Lecture Notes

On

ANTENNA & WAVE PROPAGATION

(15A04501)

B.TECH ECE III YEAR I SEMESTER

(JNTUA-R15)

by

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B. Tech III-ISem. (ECE)

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15A04501 ANTENNAS & WAVE PROPAGATION

Course Objectives:

- Fundamentals of electromagnetic radiation: Maxwell's equations, potential functions, wave equation, retarded potential, short current element, near and far fields, Poynting's theorem.
- Design of antenna arrays: principle of pattern multiplication, broadside and end fire arrays, array synthesis, coupling effects and mutual impedance, parasitic elements, Yagi-Uda antenna.

Course Outcomes:

Upon successful completion of the course, students will be able to:

- Approximate parametric equations for the calculation in the farfield region.
- Write parametric integral expressions for a given current source.
- Calculate electromagnetic fields for a given vector potential.
- Discover pattern multiplication principle for array antennas.

UNIT - I

Antenna Basics & Dipole antennas: Introduction, Basic antenna parameters- patterns, Area, Radiation Intensity, Beam Efficiency, Directivity-Gain-Resolution, Antenna Apertures, Effective height, Fields from oscillating dipole, Field Zones, Shape- Impedance considerations, Polarization – Linear, Elliptical, & Circular polarizations, Antenna temperature, Antenna impedance, Front–to-back ratio, Antenna theorems,

Radiation – Basic Maxwell"s equations, Retarded potential-Helmholtz Theorem, Radiation from Small Electric Dipole, Quarter wave Monopole and Half wave Dipole – Current Distributions, Field Components, Radiated power, Radiation Resistance, Beam width, Natural current distributions, far fields and patterns of Thin Linear Center-fed Antennas of different lengths, Illustrative problems.

UNIT- II

VHF, UHF and Microwave Antennas - I: Loop Antennas - Introduction, Small Loop, Comparison of far fields of small loop and short dipole, Radiation Resistances and Directives of small and large loops (Qualitative Treatment), Arrays with Parasitic Elements - Yagi - Uda Arrays, Folded Dipoles & their characteristics. Helical Antennas, Helical Geometry, Helix modes, Practical Design considerations for Monofilar Helical Antenna in Axial and Normal Modes. Horn Antennas- Types, Fermat"s Principle, Optimum Horns, Design considerations of Pyramidal Horns, Illustrative Problems.

UNIT - III

VHF, UHF and Microwave Antennas - II: Micro strip Antennas- Introduction, features, advantages and limitations, Rectangular patch antennas-Geometry and parameters, characteristics of Micro strip antennas, Impact of different parameters on characteristics, reflector antennas - Introduction, Flat sheet and corner reflectors, parabola reflectors- geometry, pattern characteristics, Feed Methods, Reflector Types - Related Features, Lens Antennas - Geometry of Non-metallic Dielectric Lenses, Zoning , Tolerances, Applications, Illustrative Problems.

UNIT- IV

Antenna Arrays: Point sources - Definition, Patterns, arrays of 2 Isotropic sources- Different cases, Principle of Pattern Multiplication, Uniform Linear Arrays – Broadside Arrays, Endfire Arrays, EFA with Increased Directivity, Derivation of their characteristics and comparison, BSA with Non-uniform Amplitude Distributions – General considerations and Bionomial Arrays, Illustrative problems.

Antenna Measurements: Introduction, Concepts- Reciprocity, Near and Far Fields, Coordination system, sources of errors, Patterns to be Measured, Pattern Measurement Arrangement, Directivity Measurement , Gain Measurements (by comparison, Absolute and 3-Antenna Methods).

UNIT – V

Wave Propagation: Introduction, Definitions, Characterizations and general classifications, different modes of wave propagation, Ray/Mode concepts, Ground wave propagation (Qualitative treatment) - Introduction, Plane earth reflections, Space and surface waves, wave tilt, curved earth reflections, Space wave propagation - Introduction, field strength variation with distance and height, effect of earth"s curvature, absorption, Super refraction, M-curves and duct propagation, scattering phenomena, tropospheric propagation, fading and path loss calculations, Sky wave propagation -Introduction, structure of Ionosphere, refraction and reflection of sky waves by Ionosphere, Ray path, Critical frequency, MUF, LUF, OF, Virtual height and Skip distance, Relation between MUF and Skip distance, Multi-HOP propagation, Energy loss in Ionosphere, Summary of Wave Characteristics in different frequency ranges, Illustrative problems.

TEXT BOOKS:

1. John D. Kraus and Ronald J. Marhefka and Ahmad S.Khan, "Antennas and wave propagation," TMH, New Delhi, 4th Ed., (special Indian Edition), 2010.
2. E.C. Jordan and K.G. Balmain, "Electromagnetic Waves and Radiating Systems," PHI, 2nd Edn, 2000.

REFERENCES:

1. C.A. Balanis, "Antenna Theory- Analysis and Design," John Wiley & Sons, 2nd Edn., 2001.
2. K.D. Prasad, SatyaPrakashan, "Antennas and Wave Propagation," Tech. India Publications, New Delhi, 2001.

UNIT I

Antenna Basics & Dipole antennas

1. Fundamental Concept

Introduction:

- An **antenna** (or **aerial**) is an electrical device which converts electric power into radio waves, and vice versa. It is usually used with a radio transmitter or radio receiver. In transmission, a radio transmitter supplies an oscillating radio frequency electric current to the antenna's terminals, and the antenna radiates the energy from the current as electromagnetic waves (radio waves). In reception, an antenna intercepts some of the power of an electromagnetic wave in order to produce a tiny voltage at its terminals, that is applied to a receiver to be amplified.
- Antennas are essential components of all equipment that uses radio. They are used in systems such as radio broadcasting, broadcast television, two-way radio, communications receivers, radar, cell phones, and satellite communications, as well as other devices such as garage door openers, wireless microphones, bluetooth enabled devices, wireless computer networks, baby monitors, and RFID tags on merchandise.

- Typically an antenna consists of an arrangement of metallic conductors ("elements"), electrically connected (often through a transmission line) to the receiver or transmitter.
- Antennas act as **transformers** between conducted waves and electromagnetic waves propagating freely in space.
- Their name is borrowed from zoology, in which the Latin word *antennae* is used to describe the long, thin feelers possessed by many insects.
- In wireless communication systems, signals are radiated in space as an electromagnetic wave by using a receiving transmitting antenna and a fraction of this radiated power is intercepted by using a receiving antenna.
- An antenna is a device used for radiating or receiver radio waves. An antenna can also be thought of as a transitional structure between free space and a guiding device (such as transmission line or waveguide). Usually antennas are metallic structures, but dielectric antennas are also used now a day.
- a rigid metallic structure is called an "antenna" while the wire form is called an "aerial"

With this introduction, in this first lecture let us see some common types of antennas that are in use:

Types of Antennas:

Wire antennas: (Fig. 1, 2 and Fig. 9 single element)

- dipole, monopole, loop antenna, helix
- Usually used in personal applications, automobiles, buildings, ships, aircrafts and spacecrafts.

Aperture antennas: (Fig. 3, 4)

- horn antennas, waveguide opening
- Usually used in aircrafts and space crafts, because these antennas can be flush-

mounted.

Reflector antennas: (Fig. 5)

- **parabolic** reflectors, corner reflectors
- These are high gain antennas usually used in radio astronomy, microwave communication and satellite tracking.

Lens antennas:

- convex-plane, co vex-convex , convex-concave and concave-plane lenses
- These antennas are usually used for very high frequency applications.

Microstrip antennas: (Fig. 6)

- rectan gular, circular etc. shaped metallic patch above a ground plane
- Used in aircraft, spacecraft, s atellites, mis siles, cars, mobile phones etc.

Array antennas: (Fig. 7, and 8)

- Yagi-Uda antenn a, microstrip patch array, aperture array, slotted waveguide array.
- Used for very high gain applications with added advantage, such as, controllable radiation pattern.

Fig. 1

Fig. 2

Fig. 3

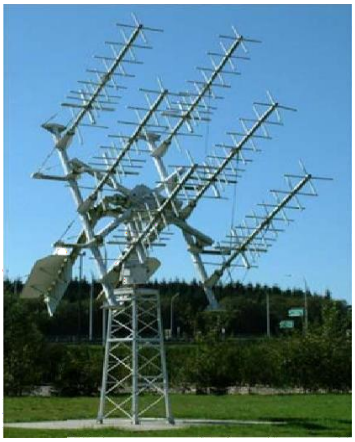


Fig. 4

Fig. 5

Fig. 6

Fig. 7

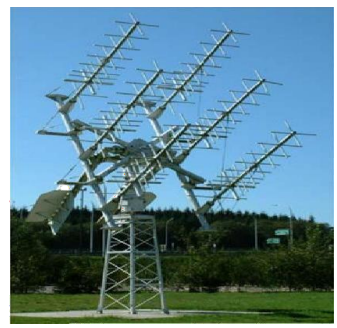


Fig. 8

Radiation Mechanism:

When electric charges undergo acceleration or deceleration, electromagnetic radiation will be produced. Hence it is the motion of charges, that is currents, is the source of radiation. Here it may be highlighted that, not all current distributions will produce a strong enough radiation for communication.

To give a mathematical flavor to it, as we know

$$A = \frac{\mu I dl}{4\pi r} \quad \text{----------1.1}$$

And

$$dl\frac{dI}{dt} = dlq\frac{dv}{dt} = dlqa \quad \text{----------1.2}$$

So

$$\boxed{E = -\nabla V - \frac{\partial A}{\partial t} = -\nabla V - \frac{\mu dl}{4\pi r}\frac{\partial I}{\partial t} = -\nabla V - \frac{\mu dl}{4\pi r}qa} \quad \text{----------1.3}$$

- As shown in these equations, to creat radiation (electric field), there must be a time-varying current dI/dt or an acceleration (or deceleration) a of a charge q.
- If the charge is not moving, a current is not created and there is no radiation.
- If a charge is moving with an uniform velocity,
  - there is no radiation if the wire is straight, and infinite in extent
  - there is radiation if the wire is curved, bent, discontinuous, terminated, or truncated
  - If the charge is oscillating in a time-motion, it radiates even if the wire is straight.

These situations are shown in Fig. 9.

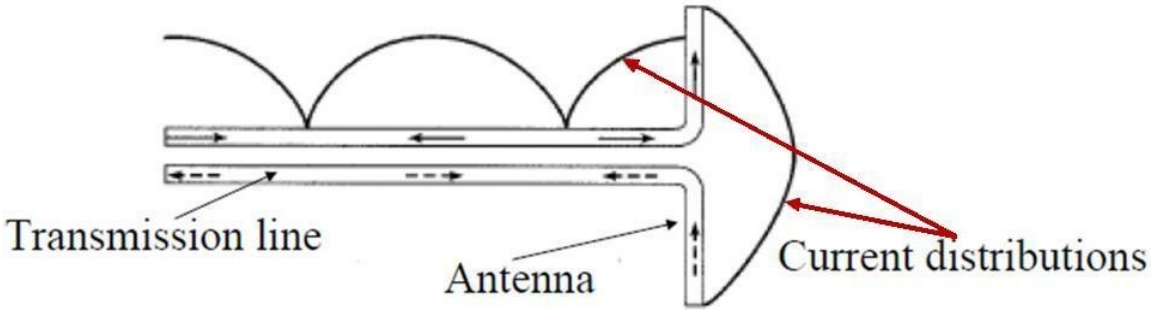


Fig. 9: Conditions for radiation

So, it is the current distribution on the antennas that produce the radiation. Usually these current distributions are excited by transmission lines and waveguides (Fig. 10)

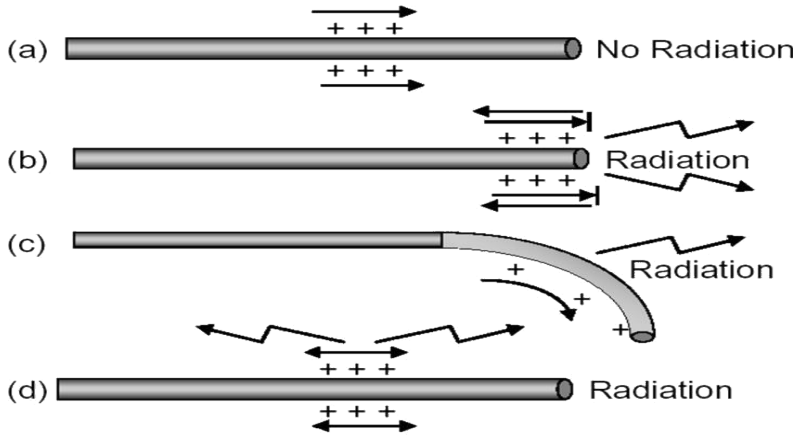


Fig. 10: Antenna radiation mechanism

Principle- Under time varying conditions , Maxwell_s equations predict the radiation of EM energy from current source(or accelerated charge). This happens at all frequencies , but is insignificant as long as the size of the source region is not comparable to the wavelength. While transmission.lines are designed to minimize this radiation loss, radiation into free space becomes main purpose in case of Antennas . For steady state harmonic variation, usually we focus on time changing current For transients or pulses ,we focus on accelerated charge The radiation is perpendicular to the acceleration. The radiated power is proportional to the square of .

I L or Q *V*

Where I = Time changing current in Amps/sec

L = Length of the current element in meters

Q= Charge in Coulombs

V= Time changing velocity

<u>Transmission line opened out in a Tapered fashion as Antenna:</u>

a) **As Transmitting Antenna**: –Here the Transmission Line is connected to source or generator at one end. Along the uniform part of the line energy is guided as Plane TEM wave with little loss. Spacing between line is a small fraction of λ. As the line is opened out and the separation b/n the two lines becomes comparable to λ, it acts like an antenna and launches a free space wave since currents on the transmission Line flow out on the antenna but fields associated with them keep on going. From the circuit point of view the antennas appear to the tr. lines As a resistance R_r, called Radiation resistance

b) **As Receiving Antenna** –Active radiation by other Antenna or Passive radiation from distant objects raises the apparent temperature of R_r .This has nothing to do with the physical temperature of the antenna itself but is related to the temperature of distant objects that the antenna is looking at. R_r may be thought of as virtual resistance that does not exist physically but is a quantity coupling the antenna to distant regions of space via a virtual transmission .line

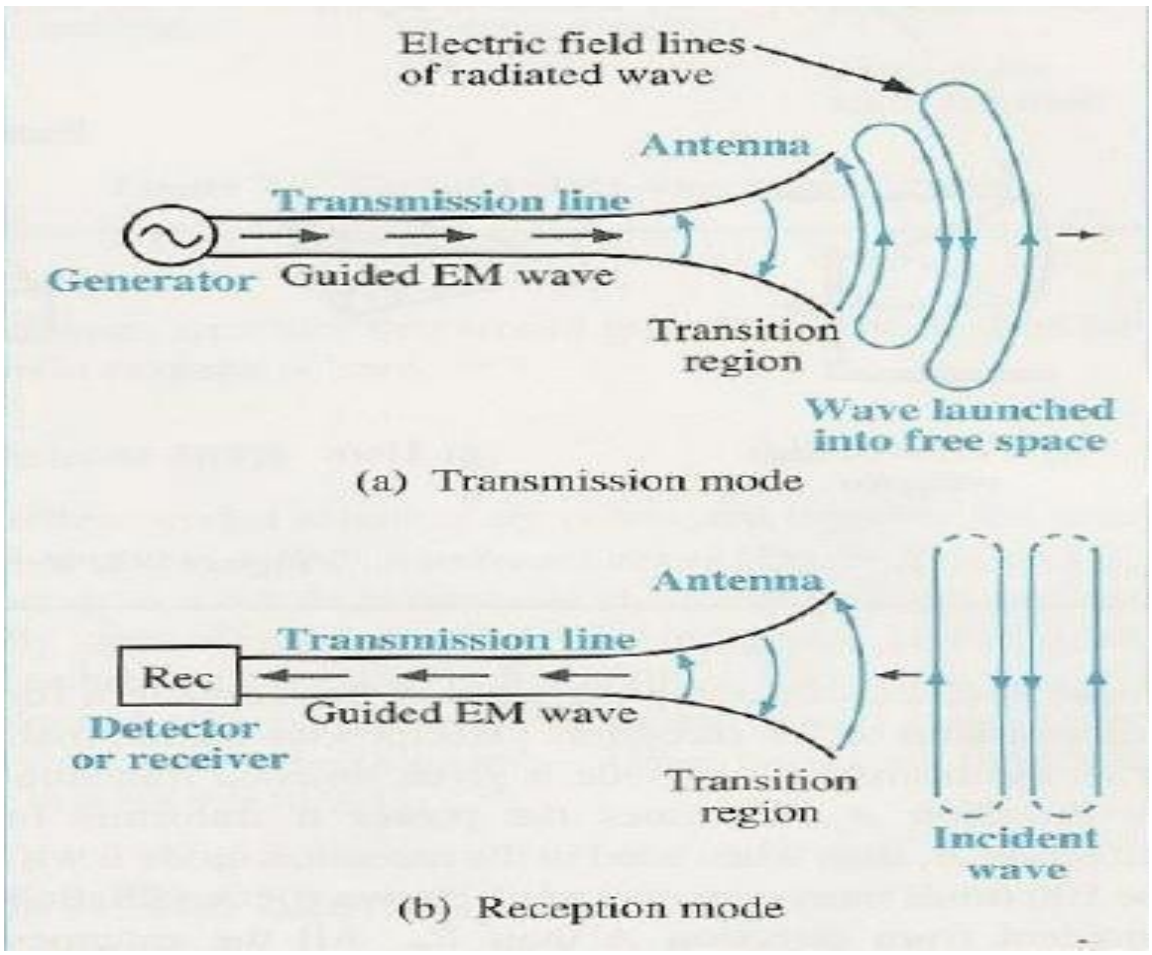


Figure11:Antenna as a a) Transmission Mode b) Receiving Mode

Reciprocity-An antenna exhibits identical impedance during Transmission or Reception, same directional patterns during Transmission or Reception, same effective height while transmitting or receiving **.** Transmission and reception antennas can be used interchangeably. Medium must be linear, passive and isotropic(physical properties are the same in different directions.) Antennas are usually optimised for reception or transmission, not both.

Current and voltage distribution.

a) A current flowing in a wire of a length related to the RF produces an electromagnetic field. This field radiates from the wire and is set free in space. The principles of radiation of electromagnetic energy are based on two laws.
   (1) A moving electric field creates a magnetic (H) field.
   (2) A moving magnetic field creates an electric (E) field.
b) In space, these two fields will be in-phase and perpendicular to each other at any given moment. Although a conductor is usually considered to be present when a moving electric or magnetic field is mentioned, the laws governing these fields do not say anything about a conductor. Thus, these laws hold true whether a conductor is present or not.
c) The current and voltage distribution on a half-wave Hertz antenna is shown in Figure 1-1. In view A, a piece of wire is cut in half and attached to the terminals of a high frequency (HF), alternating current (AC) generator. The frequency of the generator is set so each half of the wire is one-quarter wavelength of the output. The symbol for wavelength is the Greek letter lambda (D). The result is the common dipole antenna.
d) At a given moment, the generator's right side is positive and its left side is negative. A law of physics states that like charges repel each other. Consequently, electrons will flow away from the negative terminal as far as possible while the positive terminal will attract electrons. View B of Figure 1-1 shows the direction and distribution of electron flow. The distribution curve shows that most current flows in the center and none flows at the ends. The current distribution over the antenna is always the same, regardless of how much or how little current is flowing. However, current at any given point on the antenna will vary directly with the amount of voltage that the generator develops.
e) One-quarter cycle after the electrons begin to flow, the generator develops it; minimum voltage and the current decreases to zero. At that moment, the condition shown in view C of Figure 1-1 will exist. Although no current is flowing, a minimum number of electrons are at the left end of the line and a minimum number are at the right end. The charge distribution along the wire varies as the voltage of the generator varies (view C).
f)

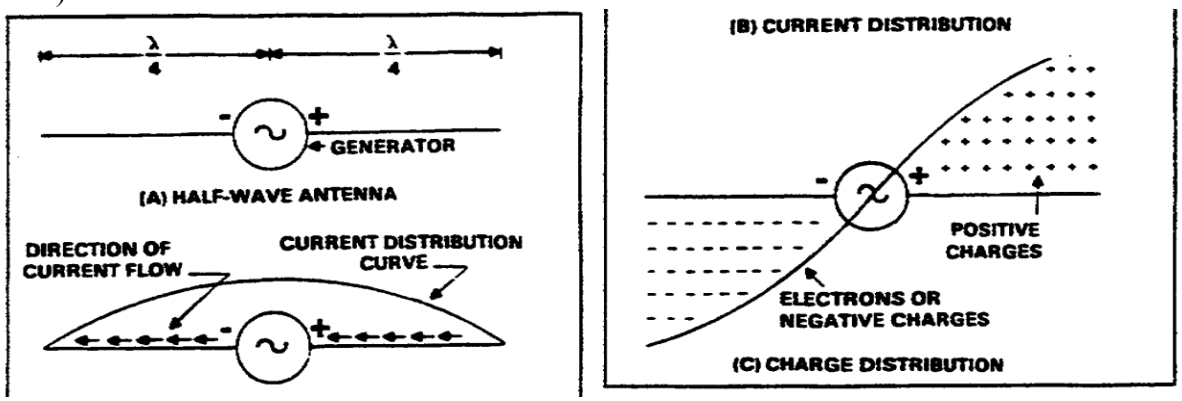


Figure 12. Current and voltage distribution on an antenna

1. A current flows in the antenna with an amplitude that varies with the generator voltage.
2. A sine wave distribution of charge exists on the antenna. The charges reverse polarity every half cycle.
3. The sine wave variation in charge magnitude lags the sine wave variation in current by one-quarter cycle.

Antenna Parameters:

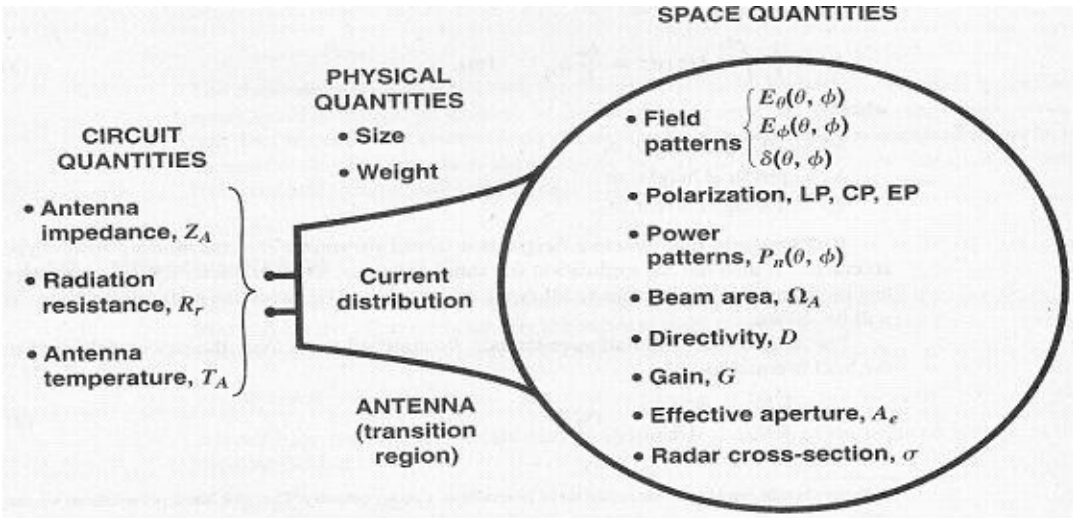


Figure 13 :Schematic diagram of basic parameters

Dual Characteristics of an Antenna

The duality of an antenna specifies a circuit device on one band and a space device on the other hand. Figure 13 shows the schematic diagram of basic antenna parameters, illustrating dual characteristics of an antenna.

Most practical transmitting antennas are divided into two basic classifications, HERTZ ANTENNAS (half-wave) and MARCONI (quarter-wave) ANTENNAS. Hertz antennas are generally installed some distance above the ground and are positioned to radiate either vertically or horizontally. Marconi antennas operate with one end grounded and are mounted perpendicular to the earth or a surface acting as a ground. The Hertz antenna, also referred to as a dipole, is the basis for some of the more complex antenna systems used today. Hertz antennas are generally used for operating frequencies of 2 MHz and above, while Marconi antennas are used for operating frequencies below 2 MHz.All antennas, regardless of their shape or size, have four basic characteristics: reciprocity, directivity, gain, and polarization.

Isotropic Radiator:An antenna does not radiate uniformly in all directions. For the sake of a reference, we consider a hypothetical antenna called an isotropic radiator having equal radiation in all directions.

Directional Antenna: A directional antenna is one which can radiate or receive electromagnetic waves more effectively in some directions than in others.

Radiation Pattern:

The relative distribution of radiated power as a function of direction in space (i.e., as function of and) is called the radiation pattern of the antenna. Instead of 3D surface, it is common practice to show planar cross section radiation pattern. E-plane and H-plane patterns give two most important views. The E-plane pattern is a view obtained from a section containing maximum value of the radiated field and electric field lies in the plane of the section. Similarly when such a section is taken such that the plane of the section contains H field and the direction of maximum radiation.A typical radiation patter plot is shown in figure 14.

The main lobe contains the direction of maximum radiation. However in some antennas, more than one major lobe may exist. Lobe other than major lobe are called minor lobes. Minor lobes can be further represent radiation in the considered direction and require to be minimized.

HPBW or half power beam width refers to the angular width between the points at which the radiated power per unit area is one half of the maximum.

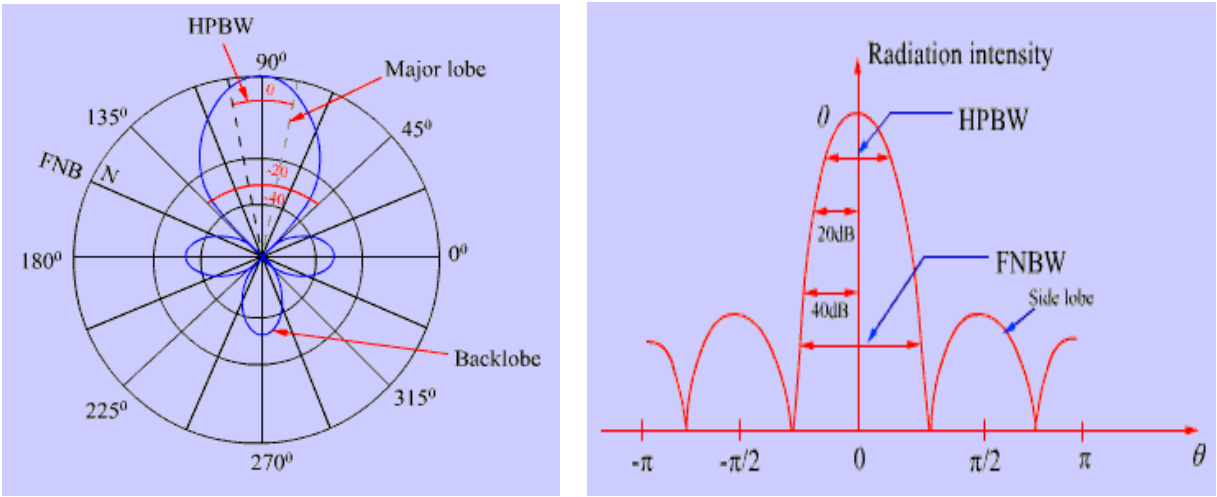


Figure 14: Radiation Pattern

Similarly FNBW (First null beam width) refers to the angular width between the first two nulls as shown in Figure 14. By the term beam width we usually refer to 3 dB beam width or HPBW.

RECIPROCITY is the ability to use the same antenna for both transmitting and receiving. The electrical characteristics of an antenna apply equally, regardless of whether you use the antenna for transmitting or receiving. The more efficient an antenna is for transmitting a certain frequency, the more efficient it will be as a receiving antenna for the same frequency. This is illustrated by figure 2-1, view A. When the antenna is used for transmitting, maximum radiation occurs at right angles to its axis. When the same antenna is used for receiving (view B), its best reception is along the same path; that is, at right angles to the axis of the antenna.

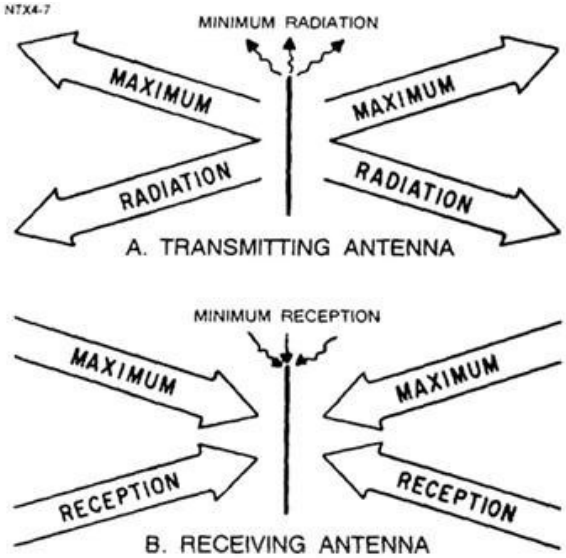
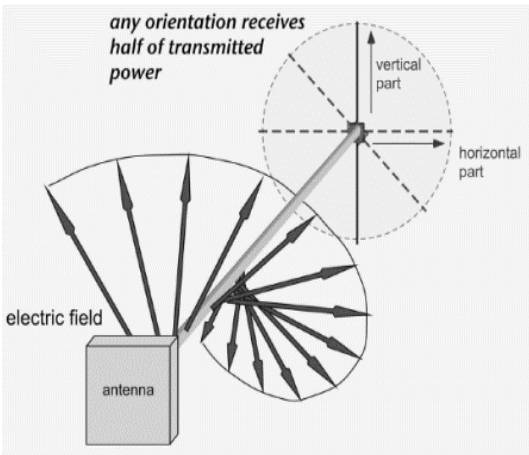


Figure 13. Reciprocity of Antenna

Polarization of an electromagnetic wave refers to the orientation of the electric field component of the wave. For a linearly polarized wave, the orientation stays the same as the wave moves through space. If we choose our axis system such that the electric field is vertical, we say that the wave is vertically polarized. If our transmitting antenna is vertically oriented, the electromagnetic wave radiated is vertically polarized since, as we saw before, the electric field is in the direction of the current in the antenna.

The convention is to refer to polarization with reference to the surface of the earth. Precise orientation is less problematic than one might think, since waves bounce of the ground and other objects so do not maintain their original orientation anyway. In space, horizontal and vertical lose their meaning, so alignment of linearly polarized sending and receiving antennas is more difficult to achieve. These difficulties are somewhat circumvented by circular polarization of waves. With circular polarization, the tip of the electric field vector traces out a circle when viewed in the direction of propagation.



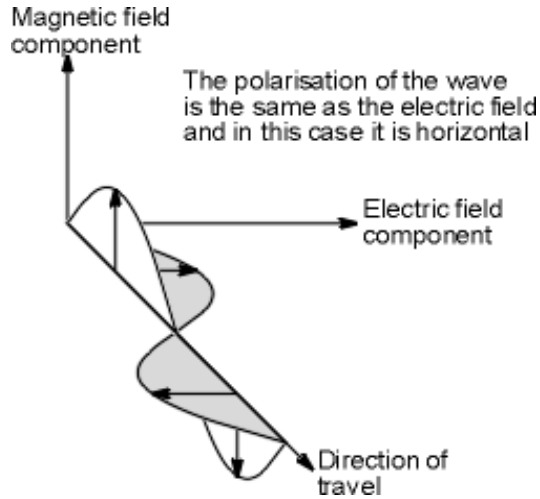


Figure 15. Polarisation

Polarization categories

Vertical and horizontal are the simplest forms of polarization and they both fall into a category known as linear polarization. However it is also possible to use circular polarization. This has a number of benefits for areas such as satellite applications where it helps overcome the effects of propagation anomalies, ground reflections and the effects of the spin that occur on many satellites. Circular polarization is a little more difficult to visualize than linear polarization. However it can be imagined by visualizing a signal propagating from an antenna that is rotating. The tip of the electric field vector will then be seen to trace out a helix or corkscrew as it travels away from the antenna. Circular polarization can be seen to be either right or left handed dependent upon the direction of rotation as seen from the transmitter.

Another form of polarization is known as elliptical polarization. It occurs when there is a mix of linear and circular polarization. This can be visualized as before by the tip of the electric field vector tracing out an elliptically shaped corkscrew.

However it is possible for linearly polarized antennas to receive circularly polarized signals and vice versa. The strength will be equal whether the linearly polarized antenna is mounted vertically, horizontally or in any other plane but directed towards the arriving signal. There will be some degradation because the signal level will be 3 dB less than if a circularly polarized

antenna of the same sense was used. The same situation exists when a circularly polarized antenna receives a linearly polarized signal.

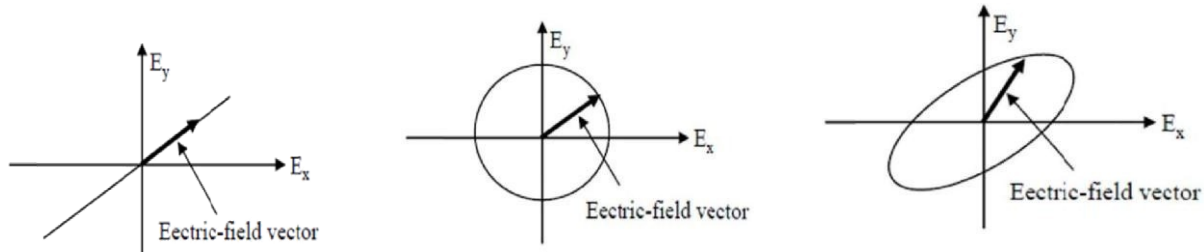


Figure: (a) Linear polarization (b) Circular polarization (c) Elliptical polarization

DIRECTIVITY

The DIRECTIVITY of an antenna or array is a measure of the antenna‘s ability to focus the energy in one or more specific directions. You can determine an antenna‘s directivity by looking at its radiation pattern. In an array propagating a given amount of energy, more radiation takes place in certain directions than in others. The elements in the array can be arranged so they change the pattern and distribute the energy more evenly in all directions. The opposite is also possible. The elements can be arranged so the radiated energy is *focused* in one direction. The elements can be considered as a group of antennas fed from a common source.

It is defined as the ratio of maximum radiation intensity of subject or test antenna to the radiation intensity of an isotropic antenna.
(or)
Directivity is defined as the ratio of maximum radiation intensity to the average radiation intensity.
Directivity (D) in terms of total power radiated is,

D = 4π x Maximum radiation intensity/ Total power radiated

Gain:

Gain is a parameter which measures the degree of directivity of the antenna's radiation pattern. A high-gain antenna will preferentially radiate in a particular direction. Specifically, the *antenna gain*, or *power gain* of an antenna is defined as the ratio of the intensity (power per unit surface) radiated by the antenna in the direction of its maximum output, at an arbitrary distance, divided by the intensity radiated at the same distance by a hypothetical isotropic antenna.

As we mentioned earlier, some antennas are highly directional. That is, they propagate more energy in certain directions than in others. The ratio between the amount of energy propagated in these directions and the energy that would be propagated if the antenna were not directional is known as antenna GAIN. The gain of an antenna is constant. whether the antenna is used for transmitting or receiving.

Directivity function $D(\theta, \phi)$ describes the variation of the radiation intensity. The directivity function $D(\theta, \phi)$ is defined by

$$D(\theta, \phi) = \frac{\text{Power radiated per unit solid angle}}{\text{Average power radiated per unit solid an gle}} \quad \text{---------- (1)}$$

If P_r is the radiated power, the $\frac{dP_r}{d\Omega}$ gives the amount of power radiated per unit solid angle. Had this power beam uniformly radiated in all directions then average power radiated per unit solid angle is $\frac{P_r}{4\pi}$

$$D(\theta,\phi) = \frac{\frac{dP_r}{d\Omega}}{\frac{P_r}{4\pi}} = \frac{\frac{dP_r}{d\Omega}}{P_r} 4\pi \quad (2)$$

The maximum of directivity function is called the directivity.

In defining directivity function total radiated power is taken as the reference. Another parameter called the gain of an antenna is defined in the similar manner which takes into account the total input power rather than the total radiated power is used as the reference. The amount of power given as input to the antenna is not fully radiated.

$$P_r = \eta P_{in} \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots (3)$$

where η is the radiation efficiency of the antenna.

