MUTHAYAMMAL ENGINEERING COLLEGE
(An Autonomous Institution)
Rasipuram - 637 408

COURSE CODE & TITLE : 19ECC08 & Antenna System Engineering

Presentation by

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SYLLABUS

- **UNIT I ANTENNA FUNDAMENTALS**

Radiation from antenna, Basic antenna parameters – Radiation pattern, Radiation intensity, Beam area, Beam solid angle, Band width, Beam width, Directivity, Gain, Antenna aperture, Effective height, Effective aperture, Radiation Resistance, Input Impedance. Matching – Baluns, Polarization, Polarization mismatch, Antenna noise temperature, Radiation from Half wave dipole, Folded dipole.

- **UNIT II ANTENNA ARRAYS**

Antenna Arrays, Expression for electric field from two element and N element Array: Broad-side array and End-Fire array - Pattern Multiplication- Concept of Adaptive array and Binomial array.

SYLLABUS

- **UNIT III APERTURE AND SLOT ANTENNAS**

Uniqueness theorem, Radiation from an elemental area of a plane wave (Huygen’s Source), Radiation from rectangular apertures, Horn antenna - Types, Parabolic reflector antennas and its feed systems, Aperture blockage, Slot antennas, Method of feeding slot antennas-Microstrip antennas – Radiation mechanism – Application, Numerical tool for antenna analysis.

- **UNIT IV SPECIAL ANTENNAS AND ANTENNA MEASUREMENTS**

Yagi-Uda Antenna - Principle of frequency independent antennas –Spiral antenna, Helical antenna, Log Periodic Dipole Array - Reconfigurable antenna, Active antenna, Antenna Measurements - Test Ranges, Measurement of Gain, Radiation pattern, Polarization, VSWR. Directivity

SYLLABUS

- **UNIT V PROPAGATION OF RADIO WAVES**

Modes of propagation, Structure of atmosphere, Ground wave propagation, Tropospheric propagation, Duct propagation, Troposcatter propagation , Flat earth and Curved earth concept Sky wave propagation – Virtual height, critical frequency , Maximum usable frequency – Skip distance, Fading , Multi hop propagation.

- **TEXT BOOKS**

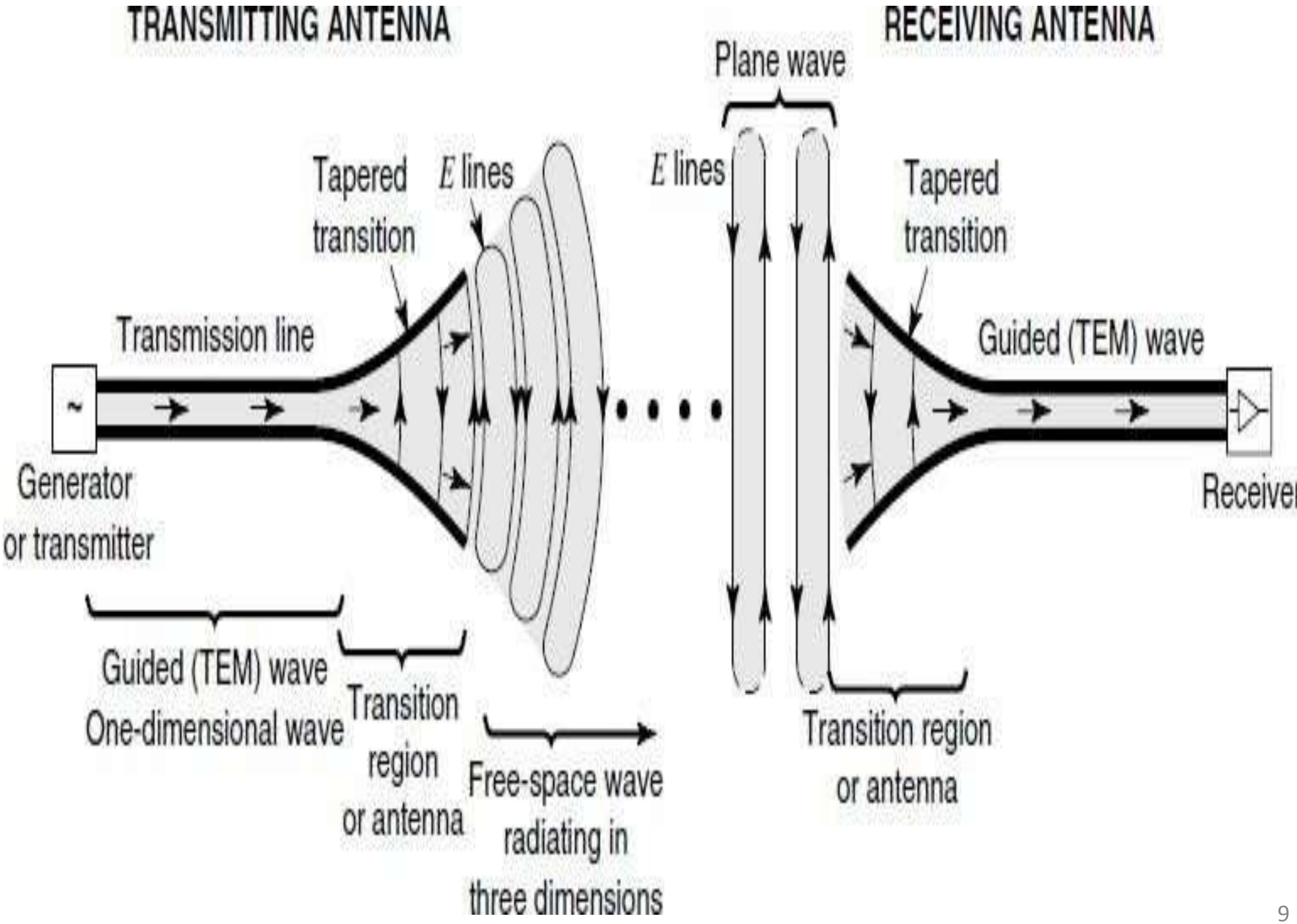
1. "Antennas for all Applications", McGraw Hill, 2005 by John D Kraus

2. "Antennas and Radiowave Propagation", McGraw Hill, 1985 by R.E.Collin

Unit – I

ANTENNA FUNDAMENTALS

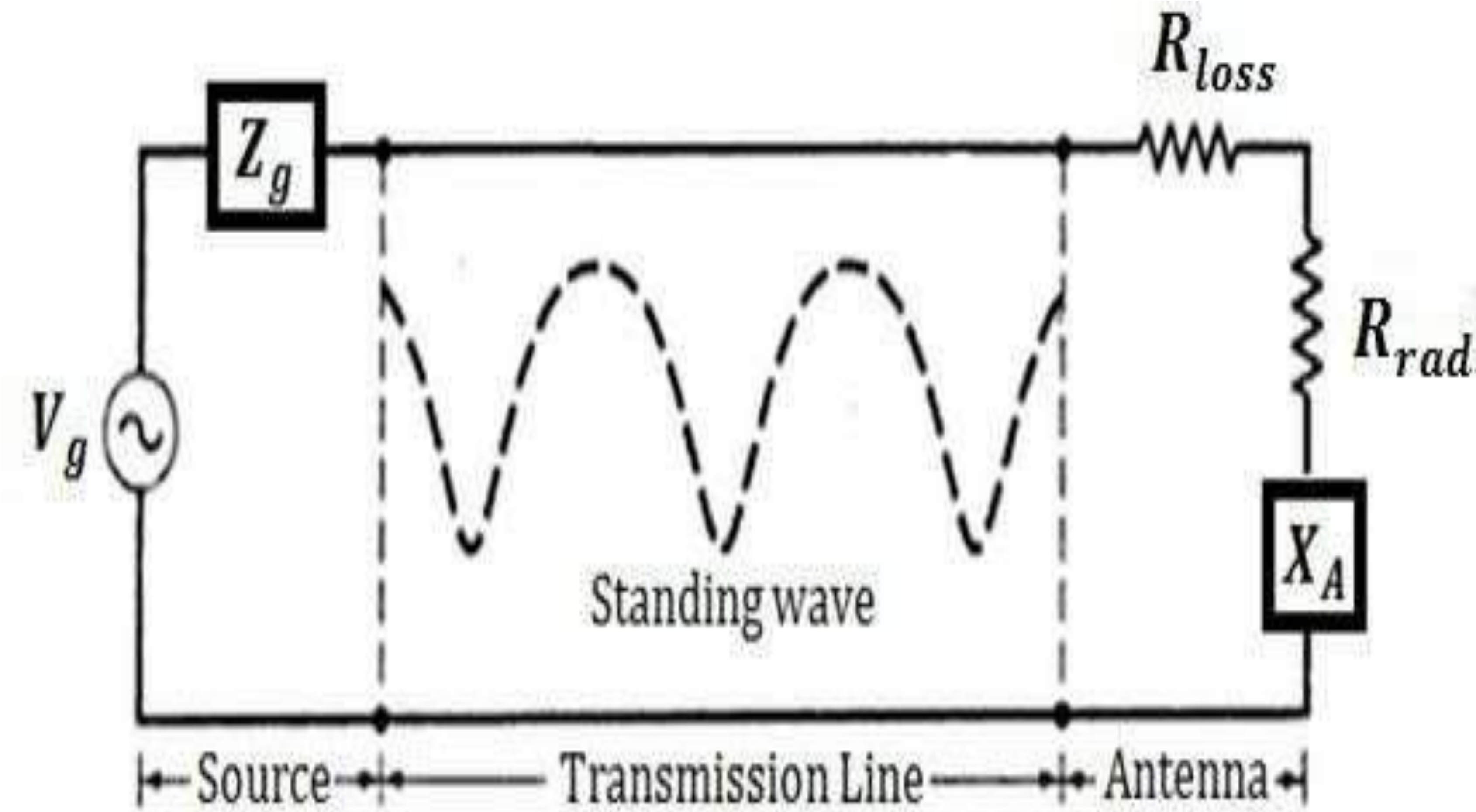
Fig 1.1 Wireless Communication Link



Definition for Antenna

- An antenna is a transitional structure (metallic device: as a rod or wire) between free-space and a guiding device , for radiating or receiving radio waves as shown in Fig. 1-1.
- The guiding device may take the form of a transmission line (coaxial line) or a waveguide (hollow pipe), and it is used to transport electromagnetic energy (from the transmitting source to the antenna, or from the antenna to the receiver).

Thevenin Equivalent of Antenna



$$[Z_A = (R_{loss} + R_{rad}) + jX_A]$$

- Thevenin equivalent of the antenna system in the transmitting mode shown in Fig. 1-2.
- The source is represented by an ideal generator (with voltage $\boldsymbol{V_g}$ and impedance $\boldsymbol{Z_g}$, [$\boldsymbol{Z_g = R_g + j X_g}$]), the transmission line is represented by a line with characteristic impedance $\boldsymbol{Z_c}$, and the antenna is represented by a load with impedance $\boldsymbol{Z_A}$, [$\boldsymbol{Z_A = (R_{loss} + R_{rad}) + j X_A}$] connected to the transmission line.

Topic 1

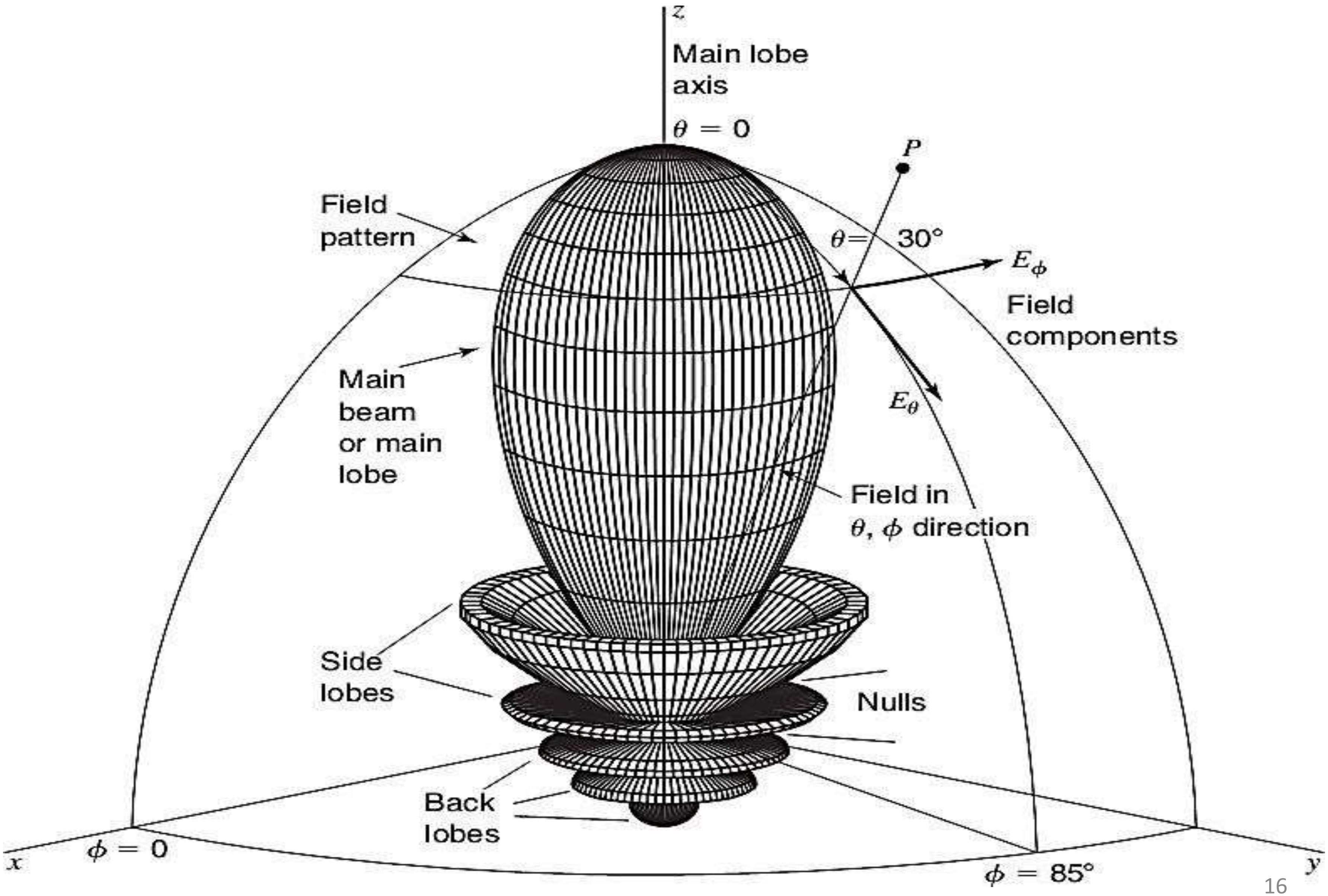
Radiation Pattern

Radiation Pattern

- An antenna radiation pattern - “a mathematical function or a graphical representation of the radiation properties of the antenna as a function of space coordinates”.
- Radiation pattern is very important characteristic of an antenna. Radiation properties include radiation density, radiation intensity, field strength, gain, directivity, effective aperture, polarization, etc.,
- If the radiation from an antenna is represented in terms of field strength (electric or magnetic), then the radiation pattern is called ***field pattern***.
- If the radiation from an antenna is represented in terms of power per unit solid angle, then the radiation pattern is called ***power pattern***.
- To completely specify the radiation pattern with respect to field intensity and polarization requires three patterns.

- The θ component of the electric field as a function of the angles θ and ϕ or $E\theta(\theta, \phi)$.
- The ϕ component of the electric field as a function of the angles θ and ϕ or $E\phi(\theta, \phi)$.
- The phases of these fields as a function of the angles θ and ϕ or $\delta\theta(\theta, \phi)$ and $\delta\varphi(\theta, \phi)$.
- Fig. 1-2. shows three-dimensional field pattern (in spherical coordinates) of a directional antenna with maximum radiation in z- direction at θ = 0°.

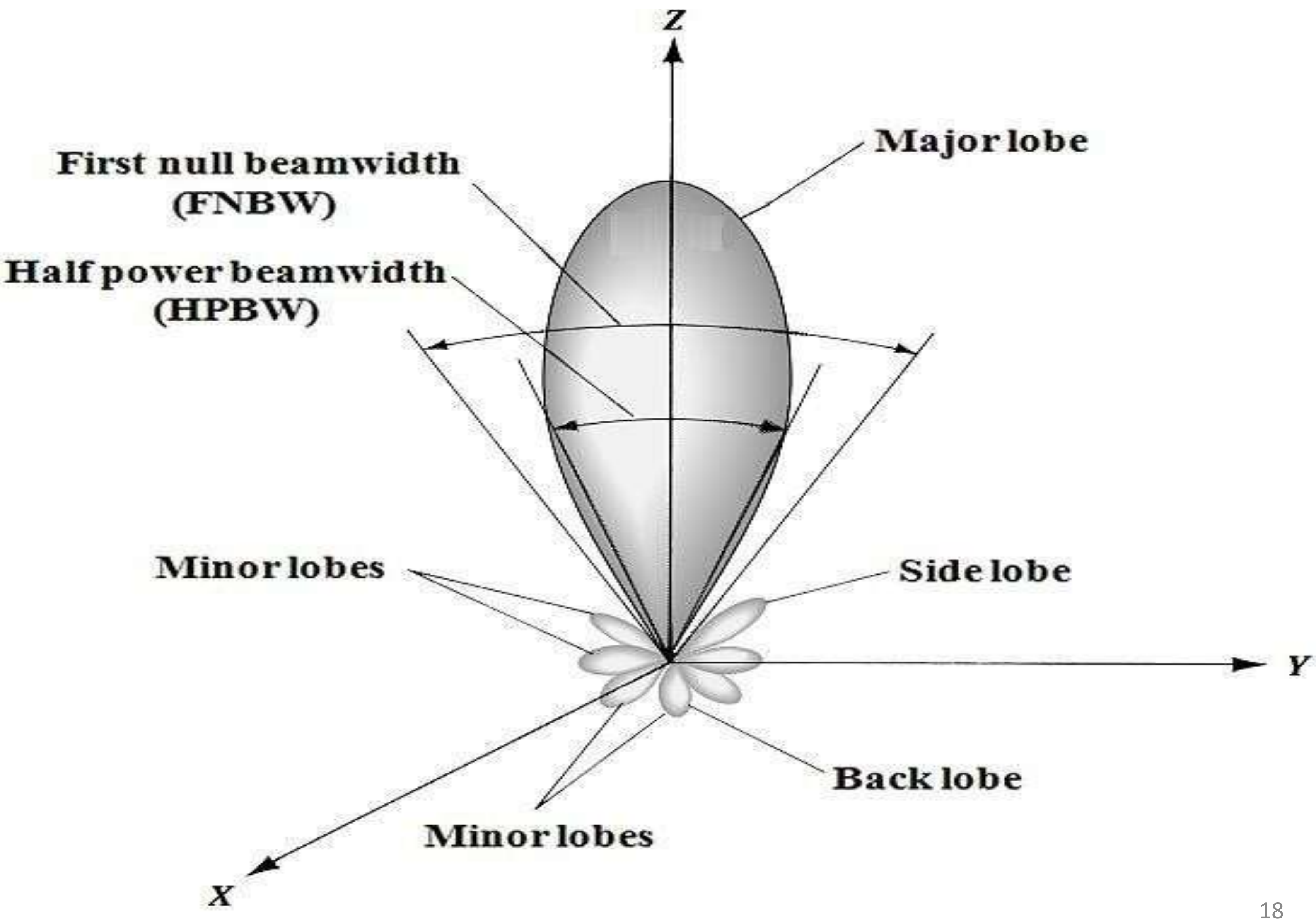
Fig 1.2 Three Dimensional Field Pattern



Radiation Pattern Lobes

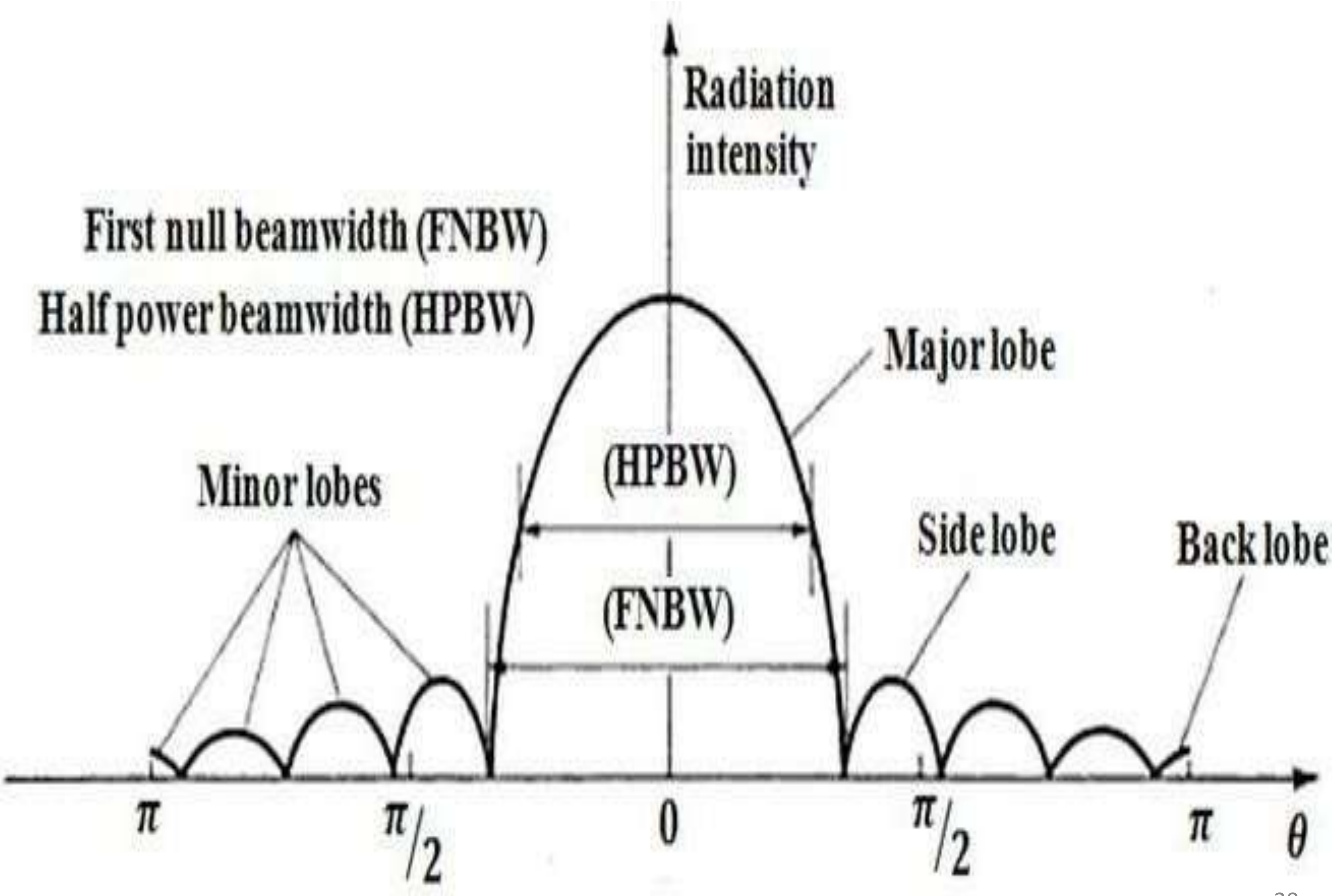
- Various parts of a radiation pattern are referred to as ***lobes***, which may be sub-classified into major or main, minor, side and back lobes.
- A major lobe is defined as the radiation lobe containing the direction of maximum radiation.
- A minor lobe is any lobe except a major lobe.
- A side lobe is a radiation lobe in any direction other than the lobe.
- A back lobe is a radiation lobe whose axis makes an angle of approximately 180° with respect to the beam of an antenna.

