of the radio spectrum, at UHF and microwave (SHF) frequencies, at which the wavelengths are small enough that conveniently-sized reflectors can be used.

Parabolic reflector basics

The RF antenna consists of a radiating system that is used to illuminate a reflector that is curved in the form of a paraboloid. A parabolic shape has the property that paths taken from the feed point at the focus to the reflector and then outwards are in parallel, but more importantly the paths taken are all the same length and therefore the outgoing waveform will form a plane wave and the energy taken by all paths will all be in phase.

This shape enables a very accurate beam to be obtained. In this way, the feed system forms the actual radiating section of the antenna, and the reflecting parabolic surface is purely passive.

When looking at parabolic reflector antenna systems there are a number of parameters and terms that are of importance:

•**Focus:** The focus or focal point of the parabolic reflector is the point at which any incoming signals are concentrated. When radiating from this point the signals will be reflected by the reflecting surface and travel in a parallel beam and to provide the required gain and beam width.

•**Vertex:** This is the innermost point at the centre of the parabolic reflector.

•**Focal length**: The focal length of a parabolic antenna is the distance from its focus to its vertex. Read more about the focal length

Design

The operating principle of a parabolic antenna is that a point source of radio waves at the focal point in front of a paraboloidal reflector of conductive material will be reflected into a collimated plane wave beam along the axis of the reflector. Conversely, an incoming plane wave parallel to the axis will be focused to a point at the focal point.

A typical parabolic antenna consists of a metal parabolic reflector with a small feed antenna suspended in front of the reflector at its focus, pointed back toward the reflector. The reflector is a metallic surface formed into a paraboloid of revolution and usually truncated in a circular rim that forms the diameter of the antenna. In a transmitting antenna, radio frequency current from a transmitter is supplied through a transmission line cable to the feed antenna, which converts it into radio waves. The radio waves are emitted back toward the dish by the feed antenna and reflect off the dish into a parallel beam. In a receiving antenna the incoming radio waves bounce off the dish and are focused to a point at the feed antenna, which converts them to electric currents which travel through a transmission line to the radio receiver.

Advantages:

- ***High gain:*** Parabolic reflector antennas are able to provide very high levels of gain. The larger the 'dish' in terms of wavelengths, the higher the gain.
- ***High directivity:*** As with the gain, so too the parabolic reflector or dish antenna is able to provide high levels of directivity. The higher the gain, the narrower the beamwidth. This can be a significant advantage in applications where the power is only required to be directed over a small area. This can prevent it, for example causing interference to other users, and this is important when communicating with satellites because it enables satellites using the same frequency bands to be separated by distance or more particularly by angle at the antenna.

Disadvantages:

Like all forms of antenna, the parabolic reflector has its, limitations and drawbacks:

- ***Requires reflector and drive element:*** the parabolic reflector itself is only part of the antenna. It requires a feed system to be placed at the focus of the parabolic reflector.
- ***Cost :*** The antenna needs to be manufactured with care. A paraboloid is needed to reflect the radio signals which must be made carefully. In addition to this a feed system is also required. This can add cost to the system
- ***Size:*** The antenna is not as small as some types of antenna, although many used for satellite television reception are quite compact.

Parabolic reflector antenna applications

There are many areas in which the parabolic / dish antenna may be used. Its performance enables it to be used almost exclusively in some areas.

- **Direct broadcast television:** Direct broadcast or satellite television has become a major form of distribution for television material. The wide and controllable coverage areas available combined with the much larger bandwidths for more channels available mean that satellite television is very attractive.

Microwave links: T

- **Satellite communications:** Many satellite uplinks, or those for communication satellites require high levels of gain to ensure the optimum signal conditions and that transmitted power from the ground does not affect other satellites in close angular proximity. Again the ideal antenna for most applications is the parabolic reflector antenna.
- **Radio astronomy:** Radio astronomy is an area where very high levels of gain and directivity are required. Accordingly the

- **Aperture** : The aperture of a parabolic reflector is what may be termed its "opening" or the area which it covers. For a circular reflector, this is described by its diameter. It can be likened to the aperture of an optical lens.
- **Gain:** The gain of the parabolic reflector is one of the key parameters and it depends on a number of factors including the diameter of the dish, wavelength and other factors.
- **Feed systems**: The parabolic reflector or dish antenna can be fed in a variety of ways. Axial or front feed, off axis, Cassegrain, and Gregorian are the four main methods.Read more about Parabolic reflector feed types.

Parabolic reflector feed types

There are several different types of parabolic reflector feed systems that can be used. Each has its own characteristics that can be matched to the requirements of the application.

- Focal feed - often also known as axial or front feed system

- Cassegrain feed system
- Gregorian feed system
- Off Axis or offset feed

Focal feed system

The parabolic reflector or dish antenna consists of a radiating element which may be a simple dipole or a waveguide horn antenna. This is placed at the focal point of the parabolic reflecting surface. The energy from the radiating element is arranged so that it illuminates the reflecting surface. Once the energy is reflected it leaves the antenna system in a narrow beam. As a result considerable levels of gain can be achieved.

Achieving this is not always easy because it is dependent upon the radiator that is used. For lower frequencies a dipole element is often employed whereas at higher frequencies a circular waveguide may be used. In fact the circular waveguide provides one of the optimum sources of illumination.

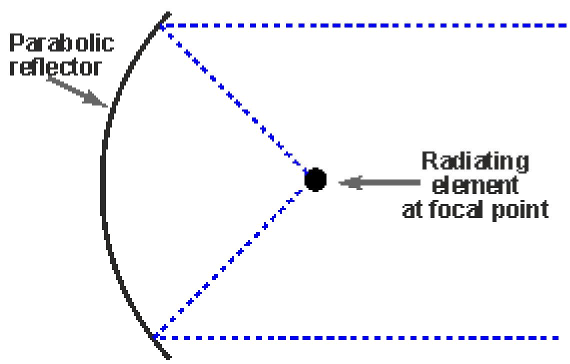


Diagram of a focal feed parabolic reflector antenna

The focal feed system is one of the most widely used feed system for larger parabolic reflector antennas as it is straightforward. The major disadvantage is that the feed and its supports block some of the beam, and this typically limits the aperture efficiency to only about 55 to 60%.

Cassegrain feed system

The Cassegrain feed system, although requiring a second reflecting surface has the advantage that the overall length of the dish antenna between the two reflectors is shorter than the length between the radiating element and the parabolic reflector. This is because there is a reflection in the focusing of the signal which shortens the physical length. This can be an advantage in some systems.

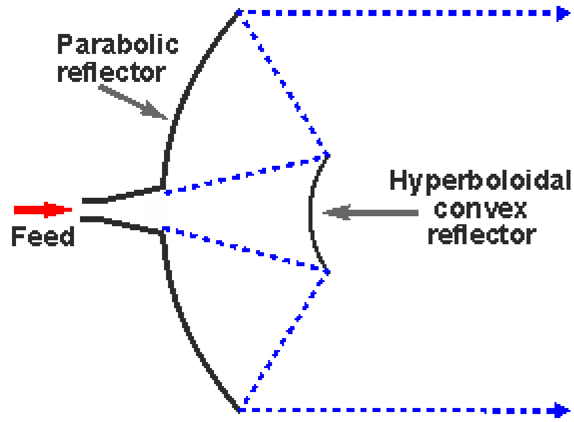


Diagram of a Cassegrain feed parabolic reflector or dish antenna

Typical efficiency levels of 65 to 70% can be achieved using this form of parabolic reflector feed system.The Cassegrain parabolic reflector antenna design and feed system gains its name because the basic concept was adapted from the Cassegrain telescope. This was reflecting telescope which was developed around 1672 and attributed to French priest Laurent Cassegrain.

Gregorian parabolic reflector feed

The Gregorian parabolic reflector feed technique is very similar to the Cassegrain design. The major difference is that except that the secondary reflector is concave or more correctly ellipsoidal in shape.

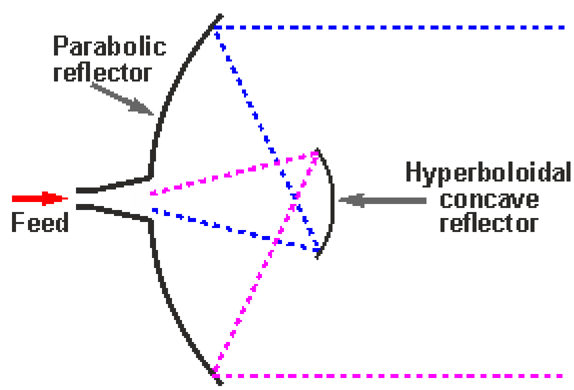


Diagram of a Gregorian feed parabolic reflector or dish antenna

Typical aperture efficiency levels of over 70% can be achieved because the system is able to provide a better illumination of all of the reflector surface.

Off axis or offset parabolic reflector antenna feed

As the name indicates this form of parabolic reflector antenna feed is offset from the centre of the actual antenna dish used.

The reflector used in this type of feed system is an asymmetrical segment of the parabolic shape normally used. In this way the focus and the feed antenna are located to one side of the reflector surface.

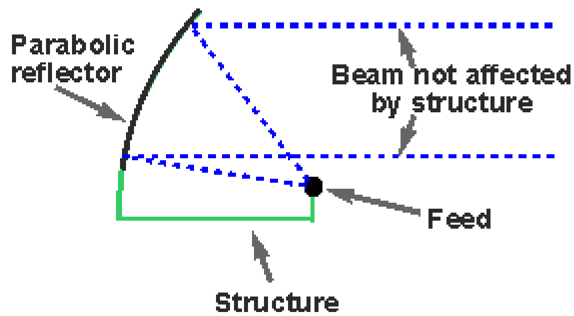


Diagram of an Offset feed parabolic reflector or dish antenna

The advantage of using this approach to the parabolic reflector feed system is to move the feed structure out of the beam path. In this way it does not block the beam.