The gain of the antenna is defined as

$$G(\theta,\phi) = \frac{\text{Radiated power per unit solid angle}}{\text{input power}} 4\pi \quad \text{---------- (4)}$$

$$G(\theta,\phi) = \eta D(\theta,\phi) \quad \text{--------(5)}$$

The maximum gain function is termed as gain of the antenna.

Another parameter which incorporates the gain is effective isotropic radiated power or EIRP which is defined as the product of the input power and maximum gain or simply the gain. An antenna with a gain of 100 and input power of 1 W is equally effective as an antenna having a gain of 50 and input power 2 W.

Radiation resistance:

The radiation resistance of an antenna is defined as the equivalent resistance that would dissipate the same amount power as is radiated by the antenna. For the elementary current element we have discussed so far. From equation (3.26) we find that radiated power density

$$P_{av} = \frac{I^2 \eta \, dl^2 \, k^2 \sin^2\theta}{\text{Radiated power}}$$

$_0$ $_0$ a_r $32 \; {}^2 r^2$ -------- (1)

$$P_r = \frac{I_0^2 (dl)^2 k_0^2}{32\pi r^2} \int_0^{2\pi} \int_0^{\pi} r^2 \sin\theta \, d\theta \, d\phi = \frac{I_0^2 (dl)^2 k_0^2}{32\pi^2} \int_0^{2\pi} \int_0^{\pi} \sin^3\theta \, d\theta \, d\phi \quad \text{---- (2)}$$

$$P = \frac{I_0^2 (dl)^2 k_0^2}{12\pi} \quad \text{..... (3)}$$

Further,

$$\frac{dP_r}{d\Omega} = \frac{r^2}{P_{av}} \sin\theta \, d\theta \, \hat{a}_r = P_{av} \, \hat{a}_r \, r^2 d\Omega$$

$$\frac{dP_r}{d\Omega} = \frac{I_0^2 (dl)^2 k_0^2}{32\pi^2} \sin^2\theta \quad (4)$$

From (3) and (4)

$$D_n(\theta, \phi) = 1.5 \sin^2\theta$$

Directivity $D = D_n(\theta, \phi)_{max}$ which occurs at $\theta = \frac{\pi}{2}$.

If R_r is the radiation resistance of the elementary dipole antenna, then

$$\frac{1}{2} I^2 R_r = P_r$$

Substituting P_r from (3) we get

$$R_r = \frac{\eta_0}{6\pi} \left(\frac{2\pi}{k_0}\right)^2 \left(\frac{dl}{\lambda_0}\right)^2 \quad \text{---------- (5)}$$

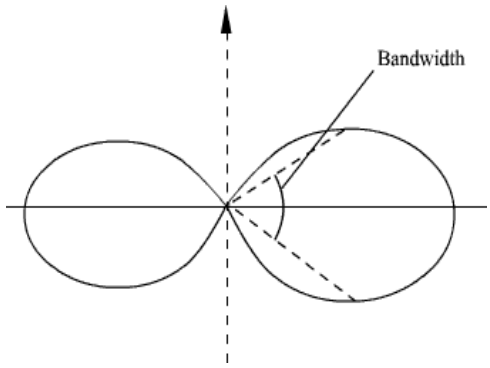
Substituting $\eta_0 = 120\pi$

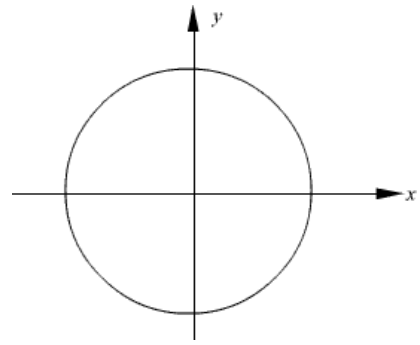
$$R_r = \frac{480\pi^3}{6\pi} \left(\frac{dl}{\lambda_0}\right)^2 \quad \text{------- (6)}$$

$$R_r = 80\pi^2 \left(\frac{dl}{\lambda_0}\right)^2 \quad \text{..... (7)}$$

For such an elementary dipole antenna the principal *E* and *H* plane pattern are shown in Fig 16(a) and (b).

Figure16 (a) Principal E plane pattern





The bandwidth (3 dB beam width) can be found to be 90^0 in the *E* plane.

Effective Area of an Antenna:

An antenna operating as a receiving antenna extracts power from an incident electromagnetic wave. The incident wave on a receiving antenna may be assumed to be a uniform plane wave being intercepted by the antenna. This is illustrated in Fig 3.5. The incident electric field sets up currents in the antenna and delivers power to any load connected to the antenna. The induced current also re-radiates fields known as scattered field. The total electric field outside the antenna will be sum of the incident and scattered fields and for perfectly conducing antenna the total tangential electric field component must vanish on the antenna surface.

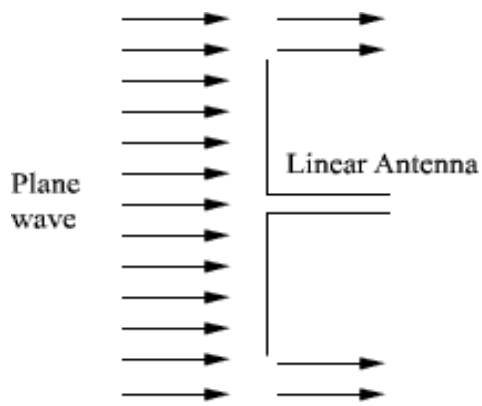


Fig 17: Plane wave intercepted by an antenna

Let P_{inc} represents the power density of the incident wave at the location of the receiving antenna and P_L represents the maximum average power delivered to the load under matched conditions with the receiving antenna properly oriented with respect to the polarization of the incident wave.

We can write,

$$\frac{P_L}{P_{inc}} = A_{em} \quad \text{..... (9)}$$

where $P_{inc} = \frac{E^2}{2\eta_0}$ and the term A_{em} is called the maximum effective aperture of the antenna. A_{em} is related to the directivity of the antenna D as,

$$D = \frac{4\pi}{\lambda^2} A_{em} \quad \text{--------- (10)}$$

If the antenna is lossy then some amount of the power intercepted by the antenna will be dissipated in the antenna.

From eqn. (2) we find that

$$G = \eta D$$

Therefore, from (5),

$$G = \frac{4\pi}{\lambda^2} \cdot \eta A_{em} = \frac{4\pi}{\lambda^2} A_e \quad \text{.....(11)}$$

$A_e = \eta A_{em}$ is called the effective aperture of the antenna (in m^2).

So effective area or aperture A_e of an antenna is defined as that equivalent area which when intercepted by the incident power density P_{in} gives the same amount of received power P_R which is available at the antenna output terminals.

If the antenna has a physical aperture A then aperture efficiency $\epsilon_a = \frac{A_e}{A}$

Effective length/height of the antenna:

When a receiving antenna intercepts incident electromagnetic waves, a voltage is induced across the antenna terminals. The effective length he of a receiving antenna is defined as the ratio of the open circuit terminal voltage to the incident electric field strength in the direction of antennas polarization.

$$h_e = \frac{V_{oc}}{E} \text{ m} \quad \text{.....(12)}$$

where V_{oc} = open circuit voltage

E = electric field strength

Effective length he is also referred to as effective height.

Radian and Steradian:

Radian is plane angle with it_s vertex a the centre of a circle of radius r and is subtended by an arc whose length is equal to r. Circumference of the circle is 2πr Therefore total angle of the circle is 2π radians.

Steradian is solid angle with it_s vertex at the centre of a sphere of radius r, which is subtended by a spherical surface area equal to the area of a square with side length r Area of the sphere is $4\pi r^2$.Therefore the total solid angle of the sphere is 4π steradians

Beam Area

In polar two-dimensional coordinates an incremental area dA on the surface of sphere is the product of the length $r\,d\theta$ in the θ direction and $r \sin\theta\, d\Phi$ in the Φ direction as shown in figure

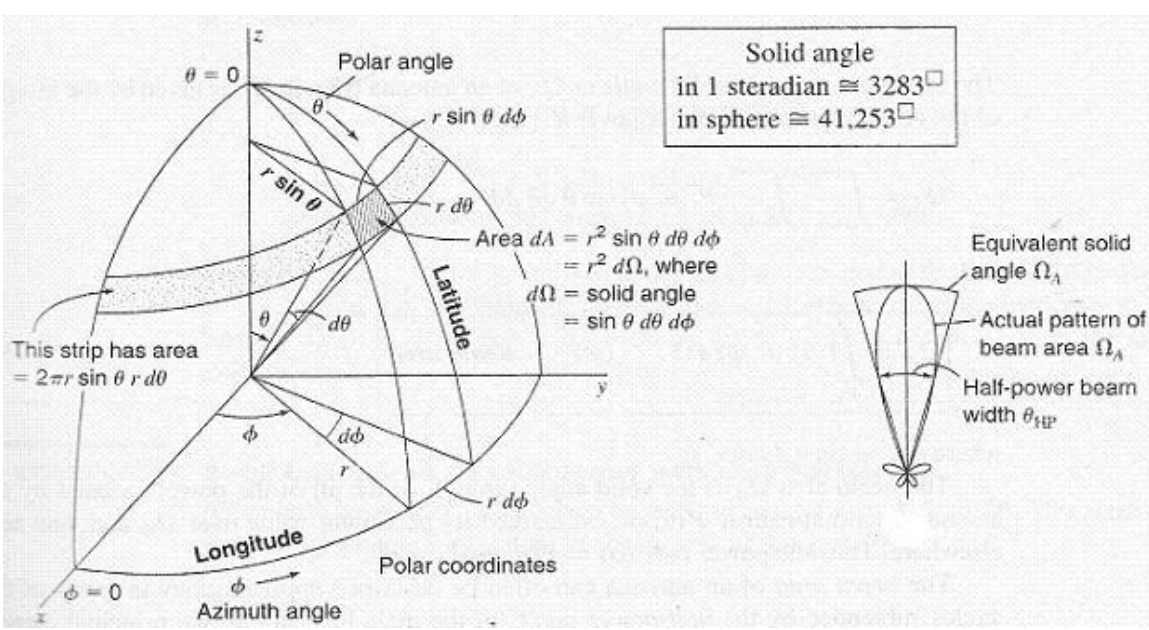


Figure 18: radian and steradian

Thus

$$dA = (rd\theta)\ (r\ sin\theta\ d\Phi) = r^2\ d\Omega$$

Where,

$d\Omega$ = solid angle expressed in steradians.

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,

Area of sphere = $2\pi r2$

$= 2\pi r^2$ [-cosθ]0

$\pi = 4\pi r^2$

Where,

4π = Solid angle subtended by a sphere

The beam area or beam solid angle or ΩA of an antenna is given by the integral of the normalized power pattern over a sphere

Beam area, ΩA = Ω (sr)

Where,

$d\Omega = sin\theta\ d\theta\ d\Phi$

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 $U\ (\theta\ ,\ \Phi\ ),$ as a function of angle, to its maximum value. Thus,

$$Pn(\theta,\Phi) = U(\theta,\Phi)/U(\theta,\Phi)max = S(\theta,\Phi)/S(\theta,\Phi)max$$

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 beam area *QA* (or beam solid angle) consists of the main beam area (or solid angle) ΩM plus the minor-lobe area (or solid angle) $\Omega\ m$. 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 εM. Thus,

Beam Efficiency = εM = ΩM/ ΩA (dimensionless)

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

εm = $\Omega\ m$ / ΩA = stray factor.

Bandwidth

Note that the system is designed for specific frequency; i.e. at any other frequency it will not be one-half wavelength. The *bandwidth* of an antenna is the range of frequencies over which the antenna gives reasonable performance. One definition of reasonable performance is that the standing wave ratio is 2:1 or less at the bounds of the range of frequencies over which the antenna is to be used.

Antenna Equivalent Circuit:

To a generator feeding a transmitting antenna, the antenna appears as a lead. In the same manner, the receiver circuitry connected to a receiving antenna's output terminal will appear as load impedance. Both transmitting and receiving antennas can be represented by equivalent circuits as shown by figure 18(a) and figure 18(b).

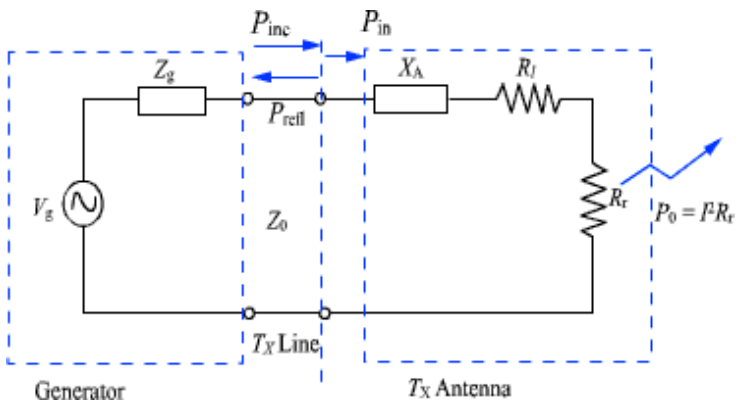


Fig 18 (a): Equivalent circuit of a T_x antenna

V_g = open circuit voltage of the generator
Z_g = antenna impedance
Z_0 = Characteristics impedance of the transmission line connecting generator to the antenna.
P_{inc} = Incident power to the antenna terminal P_{refl}
= Power reflected from the antenna terminal. P_{in}
= Input power to the antenna
X_A = Antenna reactance
R_l = Loss resistance of the antenna
R_r = Radiation resistance

Z_A ● R_l R_r jX_A ● R_A jX_A antenna impedance.

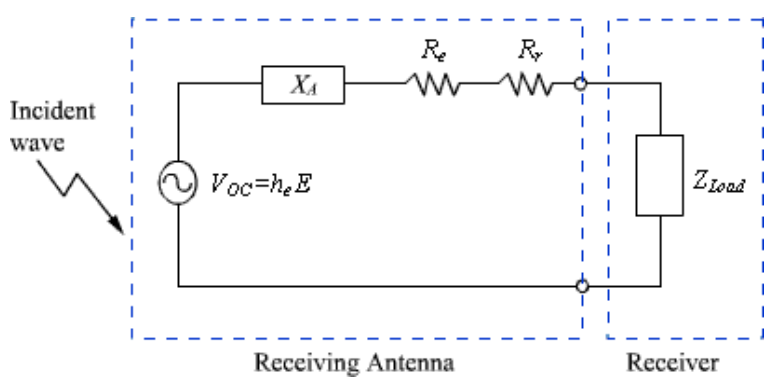
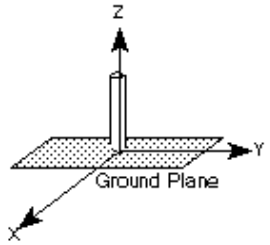
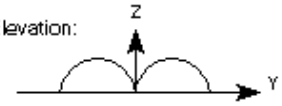
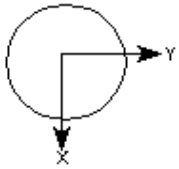
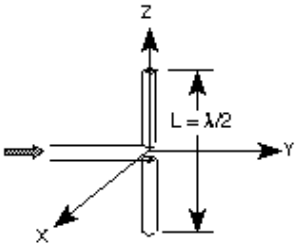
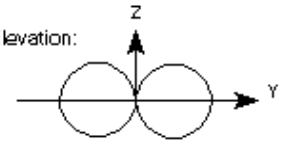
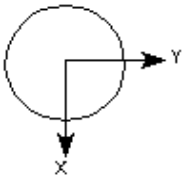
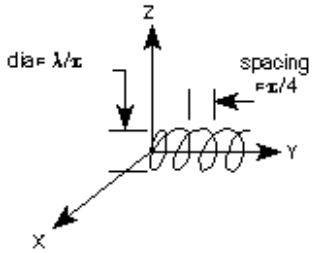
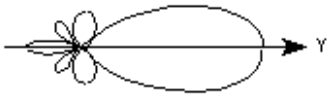
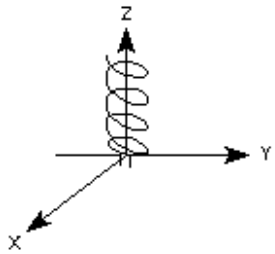
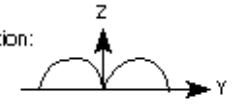
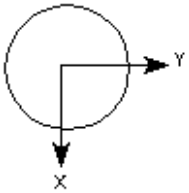


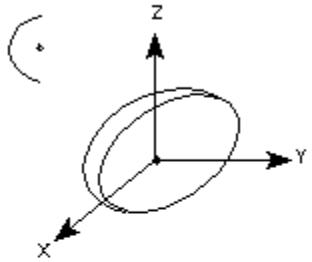
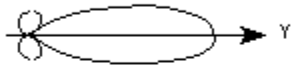
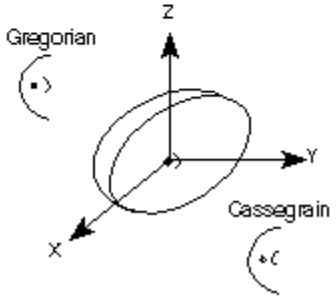
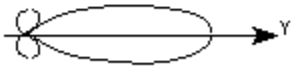
Fig 18 (b): Equivalent circuit of receiving antenna

h_e = effective length
E = incident field strength
V_{oc} = h0 E open circuit voltage
Z_{load} = Input impedance of the receiver.
R_e, R_r and X_A as defined earlier.

Antennas Radiation Patterns

Antenna Type	Radiation Pattern	Characteristics
MONOPOLE 	Elevation:  Azimuth: 	**Polarization: Linear** Vertical as shown **Typical Half-Power Beamwidth** 45 deg x 360 deg **Typical Gain:** 2-6 dB at best **Bandwidth:** 10% or 1.1:1 **Frequency Limit** **Lower:** None **Upper:** None **Remarks:** Polarization changes to horizontal if rotated to horizontal
λ/2 DIPOLE 	Elevation:  Azimuth: 	**Polarization:** Linear Vertical as shown **Typical Half-Power Beamwidth** 80 deg x 360 deg **Typical Gain:** 2 dB **Bandwidth:** 10% or 1.1:1 **Frequency Limit** **Lower:** None **Upper:** 8 GHz (practical limit) **Remarks:** Pattern and lobing changes significantly with L/f. Used as a gain reference < 2 GHz.

Antenna Type	Radiation Pattern	Characteristics
AXIAL MODE HELIX 	Elevation & Azimuth 	**Polarization:** Circular Left hand as shown **Typical Half-Power Beamwidth:** 50 deg x 50 deg **Typical Gain:** 10 dB **Bandwidth:** 52% or 1.7:1 **Frequency Limit** **Lower:** 100 MHz **Upper:** 3 GHz **Remarks:** Number of loops >3
NORMAL MODE HELIX 	Elevation:  Azimuth: 	**Polarization:** Circular - with an ideal pitch to diameter ratio. **Typical Half-Power Beamwidth:** 60 deg x 360 deg **Typical Gain:** 0 dB **Bandwidth:** 5% or 1.05:1 **Frequency Limit** **Lower:** 100 MHz **Upper:** 3 GHz

Antenna Type	Radiation Pattern	Characteristics
PARABOLIC (Prime) 	Elevation & Azimuth 	**Polarization:** Takes polarization of feed **Typical Half-Power Beamwidth:** 1 to 10 deg **Typical Gain:** 20 to 30 dB **Bandwidth:** 33% or 1.4:1 limited mostly by feed **Frequency Limit:** **Lower:** 400 MHz **Upper:** 13+ GHz
PARABOLIC 	Elevation & Azimuth 	**Polarization:** Takes polarization of feed **Typical Half-Power Beamwidth:** 1 to 10 deg **Typical Gain:** 20 to 30 dB **Bandwidth:** 33% or 1.4:1 **Frequency Limit:** **Lower:** 400 MHz **Upper:** 13+ GHz

UNIT II

VHF, UHF and Microwave Antennas – I

Types of Antennas

short dipole antenna

The **short dipole antenna** is the simplest of all antennas. It is simply an open-circuited wire, fed at its center as shown in Figure 1.

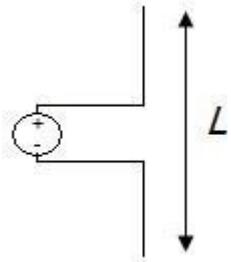


Figure 1. Short dipole antenna of length *L*.

The words "short" or "small" in antenna engineering always imply "relative to a wavelength". So the absolute size of the above dipole antenna does not matter, only the size of the wire relative to the wavelength of the frequency of operation. Typically, a dipole is short if its length is less than a tenth of a wavelength:

$$L < \frac{\lambda}{10}$$

If the short dipole antenna is oriented along the z-axis with the center of the dipole at z=0, then the current distribution on a thin, short dipole is given by:

$$I(z) = I_0(1 - \frac{2|z|}{L})$$

The current distribution is plotted in Figure 2. Note that this is the amplitude of the current distribution; it is oscillating in time sinusoidally at frequency *f*.

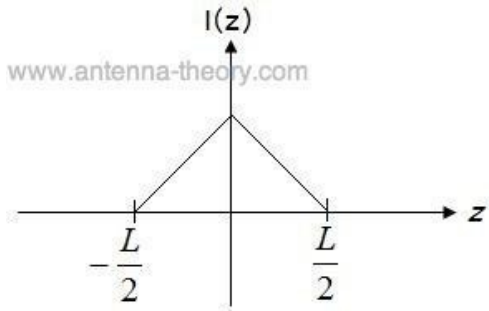


Figure 2. Current distribution along a short dipole antenna.

The fields radiated from the short dipole antenna in the far field are given by:

$$E_\theta = \frac{j\eta k I_0 L e^{-jkr}}{8\pi r}\sin\theta$$

$$H_\phi = \frac{E_\theta}{\eta}$$

$$E_r = H_r = E_\phi = H_\theta = 0$$

The above equations can be broken down and understood somewhat intuitively. First, note that in the far-field, only the E_θ and H_ϕ fields are nonzero. Further, these fields are orthogonal and in-phase. Further, the fields are perpendicular to the direction of propagation, which is always in the $\hat{r}$ direction (away from the antenna). Also, the ratio of the E-field to the H-field is given by η (the intrinsic impedance of free space).

This indicates that in the far-field region the fields are propagating like a plane-wave.
Second, the fields die off as 1/r, which indicates the power falls of as

$$P(r) \propto \frac{1}{r^2}$$

Third, the fields are proportional to *L*, indicated a longer dipole will radiate more power. This is true as long as increasing the length does not cause the short dipole assumption to become invalid. Also, the fields are proportional to the current amplitude I_0, which should make sense (more current, more power).
The exponential term:

$$e^{-jkr}$$

describes the phase-variation of the wave versus distance. The parameter k is known as the wavenumber. Note also that the fields are oscillating in time at a frequency f in addition to the above spatial variation.

Finally, the spatial variation of the fields as a function of direction from the antenna are given by $\sin\theta$. For a vertical antenna oriented along the z-axis, the radiation will be maximum in the x-y plane. Theoretically, there is no radiation along the z-axis far from the antenna.

Directivity, Impedance and other Properties of the Short Dipole Antenna

The directivity of the center-fed short dipole antenna depends only on the $\sin\theta$ component of the fields. It can be calculated to be 1.5 (1.76 dB), which is very low for realizable (physical or non-theoretical) antennas. Since the fields of the short dipole antenna are only a function of the polar angle, they have no azimuthal variation and hence this antenna is characterized as omnidirectional. The Half-Power Beamwidth is 90 degrees.
The polarization of this antenna is linear. When evaluated in the x-y plane, this antenna would be described as vertically polarized, because the E-field would be vertically oriented (along the z-axis).
We now turn to the input impedance of the short dipole, which depends on the radius a of the dipole. Recall that the impedance Z is made up of three components, the radiation resistance, the loss resistance, and the reactive (imaginary) component which represents stored energy in the fields:

$$Z = R_{rad} + R_{loss} + jX$$

The radiation resistance can be calculated to be:

$$R_{rad} = 20\pi^2\left(\frac{L}{\lambda}\right)^2$$

The resistance representing loss due to the finite-conductivity of the antenna is given by:

$$R_{loss} = \frac{L}{6\pi a}\sqrt{\frac{\pi f \mu}{2\sigma}}$$

In the above equation σ represents the conductivity of the dipole (usually very high, if made of metal). The frequency f come into the above equation because of the skin effect. The reactance or imaginary part of the impedance of a dipole is roughly equal to:

$$X = \frac{-120\lambda}{\pi L}\left(\ln\left(\frac{L}{2a}\right) - 1\right)$$

As an example, assume that the radius is 0.001 λ and the length is 0.05 λ. Suppose further that this antenna is to operate at f=3 MHz, and that the metal is copper, so that the conductivity is 59,600,000 S/m.
The radiation resistance is calculated to be 0.49 Ohms. The loss resistance is found to be 4.83 mOhms (milli-Ohms), which is approximatley negligible when compared to the radiation

resistance. However, the reactance is 1695 Ohms, so that the input resistance is Z=0.49 + j1695. Hence, this antenna would be very difficult to have proper impedance matching. Even if the reactance could be properly cancelled out, very little power would be delivered from a 50 Ohm source to a 0.49 Ohm load.

For short dipole antennas that are smaller fractions of a wavelength, the radiation resistance becomes smaller than the loss resistance, and consequently this antenna can be very inefficient.

The bandwidth for short dipoles is difficult to define. The input impedance varies wildly with frequency because of the reactance component of the input impedance. Hence, these antennas are typically used in narrowband applications.

In the next section, we'll look at general dipole antennas.

Dipole Antenna

The **dipole antenna** with a very thin radius is considered. The dipole antenna is similar to the short dipole except it is not required to be small compared to the wavelength (at the frequency the antenna is operating at).

For a dipole antenna of length *L* oriented along the z-axis and centered at z=0, the current flows in the z-direction with amplitude which closely follows the following function:

$$I(z) = \begin{cases} I_0 \sin\left[k\left(\frac{L}{2} - z\right)\right], & 0 \le z \le \frac{L}{2} \\ I_0 \sin\left[k\left(\frac{L}{2} + z\right)\right], & -\frac{L}{2} \le z \le 0 \end{cases}$$

Note that this current is also oscillating in time sinusoidally at frequency *f*. The current distributions for the quarter-wavelength (left) and full-wavelength (right) dipole antennas are given in Figure 1. Note that the peak value of the current I_0 is not reached along the dipole unless the length is greater than half a wavelength.

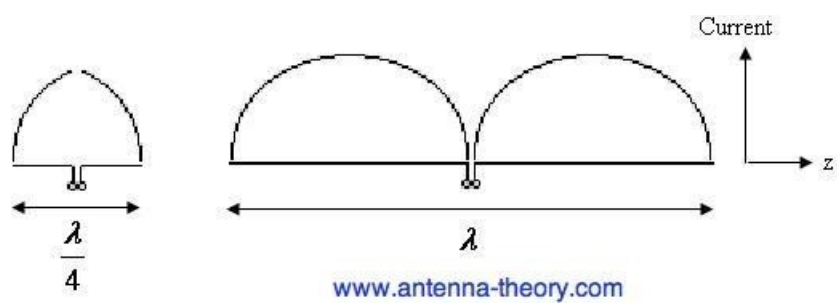


Figure 1. Current distributions on finite-length dipole antennas.

Before examining the fields radiated by a dipole antenna, consider the input impedance of a dipole as a function of its length, plotted in Figure 2 below. Note that the input impedance is specified as Z=R + jX, where R is the resistance and X is the reactance.

Radiation Patterns for Dipole Antennas

The far-fields from a dipole antenna of length *L* are given by:

$$E_\theta = \frac{j\eta I_0 e^{-jkr}}{2\pi r}\left[\frac{\cos\left(\frac{kL}{2}\cos\theta\right) - \cos\left(\frac{kL}{2}\right)}{\sin\theta}\right]$$

$$H_\phi = \frac{E_\theta}{\eta}$$

The normalized radiation patterns for dipole antennas of various lengths are shown in Figure 3.

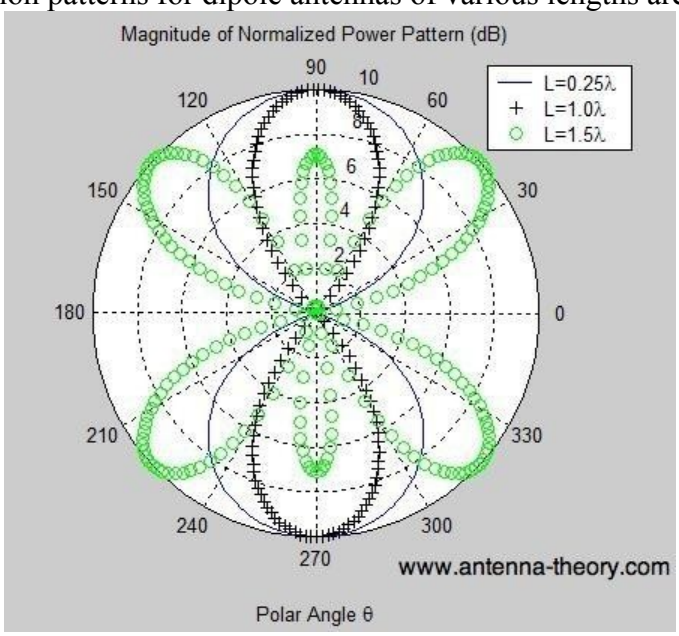


Figure 3. Normalized radiation patterns for dipole antennas of specified length.

The full-wavelength dipole antenna is more directional than the shorter quarter-wavelength dipole antenna. This is a typical result in antenna theory: it takes a larger antenna in general to increase directivity. However, the results are not always obvious. The 1.5-wavelength dipole pattern is also plotted in Figure 3. Note that this pattern is maximum at approximately +45 and -45 degrees.

The dipole antenna is symmetric when viewed azimuthally; as a result the radiation pattern is not a function of the azimuthal angle ϕ. Hence, the dipole antenna is an example of an omnidirectional antenna. Further, the E-field only has one vector component and consequently the fields are linearly polarized. When viewed in the x-y plane (for a dipole oriented along the z-axis), the E-field is in the -y direction, and consequently the dipole antenna is vertically polarized.

The 3D pattern for the 1-wavelength dipole antenna is shown in Figure 4. This pattern is similar to the pattern for the quarter- and half-wave dipole antenna.

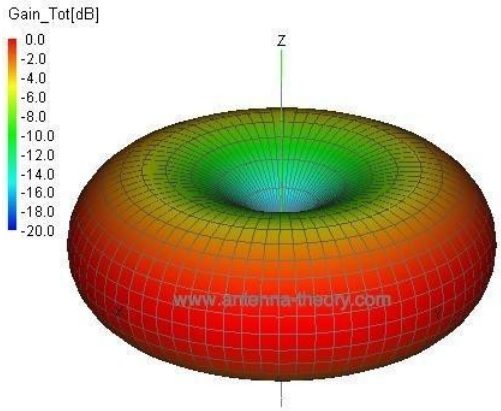


Figure 4. Normalized 3d radiation pattern for the 1-wavelength dipole antenna.

The 3D radiation pattern for the 1.5-wavelength dipole antenna is significantly different, and is shown in Figure 5.

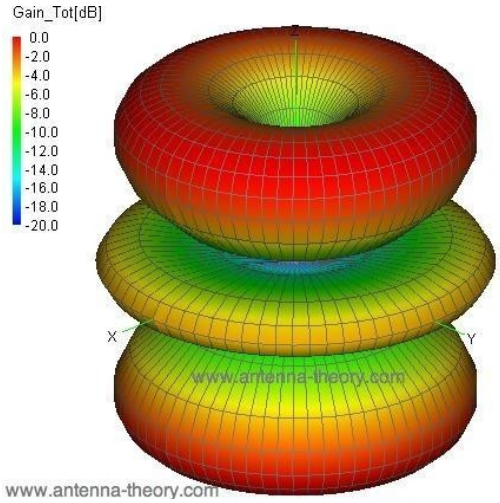


Figure 5. Normalized 3d radiation pattern for the 1.5-wavelength dipole antenna.

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Half wave Dipole

The **half-wave dipole antenna** is just a special case of the dipole antenna, but its important enough that it will have its own section. Note that the "half-wave" term means that the length of this dipole antenna is equal to a half-wavelength at the frequency of operation.

To make it crystal clear, if the antenna is to radiate at 600 MHz, what size should the half-wavelength dipole be?

One wavelength at 600 MHz is $\lambda = c / f$ = 0.5 meters. Hence, the half-wavelength dipole antenna's length is 0.25 meters.

The half-wave dipole antenna is as you may expect, a simple half-wavelength wire fed at the center as shown in Figure 1:

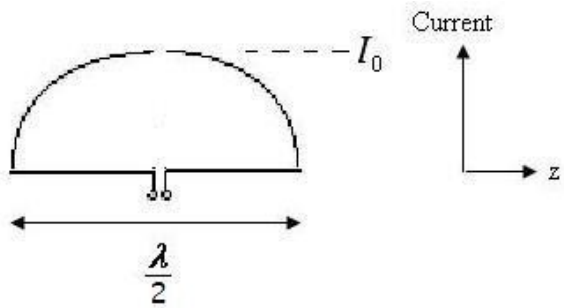


Figure 1. Electric Current on a half-wave dipole antenna.

The input impedance of the half-wavelength dipole antenna is given by Zin = 73 + j42.5 Ohms. The fields from the half-wave dipole antenna are given by:

$$E_\theta = \frac{j\eta I_0 e^{-jkr} \cos\left(\frac{\pi\cos\theta}{2}\right)}{2\pi r \sin\theta}$$

$$H_\phi = \frac{E_\theta}{\eta}$$

The directivity of a half-wave dipole antenna is 1.64 (2.15 dB). The HPBW is 78 degrees.

In viewing the impedance as a function of the dipole length in the section on dipole antennas, it can be noted that by reducing the length slightly the antenna can become resonant. If the dipole's length is reduced to 0.48 λ, the input impedance of the antenna becomes Zin = 70 Ohms, with no reactive component. This is a desirable property, and hence is often done in practice. The radiation pattern remains virtually the same.

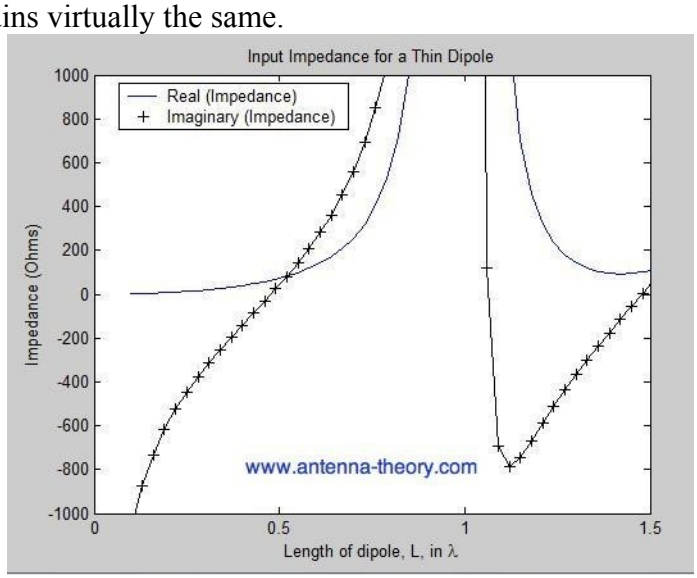


Figure 2. Input impedance as a function of the length (*L*) of a dipole antenna.

Note that for very small dipole antennas, the input impedance is capacitive, which means the impedance is dominated by a negative reactance value (and a relatively small real impedance or resistance). As the dipole gets larger, the input resistance increases, along with the reactance. At slightly less than 0.5 λ the antenna has zero imaginary component to the impedance (reactance X=0), and the antenna is said to be resonant.

If the dipole antenna's length becomes close to one wavelength, the input impedance becomes infinite. This wild change in input impedance can be understood by studying high frequency transmission line theory. As a simpler explanation, consider the one wavelength dipole shown in

Figure 1. If a voltage is applied to the terminals on the right antenna in Figure 1, the current distribution will be as shown. Since the current at the terminals is zero, the input impedance (given by Z=V/I) will necessarily be infinite. Consequently, infinite impedance occurs whenever the dipole antenna is an integer multiple of a wavelength.
In the next section, we'll consider the radiation pattern of dipole antennas.

- **Bandwidth**

The above length is valid if the dipole is very thin. In practice, dipoles are often made with fatter or thicker material, which tends to increase the bandwidth of the antenna. When this is the case, the resonant length reduces slightly depending on the thickness of the dipole, but will often be close to 0.47 λ.