Fig 1.3 Radiation Lobes & Beamwidth of Antenna



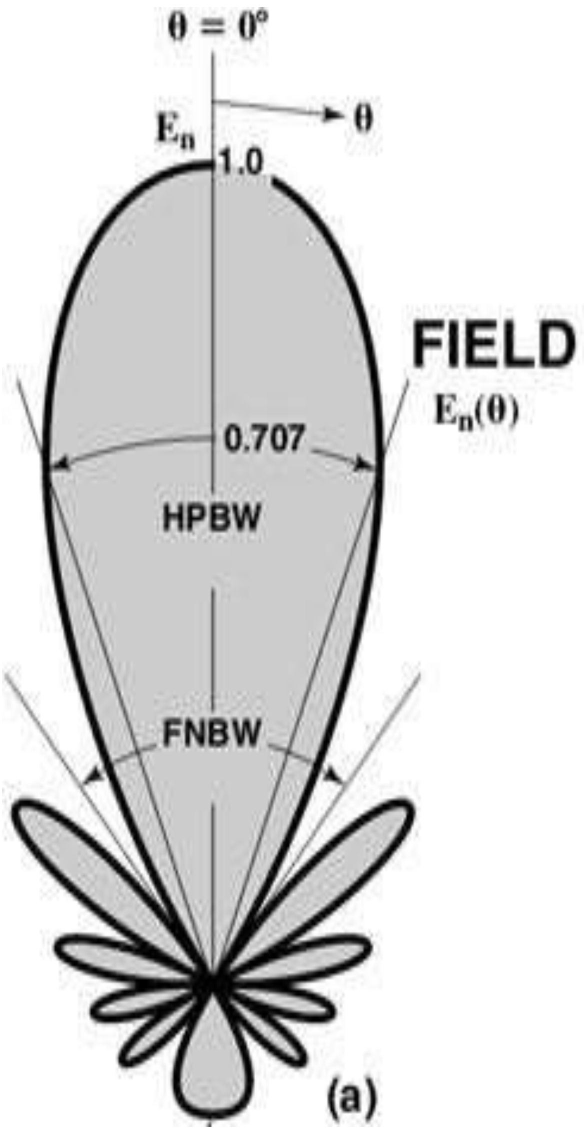
- Fig. 1-3 demonstrates a symmetrical three dimensional power pattern with a number of radiation lobes. Fig. 1-4 illustrates a linear plot of power pattern and its associated lobes and beamwidths.
- The ***half-power beamwidth*** (HPBW) is defined as the angular measurement between the directions in which the antenna is radiating half of the maximum value.
- The ***First-null beamwidth*** or ***beamwidth between first two nulls*** (FNBW) is defined as the angular measurement between the directions radiating no power.

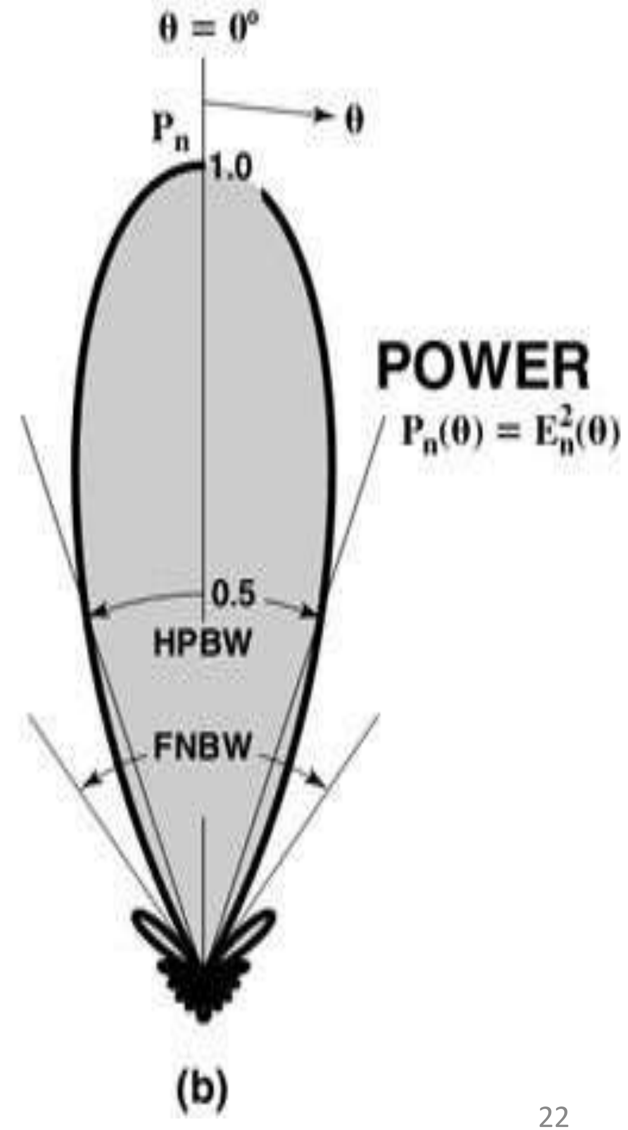
Fig 1.4 Linear Plot of Power Pattern



- Often the field and power patterns are normalized with respect to their maximum value, yielding normalized field and power patterns. Fig.1-5 shows the normalized field pattern and normalized power pattern.
- Dividing a field component by its maximum value, we obtain a *normalized* or *relative field pattern* which is a dimensionless number with maximum value of unity. The half power level occurs at those angles θ and ϕ for which $\boldsymbol{E\theta(\theta,\phi)n} = \mathbf{0.707}$ (or) $\boldsymbol{E\phi(\theta,\phi)n} = \mathbf{0.707}$.

Fig 1.5 Normalized Field and Power Patterns





- Patterns may also be expressed in terms of the power per unit area.
- Normalizing this power with respect to its maximum value yields *normalized power pattern* as a function of angle which is a dimensionless number with a maximum value of unity.
- The half power level occurs at those angles θ and ϕ for which $\boldsymbol{P}\boldsymbol{n}(\boldsymbol{\theta}, \boldsymbol{\phi})\boldsymbol{n} = \mathbf{0.5}$.

Principle Patterns

- For a linearly polarized antenna, performance is often described in terms of its principal $\boldsymbol{E}$-and $\boldsymbol{H}$-plane patterns.
- The $\boldsymbol{E}$-plane is defined as "the plane containing the electric-field vector and the direction of maximum radiation," and the $\boldsymbol{H}$-plane as "the plane containing the magnetic-field vector and the direction of maximum radiation."
- The x-z plane (elevation plane; ϕ = 0) is the principal $\boldsymbol{E}$-plane and the x-y plane (azimuthal plane; = $\pi/2$) is the principal $\boldsymbol{H}$-plane.

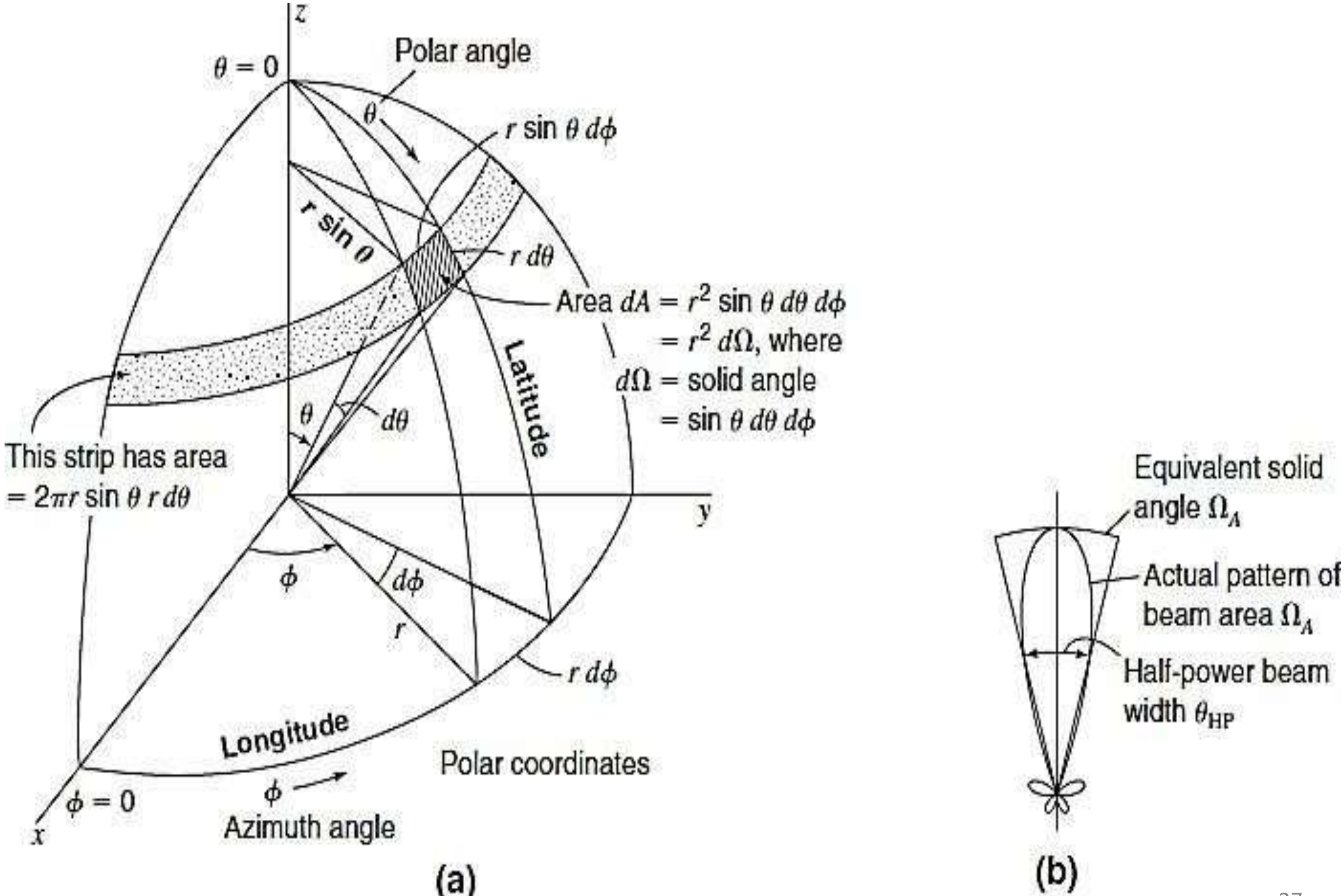
Types of Radiator

- An ***isotropic radiator*** is defined as “a hypothetical lossless antenna having equal radiation in all directions.”
- Although it is ideal and not physically realizable, it is often taken as a reference for expressing the directive properties of actual antennas.
- A ***directional antenna*** is one having the property of radiating or receiving electromagnetic waves more effectively in some directions than in others.
- An ***omnidirectional antenna*** is defined as one “having an essentially non-directional pattern in a given plane and a directional pattern in any orthogonal plane.
- An omnidirectional pattern is then a special type of a directional pattern.

Topic 2

Beam Area or Beam Solid Angle

Fig 1.6 Beam Area or Beam Solid Angle



- In polar-two dimensional coordinates an incremental area dA on the surface of a sphere is the product of the length $rd\theta$ in the θ direction and $r \sin\theta \, d\phi$ in the ϕ direction, as shown in Fig. 1-6.
- The incremental area of a sphere is given by;
- $dA = (rd\theta).(r \sin\theta \, d\phi) = r2 \sin\theta \, d\theta d\phi$ -------- (1.1)
- where $d\Omega$ = solid angle subtended by the area dA, and
  - $d\Omega = dA/r2 \sin\theta \, d\theta d\phi$
- (Generally solid angle is nothing but angle subtended by an elementary area on a sphere.)

- The beam solid angle of an antenna is given by the integral of the normalized power pattern over a sphere (4π).
- $\Omega A = \iint Pn(\theta, \phi)\, d\Omega$
- $\Omega = 4\pi$ -------- (1.2)
- where 4π = solid angle subtended by a sphere (**sr**, Sterdian)
- $2\pi \;\; \pi$

  $\Omega A = \int \int \quad Pn(\theta, \phi) \sin\theta \; d\theta d\phi$
- $\phi = 0 \;\; \theta = 0$ --------- (1.3)

- The beam area of an antenna can often be described approximately in terms of the angles subtended by the half-power points of the main lobe in the two principal planes. Thus,
- $\Omega A = \theta HP \; \phi HP$ -------- (1.4)
- where θHP and ϕHP are the half-power beam widths (HPBW) in the two principal planes, minor lobes being neglected.
  - The (total) beam area consists of the main beam area (ΩM) plus the minor-lobe area(Ωm).
- $\Omega A = \Omega M + \Omega m$ -------- (1.5)

The ratio of the main beam area to the (total) beam area is called the (main) beam efficiency.

- $eM = \Omega M / \Omega A$ -------- (1.6)
- The ratio of the minor-lobe area to the (total) beam area is called the (main) stray factor.
- $em = \Omega m / \Omega A$ -------- (1.7)
- It follows that $eM + em = 1$

Topic 3

Radiation Density

- Electromagnetic waves are used to transport information through a wireless medium or a guiding structure, from one point to the other. The quantity used to describe the power associated with an electromagnetic wave is the instantaneous Poynting vector defined as;
- $\boldsymbol{W} = \boldsymbol{E} \times \boldsymbol{H}$
- where ;
- $\boldsymbol{W}$ = instantaneous Poynting vector , $(\boldsymbol{W}/\boldsymbol{m2})$
- $\boldsymbol{E}$ = instantaneous electric-field intensity, $(\boldsymbol{V}/\boldsymbol{m})$
- $\boldsymbol{H}$ = instantaneous magnetic-field intensity , $(\boldsymbol{A}/\boldsymbol{m})$

- The total power crossing a closed surface can be obtained by integrating the normal component of the Poynting vector over the entire surface.
- $$\boldsymbol{P} = \oiint_S \boldsymbol{W} . \boldsymbol{ds}$$
- For time-varying fields, the time average power density (average Poynting vector);

$$\boldsymbol{W_{rad}} = \frac{1}{2} Re[\boldsymbol{E} \times \boldsymbol{H^*}]$$

- The average power radiated by an antenna (radiated power) can be written as

$$P_{rad} = \oiint_s W_{rad}. ds = \oiint_s \frac{1}{2} Re[E \times H^*] .ds$$

Topic 4

Radiation Intensity

- Radiation intensity in a given direction is defined as " the power radiated from an antenna per unit solid angle".
- The radiation intensity is a far-field parameter and obtained by multiplying the radiation density by the square of the distance.
- It is given by –
- $U = r2\, Wrad$ -------- (1.11)
- Where U = radiation intensity, (**W/unit solid angle**)

  $Wrad$ = radiation density, ($\boldsymbol{W/m2}$)
-

- The total power radiated by the antenna is obtained by integrating the radiation intensity over the entire solid angle of 4π. Thus –

$$P_{rad} = \oiint_{\Omega} U(\theta, \phi)\, d\Omega = \int_0^{2\pi} \int_0^{\pi} U(\theta, \phi) \sin\theta\, d\theta\, d\phi$$

- where $d\Omega$ =element of solid angle = sin $\theta\; d\theta\; d\phi$
- For an isotropic source, U will be independent of the angles θ and ϕ. Thus above exp can be written as –

$$P_{rad} = \oiint_{\Omega} U_0\, d\Omega = U_0 \int_0^{2\pi} \int_0^{\pi} d\Omega = 4\pi U_0$$

- The radiation intensity of an isotropic source as;
- $$U_0 = P_{rad}/4\pi$$

Topic 5

Directivity

- Directive gain of an antenna is defined as "the ratio of the radiation intensity in a given direction from the antenna to the radiation intensity averaged over all directions".

$$D = \frac{U(\theta,\phi)}{U(\theta,\phi)_{av}} = \frac{U}{U_0}$$

- The average radiation intensity is equal to the total power radiated divided by 4π. In mathematical form –

$$D = \frac{U}{(P_{rad}/4\pi)} = \frac{4\pi U}{P_{rad}}$$

- If the direction not specified, the direction of maximum radiation intensity (maximum directivity, D_0,) given by-

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

- ***Directivity*** D_0 : It is defined as the ratio of the maximum radiation intensity to the average radiation intensity.
- The total power radiated by the antenna is defined in terms of normalized power pattern is given by –

$$P_{rad} = U_{max} \oiint_{\Omega} P_n(\theta, \phi)_n \, d\Omega$$

- Substuiting P_{rad} in the above –

$$\boxed{D_0 = \frac{4\pi}{\Omega_A} \quad (Directivity\ from\ Beam\ Area)}$$

Topic 6

Power Gain

- Gain of an antenna is defined as "the ratio of the radiation intensity, in a given direction, to the radiation intensity obtained if the power accepted by the antenna were radiated isotropically.
- The radiation intensity corresponding to isotropically radiated power is equal to the power accepted (input) by the antenna divided by 4π."

$$G = 4\pi \frac{Radiation\ Intensity}{Total\ input\ (accepted)\ power} = 4\pi \frac{U(\theta, \phi)}{P_{in}}$$

- In most cases, the relative gain is defined as " the ratio of the power gain in a given direction to the power gain of a reference antenna in its referenced direction."

- The power input must be the same for both antennas.
- The reference antenna is usually a dipole, horn or any other antenna whose gain can be calculated or it is known.
- However, the reference antenna is a *lossless isotropic source*. Thus,

$$G = 4\pi \frac{U(\theta, \phi)}{P_{in}(lossless\ isotropic\ source)}$$

- The total power radiated, $Prad$, by the antenna is related to the input power, Pin, by:
- $P_{rad} = \kappa P_{in}$
- Substuiting P_{in} in the exp for 'G' –

$$G = 4\pi \frac{U(\theta, \phi)}{\left(\frac{P_{rad}}{\kappa}\right)}$$

$$G(\theta, \phi) = \kappa D(\theta, \phi)$$

- Where 'k' is the antenna efficiency factor.
- The gain (G) of an antenna is an actual quantity which is less than the directivity (D) due to ohmic losses in the antenna.
- The maximum value of the gain is related to maximum directivity;

$$G(\theta, \phi)_{max} = \kappa D(\theta, \phi)_{max}$$

$$\boxed{G_0 = \kappa D_0}$$

Topic 7

Antenna Efficiency Factor

- Antenna Efficiency is defined as the ratio of power radiated by the antenna to the total input power supplied by the antenna. It is denoted by κ. It value lies between $0 \le \kappa \le 1$.

$$\frac{\text{Power radiated}}{\text{Total input power}}$$

$$\frac{P_{rad}}{P_{in}} \qquad \frac{P_{rad}}{P_{rad} + P_{loss}}$$

$$\kappa = \frac{\frac{1}{2}|I|^2 R_{rad}}{\frac{1}{2}|I|^2 R_{rad} + \frac{1}{2}|I|^2 R_{loss}}$$

$$\frac{R_{rad}}{R_{rad} + R_{loss}}$$

Topic 8

Input Impedance

- Input impedance is defined as “the impedance presented by an antenna at its terminals or the ratio of the voltage to current at a pair of terminals or the ratio of the appropriate components of the electric to magnetic fields at a point.
- Consider an antenna in the transmit mode, having an input impedance of $\boldsymbol{ZA}$ = $\boldsymbol{RA}$ + $\boldsymbol{j}$ $\boldsymbol{XA}$, connected directly to a source having an equivalent Thevenin’s voltage, $\boldsymbol{Vg}$, and an internal impedance $\boldsymbol{Zg}$ = $\boldsymbol{Rg}$ + $\boldsymbol{j}$ $\boldsymbol{Xg}$, as shown in Fig.1-7.

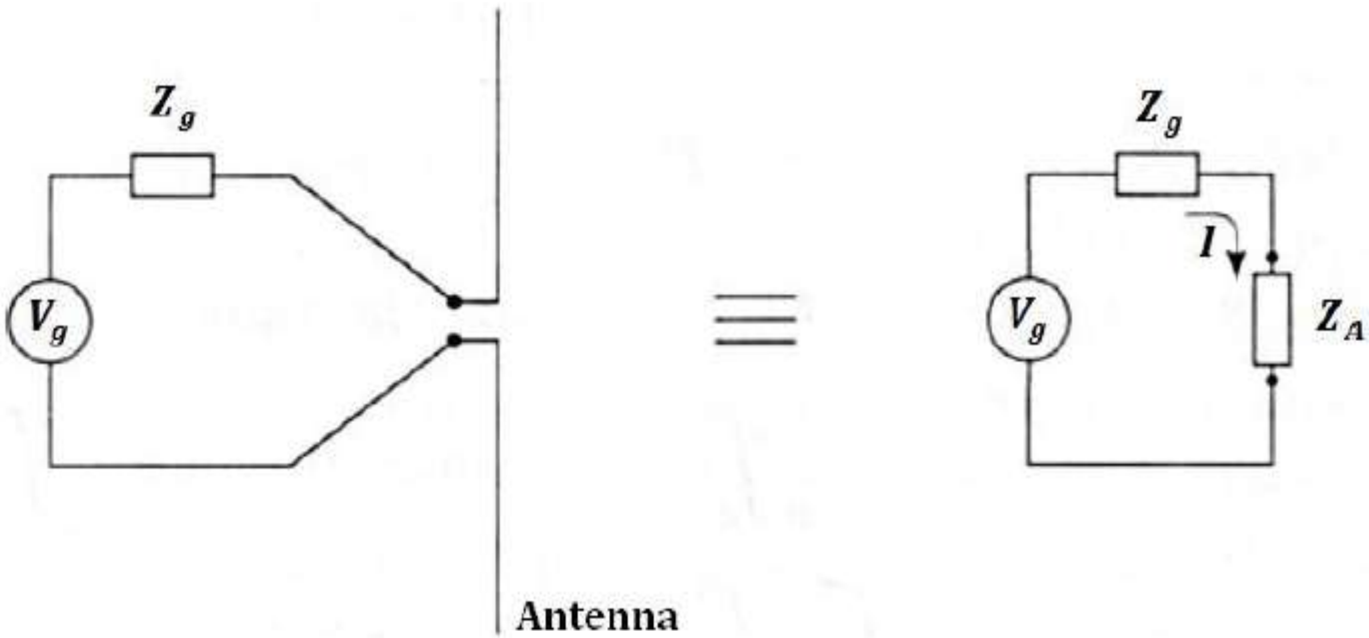


Fig. 1-7 Transmitting Antenna and its equivalent circuits.

- where $\boldsymbol{ZA}$ is the antenna impedance , $\boldsymbol{RA}$ is the antenna resistance , XA is the antenna reactance , $\boldsymbol{Rrad}$ is the radiation resistance of the antenna and $\boldsymbol{Rloss}$ is the ohmic loss resistance of the antenna.
- The maximum power transfer takes place when the antenna is conjugate-matched to the source.
- Under the complex conjugate-match condition, the antenna input current is –

$$I = \frac{V_g}{R_g + R_A} = \frac{V_g}{2R_A}$$

- The real power supplied by the source is given by;

$$P_g = \frac{1}{2} Re\{V_g I^*\} = \frac{|V_g|^2}{4 R_A}$$

- Half the power supplied by the source is lost in the source resistance, $\boldsymbol{Rg}$, and the other half gets dissipated in the antenna resistance, $\boldsymbol{RA}$.
- Thus the effective power input to the antenna is ;

$$P_{in} = \frac{1}{2}|I|^2 R_A = \frac{|V_g|^2}{8R_A}$$

- The antenna resistance, $\boldsymbol{RA}$, is comprised of two components, namely the radiation resistance, $\boldsymbol{Rrad}$, and the loss resistance, $\boldsymbol{Rloss}$;
- $RA = Rrad + Rloss$ -------- (1.32)

The total power radiated by an antenna is given by;

$$P_{rad} = \frac{1}{2}|I|^2 R_{rad}$$

The power dissipated in the antenna by ohmic losses is given by;

$$P_{loss} = \frac{1}{2}|I|^2 R_{loss}$$

For a matched antenna these are given by

$$P_{rad} = \frac{|V_g|^2}{8R_A} R_{rad}$$

$$P_{loss} = \frac{|V_g|^2}{8R_A} R_{loss}$$

Topic 9

Radiation Resistance

- Radiation resistance is the fictitious resistance such that when connected in series with an antenna will consume the same power as actually radiated by the antenna.
- The total power radiated by an antenna is given by;

$$P_{rad} = \frac{1}{2} |I|^2 R_{rad}$$

$$\therefore R_{rad} = \frac{P_{rad}}{\frac{1}{2} |I|^2}$$

- The radiation resistance is the part of an antenna’s feed point resistance caused by the radiation of electromagnetic waves from the antenna.
- The radiation resistance is determined by the geometry of the antenna.
- The energy lost by radiation resistance is converted to electromagnetic radiation.