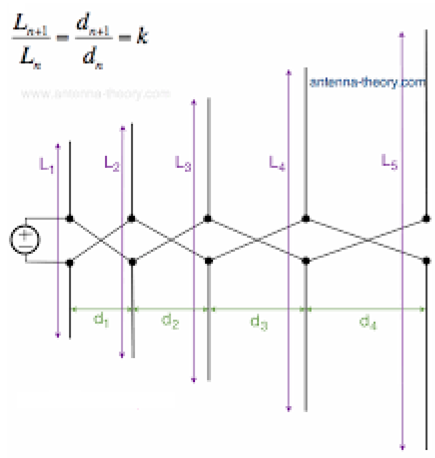
3.9 LOG PERIODIC DIPOLE ARRAY

The log periodic dipole array (LPDA) is one antenna that almost everyone over 40 years old has seen. They were used for years as TV antennas. The chief advantage of an LPDA is that it is frequency-independent. Its input impedance and gain remain more or less constant over its operating bandwidth, which can be very large. Practical designs can have a bandwidth of an octave or more.

Although an LPDA contains a large number of dipole elements, only 2 or 3 are active at any given frequency in the operating range. The electromagnetic fields produced by these active elements add up to produce a unidirectional radiation pattern, in which maximum radiation is off the small end of the array. The radiation in the opposite direction is typically 15 - 20 dB below the maximum. The ratio of maximum forward to minimum rearward radiation is called the Front-to-Back (FB) ratio and is normally measured in dB.

Operation of the Log Periodic Dipole Antenna

The log periodic dipole antenna basically behaves like a Yagi-Uda array over a wide frequency range. As the frequency varies, the active set of elements for the log periodic antenna (those elements which carry the significant current) moves from the long-element end at low frequency to the short-element end at high frequency. The director element current in the Yagi array lags that of the driven element while the reflector element current leads that of the driven element. This current distribution in the Yagi array points the main beam in the direction of the director.



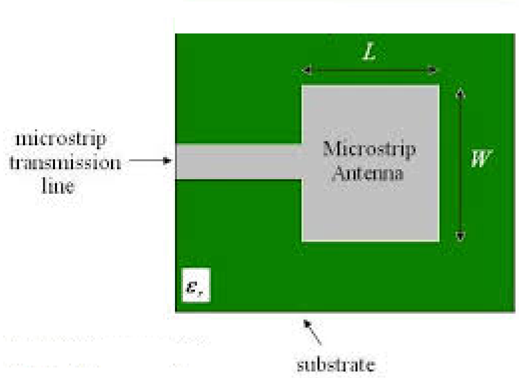
In order to obtain the same phasing in the log periodic antenna with all of the elements in parallel, the source would have to be located on the long-element end of the array.The log periodic dipole array must be driven from the short element end. But this arrangement gives the exact opposite phasing required to point the beam in the direction of the shorter elements. It can be shown that by alternating the connections from element to element, the phasing of the log periodic dipole elements points the beam in the proper direction.

3.10 Microstrip Antennas

- Also called “patch antennas”
- One of the most useful antennas at microwave frequencies (f > 1 GHz).
- It consists of a metal “patch” on top of a grounded dielectric substrate.
- The patch may be in a variety of shapes, but rectangular and circular are the most common

Basic Principles of Operation

The patch acts approximately as a resonant cavity (short circuit walls on top and bottom, open-circuit walls on the sides). In a cavity, only certain modes are allowed to exist, at different resonant frequencies. If the antenna is excited at a resonant frequency, a strong field is set up inside the cavity, and a strong current on the (bottom) surface of the patch. This produces significant radiation (a good antenna).



Advantages :

- Low profile (can even be “conformal”).
- Easy to fabricate (use etching and photolithography).
- Easy to feed (coaxial cable, micro strip line, etc.) .
- Easy to use in an array or incorporate with other microstrip circuit elements.
- Patterns are somewhat hemispherical, with a moderate directivity (about 6-8 dB is typical)

Disadvantages :

- Low bandwidth (but can be improved by a variety of techniques).
- Efficiency may be lower than with other antennas. Efficiency is limited by conductor and dielectric losses, and by surface-wave loss.
- Conductor and dielectric losses become more severe for thinner substrates.
- Surface-wave losses become more severe for thicker substrates (unless air or foam is used).

3.11 Smart antennas for mobile communications

A smart antenna consists of several antenna elements, whose signal is processed adaptively in order to exploit the spatial domain of the mobile radio channel. A smart antenna is an array of elements connected to a digital signal processor. Such a configuration dramatically enhances the capacity of a wireless link through a combination of diversity gain, array gain, and interference suppression. Increased capacity translates to higher data rates for a given number of users or more users for a given data rate per user.

Principle of Smart Antenna System

The smart antenna works as follows. Each antenna element "sees" each propagation path differently, enabling the collection of elements to distinguish individual paths to within a certain resolution. As a consequence, smart antenna transmitters can encode independent streams of data onto different paths or linear combinations of paths, thereby increasing the data rate, or they can encode data redundantly onto paths that fade independently to protect the receiver from catastrophic signal fades, thereby providing diversity gain. A smart antenna receiver can decode the data from a smart antenna transmitter this is the highest-performing configuration or it can simply provide array gain or diversity gain to the desired signals transmitted from conventional transmitters and suppress the interference. No manual placement of antennas is required. The smart antenna electronically adapts to the environment

Types of Smart Antenna Systems:

Terms commonly heard today that embrace various aspects of a smart antenna system technology include intelligent antennas, phased array, SDMA, spatial processing, digital beam forming, adaptive antenna systems, and others. Smart antenna systems are customarily categorized, however, as either switched beam or adaptive array systems. The following are distinctions between the two major categories of smart antennas regarding the choices in transmit strategy:

• Switched beam- a finite number of fixed, predefined patterns or combining strategies (sectors).

• Adaptive array - an infinite number of patterns (scenario-based) that are adjusting in real time.

Switched Beam Antenna:

Switched beam antenna systems form multiple fixed beams with heightened sensitivity in particular directions. These antenna systems detect signal strength, choose from one of several predetermined, fixed beams, and switch from one beam to another as the mobile moves throughout the sector. Instead of shaping the directional antenna pattern with the metallic properties and physical design of a single element (like a sectorized antenna), switched beam systems combine the outputs of multiple antennas in such a way as to form finely sectorized (directional) beams with more spatial selectivity than can be achieved with conventional, single-element approaches in fig 1.

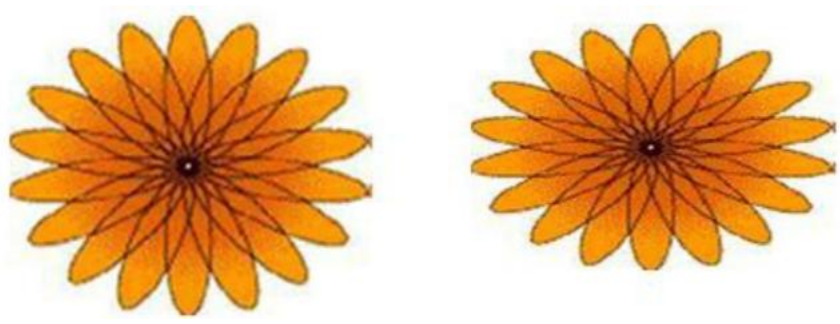


Figure 1 Switched Beam System Coverage Patterns

Adaptive Array Antenna

Adaptive antenna technology represents the most advanced smart antenna approach to date.

Using a variety of new signal-processing algorithms, the adaptive system takes advantage of its ability to effectively locate and track various types of signals to dynamically minimize interference and maximize intended signal reception.

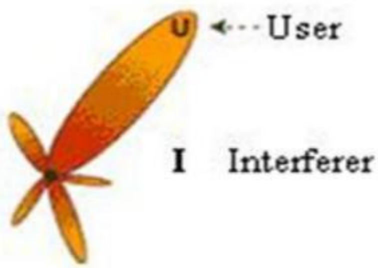


Figure 2: Adaptive array antenna

Both systems attempt to increase gain according to the location of the user; however, only the adaptive system provides optimal gain while simultaneously identifying, tracking, and minimizing interfering signals. Adaptive Array Coverage: A representative depiction of a main lobe extending toward a user with a null directed toward a co-channel interferer as shown in fig (2).

Features and Benefits:

Signal gain:

Inputs from multiple antennas are combined to optimize available power required establish given level of coverage

Better range/coverage:

Focusing the energy sent out into the cell increases base station range and coverage lower power requirements also enable a greater battery life and smaller/lighter handset size.

Interference rejection:

Antenna pattern can be generated toward co-channel interference sources improving the signal to interference ratio of the received signals. Increased capacity: Precise control of signals quality and mitigation of interference combined to frequency re use reduce distance, improving capacity. Spatial diversity: Composite information from the array is used to minimize fading, other undesirable effects of multi path propagation.

Multipath rejection:

It can reduce the effective delay spread of the channel allowing higher bit rates to be supported without the use of an equalizer. Power efficiency: It combines the input s to multiple elements to optimize available processing gain in the downlink(toward user). Reduced expence: lower amplifier costs, power consumption, and higher reliability will resul. Multipath: Multipath is a condition where the transmitted radio signal is reflected

3.12 Antennas for Infrared Detectors

Efforts are being made to develop infrared sensing devices that are cheaper, more responsive and faster. In addition, there is ongoing research in un-cooled infrared detectors. The main drive is to improve performance while reducing the processing cost, thus facilitating wide deployment of infrared detectors for a wide range of applciatons. Infrared detection has been around since the 1800s. By 1880, Langley developed bolometers that are capable of detecting thermal radiation from a cow a quarter of a mile away. There are many applications for this region of the electromagnetic spectrum ranging from imaging to infrared sensing for military applications. Recent advancement of nanofabrication technology has sparked a new vigor in this area again for further improvement in the infrared technology. Hence, this allowed us the opportunity to contribute to this field. Presently, there are many types of infrared detectors that exist such as thermocouples, bolometers, golay cells, pyroelectric detectors, and Schottky diode detectors. Semiconductor based detectors are the most stable and most sensitive among all the infrared detectors.

Fabrication and monolithic integration of infrared antenna and detector based on CMOS-compatible process has been one of the key focuses. Particularly, the compatibility between micro-bolometers and metal-insulator metal diodes with the current state of the art antenna designs has carefully considered, as the ultimate goal to seamlessly integrate infrared antenna with MIM diode. Infrared antenna designs are based on previously optimized configurations to cover the 10-30 THz range. Next, the design and fabrication process is based on standard UV photolithography based processes.

UNIT IV PROPOGATION

4.1. FACTORS AFFECTING THE RADIO WAVES PROPAGATION

There exist a number of factors which affect the propagation of radio waves in actual environment. The most important of these are –

(a) Spherical shape of the earth:- since the radio waves travel in a straight line path in free space, communication between any two points on the surface of earth is limited by the distance to horizon. Therefore, for establishing a communication link beyond the horizon, the radio waves need to undergo a change in the direction of propagation. Several mechanisms can be made use of to effect the change.