- **Broad-Dipole**

A standard rule of thumb in antenna design is: an antenna can be made more broadband by increasing the volume it occupies. Hence, a dipole antenna can be made more broadband by increasing the radius *A* of the dipole.

As an example, method of moment simulations will be performed on dipoles of length 1.5 meters. At this length, the dipole is a half-wavelength long at 100 MHz. Three cases are considered:

- *A*=0.001 m = (1/3000th) of a wavelength at 100 MHz
- *A*=0.015 m = (1/100th) of a wavelength at 100 MHz
- *A*=0.05 m = (1/30th) of a wavelength at 100 MHz

The resulting S11 for each of these three cases is plotted versus frequency in Figure 1 (assuming matched to a 50 Ohm load).

- Mono Pole

A **monopole antenna** is one half of a dipole antenna, almost always mounted above some sort of ground plane. The case of a monopole antenna of length *L* mounted above an infinite ground plane is shown in Figure 1(a).

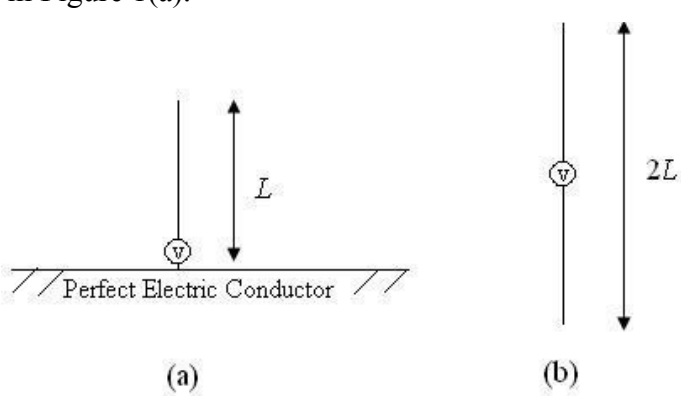


Figure 1. Monopole above a PEC (a), and the equivalent source in free space (b).

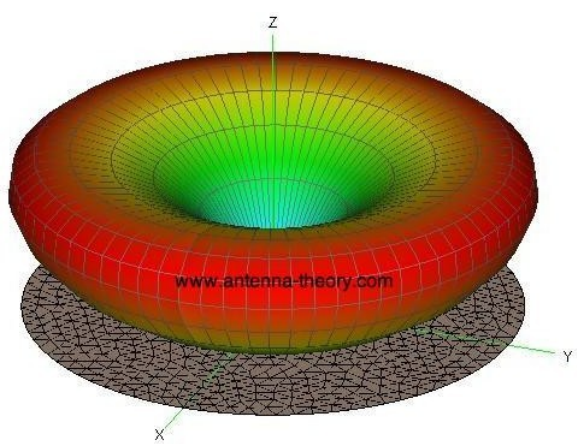
Using image theory, the fields above the ground plane can be found by using the equivalent source (antenna) in free space as shown in Figure 1(b). This is simply a dipole antenna of twice the length. The fields above the ground plane in Figure 1(a) are identical to the fields in Figure

1(b), which are known and presented in the dipole antenna section. The monopole antenna fields below the ground plane in Figure 1(a) are zero.

The radiation pattern of monopole antennas above a ground plane are also known from the dipole result. The only change that needs to be noted is that the impedance of a monopole antenna is one half of that of a full dipole antenna. For a quarter-wave monopole (L=0.25*λ), the impedance is half of that of a half-wave dipole, so Zin = 36.5 + j21.25 Ohms. This can be understood since only half the voltage is required to drive a monopole antenna to the same current as a dipole (think of a dipole as having +V/2 and -V/2 applied to its ends, whereas a monopole antenna only needs to apply +V/2 between the monopole antenna and the ground to drive the same current). Since Zin = V/I, the impedance of the monopole antenna is halved.
The directivity of a monopole antenna is directly related to that of a dipole antenna. If the directivity of a dipole of length *2L* has a directivity of D1 [decibels], then the directivity of a monopole antenna of length *L* will have a directivity of D1+3 [decibels]. That is, the directivity (in linear units) of a monopole antenna is twice the directivity of a dipole antenna of twice the length. The reason for this is simply because no radiation occurs below the ground plane; hence, the antenna is effectively twice as "directive".
Monopole antennas are half the size of their dipole counterparts, and hence are attractive when a smaller antenna is needed. Antennas on older cell phones were typically monopole antennas, with an infinite ground plane approximated by the shell (casing) of the phone.

Effects of a Finite Size Ground Plane on the Monopole Antenna

In practice, monopole antennas are used on finite-sized ground planes. This affects the properties of the monopole antennas, particularly the radiation pattern. The impedance of a monopole antenna is minimally affected by a finite-sized ground plane for ground planes of at least a few wavelengths in size around the monopole. However, the radiation pattern for the monopole antenna is strongly affected by a finite sized ground plane. The resulting radiation pattern radiates in a "skewed" direction, away from the horizontal plane. An example of the radiation pattern for a quarter-wavelength monopole antenna (oriented in the +z-direction) on a ground plane with a diameter of 3 wavelengths is shown in the following Figure:



Note that the resulting radiation pattern for this monopole antenna is still omnidirectional. However, the direction of peak-radiation has changed from the x-y plane to an angle elevated

from that plane. In general, the large the ground plane is, the lower this direction of maximum radiation; as the ground plane approaches infinite size, the radiation pattern approaches a maximum in the x-y plane.

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 1.

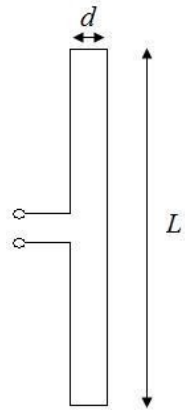


Figure 1. A Folded Dipole Antenna of length *L*.

Typically, the width *d* of the folded dipole antenna is much smaller than the length *L*.

Because the folded dipole forms a closed loop, one might expect the input impedance to depend on the input impedance of a short-circuited transmission line of length *L*. However, you can imagine the folded dipole antenna as two parallel short-circuited transmission lines of length *L*/2 (separated at the midpoint by the feed in Figure 1). 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*.

Letting Zd represent the impedance of a dipole antenna of length *L* and Zt represent the impedance of a transmission line impedance of length *L*/2, which is given by:

$$Z_t = jZ_0 \tan\frac{\beta L}{2}$$

The input impedance ZA of the folded dipole is given by:

$$Z_A = \frac{4Z_t Z_d}{Z_t + 2Z_d}$$

Folded Dipole Impedance

The folded dipole antenna is resonant and radiates well at odd integer multiples of a half-wavelength (0.5 λ, 1.5 λ...), when the antenna is fed in the center as shown in Figure 1. The input impedance of the folded dipole is higher than that for a regular dipole, as will be shown in the next section.

The folded dipole antenna can be made resonant at even multiples of a half-wavelength (1.0 λ , λ ,...) by offsetting the feed of the folded dipole in Figure 1 (closer to the top or bottom edge of the folded dipole).

Half-Wavelength Folded Dipole

The antenna impedance for a half-wavelength folded dipole antenna can be found from the above equation for ZA; the result is ZA=4*Zd. At resonance, the impedance of a half-wave dipole antenna is approximately 70 Ohms, so that the input impedance for a half-wave folded dipole antenna is roughly 280 Ohms.

Because the characteristic impedance of twin-lead transmission lines are roughly 300 Ohms, the folded dipole is often used when connecting to this type of line, for optimal power transfer. Hence, the half-wavelength folded dipole antenna is often used when larger antenna impedances (>100 Ohms) are needed.

The radiation pattern of half-wavelength folded dipoles have the same form as that of half-wavelength dipoles.

small loop antenna

The small loop antenna is a closed loop as shown in Figure 1. These antennas have low radiation resistance and high reactance, so that their impedance is difficult to match to a transmitter. As a result, these antennas are most often used as receive antennas, where impedance mismatch loss can be tolerated.

The radius is *a*, and is assumed to be much smaller than a wavelength ($a \ll \lambda$). The loop lies in the x-y plane.

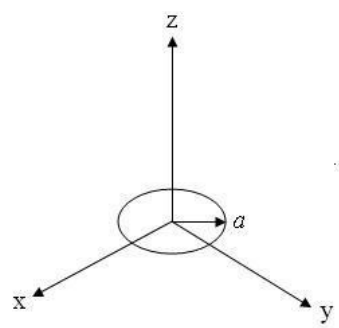


Figure 1. Small loop antenna.

Since the loop is electrically small, the current within the loop can be approximated as being constant along the loop, so that $I = I_0$.

The fields from a small circular loop are given by:

$$E_\phi = \frac{\eta k^2 a^2 I_0 e^{-jkr} \sin\theta}{4r}$$

$$H_\theta = \frac{-E_\phi}{\eta}$$

The variation of the pattern with direction is given by $\sin\theta$, so that the radiation pattern of a small loop antenna has the same power pattern as that of a short dipole. However, the fields of a small dipole have the E- and H- fields switched relative to that of a short dipole; the E-field is horizontally polarized in the x-y plane.

The small loop is often referred to as the dual of the dipole antenna, because if a small dipole had magnetic current flowing (as opposed to electric current as in a regular dipole), the fields would resemble that of a small loop.

While the short dipole has a capacitive impedance (imaginary part of impedance is negative), the impedance of a small loop is inductive (positive imaginary part). The radiation resistance (and ohmic loss resistance) can be increased by adding more turns to the loop. If there are *N* turns of a small loop antenna, each with a surface area *S* (we don't require the loop to be circular at this point), the radiation resistance for small loops can be approximated (in Ohms) by:

$$R_r = \left(\frac{177NS}{\lambda^2}\right)^2$$

For a small loop, the reactive component of the impedance can be determined by finding the inductance of the loop, which depends on its shape (then X=2*pi*f*L). For a circular loop with radius *a* and wire radius *p*, the reactive component of the impedance is given by:

$$X = 2\pi f\mu a\left[\ln\left(\frac{8a}{p}\right) - 1.75\right]$$

Small loops often have a low radiation resistance and a highly inductive component to their reactance. Hence, they are most often used as receive antennas. Exaples of their use include in pagers, and as field strength probes used in wireless measurements.

Half Wave Dipole Antenna:

Typically a dipole antenna is formed by two quarter wavelength conductors or elements placed back-to-back for a total length of $L = \lambda/2$. A standing wave on an element of length approximately $\lambda/4$ yields the greatest voltage differential, as one end of the element is at a node while the other is at an antinode of the wave. The larger the differential voltage, the greater the current between the elements.

Let us consider linear antennas of finite length and having negligible diameter. For such antennas, when fed at the center, a reasonably good approximation of the current is given by,

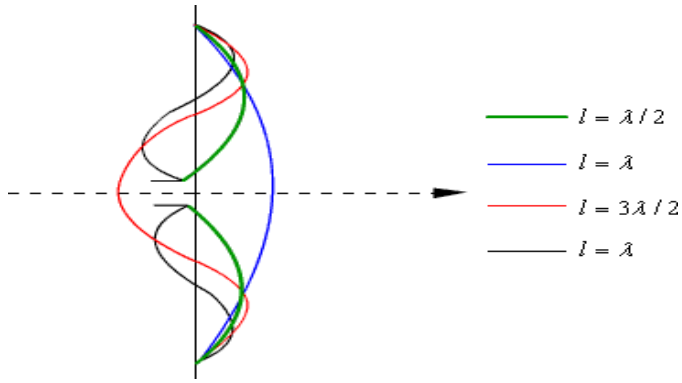


Fig 3.7: Current distribution on a center fed dipole antenna

$$I(z') = \begin{cases} I_0 \sin k_0\left(\frac{l}{2} - z'\right) & 0 \le z' \le \frac{l}{2} \\ I_0 \sin k_0\left(\frac{l}{2} + z'\right) & -\frac{l}{2} \le z' \le 0 \end{cases}$$

This distribution assumes that the current vanishes at the two end points i.e., $z' = \pm l/2$. The plots of current distribution are shown in the figure 3.7 for different 'l'.

This distribution assumes that the current vanishes at the two end points i.e. $z' = \pm l/2$. The plots of current distribution are shown in the figure 3.7 for different 's'.

For a half wave dipole, i.e., $l = \lambda/2$, the current distribution expressed as

$$I = I_0 \cos k_0 z' \qquad -\frac{\lambda_0}{4} \le z' \le \frac{\lambda_0}{4} \qquad \text{(3.39)}$$

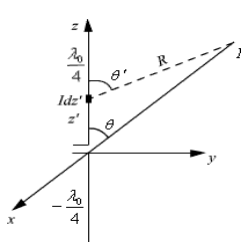


Fig 3.8(a): Half wave dipole

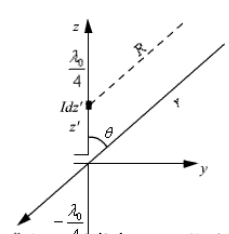


Fig 3.8(b): Far field approximation for half wave dipole

From equation (3.21) we can write

$$dA_z = \frac{\mu_0 I(z')\,dz'\,e^{-jk_0 R}}{4\pi R} \qquad \text{(3.40)}$$

From Fig 3.8(b), for the far field calculation, $R \approx r - z'\cos\theta$ for the phase variation and $R \approx r$ for amplitude term.

$$dA = \frac{\mu_0 I(z')\,dz'\,e^{-jk_0 R}}{\quad} e^{jk_0 z'\cos\theta}$$

$$a_3 \qquad 4\,r \qquad \text{..... (3.41)}$$

Substituting $I(z') = I_0 \cos k_0 z'$ from (3.39) to (3.41) we get

$$dA = a_3 \frac{\mu_0 I_0 e^{-jk_0 r}}{4\pi r} \cos k_0 z' e^{jk_0 z' \cos\theta} dz' \quad \text{.....(3.42)}$$

Therefore the vector potential for the half wave dipole can be written as:

$$A = a_3 \frac{\mu_0 I_0 e^{-jk_0 r}}{4\pi r} \int_{-\lambda/4}^{\lambda/4} \cos k_0 z' e^{jk_0 z' \cos\theta} dz'$$

$$A = a_3 \frac{\mu_0 I_0 e^{-jk_0 r}}{4\pi r} \frac{2\cos\left(\frac{k_0}{2}\cos\theta\right)}{k_0 \sin^2\theta} \quad \text{..... (7.43)}$$

From (3.37b),

$$E_\theta = -j\omega\mu \frac{\partial A_z}{\partial \theta} = -j\omega\mu \frac{\mu_0 I_0 e^{-jk_0 r}}{2\pi r} \frac{\cos\left(\frac{k_0}{2}\cos\theta\right)}{k_0 \sin^2\theta}$$

$$= -j\omega\mu \frac{\mu_0 I_0}{2\pi} \frac{e^{-jk_0 r}}{r} \frac{\cos\left(\frac{k_0}{2}\cos\theta\right)}{\sin^2\theta}$$

$$= -j\eta \frac{I_0 e^{-jk_0 r}}{2\pi r} \frac{\cos\left(\frac{k_0}{2}\cos\theta\right)}{\sin\theta} \quad \text{..... (3.44)}$$

Similarly from (3.37c)

$$E_\phi = 0 \quad \text{.....(3.45)}$$

and from (3.37e) and (3.37f)

$$H_\phi = \frac{jI_0 e^{-jk_0 r}}{2\pi r} \frac{\cos\left(\frac{k_0}{2}\cos\theta\right)}{\sin\theta} \quad \text{..... (3.46)}$$

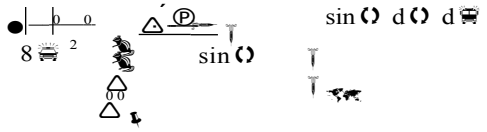
and $H_\theta = 0$ .....(3.47)

The radiated power can be computed as

$$P_r = \int_0^{2\pi}\int_0^{\pi} \frac{1}{2} \mathbf{E} \times \mathbf{H}^* \cdot r^2 \sin\theta \, d\theta \, d\phi$$

$$= \frac{I_0^2}{\ldots}$$

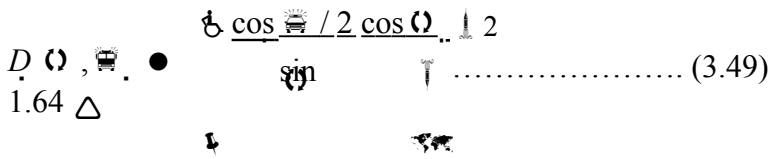
$$2 \quad \triangle \cos \quad 2$$



- 36.565 I_0 2.....(3.48)

Therefore the radiation resistance of the half wave dipole antenna is 36.565 2 = 73.13

Further, using Eqn(3.27) the directivity function for the dipole antenna can be written as



$D(\theta, \phi) = 1.64$ ……. (3.49)

Thus directivity of such dipole antenna is 1.04 as compared to 1.5 for an elementary dipole. The half power beam width in the E-plane can be found to be 780 as compared to 900 for a horizon dipole.

Quarter Wave Monopole Antenna:

A quarter wave monopole antenna is half of a dipole antenna placed over a grounded plane. The geometry of such antennas is shown in Fig 3.9(a) and equivalent half wave dipole is shown in fig 3.9(b).

If the ground plane is perfectly conducting, the monopole antenna shown in Fig 3.9(a) will be equivalent to a half wave dipole shown in Fig 3.9(b) taking image into account.
The radiation pattern above the grounded plane (in the upper hemisphere) will be same as that of a half wave dipole, however, the total radiated power will be half of that of a dipole since the field will be radiated only in the upper hemisphere.
An ideal quarter wave antenna mounted over a perfectly conducting ground plane has radiation resistance 36.56, half that of a dipole antenna, radiating in free space. The directivity of such antennas become double of that of dipole antennas.

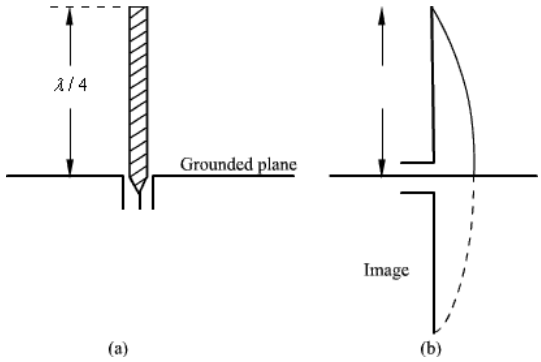


Fig 3.9 (a): Quarter wave monopole (b) Equivalent Half wave dipole

Quarter wave monopole antennas are often used as vehicle mounted antennas, the evhicle providing required ground plane for the antenna. For quarter-wave antennas mounted above earth, the poor conductivity of the soil results in excessive power loss from the induced amount in the soil.

The effect of poor ground conductivity is taken care of by installing a ground screen consisting of radial wires extending outward from the antenna base for a distance of Such arrangement is shown in Fig 3.10.

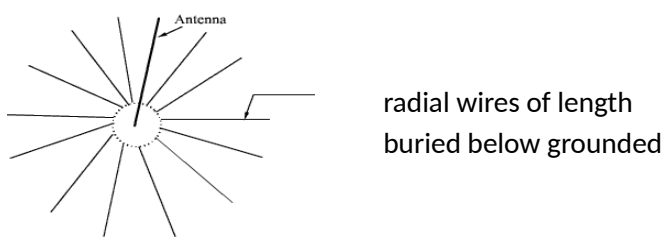


Fig 3.10: Grounded screen for improving performance of monopole antennas operating near earth surface..

Effective Length of a Half-wave Dipole

The effective length of an antenna is defined as the ratio of induced voltage at the terminal of the receiving antenna under open circuited condition to the incident electric field intensity.

i.e., Effective length, l_e = open circuited voltage/ Incident field strength

$$l_e = V / E$$

However, the included voltage _V' also depends on the effective aperture as,

$$A_e = (V_2 R_L) / \{[(R_A + R_L)^2 + (X_A + X_L)^2] P\}$$

Where,

R_L = Load resistance R_A

= Antenna resistance X_L

= Load reactance

X_A = Antenna reactance

P = Poynting vector.

$$V_2 = \{A_e [(R_A + R_L)^2 + (X_A + X_L)^2] P\} / R_L$$

Since,

$P = E_2/Z$, Where Z – Intrinsic impedance

$= 120\pi$

$$V_2 = \{A_e [(R_A + R_L)^2 + (X_A + X_L)^2] E_2\} / Z R_L$$

State reciprocity theorem for antennas. Prove that he self –impedance of an antenna in transmitting and receiving antenna are same?

Ans:Reciprocity Theorem

Statement

Reciprocity theorem states that when current I is applied at the terminals of antenna 1, an e.m.f E_{21} induces at terminals of antenna 2 and when current I applied at the terminals of antenna 2, an E_{12} induces at terminals of antenna 1, then $E_{12} = E_{21}$ provided $I_1 = I_2$

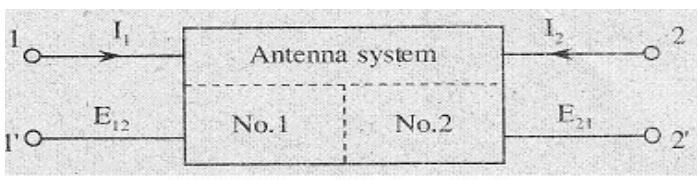


Fig 3.1 General Antenna System

Equality of Antenna Impedance

Consider, the two antennas separated with wide separation as shown below figure 3.2.

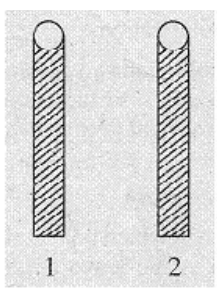


Fig 3.2 Two antennas 1 and 2 with a wide separation

The current distribution is same in case of transmitting and receiving antenna. Let antenna no. 1 is the transmitting antenna and antenna no.2 is the receiving antenna. The self impedance (Z_{11}) of transmitting antenna is given by,

$E_1 = Z_{11} I_1 + Z_{12} I_2$

Here,

Z_{11} = Self impedance of antenna1

Z_{12} = Mutual impedance between the two antennas.

Since the separation is more, mutual impedance (Z_{12}) is neglected,

$Z_{12} = 0$

$E_1 = Z_{11}I_1 + Z_{12}I_2$

$E_1 = Z_{11}I_1 + 0(I_2)$

The receiving antenna under open circuit and short circuit conditions are $Z_{11} = E_1/I_1$ shown below.

(a) Receiving Antenna under Open Circuit Condition

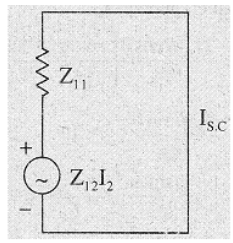


Fig 3.3 Reciving antenna under open circuit condition

$E_1 = Z_{11}I_1 + Z_{12}I_2$

When the receiving antenna is open circuited, current I_1 is zero

$E_1 = Z_{11}(0) + Z_{12}I_2$

$$E_{OC} = Z_{12}I_2$$

(b) Receiving Antenna under Short Circuit Condition

When the receiving antenna is short circuited, the voltage (E) will be zero.

$E_1 = Z_{11}I_1 + Z_{12}I_2$

$0 = Z_{11}I_{SC} + Z_{12}I_2$

$I_{SC} = - Z_{12} I_2 / Z_{11}$

From above, the term $Z_{12}I_2$ acts as a voltage source and Z_{11} as the self impedance.
Hence, impedance of the antenna is same whether it is used for transmission or reception

State the Maximum power transfer theorem and bring out their importance in antenna measurements?

Ans: Maximum Power Transfer Theorem :Statement

Maximum power transfer theorem states that, an antenna can radiated maximum power, when the terminal resistance, R_L of the antenna is same as that of finite source resistance, R_S.

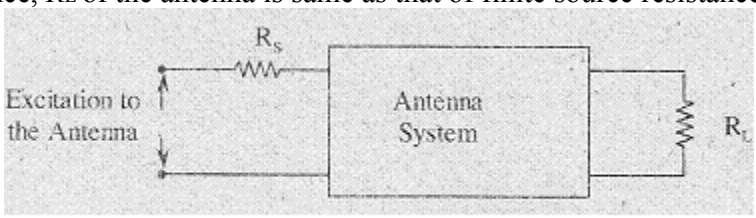


Fig 2.4.1 Maximum Power Transfer Theorem

This theorem applies to the maximum power, but not for maximum efficiency. If the antenna terminal resistance is made large than the resistance of the source, then the efficiency is more, since most of the power is generated at the terminals, but the overall power is lowered. If the internal source resistance is made larger than the terminal resistance then most of the power ends up being dissipated in the source.

Thus, the main use of maximum power transfer theorem for antennas is impedance matching i.e., maximum power transfer to and from an antenna occurs when the source or receiver impedance is same as that of antenna. But, when an antenna is not correctly matched internal reflections will occur.

Loop Antennas

LOOP ANTENNAS All antennas discussed so far have used radiating elements that were linear conductors. It is also possible to make antennas from conductors formed into closed loops. There are two broad categories of loop antennas:

1. Small loops which contain no morethan 0.086λ wavelength,s of wire
2. Large loops, which contain approximately 1 wavelength of wire.

Loop antennas have the same desirable characteristics as dipoles and monopoles in that they are inexpensive and simple to construct. Loop antennas come in a variety of shapes (circular, rectangular, elliptical, etc.) but the fundamental characteristics of the loop antenna radiation pattern (far field) are largely independent of the loop shape.

Just as the electrical length of the dipoles and monopoles effect the efficiency of these antennas, the electrical size of the loop (circumference) determines the efficiency of the loop antenna.

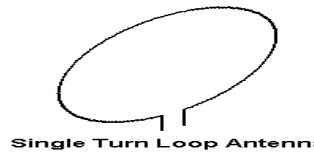
Loop antennas are usually classified as either electrically small or electrically large based on the circumference of the loop.

electrically small loop = circumference λ/10
electrically large loop - circumference λ

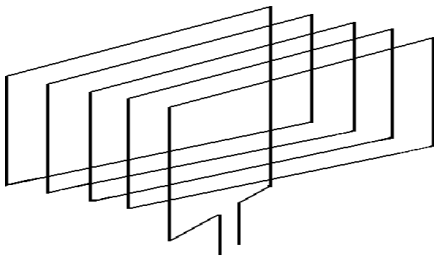
The electrically small loop antenna is the dual antenna to the electrically short dipole antenna when oriented as shown below. That is, the far-field electric field of a small loop antenna is identical to the far-field magnetic field of the short dipole antenna and the far-field magnetic field of a small loop antenna is identical to the far-field electric field of the short dipole antenna.

Small loops.

SMALL LOOP ANTENNAS A small loop antenna is one whose circumference contains no more than 0.085 wavelengths of wire. In such a short conductor, we may consider the current, at any moment in time to be constant. This is quite different from a dipole, whose current was a maximum at the feed point and zero at the ends of the antenna. The small loop antenna can consist of a single turn loop or a multi-turn loop as shown below:

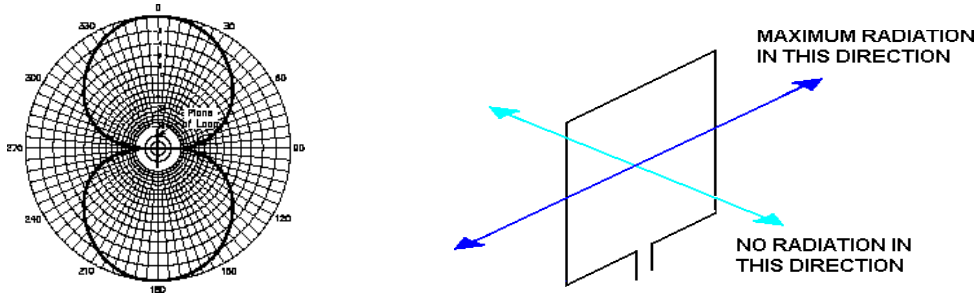


Single Turn Loop Antenna

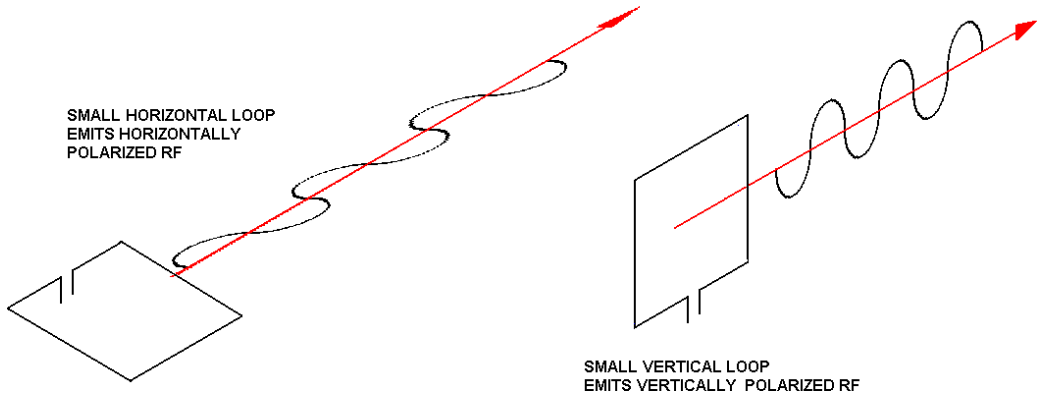


Multi-turn Loop Antenna

The radiation pattern of a small loop is very similar to a dipole. The figure below shows a 2-dimensional slice of the radiation pattern in a plane perpendicular to the plane of the loop. There is no radiation from a loop



When the loop is oriented vertically, the resulting radiation is vertically polarized and vice versa:



The input impedance of a small loop antenna is inductive, which makes sense, because the small loop antenna is actually just a large inductor. The real part of the input impedance is very small, on the order of 1 ohm, most of which is loss resistance in the conductor making up the loop. The actual radiation resistance may be 0.5 ohms or less. Because the radiation resistance is small compared to the loss resistance, the small loop antenna is not an efficient antenna and cannot be used for transmitting unless care is taken in its design and manufacture.

While the small loop antenna is not necessarily a good antenna, it makes a good receiving antenna, especially for LF and VLF. At these low frequencies, dipole antennas are too large to be easily constructed (in the LF range, a dipole's length ranges from approximately 1600 to 16,000 feet, and VLF dipoles can be up to 30 miles long!) making the small loop a good option. The small loop responds to the magnetic field component of the electromagnetic wave and is deaf to most man-made interference, which has a strong electric field. Thus the loop, although it is not efficient, picks up very little noise and can provide a better SNR than a dipole. It is possible to amplify the loop's output to a level comparable to what one might receive from a dipole.

When a small loop is used for receiving, its immunity and sensitivity may be improved by paralleling a capacitor across its output whose capacitance will bring the small loop to resonance at the desired receive frequency. Antennas of this type are used in AM radios as well as in LF and VLF direction finding equipment used on aircraft and boats.

The field pattern of a small circular loop of radius a may be determined very simple by considering a square loop of the same area, that is.

$d^2 = \pi a^2$ …(1)

where d = side length of square loop

It is assumed that the loop dimensions are small compared to the wavelength. It will be shown that the far-field patterns of circular and square loops of the same area are the same when the loops are small but different when they are large in terms of the wavelength.

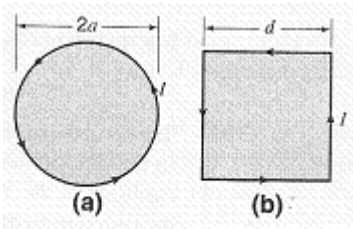


Fig 2.6.1 (a) Circular loop (b) square loop

If loop is oriented as in fig.2.6.2, its far electric field has only an E_Φ component. To find the far-field pattern in the yz plane, it is only necessary to consider two of the four small linear dipoles (2 and 4). A cross section through the loop in the yz plane is presented in Fig.2.6.3. Since the individual small dipoles 2 and 4 are nondirectional in the yz plane, fee field pattern of the loop in this plane is the same as that for two isotropic point sources. Thus,

$E_\Phi = -E_{\Phi 0}\, e^{j\psi/2} + E_{\Phi 0}\, e^{-j\psi/2}$ …(2)

Where $E_{\Phi 0}$ = electric field from individual dipole and

$\psi = (2\pi d/\lambda)\sin\theta = d_r \sin\theta$ …(3)

It follows that

$E_\Phi = -2j\, E_{\Phi 0} \sin(d_r \sin\theta/2)$ …(4)

The factor j in (4) indicates that the total field E_Φ is in phase quadrature with the field $E_{\Phi 0}$ be individual dipole.

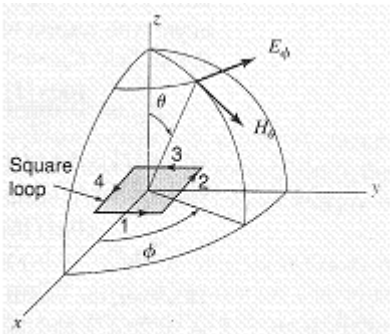


Fig 2.6.2 Relation of square loop to coordinates

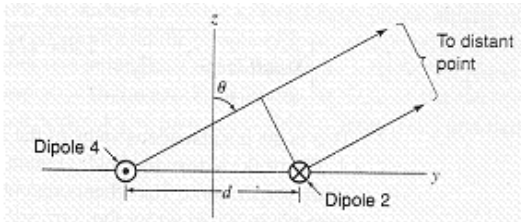


Fig 2.6.3 Construction for finding for field of dipoles 2 and dipole 4 of square loop.

However, the length L of the short dipole is the same as *d*, that is, L = *d*.

Small loop $E_\Phi = (120\pi^2 I \sin\theta A)/r\lambda^2$

This is the instantaneous value of the E_Φ, component of the far field of a small loop of area *A*. The peak value of the field is obtained by replacing [I] by I_0, where I_0 is the peak current in time on the loop.

Compare far fields of small loop and short dipole

It is of interest to compare the far-field expressions for a small loop with those for a short electric dipole. The comparison is made in table. The presence of the operator j in the dipole expressions and its absence in the loop equations indicate that the fields of the electric dipole and of the loop are in time-phase quadrature, the current I being in the same phase in both the dipole and loop. This quadrature relationship is a fundamental difference between fields of loops and dipoles.

Field	Electric Dipole	Loop
Electric	$E_\theta = (j\ 60\ \pi[\ I\ ]\ \sin\theta\ L)/(r\ \lambda)$	$E_\Phi = (120\ \pi^2[\ I\ ]\ \sin\theta\ A)\ /\ (r\ \lambda^2)$
Magnetic	$H_\Phi = (j\ [\ I\ ]\ \sin\theta\ L)\ /\ (2r\ \lambda)$	$H_\theta = (\pi\ [\ I\ ]\ \sin\theta\ A)\ /\ (r\ \lambda^2)$

Advantages
1. A small loop is generally used as magnetic dipole.
2. A loop antenna has directional properties whereas a simple vertical antenna not has the same.
3. The induced e.m.f around the loop must be equal to the difference between the two vertical sides only.
4. No e.m.f is produced in case of horizontal arms of a loop antenna.
5. The radiation pattern of the loop antenna does not depend upon the shape of the loop (for small loops).
6. The currents are at same magnitude and phase, throughout the loop.