Topic 10

Bandwidth

- The Bandwidth of an antenna is defined as “the range of frequencies within which the performance of the antenna, with respect to some characteristics, conforms to a specified standard”.
- The bandwidth - the range of frequencies, on either side of a center frequency (usually the resonance frequency for a dipole), where the antenna characteristics (input impedance, pattern, beam width, polarization, side lobe level, gain, etc) are within an acceptable value of those at the center frequency.
- For narrow band antennas, the bandwidth is percentage of the frequency difference over the center frequency of the bandwidth.

- The antenna bandwidth mainly depends on impedance and pattern of antenna.
- At low frequency, impedance variation decides the bandwidth as pattern characteristics are frequency insensitive.
- Under such condition, bandwidth of the antenna is inversely proportional to Q factor of the antenna.
- The bandwidth of the antenna can be expressed as ;

$$\text{Bandwidth }(B.W) = \Delta\omega = \omega_2 - \omega_1 = \frac{\omega_0}{Q}$$

$$\Delta f = f_2 - f_1 = \frac{f_0}{Q}$$

- where $f0$is the center frequency or resonant frequency, while Q factor of antenna is given by –

$$Q = 2\pi = \frac{Total\ energy\ stored\ by\ antenna}{Energy\ radiated\ per\ cycle}$$

- Thus for lower Q antennas, the antenna bandwidth is very high and vice versa

Topic 10

Polarization

- Polarization defined as the figure traced as a function of time by the tip of the instantaneous electric field vector at fixed location in free space observed along the direction of propagation.
- The polarization of an antenna is the polarization of the wave radiated by the antenna in the far-field.
- In the far-field region, the radiated field essentially has a spherical wave front with $\boldsymbol{E}$ and $\boldsymbol{H}$ fields transverse to the radial direction, which is the direction of propagation.

- As the radius of curvature tends to infinity, the wave front can be considered as a plane wave and the polarization of this plane wave is the polarization of the antenna.
- The polarization of the antenna is direction-dependent, i.e., polarization as a function of (θ, ϕ).
- Polarization of a plane wave describes the shape, orientation, and sense of rotation of the tip of the electric field vector as a function of time, in the direction of propagation.

Linear Polarization

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

1. Only one component,
2. Two orthogonal linear components are in time phase.

Circular Polarization

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

1. The field must have two orthogonal linear components
2. The two components must have the same magnitude.
3. The two components must have a time-phase difference of odd multiples of 90°.

Elliptical Polarization

- A time-harmonic wave is elliptically polarized at a given point in space if the electric field (or magnetic field) vector at that point traces an elliptical locus in space.
- Condition:
- 1. The field must have two orthogonal linear components,
- 2. The two components can be of the same or different magnitude, and

Elliptical Polarization 2

- 3. i). If the two components are not of the same magnitude, the time-phase between two components must not be 0° or multiples of 180°(because it will be linear).
- ii). If the two components are of the same magnitude, the time-phase between two components must not be odd multiples of 90°(because it will be circular).

Topic 11

Polarization Mismatch

- In general, the polarization of the receiving antenna will not be the same as the polarization of the incoming (incident) wave.
- This is commonly stated as “polarization mismatch.”
- The amount of power extracted by the antenna from the incoming signal will not be maximum because of the polarization loss.
- Assuming that the electric field of the incoming wave can be written as -
-
$$\mathbf{E}_i = \hat{\boldsymbol{\rho}}_w E_i$$

where $\hat{\rho}_w$ is the unit vector of the wave, and the polarization of the electric field of the receiving antenna can be expressed as

$$\mathbf{E}_a = \hat{\rho}_a E_a$$

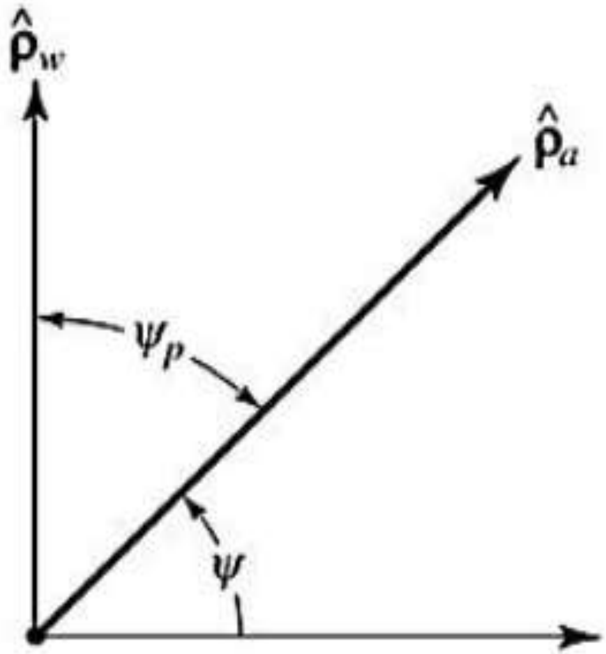
where $\hat{\rho}_a$ is its unit vector (polarization vector), the polarization loss can be taken into account by introducing a *polarization loss factor* (PLF). It is defined, based on the polarization of the antenna in its transmitting mode, as

$$\text{PLF} = |\hat{\rho}_w \cdot \hat{\rho}_a|^2 = |\cos\psi_p|^2 \text{ (dimensionless)}$$

where ψ_p is the angle between the two unit vectors.

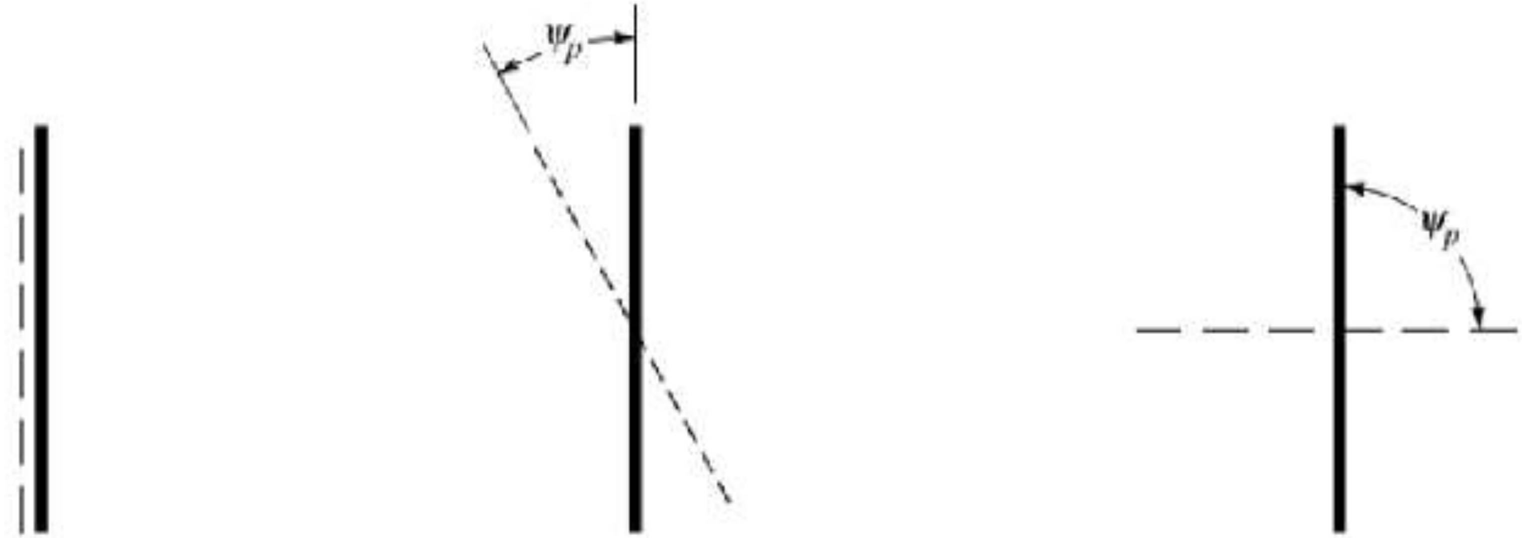
- The relative alignment of the polarization of the incoming wave and of the antenna is shown in the next Figure.
- If the antenna is polarization matched, its PLF will be unity and the antenna will extract maximum power from the incoming wave.

Polarization Unit Vectors



Polarization unit vectors of incident wave ($\hat{\rho}_w$) and antenna ($\hat{\rho}_a$), and polarization loss factor (PLF).

Polarization Loss Factor



$\text{PLF} = |\hat{\boldsymbol{\rho}}_w \cdot \hat{\boldsymbol{\rho}}_a|^2 = 1$
(aligned)

$\text{PLF} = |\hat{\boldsymbol{\rho}}_w \cdot \hat{\boldsymbol{\rho}}_a|^2 = \cos^2 \psi_p$
(rotated)

$\text{PLF} = |\hat{\boldsymbol{\rho}}_w \cdot \hat{\boldsymbol{\rho}}_a|^2 = 0$
(orthogonal)

PLF for transmitting and receiving linear wire antennas

- If the polarization of the incoming wave is orthogonal to the polarization of the antenna, then there will be no power extracted by the antenna from the incoming wave and the PLF will be zero or $-\infty$ dB.
- The previous Figure illustrates the polarization loss factors (PLF) for linear wire antenna.

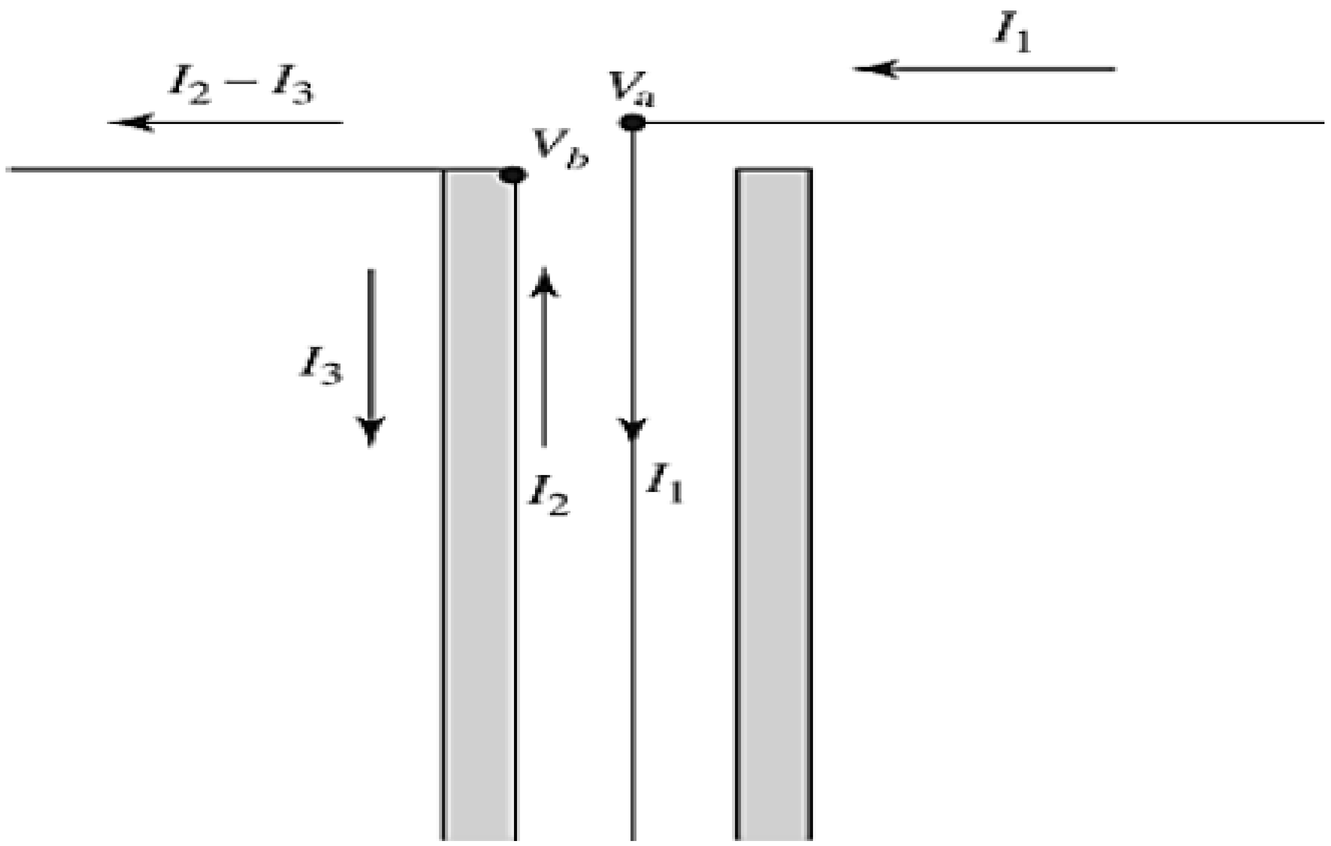
Topic 12

Matching - Baluns

- Transmission lines are referred to as balanced or unbalanced.
- Parallel wire lines are balanced in that if an incident wave is launched down the line, it will excite balanced currents on a symmetrical antenna.
- On the other hand, a coaxial transmission line is not balanced.
- A wave traveling down the coax may have a unbalanced current mode - the currents on the inner conductor and the inside of the outer conductor are equal in magnitude and opposite in direction.

- When this wave reaches a symmetrical antenna, a current may flow back on the outside of the outer conductor, which unbalances the antenna and transmission line as shown next figure.
- The currents on the two halves of the dipole are unbalanced. The current $I3$ flowing on the outside of the coax will radiate.
- The currents $I1$, and $I2$ in the coax are shielded from the external world by the thickness of the outer conductor.
- They could actually be unbalanced with no resulting radiation; it is the current on the outside surface of the outer conductor that must be suppressed.

Coaxial TX Line Feeding a Dipole

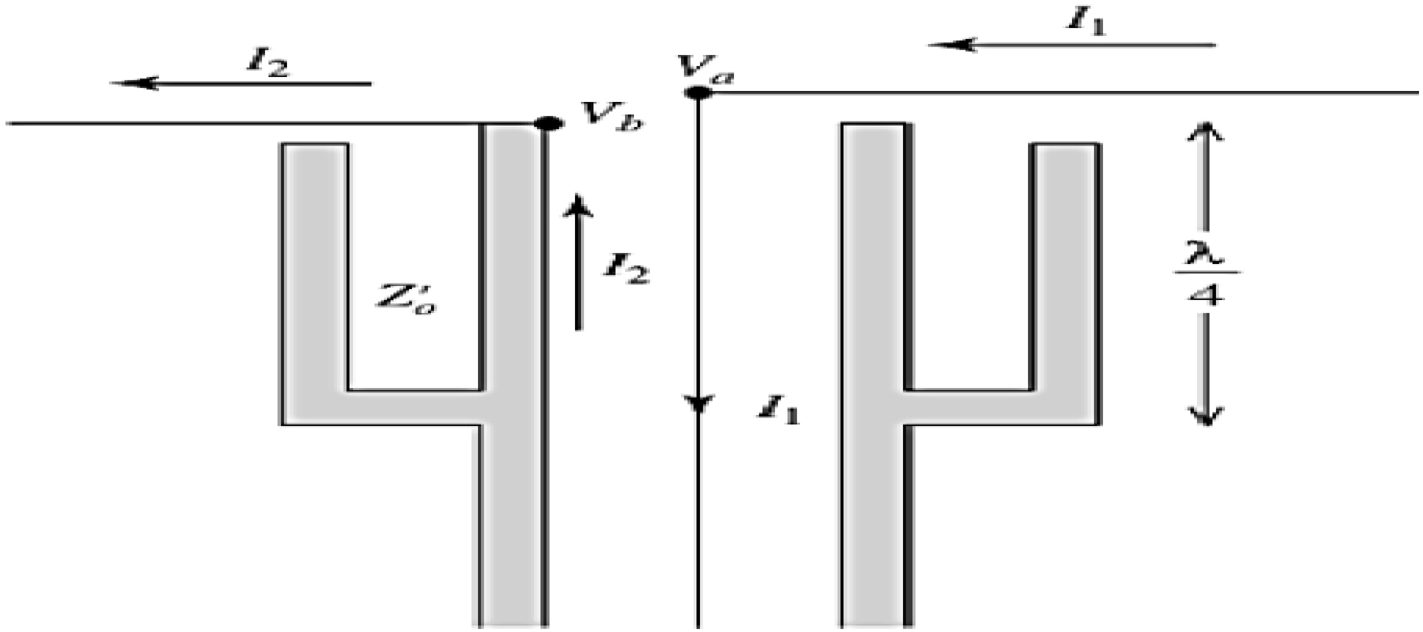


Cross section of a coaxial transmission line feeding a dipole antenna at its center.

- To suppress this outside surface current, a balun (contraction for "balanced to unbalanced transformer") is used.
- The situation in previous Figure may be understood by examining the voltages that exist at the terminals of the antenna.
- These voltages are equal in magnitude but opposite in phase (i.e., $Va = -Vb$). Both voltages act to cause a current to flow on the outside of the coaxial line.
- If the magnitude of the currents on the outside of the coax produced by both voltages are equal, the net current would be zero.

Sleeve Balun

- However, since one antenna terminal is directly connected to the outer conductor, its voltage Vb, produces a much stronger current than the other voltage Va.

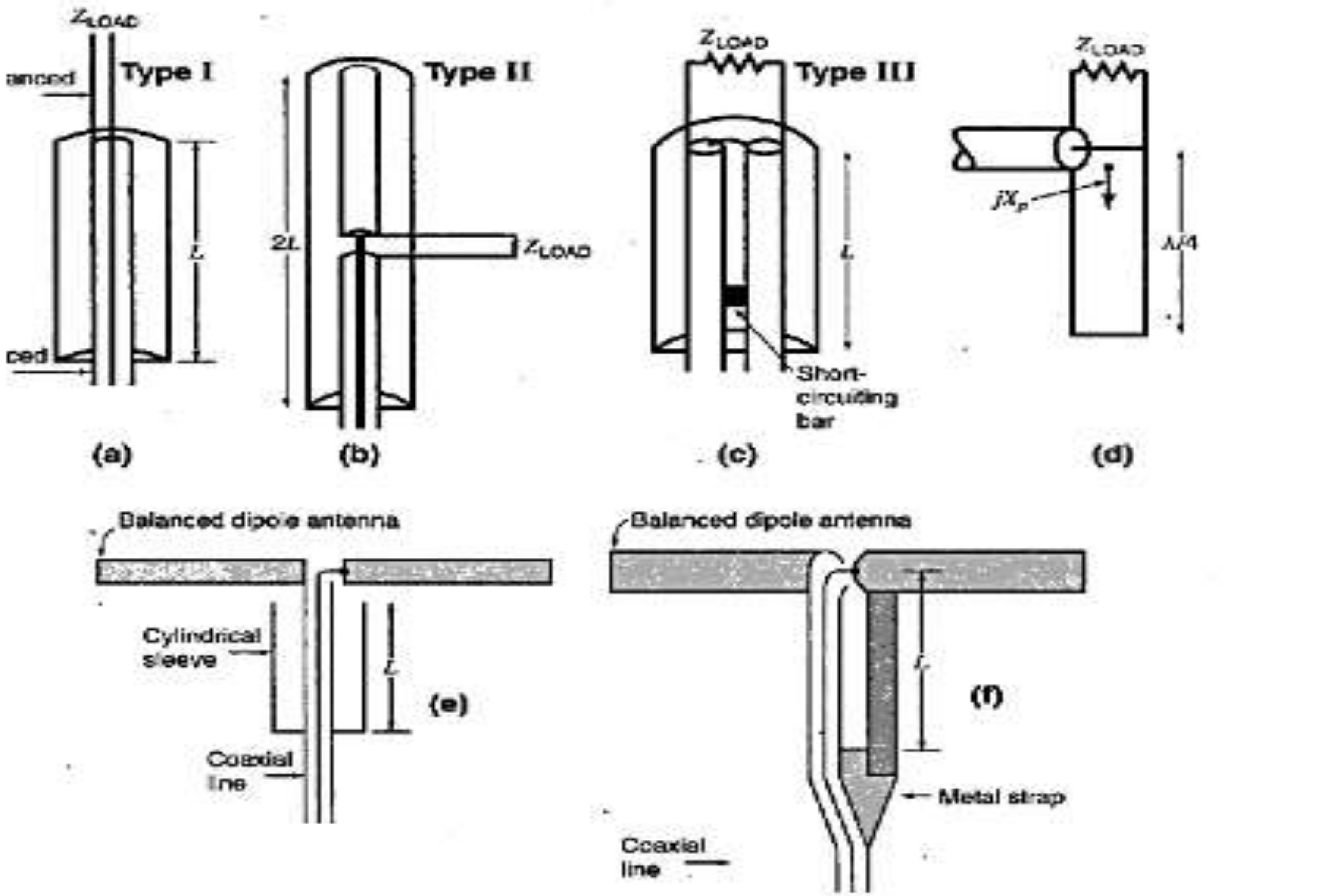


Cross section of a sleeve balun feeding a dipole antenna at its center.

Bazooka Balun

- One type of a balun is that shown in Fig. 2, referred to usually as a *bazooka* balun.
- Mechanically it requires that a $\lambda/4$ in length metal sleeve, and shorted at its one end, encapsulates the coaxial line.
- Electrically the input impedance at the open end of this $\lambda/4$ shorted transmission line, which is equivalent to
- Z' , will be very large (ideally infinity).
- Thus the current $I3$ will be choked, if not completely eliminated and system will be nearly balanced.

Three Common Types of Baluns



(*a*) Type I balun or "bazooka," (*b*) Type II balun, (*c*) Type III balun, (*d*) Type III balun equivalent circuit, (*e*) Type I balun with dipole antenna and (*f*) dipole antenna with Type III balun minus sleeve.

Topic 13

Antenna Noise Temperature

Every object with a physical temperature above absolute zero ($0K = -273°C$) radiates energy.

The amount of energy radiated is usually represented by an equivalent temperature TB, better known as brightness temperature, and it is defined as

$$T_B(\theta, \phi) = \xi(\theta, \phi) T_m$$

Where ; T_B = brightness temperature (equivalent temperature; K)

ξ = emissivity (dimensionless)

T_m = molecular (physical) temperature (K)

- Since the values of emissivity are $0 \le x \le 1$, the maximum value the brightness temperature can achieve is equal to the molecular temperature.
- Some of the better natural emitters of energy at microwave frequencies are (a) the ground with equivalent temperature of about 300 K and (b) the sky with equivalent temperature of about 5 K when looking toward zenith and about 100–150 K toward the horizon.
- The brightness temperature emitted by the different sources is intercepted by antennas and appears at their terminals as an antenna temperature.

- It is defined as –

$$T_A = \frac{\int_0^{2\pi}\int_0^{\pi} T_B(\theta,\phi)\, G(\theta,\phi) \sin\theta \, d\theta \, d\varphi}{\int_0^{2\pi}\int_0^{\pi} G(\theta,\phi) \sin\theta \, d\theta \, d\varphi}$$

Where ; T_A = antenna temperature (effective noise temperature of the antenna radiation resistance ; K) , $G(\theta,\phi)$ = gain (power) pattern of the antenna.

- If the antenna and transmission line are maintained at certain physical temperatures & transmission line between the antenna and receiver is lossy, the antenna temperature TA as seen by the receiver must be modified to include the other contributions and the line losses.