(B) The atmosphere:- The earth's atmosphere extends all the way up to about 600 km. The atmosphere is divided into several layers, viz., troposphere, stratosphere, mesosphere, and ionosphere. The propagation of the radio waves near the surface of earth is affected mostly by the troposphere which extends up to height of 8-15 km. Higher up in the atmosphere; it is the ionosphere which interacts with radio waves.

(C) Interaction with the objects on the ground:- The radio waves travelling close to the surface of earth encounter many obstacles such as building, trees, hills, valleys, water bodies, etc. The interaction of such objects with the radio waves is mostly manifested as the phenomena of reflection, refraction, diffraction, and scattering.

4.2. GROUND WAVE PROPAGATION

The ground wave is a wave that is guided along the surface of the earth just as an electromagnetic wave is guided by a wave guide or transmission line. This ground wave propagation takes place around the curvature of the earth in the frequency bands up to 2 MHz This also called as surface wave propagation The ground wave is vertically polarized, as any horizontal component of the E field in contact with the earth is short-circuited by it. In this mode, the wave glides over the surface of the earth and induces charges in the earth which travel with the wave, thus constituting a current, While carrying this current, the earth acts as a leaky capacitor. Hence it can be represented by a resistance or conductance shunted by a capacitive reactance.

As the ground wave passes over the surface of the earth, it is weakened due to the absorption of its energy by the earth. The energy loss is due to the induced current flowing through the earth's resistance and is replenished partly, by the downward diffraction of additional energy, from the portions of the wave in the immediate vicinity of the earth's surface.

Applications

Ground wave propagation is generally used in TV, radio broadcasting etc.

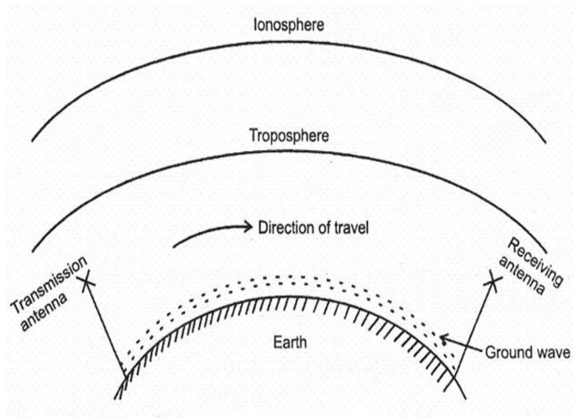


Fig 4.1 Ground wave Propagation

4.3 STRUCTURE OF THE IONOSPHERE

As the medium between the transmitting and receiving antennas plays a significant role, it is essential to study the medium above the earth, through which the radio waves propagate. The various regions above the earth's surface are illustrated in Fig.

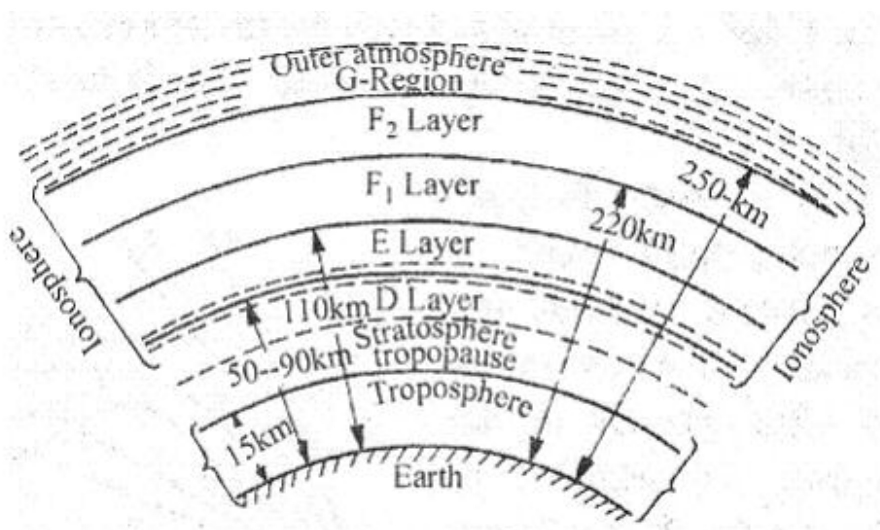


Fig 4.2 Structure of the ionosphere

The portion of the atmosphere, extending up to a height (average of 15 Km) of about 16 to 18 Kms from the earth's surface, at the equator is termed as troposphere or region of change. Tropopause starts at the top of the troposphere and ends at the beginning of or region of calm. Above the stratosphere, the upper stratosphere parts of the earth’s atmosphere absorb large quantities of radiant energy from the sun. This not only heats up the atmosphere, but also produces some ionization in the form of free electrons, positive and negative ions. This part of the atmosphere where the ionization is appreciable, is known as the ionosphere. The most important ionizing agents are ultraviolet UV radiation, ά , β and cosmic rays and meteors. The ionization tends to be stratified due to the differences in the physical properties of the atmosphere at different heights and also because various kinds of radiation are involved.

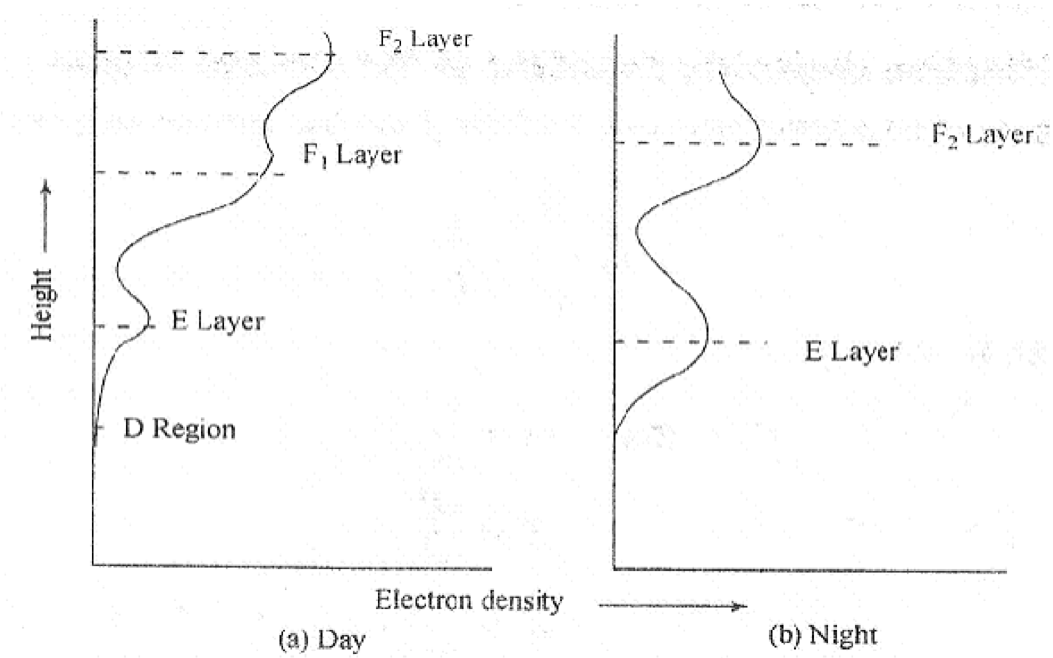


Fig. 4.3 Electron Density Layers

The levels, at which the electron density reaches maximum, are called as layers. The three principal day time maxima are called E, F1, and F2 layers. In addition to these three regular layers, there is a region (below E) responsible for much of the day time attenuations of HF radio waves, called D region. It lies between the heights of 50 and 90 Km.The heights of maximum density of regular layers E and F1are relatively constant at about 110 Km and 220Km respectively. These have little or no diurnal variation, whereas the F2 layer is more variable, with heights in the range of 250 to 350 Km.

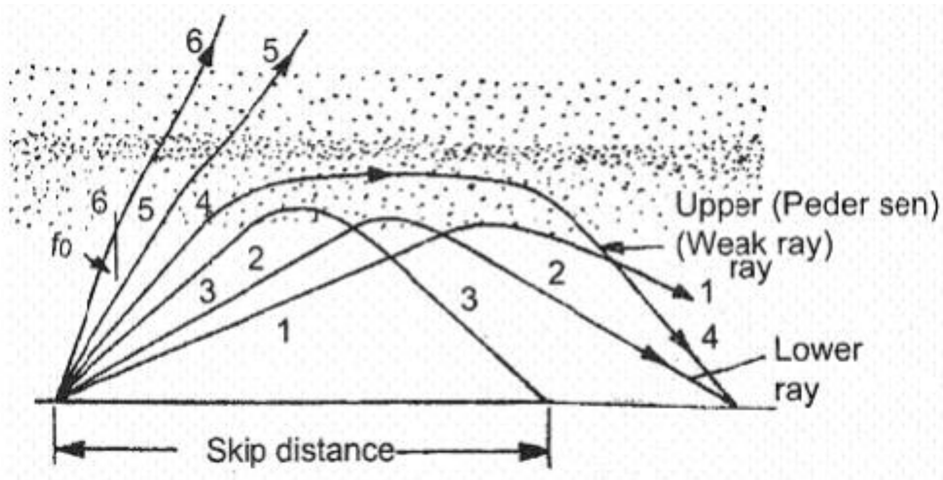


Fig 4.4 Effect of ionosphere on rays

At night F1 and F2 layers combine to form a single night time F2 layer .The E layer is governed closely by the amount of UV light from the sun and at night tends to decay uniformly with time. The D layer ionization is largely absent during night A sporadic E layer is not a thick layer. It is formed without any cause. The ionization is often present in the region, in addition to the regular E ionization. Sporadic E exhibits the characteristics of a very thin layer appearing at a height of about 90 to 130 Kms. Often, it occurs in the form of clouds, varying in size from 1 Km to several 100 Kms across and its occurrence is quite unpredictable. It may be observed both day and night and its cause is still uncertain.

Basically the troposphere is the region atmosphere. It is adjacent to the earth and is located up to about 1 kilometers with the height temperature of this region decreases 6.5▫c per kilometer it is observed that up to upper boundary of the troposphere, temperature may decreases up to 5 in this region the clouds are formed next to the troposphere. troposphere exists. The propagation through the troposphere takes place due to mechanisms such as diffraction normal refraction, abnormal reflection and refraction and troposphere scattering. Let us consider few of them in brief it clear that the radius of curvature depends on the rate of change of the dielectric constant with the height .thus it is observed that the radius of curvature varies from hour to hour, day to day and season.eventhough there is such a variation in radius of curvature, for the practical calculation, average value of four times the radius of earth is used.