Disadvantages
1. Transmission efficiency of the loop is very poor.
2. It is suitable for low and medium frequencies and not for high frequencies.
3. In loop antenna, the two nulls of the pattern result in 180° ambiguity.
4. Loop antennas used as direction finders are unable to distinguish between bearing of a distant transmitter and its reciprocal bearing.

far field patterns of loops of 0.1λ, λ and 3λ/2 diameter

The far field of loop antenna is, $E_\Phi = (\mu\omega[I]a\ I_1(\beta\ a\ \sin\theta))/2r$

$H_\theta = (\beta\ a[I]\ J_1(\beta\ a\ \sin\theta))/2r$

The above expression shows the far field pattern for loop of any size. The far field expressions E_Φ and H_θ as a function of θ is given by $J_1\ (C_\lambda \sin\theta)$

Here,

C_λ = Circumference of the loop

$C_\lambda = a\beta$

$\beta = (2\pi/\lambda)$

$C_\lambda = (2\pi/\lambda)\ a$

Far Field Patterns of Loops of 0.1λ, λ and 3λ/4 diameters

(i) Field patterns of 0.1λ

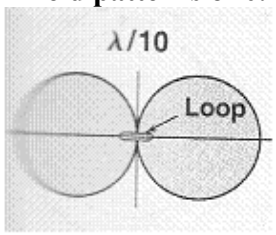


fig 9.1 Field pattern of 0.1λ

(ii) Field pattern of λ

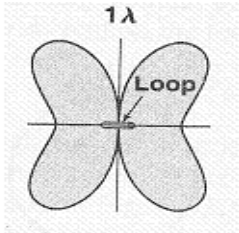


fig 9.2 Field Pattern of λ

(iii) Field pattern of 3λ/2 Diameter

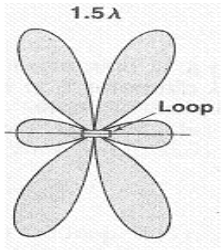
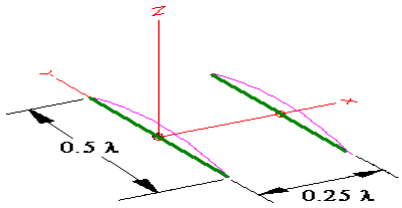


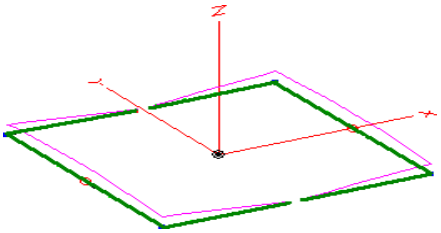
fig 9.3 Field pattern of 3λ/2

LARGE LOOP ANTENNAS A large loop antenna consists of approximately 1 wavelength of wire. The loop may be square, circular, triangular or any other shape. Because the loop is relatively long, the current distribution along the antenna is no longer constant, as it was for the small loop. As a result, the behavior of the large loop is unlike its smaller cousin.

The current distribution and radiation pattern of a large loop can be derived by folding two half wave dipoles and connecting them as shown in the diagrams below:

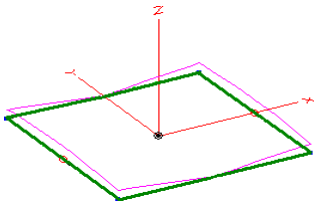
We begin with two λ/2 dipoles separated by λ/4. RF is fed in center of dipole. The resulting current distribution is shown below as a pink line. Note that the current is zero at the dipoles' ends,



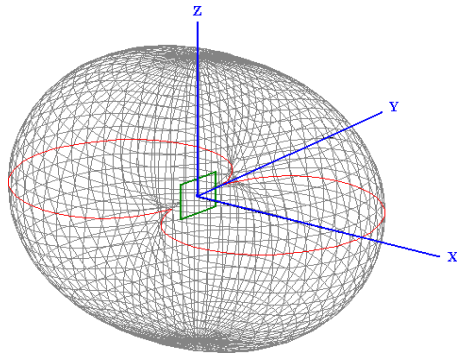


Now each dipole is folded in towards the other in a "U" shape as shown below. The current distribution has not changed - the antenna current is still zero at the ends.

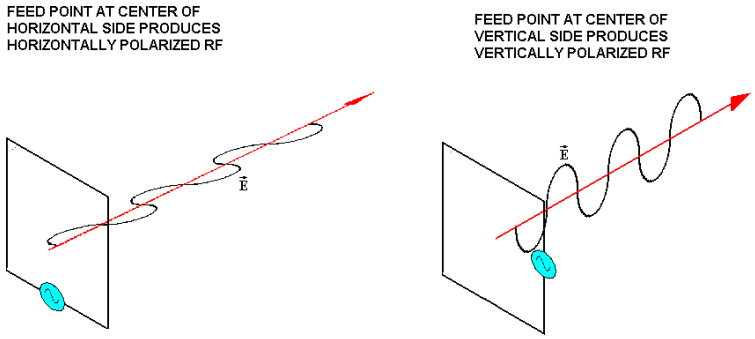
Since the current at the ends is zero, it would be OK to connect the ends to make a loop as shown below.



We have now created a square loop of wire whose circumference is 1 wavelength. From an electrical point of view, we have just shown that the large loop is equivalent to two bent dipole antennas. The radiation pattern of a loop antenna is shown below:



A horizontal slice of the radiation pattern in the XY plane is highlighted in red. It is similar to the figure-8 pattern of a dipole. It is possible to create either horizontally or vertically polarized radiation with a large loop antenna. The polarization is determined by the location of the feed point as shown below. If the feed point is in a horizontal side of the loop, the polarization is horizontal. If the feed point is in a vertical side of the loop, the polarization is vertical.



So far we have looked at square loop antennas. One of the interesting things about the large loop antenna is that the shape is not important. As long as the perimeter of the antenna is approximately 1 wavelength, the loop antenna will produce a radiation pattern very similar to the one shown above. The shape of the loop may be circular, square, triangular, rectangular, or any other polygonal shape. While the shape of the radiation pattern is not dependent on the shape of the loop, the gain of the loop does depend on the shape. In particular, the gain of the loop is dependent on the area enclosed by the wire. The greater the enclosed area, the greater the gain. The circular loop has the largest gain and the triangular loop has the least. The actual difference between the gain of the circular loop and triangular loop is less than 1 dB, and is usually unimportant. Loop antennas may be combined to form arrays in the same manner as dipoles. Arrays of loop antennas are called "quad arrays" because the loops are most often square. The most common type of quad array is a Yagi-Uda array using loops rather than dipoles as elements. This type of array is very useful at high elevations, where the combination of high voltage at the element tips of the dipoles in a standard Yagi array and the lower air pressure lead to corona discharge and erosion of the element . In fact, the first use of a quad array was by a broadcaster located in Quito, Ecuador (in the Andes Mountains) in the 1930's.

The input impedance of a loop depends on its shape. It ranges from approximately 100 ohms for a triangular loop to 130 ohms for a circular loop. Unlike the dipole, whose input impedance presents a good match to common 50 or 75 ohm transmission lines, the input impedance of a loop is not a good match and must be transformed to the appropriate impedance.

Helical Antenna

Helical antenna is useful at very high frequency and ultra high frequencies to provide circular polarization.

Consider a helical antenna as shown in figure 4.6.1.

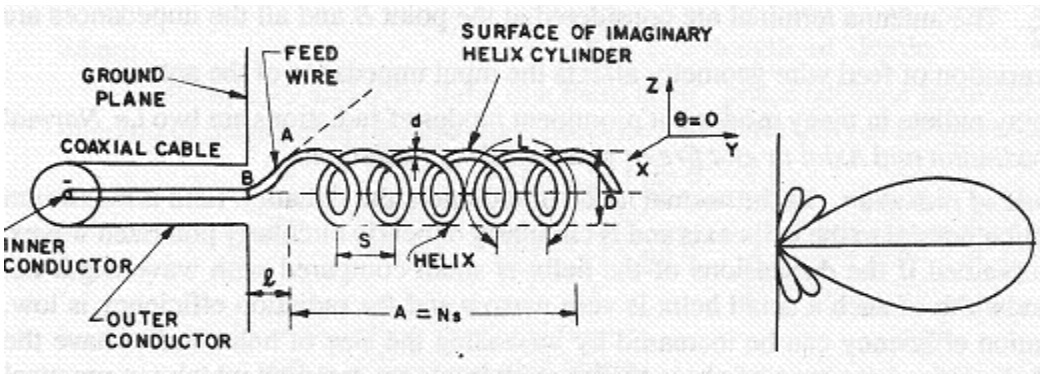


Fig 4.6.1 Helical antenna and its radiation pattern

Here helical antenna is connected between the coaxial cable and ground plane. Ground plane is made of radial and concentric conductors. The radiation characteristics of helical antenna depend upon the diameter (D) and spacing S.

In the above figure,

L = length of one turn = $\sqrt{S^2+(\pi D)^2}$

N = Number of turns

D = Diameter of helix = πD

α = Pitch angle = tan-1($S/\pi D$)

l = Distance between helix and ground plane.

Helical antenna is operated in two modes. They are,

1. Normal mode of radiation
2. Axial mode of radiation.

1. Normal mode of radiation

Normal mode of radiation characteristics is obtained when dimensions of helical antenna are very small compared to the operating wavelength. Here, the radiation field is maximum in the direction normal to the helical axis. In normal mode, bandwidth and efficiency are very low. The above factors can be increased, by increasing the antenna size. The radiation fields of helical antenna are similar to the loops and short dipoles. So, helical antenna is equivalent to the small loops and short dipoles connected in series.

We know that, general expression for far field in small loop is,

$E\Phi = \{120\ \pi 2[I]\ \sin\theta/r\}[A/\lambda 2]$

Where,

r = Distance

I = I0 sin ω(t-r/C) = Retarded current

A = Area of loop = $\pi D2/4$

D = Diameter

λ = Operating wavelength.

The performance of helical antenna is measured in terms of Axial Ratio *(AR).* Axial ratio is defined as the ratio of far fields of short dipole to the small loop.

Axial Ratio, AR =(EØ)/(EΦ)

2. Axial mode of radiation

Helical antenna is operated in axial mode when circumference C and spacing S are in the order of one wavelength. Here, maximum radiation field is along the helical axis and polarization is circular. In axial mode, pitch angle lies between 12° to 18° and beam width and antenna gain depends upon helix length NS.

General expression for terminal impedance is,

$$R = 140C/\lambda \text{ ohms}$$

Where,

R = Terminal impedance

C = Circumference.

In normal mode, beam width and radiation efficiency is very small. The above factors increased by using axial mode of radiation. Half power beam width in axial mode is,

$$HPBW = 52/C\sqrt{\lambda 3/NS} \text{ Degrees.}$$

Where,

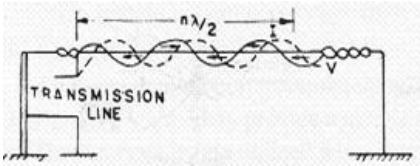
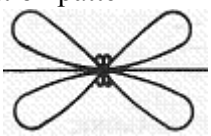
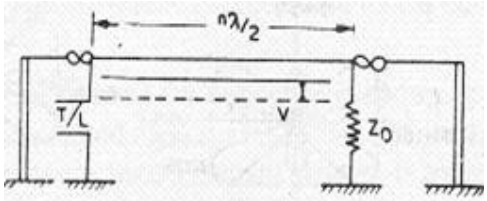
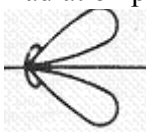
λ = Wavelength

C = Circumference

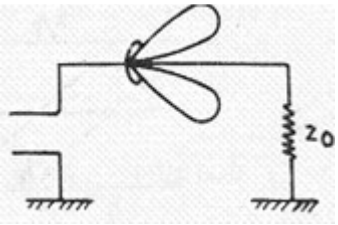
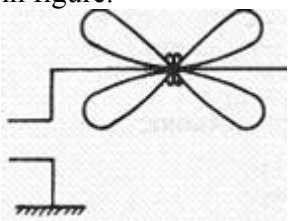
N = Number of turns

S = Spacing.

Axial Ratio, AR = 1 + 1/2N

Resonant Antenna	**Non-resonant Antenna**
1. These correspond to a resonant transmission line that is an exact number of half wave length long and is open at both ends. 2. Because of incident and reflected waves, standing waves exist. 3. The radiation pattern of this antenna is bi-directional. 4. These antennas are used for fixed frequency operations. 5. Resonant antenna  Long wire resonant antenna. 6.Radiation pattern  Bi-directional radiation pattern	1. These correspond to a transmission line that is exited at one end and terminated with characteristic impedance at the other end. 2. Due to the absence of reflected waves, standing waves do not exist. 3. The radiation pattern of this antenna is uni-directional. 4. These antennas are used for variable and wide frequency operations. 1. Non-resonant antenna  Long wire non-resonant antenna. 2. Radiation pattern  Uni-directional radiation pattern

Travelling Wave Antennas	**Standing Wave Antennas**
which standing waves does not exist. 2. Travelling wave antennas are also known as aperiodic or non-resonant	In standing wave antenna, standing wave exists. 2. Standing wave antennas are also known

antenna. 3. Reflected wave does not appear in travelling wave antennas. 4. Radiation pattern of travelling wave antenna is uni-directional. 5. Uni-directional pattern for ***n* = 4** is shown in figure. Here, ***n*** = Number of wave lengths.  6. Directivity is more. 3. The length of wire increases, major lobes get closer and narrower to the wire axis	as periodic or resonant antennas. 3. Reflected wave appears in standing wave antenna. 4. Radiation pattern of standing wave antenna is bi-directional. 5. Bi-directional pattern for ***n*** = 3 is shown in figure.  6. Directivity is less. 3. Length of wire does not depend upon the lobes

Narrow Band Antennas	**Wide Band Antennas**
1.Since, the bandwidth of receiving antenna is narrow, it is difficult for high-speed data communication. 2. These are bigger in size. 3. Because of the constitution of narrow band radio module, these are more expensive. 4. These antennas can realize stable long range communication. 5. These antennas lead to the high efficiency of radio wave use within same frequency band.	1.Since, the bandwidth of receiving antenna is very high, it is very easy for high-speed data communication. 2. These are small in size. 3. These are less expensive than narrow band antennas. 4. Because of large bandwidth, these are not suitable for long range communication. 5. These antennas lead to the less efficiency of radio wave use within same frequency band

UNIT III

VHF, UHF AND Microwave Antennas -II

REFLECTOR ANTENNAS

INTRODUCTION

The radiation pattern of a radiating antenna element is modified using reflectors. A simple example is that the backward radiation from an antenna may be eliminated with a large metallic plane sheet reflector. So, the desired characteristics may be produced by means of a large, suitably shaped, and illuminated reflector surface. The characteristics of antennas with sheet reflectors or their equivalent are considered in this chapter.

Some reflectors are illustrated in Figure 3.1. The arrangement in Figure 3.1a has a large, flat sheet reflector near a linear dipole antenna to reduce the backward radiation. With small spacing between the antenna and sheet this arrangement also yields an increase in substantial gain in the forward radiation. The desirable properties of the sheet reflector may be largely preserved with the reflector reduced in size as long as its size is greater than that of the antenna.

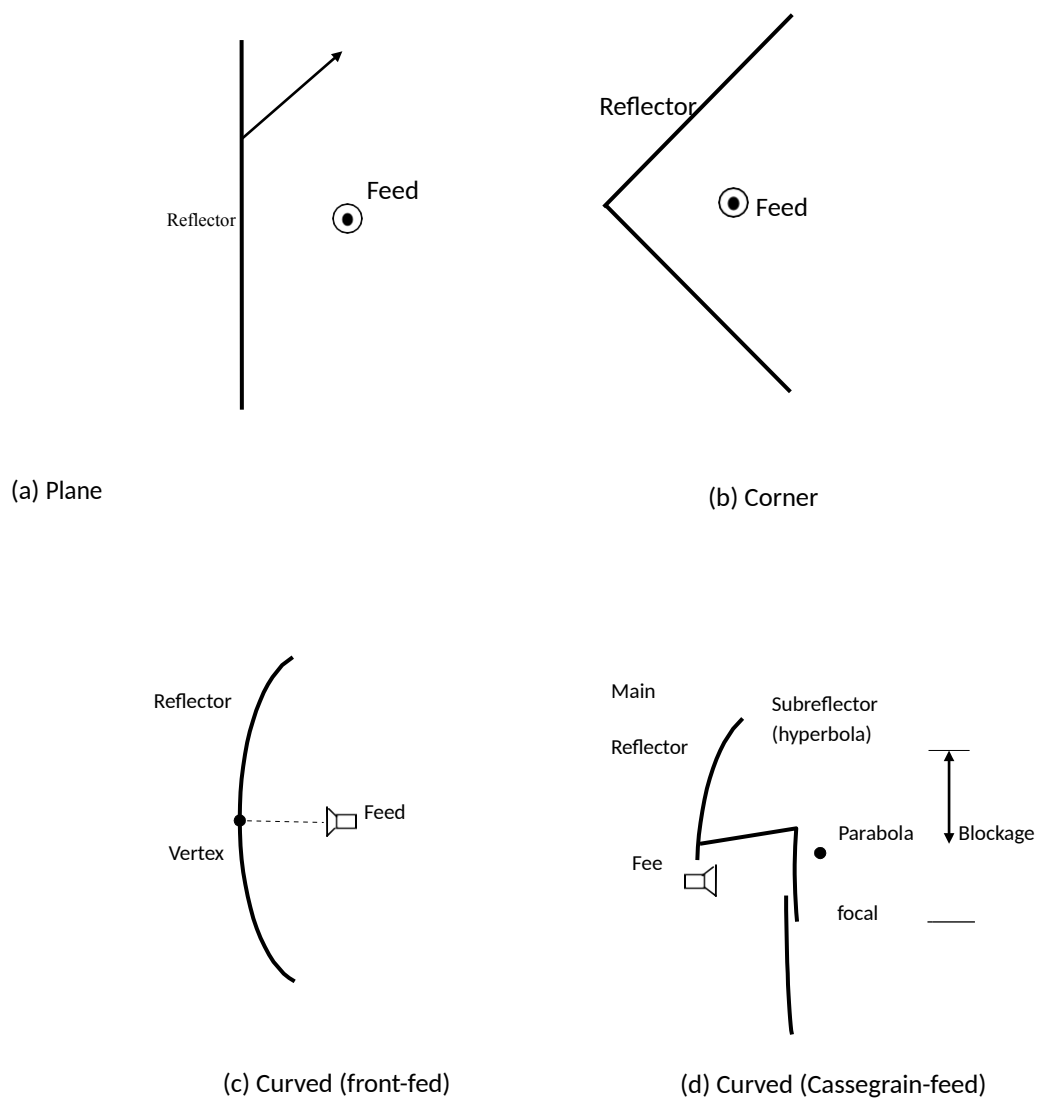


Figure 3.1 Some configurations of reflector antennas

With two flat sheets intersecting at an angle ✔ (<180☛) as in Figure 3.1b, a sharper radiation pattern than from a flat sheet reflector (✔ =180☛) can be obtained. This arrangement, called *corner reflector antenna*, is most practical where apertures of 1 or 2✢ are of convenient size. A corner reflector without an exciting antenna can be used as a *passive reflector* or target for radar waves. In this application the aperture may be many wavelengths, and the corner angle is *always* 90☛ . Reflectors with this angle have the property that an incidence wave is reflected back toward its source, the corner acting as a *retroreflector*.

When it is feasible to build antennas with apertures of many wavelengths, parabolic reflectors can be used to provide highly directional antennas. A parabolic reflector antenna is shown in Figure 3.1c. The parabola reflects the waves originating from a source at the focus into a parallel beam, the parabola transforming the curved wave front from the feed antenna at the focus into a plane wave front. A front fed and a cassegrain –feed parabolic reflectors are depicted in Figures 3.1c and d. Many other shapes of reflectors can be employed for special applications. For instance, with an antenna at one focus, the elliptical reflector produces a diverging beam with all reflected waves passing through the second focus of the ellipse. Examples of reflectors of other shapes are the hyperbolic and the spherical reflectors.

The plane sheet reflector, the corner reflector, the parabolic reflector and other reflectors are discussed in more detail in the following sections. In addition, feed systems, aperture blockage, aperture efficiency, diffraction, surface irregularities, gain and frequency-selective surfaces are considered.

PLANE REFLECTORS

Let an omnidirectional antenna is placed at a distance h above an infinite, flat, perfect electric conductor as shown in Figure 3.2. Power from the actual source is radiated in all directions in a manner determined by its unbounded medium directional properties. For an observation point p_1, there is a direct wave. In addition, a wave from the actual source radiated toward point R_1 of the interface undergoes a reflection. The direction is determined by the law of reflection ▪ ❍ $^{i}{}_{1}$● ❍ $^{r}{}_{1}$ ▪ which assures that the energy in homogeneous media travels in straight lines along the shortest paths. This wave will pass through the observation point p_1. By extending its actual path below the interface, it will seem to originate from a virtual source positioned a distance h below the boundary. For another observation point p_2 the point of reflection is R_2, but the virtual source is the same as before. The same is concluded for all other observation points above the interface.

The amount of reflection is generally determined by the respective constitutive parameters of the media below and above the interface. For a perfect electric conductor below the interface, the incidence wave is completely reflected and the field below the boundary is zero. According to the boundary conditions, the tangential components of the electric field must vanish at all points along the interface. Thus for an incident electric field with vertical polarization shown by the arrows, the polarization of the reflected waves must be as indicated in the figure to satisfy the boundary conditions.

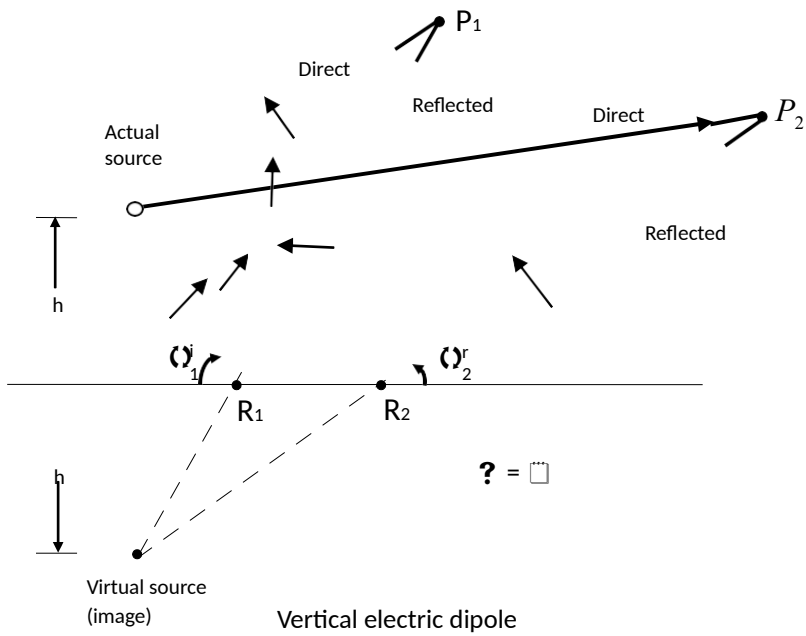
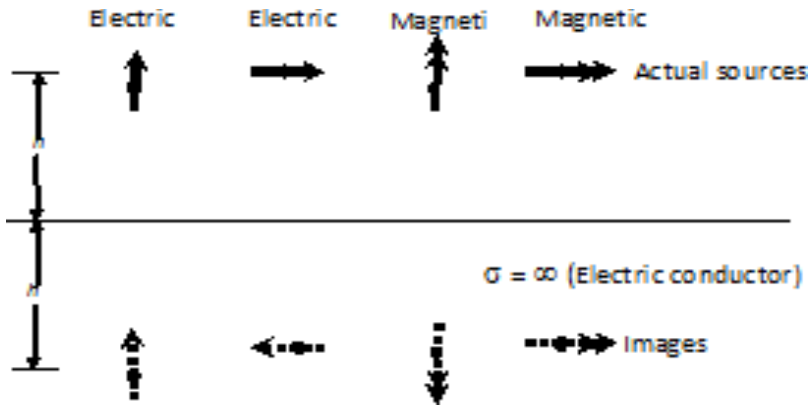


Figure 3.2 Antenna above an infinite, flat, perfect electric conductor.

For a vertical dipole, to excite the polarization of the reflected waves, the virtual source must also be vertical and with a polarity in the same direction as that of the actual source (thus a reflection coefficient of +1). Another orientation of the source will be to have the radiating element in a horizontal position, as shown in Figure 3.3. As shown in Figures 3.3, the virtual source (image) is also placed at a distance *h* below the interface. For horizontal polarized antenna, the image will have a 180° polarity difference relative to the actual source (thus a reflection coefficient of -1).

In addition to electric sources, artificial equivalent –magnetic || sources have been introduced to aid in the analyses of electromagnetic boundary value problems. Figure 3.3 displays the sources and their images for an electric plane conductor. The single arrow indicates an electric element and the double a magnetic one. The direction of the arrow identifies the polarity.



Electric conductor

Figure 3.3 Electric and magnetic sources and their images near electric conductors.

3.2.1 Vertical Electric Dipole

The analysis procedure for vertical and horizontal electric and magnetic elements near infinite electric plane conductors, using image theory, was illustrated graphically in the previous section. Based on the graphical model of Figure 3.2, the mathematical expressions for the fields of a vertical linear element near a perfect electric conductor will now be developed. For simplicity, only far-field observations will be considered.

Referring to the geometry of Figure 3.4(a), the far-zone direct component of the electric field of the infinitesimal dipole of length l, constant current I_0, and observation point P is given by

$$E_{\theta}^{d} = j\frac{\eta k I_0 l}{4\pi r_1}\sin\theta_1 e^{-jkr_1} \tag{3.1}$$

The reflected component can be accounted for by the introduction of the virtual source (image) as shown in Figure 3.4(a), and it can be written as

$$E_{\theta}^{r} = jR_v\frac{\eta k I_0 l}{4\pi r_2}\sin\theta_2 e^{-jkr_2} \tag{3.2}$$

or

$$E_{\theta}^{r} = j\frac{\eta k I_0 l}{4\pi r_2}\sin\theta_2 e^{-jkr_2} \tag{3.2a}$$

Since the reflection coefficient R_v is equal to unity.

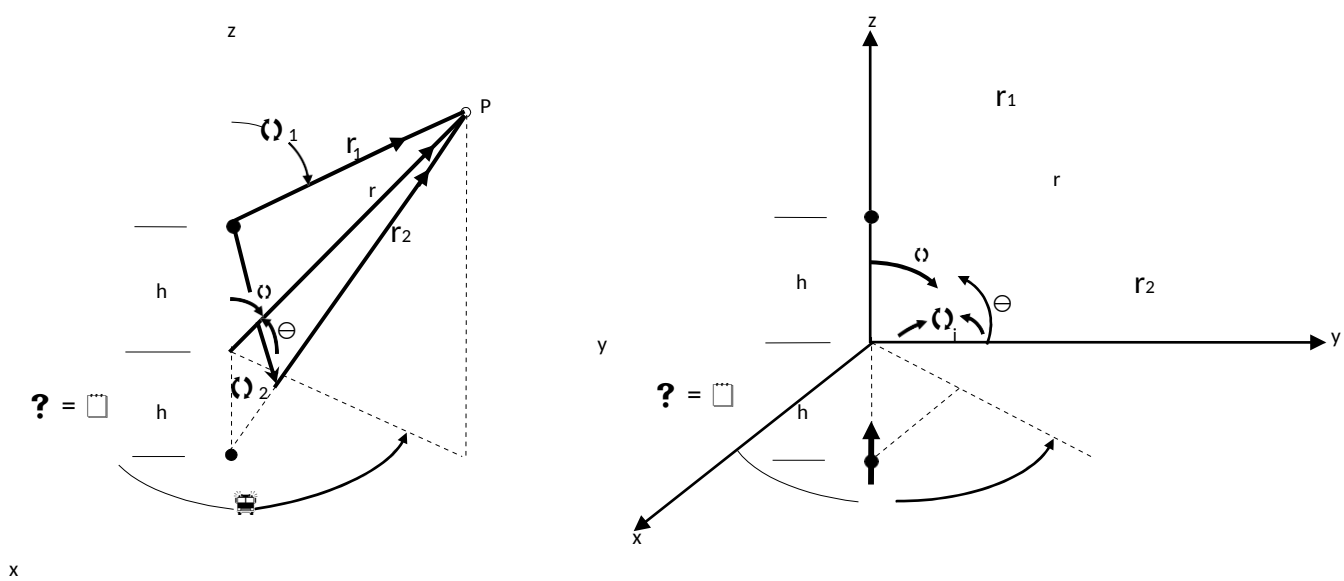


Figure 3.4 (a) Vertical electric dipole above infinite perfect electric conductor and its
(b) Far-field observations

The total field above the interface (z ≥ 0) is equal to the sum of the direct and reflected components as given by (3.1) and (3.2a). Since a field cannot exist inside a perfect electric conductor, it is equal to zero below the interface. To simplify the expression for the total electric field, it is referred to the origin of the coordinate system (z = 0).

In general, we can write that

$$r_1 = \left[r^2 + h^2 - 2rh\cos(\pi - \theta) \right]^{1/2} \tag{3.3a}$$

$$r_2 = \left[r^2 + h^2 - 2rh\cos(\theta) \right]^{1/2} \tag{3.3b}$$

For far-field observations $r \gg h$, (3.3a) and (3.3b) reduce using the binomial expansion to

$$r_1 \approx r - h\cos(\pi - \theta) \tag{3.4a}$$

$$r_2 \approx r - h\cos(\theta) \tag{3.4b}$$

As shown in Figure 3.4(b), geometrically (3.4a) and (3.4b) represent parallel lines. Since the amplitude variations are not as critical

$$r_1 \approx r_2 \approx r \qquad \text{for amplitude variations} \tag{3.5}$$

Using (3.4a)-(3.5), the sum of (3.1) and (3.2a) can be written as

$$E_\theta \approx \frac{j\eta k I_0 l e^{-jkr_2}}{4\pi r}\sin(\theta)\left[1 + \cos(\theta)\right]\cos(kh) \qquad z \geq 0 \tag{3.6}$$

$$E_\theta = 0 \qquad z \leq 0$$

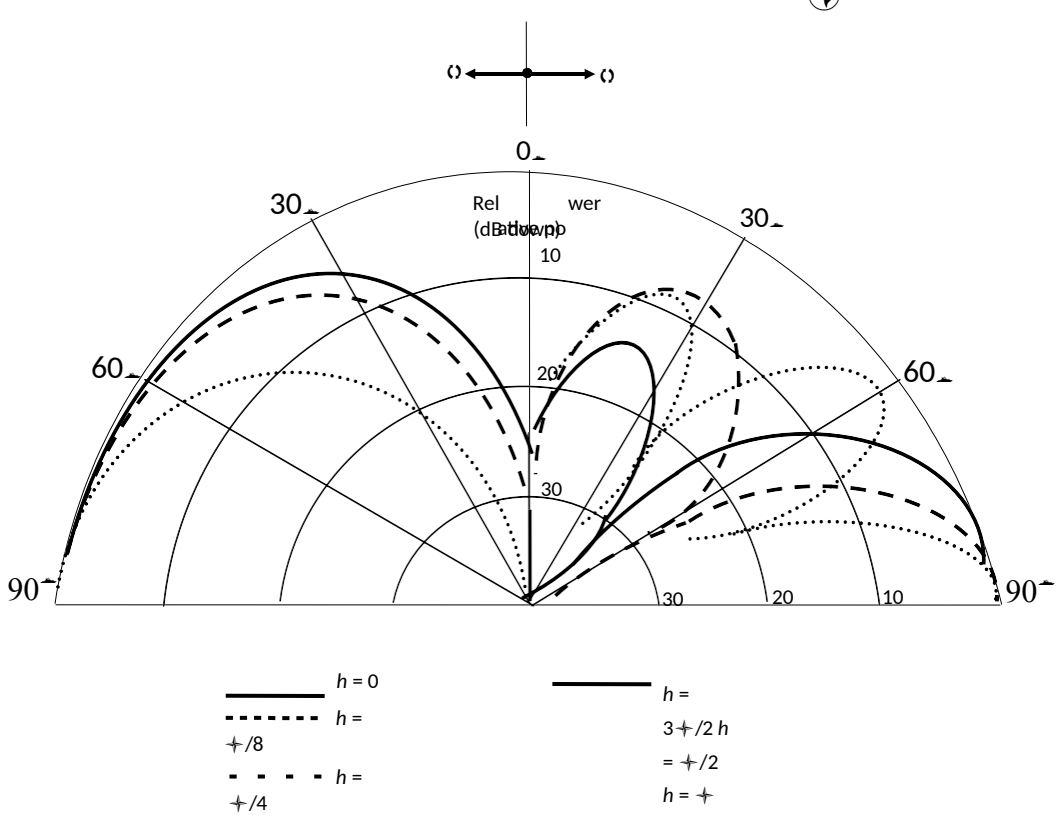


Figure 3.5 Elevation plane amplitude patterns of a vertical infinitesimal electric dipole for different heights above an infinite perfect electric conductor.

The shape and amplitude of the field is not only controlled by the field of the single element but also by the positioning of the element relative to the ground. To examine the field variations as a

function of the height h, the normalized (to 0 dB) power patterns for h = 0, ✦/8, ✦/4, 3✦/8, ✦/2, and ✦ are plotted in Figure 3.5. Because of symmetry, only half of each pattern is shown. For h >
✦/4 more minor lobes, in addition to the major ones, are formed. As h attains values greater than ✦, an even greater number of minor lobes are introduced. The introduction of the additional lobes is usually called scalloping. In general, the total number of lobes is equal to the integer that is close to

$$\text{number of lobes} \quad \frac{2h}{\lambda} \quad 1 \tag{3.7}$$

Since the total field of the antenna system is different from that of a singe element, the directivity and radiation resistance are also different. The directivity can be written as

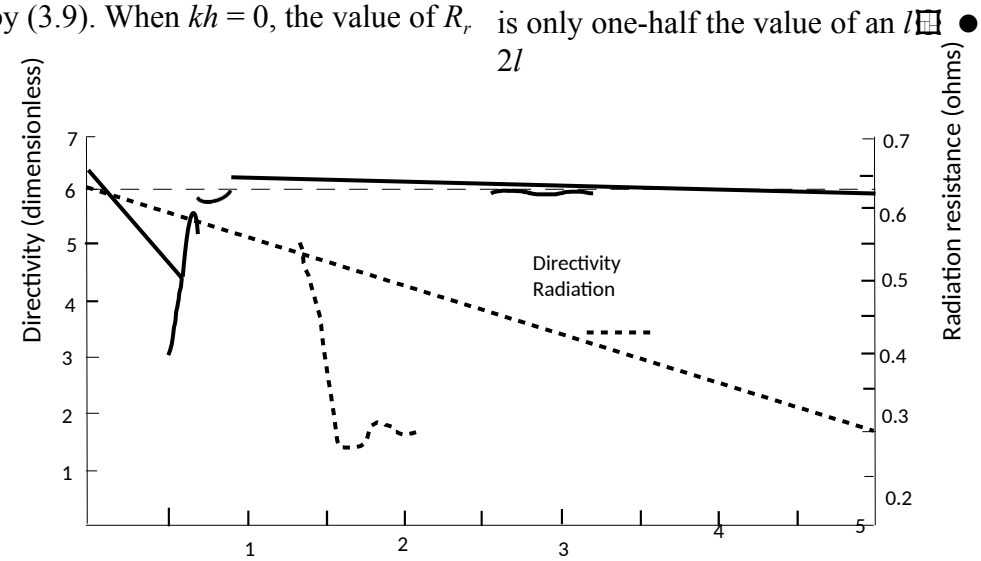
$$D_0 = \frac{4\pi U_{max}}{P_{rad}} = \frac{4\pi}{2\left[1 + \frac{\sin(2kh)}{(2kh)} - \frac{\cos(2kh)}{(2kh)^2} + \frac{\sin(2kh)}{(2kh)^3}\right]} \tag{3.8}$$

Whose value for kh = 0 is 3. The maximum value occurs when kh = 2.881 (h = 0.4585✦), and it is equal to 6.566 which is greater than four times that of an isolated dipole element (1.5). The directivity is displayed in Figure 3.6 for 0 ☒ h ☒ 5✦.

Similarly, from radiated power, the radiation resistance can be written as

$$R_r = \frac{2P_{rad}}{|I_0|^2} = \eta\frac{l}{2\pi}\left[1 + \frac{\sin(2kh)}{(2kh)} - \frac{\cos(2kh)}{(2kh)^2} + \frac{\sin(2kh)}{(2kh)^3}\right] \tag{3.9}$$

Whose value for kh ☒ ☐ is the same and for kh = 0 is twice that of the isolated element as given by (3.9). When kh = 0, the value of R_r is only one-half the value of an l☐ ● isolated element. 2l



Height (wavelengths)

Figure 3.6 Directivity and radiation resistance of a vertical infinitesimal electric dipole as a function of its height above an infinite perfect electric conductor.

3.2.2 Horizontal Electric Dipole

Another dipole configuration is when the linear element is placed horizontally relative to the infinite electric ground plane, as shown in Figure 3.3. The analysis procedure of this is identical to the one of the vertical dipole. Introducing an image and assuming far-field observations, as shown in Figure 3.7(a, b), the direct component can be written as

$$E_{\theta}^{d} = j\frac{kI_0 l e^{-jkr_1}}{4\pi r_1}\sin\theta \tag{3.10}$$

and the reflected one by

$$E_{\theta}^{r} = -jR_h\frac{kI_0 l e^{-jkr_2}}{4\pi r_2}\sin\theta \tag{3.11}$$

or

$$E_{\theta}^{r} = -j\frac{kI_0 l e^{-jkr_2}}{4\pi r_2}\sin\theta \tag{3.11a}$$

since the reflection coefficient is equal to $R_h = -1$.