- If the antenna itself is maintained at a certain physical temperature Tp and a transmission line of length l, constant physical temperature $T0$ throughout its length, and uniform attenuation of α (Np/m) is used to connect an antenna to a receiver, as shown in Fig. 1, the effective antenna temperature at the receiver terminals is given by ;

$$T_a = T_A e^{-2\alpha l} + T_{AP} e^{-2\alpha l} + T_0(1 - e^{-2\alpha l})$$

$$T_{AP} = (\frac{1}{e_A} - 1)\,T_p$$

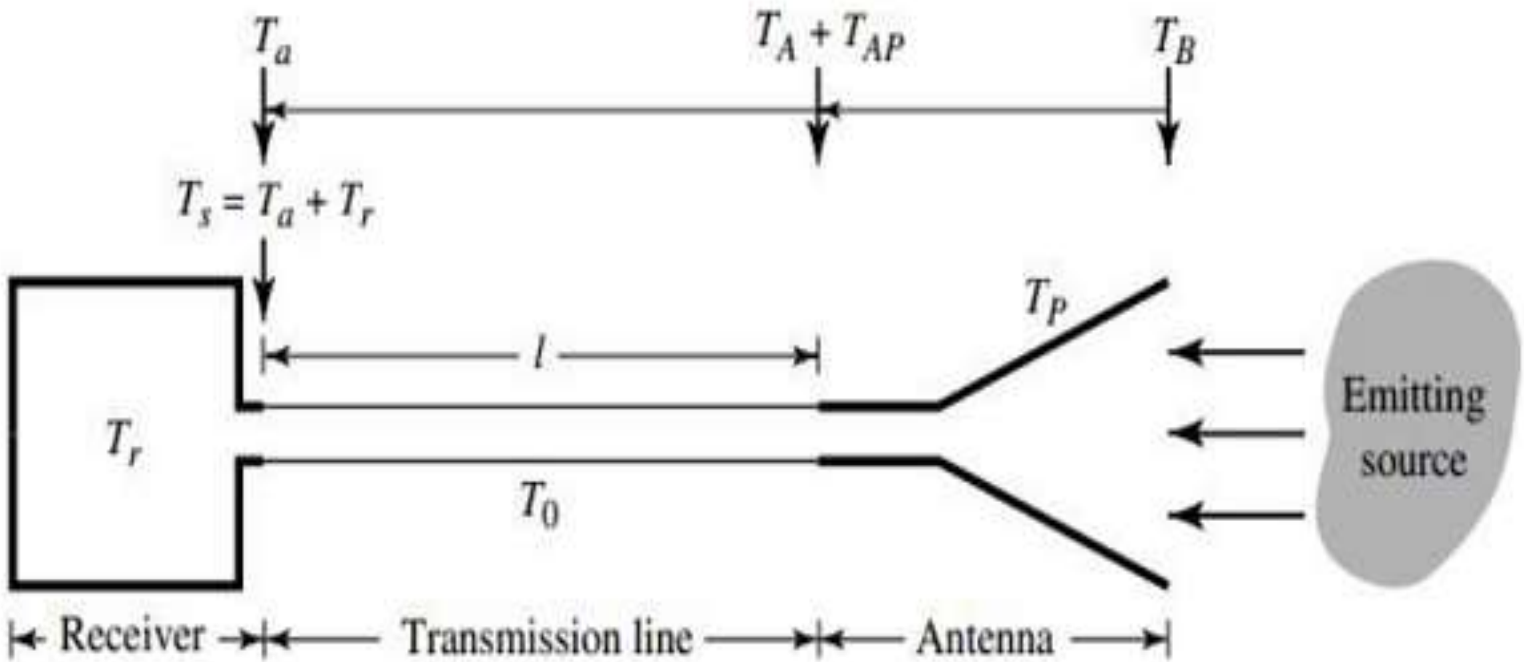


Fig. 1 Antenna, transmission line, and receiver arrangement for system noise power calculation.

where ; T_a = antenna temperature at the receiver terminals

T_A = antenna noise temperature at the antenna terminals

T_{AP} = antenna temperature at the antenna terminals due to physical temperature

α = attenuation coefficient of transmission line

T_0 = physical temperature of the transmission line

T_p = antenna physical temperature

e_A = themal efficiency of antenna l= length of the transmission line

The antenna noise power of Eqn. (1.52) must also be modified and written as;

$$P_r = kT_a\Delta f$$ -------- (

If the receiver itself has a certain noise temperature T_r (due to thermal noise in the rec components), the system noise power at the receiver terminals is given by ;

$$P_s = k(T_a + T_r)\Delta f = kT_s\Delta f$$ -------- (

where ; P_s = system noise power (at receiver terminals)

T_a = antenna noise temperature (at receiver terminals)

T_r = receiver noise temperature (at receiver terminals)

T_s = $T_a + T_r$ =effective system noise temperature (at receiver terminals)

Topic 14

Effective Aperture

- The effective aperture (also known as the effective area) of an antenna is the area over which the antenna collects energy from the incident wave and delivers it to the receiver load
- If the power density in the wave incident from the (θ, ϕ) direction is $\boldsymbol{W}$ at the antenna and $Pr(\theta,\ \phi)$ is the power delivered to the load connected to the antenna, then the effective aperture, Ae,is defined as ;

$$A_e(\theta, \phi) = \frac{P_r(\theta, \phi)}{W} \quad m^2$$

Receiving Antenna and Load

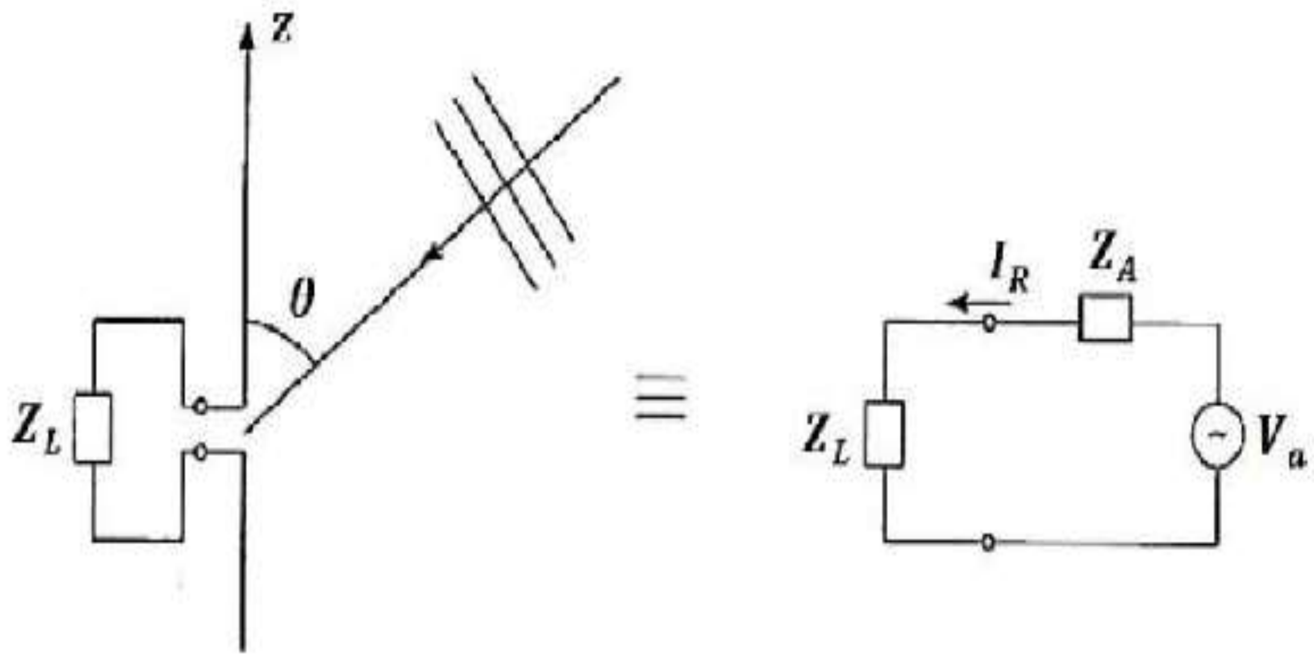


Fig. 1 shows the equivalent circuit of the receiving antenna and the load.

Fig. 1-: shows the equivalent circuit of the receiving antenna and the load, the power delivered to the load, Z_L, connected to the antenna terminals is ;

$$P_r = \frac{1}{2} |I_R|^2 R_L$$

where ; P_r = Power delivered to the load Z_L

R_L = Real part of the load impedance, ($Z_L = R_L + j X_L$)

Let $Z_A = R_A + j X_A$ be the antenna impedance and V_a be Thevenin's equivalent source corresponding to the incident plane wave. The real part of the antenna impedance can be further divided into two parts, i.e., $R_A = R_{rad} + R_{loss}$.

If the antenna is conjugate-matched to the load so that maximum power can be transferred to the load, we have $Z_L = Z_A^*$ or $R_L = R_A$ and $X_L = -X_A$.

It is seen from the equivalent circuit that the power collected from the plane wave is dissipated in the three resistances, the receiver load, the radiation resistance, and the loss resistance.

For a conjugate-match, the current through all three resistances is ;

$$I = \frac{V_a}{R_L + R_{rad} + R_{loss}} = \frac{V_a}{R_L + R_A} = \frac{V_a}{2R_L}$$

and the three powers are computed using the formulae ;

$$P_r = \frac{1}{2}|I|^2 R_L = \frac{1}{2}\frac{|V_a|^2}{(2R_L)^2} R_L = \frac{|V_a|^2}{8R_L}$$

$$P_{scat} = \frac{1}{2}|I|^2 R_{rad} = \frac{|V_a|^2}{8R_L{}^2} R_{rad}$$

$$P_{loss} = \frac{1}{2}|I|^2 R_{loss} = \frac{|V_a|^2}{8R_L{}^2} R_{loss}$$

- where Pr is the power delivered to the receiver load, $Ploss$ is the power dissipated in the antenna, and $Pscat$ is the power scattered, since there is no physical resistance corresponding to the radiation resistance.
- The total power collected by the antenna is the sum of the three powers.

$$P_c = P_r + P_{scat} + P_{loss}$$

- If the power density in the incident wave is $\boldsymbol{W}$, then the effective collecting aperture, Ac,of the antenna is the equivalent area from which the power is collected.

$$A_c(\theta, \phi) = \frac{P_c(\theta, \phi)}{W} \quad m^2$$

- Consider now an antenna with an effective aperture Ae, which radiates all of its power in a conical pattern of beam area ΩA, as suggested in Fig. 1.

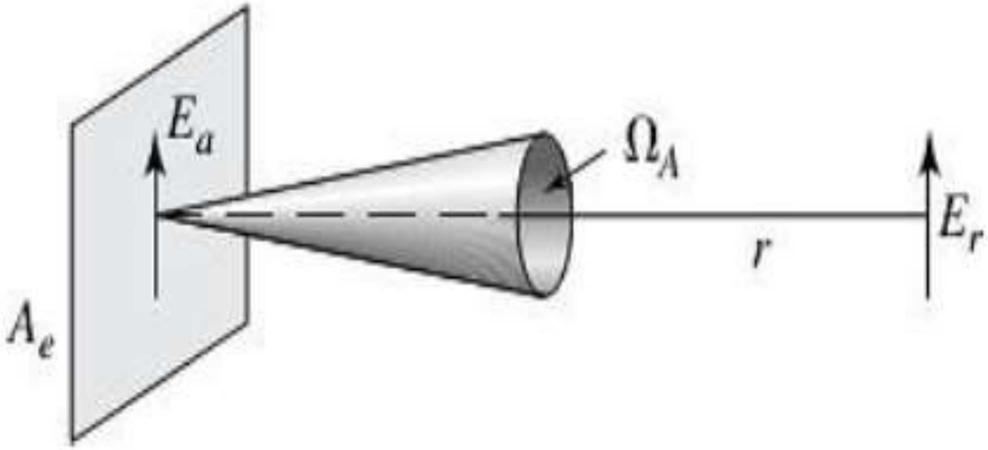


Fig. 1 Radiation over beam area Ω_A from aperture A_e

Assuming a uniform field E_a, over the aperture, the power radiated is ;

$$P = \frac{E_a^2}{\eta} A_e \qquad \text{-------- (1)}$$

Assuming a uniform field E_r in the farfield at a distance r, the power radiated is also given by

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

where ; $E_r = (E_a A_e r)/\lambda$

Eqn. (1.) and (2) yield aperture-beam area relationship ;

$$\lambda^2 = A_e \Omega_A \qquad \text{-------- (3)}$$

We know that directive gain (D) and beam-area (Ω_A) relationship ;

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

Comparing the above equations ;

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

Maximum effective aperture

$$\boxed{A_{em} = \frac{\lambda^2}{4\pi} D_0}$$

Topic 15

Radiation from Oscillating Dipole

To calculate the electromagnetic field radiated in the space by a short dipole, the retarded potential is used. A short dipole is an **alternating current element.** It is also called an **oscillating current element.**

In general, a current element IdL is nothing but an element of length dL carrying filamentary current I. This length of a thin wire is assumed to be very short, so that the filamentary current can be considered as constant along the length of an element.

Let us study how to calculate the electromagnetic field due to an alternating current element. Consider **spherical co-ordinate system.** Consider that an alternating current element IdL cos ωt is located at the centre as shown in the Fig. 1.3. The aim is to calculate electromagnetic field at a point P placed at a distance R from the origin. The current element I dL cosω t is placed along the z-axis.

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

$$\bar{A}(r) = \frac{\mu}{4\pi}\int_{v} \frac{\bar{J}\left(t-\frac{r}{v}\right)}{R}\, dv' \qquad \ldots (1)$$

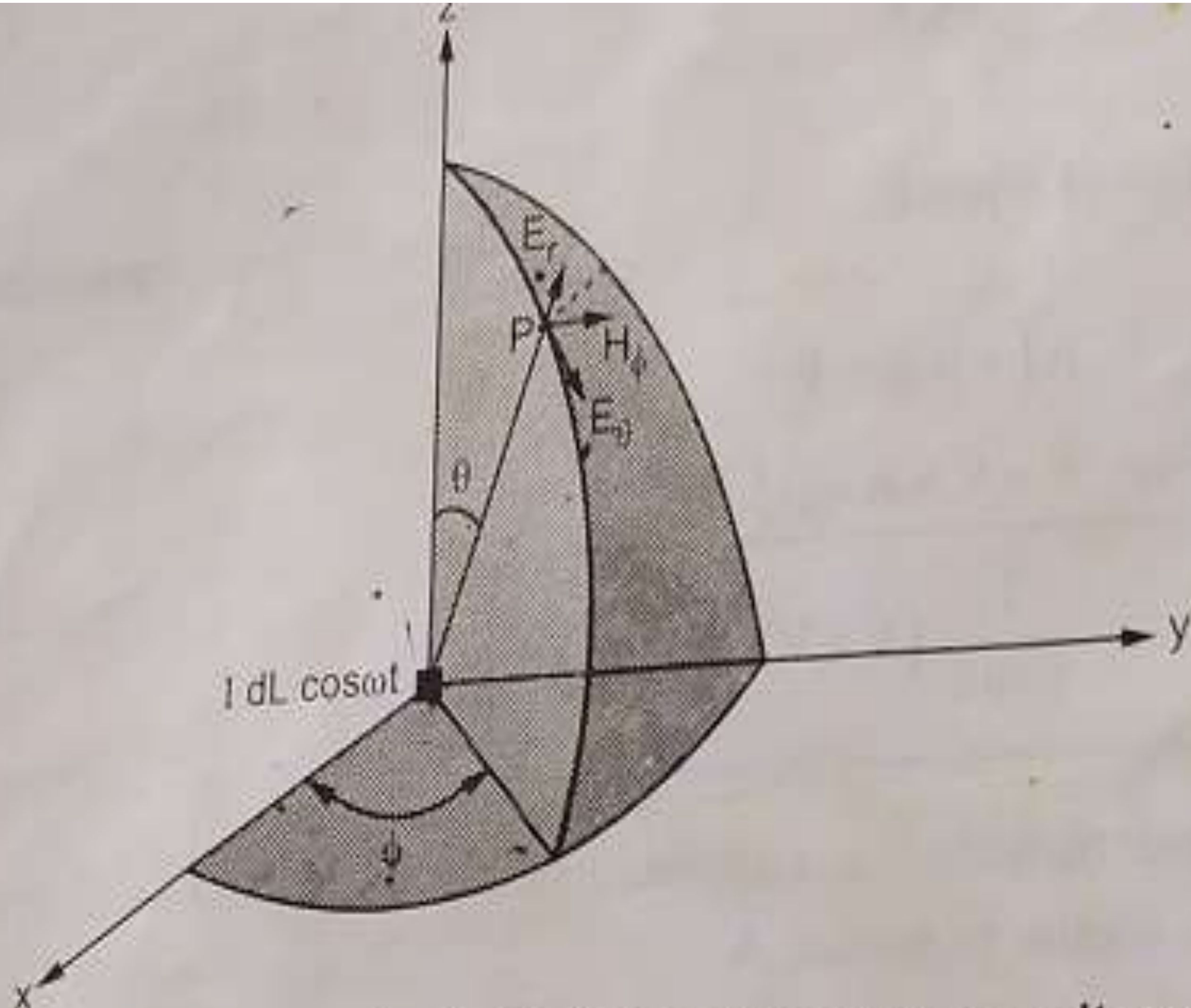


Fig. 1.3 Electromagnetic field at point P when an alternating current element I dL cos ωt placed at origin

Here the vector potential is retarded in time by $\frac{r}{v}$ sec, where v is the velocity

propagation. As the current element is placed along the z-axis, the vector poten will also have only one component in positive z-direction. Hence we can write,

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

From equation (2) it is clear that the component of vector potential A_z can obtained by integrating the current density $\bar{J}$ over the volume. This inclu integration over the cross section area of an element of wire and integration along length. But the integration of the current density $\bar{J}$ over a cross-section area yie current I. Now this current is assumed to be constant along the length dL, integration of $\bar{J}$ over the length dL gives value IdL. Thus mathematically we can wr

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

Substituting the value of integration from equation (3) in equation (2), the ve
potential in z-direction is given by,

$$A_z = \frac{\mu}{4\pi} \frac{I \, dL \cos\omega(t - r/v)}{r} \quad \text{--- (4)}$$

Now the magnetic field is given by

$$\mu \bar{H} = \nabla \times \bar{\mathbf{A}} \quad \text{... (5)}$$

As we are using spherical co-ordinate system, to find the curl of $\bar{\mathbf{A}}$, we must find the component of $\bar{\mathbf{A}}$ in r, θ and φ directions. From the Fig. 1.3, it is clear that,

$$A_r = A_z \cos\theta$$

$$A_\theta = -A_z \sin\theta \quad \text{... (6)}$$

Polar form, $A_\phi = 0$

~~Hence~~ $\bar{A}$ is given by,

$$\nabla \times \bar{\mathbf{A}} = \frac{1}{r\sin\theta}\left[\frac{\partial}{\partial\theta}(A_\phi \sin\theta) - \frac{\partial A_\theta}{\partial\phi}\right]\bar{\mathbf{a}}_r +$$

$$\frac{1}{r}\left[\frac{1}{\sin\theta}\frac{\partial}{\partial\phi}A_r - \frac{\partial}{\partial r}(r A_\phi)\right]\bar{\mathbf{a}}_\theta +$$

$$\frac{1}{r}\left[\frac{\partial}{\partial r}(r A_\theta) - \frac{\partial}{\partial\theta}(A_r)\right]\bar{\mathbf{a}}_\phi \quad \text{... (7)}$$

Now note that $A_\phi = 0$ and because of symmetry $\frac{\partial}{\partial \phi} = 0$ as no variation along ϕ direction. Thus first two terms in equation (7) can be neglected being zero.

$$\therefore \quad \nabla \times \bar{\mathbf{A}} = \frac{1}{r}\left[\frac{\partial (r A'_\theta)}{\partial r} - \frac{\partial A_r}{\partial \theta}\right]\bar{\mathbf{a}}_\phi \qquad \dots (8)$$

Putting values of A_θ and A_r, from equation (6) we get,

$$\nabla \times \bar{\mathbf{A}} = \frac{1}{r}\left[\frac{\partial}{\partial r}\{r(-A_z \sin\theta)\} - \frac{\partial}{\partial \theta}\{A_z \cos\theta\}\right]\bar{\mathbf{a}}_\phi$$

Substituting value of A_z,

$$\nabla \times \bar{\mathbf{A}} = \frac{\mu}{r}\left[\frac{\partial}{\partial r}\left\{(-r)\sin\theta \frac{I\, dL \cos\omega\left(t - \frac{r}{v}\right)}{4\pi r}\right\} - \right.$$

$$\left.\frac{\partial}{\partial \theta}\left\{\cos\theta \frac{I\, dL \cos\omega\left(t - \frac{r}{v}\right)}{4\pi r}\right\}\right]\bar{\mathbf{a}}_\phi$$

$$\nabla \times \bar{\mathbf{A}} = \frac{\mu\, I\, dL}{4\pi r}\left\{\left[(+\sin\theta)\left\{\frac{-\omega \sin\omega\left(t-\frac{r}{v}\right)}{+v}\right\}\right] - \left[\frac{(-\sin\theta)}{r}\left\{\cos\omega\left(t-\frac{r}{v}\right)\right\}\right]\right\}\bar{\mathbf{a}}_\phi$$

$$\nabla \times \bar{\mathbf{A}} = \frac{\mu\, I\, dL \sin\theta}{4\pi r}\left[\frac{-\omega \sin\omega\left(t-\frac{r}{v}\right)}{v} + \frac{\cos\omega\left(t-\frac{r}{v}\right)}{r}\right]\bar{\mathbf{a}}_\phi$$

$$\nabla \times \bar{\mathbf{A}} = \frac{\mu\, I\, dL \sin\theta}{4\pi}\left[\frac{-\omega \sin\omega\left(t-\frac{r}{v}\right)}{r v} + \frac{\cos\omega\left(t-\frac{r}{v}\right)}{r^2}\right]\bar{\mathbf{a}}_\phi$$

Hence the magnetic field $\overline{\mathbf{H}}$ is given by

$$\overline{\mathbf{H}} = \frac{1}{\mu}\left[\nabla \times \overline{\mathbf{A}}\right]$$

Putting value of $\nabla \times \overline{\mathbf{A}}$ from equation (9), we get,

$$\overline{\mathbf{H}} = \frac{I\,dL\sin\theta}{4\pi}\left[\frac{-\omega\sin\omega\left(t-\frac{r}{v}\right)}{rv}+\frac{\cos\omega\left(t-\frac{r}{v}\right)}{r^2}\right]\overline{\mathbf{a}}_\phi$$

Equation (10) indicates that the magnetic field $\overline{\mathbf{H}}$ exists only in ϕ direction.

$$\therefore \quad H_\phi = \frac{I\,dL\sin\theta}{4\pi}\left[\frac{-\omega\sin\omega\left(t-\frac{r}{v}\right)}{rv}+\frac{\cos\omega\left(t-\frac{r}{v}\right)}{r^2}\right]$$

Let $\left(t-\frac{r}{v}\right)=t'$, substituting the value in equation (11), we get,

$$H_\phi = \frac{I\,dL\sin\theta}{4\pi}\left[\frac{-\omega\sin\omega t'}{rv}+\frac{\cos\omega t'}{r^2}\right]$$

After calculating the magnetic field, now let us calculate the electric field given by,

$$\nabla \times \overline{H} = \varepsilon \frac{\partial \overline{E}}{\partial t}$$

$$\therefore \quad \partial \overline{E} = \frac{1}{\varepsilon}\left(\nabla \times \overline{H}\right) dt \qquad \text{... separating variables}$$

Integrating with respect to corresponding variables, we get,

$$\overline{E} = \frac{1}{\varepsilon}\int \nabla \times \overline{H}\ dt \qquad \text{... (13)}$$

Let us calculate each term of $\nabla \times \overline{H}$ separately.