In the analysis of propagation problems practically ray in the straight path is considered. then to compensate for the curvature, the effective radius of the earth is selected very large. the actual path of radius a is .the imagined straight line path.Thus when the radius of curvature p equals to four times radius of the earth, then the effective radius of the earth equals to 4/3 times actual radius of the earth.

Abnormal Reflection and Refraction

As discussed previously, the refraction of waves take place in the troposphere even under normal conditions. along with this, there are chances of further refractions and reflections which are due to the abrupt variation in the refractive index and its gradient. The important point here is that where the permittivity of the medium changes abruptly, the reflections are resulted can produce usable signal beyond the range compared with only ground wave propagation

Reflection

Reflection occurs when an electromagnetic wave falls on an object, which has very large dimensions as compared to the wavelength of the propagating wave. For example, such objects can be the earth, buildings and walls. When a radio wave falls on another medium having different electrical properties, a part of it is transmitter into it, while some energy is reflected back. Let us see some special cases. If the medium on which the e.m. wave is incident is a dielectric, some energy is reflected back and some energy is transmitted. If the medium is a perfect conductor, all energy is reflected back to the first medium. The amount of energy that is reflected back depends on the polarization of the e.m. wave. particular case of interest arises in parallel polarization, when no reflection occurs in the medium of origin. This would occur, when the incident angle would be such that the reflection coefficient is equal to zero. This angle is the Brewster’s angle. By applying laws of electro-magnetic, it is found to be sin(θB) =_1_1 + _2Further, considering perfect conductors, the electric field inside the conductor always zero. Hence all energy is reflected back. Boundary conditions require that θi = θr and Ei = Er for vertical polarization, and Ei = −Er for horizontal polarization.

Mechanism of ionospheric propagation _reflection &reflection

Ionospheric propagation involves reflection of wave by the ionosphere .In actual mechanism, refraction takes place as shown in fig. 4.5

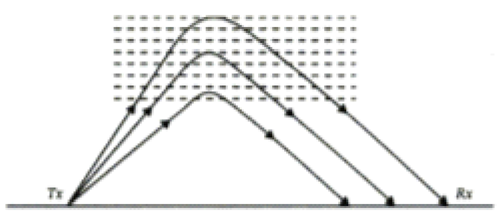


Fig 4.5 Ionospheric reflection and refraction

At ionization density increases at an angle for the incoming wave, the refraction index of the layer decreases and the dielectric constant also decreases; hence the incident wave is gradually bent away from the normal. Sufficient, the refracted ray finally becomes parallel to the layer .then it bends downwards and returns from the ionized at an equal to the angle of incidence. Although, some absorption takes place depending on the frequency the wave is returned by the ionosphere to the receiver on earth. As a result ionosphere propagation takes place through reflection and refraction of EM waves in the ionosphere.

The bending of wave produced by the ionosphere follows the optical laws the direction of propagating wave at a point in the ionosphere is given by Snell's law that is,

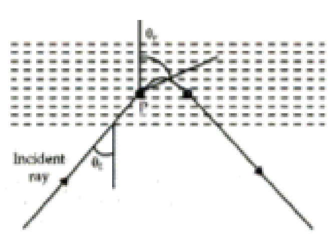


Fig 4.6 Refraction of EM waves in ionosphere

The skip distance defined as the shortest Distance from the transmitter that is covered by a fixed Frequency (>fc).When the angle of incident is large ray 1 returns to ground at a long distance from the transmitter. if the angle is reduced.ray 2 returns to a point closer to the transmitter .so there is always possibility that short distance may not be covered by sky-wave propagation under certain conditions. The Transmission path is limited by the skip distance and curvature of the earth.

4.4 SKY WAVE PROPAGATION

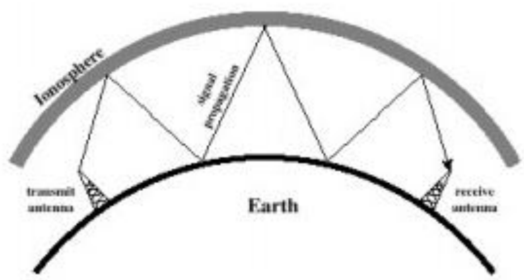


Fig 4.7 Sky wave propogation

When the critical angle is less than 90 degree there will always be a region around the transmitting site where the ion spherically propagated signal cannot be heard, or is heard weakly. This area lies between the outer limit of the ground-wave range and the inner edge of energy return from the ionosphere. It is called the skip zone, and the distance between the originating site and the beginning of the ionosphere return is called the skip distance. This terminology should not to be confused with ham jargon such as "the skip is in," referring to the fact that a band is open for sky-wave propagation. The signal may often be heard to some extent within the skip zone, through various forms of scattering, but it will ordinarily be marginal in strength. When the skip distance is short, both ground wave and sky-wave signals may be received near the transmitter. In such instances the sky wave frequently is stronger than the ground wave, even as close as a few miles from the transmitter. The ionosphere is an efficient communication medium under favorable conditions. Comparatively, the ground wave is not.

Sky wave propagation is practically important at frequencies between 2 to 30 MHz Here the electromagnetic waves reach the receiving point after reflection from an atmospheric layer known as ionosphere. Hence, sky wave propagation is also known as 'ionospheric wave propagation'. It can provide communication over long distances.Hence, it is also known as point-to-point propagation or point-to-point communication.

Virtual heights: The virtual height (h) has the great advantage of being easily measured, and it is very useful in transmission path calculations. For fiat earth approximation and assuming that ionosphere conditions are symmetrical for incident and refracted waves,

The transmission path distance,

$$TR=2h/\tan \beta$$

Where β=Angle of elevation

h =Virtual height

Critical frequency: When the refractive index, n has decreased to the point where n = sin φi the angle of refraction φ will be 90° and wave will be travelling horizontally. The higher point reached by the wave is free. If the electron density at some level in a layer is sufficient great to satisfy the above condition. Then the wave will be returned to earth from that level. If maximum electron density in a layer is less than n', the wave will penetrate the layer (Though it may be reflected back from a higher layer for which N is greater). The largest electron density required for reflection occurs when the angle of incident φi is zero, i.e., for vertical incidence. For any given layer the highest frequency that will be reflected back for vertical incidence.

The characteristics of the ionosphere layers are usually described in terms of their virtual heights and critical frequencies, as these quantities can be readily measured. The virtual height is the height that would be reached by a short pulse of energy showing the same time delay as the actual pulse reflected from the layer travelling with the speed of light. The virtual height is always greater than the true height of reflection, because the interchange of energy taking place between the wave and electrons of the ionosphere causes the velocity of propagation to be reduced. The extent of this difference is influenced, by the electron distributions in the regions below the level of reflection. It is usually very small, but on occasions may be as large as 100 Kms or so.

The critical frequency is the highest frequency that is returned by a layer at vertical incidence. For regular layers,

fc =√max electron density in the layer

i.e. fc =√Ne

The critical frequencies of the E and F1 layers primarily depend on the zenith angle of the sun. It, therefore, follows a regular diurnal cycle, being maximum at noon and tapering off an either side. The fc of the F2 layer shows much larger seasonal variation and also changes more from day to day. It can be seen that the critical frequencies of the regular layers decrease greatly during night as a result of recombination in the absence of solar radiation. But the fc of sporadic E shows regular variation throughout the day and night suggesting that sporadic E is affected strongly by factors other than solar radiation. There is a long term variation in all ionosphere characteristics closely associated with the 11 year sunspot cycle. From the minimum to maximum of the cycle, fc of F2 layer varies from about 6 to 11 MHz (ratio of 1:1.8), fc of E layer varies from 3.1 to 3.8 MHz (a ratio of mere 1 to 1.2). Long term predictions of ionosphere characteristics are based on predictions of the sunspot number.

Maximum usable Frequency: Although the critical frequency for any layer represents the highest frequency that will be reflected back from that layer at vertical incidence, it is not the highest frequency that can be reflected from the layer. The highest frequency that can be reflected depends also upon the angle of incidence, and hence, for a given layer height, upon the distance between the transmitting and receiving points. The maximum, frequency that can be reflected back for a given distance of transmission is called the maximum usable frequency (MUF) for that distance. It is seen that the MUF is related to the critical frequency and the angle of incidence by the simple expression

MUF =f cr secφi

The MUF for a layer is greater than the critical frequency by the factor sec φi the largest angle of incidence φi that can be obtained in F-layer reflection is of the order of 74°. This occurs for a ray that leaves the earth at the grazing angle Where φimax =sin - 1 (r/r+h)

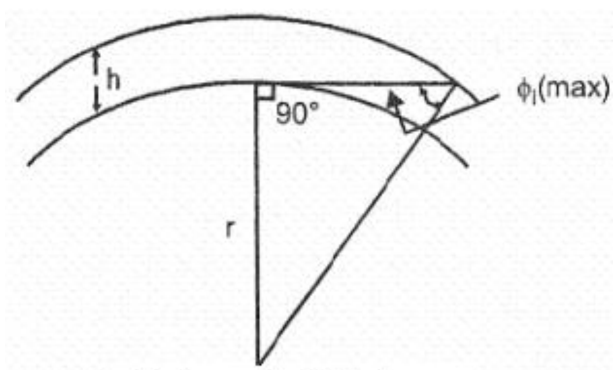


Fig. 4.8 Geometry of MUF

Geometry of MUF

The MUF at this limiting angle is related to the critical frequency of the layer by MUFmax = f cr/ cos 74 o =3.6 f cr

Disadvantage

Sky wave propagation suffers, from fading due to reflections from earth surface; fading can be reduced with the help of diversity reception.

Applications

1. It can provide communication over long distances.
2. Global communication is possible.

4.5 FADING

The term fading, or, small-scale fading, means rapid fluctuations of the amplitudes, phases, or multipath delays of a radio signal over a short period or short travel distance. This might be so severe that large scale radio propagation loss effects might be ignored.

Factors Influencing Fading

The following physical factors influence small-scale fading in the radio propagation channel:

(1) Multipath propagation – Multipath is the propagation phenomenon that results in radio signals reaching the receiving antenna by two or more paths. The effects of multipath include constructive and destructive interference, and phase shifting of the signal.