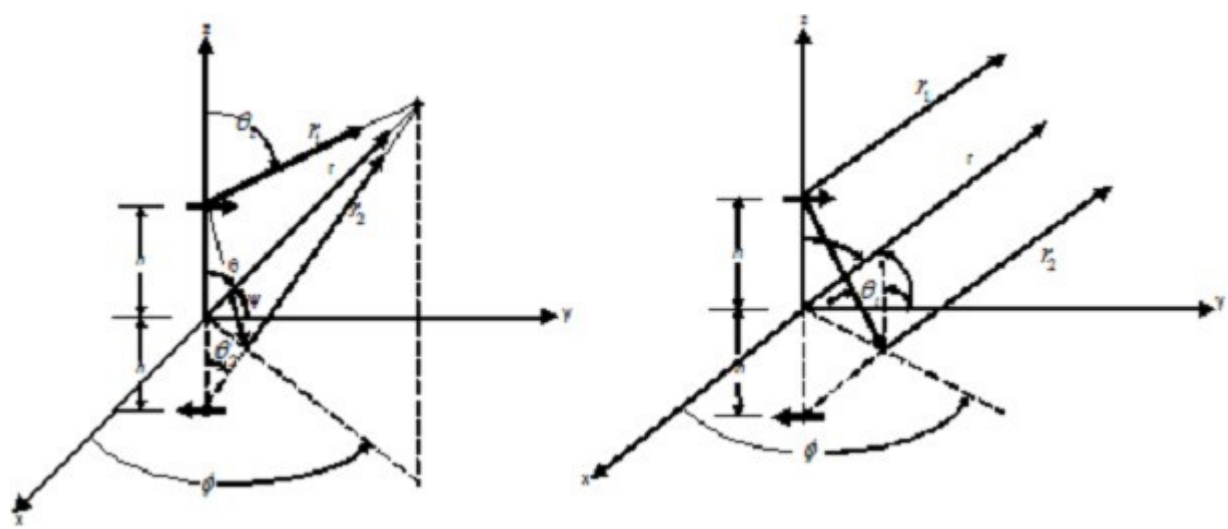


Figure 3.7 Horizontal electric dipole above an infinite perfect electric conductor and its far-field observations.

To find the angle θ, which is measured from the y-axis toward the observation point, we first form

$$\cos\theta = \hat{a}_y \cdot \hat{a}_r = \sqrt{1-\cos^2\theta}\,\sqrt{1-\sin^2\phi\sin^2\theta}\,\left(\hat{a}_x \sin\phi\cos\theta + \hat{a}_y \sin\phi\sin\theta + \hat{a}_z \cos\phi\right) \tag{3.12}$$

from which we find

$$\sin\theta = \sqrt{\quad}\sqrt{\quad}$$

(3.13)

Since for far-field observations

$$r_1 \approx \tilde{r} - h\cos\theta \qquad \text{for phase variations} \qquad (3.14a)$$

$$r_2 \approx \tilde{r} + h\cos\theta$$

$$r_1 \approx r_2 \approx r \qquad \text{for amplitude variations} \qquad (3.14b)$$

The total field, which is valid only above the ground plane $z \geq 0; 0 \leq \theta \leq \pi/2, 0 \leq \phi \leq 2\pi$, can be written as

$$E_\theta = E_\theta^d + E_\theta^r = j\frac{\eta I_e e^{-jkr_1}}{4\pi r}\left[\sqrt{1+\sin^2\theta\sin^2\phi}\;\; 2j\sin\left(kh\cos\theta\right)\right] \qquad (3.15)$$

Equation (3.15) again consists of the product of the field of a single isolated element placed symmetrically at the origin and a factor (within the brackets) known as the array factor. This again is the pattern multiplication rule.

To examine the variations of the total field as a function of the element height above the ground plane, the two dimensional elevation plane patterns (normalized to 0 dB) for $\phi = 90^\circ$ (y-z plane) when h = 0, $\lambda/8$, $\lambda/4$, $3\lambda/8$, $\lambda/2$, and λ are plotted in Figure 3.8. Since this antenna system is not symmetric, the azimuthal plane (x-y plane) pattern will not be isotropic.

As the height increases beyond one wavelength ($h > \lambda$), a larger number of lobes is again formed. The total number of lobes is equal to the integer that most closely is equal to

$$\text{number of lobes} \approx 2\left(\frac{h}{\lambda}\right) + 1 \qquad (3.16)$$

With unity being the smallest number.

The directivity can be written as

$$D_0 = \frac{4\pi U_{max}}{P_{rad}} = \frac{4\pi R(kh)}{R(kh)} \qquad kh \leq \pi/2 \;\; h \leq \lambda/4 \qquad (7.17a)$$

$$D_0 = \frac{4\sin^2(kh)}{R(kh)} \qquad kh \geq \pi/2 \;\; h \geq \lambda/4 \qquad (7.17b)$$

where

$$R(kh) = \left[2 - \frac{\sin(2kh)}{2kh} - \frac{\cos(2kh)}{(2kh)^2} + \frac{\sin(2kh)}{(2kh)^3}\right] \qquad (3.17c)$$

For small values of kh $kh \ll 0$, (3.17a) reduces to

$$D_0 = \frac{4\sin^2(kh)}{\left[2 - \frac{2}{8}\right]} = 7.5\left(\frac{\sin kh}{kh}\right)^2 \qquad (3.18)$$

$$\Delta \approx \frac{3}{15}(kh)^2$$

For h = 0 the element is shorted and it does not radiate. The directivity is plotted for $0 \le h \le 5\pi$ in Figure 3.9. It exhibits a maximum value of 3.5 for small values of h. A values of 6 occurs when $h \approx 0.725\, n/2\pi$, n = 1, 2, 3,…

The conductivity has a more pronounced effect on the impedance values, compared to those of the vertical dipole on input impedance. The values of the resistance and reactance approach, as the height increases, to the corresponding values of the isolated element of length λ/2 (73 ohms for the resistance and 42.5 ohms for the reactance).

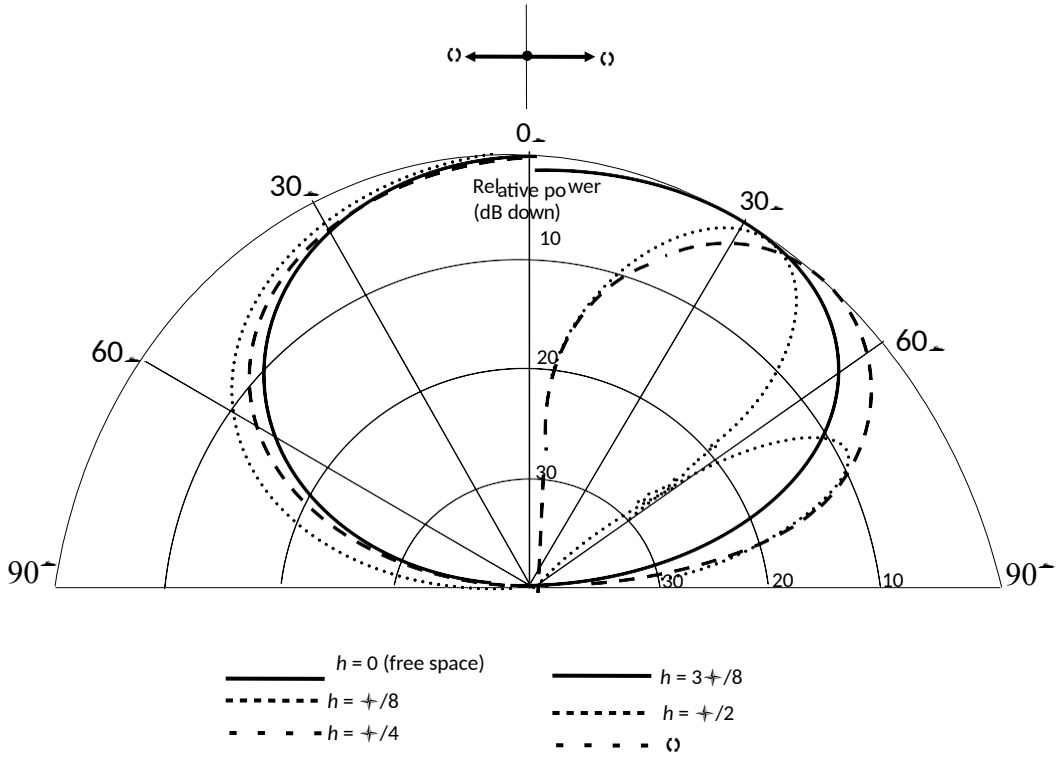


Figure 3.8 Elevation plane (φ = 90°) amplitude patterns of a horizontal infinitesimal electric dipole for different heights above an infinite perfect electric conductor.

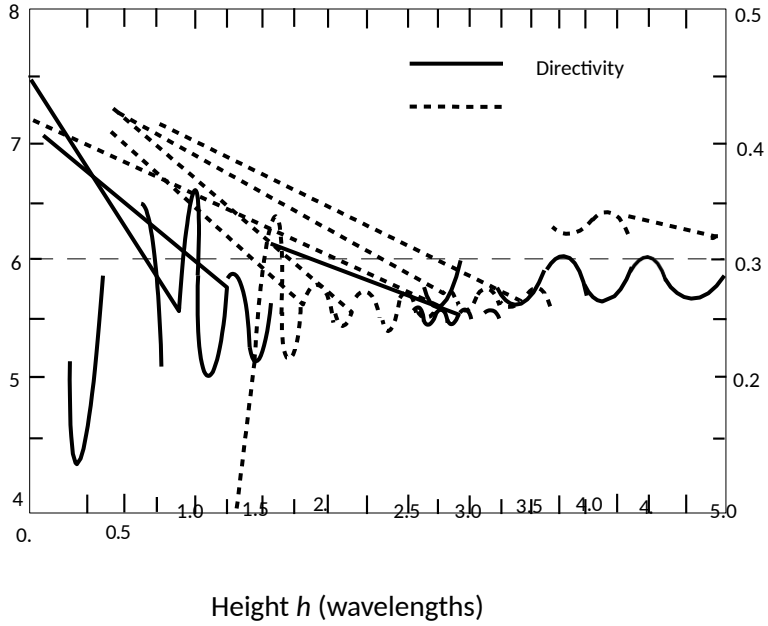


Figure 3.9 Radiation resistance and directivity of a horizontal infinitesimal electric dipole as a function of its height above an infinite perfect electric conductor.

CORNER REFLECTOR

For better collimination of the power in the forward directions, an arrangement can be made with two plane reflectors joined so as to form a corner, as shown in Figure 3.10 (a). This is known as the corner reflector. Because of its simplicity in construction, it has many unique applications. For example, if the reflector is used as a passive target for radar or communication applications, it will return the signal exactly in the same direction as it received it when its included angle is 90° . This is illustrated geometrically in Figure 3.10(b). Because of this unique feature, military ships and vehicles are designed with minimum sharp corners to reduce their detection by enemy radar.

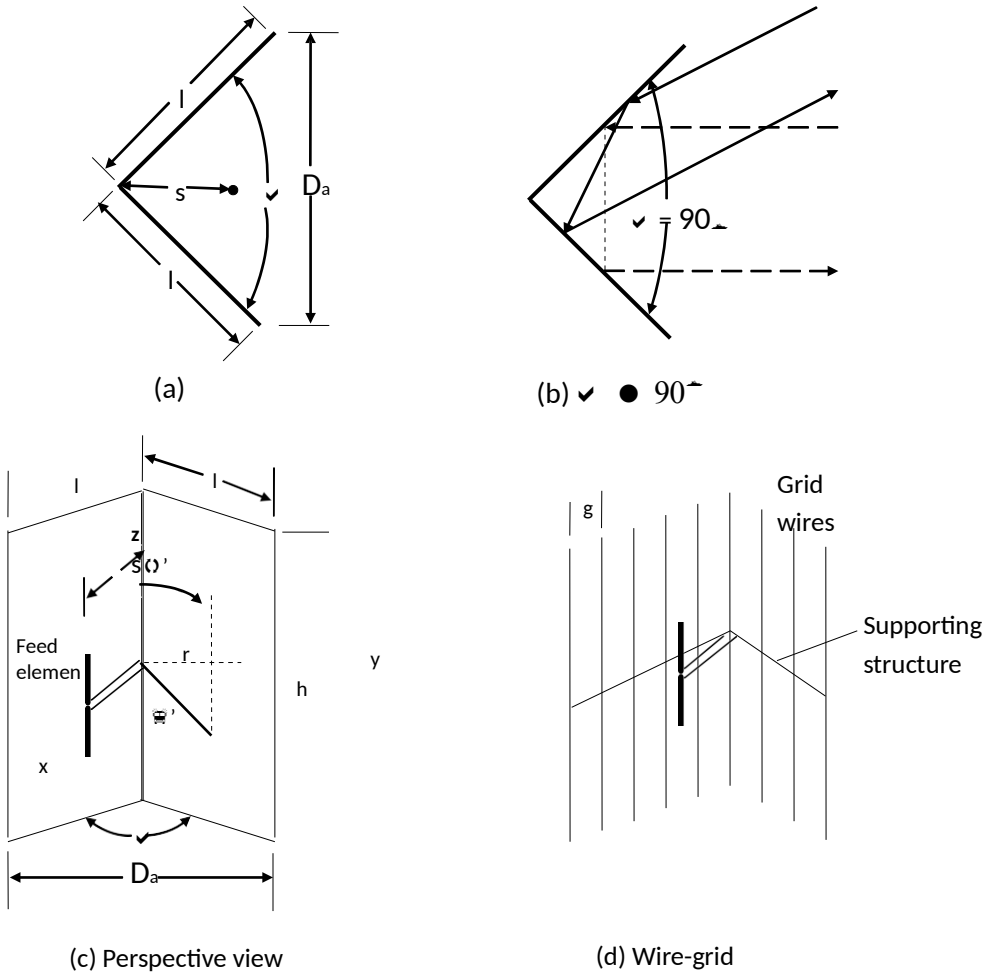


Figure 3.10 Side and perspective views of solid and wire-grid corner reflectors.

In most practical applications, the included angle formed by the plates is usually 90° ; however other angles are also used. To maintain a given system efficiency, the spacing between the vertex and the feed element must increase as the included angle of the reflector decreases, and vice-versa. For reflectors with infinite sides, the gain increases as the included angle between the planes decreases. This, however, may not be true for finite size plates. For simplicity, in this chapter it will be assumed that the plates themselves are infinite in extent (l = ∞). However, since

in practice the dimensions must be finite, guidelines on the size of aperture D_a , length (l), height (h) is given.

The feed element for a corner reflector is almost always a dipole or an array of collinear dipoles placed parallel to the vertex distance s away, as shown in in Figure 3.10 (c). Greater bandwidth is obtained when the feed elements are cylindrical or biconical dipoles instead of thin wires. In many applications, especially when the wavelength is large compared to tolerable physical dimensions, the surfaces of the corner reflector are frequently made of grid wires rather than solid sheet metal, as shown in Figure 3.10 (d). One of the reasons for doing that is to reduce wind resistance and overall system weight. The spacing g between wires is made a small fraction of a wavelength (usually g ☒ ✦/10). For wires that are parallel to the length of the dipole, as is the case for the arrangement of Figure 3.10(d), the reflectivity of the grid-wire surface is as good as that of a solid surface.

In practice, the aperture of the corner reflector (D_a) is usually made between one and two wavelengths: ✦ ■ D_a ■ 2✦ . The length of the sides of a 90☛ corner reflector is most commonly taken to be about twice the distance from the vertex to the feed: l ॥ 2s . For reflectors with smaller included angles, the sides are made larger. The feed-to-vertex distance (s) is usually taken to be between ✦/3 and 2✦/3 (✦/3 < s < 2✦/3). For each reflector, there is an optimum feed- to-vertex spacing. If the spacing becomes too small, the radiation resistance decreases and becomes comparable to the loss resistance of the system which leads to an inefficient antenna. For very large spacing, the system produces undesirable multiple lobes, and it loses its directional characteristics. It has been experimentally observed that increasing the size of the sides does not greatly affect the beamwidth and directivity, but it increases the bandwidth and radiation resistance. The main lobe is somewhat broader for reflectors with finite sides compared to that of infinite dimensions. The height (h) of the reflector is usually taken to be about 1.2 to 1.5 times greater than the total length of the feed element, in order to reduce radiation toward the back region from the ends.

The analysis for the field radiated by a source in the presence of a corner reflector is facilitated when the included angle (✔) of the reflector is ✔ = ✉/n, where n is an integer (✔ = ✉, ✉/2, ✉/3,

✉/4, etc.). For these cases: ✔ ● 180☛,90☛, 60☛, 45☛, *etc*: it is possible to find a system of images, which when properly placed in the absence of the reflector plates, form an array that yields the same field within the space formed by the reflector plates as the actual system. The number of images, polarity, and position is controlled by included angle and the polarization of the feed element. The geometrical and electrical arrangement of the images for corner reflectors with included angles of 90☛ , 60☛ , 45☛ *and* 30☛ and feed with perpendicular polarization are displayed in

Figure 4.11.

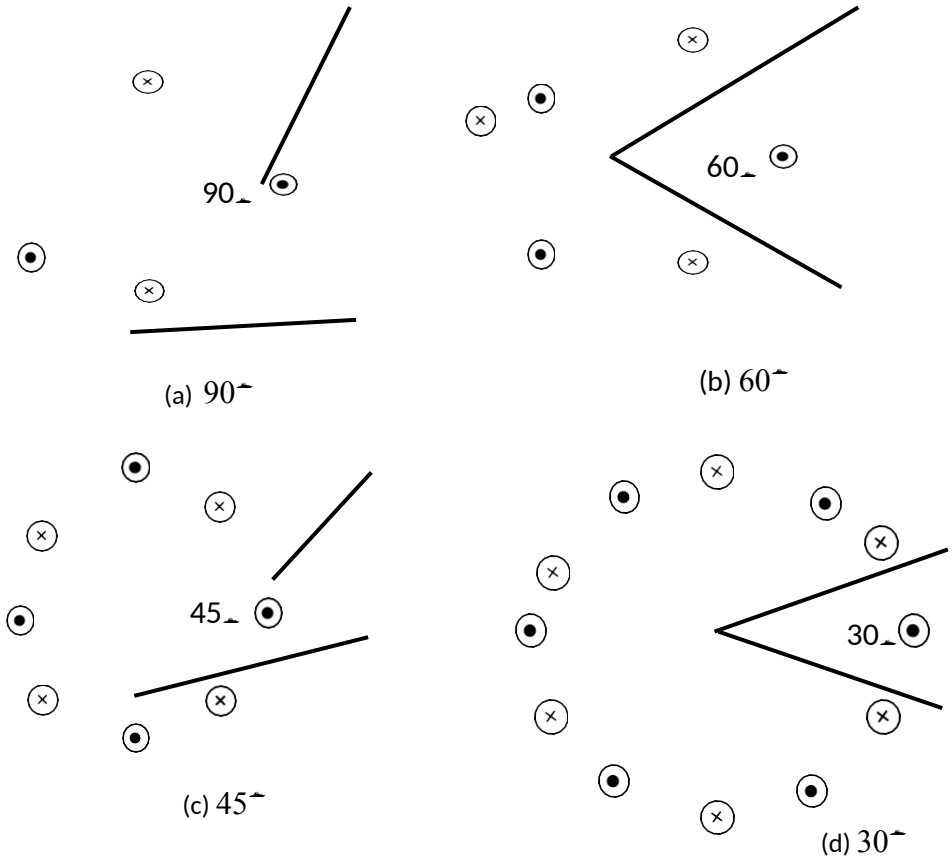


Figure3.11 Corner reflectors and their images (with perpendicularly polarized feeds) for angles of 90 , 60 , 45 *and* 30 .

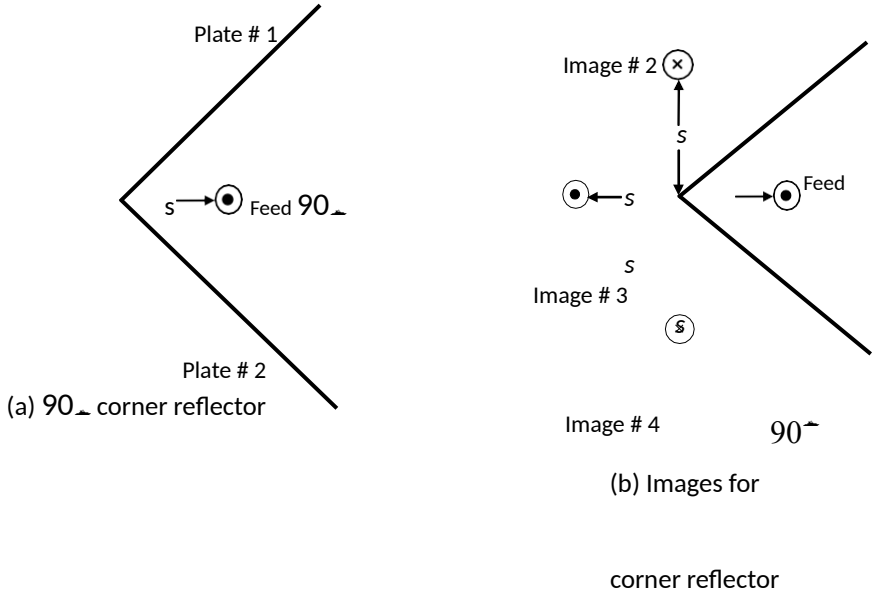


Figure4.12 Geometrical placement and electrical polarity of images for a 90 corner reflector with a parallel polarized feed.

The procedure for finding the number, location, and polarity of the images is demonstrated graphically in Figure 3.12 for a corner reflector with a 90° included angle. It is assumed that the feed element is a linear dipole placed parallel to the vertex. A similar procedure can be followed for all other reflectors with an included angle of π / n, where n is an integer.

4.3.1 A 90° Corner Reflector

For the corner reflector with an included angle of 90°, the total field of the system can be derived by summing the contributions from the feed and its images. Thus

$$E(r,\theta,\phi) = E_1(r_1,\theta,\phi) + E_2(r_2,\theta,\phi) + E_3(r_3,\theta,\phi) + E_4(r_4,\theta,\phi) \tag{3.18}$$

In the far-zone, the normalized scalar field can be written as

$$E(r,\theta,\phi) = f(\theta,\phi)\frac{e^{-jkr_1}}{r_1} + f(\theta,\phi)\frac{e^{-jkr_2}}{r_2} + f(\theta,\phi)\frac{e^{-jkr_3}}{r_3} + f(\theta,\phi)\frac{e^{-jkr_4}}{r_4}$$

$$E(r,\theta,\phi) = \left[e^{+jks\cos\Theta_1} + e^{+jks\cos\Theta_2} + e^{+jks\cos\Theta_3} + e^{+jks\cos\Theta_4}\right] f(\theta,\phi)\frac{e^{-jkr}}{r} \tag{3.19}$$

where

$$\cos\Theta_1 = \hat{a}_x \cdot \hat{a}_r = \sin\theta\cos\phi \tag{3.19a}$$

$$\cos\Theta_2 = \hat{a}_y \cdot \hat{a}_r = \sin\theta\sin\phi \tag{3.19b}$$

$$\cos\Theta_3 = -\hat{a}_x \cdot \hat{a}_r = -\sin\theta\cos\phi \tag{3.19c}$$

$$\cos\Theta_4 = -\hat{a}_y \cdot \hat{a}_r = -\sin\theta\sin\phi \tag{3.19d}$$

since $\hat{a}_r = \hat{a}_x \sin\theta\cos\phi + \hat{a}_y \sin\theta\sin\phi + \hat{a}_z \cos\theta$. Equation (3.18) can also be written, using (4.19a)-(4.19d), as

$$E(r,\theta,\phi) = 2[\cos(ks\sin\theta\cos\phi) + \cos(ks\sin\theta\sin\phi)]\, f(\theta,\phi)\frac{e^{-jkr}}{r} \tag{3.20}$$

where $0 \le \phi \le 2\pi$, $0 \le \theta \le \pi/2$ $\tag{3.21a}$

Letting the field of a single isolated (radiating in free-space) element to be

$$e^{-jkr}$$

$$E_0 \quad f(\theta, \phi) \frac{}{r} \qquad (3.22)$$

(4.20) can be rewritten as

$$\frac{E}{E_0} = AF(\theta, \phi) = 2\left[\cos\left(\frac{ks}{2}\sin\theta\cos\phi\right) - \cos\left(\frac{ks}{2}\sin\theta\sin\phi\right)\right] \qquad (3.23)$$

Equation (4.23) represents not only the ratio of the total field to that of an isolated element at the origin but also the array factor of the entire reflector system. In the azimuthal plane ($\theta = \pi/2$), (4.23) reduces to

$$\frac{E}{E_0} = AF(\phi) = \phi / 2, \phi = 2 \cos ks \cos \phi \cos ks \sin \phi \quad (3.24)$$

The normalized patterns for a $\psi = 90^\circ$ corner reflector for spacings of $s = 0.1\lambda$, 0.7λ, 0.8λ, 0.9λ, and 1.0λ is shown in Figure 3.13. It is evident that for the small spacing the pattern consists of a single major lobe whereas multiple lobes appear for the larger spacings ($s > 0.7\lambda$). For $s = \lambda$ the pattern exhibits two lobes separated by a null along the $\phi = 0^\circ$ axis.

Another parameter of performance for the corner reflector is the field strength along the symmetry axis $(\phi = 90^\circ, \phi = 0^\circ)$ as a function of feed-to-vertex distance s. The normalized (relative to the field of a single isolated element) absolute field strength peaks when $s = 0.5\lambda$, and it is equal to 4. The field is also periodic with a period of $\Delta s / \lambda = 1$.

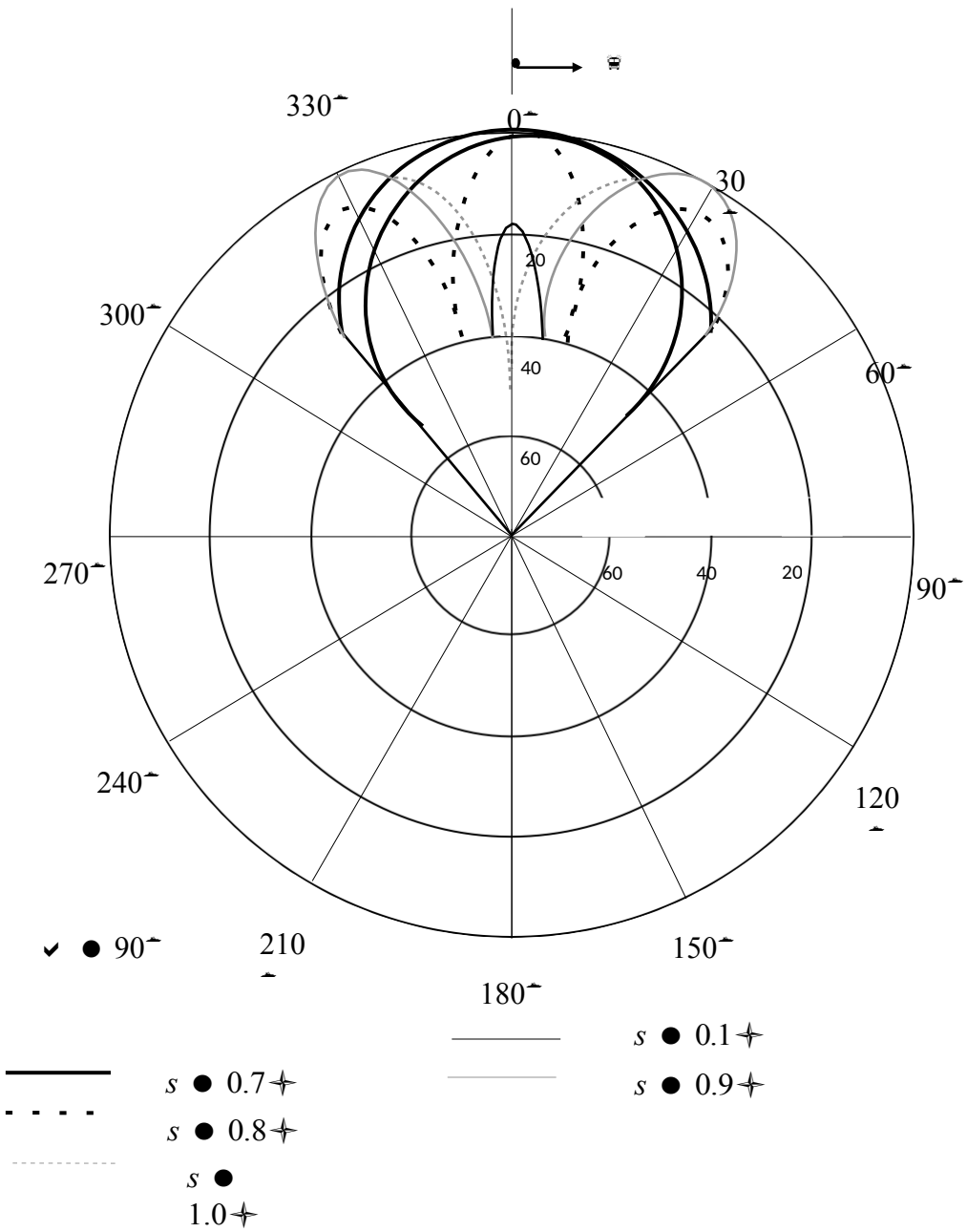


Figure 3.13 Normalized radiation patterns for $\psi = 90^\circ$ corner reflector for various values of s.

3.3.2 Other Corner Reflectors

A similar procedure can be used to derive the array factors and total fields for all other corner reflectors with included angles of ✔ ● $180^{\circ} / n$. Referring to Figure 3.11, it can be shown that the array factors for ✔ ● 60°, 45° *and* 30° can be written as

✔ ● 60°

$$AF(X,Y) = 4\sin\left(\frac{X}{2}\right)\cos\left(\frac{Y}{2}\right) \tag{3.25}$$

✔ ●

$$AF(X,Y) = 2\cos\left(\frac{45^{\circ}}{2}\right)\cos(X)\cos(Y) \tag{3.26}$$

$$= 2\cos\left(\frac{X}{2}\right)$$

$$= 4\sqrt{2}$$

$$= \sqrt{2}$$

✔ ● 30°

$$AF(\theta,\phi) = 2\left[\cos(X) + 2\cos\left(\frac{X}{2}\right)\cos\left(\frac{3Y}{2}\right) + \cos(Y) + 2\cos\left(\frac{X}{2}\right)\cos\left(\frac{3Y}{2}\right)\right] \tag{3.27}$$

where

$$X = ks\sin(\theta)\cos(\phi) \tag{3.28}$$

$$Y = ks\sin(\theta)\sin(\phi) \tag{3.29}$$

The array factor for a corner reflector has a form that is similar to the array factor for a uniform circular array. This should be expected since the feed sources and their images in Figure 3.11 form a circular array. The number of images increases as the included angle of the corner reflector decreases.

Patterns have been computed for corner reflectors with included angles of 60°, 45° *and* 30°. It has been found that these corner reflectors have also single-lobed patterns for the smaller values of s, and they become narrower as the included angle decreases. Multiple lobes begin to appear when

$s \le 0.95\lambda$	for $\alpha = 60^\circ$
$s \le 1.2\lambda$	for $\alpha = 45^\circ$
$s \le 2.5\lambda$	for $\alpha = 30^\circ$

The maximum field strength increases as the included angle of the reflector decreases. This is expected since a smaller angle reflector exhibits better directional characteristics because of the narrowness of its angle. The maximum values of $|E/E_0|$ for $\alpha = 60^\circ$, 45°, *and* 30° are approximately 5.2, 8, and 9, respectively. The first field strength peak, is achieved when

$s \le 0.65\lambda$	for $\alpha = 60^\circ$
$s \le 0.85\lambda$	for $\alpha = 45^\circ$
$s \le 1.20\lambda$	for $\alpha = 30^\circ$

The gain in the direction $\theta = 0$ are shown in Figure 3.12 for each corner angle. The solid curve in each case is computed for zero losses ($R_{1L} = 0$), while the dashed curve is for an assumed loss resistance $R_{1L} = 1\Omega$. It is apparent that for efficient operation too small spacing should be avoided. A small spacing is also objectionable because of narrow bandwidth. On the other hand, too large a spacing results in less gain.

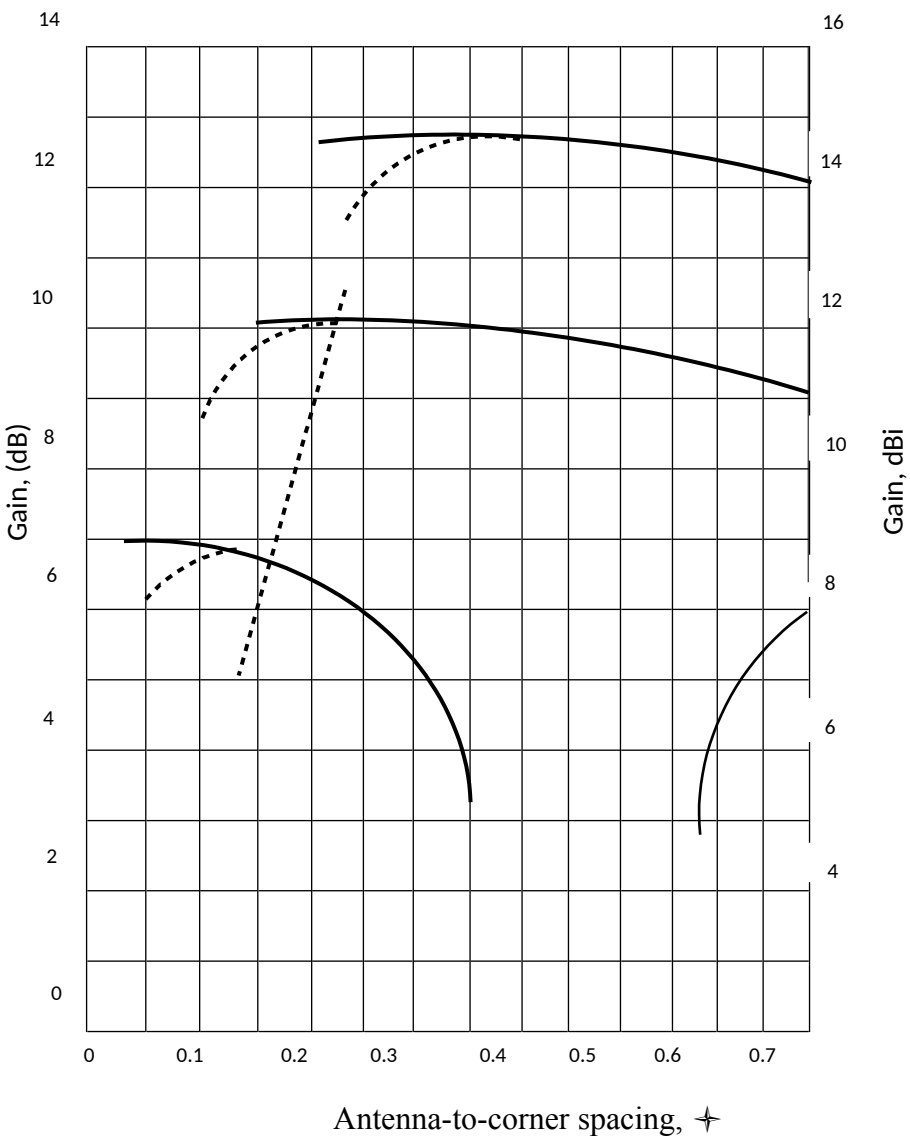


Figure 3.14 Gain of corner reflector antennas over a $\lambda/2$ dipole antenna in free space with the same power input as a function of the antenna-to-corner spacing. Gain is in the direction $\theta = 0$ and is shown for zero loss resistance (solid curves) and for an assumed loss resistance of 1Ω $R_{1L} = 1\Omega$ (dashed curves).

The gain of a 90° corner reflector with antenna-to-corner spacing $S = 0.5\lambda$ is nearly 10 dB over a reference $\lambda/2$ antenna or 12 dBi.

Restricting patterns to the lower-order radiation mode (no minor lobes), it is generally desirable that S lie between the following limits:

	S
90°	$0.25\text{-}0.7\lambda$
180° (flat sheet)	$0.1\text{-}0.3\lambda$

In the above discussions, it is assumed that the reflectors are perfectly conducting and of infinite extent, with the exception that the gains with a finitely conducting reflector may be approximated with a proper choice of R_{1L}.