From the definition of curl of a vector, the component in $\bar{a}_r$ direction is given by

$$\left(\nabla \times \overline{H}\right)_r = \frac{1}{r \sin\theta}\left[\frac{\partial H_\phi \sin\theta}{\partial \theta} - \frac{\partial H_\theta}{\partial \phi}\right]$$

But $\quad \dfrac{\partial}{\partial \phi} = 0$

$$\therefore \quad \left(\nabla \times \overline{H}\right)_r = \frac{1}{r \sin\theta}\left[\frac{\partial}{\partial \theta}\{H_\phi \sin\theta\}\right]$$

Substituting value of H_ϕ from equation (12),

$$\left(\nabla \times \overline{\mathbf{H}}\right)_r = \frac{1}{r\sin\theta}\left[\frac{\partial}{\partial\theta}\left\{\frac{I\,dL\sin\theta}{4\pi}\left[\frac{-\omega\sin\omega t'}{r v}+\frac{\cos\omega t'}{r^2}\right]\sin\theta\right\}\right]$$

$$\therefore \quad \left(\nabla \times \overline{\mathbf{H}}\right)_r = \frac{1}{r\sin\theta}\cdot\frac{I\,dL}{4\pi}\left[\frac{-\omega\sin\omega t'}{r v}+\frac{\cos\omega t'}{r^2}\right]\left\{\frac{\partial}{\partial\theta}\sin^2\theta\right\}$$

$$\therefore \quad \left(\nabla \times \overline{\mathbf{H}}\right)_r = \frac{I\,dL}{(r\sin\theta)\,4\pi}\left[\frac{-\omega\sin\omega t'}{r v}+\frac{\cos\omega t'}{r^2}\right](2\sin\theta\cos\theta)$$

$$\therefore \quad \left(\nabla \times \overline{\mathbf{H}}\right)_r = \frac{2\,I\,dL\cos\theta}{4\pi}\left[\frac{-\omega\sin\omega t'}{v r^2}+\frac{\cos\omega t'}{r^3}\right] \qquad \ldots (14)$$

Let us calculate the component in $\overline{\mathbf{a}}_\theta$ direction

$$\left(\nabla \times \overline{\mathbf{H}}\right)_\theta = \frac{1}{r}\left[\frac{1}{\sin\theta}\frac{\partial H_r}{\partial\phi}-\frac{\partial(r H_\phi)}{\partial r}\right]$$

But again $\frac{\partial}{\partial\phi} = 0$

$$\therefore \quad \left(\nabla \times \overline{\mathbf{H}}\right)_\theta = \frac{1}{r}\left[-\frac{\partial}{\partial r}\{r H_\phi\}\right]$$

Substituting value of H_ϕ from equation (11),

$$(\nabla \times \overline{H})_\theta = \frac{-I\,dL\sin\theta}{4\pi r}\frac{\partial}{\partial r}\left\{\frac{-r\omega\sin\omega\left(t-\frac{r}{v}\right)}{rv}+\frac{r\cos\omega\left(t-\frac{r}{v}\right)}{r^2}\right\}$$

$$\therefore (\nabla \times \overline{H})_\theta = \frac{-I\,dL\sin\theta}{4\pi r}\frac{\partial}{\partial r}\left\{\frac{-\omega\sin\omega\left(t-\frac{r}{v}\right)}{v}+\frac{\cos\omega\left(t-\frac{r}{v}\right)}{r}\right\}$$

$$(\nabla \times \overline{H})_\theta = -\frac{I\,dL\sin\theta}{4\pi r}\left\{\left[\frac{-\omega\cos\omega\left(t-\frac{r}{v}\right)}{v}\right]\left[-\frac{\omega}{v}\right]+\right.$$

$$\left.\frac{1}{r^2}\left[(r)\sin\omega\left(t-\frac{r}{v}\right)\left(\frac{\omega}{v}\right)-\cos\omega\left(t-\frac{r}{v}\right)\right]\right\}$$

$$\therefore (\nabla \times \overline{H})_\theta = -\frac{I\,dL\sin\theta}{4\pi}\left[\frac{\omega^2\cos\omega\left(t-\frac{r}{v}\right)}{v^2 r}+\frac{\omega\sin\omega\left(t-\frac{r}{v}\right)}{v\,r^2}-\frac{\cos\omega\left(t-\frac{r}{v}\right)}{r^3}\right]$$

Finally the component of $\left(\nabla \times \overline{H}\right)$ in $\bar{a}_\phi$ direction is zero.

From equation (13), the component of $\overline{E}$ in $\bar{a}_r$ direction is given by

$$E_r = \frac{1}{\varepsilon}\int \left(\nabla \times \overline{H}\right)_r dt$$

Putting value of $\left(\nabla \times \overline{H}\right)_r$ from equation (14),

$$E_r = \frac{1}{\varepsilon}\int \frac{2\,I\,dL\cos\theta}{4\pi}\left[\frac{-\omega\sin\omega t'}{v\,r^2}+\frac{\cos\omega t'}{r^3}\right]dt$$

$$= \frac{1}{\varepsilon}\int \frac{2\,I\,dL\cos\theta}{4\pi}\left[\frac{-\omega\sin\omega\left(t-\frac{r}{v}\right)}{v\,r^2}+\frac{\cos\omega\left(t-\frac{r}{v}\right)}{r^3}\right]dt$$

$$= \frac{2\,I\,dL\cos\theta}{4\pi\varepsilon}\left[\frac{\omega\cos\omega\left(t-\frac{r}{v}\right)}{v\,r^2}\left(\frac{1}{\omega}\right)+\frac{\sin\omega\left(t-\frac{r}{v}\right)}{r^3}\left(\frac{1}{\omega}\right)\right]$$

$$= \frac{2\,I\,dL\cos\theta}{4\pi\varepsilon}\left[\frac{\cos\omega\left(t-\frac{r}{v}\right)}{v\,r^2}+\frac{\sin\omega\left(t-\frac{r}{v}\right)}{\omega\,r^3}\right]$$

Put $\left(t-\frac{r}{v}\right)=t'$

$$\therefore \qquad E_r = \frac{2\,I\,dL\cos\theta}{4\pi\varepsilon}\left[\frac{\cos\omega t'}{v r^2}+\frac{\sin\omega t'}{\omega r^3}\right] \qquad \ldots (16)$$

Similarly from equation (13), the component of $\bar{\mathbf{E}}$ in $\bar{\mathbf{a}}_\theta$ direction is given by,

$$E_\theta = \frac{1}{\varepsilon}\int \left(\nabla\times\overline{\mathbf{H}}\right)_\theta dt$$

Substituting value of $\left(\nabla\times\overline{\mathbf{H}}\right)_\theta$ from equation (15),

$$E_\theta = \frac{1}{\varepsilon}\int \frac{-I\,dL\sin\theta}{4\pi}\left[\frac{\omega^2\cos\omega\left(t-\frac{r}{v}\right)}{v^2 r}+\frac{\omega\sin\omega\left(t-\frac{r}{v}\right)}{v r^2}-\frac{\cos\omega\left(t-\frac{r}{v}\right)}{r^3}\right]dt$$

$$\therefore E_\theta = \frac{-I\,dL\sin\theta}{4\pi\varepsilon}\int\left[\frac{\omega^2\cos\omega\left(t-\frac{r}{v}\right)}{v^2 r}+\frac{\omega\sin\omega\left(t-\frac{r}{v}\right)}{v r^2}-\frac{\cos\omega\left(t-\frac{r}{v}\right)}{r^3}\right]dt$$

$$\therefore E_\theta = \frac{-I\,dL\sin\theta}{4\pi\varepsilon}\left[\frac{\omega^2\sin\omega\left(t-\frac{r}{v}\right)}{v^2 r}\left(\frac{1}{\omega}\right)+\frac{-\omega\cos\omega\left(t-\frac{r}{v}\right)}{v r^2}\left(\frac{1}{\omega}\right)-\frac{\sin\omega\left(t-\frac{r}{v}\right)}{r^3}\left(\frac{1}{\omega}\right)\right]$$

$$\therefore E_\theta = \frac{-I\,dL\sin\theta}{4\pi\varepsilon}\left[\frac{\omega\sin\omega\left(t-\frac{r}{v}\right)}{v^2 r}-\frac{\cos\omega\left(t-\frac{r}{v}\right)}{v r^2}-\frac{\sin\omega\left(t-\frac{r}{v}\right)}{\omega r^3}\right]$$

$$\therefore E_\theta = \frac{I\,dL\sin\theta}{4\pi\varepsilon}\left[\frac{-\omega\sin\omega t'}{v^2 r}+\frac{\cos\omega t'}{v r^2}+\frac{\sin\omega t'}{\omega r^3}\right] \qquad \ldots (17)$$

Topic 16

Radiation from Half wave Dipole/Monopole

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

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

1.8.1 Power Radiated by the Half Wave Dipole and the Monopole

A **dipole antenna** is a vertical radiator fed in the centre. It produces maxim radiation in the plane normal to the axis. For such a dipole antenna, the len specified is the overall length.

The vertical antenna of height $H = \frac{L}{2}$ produces the radiation characteristics ab the plane which is similar to that produced by the dipole antenna of length L= The vertical antenna is referred as a monopole.

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

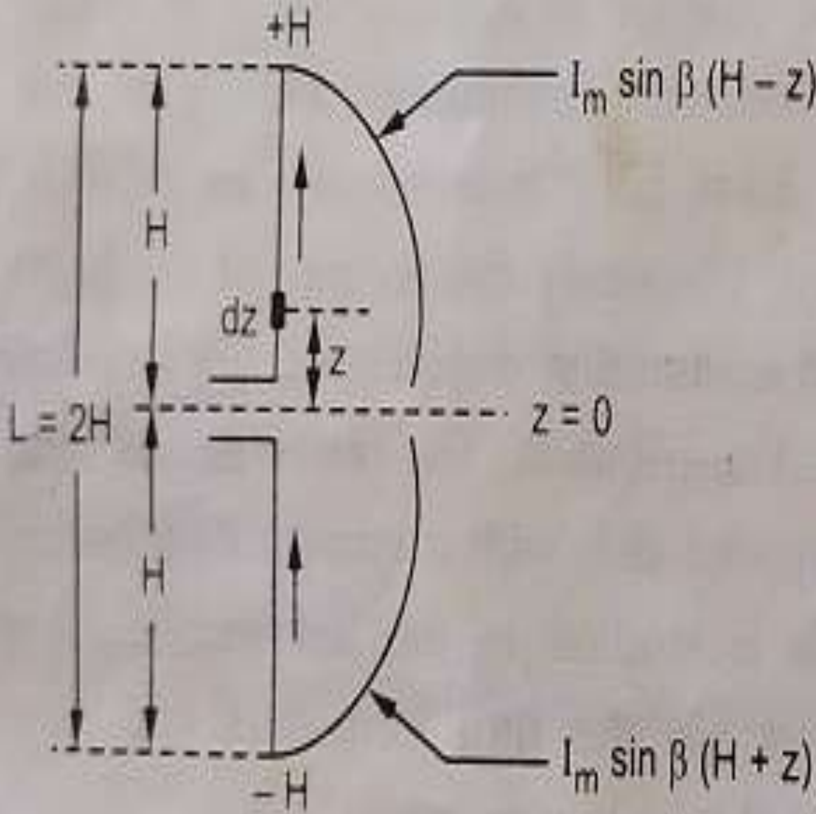


Fig. 1.9 Assumed sinusoidal current distribution in half wave dipole

The **half wave dipole** consists two legs each of length $\frac{L}{2}$. The physical length of the half wave dipole at the frequency of operation is $\frac{\lambda}{2}$ in the free space.

The **quarter wave monopole** consists of a single vertical leg erected on the perfect ground i.e. perfect conductor. The length of the leg of the quarter wave monopole is $\frac{\lambda}{4}$.

For the calculation of the electromagnetic fields, the assumed sinusoidal current distributions along the half wave dipole and quarter wave monopole are as shown in the Fig. 1.9 and Fig. 1.10 respectively.

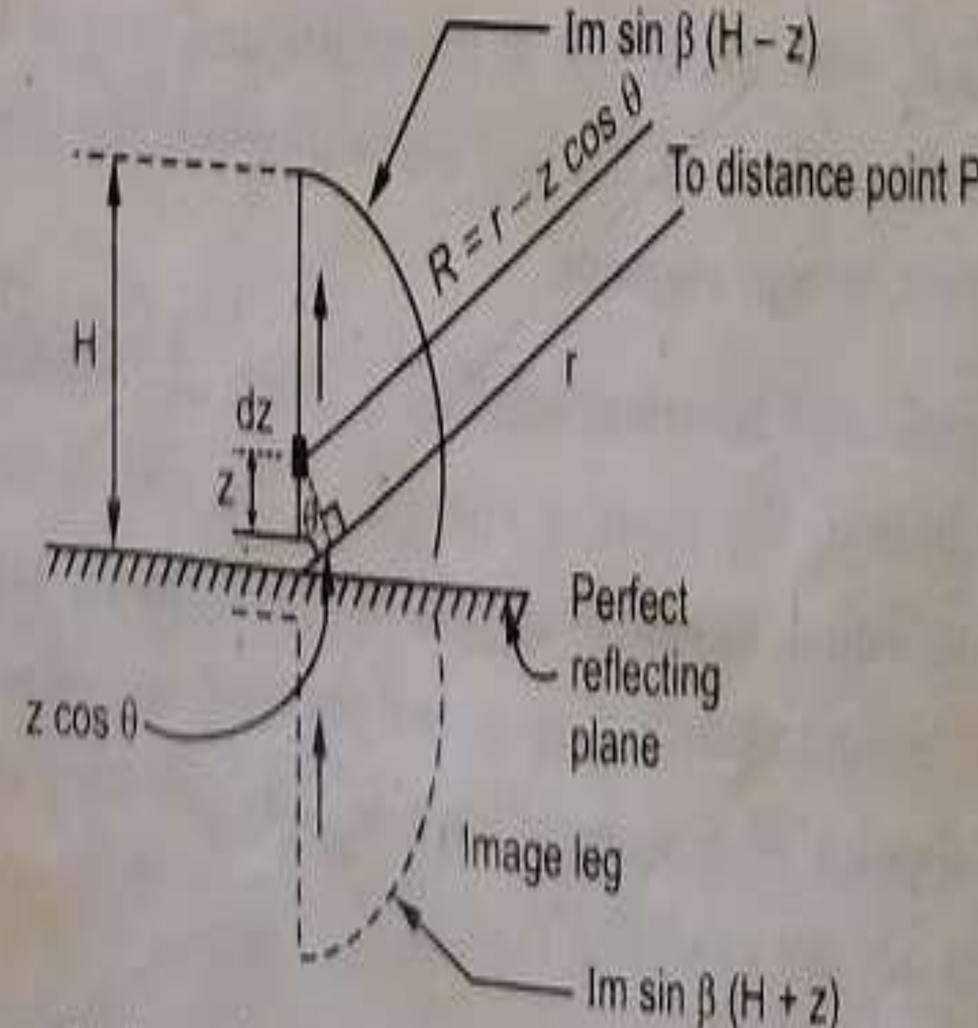


Fig. 1.10 Assumed sinusoidal current distribution in quarter wave monopole

Consider the assumed sinusoidal current distribution in the quarter wave monopole and half wave dipole. The current element Idz is placed at a distance z from z = 0 plane. Let I_m be the maximum value of the current in the current element. Then the sinusoidal current distribution is given by,

$$I = I_m \sin\beta(H - z) \quad \text{for} \quad z > 0 \qquad \ldots (1)$$

and

$$I = I_m \sin\beta(H + z) \quad \text{for} \quad z < 0 \qquad \ldots (2)$$

Consider a point P located at a far distance from the current element I dz. Then the vector potential at point P due to the current element I dz is given by,

$$d A_z = \frac{\mu I}{4\pi R} e^{-j\beta R} \, dz \qquad \ldots (3)$$

R is the distance of point P from the current element. The total vector potential at point P due to all such currents can be obtained by integrating the vector potential $d A_z$ (due to the current element I dz) over the total length of the antenna.

$$\therefore \quad A_z = \frac{\mu}{4\pi} \int_{-H}^{H} \frac{I}{R} e^{-j\beta R} \, dz$$

$$\therefore \quad A_z = \frac{\mu}{4\pi} \int_{-H}^{0} \frac{I}{R} e^{-j\beta R} \, dz + \frac{\mu}{4\pi} \int_{0}^{+H} \frac{I}{R} e^{-j\beta R} \, dz \qquad \ldots (4)$$

From equations (1) and (2) we can get the current distributions for $z < 0$ and $z > 0$. Thus substituting the values of I from equations (2) and (1) in the first and second term of equation (4) respectively.

$$\therefore \; A_z = \frac{\mu}{4\pi}\int_{-H}^{0} \frac{I_m \sin\beta(H+z)}{R} e^{-j\beta R}\, dz + \frac{\mu}{4\pi}\int_{0}^{+H} \frac{I_m \sin\beta(H-z)}{R} e^{-j\beta R}\, dz \qquad \ldots (5)$$

At this point we have to make certain assumptions so as to calculate distant or radiation field. Assuming $R \approx r$, replace R in the denominators only by r. For the large distances of point P from the current element, all the lines to the point P can be assumed to be parallel.

Then we can write,

$$R = r - z\cos\theta$$

So replace R in the numerator terms by $(r - z\cos\theta)$. Then the equation (5) changes to

$$A_z = \frac{\mu}{4\pi}\int_{-H}^{0} \frac{I_m \sin\beta(H+z)}{r} e^{-j\beta(r - z\cos\theta)}\, dz + \frac{\mu}{4\pi}\int_{0}^{+H} \frac{I_m \sin\beta(H-z)}{r} e^{-j\beta(r - z\cos\theta)}\, dz$$

$$\therefore \quad A_z = \frac{\mu I_m}{4\pi r}\left[\int_{-H}^{0} \sin\beta(H+z)\, e^{-j\beta r}\, e^{j\beta z\cos\theta}\, dz + \int_{0}^{+H} \sin\beta(H-z)\, e^{-j\beta r}\, e^{j\beta z\cos\theta}\, dz\right]$$

$$\therefore \ A_z = \frac{\mu I_m\, e^{-j\beta r}}{4\pi r}\left[\int_{-H}^{0} \sin\beta(H+z)\, e^{j\beta z\cos\theta}\, dz + \int_{0}^{+H} \sin\beta(H-z)\, e^{j\beta z\cos\theta}\, dz\right]$$

For quarter wave monopole,

$$H = \frac{\lambda}{4} \text{ and}$$

$$\beta = \frac{2\pi}{\lambda}$$

$$\therefore \quad \sin\beta(H+z) = \sin(\beta H + \beta z) = \sin\left(\frac{\pi}{2} + \beta z\right)$$

and $$\sin\beta(H-z) = \sin(\beta H - \beta z) = \sin\left(\frac{\pi}{2} - \beta z\right)$$

But $$\sin\left(\frac{\pi}{2} + \beta z\right) = \sin\left(\frac{\pi}{2} - \beta z\right) = \cos\beta z$$

Substituting values of sine terms in equation (6) we get,

$$A_z = \frac{\mu I_m e^{-j\beta r}}{4\pi r}\left[\int_{-H}^{0} \cos\beta z\, e^{j\beta z\cos\theta}\, dz + \int_{0}^{H} \cos\beta z\, e^{j\beta z\cos\theta}\, dz\right]$$

Now $\int_{-H}^{0} e^{+j\theta}\, d\theta = \int_{0}^{H} e^{-j\theta} d\theta$ Hence using this property, changing limits of integration of the first term.

$$\therefore \quad A_z = \frac{\mu I_m e^{-j\beta r}}{4\pi r}\left[\int_{0}^{H} \cos(-\beta z)\, e^{-j\beta z\cos\theta}\, dz + \int_{0}^{H} \cos\beta z\, e^{j\beta z\cos\theta}\, dz\right]$$

$$\therefore \quad A_z = \frac{\mu I_m e^{-j\beta r}}{4\pi r}\left[\int_{0}^{H} \cos\beta z\, e^{-j\beta z\cos\theta}\, dz + \int_{0}^{H} \cos\beta z\, e^{j\beta z\cos\theta}\, dz\right]$$

$$\ldots \cos(-\theta) = \cos\theta$$

$$\therefore \quad A_z = \frac{\mu I_m e^{-j\beta r}}{4\pi r}\left[\int_{0}^{H} \cos\beta z\left(e^{j\beta z\cos\theta} + e^{-j\beta z\cos\theta}\right) dz\right] \quad \ldots (8)$$

By Euler's identity,

$$e^{j\beta z\cos\theta} = \cos(\beta\, z\cos\theta) + j\sin(\beta\, z\cos\theta)$$

$$e^{-j\beta z\cos\theta} = \cos(\beta\, z\cos\theta) - j\sin(\beta\, z\cos\theta)$$

$$e^{j\beta z\cos\theta} + e^{-j\beta z\cos\theta} = 2\cos(\beta\, z\cos\theta) \qquad \dots (9)$$

Putting the value of the term inside the bracket in equation (8), we get,

$$A_z = \frac{\mu\, I_m\, e^{-j\beta r}}{4\pi r}\int_0^H (\cos\beta\, z)\, 2[\cos(\beta\, z\cos\theta)]\,dz$$

$$\therefore \quad A_z = \frac{\mu\, I_m\, e^{-j\beta r}}{4\pi r}\int_0^H 2\cos\beta\, z\cos(\beta\, z\cos\theta)\,dz \qquad \dots (10)$$

From the trigonometric identity,

$$2\cos A\cos B = \cos(A - B) + \cos(A + B)$$

Using this property in equation (10), we get,

$$A_z = \frac{\mu\, I_m\, e^{-j\beta r}}{4\pi r}\int_0^H \{\cos[\beta\, z + \beta\, z\cos\theta] + \cos[\beta\, z - \beta\, z\cos\theta]\}\,dz$$

$$\therefore \quad A_z = \frac{\mu\, I_m\, e^{-j\beta r}}{4\pi r}\cdot\int_0^{H=\lambda/4} \{\cos\beta\, z(1+\cos\theta) + \cos\beta\, z(1-\cos\theta)\}\,dz \qquad \dots (11)$$

Integrating with respect to z and putting value of $H = \lambda / 4$ as limit, we get,

$$A_z = \frac{\mu I_m e^{-j\beta r}}{4\pi r}\left[\frac{\sin\beta z(1+\cos\theta)}{\beta(1+\cos\theta)} + \frac{\sin\beta z(1-\cos\theta)}{\beta(1-\cos\theta)}\right]_0^{\lambda/4}$$

Finding LCM,

$$A_z = \frac{\mu I_m e^{-j\beta r}}{4\pi r}\left[\frac{[\sin\beta z(1+\cos\theta)](1-\cos\theta) + [\sin\beta z(1-\cos\theta)](1+\cos\theta)}{\beta\left(1-\cos^2\theta\right)}\right]_0^{\lambda/4}$$

$$\therefore A_z = \frac{\mu I_m e^{-j\beta r}}{4\pi\beta r}\left[\frac{(1-\cos\theta)\left[\sin\frac{\pi}{2}(1+\cos\theta)\right] + (1+\cos\theta)\left[\sin\frac{\pi}{2}(1-\cos\theta)\right]}{\sin^2\theta}\right] \quad \ldots (12)$$

$$\ldots \beta z = \frac{\pi}{2} \text{ and } \left(1-\cos^2\theta\right) = \sin^2\theta$$

Again using property,

$$\sin(\pi/2+\theta) = \sin(\pi/2-\theta) = \cos\theta$$

$$\sin\left(\frac{\pi}{2}+\frac{\pi}{2}\cos\theta\right) = \sin\left(\frac{\pi}{2}-\frac{\pi}{2}\cos\theta\right) = \cos\left(\frac{\pi}{2}\cos\theta\right)$$

Substituting values of the sine terms from equation (13) in equation (12), we ge

$$A_z = \frac{\mu I_m e^{-j\beta r}}{4\pi\beta r}\left[\frac{(1-\cos\theta)\cdot\cos\left(\frac{\pi}{2}\cos\theta\right)+(1+\cos\theta)\cdot\cos\left(\frac{\pi}{2}\cos\theta\right)}{\sin^2\theta}\right]$$

$$\therefore \quad A_z = \frac{\mu I_m e^{-j\beta r}}{4\pi\beta r}\left[\frac{2\cos\left(\frac{\pi}{2}\cos\theta\right)}{\sin^2\theta}\right]$$

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

After finding the vector potential, the next step is to find the magnetic field us Maxwell's equation. The ϕ components of $\overline{\mathbf{H}}$ is given by,

$$H_\phi = \frac{1}{\mu}(\nabla \times \overline{\mathbf{A}})\phi$$

But $$(\nabla \times \overline{\mathbf{A}})\phi = \frac{1}{r}\left[\frac{\partial}{\partial r}(r A_\theta) - \frac{\partial A_r}{\partial \theta}\right]$$

Now the current element is placed along z-axis,

$$\therefore \quad A_\theta = -A_z \sin\theta ,$$

$$A_r = 0$$

$$(\nabla \times \overline{\mathbf{A}})_\phi = \frac{1}{r}\left[\frac{\partial}{\partial r}(r)(-A_z \sin\theta)\right]$$

Substituting value of $(\nabla \times \overline{\mathbf{A}})_\phi$ in equation (15), we get,

$$H_\phi = \frac{1}{\mu}\left[\frac{1}{r}\frac{\partial}{\partial r}(-r A_z \sin\theta)\right]$$

Substituting value of A_z from equation (14) in equation (17), we get,

$$H_\phi = \frac{1}{\mu}\left[\frac{1}{r}\frac{\partial}{\partial r}\left\{-r\,\frac{\mu I_m e^{-j\beta r}}{2\pi\beta r}\left[\frac{\cos\left(\frac{\pi}{2}\cos\theta\right)}{\sin^2\theta}\right]\sin\theta\right\}\right]$$

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

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

$$\therefore\ H_\phi = \frac{j I_m e^{-j\beta r}}{2\pi r}\left[\frac{\cos\left(\frac{\pi}{2}\cos\theta\right)}{\sin\theta}\right] \quad \text{... (18)}$$

The magnitude of the magnetic field strength for the radiation field of a half wave dipole or quarter wave monopole is given by,

$$|H_\phi| = \frac{I_m}{2\pi r}\left[\frac{\cos\left(\frac{\pi}{2}\cos\theta\right)}{\sin\theta}\right] \quad \text{... (19)}$$

The electric field strength is related to the magnetic field strength by the relation,

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

For free space, $\eta = \eta_0 = 120\pi$

$$\therefore \quad E_\theta = (120\pi) H_\phi$$

Substituting value of H_ϕ from equation (18), we get,

$$E_\theta = (120\pi)\left\{\frac{j I_m\, e^{-j\beta r}}{2\pi r}\left[\frac{\cos\left(\frac{\pi}{2}\cos\theta\right)}{\sin\theta}\right]\right\}$$

$$\therefore \quad E_\theta = \frac{j60\, I_m\, e^{-j\beta r}}{r}\left[\frac{\cos\left(\frac{\pi}{2}\cos\theta\right)}{\sin\theta}\right]$$

The magnitude of the electric field strength for the radiation field of a half dipole or a quarter wave monopole is given by

$$|E_\theta| = \frac{60\, I_m}{r}\left[\frac{\cos\left(\frac{\pi}{2}\cos\theta\right)}{\sin\theta}\right]$$

Topic 17

Folded Dipole

- To achieve good matching to practical coaxial lines with 50 Ω or 75 Ω characteristic impedances, the most widely used dipole is half wavelength $\lambda/2$.
- It has an input impedance of Zin = 73 + j42.5 and directivity of $Dmax$ = 1.643.
- In practice, there are other very common transmission lines whose characteristic impedance is much higher than 50 Ω or 75 Ω.
- For example, a "twin-lead" transmission line is widely used for TV applications and has a characteristic impedance of about 300 Ω.
- One simple geometry that can achieve this is a folded wire which forms a very thin rectangular loop as shown in Fig. 1.