(2) Speed of the mobile – The relative motion between the base station and the mobile results in random frequency modulation due to different doppler shifts on each of the multipath components.

(3) Speed of surrounding objects – If objects in the radio channel are in motion, they induce a time varying Doppler shift on multipath components. If the surrounding objects move at a greater rate than the mobile, then this effect dominates fading.

(4) Transmission Bandwidth of the signal – If the transmitted radio signal bandwidth is greater than the "bandwidth" of the multipath channel (quantified by coherence bandwidth), the received signal will be distorted.

Selective Fading

This type of fading produces serious distortion in modulated signal. Selective fading is important at higher frequencies. Selective fading generally occurs in amplitude modulated signals. SSB signals become less distorted compared to the AM signals due to selective fading.

Interference Fading

Interference fading occurs due to the variation in different layers of ionosphere region. This type of fading is very serious and produces interference between the upper and lower rays of sky wave propagation. Interference fading can be reduced with the help of frequency and space diversity reception.

4.6 DIVERSITY RECEPTION

- To reduce fading effects, diversity reception techniques are used. Diversity means the provision of two or more uncorrelated (independent) fading paths from transmitter to receiver
- These uncorrelated signals are combined in a special way, exploiting the fact that it is unlikely that all the paths are poor at the same time. The probability of outage is thus reduced.
- Uncorrelated paths are created using polarization, space, frequency, and time diversity

Frequency Diversity

- Different frequencies mean different wavelengths. The hope when using frequency diversity is that the same physical multipath routes will not produce simultaneous deep fades at two separate wavelengths.

Space Diversity

Deep multipath fade have unlucky occurrence when the receiving antenna is in exactly in the 'wrong' place. One method of reducing the likelihood of multipath fading is by using two receive antennas and using a switch to select the better signal. If these are physically separated then the probability of a deep fade occurring simultaneously at both of these antennas is significantly reduced. ·

Angle Diversity: In this case the receiving antennas are co-located but have different principal directions. ·

Polarization Diversity: This involves simultaneously transmitting and receiving on two orthogonal polarizations (e.g. horizontal and vertical). The hope is that one polarization will be less severely affected when the other experiences a deep fade. ·

Time Diversity: This will transmit the desired signal in different periods of time. The intervals between transmissions of the same symbol should be at least the coherence time so that different copies of the same symbol undergo independent fading.

4.7 FREE SPACE RADIO WAVE PROPAGATION

There are two basic ways of transmitting an electro-magnetic (EM) signal, through a guided medium or through an unguided medium. Guided mediums such as coaxial cables and fiber optic cables are far less hostile toward the information carrying EM signal than the wireless or the unguided medium. It presents challenges and conditions which are unique for this kind of transmissions. A signal, as it travels through the wireless channel, undergoes many kinds of propagation effects such as reflection, diffraction and scattering, due to the presence of buildings, mountains and other such obstructions. Reflection occurs when the EM waves impinge on objects which are much greater than the wavelength of the traveling wave. Diffraction is a phenomena occurring when the wave interacts with a surface having sharp irregularities. Scattering occurs when the medium through the wave is traveling contains objects which are much smaller than the wavelength of the EM wave. These varied phenomena's lead to large scale and small scale propagation losses. Due to the inherent randomness associated with such channels they are best described with the help of statistical models. Models which predict the mean signal strength for arbitrary transmitter receiver distances are termed as large scale propagation models. These are termed so because they predict the average signal strength for large Tx-Rx separations, typically for hundreds of kilometers.

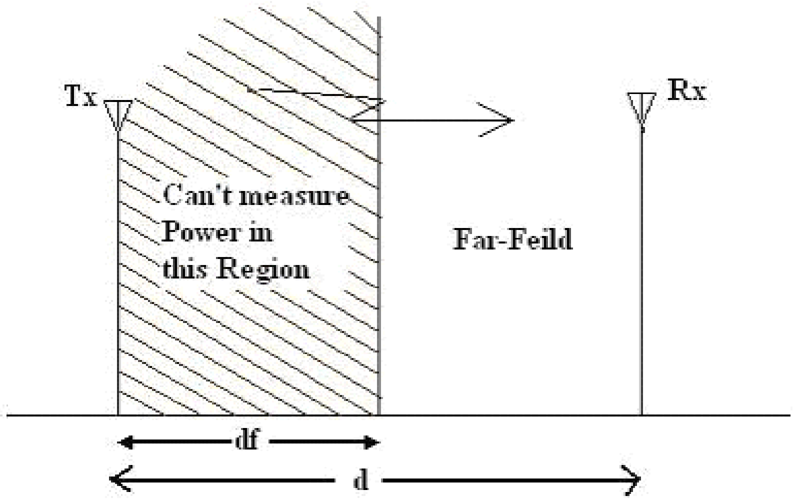


Fig 4.9 Space wave propagation

4.8 CONSIDERATION IN SPACE WAVE PROPAGATION

The space wave field strength is affected by the following

1. Curvature of the earth
2. Earth's imperfections and roughness
3. Hills, tall buildings and other obstacles
4. Height above the earth.
5. Transition between ground and space wave
6. Polarization

4.9 ATMOSPHERIC EFFECTS IN SPACE WAVE PROPAGATION

There is a significant effect of the atmosphere through which the space wave travels on the propagation. this is basically because of presence of gas molecules particularly of a water vapor. Water vapor has a high dielectric constant and its presence causes the air of the troposphere to have a dielectric constant and its presence causes the air of troposphere to have a dielectric constant slightly greater than unity. the distribution of water vapor is not uniform through out of the air and along with it the density of the air varies with height .As a consequence of the dielectric constant and in turn the refractive index of air also depend upon the height it is in general observed to be deceasing with increasing height gives rise to a variety of phenomena like reflection,refraction ,scattering, duct transmission and fading of signals. The behavior of the space wave under the different conditions can be better studied by changing the co - ordinates in such a manner that the particular ray path of interest is a straight line instead of a curve. for this , the radius of curvature of the earth is required to be simultaneously readjusted to preserve the correct relative relation.

4.10 DUCT PROPAGATION

Duct propagation is phenomenon of propagation making use of the atmospheric duct region. The duct region exits between two levels where the variation of modified refractive index with height is minimum . It is also said to exist between a level .where the variation of modified refractive index and a surface bounding the atmosphere. The higher frequencies or microwaves are continuously reflected in the duct and reflected by the ground.So that they propagate around the curvature for beyond the line of sight. This special refraction electromagnetic waves is called super refraction and the process is called duct propagation. Duct propagation is also known as super refraction. Consider the figure,

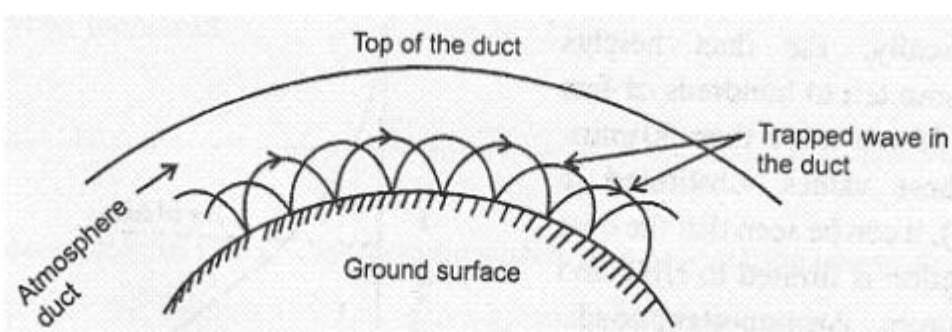


Fig. 4.10 Duct Propagation

Here, two boundary surfaces between layers of air form a duct or a sort of wave guide which guides the electromagnetic waves between the walls. Temperature inversion is one of the important factor for the formation of duct. For proper value of curvature, the refractive index (n) must be replaced by a modified refractive index (N). N = n + (h/r) The term modified index of refractive modules (m) is related to N as N = n + (h/r) (N-1) = n-1+ h/r (N-1) x 106 = [n-1+ h/r] x 106 m = (N-1) x 106 = [n-1+ h/r] x 106 Where, n = Refractive index h = Height above ground r = Radius of the earth = 6370 km Duct can be used at VHF, UHF and microwave frequencies. Because, these waves are neither reflected nor propagated along earth surface. So, the only possible way to transmit such signal is to utilize the phenomenon of refraction in the troposphere.

UNIT-V MEASUREMENT

5.1. Impedance Measurement

Impedance Measurement are done according to frequency involved. For Radio frequencies below 30MHz(low frequency).Usually impedance Bridge method is employed for frequencies above 1000 MHz(High frequency)"slotted line" measurement is almost invariably used. However between frequencies 30MHz-1000MHz either method can be used depending on the applications, convenience or availability of equipment. Impedance at a pair of electrical terminals is the ratio of current (I) which flows when a voltage V is applied between the terminals. By generalization of Ohm's law

$$Z=\frac{V}{I} \quad (1)$$

Where Z=$R + jX$
R=Resistive components
X=Reactive components
If there is a phase difference θbetween I&V, then

$$\theta=\tan^{-1}\frac{X}{R} \quad (2)$$

The voltage V and current I will be in phase when reactive component X=0.Thus the impedance can be determined by measuring the voltage and current at the terminals.

5.1.1 Impedance Bridge Method for Low Frequency

Wheat stone bridge is used to measure unknown impedance(resistance, inductance or capacitance) by comparison with known impedances. It consists of 4 impedances connected in four arms of the bridge as shown in figure below.