Although the gain of a corner reflector with infinite sides can be increased by reducing the corner angle, it does not follow that the gain of a corner reflector with finite sides of fixed length will increase as the corner angle is decreased. To maintain a given efficiency with a smaller corner angle requires that S be increased. Also on a 60° reflector, for example, the point at which a wave is reflected parallel to the axis is at a distance of $1.73S$ from the corner as compared to $1.41S$ for the 90^0 type. Hence, to realize the increase in gain requires that the length of the reflector sides be much larger than for a 90° corner reflector designed for the same frequency. Usually this is a practical disadvantage in view of the relatively small increase to be expected in gain.

If the length or arm of the reflector is reduced to values of less than 0.6λ, radiation to the sides and rear tends to increase and the gain decreases. When R is decreased to as little as 0.3λ, the strongest radiation is no longer forward and the –reflector || acts as a director.

PARABOLIC REFLECTOR

If a beam of parallel rays is incident upon a reflector whose geometrical shape is a parabola, the radiation will converge or get focused at a spot which is known as the *focal point*. In the same manner if a point source is placed at the focal point, the rays reflected by a parabolic reflector will emerge as a parallel beam. The symmetrical point on the parabolic surface is known as the *vertex*. Rays that emerge in a parallel formation are usually said to be *collimated*. In practice, collimation is often used to describe the highly directional characteristics of an antenna even though the emanating rays are not exactly parallel. Since the transmitter (receiver) is placed at the focal point of the parabola, the configuration is usually known as *front fed*.

A parabolic reflector can take two different forms. One configuration is that of the parabolic right cylinder, whose energy is collimated at a line that is parallel to the axis of the cylinder through the focal point of the reflector. The most widely used feed for this type of a reflector is a linear dipole, a linear array, or a slotted waveguide. The other reflector configuration is that which is formed by rotating the parabola around its axis, and it is referred to as a *paraboloid*

(parabola of revolution). A pyramidal or a conical horn has been widely utilized as a feed for this arrangement.

Front-Fed Parabolic Reflector

Parabolic cylinders have widely been used as high-gain apertures fed by line sources. The analysis of a parabolic cylinder (single curved) reflector is similar, but considerably simpler than that of a paraboidal (double curved) reflector. The principle characteristics of aperture amplitude, phase, and polarization for a parabolic cylinder, as contrasted to those of a paraboloid, are as follows:

1. The amplitude taper, due to variations in distance from the feed to the surface of the reflector, is proportional to 1/✕ in a cylinder compared to 1/ r^2 in a paraboloid.

2. The focal region, where incident plane waves converge, is a line-source for a cylinder and a point source for a paraboloid.

3. When the fields of the feed are linearly polarized parallel to the axis of the cylinder, no cross-polarized components are produced by the parabolic cylinder. That is not the case for a paraboloid.

The surface of a paraboloidal reflector is formed by rotating a parabola about its axis. Its surface must be a paraboloid of revolution so that rays emanating from the focus of the reflector are transformed into plane waves. The design is based on optical techniques, and it does not take into account any deformations (diffractions) from the rim of the reflector. Referring to Figure 3.15 and choosing a plane perpendicular to the axis of the reflector through the focus, it follows that

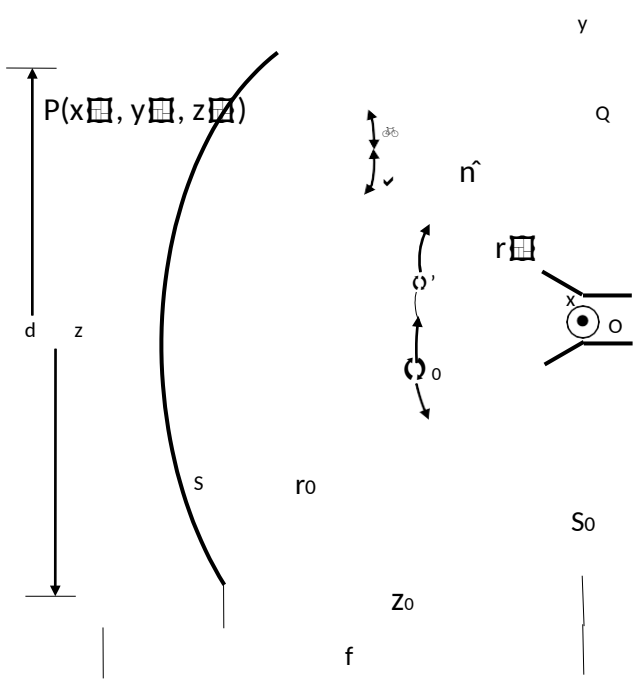


Figure 3.15 Two-dimensional configuration of a paraboloidal reflector.

$$OP + PQ = \text{constant} = 2f \tag{3.30}$$

Since

$$OP = r' \tag{3.31}$$

$$PQ = r'\cos\theta'$$

(3.30) can be written as

$$r'(1+\cos\theta') = 2f \tag{3.32}$$

or

$$r = \frac{2f}{1+\cos\theta} = f\sec^2\frac{\theta}{2} \qquad \theta \le \theta_0 \tag{3.32a}$$

Since a paraboloid is a parabola of revolution (about its axis), (3.32a) is also the equation of a paraboloid in terms of the spherical coordinates r, θ, ϕ. Because of its rotational symmetry, there are no variations with respect to ϕ.

Another expression that is usually very prominent in the analysis of reflectors is that relating the subtended angle θ_0 to the *f/d* ratio. From the geometry of Figure 3.15

$$\theta_0 = \tan^{-1}\frac{d/2}{z_0} \tag{3.33}$$

where z_0 is the distance along the axis of the reflector from the focal point to the edge of the rim.

$$z_0 = \tan^{-1}\left[\frac{d}{2f}\right] = \left[\frac{1}{2}\frac{f}{d}\right] \tag{3.34}$$

$$f \approx \frac{d^2}{16f} \qquad \left[\frac{f}{d}\right] \approx \frac{1}{16}$$

It can also be shown that another form of (3.34) is

$$f = \frac{d}{4}\cot\frac{\theta_0}{2} \tag{3.35}$$

Aperture antennas usually have an obvious physical aperture of area A_p through which energy passes on its way to the far field. The maximum achievable gain for an aperture antenna is

$$G_{max} = D_u = \frac{4\pi}{\lambda^2}A$$

$2 \quad p$

This gain is possible only under the ideal circumstances of a uniform amplitude, uniform phase antenna with no spillover or ohmic losses present. In practice, these conditions are not satisfied and gain is decreased from ideal, as represented through the following:

$$G \bullet$$

$$D \bullet \quad \frac{4}{A} \tag{3.37}$$

$$ap \quad u \qquad ap \quad 2 \quad p$$

It is found that for a given feed pattern

- There is only one reflector with a given angular aperture or *f/d* ratio which leads to a maximum aperture efficiency.
- Each maximum aperture efficiency is in the neighborhood of 82-83%.
- Each maximum aperture efficiency, for any one of the given patterns, is almost the same as that of any of the others.
- As the feed pattern becomes more directive, the angular aperture of the reflector that leads to the maximum efficiency is smaller.

The aperture efficiency is generally the product of the

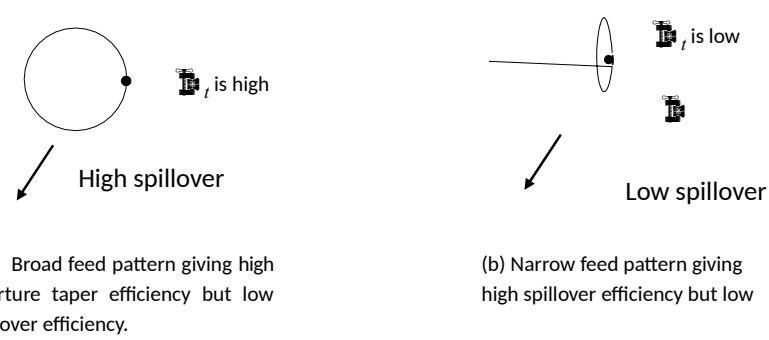
- fraction of the total power that is radiated by the feed, intercepted, and collimated by the reflecting surface (generally known as *spillover efficiency* e_s)
- uniformity of the amplitude distribution of the feed pattern over the surface of the reflector (generally known as *taper efficiency* e_t)
- phase uniformity of the field over the aperture plane (generally known as *phase efficiency* e_p)
- polarization uniformity of the field over the aperture plane (generally known as polarization efficiency e_x)
- blockage efficiency e_b
- random error efficiency e_r over the reflector

surface This in general

$$e_{ap} = e_s \, e_t \, e_p \, e_x \, e_b \, e_r \tag{3.38}$$

An additional factor that reduces the antenna gain is the attenuation in the antenna feed and associated transmission line.

The two main factors that contribute to the aperture efficiency are the spillover and nonuniform amplitude distribution losses. Because these losses depend primarily on the feed pattern, a compromise between spillover and taper efficiency must emerge. It has been depicted pictorially in Figure 3.16.



(a) Broad feed pattern giving high aperture taper efficiency but low spillover efficiency.

(b) Narrow feed pattern giving high spillover efficiency but low

Figure 3.16 Illustration of the influence of the feed antenna pattern on reflector aperture taper and spillover.

Very high spillover efficiency can be achieved by a narrow beam pattern with low major lobes at the expense of a very low taper efficiency. Uniform illumination and ideal taper efficiency can be obtained when the feed power pattern G_f ( ) is given by

(3.39)

which is plotted in Figure 5.13. Although such a pattern is ‗ideal‖ and impractical to achieve, much effort has been devoted to develop feed designs which attempt to approximate it.

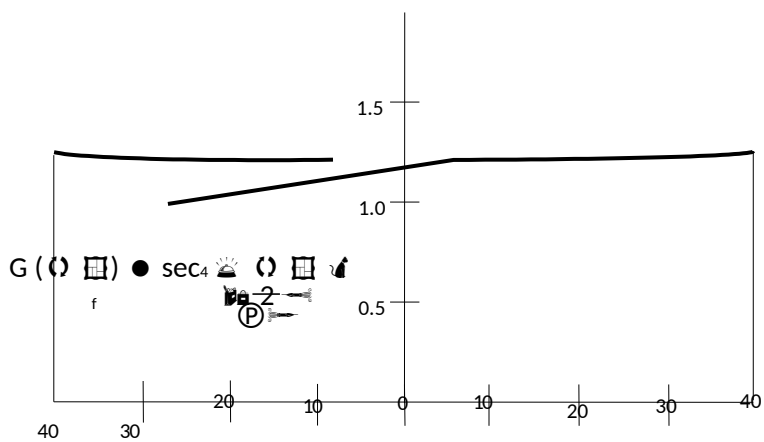
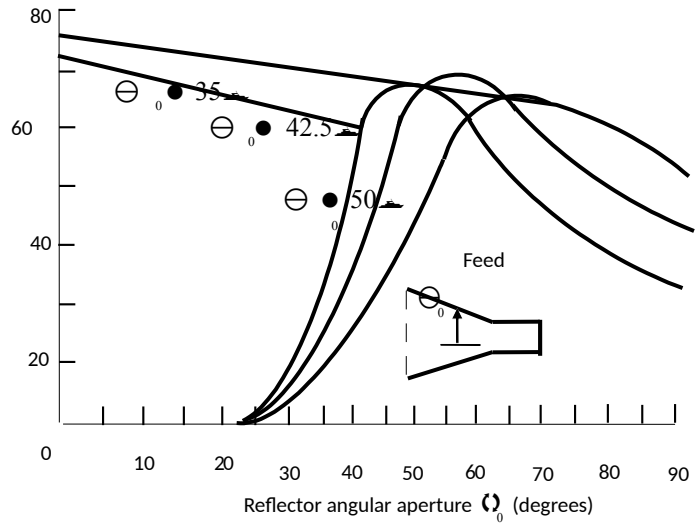


Figure 3. 17 Normalized gain pattern of feed for uniform amplitude illumination of paraboloidal reflector with a total subtended angle of 80°.

To develop guidelines for designing practical feeds which results in high aperture efficiencies, it is instructive to examine the relative field strength at the edges of the reflector‘s bounds ( ) for patterns that lead to optimum efficiencies.

In practice, maximum reflector efficiencies are in the 65-80% range. To demonstrate that paraboloidal reflector efficiencies for square corrugated horns feeds were computed and are shown in Figure 3.18(a). For the data of Figures 3.18 (a)and (b), each horn had aperture dimensions of 8λ × 8λ, their patterns were assumed to be symmetrical (by averaging the E- and H-planes). From the plotted data, it is apparent that the maximum aperture efficiency for each feed pattern is in the range of 74-79%, and that the product of the taper and spillover efficiencies is approximately equal to the total aperture efficiency.



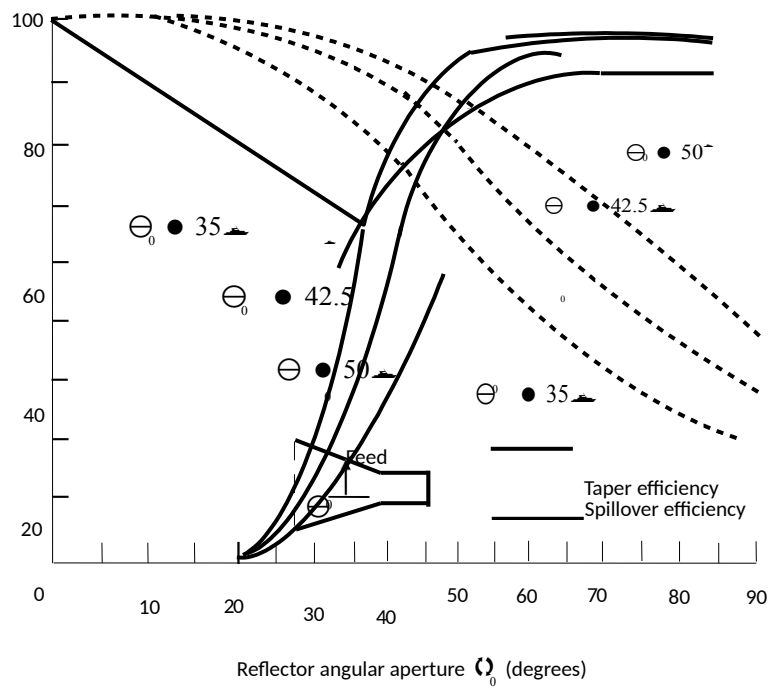


Figure 3.18 Parabolic reflector aperture efficiency as a function of angular aperture for 8λ × 8λ square corrugated horn feed with total flare angles of $2\theta_0$ = 70°, 85°, and 100°.

Phase Errors

Any departure of the phase, over the aperture of the antenna, from uniform can lead to a significant decrease in its directivity. For a paraboloidal reflector system, phase errors result from

1. displacement (defocusing) of the feed phase center from the focal point
2. deviation of the reflector surface from a parabolic shape or random errors at the surface of the reflector
3. departure of the feed wave fronts from spherical shape

The defocusing effect can be reduced by first locating the phase center of the feed antenna and then placing it at the focal point of the reflector. It is found that the phase center for horn antennas, which are widely utilized as feeds for reflectors, is located between the aperture of the horn and the apex formed by the intersection of the inclined walls of the horn. Very simple theory has been derived to predict the loss in directivity for rectangular and circular apertures when the peak values of the aperture phase deviation is known. When the phase errors are assumed to be relatively small, it is not necessary to know the exact amplitude or phase distribution function over the aperture.

CASSEGRAIN REFLECTORS

The disadvantage of the front-fed arrangement is that the transmission line from the feed must usually be long enough to reach the transmitting or the receiving equipment, which is usually placed behind or below the reflector. This may necessitate the use of long transmission lines whose losses may not be tolerable in many applications, especially in low-noise receiving systems. In some applications, the transmitting or receiving equipment is placed at the focal point to avoid the need for long transmission lines. However, in some of these applications, especially for transmission that may require large amplifiers and for low-noise receiving systems where cooling and weatherproofing may be necessary, the equipment may be too heavy and bulky and will provide undesirable blockage.

The arrangement that avoids placing the feed (transmitter and/or receiver) at the focal point is that shown in Figure 3.1(d) and it is known as the Cassegrain feed. Through geometrical optics, Cassegrain, a famous astronomer (N. Cassegrain of France, hence its name), showed that incident parallel rays can be focused to a point by utilizing two reflectors. To accomplish this, the main (primary) reflector must be a parabola, the secondary reflector (Subreflector) a hyperbola, and the feed placed along the axis of the parabola usually at or near the vertex. Cassegrain used this scheme to construct optical telescopes, and then its design was copied for use in radio frequency systems. For this arrangement, the rays that emanate from the feed illuminate the Subreflector and are reflected by it in the direction of the primary reflector, as if they originated at the focal point of the parabola (primary reflector). The rays are then reflected by the primary reflector and are converted to parallel rays, provided the primary reflector is a parabola and the subreflector is a hyperbola. Diffraction occurs at the edges of the subreflector and primary reflector and they must be taken into account to accurately predict the overall system pattern, especially in regions of low intensity. Even in regions of high intensity, diffraction must be included if an accurate formation of the fine ripple structure of the pattern is desired. With the Cassegrain-feed arrangement, the transmitting and/or receiving equipment can be placed behind the primary reflector. This scheme makes the system relatively more accessible for servicing and adjustments.

Cassegrain designs, employing dual reflector surfaces, are used in applications where pattern control is essential, such as in satellite ground-based systems, and have efficiencies of 65-80%. They supersede the performance of the single-reflector front-fed arrangement by about 10%. Using geometrical optics, the classical Cassegrain configuration, consisting of a paraboloid and

hyperboloid, is designed to achieve a uniform phase front in the aperture of the paraboloid. By employing good feed designs, this arrangement can achieve lower spillover and more uniform illumination of the main reflector. In addition, slight shaping of one or both of the dual-reflector‘s surfaces can lead to an aperture with almost uniform amplitude and phase with substantial enhancement in gain. These are referred to as shaped reflectors. Shaping techniques have been employed in dual-reflectors used in earth station applications.

Two reflectors with ray geometry, with concept of equivalent parabola, are shown in Figure 3.19 The use of a second reflector, which is usually referred to as the subreflector or subdish, gives an additional degree of freedom for achieving good performance in a number of different applications. For an accurate description of its performance, diffraction techniques must be used to take into account diffractions from the edges of the subreflector, especially when its diameter is small.

In general, the Cassegrain arrangement provides a variety of benefits, such as the

1. ability to place the feed in a convenient location
2. reduction of spillover and minor lobe radiation
3. ability to obtain an equivalent focal length much greater than the physical length
4. capability for scanning and/or broadening of the beam by moving one of the reflecting surfaces

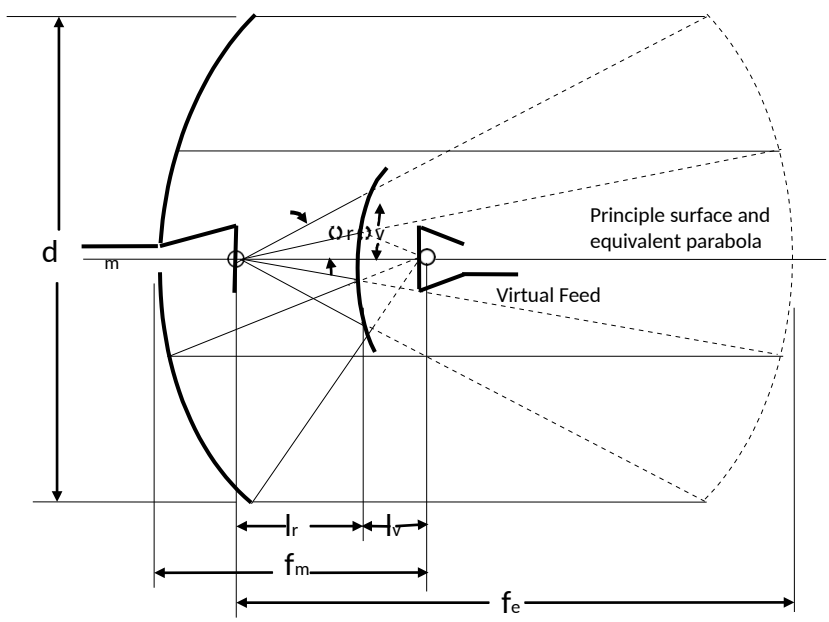


Figure 3.19 Equivalent parabola concepts.

To achieve good radiation characteristics, the subreflector must be few wavelengths in diameter. However, its presence introduces shadowing which is the principle limitation of its use as a microwave antenna. The shadowing can significantly degrade the gain of the system, unless the main reflector is several wavelengths in diameter. Therefore the Cassegrain is usually attractive for applications that require gains of 40 dB or greater. There are, however, a variety of

techniques that can be used to minimize the aperture blocking by the subreflector. Some of them are minimum blocking with simple Cassegrain, and twisting Cassegrains for least blocking.

Sub-reflectors offer flexibility of design for reflecting telescopes. Referring to Figure 3.20, it is required that all rays from the focal point F form a spherical wave front (circle of radius CF‘) on reflection from the (hyperbolic) subreflector (as though radiating isotropically from the parabola focus F‘) or by Fermat‘s principle of equality of path length that

$$CA + AF = FA + CA \tag{3.40}$$

Noting that $CA = CF + AF$ and that $FA + AF = 2OA$ we obtain

$$FA + AF = 2OA = BA \tag{3.41}$$

which is the relation for an hyperbola with standard form

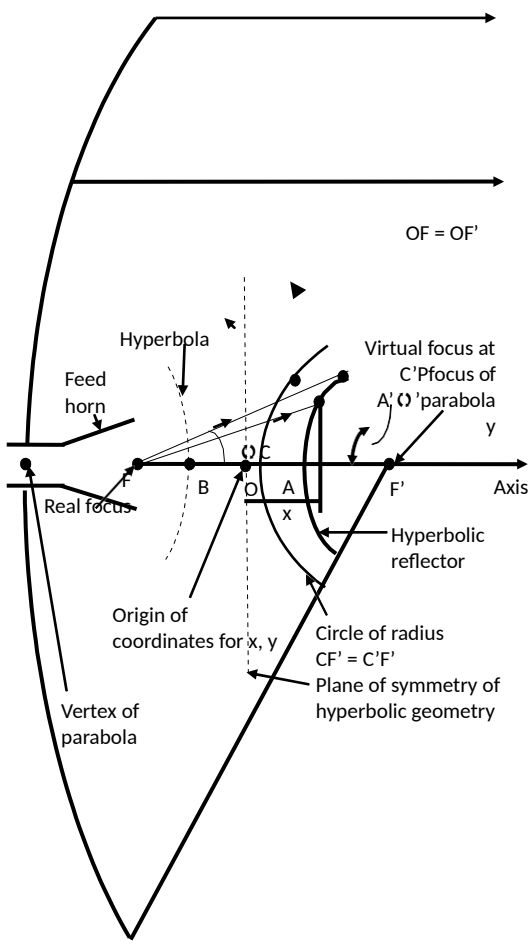


Figure 3.20 Geometry for Cassegrain reflector.

$$\frac{x^2}{a^2} - \frac{y^2}{f^2 - a^2} = 1 \tag{3.42}$$

where $a = OA = OB, f = OF$ [illegible] $= OF$, and x and y are as shown in Figure 3.20. Or

$$y = \left(\frac{1}{a^2}\right)\left(f^2 - a^2\right) x^2 \qquad (3.43)$$

The parabolic sub-reflector is then truncated at point P for which a ray reflected from the hyperbola hits the edge of the parabolic reflector. The hyperbolic reflector then subtends an angle [illegible] from the feed location at the focal point F while the (main) parabolic reflector subtends an angle [illegible] from the focal point F' of the parabola. Thus, the feed horn beam angle is increased in the ratio [illegible] to fill the parabola aperture.

The surface of the hyperbola is deformed to enlarge or restrict the incremental ray bundle, thereby decreasing or increasing the watts per steradian in the bundle and finally the watts per square meter in the aperture plane of the parabola. This shaping technique may be extended over the entire sub-reflector and often both sub-reflector and parabola are shaped. As a result a more uniform aperture distribution and higher aperture efficiency can be achieved but with higher first sidelobes and also more rapid loss as the feed is moved off-axis to squint the beam.

A constraint on the Cassegrain arrangement is that to minimize blockage the sub-reflector should be small compared to the parabola, yet the sub-reflector must be large compared to the wavelength.

In Figure 3.21 a *parabola* is given by

$$y^2 = 4fx \qquad (3.44)$$

where f = focal distance = VF

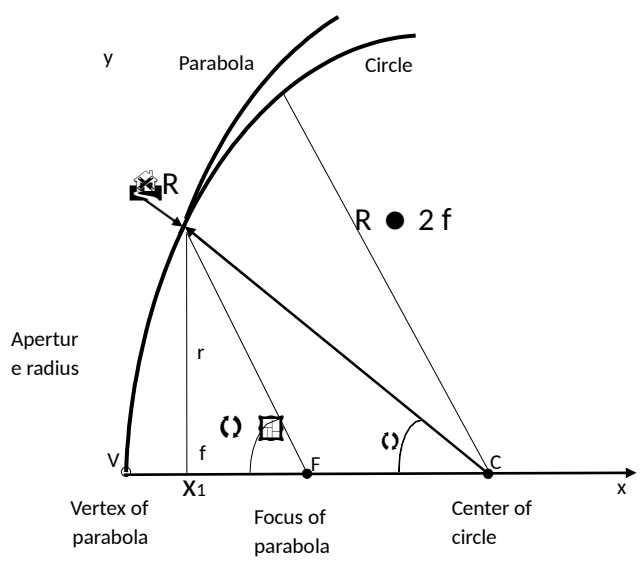


Figure 3.21 Circle and parabola compared, with radius of circle equal to twice the focal length of the parabola.

This parabola is compared with a *circle* of radius $R = VC$. It may be shown that for small values of x, the circle is of nearly the same form as the parabola when

$$R = 2f \tag{4.45}$$

Over an angle ❍ and *aperture radius*

$$r \bullet R \sin ❍ \tag{4.46}$$

The circle differs from the parabola by less than ✍R . If ✍R ❚❚ ✦ (or specifically ■ ✦ /16) the field radiated from a point source at F within an angle ❍ ⊞ and reflected from the circle will be within 45☛ (● 2 ✐ 360☛ /16) of the phase of a field radiated from F and reflected from the parabola. Then a feed antenna at the focal point F which illuminates the sphere only within the angle ❍ ⊞ will produce a plane wave over the aperture of diameter 2r having a phase deviation of less than 45☛, this amount of deviation occurring only near the edge of the aperture.

UNIT IV

Antenna Arrays & Measurements

Antenna Array

INTRODUCTION

Usually a single element provides wide radiation and low directivity (gain).
In many applications it is necessary to design antennas with very directive characteristics to meet the demands of long distance communication.

- Enlarging the dimensions of single elements.
- Enlarging the dimensions of the antenna, without increasing the size of the individual elements, is to form an assembly of radiating elements in an electrical and geometrical configuration.

This new antenna, formed by multi-elements, is referred to as an ***array***.
In most cases, the elements of an array are identical. This is not necessary, but it is often convenient, simpler, and more practical. The individual elements of an array may be of any form (wires, apertures, etc).

- **The total field of the array is determined by the vector addition of the fields radiated by the individual elements**
  - During the summation, the current in each element is assumed to be the same as that of the isolated element (neglecting coupling).
  - This is usually not the case and depends on the separation between the elements.

In an array of identical elements, there are at least five controls that can be used to shape the overall pattern of the antenna.
1. The geometrical configuration of the overall array (linear, circular, rectangular, spherical, etc.)
2. The relative displacement between the elements
3. The excitation amplitude of the individual elements
4. The excitation phase of the individual elements
5. The relative pattern of the individual elements

This is one of the common methods of combining the radiations from a group of similar antennas in which the wave-interference phenomenon is involved. The field strength can be increased in preferred directions by properly exciting group or array of antennas simultaneously, such as arrangement is known as antenna array. Array of antenna is an arrangement, of several individual antennas so spaced and phased that their individual contributions coming in one preferred direction and cancel in all other directions, which will be going to increase the directivity of the system.
The different types of arrays with regard to beam pointing direction are as follows,
1. Broadside array
2. End fire array
3. Collinear array.

1 . Broadside Array

Broadside array is one of the most commonly used antenna array in practice. The array in which a number of identical parallel antennas are arranged along a line perpendicular to the line of array axis is known as broadside array, which is shown in figure (2.1). In this, the individual antennas are equally spaced along a line and each element is fed with current of equal magnitude, all in the same phase.

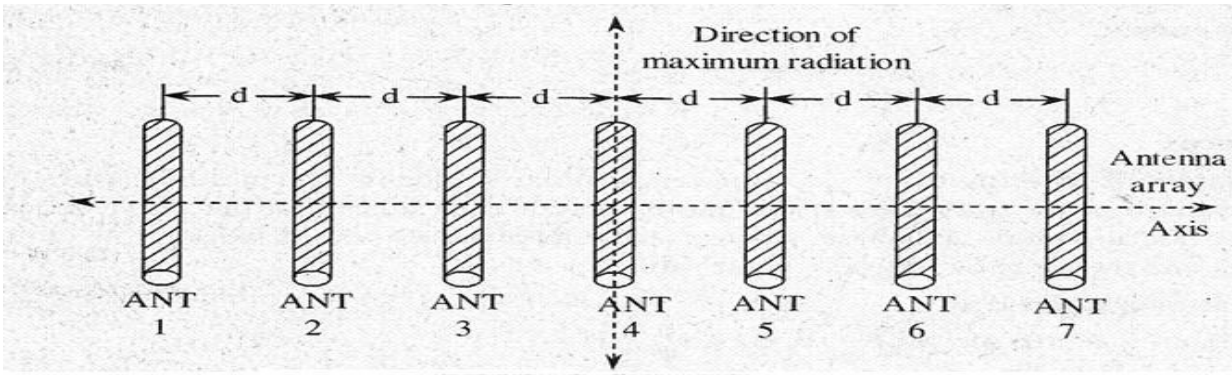


fig 2.1 Board side array

The radiation pattern of broadside array is bidirectional, which radiates equally well in either direction of maximum radiation.

2 . End Fire Array

The array in which a number of identical antennas are spaced equally along a line and Individual elements are fed with currents of unequal phases (i.e., with a phase shift of 180°) is known as end fire array .This array is similar to that of broadside array except that individual elements are fed in with, a phase shift of 180 .In this, the direction of radiation is coincides with the direction of the array axis, which is shown in figure (2.2).

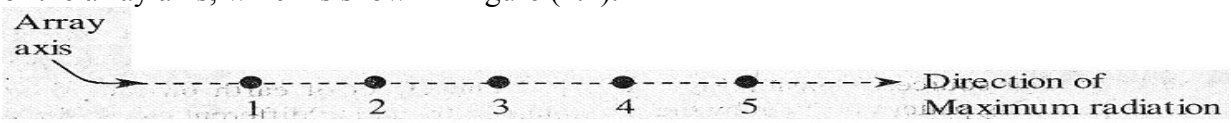


fig 2.2 End fire array

The radiation pattern of end fire array is unidirectional. But, the end fire array may be bidirectional also. One such example is a two element array, fed with equal current, 180° out of phase**.**

3. Collinear Array

The array in which antennas are arranged end to end in a single line is known as collinear array. Figure (2.3), shows the arrangement of collinear array, in which one antenna is stacked over another antenna. Similar to that of broadside array, the individual elements of the collinear array are fed with equal in phase currents. A collinear array is a broadside radiator, in which the direction of maximum radiation is perpendicular to the line of antenna. The collinear array is sometimes called as broadcast or Omni directional arrays because its radiation pattern has circular symmetry with its main to be everywhere perpendicular to the principal axis.

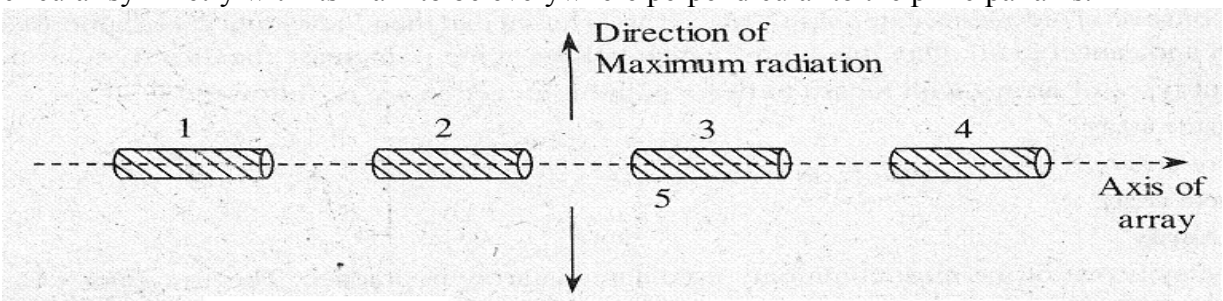


fig 2.3 Collinear array

3. **Arrays of two point sources with equal amplitude and opposite phase:**

In this**,** point source 1 is out of phase or opposite phase (180°) to source 2 i.e. when there is

maximum in source 1 at one particular instant, and then there is minimum in source 2 at that instant and vice-versa.

Referring to Fig.1.1 the total far field at distant point ***P*** **, is** given by

E = (-Ei e - φ j / 2) + (+ E 2 e + φ i / 2)

But *E1=E2=E0* (say)

Then E=E0 2j

E=2jE0 sinφ/2….....(1.1a)

E=2jE0sin

Cosθ.....(1.1b)

Let ***d*** = λ/2 and ***2E0 j* = 1**

Enorm, = sin (π/2 COS θ).....(1.2)

Maximum directions: Maximum value of sine function is ± 1

sin (**π**/2 cosθ**) = ± 1**

(Π/2 cosθmax) = ± (2 n + **1) *n / 2*** where ***n* = 0, 1,** 2

(cosθmax**) = ± 1** if ***n* = 0**

θmax= 0° **and 180°….....(1.3 a)**

Minima directions: Minimum value of a sine function is 0

sin (π/2 cosθ) = 0

π/2 cosθmin = ± nπ where n = 0, 1, 2……..

COS θmin = 0

Therefore θmin= 90° and -90°.....(1.3b)

Half power point directions:

sin (Π /2 cosθ) = ±

√

Π/2 cosθ*HPPD* = ± ***(2 n*** + 1) π/4

Π/2 cosθ*HPPD* = ± π / 4 i f n = 0

cosθ*HPPD* = ±

θ*HPPD* = **60°, ± 120°.....**(1.3c)

From these, it is possible to draw the field pattern which is as shown in Fig.1.2

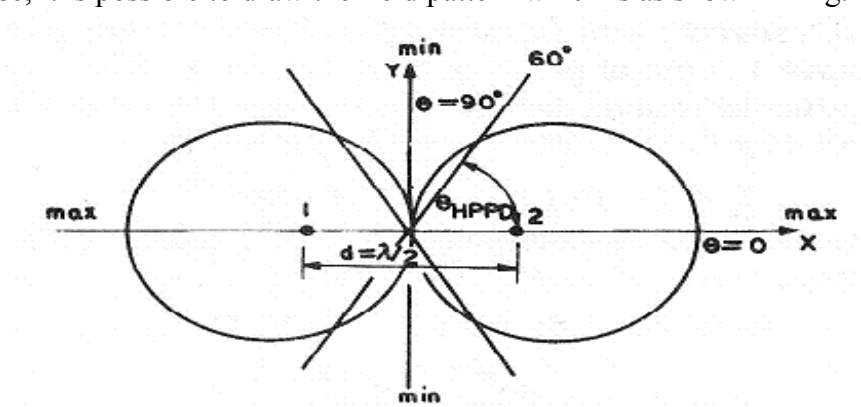


fig 1.2 Two Point sources with equal amplitude and opposite phase spacing λ/2

It is seen that maxima have shifted 90° along X-axis in comparison to in-phase field pattern. The figure is horizontal figure of 8 and 3-dimensional space pattern is obtained by rotating it along X -axis. Once the arrangement gives maxima along line joining the two sources and hence this is one of the simpl e s t t y p e o f "End fire" 'Array'.

What is uniform linear array? Discuss the application of linear array? and also explain the advantages and disadvantage of linear array?