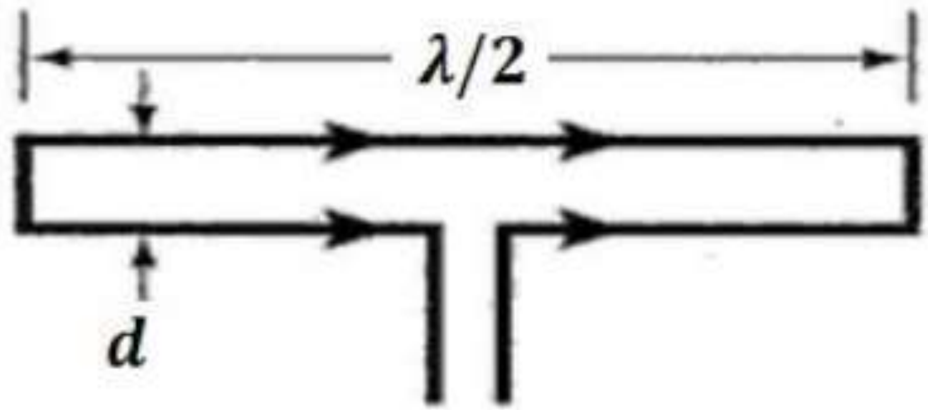


Fig. 1 2 Wire folded dipole antenna

- A folded dipole consists of two parallel $\lambda/2$ dipoles connected to each other at the ends. It is fed at the centre of one of the dipoles and the other dipole is shorted.
- The impedance of the folded dipole is four times greater than that of an isolated dipole of the same length, i.e., $\boldsymbol{Zf} = \boldsymbol{4Zd}$.

- **Advantages:**
  - Very high input impedance
  - Inherent impedance transformation property
  - Wide bandwidth
  - Acts as reactance compensation network

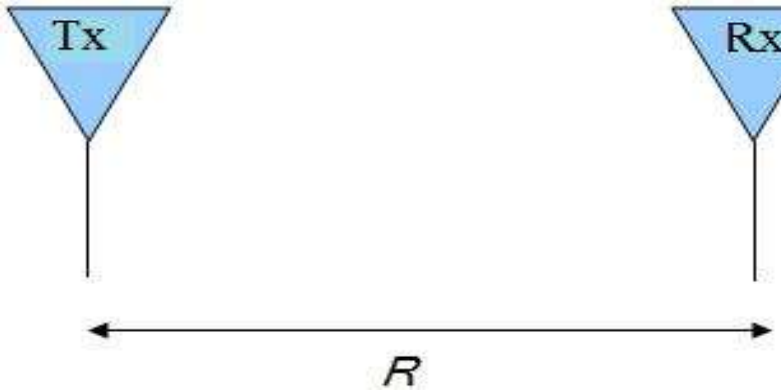
- **Applications:**
- The folded dipoles with parasitic elements can be used for wideband transmissions such as TV signals.
- It is used in Yagi-uda antenna as a driven element.

Topic 18

Friss Transmission Formula

Two Antennas in Free Space

- One of the most fundamental equations in antenna theory is the **Friis Transmission Equation**.
- The Friis Transmission Equation is used to calculate the power received from one antenna (with gain *G1*), when transmitted from another antenna (with gain *G2*) separated by a distance *R*, and operating at frequency *f*.
- Consider two antennas in free space (no obstructions nearby) separated by a distance *R*:



Power Density

- Assume that, P_T Watts of total power are delivered to the transmit antenna which is omni-directional, lossless and that the receive antenna is in the far field of the transmit antenna.
- Then the power density *p* (Watts per square meter) of the plane wave incident on the receive antenna a distance '*R*' from the transmit antenna is given by –

$$p = \frac{P_T}{4\pi R^2}$$

- If the transmit antenna has an antenna gain in the direction of the receive antenna given by G_T, then the power density becomes

$$p = \frac{P_T}{4\pi R^2} G_T$$

Power Received

- The gain term factors in the directionality and losses of a real antenna. Assume now that the receive antenna has an effective aperture given by A_{ER}. Then the power received by this antenna is given by -

$$P_R = \frac{P_T}{4\pi R^2} G_T A_{ER}$$

- Since the effective aperture for any antenna can also be expressed as –

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

- The resulting received power can be written as
- $$P_R = \frac{P_T G_T G_R \lambda^2}{(4\pi R)^2} \qquad \text{-------} \quad (1)$$
- This is known as the **Friis Transmission Formul*a***. It relates the free space path loss, antenna gains and wavelength to the received and transmitted powers.

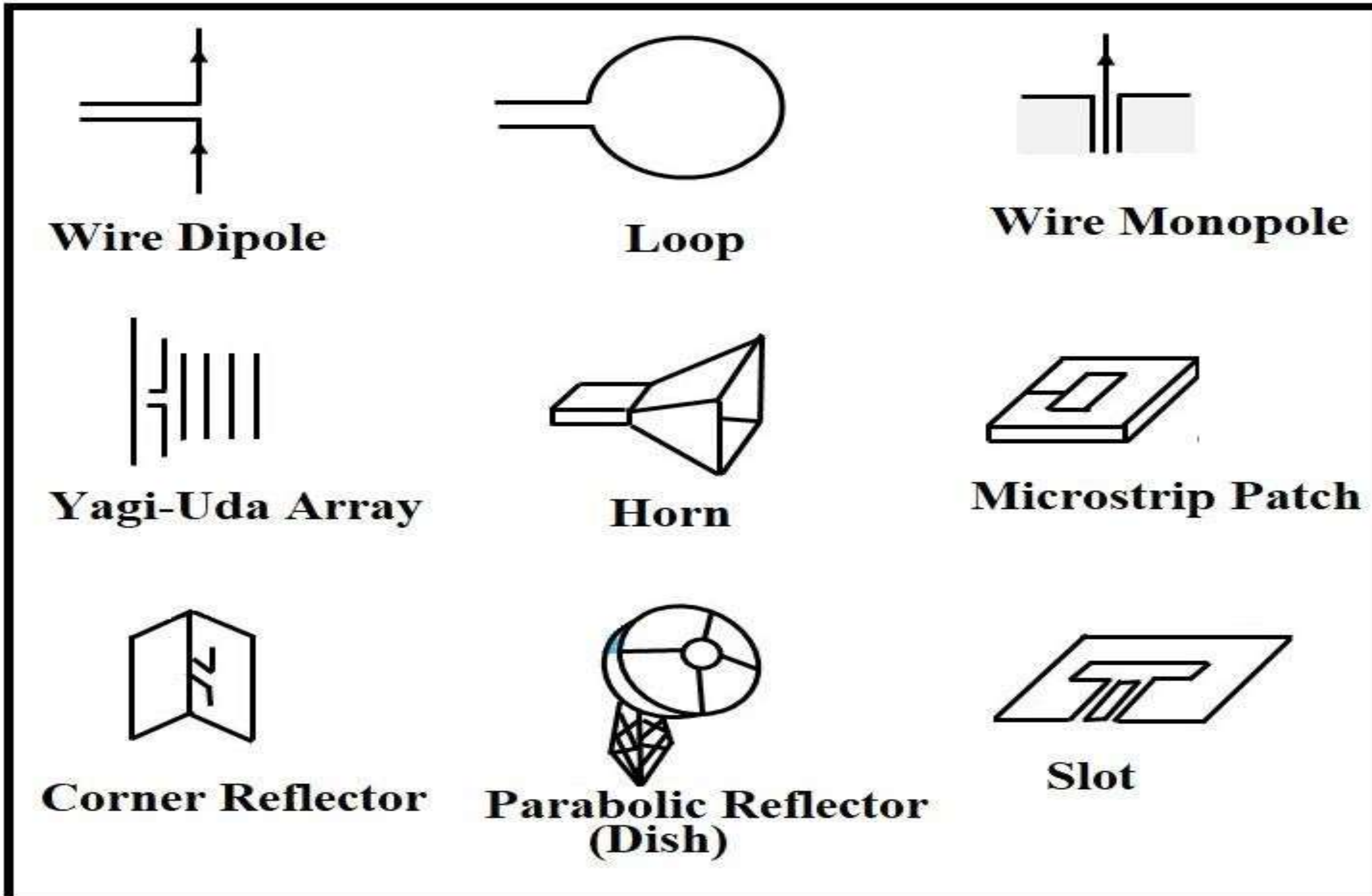
Second Form of Friss Transmission Formula

- Another useful form of Friis Transmission Equation is given in Equation [2]. Since wavelength and frequency are related by speed of light 'C', the Friis Transmission Formula can be expressed as –
- $$P_R = \frac{P_T G_T G_R c^2}{(4\pi R f)^2} \quad \text{------- (2)}$$
- The above equation shows that more power is lost at higher frequencies.
- For antennas with specified gains, the energy transfer will be highest at lower frequencies. The difference between the power received and the power transmitted is known as path loss.

Generalized Form of Friss TX Formula

- The Friis transmission equation indicates that the path loss is higher for higher frequencies.
- If the antennas are not polarization matched, the above received power can be multiplied by the polarization loss factor (PLF) to properly account for this mismatch. The second equation can be altered to produce a generalized Friis Transmission formula which includes polarization mismatch.
- $$P_R = (PLF) \cdot \frac{P_T G_T G_R c^2}{(4\pi R f)^2} \qquad \text{------ (3)}$$

Applications



Applications

- An antenna is mainly used as a metallic device for radiating or receiving **radio waves** which is basically used for transmitting signals, transmitting antenna is used to transmit information and for receiving signal, receiving antenna is used at receiver end to receive signals

UNIT – II

ANTENNA ARRAYS

Introduction to Antenna Arrays

Introduction to Array Antennas

- The radiation pattern of a single element is relatively wide & each element provides low values of directivity (gain).
- In many applications, it is necessary to design antennas with very directive characteristics (very high gains) to meet the demands of long distance communication.
- This can only be accomplished by increasing the electrical size of the antenna.
- Higher directivity is the basic requirement in point-to-point communication, radars and space applications.

- Enlarging the dimensions of single elements often leads to more directive characteristics.
- Another way to enlarge 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.
- Thus antenna array can be defined as the system of similar antennas directed to get required high directivity in the desired direction.

- The antenna array is said to be *linear* if the elements of the antenna array are equally spaced along a straight line.
- It is said to be uniform linear array if all the elements are fed with a current of equal magnitude with progressive uniform phase shift along the line.
- Five controls used to shape the overall pattern of the antenna. - Geometrical configuration of the overall array (linear, circular, rectangular, spherical, etc.)
- Relative displacement between the elements
- Excitation amplitude of the individual elements
- Excitation phase of the individual elements
- Relative pattern of the individual elements

Types of Antenna Arrays

Types of Antenna Arrays

- Practically various forms of the antenna array are used as radiating systems. They are;

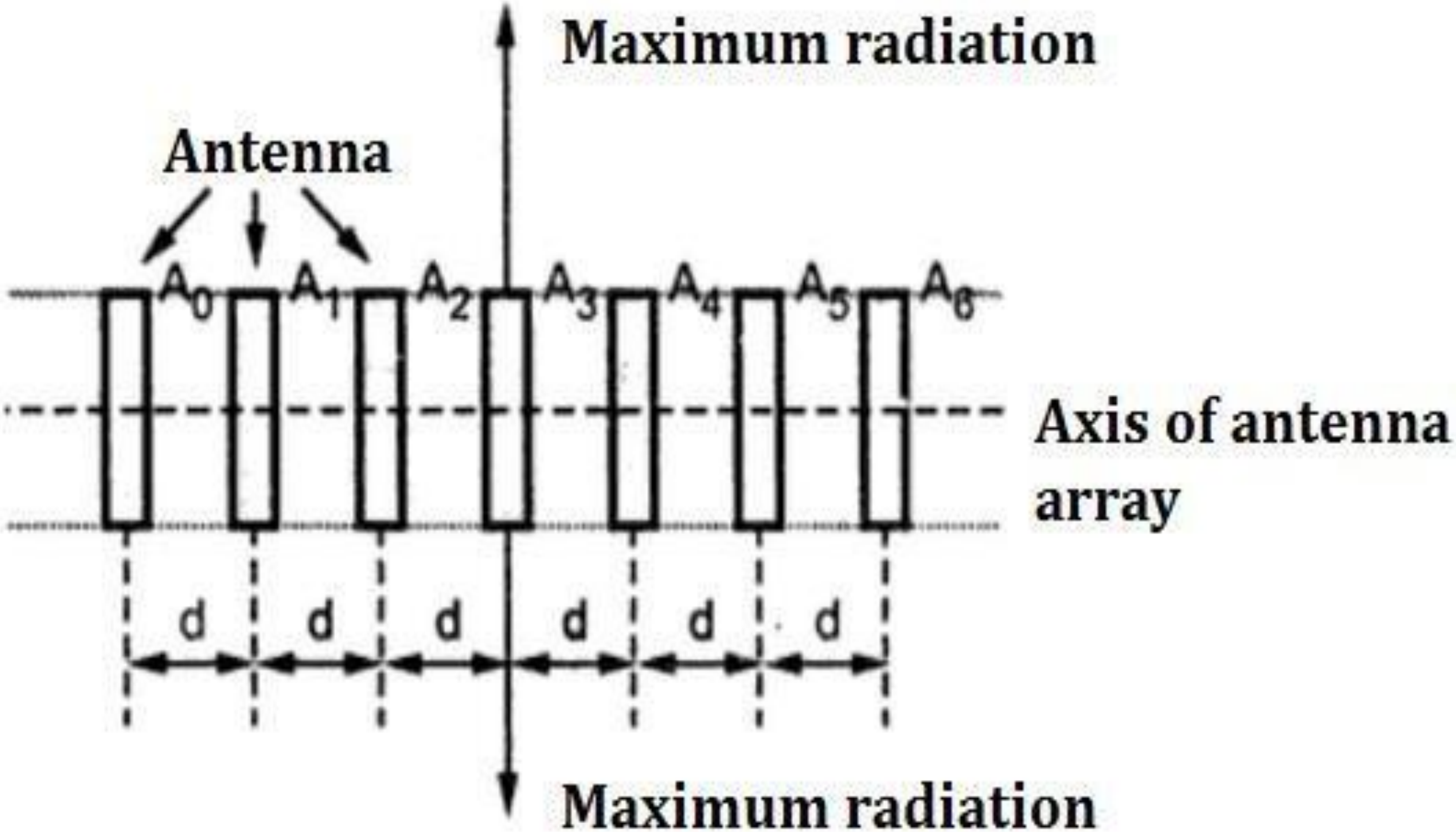
1. Broadside Array (BSA)
2. End-Fire Array (EFA)
3. Collinear Array
4. Parasitic Array

Broadside Array

- A typical arrangement of a Broadside array is shown in Figure 1.
- A broadside array consists number of identical antennas placed parallel to each other along a straight line and the direction of maximum radiation is always perpendicular to the plane consisting elements.
- This straight line is perpendicular to the axis of individual antenna. It is known as *axis of antenna array.*
- Thus each element is perpendicular to the axis of antenna array.

- All the individual antennas are spaced equally along the axis of antenna array.
- All the elements are fed with currents with equal magnitude and same phase.
- As the maximum radiation is directed in broadside direction i.e. perpendicular to the line of axis of array, the radiation pattern for the broadside array is bidirectional.
- Thus broadside array can be defined as the arrangement of antennas in which maximum radiation is in the direction perpendicular to the axis of array and plane containing the elements of array

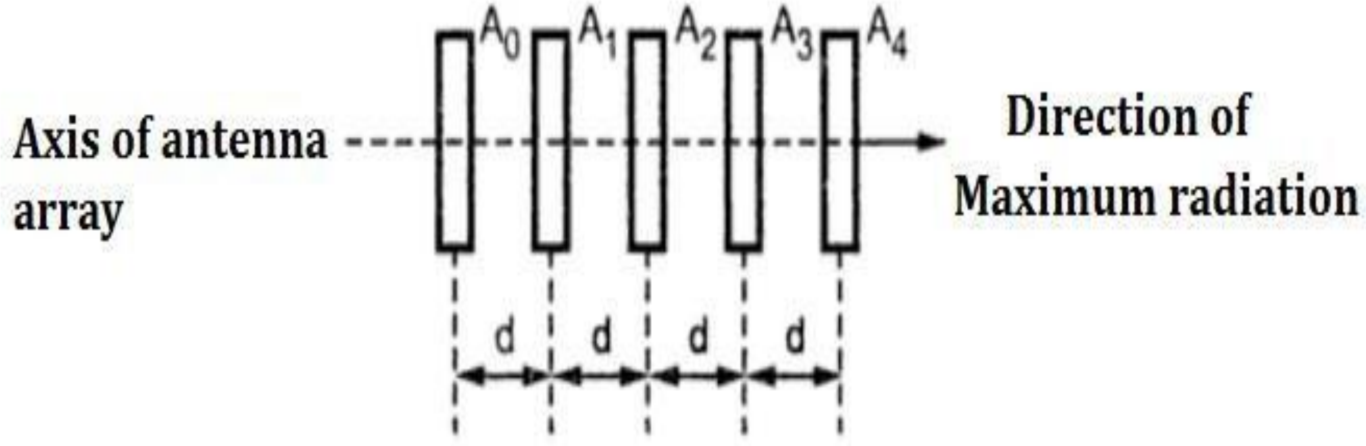
Figure 1 Broadside Array



End Fire Array

- The end fire array is very much similar to the broadside array from the point of view of arrangement.
- But the main difference is in the direction of maximum radiation.
- In broadside array, the direction of the maximum radiation is perpendicular to the axis of array;
- while in the end fire array, the direction of the maximum radiation is along the axis of array.
- Thus in the end fire array number of identical antennas are spaced equally along a line.

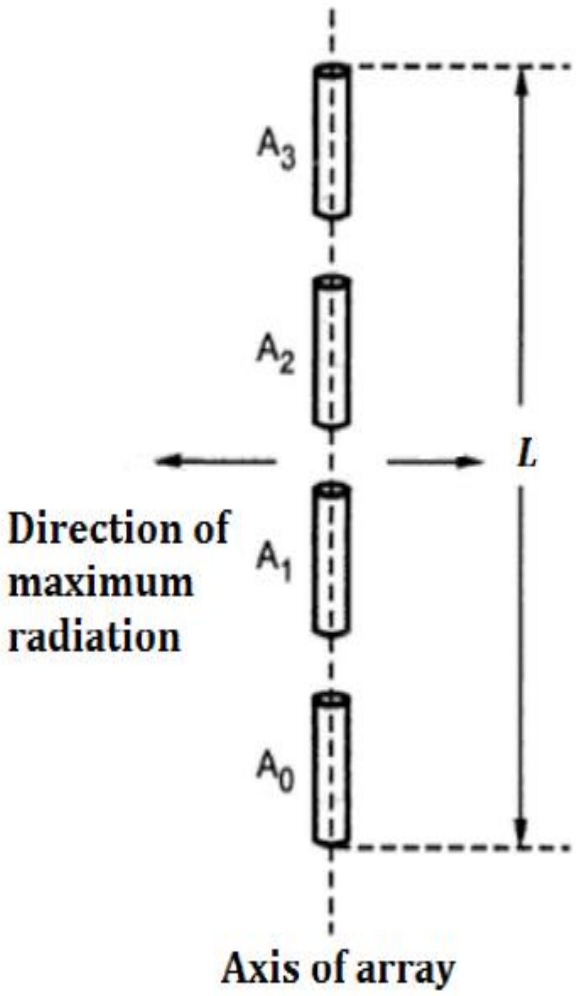
- All the antennas are fed individually with currents of equal magnitudes but their phases vary progressively along the line to get entire arrangement unidirectional finally. i.e. maximum radiation along the axis of array as shown in Figure 2.
- Thus end fire array can be defined as an array with direction of maximum radiation coincides with the direction of the axis of array to get unidirectional radiation.



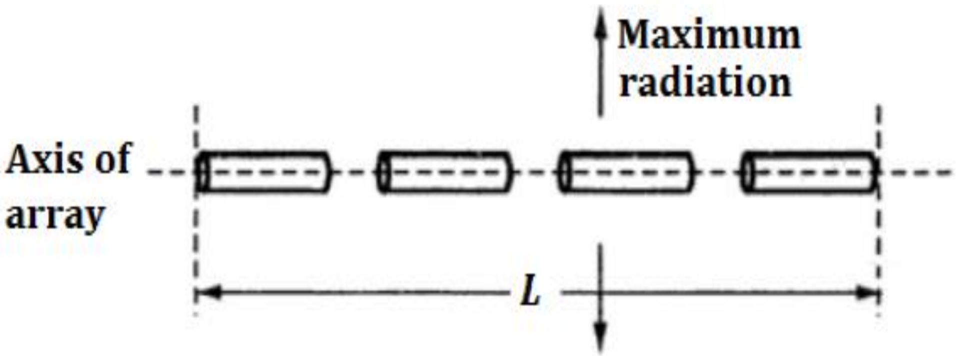
Co-linear Array

- As the name indicates, in the collinear array, the antennas are arranged co-axially i.e. the antennas are arranged end to end along, a single line as shown in Fig. 3-3 (a) and (b).
- The individual elements in the collinear array are fed with currents equal in magnitude and phase.
- This condition is similar to the broadside array.
- In collinear array the direction of maximum radiation is perpendicular to the axis of array.