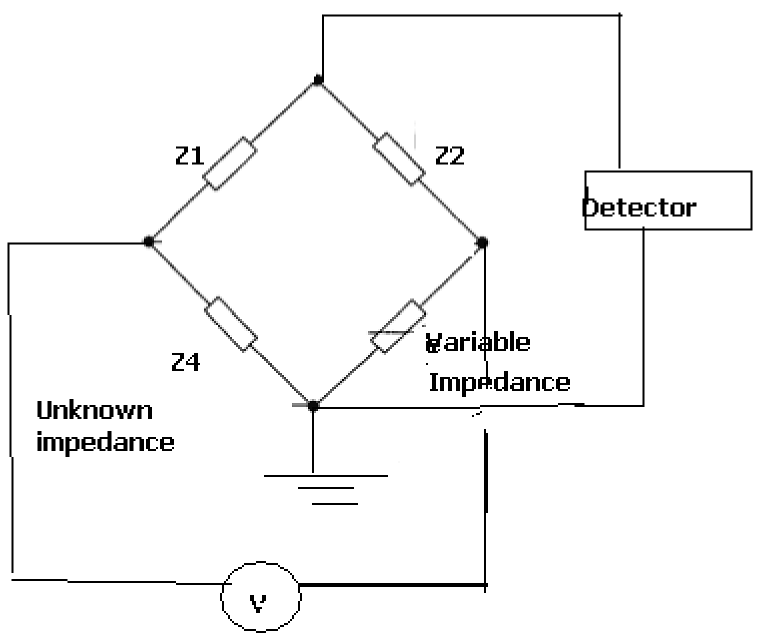


Fig: 5.1.Wheat stone Bridge for impedance measurement

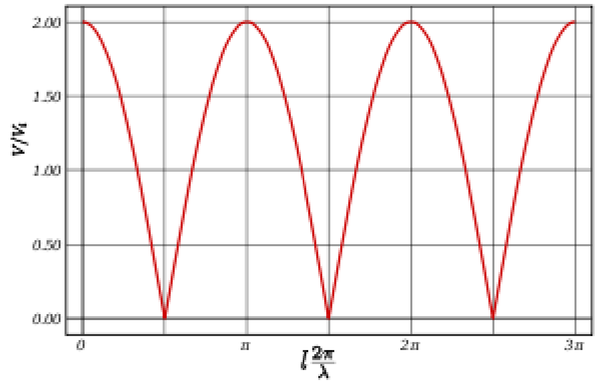


Fig:5. 2.Standing wave along the line

It may be noted however the bridge is balanced not for impedance magnitude but for also a phase balance. Thus writing in polar form we have

$$\frac{Z1\theta1}{Z2\theta2}=\frac{Z4\theta4}{Z3\theta3} \quad \textbf{(3)}$$

Thus there are two balance conditions which must be satisfied simultaneously

$Z_1Z_3=Z_2Z_4$ For magnitude balance

Angle$(\theta_1 + \theta_3)= (\theta_2+\theta_4)$ For phase angle balance

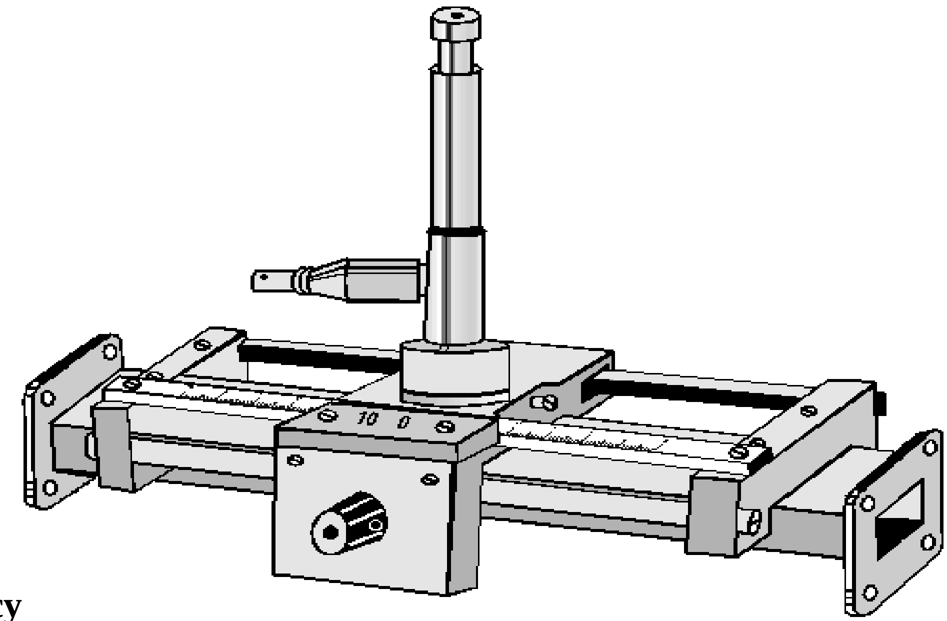
Unknown impedance Z_4 is calculated as

$$Z_4 =Z_3(\frac{Z_1}{Z_2}) \quad \textbf{(4)}$$

For antenna input impedance measurement the antenna input terminals are connected as unknown impedance between point A and D is grounded so it is suitable for a low frequency grounded vertical antenna. For balanced antenna one should see that points A and D of bridge are balanced w.r.t. ground.

The measurements usually are preceded by calibration, the bridge is balanced with unknown impedance terminal short circuited or open circuited. Then the short is removed and unknown impedance is inserted in unknown impedance arm between A and D and the bridge is re-balanced. The unknown impedance is now determined by impedance equation.

5.1.2 Standing Wave Ratio Method or Slotted Line Method for Impedance Measurement at High



Frequency

Fig: 3.Prespective view of slotted line arrangement

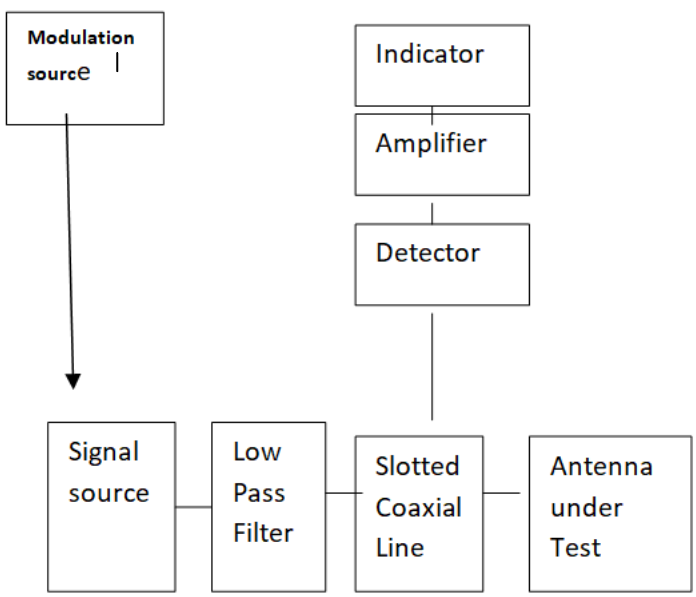


Fig5.4.Block diagram of making slotted line measurement

The Experimental set up for determination of standing wave ratio and hence the input impedance is shown in the above figure.

A Transmission line system terminated in an antenna, if not perfectly matched to this feeding transmission line, the incident and reflected waves and consequently produce standing wave will be set up along the transmission line. A part of transmission line is replaced by an axial slotted line, VSWR or impedance measuring set up. As shown the slotted line arrangement consists of a length of a transmission line with an axial slot, along which moves a travelling carriage carrying probe. The probe project through the slot. A voltage measuring device may be in simplest case, a crystal detector and a micro ammeter. A signal source may be a transmitter or oscillator which is connected to left end and right end is connected to the unknown impedance being, measured. The standing wave pattern is obtained by moving the probe along the carriage and observing the resulting variation in the crystal detector output.Infact this device measures electric field intensity but since it is proportional to the voltage between conductors, therefore standing wave indicator is assumed to be a voltage measuring device. The probe in the slotted coaxial cable is moved and two consecutive points of V_{max} and V_{min} are noted. Their ratio will give the VSWR and hence the input impedance.

The probe is inserted deeply into the axial slot line in order to sample the standing wave pattern. Commercial standing wave detectors, a high or low pass filter is used to avoid harmonics spurious signal sources and unwanted signals between slotted line and signal source as shown in block diagram. The modulation source is for modulating signal source with a square or pulse.

5.2. Radiation pattern Measurement

Radiation pattern of transmitting antenna is described as the field strength or power density at a fixed distance from the antennas as a function of direction. The Radiation pattern of an antenna is a three dimensional figure and it needs measurements of field intensity all over the spatial angles. Hence for radiation pattern of antenna under test the various spatial angles must be specified. The test antenna is assumed to be placed at the origin of spherical coordinate as shown in fig. below

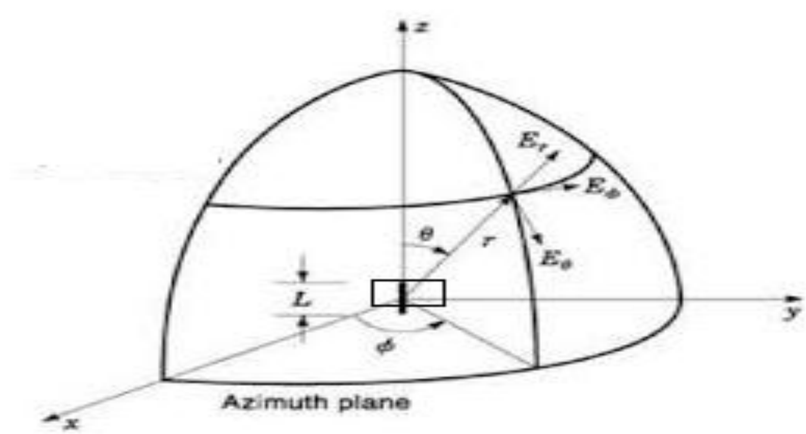


Fig:5.5 Spherical coordinate system for pattern measurement

XY-Plane is horizontal plane and XZ-plane is vertical plane the radiation pattern is accordingly taken either along latitude as a function of Azimuth angle ∅ as a function of

θ depending upon the application and the information needed.
For most antennas if it is generally necessary to take radiation pattern in XY-Plane (Horizontal plane) and XZ-Plane (Vertical Plane).

For horizontal antenna two patterns are sufficient

(i) The $\emptyset$ component of electric field is measured as the function of $\emptyset$ in XY-Plane ($\theta = 90°$). It is represented as $E_{\emptyset}(\theta = 90°, \emptyset)$ and is called as E-Plane Pattern.

(ii)The $\emptyset$ component of the field is measured as the function of θ in the XZ-Plane ($\emptyset = 0°$). It is represented as $E_{\emptyset}(\theta, \emptyset = 0°)$ is called as H-Plane Pattern.

These two patterns bisect the major lobe in mutually perpendicular planes and hence provide enough information's for a number of applications.

Similarly for vertically polarized antennas:

(i) The θ component of electrical field is measured as function of $\emptyset$ in XY-Plane ($\theta = 90°$). It is represented as $E_{\theta}(\theta = 90°, \emptyset)$ and is called as H-Plane Pattern.