Ans: In general single element antennas having non uniform radiation pattern are used in several broadcast services. But this type of radiation pattern is not useful in point-to-point communication and services that require to radiate most of the energy in one particular direction i.e., there are applications where we need high directive antennas. This type of radiation pattern is achieved by a mechanism called antenna array. An antenna array consists of identical antenna elements with identical orientation distributed in space. The individual antennas radiate and their radiation is coherently added in space to form the antenna beam.

In a linear array, the individual antennas of the array are equally spaced along a straight line. This individual antennas of an array are also known as elements. A linear array is said to be uniform linear array, if each element in the array is.fed with a current of equal magnitude with progressive phase shift (phase shift between adjacent antenna elements).

Application of Linear Array

1. Adaptive linear arrays are used extensively in wireless communication to reduce interference between desired users and interfering signals.
2. Many linear arrays spaced parallel on the common plane create a planar array antenna. These are used in mobile radar equipment.
3. The linear array is most often used to generate-a fan beam and is useful where broad coverage in one plane and narrow beam width in the orthogonal plane are desired.
4. Linear arrays can be made extremely compact and .are therefore very attractive for shipboard applications.

The advantages and disadvantages of linear arrays are as follows.

Advantages

1. Increases the overall gain. .
2. Provide diversity receptions.
3. Cancel out interference from a particular set of directions.
4."Steer" the array so that it is more sensitive in a particular direction.
5. Determines the direction of arrival of the incoming signals.
6. It maximize the Signal to Interference plus Noise ratio

Disadvantages

1. Ray deflection only in a single plane possible.
2. Complicated arrangement and more electronically controlled phase shifter needed. ;
3. Field view is restricted.
4. Considerable minor lobes are formed.
5. Large power loss due to current flowing in all elements.
6. Overall efficiency decreases.
3. Costly to implement.

What is linear array? Compare Broad side array and End fire array?

Ans : Linear arrays: The arrays in which the individual antennas (called as elements) are equally spaced along a straight line are called as linear arrays. Thus, linear antenna array is a system of equally spaced elements.

Broad side array	End fire array
1. The array is said to be broad side array, if the direction of maximum radiation is perpendicular to the array axis. 2. In broad side, phase difference $\alpha = 0$	1. The array is said to be end fire array, if maximum radiation is along the array axis. 2. In end fire, phase difference between adjacent element is $\alpha = -\beta d$

3.General equation for pattern maxima is (θmax)minor=cos-1 4.general Expression for pattern minima is, (θmax)minor=cos-1 *Π* 5.Half power beam width is given by, HPBW= degree 6.Directivity of broad side array is , D=2 L=Length of array 3.Beam width between first nulls is, BWFN= degree 8. In broad side array, all elements are equally spaced along the array axis and fed with current of equal magnitude and same phase. 9. Radiation pattern of broad side array is bidirectional 10.In broad side array, φ =βdcosθ+α (since α=0) Therefore φ =βdcosθ	3. General expression for pattern maxima is, (θmax)minor=cos- 4.Genaral expression for pattern minima is, θ min =2sin-1 5.Halfpower beam width is given by, HPBW=53.3 degree 6.Directivity of end fire array is, D= 4 L=Length of array 3.Beam width between first nulls is BWFN=114.6 degree 8. In End fire array, all elements are equally spaced along the array axis and fed with current of equal magnitude but their phases are different. 9. Radiation pattern of broad side array is Unidirectional 10.In end fire array, φ =βdcosθ+α (since α=-βd) Therefore φ =βd(cosθ-1)

4. Explain the principal of pattern multiplication .What is the effect of earth of radiation pattern of antennas?

Ans: Multiplication of Patterns ,

The total field pattern of an array of non-isotropic but similar sources is the multiplication of the individual source pattern and the pattern of an array of isotropic point sources each located at the phase centre of individual source and having the relative amplitude and phase, where as the total phase patterns is the addition of the phase pattern of the individual sources and the array of isotropic point sources. Total field by an array is defined as

E = { E0(θ,φ) x Ei(θ,φ)} x { Epi(θ,φ)+ Epa(θ,φ) }

= (Multiplication of field patterns) (Addition of phase patterns)

Where

E - Total field

E0(θ,φ) = Field pattern of individual source

Ei(θ,φ) = Field pattern of array of isotropic point source

Epi(θ,φ)= Phase pattern of individual source

Epa(θ,φ)= Phase pattern of array of isotropic point sources.

Hence, θ and φ are polar and azimuth angles respectively.

The principle of multiplication of pattern is best suited for any number of similar sources.

Considering a two dimensional case, the resulting pattern is given by the equation,

E = 2 E0 cosφ /2

E = 2 E1 sinθ cosφ /2.

E = E (θ) cosφ /2

It can be seen that E0 is a function of E (θ). In the above equation the total field pattern is equal to the product of primary pattern E (θ) and a secondary pattern cosφ/2.

Effect of Earth on the Radiation Pattern

The effect of earth on the radiation pattern can be obtained by using an image principle. In image principle, earth is considered as an image antenna of same length and current as shown in the figure (3.1).

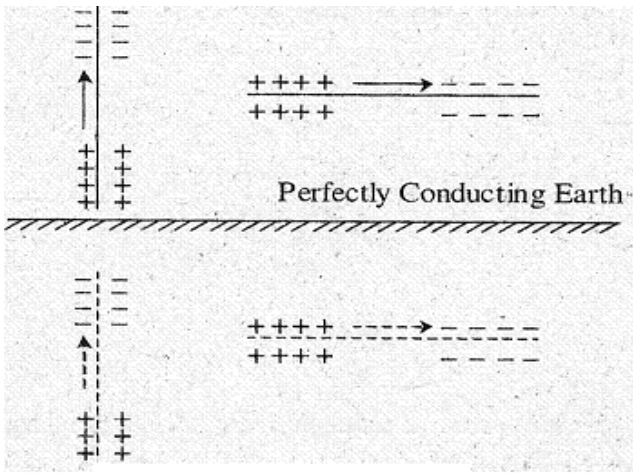


fig 3.1 Earth as Image Antenna

For vertical antenna, currents in actual and image antennas are equal and have same direction, whereas opposite direction for horizontal antenna. The resultant field is obtained by the addition of field of an image antenna to that of an actual antenna. The shape of the vertical pattern is affected more than the horizontal pattern

Effect of Earth on the Radiation Pattern of Vertical Antenna

The ground-effect factor of a perfectly conducting earth is given as,

2cos sinφ

Where,

h = Height of the center of antenna above earth

φ = Elevation angle above horizontal.

But, for the case of finite conducting of earth, the above given expression is valid for large angles of φ0. Whereas, for low angles of φ0, less than 15° known as "Pseudo-Brewster angle", the phase of the reflection factor is nearer to 180° than it is to 0° and the use of above equation would lead to erroneous result.

The effect of earth on radiation pattern can be explained by taking different cases of onductivities (σ). The function _n‘ is defined as,

n =

Where, x=σ/ω

σ = conductivity of the earth in mho/meter

►►=15, Relative dielectric of the earth.\

The vertical radiation pattern of a vertical dipole is a shown in the fig 3.2

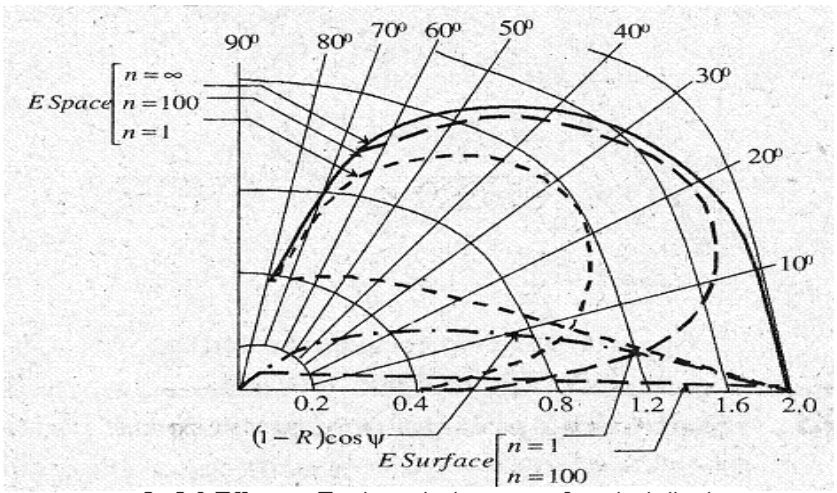


fig 3.2 Effect on Earth vertical pattern of vertical dipole

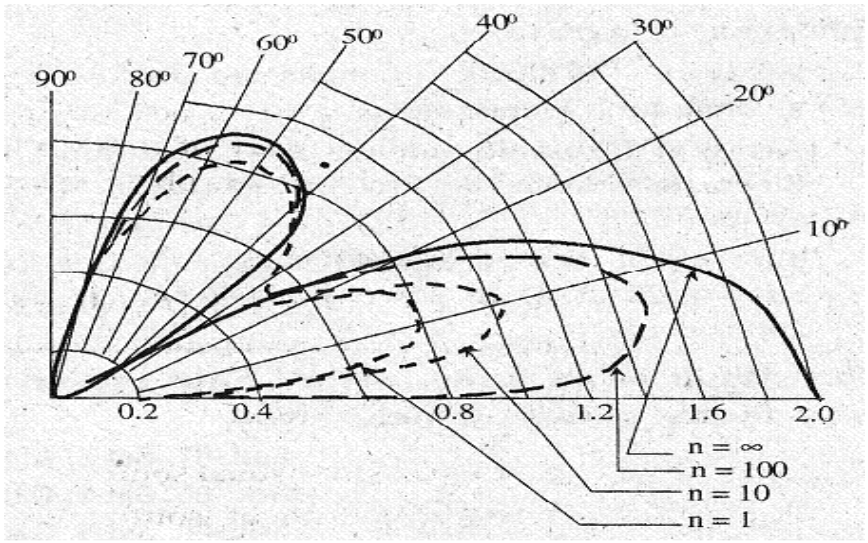


fig 3.3 Vertical pattern of a vertical dipole placed at a certain height above the earth

Effect of Earth on the Radiation Pattern of Horizontal Antenna

The effect of ground is obtained by multiplying free-space pattern and ground factor, i.e.
2cos
Sinφ

The first maxima in this pattern occurs at,

Sinφ = λ/4h (h> λ/4)

The effect of earth on the vertical pattern perpendicular to the axis of dipole is as shown in figure3.4.

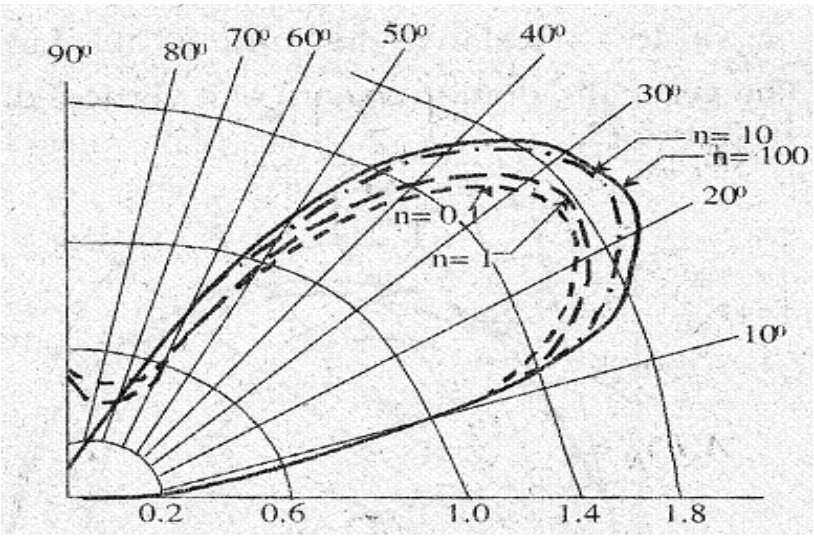


fig 3.4 Effect of the earth on vertical pattern of horizontal antenna

Antenna Measurements

Testing of real antennas is fundamental to antenna theory. All the antenna theory in the world doesn't add up to a hill of beans if the antennas under test don't perform as desired. **Antenna Measurements** is a science unto itself; as a very good antenna measurer once said to me "good antenna measurements don't just happen".

What exactly are we looking for when we test or measure antennas?
Basically, we want to measure many of the fundamental parameters listed on the Antenna Basics page. The most common and desired measurements are an antenna's radiation pattern including antenna gain and efficiency, the impedance or VSWR, the bandwidth, and the polarization.

The procedures and equipment used in antenna measurements are described in the following sections:

1. Required Equipment and Ranges
In this first section on Antenna Measurements, we look at the required equipment and types of "antenna ranges" used in modern antenna measurement systems.

2. Radiaton Pattern and Gain Measurements
The second antenna measurements section discusses how to perform the most fundamental antenna measurement - determining an antenna's radiation pattern and extracting the antenna gain.

3. Phase Measurements
The third antenna measurements section focuses on determining phase information from an antenna's radiation pattern. The phase is more important in terms of 'relative phase' (phase relative to other positions on the radiation pattern), not 'absolute phase'.

4. Polarization Measurements
The fourth antenna measurements section discusses techniques for determining the polarization of the antenna under test. These techniques are used to classify an antenna as linearly, circularly or elliptically polarized.

5. Impedance Measurements
The fifth antenna measurement section illustrates how to determine an antenna's impedance as a function of frequency. Here the focus is on the use of a Vector Network Analyzer (VNA).

6. Scale Model Measurements
The sixth antenna measurement section explains the useful concept of scale model measurements. This page illustrates how to obtain measurements when the physical size of the desired test is too large (or possibly, too small).

3. SAR (Specific Absorption Rate) Measurements
The final antenna measurement section illustrates the new field of SAR measurements and explains what SAR is. These measurements are critical in consumer electronics as antenna design consistently needs altered (or even degraded) in order to meet FCC SAR requirements.

Required Equipment in Antenna Measurements

For antenna test equipment, we will attempt to illuminate the test antenna (often called an Antenna-Under-Test) with a plane wave. This will be approximated by using a source (transmitting) antenna with known radiation pattern and characteristics, in such a way that the fields incident upon the test antenna are approximately plane waves. More will be discussed about this in the next section. The required equipment for antenna measurements include:

- A source antenna and transmitter- This antenna will have a known pattern that can be used to illuminate the test antenna
- A receiver system- This determines how much power is received by the test antenna
- A positioning system - This system is used to rotate the test antenna relative to the source antenna, to measure the radiation pattern as a function of angle.

A block diagram of the above equipment is shown in Figure 1.

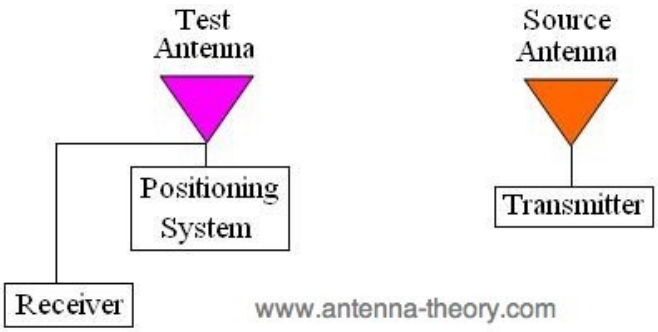


Figure 1. Diagram of required antenna measurement equipment.

These components will be briefly discussed. The **Source Antenna** should of course radiate well at the desired test frequency. It must have the desired polarization and a suitable beamwidth for the given antenna test range. Source antennas are often horn antennas, or a dipole antenna with a parabolic reflector.

The **Transmitting System** should be capable of outputing a stable known power. The output frequency should also be tunable (selectable), and reasonably stable (stable means that the frequency you get from the transmitter is close to the frequency you want).

The **Receiving System** simply needs to determine how much power is received from the test antenna. This can be done via a simple bolometer, which is a device for measuring the energy of incident electromagnetic waves. The receiving system can be more complex, with high quality amplifiers for low power measurements and more accurate detection devices.

The **Positioning System** controls the orientation of the test antenna. Since we want to measure the radiation pattern of the test antenna as a function of angle (typically in spherical coordinates), we need to rotate the test antenna so that the source antenna illuminates the test antenna from different angles. The positioning system is used for this purpose.

Once we have all the equipment we need (and an antenna we want to test), we'll need to place the equipment and perform the test in an antenna range, the subject of the next section.

The first thing we need to do an antenna measurement is a place to perform the measurement. Maybe you would like to do this in your garage, but the reflections from the walls, ceilings and

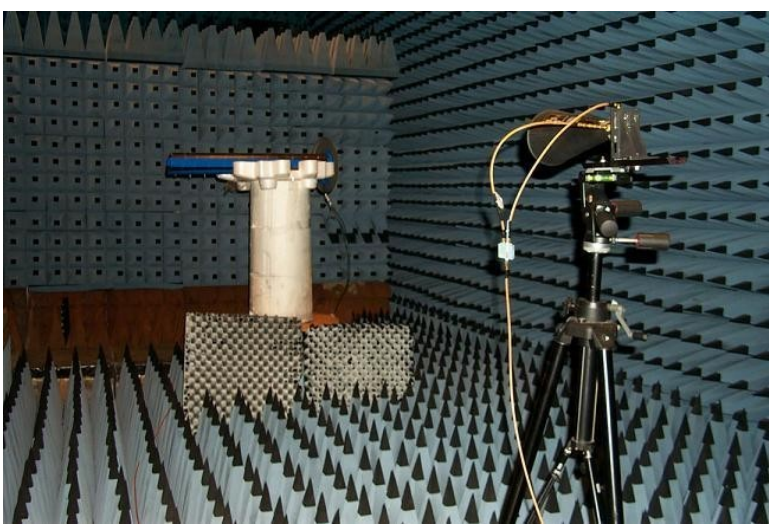
floor would make your measurements inaccurate. The ideal location to perform antenna measurements is somewhere in outer space, where no reflections can occur. However, because space travel is currently prohibitively expensive, we will focus on measurement places that are on the surface of the Earth. There are two main types of ranges, *Free Space Ranges* and *Reflection Ranges*. Reflection ranges are designed such that reflections add together in the test region to support a roughly planar wave. We will focus on the more common free space ranges.

Free Space Ranges

Free space ranges are antenna measurement locations designed to simulate measurements that would be performed in space. That is, all reflected waves from nearby objects and the ground (which are undesirable) are suppressed as much as possible. The most popular free space ranges are anechoic chambers, elevated ranges, and the compact range.

Anechoic Chambers

Anechoic chambers are indoor antenna ranges. The walls, ceilings and floor are lined with special electromagnetic wave absorbering material. Indoor ranges are desirable because the test conditions can be much more tightly controlled than that of outdoor ranges. The material is often jagged in shape as well, making these chambers quite interesting to see. The jagged triangle shapes are designed so that what is reflected from them tends to spread in random directions, and what is added together from all the random reflections tends to add incoherently and is thus suppressed further. A picture of an anechoic chamber is shown in the following picture, along with some test equipment:



The drawback to anechoic chambers is that they often need to be quite large. Often antennas need to be several wavelenghts away from each other at a minimum to simulate far-field conditions. Hence, it is desired to have anechoic chambers as large as possible, but cost and practical constraints often limit their size. Some defense contracting companies that measure the Radar Cross Section of large airplanes or other objects are known to have anechoic chambers the size of basketball courts, although this is not ordinary. universities with anechoic chambers typically have chambers that are 3-5 meters in length, width and height. Because of the size constraint, and because RF absorbing material typically works best at UHF and higher, anechoic chambers are most often used for frequencies above 300 MHz. Finally, the chamber should also be large enough that the source antenna's main lobe is not in view of the side walls, ceiling or floor.

Elevated Ranges

Elevated Ranges are outdoor ranges. In this setup, the source and antenna under test are mounted above the ground. These antennas can be on mountains, towers, buildings, or wherever one finds that is suitable. This is often done for very large antennas or at low frequencies (VHF and below, <100 MHz) where indoor measurements would be intractable. The basic diagram of an elevated range is shown in Figure 2.

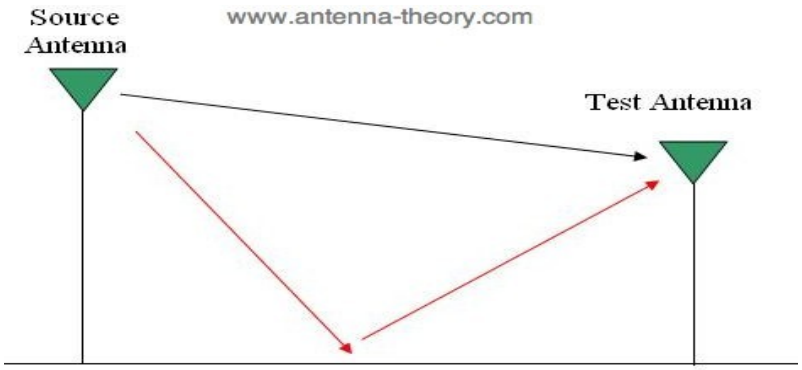


Figure 2. Illustration of elevated range.

The source antenna is not necessarily at a higher elevation than the test antenna, I just showed it that way here. The line of sight (LOS) between the two antennas (illustrated by the black ray in Figure 2) must be unobstructed. All other reflections (such as the red ray reflected from the ground) are undesirable. For elevated ranges, once a source and test antenna location are determined, the test operators then determine where the significant reflections will occur, and attempt to minimize the reflections from these surfaces. Often rf absorbing material is used for this purpose, or other material that deflects the rays away from the test antenna.

Compact Ranges

The source antenna must be placed in the far field of the test antenna. The reason is that the wave received by the test antenna should be a plane wave for maximum accuracy. Since antennas radiate spherical waves, the antenna needs to be sufficiently far such that the wave radiated from the source antenna is approximately a plane wave - see Figure 3.

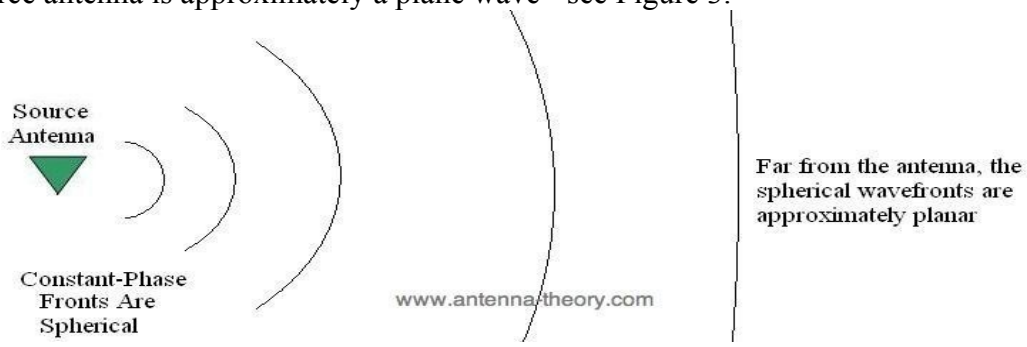


Figure 3. A source antenna radiates a wave with a spherical wavefront.

However, for indoor chambers there is often not enough separation to achieve this. One method to fix this problem is via a compact range. In this method, a source antenna is oriented towards a reflector, whose shape is designed to reflect the spherical wave in an approximately planar manner. This is very similar to the principle upon which a dish antenna operates. The basic operation is shown in Figure 4.

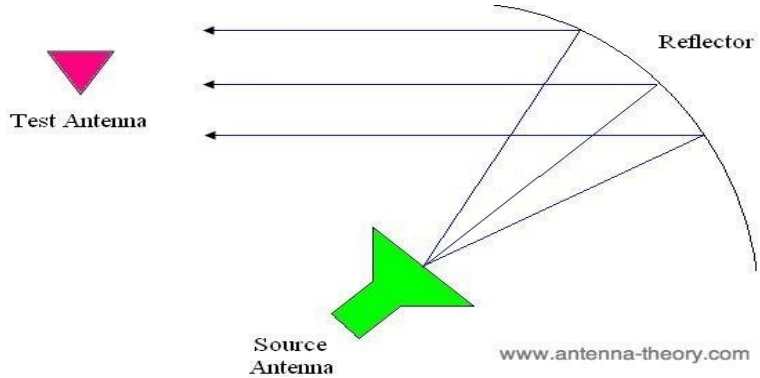


Figure 4. Compact Range - the spherical waves from the source antenna are reflected to be planar (collimated).

The length of the parabolic reflector is typically desired to be several times as large as the test antenna. The source antenna in Figure 4 is offset from the reflector so that it is not in the way of the reflected rays. Care must also be exercised in order to keep any direct radiation (mutual coupling) from the source antenna to the test antenna.

1. Antenna Radiation Pattern measurement

Now that we have our measurement equipment and an antenna range, we can perform some antenna measurements. We will use the source antenna to illuminate the antenna under test with a plane wave from a specific direction. The polarization and antenna gain (for the fields radiated toward the test antenna) of the source antenna should be known.

Due to reciprocity, the radiation pattern from the test antenna is the same for both the receive and transmit modes. Consequently, we can measure the radiation pattern in the receive or transmit mode for the test antenna. We will describe the receive case for the antenna under test.

The test antenna is rotated using the test antenna's positioning system. The received power is recorded at each position. In this manner, the magnitude of the radiation pattern of the test antenna can be determined. We will discuss phase measurements and polarization measurements later.

The coordinate system of choice for the radiation pattern is spherical coordinates.

Measurement Example

An example should make the process reasonably clear. Suppose the radiation pattern of a microstrip antenna is to be obtained. As is usual, lets let the direction the patch faces ('normal' to the surface of the patch) be towards the z-axis. Suppose the source antenna illuminates the test antenna from +y-direction, as shown in Figure 1.

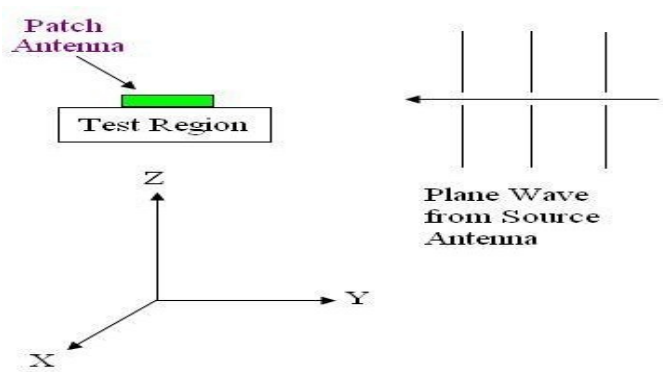


Figure 1. A patch antenna oriented towards the z-axis with a Source illumination from the +y-direction.

In Figure 1, the received power for this case represents the power from the angle: $(\theta,\phi) = (90°,90°)$. We record this power, change the position and record again. Recall that we only rotate the test antenna, hence it is at the same distance from the source antenna. The source power again comes from the same direction. Suppose we want to measure the radiation pattern normal to the patch's surface (straight above the patch). Then the measurement would look as shown in Figure 2.

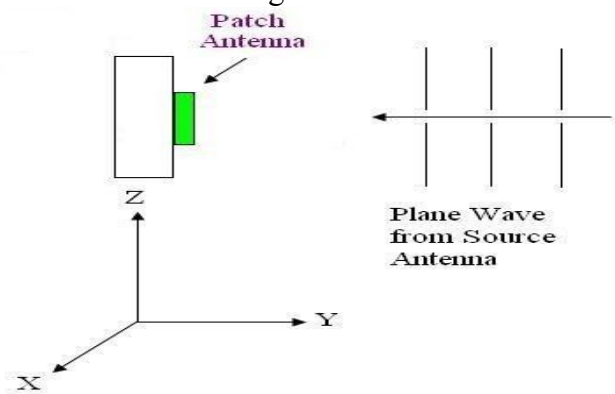


Figure 2. A patch antenna rotated to measure the radiation power at normal incidence.

In Figure 2, the positioning system rotating the antenna such that it faces the source of illumination. In this case, the received power comes from direction $(\theta,\phi) = (0°,0°)$. So by rotating the antenna, we can obtain "cuts" of the radiation pattern - for instance the E-plane cut or the H-plane cut. A "great circle" cut is when θ=0 and ϕ is allowed to vary from 0 to 360 degrees. Another common radiation pattern cut (a cut is a 2d 'slice' of a 3d radiation pattern) is when ϕ is fixed and θ varies from 0 to 180 degrees. By measuring the radiation pattern along certain slices or cuts, the 3d radiation pattern can be determined.

It must be stressed that the resulting radiation pattern is correct for a given polarization of the source antenna. For instance, if the source is horizontally polarized (see polarization of plane waves), and the test antenna is vertically polarized, the resulting radiation pattern will be zero everywhere. Hence, the radiation patterns are sometimes classified as H-pol (horizontal polarization) or V-pol (vertical polarization). See also cross-polarization.

In addition, the radiation pattern is a function of frequency. As a result, the measured radiation pattern is only valid at the frequency the source antenna is transmitting at. To obtain broadband measurements, the frequency transmitted must be varied to obtain this information.

2. Antenna Impedence Measurement

The impedance is fundamental to an antenna that operates at RF frequencies (high frequency). If the impedance of an antenna is not "close" to that of the transmission line, then very little power will be transmitted by the antenna (if the antenna is used in the transmit mode), or very little power will be received by the antenna (if used in the receive mode). Hence, without proper impedance (or an impedance matching network), out antenna will not work properly.

Before we begin, I'd like to point out that object placed around the antenna will alter its radiation pattern. As a result, its input impedance will be influenced by what is around it - i.e. the environment in which the antenna is tested. Consequently, for the best accuracy the impedance should be measured in an environment that will most closely resemble where it is intended to operate. For instance, if a blade antenna (which is basically a dipole shaped like a paddle) is to be

utilized on the top of a fuselage of an aiprlane, the test measurement should be performed on top of a cylinder type metallic object for maximum accuracy. The term driving point impedance is the input impedance measured in a particular environment, and self-impedance is the impedance of an antenna in free space, with no objects around to alter its radiation pattern.

Fortunately, impedance measurements are pretty easy if you have the right equipment. In this case, the right equipment is a Vector Network Analyzer (VNA). This is a measuring tool that can be used to measure the input impedance as a function of frequency. Alternatively, it can plot S11 (return loss), and the VSWR, both of which are frequency-dependent functions of the antenna impedance. The Agilent 8510 Vector Network Analyzer is shown in Figure 1.

Figure 1. The popular Agilent (HP) 8510 VNA.

Let's say we want to perform an impedance measurement from 400-500 MHz. Step 1 is to make sure that our VNA is specified to work over this frequency range. Network Analyzers work over specified frequency ranges, which go into the low MHz range (30 MHz or so) and up into the high GigaHertz range (110 GHz or so, depending on how expensive it is). Once we know our network analyzer is suitable, we can move on.

Next, we need to calibrate the VNA. This is much simpler than it sounds. We will take the cables that we are using for probes (that connect the VNA to the antenna) and follow a simple procedure so that the effect of the cables (which act as transmission lines) is calibrated out. To do this, typically your VNA will be supplied with a "cal kit" which contains a matched load (50 Ohms), an open circuit load and a short circuit load. We look on our VNA and scroll through the menus till we find a calibration button, and then do what it says. It will ask you to apply the supplied loads to the end of your cables, and it will record data so that it knows what to expect with your cables. You will apply the 3 loads as it tells you, and then your done. Its pretty simple, you don't even need to know what you're doing, just follow the VNA's instructions, and it will handle all the calculations.

Now, connect the VNA to the antenna under test. Set the frequency range you are interested in on the VNA. If you don't know how, just mess around with it till you figure it out, there are only so many buttons and you can't really screw anything up.

If you request output as an S-parameter (S11), then you are measuring the return loss. In this case, the VNA transmits a small amount of power to your antenna and measures how much power is reflected back to the VNA. A sample result (from the slotted waveguides page) might look something like:

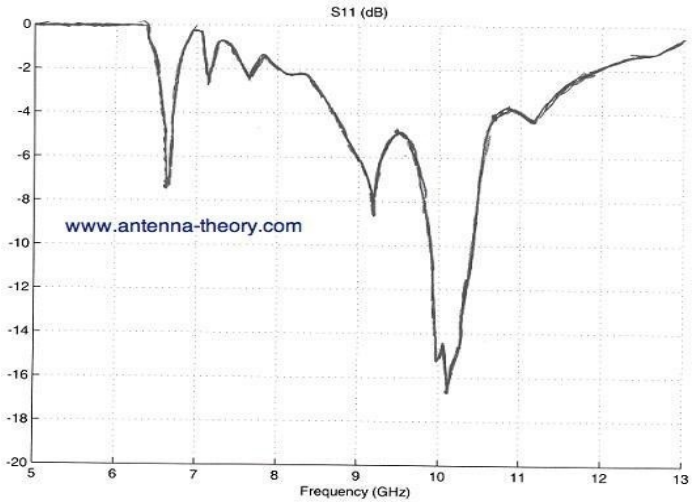


Figure 2. Example S11 measurement.

Note that the S-parameter is basically the magnitude of the reflection coefficient, which depends on the antenna impedance as well as the impedance of the VNA, which is typically 50 Ohms. So this measurement typically measures how close to 50 Ohms the antenna impedance is.

Another popular output is for the impedance to be measured on a Smith Chart. A Smith Chart is basically a graphical way of viewing input impedance (or reflection coefficient) that is easy to read. The center of the Smith Chart represented zero reflection coefficient, so that the antenna is perfectly matched to the VNA. The perimeter of the Smith Chart represents a reflection coefficient with a magnitude of 1 (all power reflected), indicating that the antenna is very poorly matched to the VNA. The magnitude of the reflection coefficient (which should be small for an antenna to receive or transmit properly) depends on how far from the center of the Smith Chart you are. As an example, consider Figure 3. The reflection coefficient is measured across a frequency range and plotted on a Smith Chart.

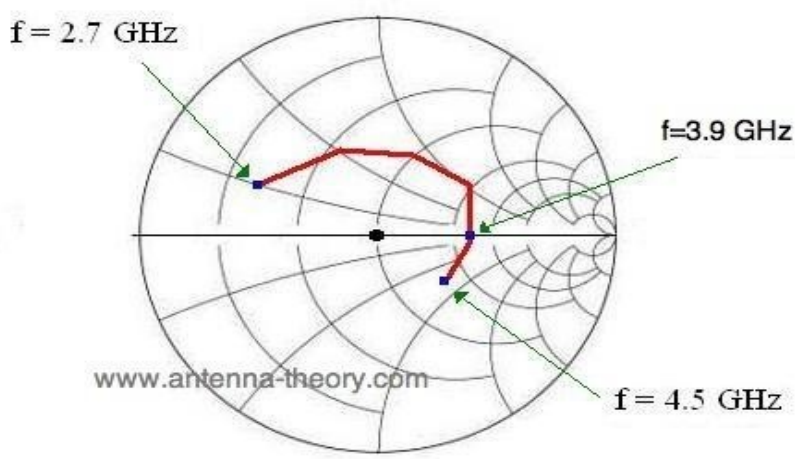


Figure 3. Smith Chart Graph of Impedance Measurement versus Frequency.

In Figure 3, the black circular graph is the Smith Chart. The black dot at the center of the Smith Chart is the point at which there would be zero reflection coefficient, so that the antenna's impedance is perfectly matched to the generator or receiver. The red curved line is the measurement. This is the impedance of the antenna, as the frequency is scanned from 2.7 GHz to 4.5 GHz. Each point on the line represents the impedance at a particular frequency. Points above the equator of the Smith Chart represent impedances that are inductive - they have a positive reactance (imaginary part). Points below the equator of the Smith Chart represent impedances that are capacitive - they have a negative reactance (for instance, the impedance would be something like Z = R - jX).

To further explain Figure 3, the blue dot below the equator in Figure 3 represents the impedance at f=4.5 GHz. The distance from the origin is the reflection coefficient, which can be estimated to have a magnitude of about 0.25 since the dot is 25% of the way from the origin to the outer perimeter.

As the frequency is decreased, the impedance changes. At f = 3.9 GHz, we have the second blue dot on the impedance measurement. At this point, the antenna is resonant, which means the impedance is entirely real. The frequency is scanned down until f=2.7 GHz, producing the locus of points (the red curve) that represents the antenna impedance over the frequency range. At f = GHz, the impedance is inductive, and the reflection coefficient is about 0.65, since it is closer to the perimeter of the Smith Chart than to the center.

In summary, the Smith Chart is a useful tool for viewing impedance over a frequency range in a concise, clear form.