Collinear Array



(a) Vertical



(b) Horizontal

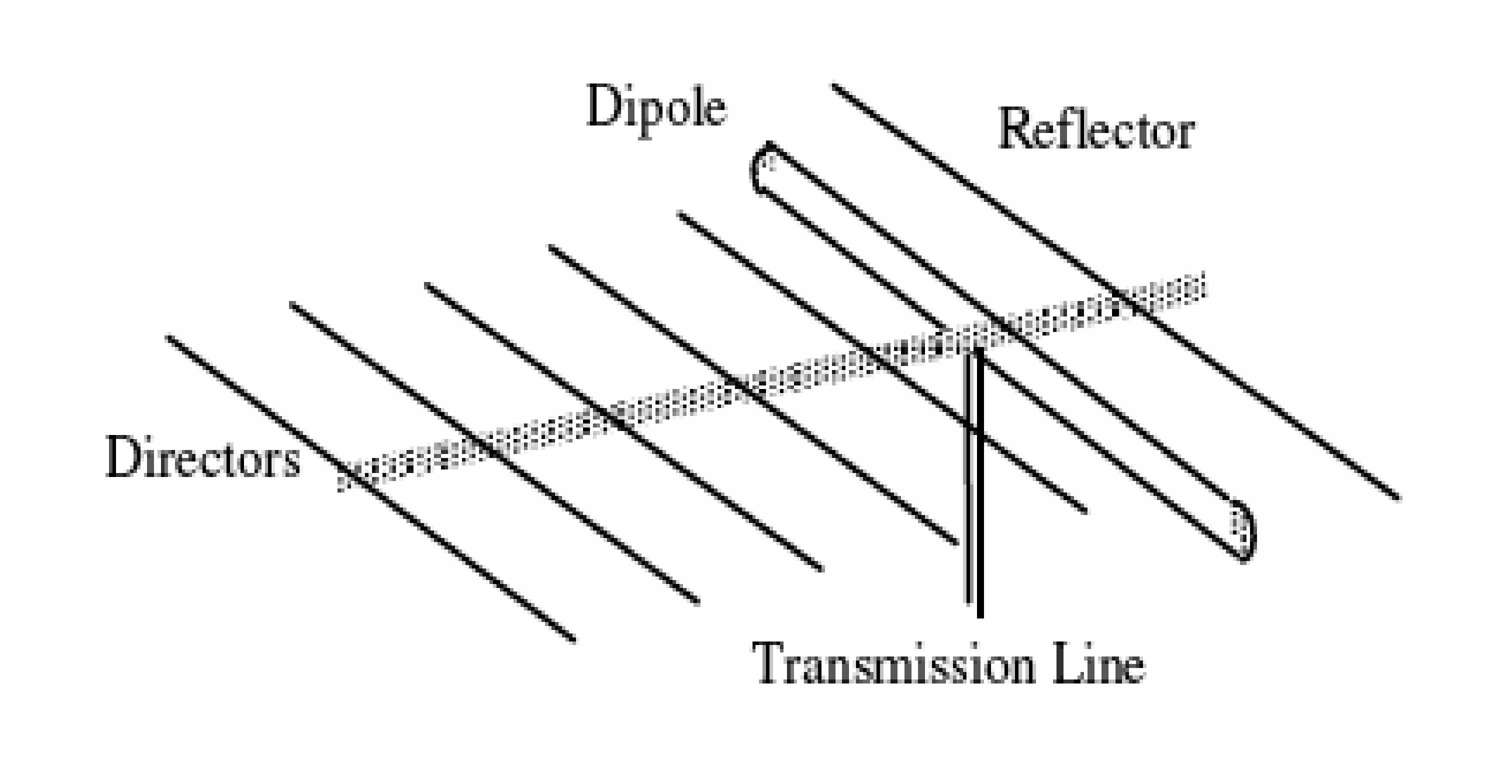
- So the radiation pattern of the collinear array and the broadside array is very much similar but the radiation pattern of the collinear array has circular symmetry with main lobe perpendicular everywhere to the principle axis.
- Thus the collinear array is also called omni directional array or broadcast array.
- The gain of the collinear array is maximum if the spacing between the elements is of the order of 0.3 λ to 0.5 λ.

Parasitic Array

- In order to overcome feeding problems of the antenna, sometimes, the elements of the array are fed through the radiation from the nearby element.
- The array of antennas in which the parasitic elements get the power through electromagnetic coupling with driven element in proximity with the parasitic element is known as parasitic array.
- The simplest form of the parasitic array consists one driven element and one parasitic element.
- In multi-element parasitic array - one or more driving elements and also one or more parasitic elements.
- In general the multi-element parasitic array - at least one driven element and one or more parasitic elements.

- The common example of the parasitic array with linear half wave dipoles as elements of array is Yagi-Uda array or simply Yagi antenna.
- The amplitude and the phase of the current induced in the parasitic element depends on the spacing between the driven element and parasitic element.
- To make the radiation pattern unidirectional, the relative phases of the currents are changed by adjusting the spacing between the elements.
- This is called tuning of array.
- For a spacing between the driven and parasitic element equal to $\lambda/4$ and phase difference of $\pi/2$ radian, unidirectional radiation pattern is obtained.

Parasitic Array / Yagi Antenna



Topic 1

Array of Two Point Sources

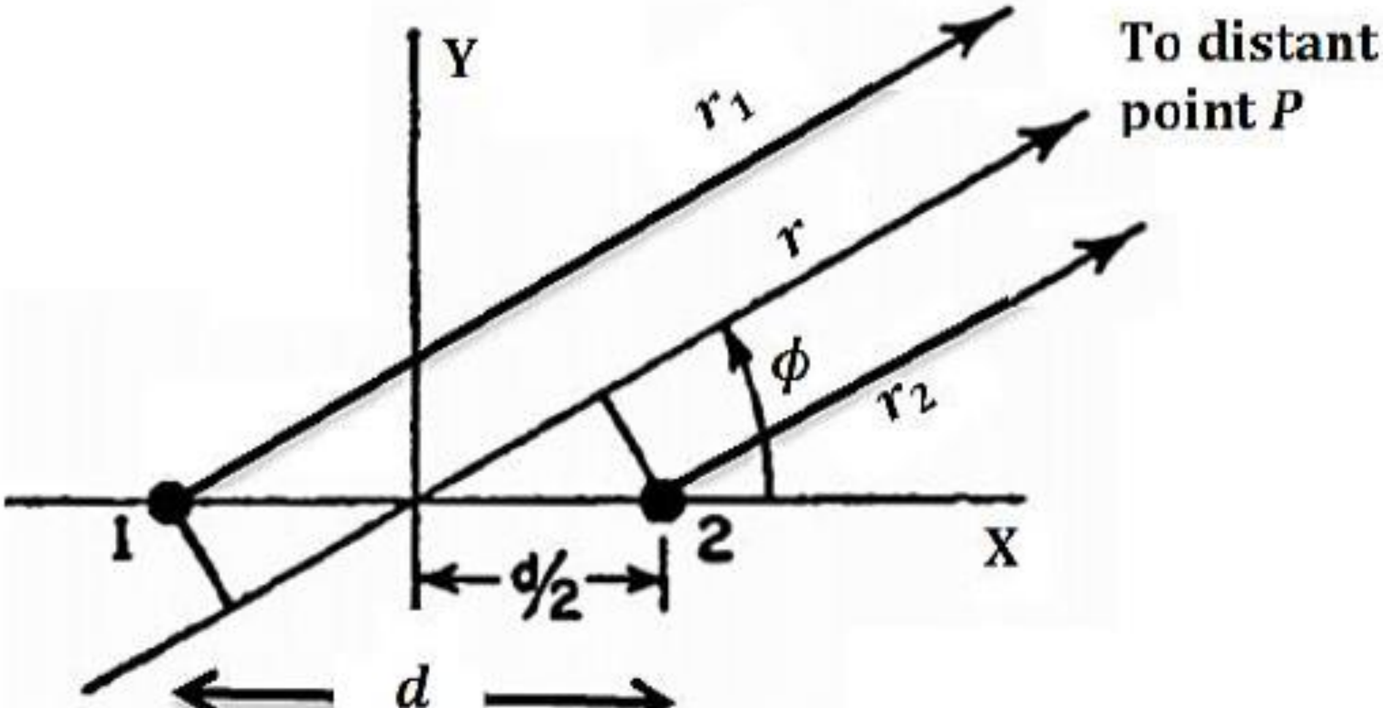
- The array of point sources is nothing but the array of an isotropic radiators occupying zero volume.
- For the greater number of point source in the array, the analysis of antenna array becomes complicated and time consuming.
- Also the simplest condition of number of point sources in the array is two.
- Then conveniently analysis is done by considering first two point sources, which are separated by distance and having same polarization.
- The results obtained for only two point sources can be further extended for ′′ number of point sources in the array.

- Let us consider the array of two isotropic point sources, with a distance of separation ‘d’ between them. The polarization of two isotropic point sources is assumed to be the same. To derive different expressions following conditions can be applied to the antenna array ;
- 1. Two point sources with currents of equal magnitudes and with same phase.
- 2. Two point sources with currents of equal magnitude but with opposite phase.
- 3. Two point sources with currents of unequal magnitudes and with any phase.

Case 1 / Currents with Equal magnitude & Phase

- Consider two point-sources 1 and 2 separated by distance – d and both the point sources are supplied with currents equal in magnitude and phase as shown in Figure 1.
- Let point P far away from the array and the distance between point P and point sources 1 and 2 be r1 and r2 respectively.
- Assuming far-field observations - r1 = r2 = r
- The radiation from the point source 2 will reach earlier at point P than that from point source 1 because of the path difference.

Figure 1 – Two Element Array



- The extra distance is travelled by the radiated wave from point source 1 than that by the wave radiated from point source 2.
- Hence path difference -

$$Path\ difference = \frac{d}{2}\cos\phi + \frac{d}{2}\cos\phi = d\cos\phi \qquad \text{--------} (3.2)$$

The path difference can be expressed in terms of wavelength as ;

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

Hence the phase difference $'\psi'$ is given by ;

$$Phase\ difference\ \ \psi = 2\pi \times Path\ difference$$

$$\psi = 2\pi \times \frac{d}{\lambda}\cos\phi = \frac{2\pi}{\lambda} d\cos\phi$$

$$k = \frac{2\pi}{\lambda}$$

$$\psi = kd\cos\phi \qquad \text{--------} (3.3)$$

Let $E_1 = E_0.e^{-j\frac{\psi}{2}}$ is field component due topoint source 1. Similarly, let $E_2 = E_0.e^{j\frac{\psi}{2}}$ is field component due to point source 2. Therefore, the total far-field at a distant point P is ;

$$E_T = E_1 + E_2 = E_0.e^{-j\frac{\psi}{2}} + E_0.e^{j\frac{\psi}{2}}$$

$$E_T = E_0\left(e^{-j\frac{\psi}{2}} + e^{j\frac{\psi}{2}}\right) = 2E_0 \cos\frac{\psi}{2} \qquad \text{-------- (3.4)}$$

Note that the amplitude of both the field components is E_0 as currents are same and the point sources are identical.

Substituting value of ψ from Eqn. (3.3), we get,

$$E_T = 2E_0 \cos\left[\frac{kd\cos\phi}{2}\right] \qquad \text{-------- (3.5)}$$

- Above equation represents total field intensity at point P, due to two point sources having currents of same amplitude and phase.
- The total amplitude of the field at point P is 2E0 while the phase shift is (kdcos ϕ)/2 .
- By putting 2E0 = 1 , then the pattern is said to be normalized.

Maxima Direction

From Eqn. (3.5), the total field is maximum when $\cos\left[\frac{kd\cos\phi}{2}\right]$ is maximum. Maximum value of cosine function is ±1. Hence the condition for maxima is given by,

$$\cos\left[\frac{kd\cos\phi}{2}\right] = \pm 1 \qquad \text{-------- (3.6)}$$

$$k = \frac{2\pi}{\lambda} \;;\; d = \frac{\lambda}{2}$$

Let spacing between the two point sources be $\lambda/2$, then ;

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

$$i.e, \quad \frac{\pi}{2}\cos\phi_{max} = \cos^{-1}(\pm 1) = \pm n\pi,\ where, n = 0,1,2, \ldots$$

If $n = 0$, then ; $\quad \frac{\pi}{2}\cos\phi_{max} = 0$

$$\cos\phi_{max} = 0 \quad ; \qquad \phi_{max} = 90° \text{ or } 270° \qquad \text{-------- (3.7)}$$

Minima Direction

From Eqn. (3.5), the total field is minimum when $\cos\left[\frac{kd\cos\phi}{2}\right]$ is minimum. Minimum value of cosine function is 0. Hence the condition for maxima is given by,

$$\cos\left[\frac{kd\cos\phi}{2}\right] = 0 \qquad \text{-------- (3.8)}$$

Let spacing between the two point sources be $\lambda/2$, then ;

$$\cos\left[\frac{\pi}{2}\cos\phi\right] = 0$$

$$i.e, \qquad \frac{\pi}{2}\cos\phi_{min} = \cos^{-1}(0) = \pm(2n+1)\frac{\pi}{2}, \;\; where, n = 0,1,2, ...$$

If $n = 0$, then ; $\quad \frac{\pi}{2}\cos\phi_{min} = \pm\frac{\pi}{2}$

$$\cos\phi_{min} = \pm 1 \quad ; \quad \phi_{min} = 0° \text{ or } 180° \qquad \text{-------- (3.9)}$$

Half Power Point Direction

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

$$\cos\left[\frac{kd\cos\phi}{2}\right] = \pm\frac{1}{\sqrt{2}} \qquad \text{-------- (3.10)}$$

Let spacing between the two point sources be $\lambda/2$, then ;

$$\cos\left[\frac{\pi}{2}\cos\phi\right] = \pm\frac{1}{\sqrt{2}}$$

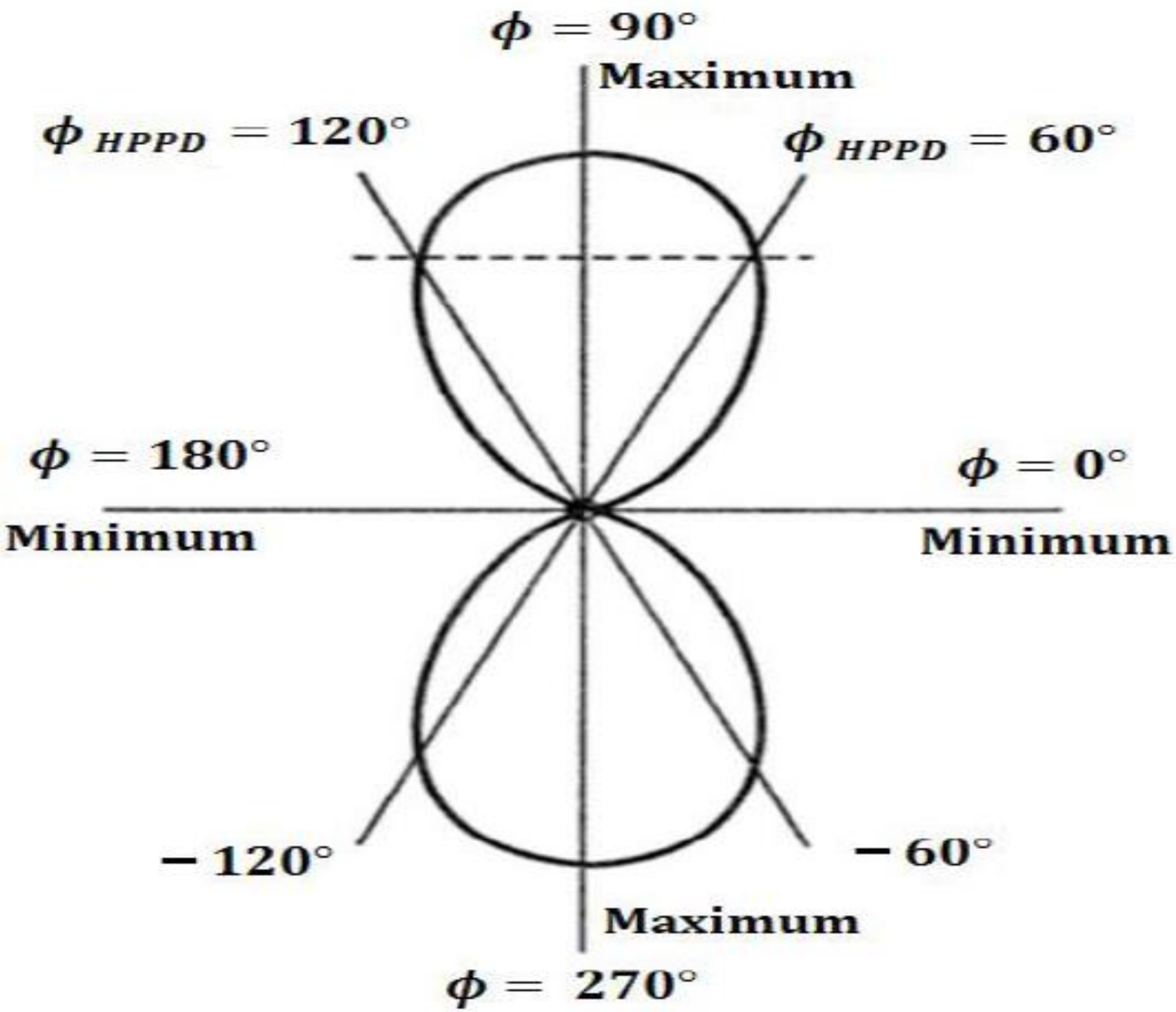
$$i.e, \qquad \frac{\pi}{2}\cos\phi_{HPPD} = \cos^{-1}\left(\pm\frac{1}{\sqrt{2}}\right) = \pm\frac{\pi}{4}, where, n = 0,1,2, \ldots.$$

If $n = 0$, then ; $\quad \frac{\pi}{2}\cos\phi_{HPPD} = \pm\frac{\pi}{4}$

$$\cos\phi_{HPPD} = \pm\frac{1}{2} \quad ; \quad \phi_{HPPD} = \pm 60° \text{ or } \pm 120° \qquad \text{-------- (3.11)}$$

- The field pattern drawn with T against ϕ for $= \lambda/2$, then the pattern is bidirectional as shown in Fig. 2.
- The field pattern obtained is bidirectional and it is a figure of eight (8).
- If this patterns is rotated by 360° about axis, it will represent three dimensional doughnut shaped space pattern

Case 1 / Equal in Amplitude & Phase



Case 2 / Currents Equal in Magnitude & Opposite in Phase

- Consider two point sources separated by distance and supplied with currents equal in magnitude but opposite phase.
- Consider Figure 2, all the conditions are exactly same except the phase of the currents is opposite i.e. 180°.
- With this condition, the total field at far point P is given by,

$$E_T = (-E_1) + E_2 \qquad \text{-------- (3.12)}$$

Assuming equal magnitudes of currents, the fields at point P due to the point sources 1 and 2 can be written as ; $E_1 = E_0 . e^{-j\frac{\psi}{2}}$

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

Therefore, the total far-field at a distant point P is ;

$$E_T = (-E_1) + E_2 = -E_0 . e^{-j\frac{\psi}{2}} + E_0 . e^{j\frac{\psi}{2}}$$

$$E_T = E_0 \left(e^{j\frac{\psi}{2}} - e^{-j\frac{\psi}{2}} \right) = j\,(2E_0)\sin\frac{\psi}{2} \qquad \text{-------- (3.13)}$$

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

$$\psi = kd\cos\phi \qquad \text{-------- (3.14)}$$

Substituting value of ψ in Eqn. (2.93), we get,

$$E_T = j(2E_0)\sin\left[\frac{kd\cos\phi}{2}\right] \qquad \text{-------- (3.15)}$$

By putting $j(2E_0) = 1$, then the pattern is said to be normalized.

Maxima Directions

- From Eqn. (3.15), the total field is maximum when $\sin\left[\frac{kd\cos\phi}{2}\right]$ is maximum. Maximum value of sine function is ±1. Hence the condition for maxima is given by,

$$\sin\left[\frac{kd\cos\phi}{2}\right] = \pm 1 \qquad \text{-------- (3.16)}$$

- Let spacing between the two point sources be $\lambda/2$, then ;

$$k = \frac{2\pi}{\lambda} \; ; \; d = \frac{\lambda}{2}$$

$$\sin\left[\frac{\pi}{2}\cos\phi\right] = \pm 1$$

$$i.e, \quad \frac{\pi}{2}\cos\phi_{max} = \sin^{-1}(\pm 1) = \pm(2n+1)\frac{\pi}{2}, where, n = 0,1,2, \ldots.$$

If $n = 0$, then ; $\frac{\pi}{2}\cos\phi_{min} = \pm\frac{\pi}{2}$

$$\cos\phi_{max} = \pm 1 \quad ; \quad \phi_{max} = 0° \text{ or } 180° \qquad \text{-------- (3.17)}$$

Minima Directions

- From Eqn. (3.15), the total field is minimum when $\sin\left[\frac{kd\cos\phi}{2}\right]$ is minimum. Minimum value of sine function is 0.Hence the condition for maxima is given by,

$$\sin\left[\frac{kd\cos\phi}{2}\right] = 0 \qquad \text{-------- (3.18)}$$

- Let spacing between the two point sources be $\lambda/2$, then ;

$$\sin\left[\frac{\pi}{2}\cos\phi\right] = 0$$

$$i.e, \quad \frac{\pi}{2}\cos\phi_{min} = \sin^{-1}(0) = \pm n\pi, where, n = 0,1,2, ...$$

$$\text{If } n = 0, \text{ then ;} \quad \frac{\pi}{2}\cos\phi_{min} = 0$$

$$\cos\phi_{min} = 0 \quad ; \quad \phi_{min} = 90° \text{ or } 270° \qquad \text{-------- (3.19)}$$

Half Power Point Directions

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

$$\sin\left[\frac{kd\cos\phi}{2}\right] = \pm\frac{1}{\sqrt{2}} \qquad \text{-------- (3.20)}$$

- Let spacing between the two point sources be $\lambda/2$, then ;

$$\sin\left[\frac{\pi}{2}\cos\phi\right] = \pm\frac{1}{\sqrt{2}}$$

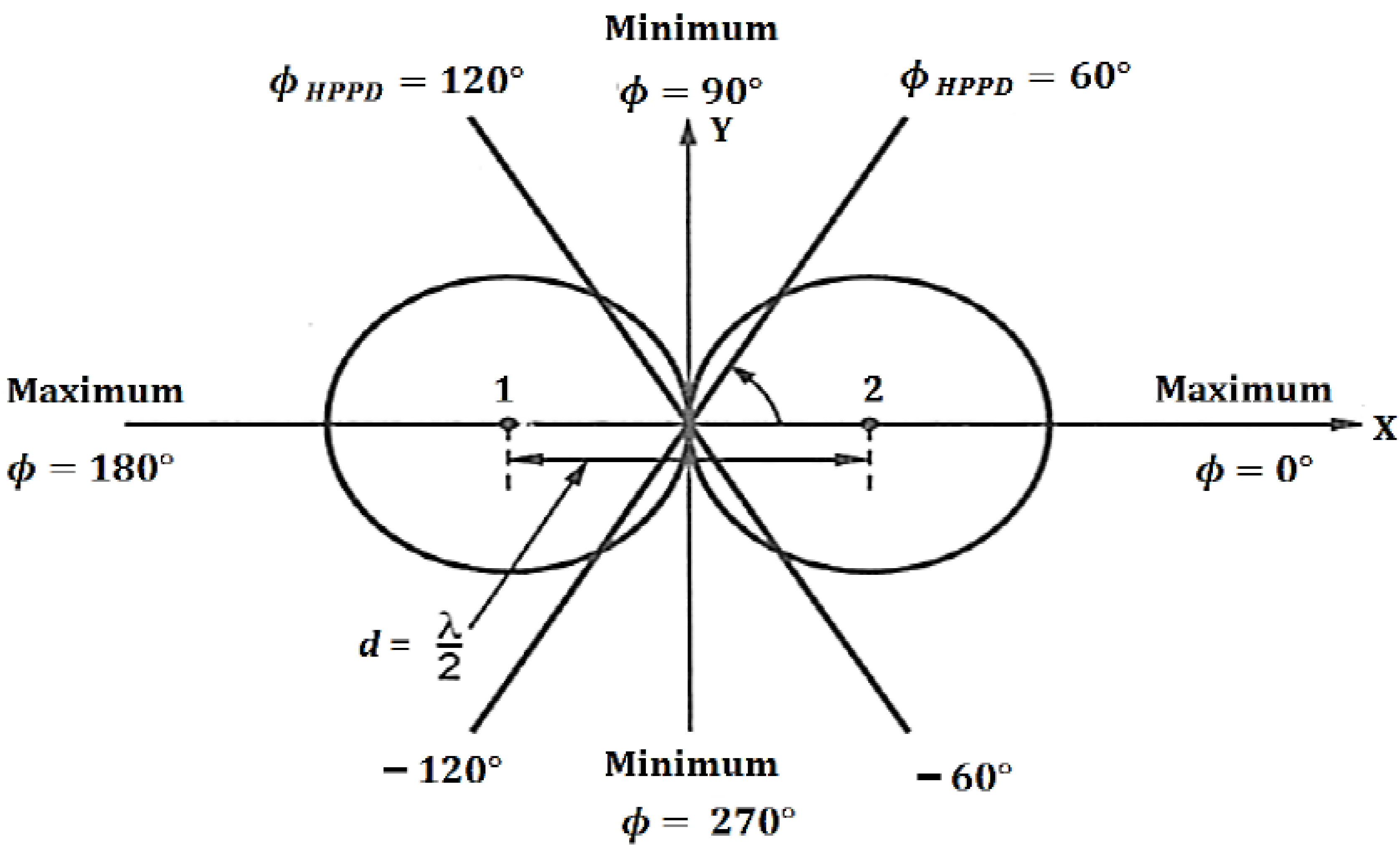
$$i.e, \qquad \frac{\pi}{2}\cos\phi_{HPPD} = \sin^{-1}\left(\pm\frac{1}{\sqrt{2}}\right) = \pm\frac{\pi}{4}, where, n = 0,1,2, \ldots.$$