(ii)The θ component of electrical field is measured as function of $\emptyset$ in XZ-Plane (($\emptyset = 0°$). It is represented as $E_{\theta}(\theta,, \emptyset = 0°)$ and is called as E-Plane Pattern. For circularly and elliptically polarized antenna measurement of these four patterns would be needed.

5.2.1 Arrangement for Radiation pattern Measurements

There is a transmitting antenna (Primary Antenna) and the antenna under test is secondary antenna a mount for rotating the primary antenna, a detector and an indicator for indicating the relative magnitude of received field shows the arrangement of radiation pattern measurement the equipment may be entirely automatic or point to point plot.

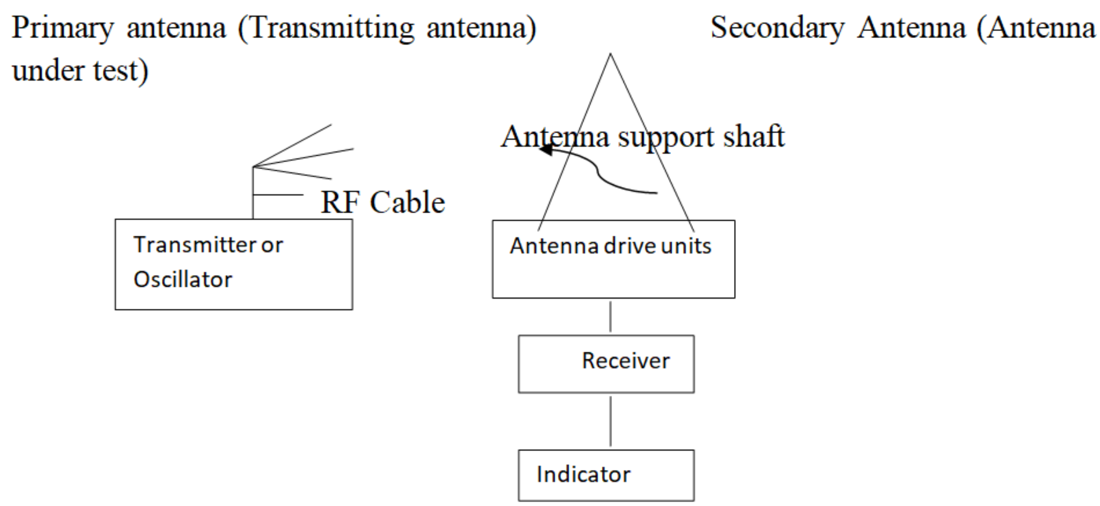


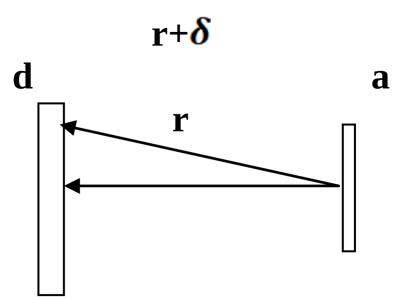
Fig:5.6.Radiation pattern measuring set up

It is usual to operate antenna under test as a receiver, placing it under proper illumination by primary antenna. The primary antenna is fixed and the secondary is rotated on a vertical axis by antenna support shaft. If large number of patterns are to be taken" automatic pattern recorder may also be used which is commercially available.

Two methods (a) Distance requirement and

(b) Uniform illumination requirement

(a) Distance Requirement:



Transmitting antenna aperture   Receiving Antenna Aperture

Fig: 5.7.Phase difference between centre and edges of receiving array for distance requirement

In order to obtain accurate far-field the distance between primary and secondary antenna must be large if the distance between the two antennas is very much small the near field is obtained for accurate far-field pattern measurement the secondary antenna should be illuminated by a plane wave front and plane wave front is obtained only at infinite distance thus the limit specified is that the phase difference between the centre and edge of the antenna under test should not exceed $\frac{\lambda}{16}$ under this condition the distance between primary and secondary antenna should be

$$r \geq \frac{2d^2}{\lambda}$$

d- Maximum linear dimension of either antenna, λ wavelength
r- Distance between TX and RX
The value of r may be calculated in terms of receiving aperture'd' and distance 'r'

$$(r+\delta)^2 = \left(\frac{d}{2}\right)^2 + r^2 \quad (5)$$

$$r = \frac{d^2}{8\delta} \quad (6)$$

Uniform illumination requirement

The other requirement for an accurate field pattern is that primary antenna (transmitting) should produce a plane wave of uniform amplitude and phase over the distance at least equal to 'r'. The interference between direct rays and indirect rays should be avoided as far as possible. Besides the reflections from surrounding objects like buildings, trees, etc., should be avoided. Test should be conducted in open plane area and antennas should be directional , installed on higher towers or top of high buildings.

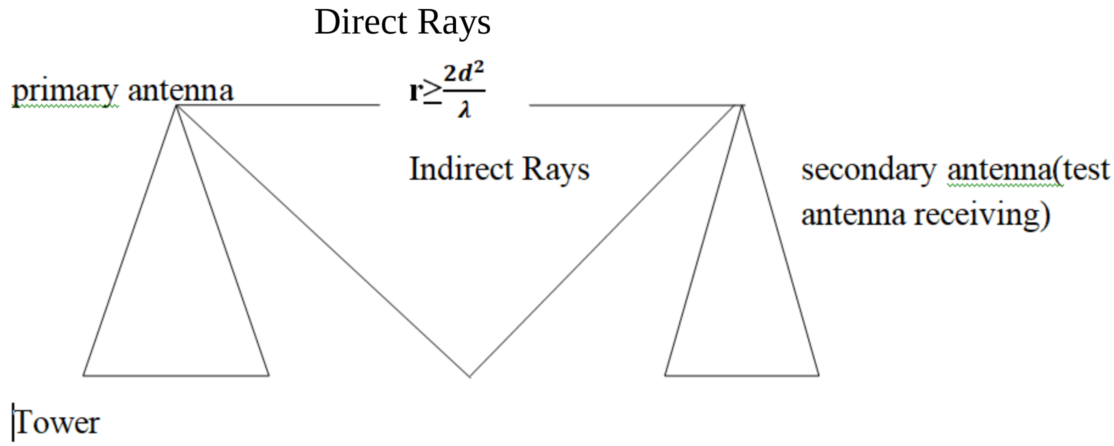


Fig:5. 8.Experimental set up for antenna test

5.3. Measurement of gain

Gain = Maximum radiation intensity (subject or test antenna/maximum radiation intensity (reference antenna)
For the same input power of both antennas
Directivity =Maximum radiation intensity/average radiation intensity
Thus again compares the actual antenna with any reference antenna while the directivity is concerned only with a hypothetical isotropic loss less antenna.
Gain of an antenna over an isotropic loss less antenna is given by

$$G0 = \alpha D \quad (7)$$

G0- Gain with respect to isotropic antenna
D- Directivity α effectiveness ratio

5.3.1 Measurement of gain by direct comparison method

Gain is a comparison of two antennas and hence gain measurement by comparison is done. At high frequencies the comparison method is one which is commonly used, gain is done by comparing the signal strength transmitted or received with the unknown gain antenna and a standard gain antenna. A standard gain antenna whose gain is accurately known so that it can be used in measurement of other antenna. Electromagnetic horn antenna at microwave frequencies is mostly used as standard gain antenna. The secondary antenna may be an arbitrary transmitting antenna and it is not necessary to know its gain. In place of primary antenna there will be two antennas one the subject antenna under test and the other antenna at a considerable distance so that the coupling or interaction between two antennas can be avoided the experimental setup is a pattern measurement which is shown in fig. below.,

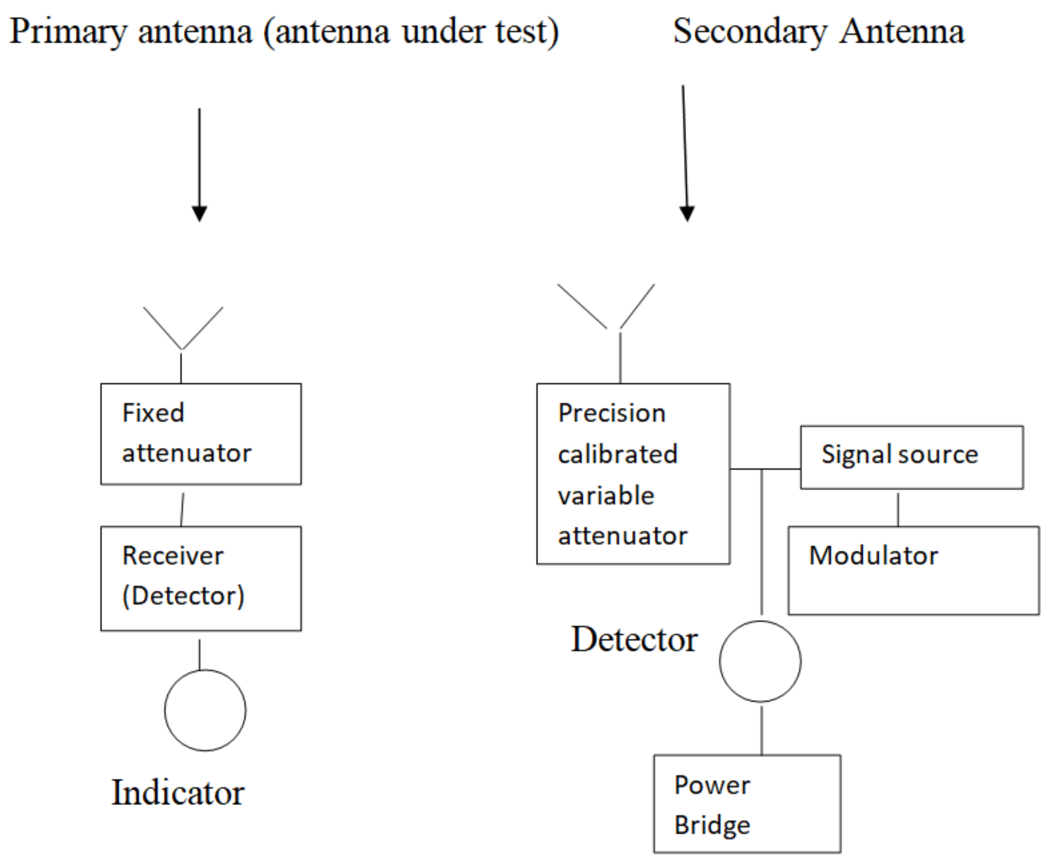


Fig:5.9 Set up for Gain measurement by comparison method

The distance between the primary and secondary antenna should be $\geq \frac{2d^2}{\lambda}$ and reflection between them should be minimized to the extent possible. The following procedure is adopted for gain measurement