Finally, the magnitude of the impedance could also be measured by measuring the VSWR (Voltage Standing Wave Ratio). The VSWR is a function of the magnitude of the reflection coefficient, so no phase information is obtained about the impedance (relative value of reactance divided by resistance). However, VSWR gives a quick way of estimated how much power is reflected by an antenna. Consequently, in antenna data sheets, VSWR is often specified, as in "VSWR: < 3:1 from 100-200 MHz". Using the formula for the VSWR, you can figure out that this menas that less than half the power is reflected from the antenna over the specified frequency range.
In summary, there are a bunch of ways to measure impedance, and a lot are a function of reflected power from the antenna. We care about the impedance of an antenna so that we can properly transfer the power to the antenna.
In the next Section, we'll look at scale model measurements.

3.Gain Measurement

On the previous page on measuring radiation patterns, we saw how the radiation pattern of an antenna can be measured. This is actually the "relative" radiation pattern, in that we don't know what the peak value of the gain actually is (we're just measuring the received power, so in a sense can figure out how directive an antenna is and the shape of the radiation pattern). In this page, we will focus on measuring the peak gain of an antenna - this information tells us how much power we can hope to receive from a given plane wave.
We can measure the peak gain using the Friis Transmission Equation and a "gain standard" antenna. A gain standard antenna is a test antenna with an accurately known gain and

polarization (typically linear). The most popular types of gain standard antennas are the thin half-wave dipole antenna (peak gain of 2.15 dB) and the pyramidal horn antenna (where the peak gain can be accurately calculated and is typically in the range of 15-25 dB). Consider the test setup shown in Figure 1. In this scenario, a gain standard antenna is used in the place of the test antenna, with the source antenna transmitting a fixed amount of power (*PT*). The gains of both of these antennas are accurately known.

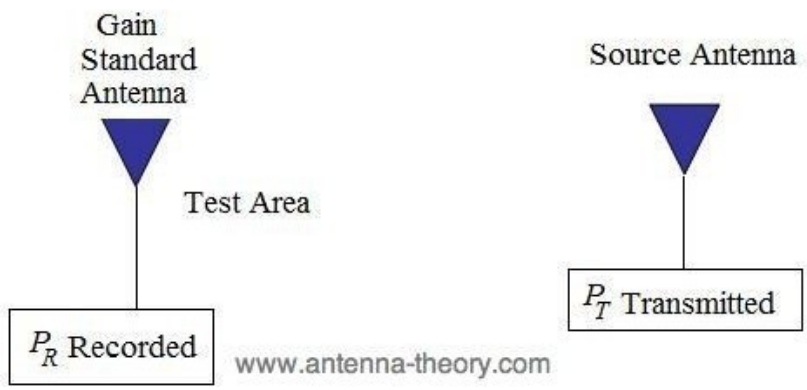


Figure 1. Record the received power from a gain standard antenna.

From the Friis transmission equation, we know that the power received (*PR*) is given by:

$$P_R = \frac{P_T G_T G_R \lambda^2}{(4\pi R)^2}$$

If we replace the gain standard antenna with our test antenna (as shown in Figure 2), then the only thing that changes in the above equation is *GR* - the gain of the receive antenna. The separation between the source and test antennas is fixed, and the frequency will be held constant as well.

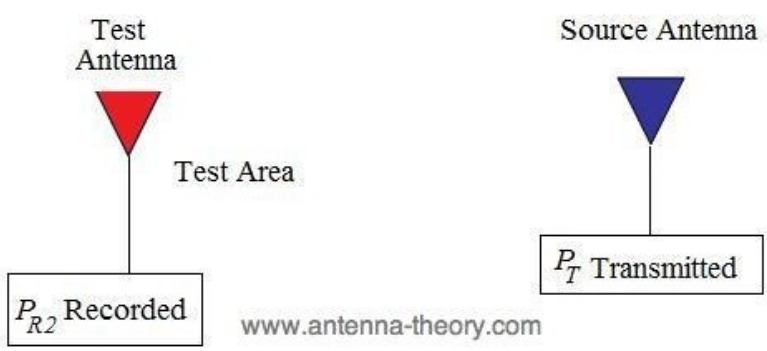


Figure 2. Record the received power with the test antenna (same source antenna).

Let the received power from the test antenna be *PR2*. If the gain of the test antenna is higher than the gain of the "gain standard" antenna, then the received power will increase. Using our measurements, we can easily calculate the gain of the test antenna. Let *Gg* be the gain of the "gain standard" antenna, *PR* be the power received with the gain antenna under test, and *PR2* be the power received with the test antenna. Then the gain of the test antenna (*GT*) is (in linear units):

$$G_T = G_G \frac{P_{R2}}{P_R}$$

The above equation uses linear units (non-dB). If the gain is to be specified in decibels, (power received still in Watts), then the equation becomes:

$$[G_T]_{dB} = [G_G]_{dB} + 10\log_{10}\left(\frac{P_{R2}}{P_R}\right)$$

And that is all that needs done to determine the gain for an antenna in a particular direction.

Efficiency and Directivity

Recall that the directivity can be calculated from the measured radiation pattern without regard to what the gain is. Typically this can be performed by approximated the integral as a finite sum, which is pretty simple.

Recall that the efficiency of an antenna is simply the ratio of the peak gain to the peak directivity:

$$\varepsilon = \frac{G}{D}$$

Hence, once we have measured the radiation pattern and the gain, the efficiency follows directly from these.

antenna temperature

To estimate antenna temperature one should know the power at antenna terminals. For this, a simple experiment is carried out using the spectrum analyzer and Low noise RF amplifier. The low noise RF amplifier in front of spectrum analyzer reduces the noise temperature of the receiving system, and then with this the estimated system temperature is mainly conributed by antenna temperature.

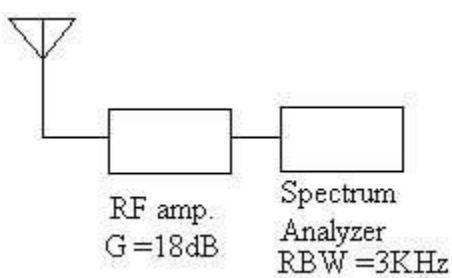


Fig: - Antenna temperature measurement setup

The RF amplifier used here is actually two cascaded stages of J310 amplifier used in jove receiver. The amplifier has gain of 18dB and noise figure of 3.71dB. Spectrum analyzer usually has high noise figure of the order of 20dB.

Let's assume the spectrum analyzer noise figure of 25dB. Then the overall receiving system (as shown in figure) will have the noise figure of 8.66dB. This will lead to noise temperature of 1840.1 Kelvin@ 290Kelvin. Thus the main cotributor to the system temperature in this setup will be the antenna itself!

It is observed that the Noise floor seen on spectrum analyzer is typically -94dBm.

So considering the 18dB gain of low noise amplifier in front of the spectrum analyzer,

power at antenna terminals = -94-18 = -112dBm = 6.3 x 10^-15 Watt.
Equating power to K*T*B we get,

Tsys= P/(K*B)= 0.152 x 10^6 Kelvin= 0.152 Million Kelvin.

Since the contribution to system temperature by the receiver is much less (~1800 Kelvin). so the system temperature can be approximated as antenna temperature at 20.1MHz. Tant=0.152 Million Kelvin.

The antenna temperature of 0.152 Million Kelvin is high, typically galactic background contribution to antenna temperature can be 50000 Kelvin. The high temperature of 0.152 Million Kelvin suggest some local noise causing increase in temperature. With this temperature one can receive only strong bursts of Jupiter. As for 10^6 Jansky Jupiter bursts power at antenna terminals will be -115dBm, still 3dB less than the antenna noise!

The proper way to measure antenna tempearture is the use of the noise sourses. So the above method will give a rough estimate.

UNIT V

Wave propagation

POLARIZATION

For maximum absorption of energy from the electromagnetic fields, the receiving antenna must be located in the plane of polarization. This places the conductor of the antenna at right angles to the magnetic lines of force moving through the antenna and parallel to the electric lines, causing maximum induction.

Normally, the plane of polarization of a radio wave is the plane in which the E field propagates with respect to the Earth. If the E field component of the radiated wave travels in a plane perpendicular to the Earth's surface (vertical), the radiation is said to be VERTICALLY POLARIZED, as shown in figure 2-5, view A. If the E field propagates in a plane parallel to the Earth's surface (horizontal), the radiation is said
to be HORIZONTALLY POLARIZED,

ATMOSPHERIC PROPAGATION

Within the atmosphere, radio waves can be reflected, refracted, and diffracted like light and heat waves.

Reflection

Radio waves may be reflected from various substances or objects they meet during travel between the transmitting and receiving sites. The amount of reflection depends on the reflecting material. Smooth metal surfaces of good electrical conductivity are efficient reflectors of radio waves. The surface of the Earth itself is a fairly good reflector. The radio wave is not reflected from a single point on the reflector but rather from an area on its surface. The size of the area required for reflection to take place depends on the wavelength of the radio wave and the angle at which the wave strikes the reflecting substance.

When radio waves are reflected from flat surfaces, a phase shift in the alternations of the wave occurs. Figure 2-7 shows two radio waves being reflected from the Earth's surface. Notice that the positive and negative alternations of radio waves (A) and (B) are in phase with each other in their paths toward the Earth's surface. After reflection takes place, however, the waves are approximately 180 degrees out of phase from their initial relationship. The amount of phase shift that occurs is not constant. It depends on the polarization of the wave and the angle at which the wave strikes the reflecting surface.

Radio waves that keep their phase relationships after reflection normally produce a stronger signal at the receiving site. Those that are received out of phase produce a weak or fading signal. The shifting in the phase relationships of reflected radio waves is one of the major reasons for fading. Fading will be discussed in more detail later in this chapter.

Refraction

Another phenomenon common to most radio waves is the bending of the waves as they move from one medium into another in which the velocity of propagation is different. This bending of the waves is called refraction. For example, suppose you are driving down a smoothly paved road at a constant speed and suddenly one wheel goes off onto the soft shoulder. The car tends to veer off to one side. The change of medium, from hard surface to soft shoulder, causes a change in speed or velocity. The tendency is for the car to change direction. This same principle applies to radio waves as changes occur in the medium through which they are passing. As an example, the radio wave shown in figure 2-8 is traveling through the Earth's atmosphere at a constant speed. As the wave enters the dense layer of electrically charged ions, the part of the wave that enters the new medium first travels faster than the parts of the wave that have not yet entered the new medium. This abrupt increase in velocity of the upper part of the wave causes the wave to bend back toward the Earth. This bending, or change of direction, is always toward the medium that has the lower velocity of propagation.

Radio waves passing through the atmosphere are affected by certain factors, such as temperature, pressure, humidity, and density. These factors can cause the radio waves to be refracted. This effect will be discussed in greater detail later in this chapter.

Diffraction

A radio wave that meets an obstacle has a natural tendency to bend around the obstacle as illustrated in figure 2-9. The bending, called diffraction, results in a change of direction of part of the wave energy from the normal line-of-sight path. This change makes it possible to receive energy around the edges of an obstacle as shown in view A or at some distances below the highest point of an obstruction, as shown in view B. Although diffracted rf energy usually is weak, it can still be detected by a suitable receiver. The principal effect of diffraction extends the radio range beyond the visible horizon. In certain cases, by using high power and very low frequencies, radio waves can be made to encircle the Earth by diffraction.

What is one of the major reasons for the fading of radio waves which have been reflected from a surface?

THE EFFECT OF THE EARTH'S ATMOSPHERE ON RADIO WAVES

This discussion of electromagnetic wave propagation is concerned mainly with the properties and effects of the medium located between the transmitting antenna and the receiving antenna. While radio waves traveling in free space have little outside influence affecting them, radio waves traveling within the Earth's atmosphere are affected by varying conditions. The influence exerted on radio waves by the

Earth's atmosphere adds many new factors to complicate what at first seems to be a relatively simple problem. These complications are because of a lack of uniformity within the Earth's atmosphere. Atmospheric conditions vary with changes in height, geographical location, and even with changes in time (day, night, season, year). A knowledge of the composition of the Earth's atmosphere is extremely important for understanding wave propagation.

The Earth's atmosphere is divided into three separate regions, or layers. They are the TROPOSPHERE, the STRATOSPHERE, and the IONOSPHERE.

TROPOSPHERE

The troposphere is the portion of the Earth's atmosphere that extends from the surface of the Earth to a height of about 3.7 miles (6 km) at the North Pole or the South Pole and 11.2 miles (18 km) at the equator. Virtually all weather phenomena take place in the troposphere. The temperature in this region decreases rapidly with altitude, clouds form, and there may be much turbulence because of variations in temperature, density, and pressure. These conditions have a great effect on the propagation of radio waves, which will be explained later in this chapter.

STRATOSPHERE

The stratosphere is located between the troposphere and the ionosphere. The temperature throughout this region is considered to be almost constant and there is little water vapor present. The stratosphere has relatively little effect on radio waves because it is a relatively calm region with little or no temperature changes.

IONOSPHERE

The ionosphere extends upward from about 31.1 miles (50 km) to a height of about 250 miles (402 km). It contains four cloud-like layers of electrically charged ions, which enable radio waves to be propagated to great distances around the Earth. This is the most important region of the atmosphere for long distance point-to-point communications. This region will be discussed in detail a little later in this chapter.

RADIO WAVE TRANSMISSION

There are two principal ways in which electromagnetic (radio) energy travels from a transmitting antenna to a receiving antenna. One way is by GROUND WAVES and the other is by SKY WAVES.

Ground waves are radio waves that travel near the surface of the Earth (surface and space waves). Sky waves are radio waves that are reflected back to Earth from the ionosphere.

Ground Waves The ground wave is actually composed of two separate component waves. These are known as the SURFACE WAVE and the SPACE WAVE (fig. 2-11). The determining factor in whether a ground wave component is classified as a space wave or a surface wave is simple. A surface wave travels along the surface of the Earth. A space wave travels over the surface.

SURFACE WAVE.—The surface wave reaches the receiving site by traveling along the surface of the ground as shown in figure 2-12. A surface wave can follow the contours of the Earth because of the process of diffraction. When a surface wave meets an object and the dimensions of the object do not exceed its wavelength, the wave tends to curve or bend around the object. The smaller the object, the more pronounced the diffractive action will be.

As a surface wave passes over the ground, the wave induces a voltage in the Earth. The induced voltage takes energy away from the surface wave, thereby weakening, or attenuating, the wave as it moves away from the transmitting antenna. To reduce the attenuation, the amount of induced voltage must be reduced. This is done by using vertically polarized waves that minimize the extent to which the electric field of the wave is in contact with the Earth. When a surface wave is horizontally polarized, the electric field of the wave is parallel with the surface of the Earth and, therefore, is constantly in contact with it. The wave is then completely attenuated within a short distance from the transmitting site. On the other hand, when the surface wave is vertically polarized, the electric field is vertical to the Earth and merely dips into and out of the Earth's surface. For this reason, vertical polarization is vastly superior to horizontal polarization for surface wave propagation.

The attenuation that a surface wave undergoes because of induced voltage also depends on the electrical properties of the terrain over which the wave travels. The best type of surface is one that has good electrical conductivity. The better the conductivity, the less the attenuation. Table 2-2 gives the relative conductivity of various surfaces of the Earth.

Table 2-2.—Surface Conductivity

SURFACE	RELATIVE CONDUCTIVITY
Sea water	Good
Flat, loamy soil	Fair
Large bodies of fresh water	Fair
Rocky terrain	Poor
Desert	Poor
Jungle	Unusable

Another major factor in the attenuation of surface waves is frequency. Recall from earlier discussions on wavelength that the higher the frequency of a radio wave, the shorter its wavelength will be. These high frequencies, with their shorter wavelengths, are not normally diffracted but are absorbed by the Earth at points relatively close to the transmitting site. You can assume, therefore, that as the frequency of a surface wave is increased, the more rapidly the surface wave will be absorbed, or attenuated, by the Earth. Because of this loss by attenuation, the surface wave is impractical for long-distance transmissions at frequencies above 2 megahertz. On the other hand, when the frequency of a surface wave is low enough to have a

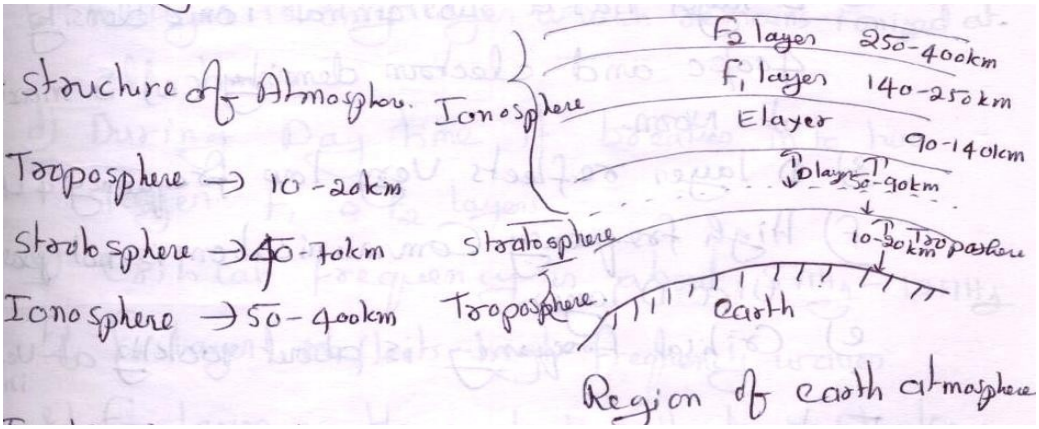
very long wavelength, the Earth appears to be very small, and diffraction is sufficient for propagation well beyond the horizon. In fact, by lowering the transmitting frequency into the very low frequency (vlf) range and using very high-powered transmitters, the surface wave can be propagated great distances. The Navy's extremely high-powered vlf transmitters are actually capable of transmitting surface wave signals around the Earth and can provide coverage to naval units operating anywhere at sea.

SPACE WAVE.—The space wave follows two distinct paths from the transmitting antenna to the receiving antenna—one through the air directly to the receiving antenna, the other reflected from the ground to the receiving antenna. This is illustrated in figure 2-13. The primary path of the space wave is directly from the transmitting antenna to the receiving antenna. So, the receiving antenna must be located within the radio horizon of the transmitting antenna. Because space waves are refracted slightly, even when propagated through the troposphere, the radio horizon is actually about one-third farther than the line-of-sight or natural horizon.
Although space waves suffer little ground attenuation, they nevertheless are susceptible to fading. This is because space waves actually follow two paths of different lengths (direct path and ground reflected path) to the receiving site and, therefore, may arrive in or out of phase. If these two component waves are received in phase, the result is a reinforced or stronger signal. Likewise, if they are received out of phase, they tend to cancel one another, which results in a weak or fading signal.

Sky Wave
The sky wave, often called the ionospheric wave, is radiated in an upward direction and returned to Earth at some distant location because of refraction from the ionosphere. This form of propagation is relatively unaffected by the Earth's surface and can propagate signals over great distances. Usually the high frequency (hf) band is used for sky wave propagation. The following in-depth study of the ionosphere and its effect on sky waves will help you to better understand the nature of sky wave propagation.

STRUCTURE OF THE IONOSPHERE
As we stated earlier, the ionosphere is the region of the atmosphere that extends from about 30 miles above the surface of the Earth to about 250 miles. It is appropriately named the ionosphere because it consists of several layers of electrically charged gas atoms called ions. The ions are formed by a process called ionization.
Ionization
Ionization occurs when high energy ultraviolet light waves from the sun enter the ionospheric region of the atmosphere, strike a gas atom, and literally knock an electron free from its parent atom. A normal atom is electrically neutral since it contains both a positive proton in its nucleus and a negative orbiting electron. When the negative electron is knocked free from the atom, the atom becomes positively charged (called a positive ion) and remains in space along with the free electron, which is negatively charged. This process of upsetting electrical neutrality is known as IONIZATION. The free negative electrons subsequently absorb part of the ultraviolet energy, which initially freed them from their atoms. As the ultraviolet light wave continues to produce positive ions and negative electrons, its intensity decreases because of the absorption of energy by the free electrons, and an ionized layer is formed. The rate at which ionization occurs depends on the density of atoms in the atmosphere and the intensity of the ultraviolet light wave, which varies with the activity of the sun.
Since the atmosphere is bombarded by ultraviolet light waves of different frequencies, several ionized layers are formed at different altitudes. Lower frequency ultraviolet waves penetrate the

atmosphere the least; therefore, they produce ionized layers at the higher altitudes. Conversely, ultraviolet waves of higher frequencies penetrate deeper and produce layers at the lower altitudes.



An important factor in determining the density of ionized layers is the elevation angle of the sun, which changes frequently. For this reason, the height and thickness of the ionized layers vary, depending on the time of day and even the season of the year. Recombination Recall that the process of ionization involves ultraviolet light waves knocking electrons free from their atoms. A reverse process called RECOMBINATION occurs when the free electrons and positive ions collide with each other. Since these collisions are inevitable, the positive ions return to their original neutral atom state.

The recombination process also depends on the time of day. Between the hours of early morning and late afternoon, the rate of ionization exceeds the rate of recombination. During this period, the ionized layers reach their greatest density and exert maximum influence on radio waves. During the late afternoon and early evening hours, however, the rate of recombination exceeds the rate of ionization, and the density of the ionized layers begins to decrease. Throughout the night, density continues to decrease, reaching a low point just before sunrise.

Four Distinct Layers

The ionosphere is composed of three layers designated D, E, and F, from lowest level to highest level as shown in figure, The F layer is further divided into two layers designated F1 (the lower layer) and F2 (the higher layer). The presence or absence of these layers in the ionosphere and their height above the Earth varies with the position of the sun. At high noon, radiation in the ionosphere directly above a given point is greatest. At night it is minimum. When the radiation is removed, many of the particles that were ionized recombine. The time interval between these conditions finds the position and number of the ionized layers within the ionosphere changing. Since the position of the sun varies daily, monthly, and yearly, with respect to a specified point on Earth, the exact position and number of layers present are extremely difficult to determine. However, the following general statements can be made:

The D layer ranges from about 30 to 55 miles. Ionization in the D layer is low because it is the lowest region of the ionosphere. This layer has the ability to refract signals of low frequencies. High frequencies pass right through it and are attenuated. After sunset, the D layer disappears because of the rapid recombination of ions.

b. The E layer limits are from about 55 to 90 miles. This layer is also known as the Kennelly-Heaviside layer, because these two men were the first to propose its existence. The rate of ionic recombination in this layer is rather rapid after sunset and the layer is almost gone by midnight. This layer has the ability to refract signals as high as 20 megahertz. For this reason, it is valuable

for communications in ranges up to about 1500 miles.

c. The F layer exists from about 90 to 240 miles. During the daylight hours, the F layer separates into two layers, the F1 and F2 layers. The ionization level in these layers is quite high and varies widely during the day. At noon, this portion of the atmosphere is closest to the sun and the degree of ionization is maximum. Since the atmosphere is rarefied at these heights, recombination occurs slowly after sunset. Therefore, a fairly constant ionized layer is always present. The F layers are responsible for high-frequency, long distance transmission.

REFRACTION IN THE IONOSPHERE

When a radio wave is transmitted into an ionized layer, refraction, or bending of the wave, occurs. As we discussed earlier, refraction is caused by an abrupt change in the velocity of the upper part of a radio wave as it strikes or enters a new medium. The amount of refraction that occurs depends on three main factors: (1) the density of ionization of the layer, (2) the frequency of the radio wave, and (3) the angle at which the wave enters the layer.

Density of Layer

Figure 2-15 illustrates the relationship between radio waves and ionization density. Each ionized layer has a central region of relatively dense ionization, which tapers off in intensity both above and below the maximum region. As a radio wave enters a region of INCREASING ionization, the increase in velocity of the upper part of the wave causes it to be bent back TOWARD the Earth. While the wave is in the highly dense center portion of the layer, however, refraction occurs more slowly because the density of ionization is almost uniform. As the wave enters into the upper part of the layer of DECREASING ionization, the velocity of the upper part of the wave decreases, and the wave is bent AWAY from the Earth.

If a wave strikes a thin, very highly ionized layer, the wave may be bent back so rapidly that it will appear to have been reflected instead of refracted back to Earth. To reflect a radio wave, the highly ionized layer must be approximately no thicker than one wavelength of the radio wave. Since the ionized layers are often several miles thick, ionospheric reflection is more likely to occur at long wavelengths (low frequencies).

Frequency

For any given time, each ionospheric layer has a maximum frequency at which radio waves can be transmitted vertically and refracted back to Earth. This frequency is known as the CRITICAL FREQUENCY. It is a term that you will hear frequently in any discussion of radio wave propagation.

Radio waves transmitted at frequencies higher than the critical frequency of a given layer will pass through the layer and be lost in space; but if these same waves enter an upper layer with a higher critical frequency, they will be refracted back to Earth. Radio waves of frequencies lower than the critical frequency will also be refracted back to Earth unless they are absorbed or have been refracted from a lower layer. The lower the frequency of a radio wave, the more rapidly the wave is refracted by a given degree of ionization. Figure 2-16 shows three separate waves of different frequencies entering an ionospheric layer at the same angle. Notice that the 5-megahertz wave is refracted quite sharply. The 20-megahertz wave is refracted less sharply and returned to Earth at a greater distance. The 100-megahertz wave is obviously greater than the critical frequency for that ionized layer and, therefore, is not refracted but is passed into space.

Angle of Incidence

The rate at which a wave of a given frequency is refracted by an ionized layer depends on the angle at which the wave enters the layer. Figure 2-17 shows three radio waves of the same frequency entering a layer at different angles. The angle at which wave A strikes the layer is too nearly vertical for the wave to be refracted to Earth. As the wave enters the layer, it is bent

slightly but passes through the layer and is lost. When the wave is reduced to an angle that is less than vertical (wave B), it strikes the layer and is refracted back to Earth. The angle made by wave B is called the CRITICAL ANGLE for that particular frequency. Any wave that leaves the antenna at an angle greater than the critical angle will penetrate the ionospheric layer for that frequency and then be lost in space. Wave C strikes the ionosphere at the smallest angle at which the wave can be refracted and still return to Earth. At any smaller angle, the wave will be refracted but will not return to Earth.

As the frequency of the radio wave is increased, the critical angle must be reduced for refraction to occur. This is illustrated in figure 2-18. The 2-megahertz wave strikes the layer at the critical angle for that frequency and is refracted back to Earth. Although the 5-megahertz wave (broken line) strikes the ionosphere at a lesser angle, it nevertheless penetrates the layer and is lost. As the angle is lowered from the vertical, however, a critical angle for the 5-megahertz wave is reached, and the wave is then refracted to Earth

Skip Distance/Skip Zone

In figure 2-19, note the relationship between the sky wave skip distance, the skip zone, and the ground wave coverage. The SKIP DISTANCE is the distance from the transmitter to the point where the sky wave is first returned to Earth. The size of the skip distance depends on the frequency of the wave, the angle of incidence, and the degree of ionization present.

The SKIP ZONE is a zone of silence between the point where the ground wave becomes too weak for reception and the point where the sky wave is first returned to Earth. The size of the skip zone depends on the extent of the ground wave coverage and the skip distance. When the ground wave coverage is great enough or the skip distance is short enough that no zone of silence occurs, there is no skip zone. Occasionally, the first sky wave will return to Earth within the range of the ground wave. If the sky wave and ground wave are nearly of equal intensity, the sky wave alternately reinforces and cancels the ground wave, causing severe fading. This is caused by the phase difference between the two waves, a result of the longer path traveled by the sky wave.

PROPAGATION PATHS

The path that a refracted wave follows to the receiver depends on the angle at which the wave strikes the ionosphere. You should remember, however, that the rf energy radiated by a transmitting antenna spreads out with distance. The energy therefore strikes the ionosphere at many different angles rather than a single angle. After the rf energy of a given frequency enters an ionospheric region, the paths that this energy might follow are many. It may reach the receiving antenna via two or more paths through a single layer. It may also, reach the receiving antenna over a path involving more than one layer, by multiple hops between the ionosphere and Earth, or by any combination of these paths.

When the angle is relatively low with respect to the horizon (ray 1), there is only slight penetration of the layer and the propagation path is long. When the angle of incidence is increased (rays 2 and 3), the rays penetrate deeper into the layer but the range of these rays decreases. When a certain angle is reached (ray 3), the penetration of the layer and rate of refraction are such that the ray is first returned to Earth at a minimal distance from the transmitter. Notice, however, that ray 3 still manages to reach the receiving site on its second refraction (called a hop) from the ionospheric layer.

As the angle is increased still more (rays 4 and 5), the rf energy penetrates the central area of maximum ionization of the layer. These rays are refracted rather slowly and are eventually returned to Earth at great distances. As the angle approaches vertical incidence (ray 6), the ray is not returned at all, but passes on through the layer.

ABSORPTION IN THE IONOSPHERE

Many factors affect a radio wave in its path between the transmitting and receiving sites. The factor that has the greatest adverse effect on radio waves is ABSORPTION. Absorption results in the loss of energy of a radio wave and has a pronounced effect on both the strength of received signals and the ability to communicate over long distances.

You learned earlier in the section on ground waves that surface waves suffer most of their absorption losses because of ground-induced voltage. Sky waves, on the other hand, suffer most of their absorption losses because of conditions in the ionosphere. Note that some absorption of sky waves may also occur at lower atmospheric levels because of the presence of water and water vapor. However, this becomes important only at frequencies above 10,000 megahertz. 2-26

Most ionospheric absorption occurs in the lower regions of the ionosphere where ionization density is greatest. As a radio wave passes into the ionosphere, it loses some of its energy to the free electrons and ions. If these high-energy free electrons and ions do not collide with gas molecules of low energy, most of the energy lost by the radio wave is reconverted into electromagnetic energy, and the wave continues to be propagated with little change in intensity. However, if the high-energy free electrons and ions do collide with other particles, much of this energy is lost, resulting in absorption of the energy from the wave. Since absorption of energy depends on collision of the particles, the greater the density of the ionized layer, the greater the probability of collisions; therefore, the greater the absorption. The highly dense D and E layers provide the greatest absorption of radio waves.

Because the amount of absorption of the sky wave depends on the density of the ionosphere, which varies with seasonal and daily conditions, it is impossible to express a fixed relationship between distance and signal strength for ionospheric propagation. Under certain conditions, the absorption of energy is so great that communicating over any distance beyond the line of sight is difficult.

FADING

The most troublesome and frustrating problem in receiving radio signals is variations in signal strength, most commonly known as FADING. There are several conditions that can produce fading. When a radio wave is refracted by the ionosphere or reflected from the Earth's surface, random changes in the polarization of the wave may occur. Vertically and horizontally mounted receiving antennas are designed to receive vertically and horizontally polarized waves, respectively. Therefore, changes in polarization cause changes in the received signal level because of the inability of the antenna to receive polarization changes Fading also results from absorption of the rf energy in the ionosphere. Absorption fading occurs for a longer period than other types of fading, since absorption takes place slowly.

Usually, however, fading on ionospheric circuits is mainly a result of multipath propagation.

Multipath Fading

MULTIPATH is simply a term used to describe the multiple paths a radio wave may follow between transmitter and receiver. Such propagation paths include the ground wave, ionospheric refraction, reradiation by the ionospheric layers, reflection from the Earth's surface or from more than one ionospheric layer, etc. Figure 2-21 shows a few of the paths that a signal can travel between two sites in a typical circuit. One path, XYZ, is the basic ground wave. Another path, XEA, refracts the wave at the E layer and passes it on to the receiver at A. Still another path, XFZFA, results from a greater angle of incidence and two refractions from the F layer. At point Z, the received signal is a combination of the ground wave and the sky wave. These two signals having traveled different paths arrive at point Z at different times. Thus, the arriving waves may or may not be in phase with each other. Radio waves that are received in phase reinforce each other and produce a stronger signal at the receiving site. Conversely, those that are received out of phase produce a weak or fading signal. Small alternations in the transmission path may change the phase relationship of the two signals, causing periodic fading. This condition occurs at point

A. At this point, the double-hop F layer signal may be in or out of phase with the signal arriving from the E layer.

Multipath fading may be minimized by practices called SPACE DIVERSITY and FREQUENCY DIVERSITY. In space diversity, two or more receiving antennas are spaced some distance apart. Fading does not occur simultaneously at both antennas; therefore, enough output is almost always available from one of the antennas to provide a useful signal. In frequency diversity, two transmitters and two receivers are used, each pair tuned to a different frequency, with the same information being transmitted simultaneously over both frequencies. One of the two receivers will almost always provide a useful signal.

Selective Fading

Fading resulting from multipath propagation is variable with frequency since each frequency arrives at the receiving point via a different radio path. When a wide band of frequencies is transmitted simultaneously, each frequency will vary in the amount of fading. This variation is called SELECTIVE FADING. When selective fading occurs, all frequencies of the transmitted signal do not retain their original phases and relative amplitudes. This fading causes severe distortion of the signal and limits the total signal transmitted.

Maximum Usable Frequency

As we discussed earlier, the higher the frequency of a radio wave, the lower the rate of refraction by an ionized layer. Therefore, for a given angle of incidence and time of day, there is a maximum frequency that can be used for communications between two given locations. This frequency is known as the MAXIMUM USABLE FREQUENCY (muf).

Waves at frequencies above the muf are normally refracted so slowly that they return to Earth beyond the desired location, or pass on through the ionosphere and are lost. You should understand, however, that use of an established muf certainly does not guarantee successful communications between a transmitting site and a receiving site. Variations in the ionosphere may occur at any time and consequently raise or lower the predetermined muf. This is particularly true for radio waves being refracted by the highly variable F2 layer. The muf is highest around noon when ultraviolet light waves from the sun are the most intense. It then drops rather sharply as recombination begins to take place.

the refractive index given by.

$$n = \frac{\sin\phi_i}{\sin\phi_r} = \sqrt{1 - \frac{81N}{f^2}}.$$

* for radio wave to return back to the earth angle of refraction must be 90° i.e $\phi_r = 90^\circ$

for this Condition N becomes N_{Max} and f becomes f_{muf}

$$\therefore \frac{\sin\phi_i}{\sin 90^\circ} = \sqrt{1 - \frac{81N_{Max}}{f_{muf}^2}}$$

$$\therefore \sin\phi_i = \sqrt{1 - \frac{81 N_{max}}{f_{Muf}^2}}$$

Squaring both sides

$$\sin^2\phi_i = 1 - \frac{81 N_{max}}{f_{Muf}^2}$$

$$\therefore \frac{81 N_{max}}{f_{Muf}^2} = 1 - \sin^2\phi_i = \cos^2\phi_i.$$

but $f_{cr} = \sqrt{81 N_{max}}$

$$\therefore \frac{f_{cr}^2}{f_{Muf}^2} = \cos^2\phi_i$$

$$\therefore f_{Muf} = \frac{f_{cr}}{\cos\phi_i} = \sec\phi_i \, f_{cr}$$

$\therefore$ The f_{Muf} is always greater than f_{cr} of the layer by the factor $\sec\phi_i$. This is called Secant law.

Lowest Usable Frequency

As there is a maximum operating frequency that can be used for communications between two points, there is also a minimum operating frequency. This is known as the LOWEST USABLE FREQUENCY (luf). As the frequency of a radio wave is lowered, the rate of refraction increases. So a wave whose frequency is below the established luf is refracted back to Earth at a shorter distance than desired, as shown in figure

The transmission path that results from the rate of refraction is not the only factor that determines the luf. As a frequency is lowered, absorption of the radio wave increases. A wave whose frequency is too low is absorbed to such an extent that it is too weak for reception. Likewise, atmospheric noise is greater at lower frequencies; thus, a low-frequency radio wave may have an unacceptable signal-to-noise ratio. For a given angle of incidence and set of ionospheric conditions, the luf for successful communications between two locations depends on the refraction properties of the ionosphere, absorption considerations, and the amount of atmospheric noise present.

Optimum Working Frequency

Neither the muf nor the luf is a practical operating frequency. While radio waves at the luf can be refracted back to Earth at the desired location, the signal-to-noise ratio is still much lower than at the higher frequencies, and the probability of multipath propagation is much greater. Operating at or near the muf can result in frequent signal fading and dropouts when ionospheric variations alter the length of the transmission path. The most practical operating frequency is one that you can rely on with the least amount of problems. It should be high enough to avoid the problems of multipath, absorption, and noise encountered at the lower frequencies; but not so high as to result in the adverse effects of rapid changes in the ionosphere. A frequency that meets the above criteria has been established and is known as the OPTIMUM WORKING FREQUENCY. It is abbreviated "fot" from the initial letters of the French words for optimum working frequency,

"frequence optimum de travail." The fot is roughly about 85 percent of the muf but the actual percentage varies and may be either considerably more or less than 85 percent.