(i) At first the standard antenna is connected to the receiver with the help of switch(S) and the antenna is named at secondary antenna in the direction of maximum signal intensity. The input to the transmitting antenna (secondary antenna) is adjusted to a convenient level and corresponding reading at the receiver (primary antenna circuit)is recorded. The attenuator dial setting and the power bridge reading are also recorded. Say it is W_1 and P_1 respectively

(ii) Now connect the subject antenna whose gain is to be measured in place of standard gain antenna the attenuator dial is adjusted such that receiver indicates the same previous reading with standard gain antenna. Let the attenuator dial setting be W_2 and P_2 two cases arise

Case 1:- When P_1=P_2

If $P_1=P_2$ then no correction need to be applied and the gain of the subject antenna under measurement W.r.to standard gain antenna is given by

$$\text{Power gain} = W_2/W_1 \qquad (8)$$

W_1 and W_2 are the relative power levels

$$\text{Log } G_p = \log W_2 - \log W_1$$

$$G_p(db) = W_2(db) - W_1(db)$$

Case 2: When $P_1 \neq P_2$

$$P_1/P_2 = P \text{ (Say)}$$

$$10\log P_1/P_2 = P(db)$$

$$G = G_p * P$$

$$G(db) = G_p(db) + P(db)$$

$$G_p(db) = W_2(db) - W_1(db) + P(db) \qquad (9)$$

5.4. Ionospheric measurements – Vertical incidence measurements of the ionosphere:

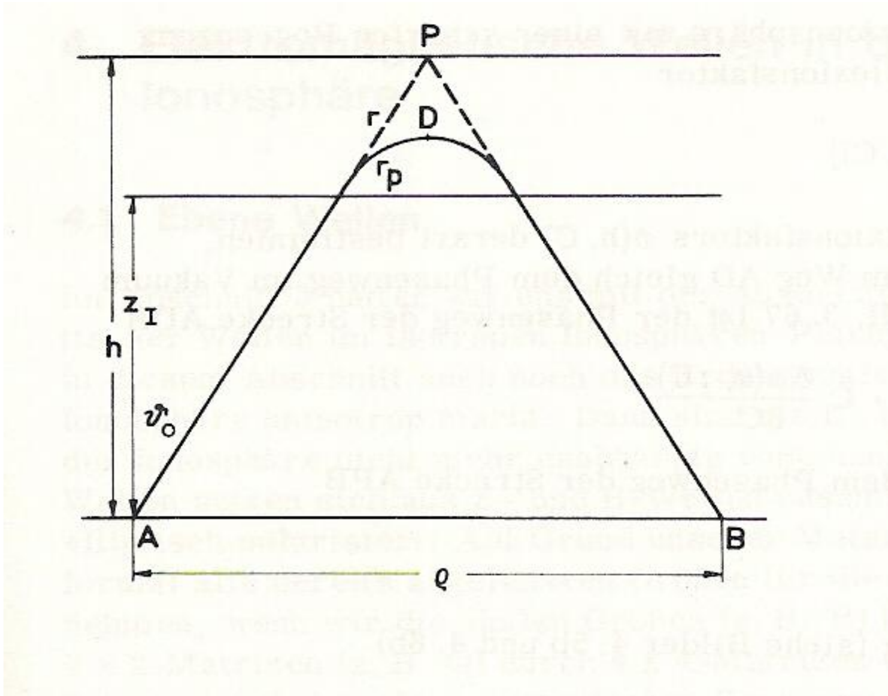


Fig: 10 Vertical incidence measurements of the ionosphere

Waves with frequencies smaller than fc are reflected within the ionospheric D-, E-, and F-layers. fc is of the order of 8–15 MHz during day time conditions. For oblique incidence, the critical frequency becomes larger. Very low frequencies (VLF: 3–30 kHz), and extremely low frequencies (ELF: <3 kHz) are reflected at the ionospheric D- and lower E-layer. An exception is whistler propagation of lightning signals along the geomagnetic field lines

The wavelengths of VLF waves (10–100 km) are already comparable with the height of the ionospheric D-layer (about 70 km during the day, and 90 km during the night). Therefore, ray theory is only applicable for propagation over short distances, while mode theory must be used for larger distances. The region between Earth's surface and the ionospheric D-layer behaves thus like a waveguide for VLF- and ELF-waves.

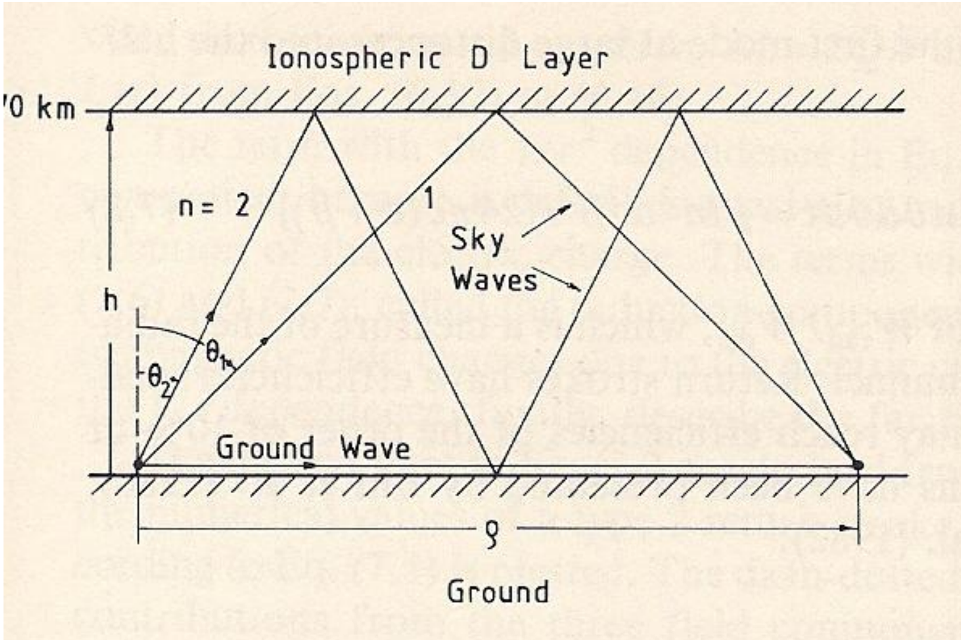


Fig: 11Ionosheric Measurement

In the VLF range, the transfer function is the sum of a ground wave which arrives directly at the receiver and multihop sky waves reflected at the ionospheric D-layer (Figure 10).

For the real Earth's surface, the ground wave becomes dissipated and depends of the orography along the ray path. For VLF waves at shorter distances, this effect is, however, of minor importance, and the reflection factor of the Earth is $R_e = 1$, in a first approximation.

At shorter distances, only the first hop sky wave is of importance. The D-layer can be simulated by a magnetic wall ($R_i = -1$) with a fixed boundary at a virtual height h, which means a phase jump of 180° at the reflection point.[2][5] In reality, the electron density of the D-layer increases with altitude, and the wave is bounded as shown in Figure 11.

5.5. Antenna Noise Temperature and System Signal-to Noise Ratio

Antenna temperature :

The performance of a telecommunication system depends very much on the signal-to-noise ratio (SNR) at the receiver's input. The electronic circuitry of the receiver (amplifiers, mixers, etc.) has its own contribution to the noise generation. However, the antenna itself is a significant source of noise. The antenna noise can be divided into two types of noise according to its physical source: - noise due to the loss resistance of the antenna itself; and - noise, which the antenna picks up from the surrounding environment. Any object whose temperature is above the absolute zero radiates EM energy. Thus, each antenna is surrounded by noise sources, which create noise power at the antenna terminals. Here, we will not be concerned with technological sources of noise, which are a subject of the electromagnetic interference science. We are also not concerned with intentional sources of electromagnetic interference. We are concerned with natural sources of EM noise, such as sky noise and ground noise. The concept of antenna temperature is not only associated with the EM noise. The relation between the object's temperature and the power it can create at the antenna terminals is used in passive remote sensing (radiometry). A radiometer can create temperature images of objects. Typically, the remote object's temperature is measured by comparison with the noise due to background sources and the receiver itself.

System signal-to-noise ratio (SNR)

Signal-to-noise ratio (abbreviated SNR or S/N) is a measure used in science and engineering that compares the level of a desired signal to the level of background noise. It is defined as the ratio of signal power to the noise power, often expressed in decibels. A ratio higher than 1:1 (greater than 0 dB) indicates more signal than noise. While SNR is commonly quoted for electrical signals, it can be applied to any form of signal (such as isotope levels in an ice core or biochemical between cells).

Signal-to-noise ratio is defined as the ratio of the power of a signal (meaningful information) and the power of background noise (unwanted signal)

$$\mathbf{SNR} = \frac{P_{signal}}{p_{noise}} \qquad (10)$$

A link budget is accounting of all of the gains and losses from the transmitter, through the medium (free space, cable, waveguide, fiber, etc.) to the receiver in telecommunication system. It accounts for the attenuation of the transmitted signal due to propagation, as well as the antenna gains, feed line and miscellaneous losses. Randomly varying channel gains such as fading are taken into account by adding some margin depending on the anticipated severity of its effects. The amount of margin required can be reduced by the use of mitigating techniques such as antenna diversity or frequency hopping.

A simple link budget equation looks like this:

Received Power (dB) = Transmitted Power (dB) + Gains (dB) − Losses (dB)

A link budget equation including all these effects, expressed logarithmically, might be

$P_{RX}=P_{TX}+G_{TX}-L_{TX}-L_{FS}-L_M+G_{RX}-L_{RX}$ which is given by

P_{RX}= received power (dBm)

P_{Tx}= transmitter output power (dBm)

G_{TX} = transmitter antenna gain (dBi)

L_{TX} = transmitter losses (coax, connectors...) (dB)

L_{Fs} = path loss, usually free space loss (dB)

L_m = miscellaneous losses (fading margin, body loss, polarization mismatch, other losses...) (dB)

G_{RX} = receiver antenna gain (dBi)

L_{RX} = receiver losses (coax, connectors...) (dB)