If $n = 0$, then ; $\frac{\pi}{2}\cos\phi_{HPPD} = \pm\frac{\pi}{4}$

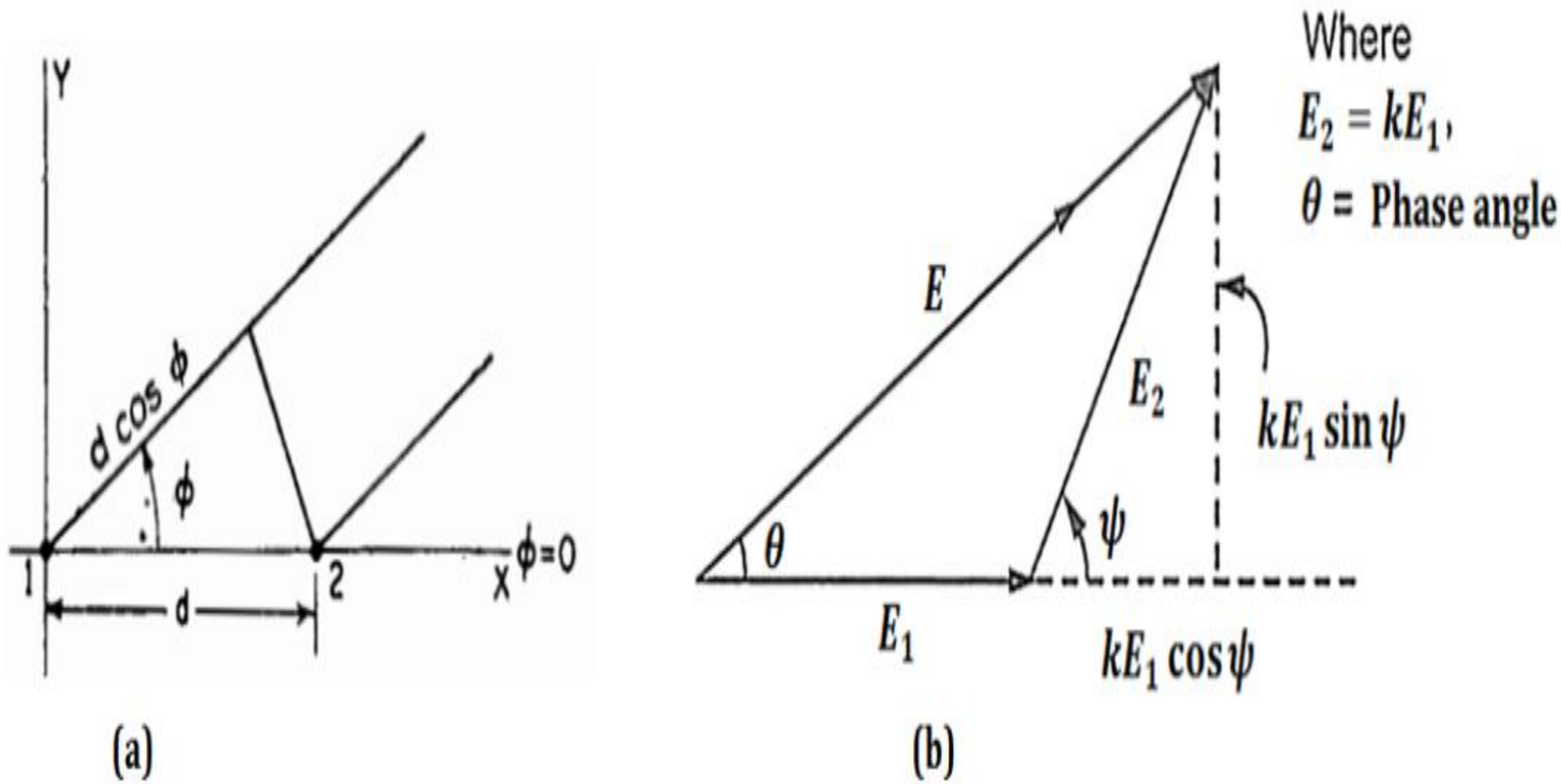
$$\cos\phi_{HPPD} = \pm\frac{1}{2}\ ; \quad \phi_{HPPD} = \pm 60° \text{ or } \pm 120° \quad \text{........} \ (3.21)$$

Thus from the conditions of maxima, minima and half power points, the field pattern can be drawn with E_T against ϕ for $d = \lambda/2$ as shown in Fig. 3-6.

Case 2 / Equal in amplitude & Opposite in Phase



Case 3 / Unequal in Amplitude & Any Phase



Case 3 / Unequal in Amplitude & Any Phase

- Two point sources are separated by distance and supplied with currents which are different in magnitudes and with any phase difference say α, as shown in Figure 3.
- Assume that source 1 is taken as reference for phase. The amplitude of the fields due to source 1 and source 2 at the distant point P is E1 and E2 respectively, in which E1 is greater than E2 , as shown in the vector diagram in Figure 3.

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

$$\psi = kd\cos\phi + \alpha \qquad \text{-------- (3.22)}$$

where α is the phase angle with which current I_2 leads current I_1.

Then the resultant field at point P is given by,

$$E_T = E_1 e^{j0} + E_2 e^{j\psi} \qquad \text{-------- (3.23)}$$

(Source 1 is assumed to be reference hence phase angle is 0)

$$E_T = E_1 + E_2 e^{j\psi} = E_1\left(1 + \frac{E_2}{E_1} e^{j\psi}\right)$$

Let $\frac{E_2}{E_1} = k$

$$E_T = E_1[1 + k(\cos\psi) + j\sin\psi] \qquad \text{-------- (3.24)}$$

Note that $E_1 > E_2$, the value of k is less than unity and varies from $0 \leq k \leq 1$.

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

$$|E_T| = E_1\sqrt{(1+k\cos\psi)^2 + j(k\sin\psi)^2} \quad \text{--------} \; (3.25)$$

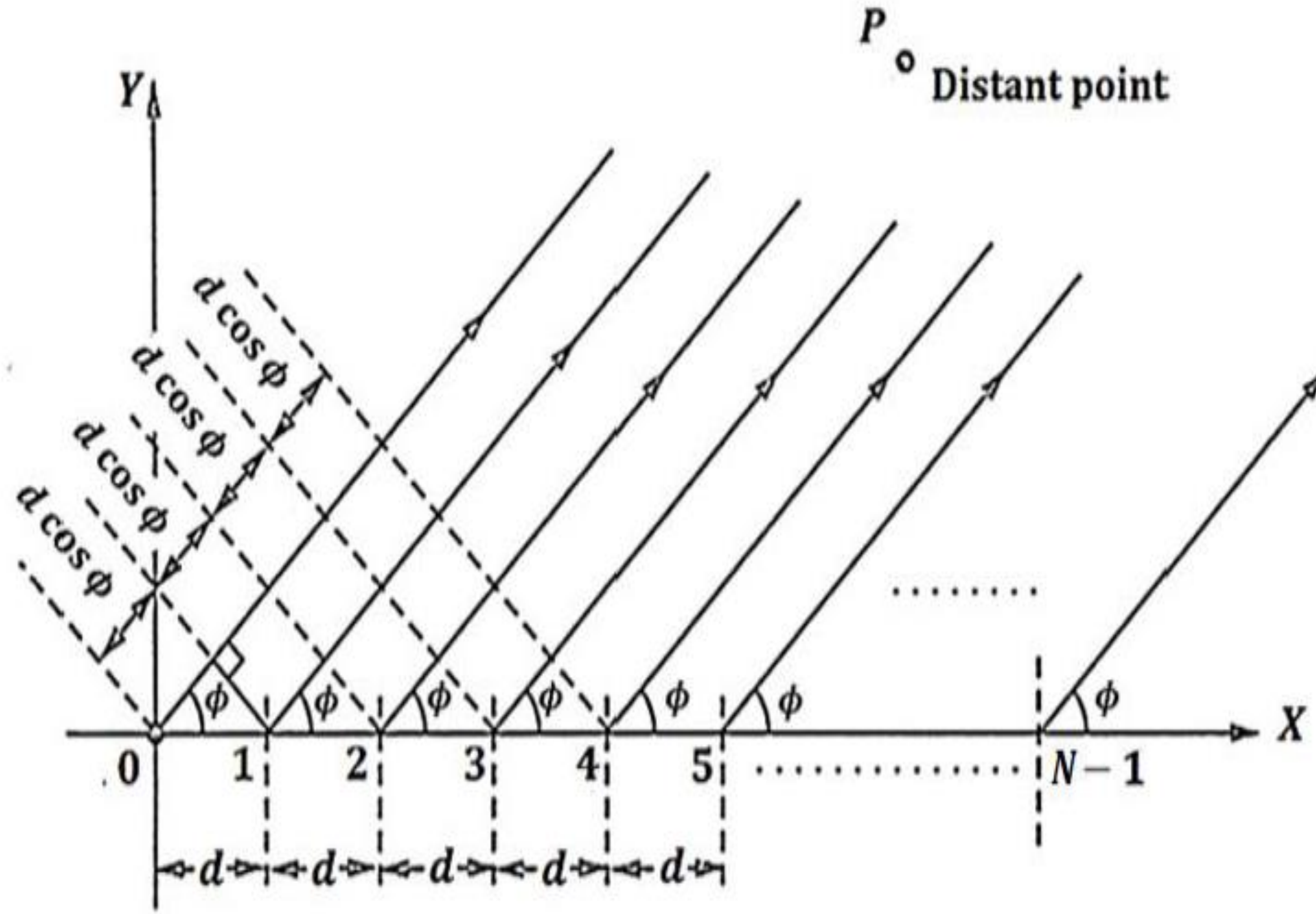
$$\theta = \tan^{-1}\frac{k\sin\psi}{1+k\cos\psi} \quad \text{--------} \; (3.26)$$

Topic 2

N- Element Uniform Linear Antenna Array

- An array of elements is said to be linear array if all the individual elements are spaced equally along a line.
- An array is said to be uniform array if the elements in the array are fed with currents with equal magnitudes and uniform progressive phase shift along the line.
- Consider uniform linear array of $\boldsymbol{N}$ isotropic point sources with all the individual elements spaced equally at distance from each other and all elements are fed with currents equal in magnitude and uniform progressive phase shift along line as shown in Fig. 1.

N - Element Uniform Array



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

$$E_T = E_0 e^{j0} + E_0 e^{j\psi} + E_0 e^{2j\psi} + \dots\dots\dots + E_0 e^{(N-1)j\psi}$$

$$E_T = E_0\left(1 + e^{j\psi} + e^{2j\psi} + \dots\dots\dots + e^{(N-1)j\psi}\right) \qquad \text{-------} (3.27)$$

Note that $\psi = kd\cos\phi + \alpha$ indicates the total phase difference of the fields from adjacent sources calculated at point P. Simlarly α is the progressive phase shift betweentwo adjacentpoint sources. The value of α may lie between $0°$and $180°$. If $\alpha = 0°$ we get N element uniform linear broadside array. If $\alpha = 180°$, we get N element uniformlinear end fire array.

Multiply by $e^{j\psi}$ on both sides ;

$$E_T e^{j\psi} = E_0\left(e^{j\psi} + e^{2j\psi} + e^{3j\psi} + \dots\dots\dots + e^{Nj\psi}\right) \qquad \text{-------} (3.28)$$

Subtract Eqn. (3.27) and (3.28) ;

$$E_T(1 - e^{j\psi}) = E_0(1 - e^{Nj\psi})$$

$$\frac{E_T}{E_0} = \frac{(1 - e^{Nj\psi})}{(1 - e^{j\psi})} = e^{j[(N-1)/2]\psi}\left[\frac{e^{j(N/2)\psi} - e^{-j(N/2)\psi}}{e^{j(1/2)\psi} - e^{-j(1/2)\psi}}\right] \quad \text{-------- (3.29)}$$

If the reference point is the physical center of the array, then Eqn. (2.94)reduces to ;

$$\frac{E_T}{E_0} = AF = \left[\frac{\sin\left(\frac{N}{2}\psi\right)}{\sin\left(\frac{1}{2}\psi\right)}\right]$$

which is the ***antenna array factor.***

If $\psi = 0$, the maximum value of E_T is determined using L'Hospital's rule ;

$$E_{T\,max} = E_0\,N$$

Thus the maximum value of E_T is N times the field from a single source. To normalize the field pattern , so that the maximum value of each is equal to unity. The normalized field pattern is given by ;

$$(E_T)_N = \frac{E_T}{E_{T\,max}} = \left[\frac{\sin\left(\frac{N}{2}\psi\right)}{N\,\sin\left(\frac{1}{2}\psi\right)} \right] \quad \text{--------}\ (3.30)$$

Eqn. (3.30) is the normalized array factor ;

$$(AF)_N = \left[\frac{\sin\left(\frac{N}{2}\psi\right)}{N\sin\left(\frac{1}{2}\psi\right)} \right] \quad \text{--------}\ (3.31)$$

Topic 3

Broadside Array (BSA)

- An array is said to be broadside array, if maximum radiation occurs in direction perpendicular to array axis.
- In broadside array, individual elements are equally spaced along a line and each element is fed with current of equal magnitude and same phase.
- The total phase difference of the fields at point P from adjacent sources is given by -

$$\psi = kd\cos\phi + \alpha$$

The normalized array factor for $'N'$ elements ;

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

Major lobe

In broadside array sources should be in phase i.e., $\alpha = 0°$ and $\psi = 0$ for maximum must be satisfied.

$$\psi = kd \cos\phi + \alpha = 0$$

$$kd \cos\phi + \alpha = 0 \qquad \because \alpha = 0$$

$$\cos\phi = 0 \quad ; \qquad \phi_m = 90° \text{ or } 270°$$

Nulls

To find the nulls of the array Eqn. (3.33) is set to zero ;

$$k = \frac{2\pi}{\lambda}$$

$$\sin\left(\frac{N}{2}\psi\right) = 0 \quad \Rightarrow \quad \left.\frac{N}{2}\psi\right|_{\phi=\phi_n} = \pm n\pi$$

For BSA $\alpha = 0°$

$$\frac{N}{2}(kd \cos\phi_n + \alpha) = \pm n\pi \quad \Rightarrow \quad \phi_n = \cos^{-1}\left(\pm\frac{n\lambda}{Nd}\right) \quad \text{-------- (3.34)}$$

where ; $n = 1, 2, 3 \ldots$

Maxima of minor lobes (secondary maxima)

The maximum value of Eqn. (3.33) occur when ;

$$\sin\left(\frac{N}{2}\psi\right) = 1 \quad \Rightarrow \quad \frac{N}{2}\psi\bigg|_{\phi=\phi_s} = \pm(2s+1)\frac{\pi}{2}$$

$$k = \frac{2\pi}{\lambda}$$

$$\frac{N}{2}(kd\cos\phi_s + \alpha) = \pm(2s+1)\frac{\pi}{2}$$

For BSA $\alpha = 0°$

$$\phi_s = \cos^{-1}\left\{\frac{1}{kd}\left[\pm\frac{(2s+1)\pi}{N} - \alpha\right]\right\}$$

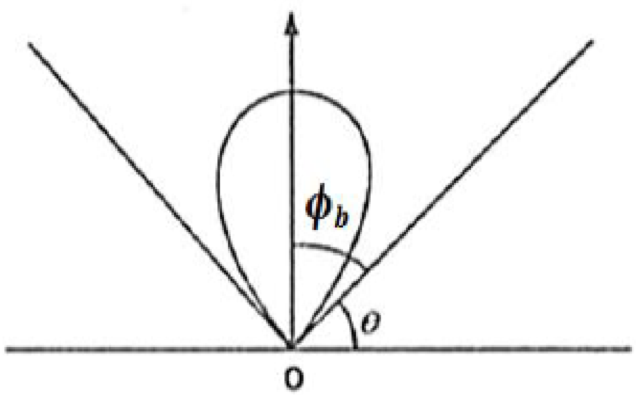
$$\phi_s = \cos^{-1}\left\{\frac{1}{kd}\left[\pm\frac{(2s+1)\pi}{N}\right]\right\} \qquad s = 1,2,3,\ldots$$

$$\phi_s = \cos^{-1}\left[\pm\frac{(2s+1)\lambda}{2Nd}\right] \qquad \text{-------- (3.35)}$$

Beamwidth of major lobe

Beamwidth is defined as angle between first null and maximum of major lobe (or) Beamwidth is the angle equal to twice the angle between first null and the major lobe maximum. $BWFN = 2 \times \phi_b = 2 \times (90 - \phi_n)$

$$90 - \phi_n = \phi_b \quad \Rightarrow \quad 90 - \phi_b = \phi_n \qquad \text{-------- (3.36)}$$



$$90 - \phi_b = \cos^{-1}\left(\pm\frac{n\lambda}{Nd}\right)$$

Take cosine on both sides ;

$$\cos(90 - \phi_b) = \cos\left(\cos^{-1}\left(\pm\frac{n\lambda}{Nd}\right)\right)$$

$$\sin\phi_b = \pm\frac{n\lambda}{Nd}$$

For first null $n = 1$

$$\sin\phi_b = +\frac{\lambda}{Nd}$$

Nd indicates the total length of the array L

$$BWFN = 2 \times \phi_b = +\frac{2\lambda}{Nd} \quad \text{-------- (3.37)}$$

$$BWFN = \frac{2\lambda}{L} = \frac{2}{(L/\lambda)} \quad rad$$

$$BWFN = \frac{114.6^\circ}{(L/\lambda)} \quad deg \quad \text{-------- (3.38)}$$

Half power beamwidth (HPBW)

$$HPBW = \frac{BWFN}{2} = \frac{1}{(L/\lambda)} \quad rad$$

$$HPBW = \frac{57.3^\circ}{(L/\lambda)} \quad deg \quad \text{-------- (3.39)}$$

Directivity

Directivity can be expressed in terms of the total length of the array ;

$$D_{max} = 2(L/\lambda) \quad \text{-------- (3.40)}$$

Topic 4

End-fire Array (EFA)

- An array is said to be end-fire array, if the direction of maximum radiation coincides with the array axis.
- In end-fire array, individual elements are equally spaced along a line and each element is fed with current of equal magnitude and opposite phase.
- The total phase difference of the fields at point P from adjacent sources is given by,
- $\psi = kd \cos \phi + \alpha$

The normalized array factor for $'N'$ elements ;

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

Major lobe

In end-fire array $\psi = 0$ and $\phi = 0°$ or $180°$

$$\psi = kd\cos\phi + \alpha = 0$$

$\psi = 0$ and $\phi = 0°$ ⇨ $\alpha = -kd$

$\psi = 0$ and $\phi = 180°$ ⇨ $\alpha = kd$

$$\phi_m = 0° \text{ or } 180°$$

Nulls

To find the nulls of the array Eqn. (3.42) is set to zero ;

$$k = \frac{2\pi}{\lambda}$$

$$\sin\left(\frac{N}{2}\psi\right) = 0 \quad \Rightarrow \quad \frac{N}{2}\psi\Big|_{\phi=\phi_n} = \pm n\pi$$

For EFA $\alpha = -kd$

$$\frac{N}{2}(kd\cos\phi_n + \alpha) = \pm n\pi$$

$$\frac{N}{2}(kd\cos\phi_n - kd) = \pm n\pi$$

where ; $n = 1, 2, 3 \ldots$

$$\frac{Nd}{\lambda}(\cos\phi_n - 1) = \pm n \qquad \text{-------- (3.45)}$$

$$2\sin^2\frac{\phi_n}{2} = \pm\frac{n\lambda}{Nd}$$

$$\phi_n = 2\sin^{-1}\left(\pm\sqrt{\frac{n\lambda}{2Nd}}\right) \qquad \text{-------- (3.46)}$$

Maxima of minor lobes (secondary maxima)

The maximum value of Eqn. (3.42) occur when ;

$$\sin\left(\frac{N}{2}\psi\right) = 1 \quad \Rightarrow \quad \left.\frac{N}{2}\psi\right|_{\phi=\phi_s} = \pm(2s+1)\frac{\pi}{2}$$

$$k = \frac{2\pi}{\lambda}$$

$$\frac{N}{2}(kd\cos\phi_s + \alpha) = \pm(2s+1)\frac{\pi}{2}$$

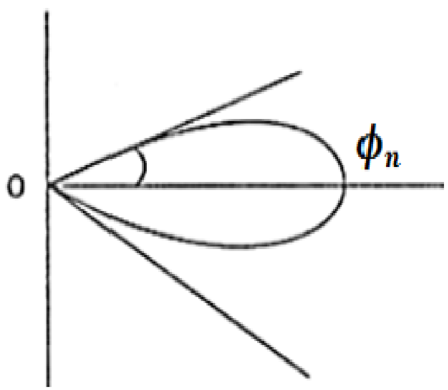
For EFA $\alpha = -kd$

$$(\cos\phi_s - 1) = \pm(2s+1)\frac{\lambda}{2Nd} \quad s = 1,2,3,\ldots \qquad \text{-------- (3.48)}$$

$$\phi_s = \cos^{-1}\left[\pm(2s+1)\frac{\lambda}{2Nd} + 1\right] \qquad \text{-------- (3.49)}$$

Beamwidth of major lobe

Beamwidth is defined as angle between first null and maximum of major lobe (or) Beamwidth is the angle equal to twice the angle between first null and the major lobe maximum. $BWFN = 2 \times \phi_n$



$$\phi_n = 2\sin^{-1}\left(\pm\sqrt{\frac{n\lambda}{2Nd}}\right)$$

$$\sin\frac{\phi_n}{2} = \pm\sqrt{\frac{n\lambda}{2Nd}}$$

For small angles ; $\sin\phi_n \approx \phi_n$

$$\frac{\phi_n}{2} \approx \pm\sqrt{\frac{n\lambda}{2Nd}}$$

$$\phi_n = \pm\sqrt{\frac{2n\lambda}{Nd}} \quad \text{-------- (3.51)}$$

Nd indicates the total length of the array L

$$\phi_n = \pm\sqrt{\frac{2n\lambda}{Nd}}$$

For first null $n = 1$

$$\phi_n = \pm\sqrt{\frac{2\lambda}{L}}$$

$$BWFN = 2 \times \phi_n = \pm 2\sqrt{\frac{2\lambda}{L}} \quad \text{-------- (3.52)}$$

$$BWFN = \pm 2\sqrt{\frac{2}{(L/\lambda)}} \quad rad$$

$$BWFN = 114.6^\circ\sqrt{\frac{2}{(L/\lambda)}} \quad deg \quad \text{-------- (3.53)}$$

Half power beamwidth (HPBW)

$$HPBW = \frac{BWFN}{2} = \pm\sqrt{\frac{2}{(L/\lambda)}} \quad rad$$

$$HPBW = 57.3^\circ \sqrt{\frac{2}{(L/\lambda)}} \quad deg \qquad \text{-------- (3.54)}$$

Directivity

Directivity can be expressed in terms of the total length of the array ;

$$D_{max} = 4(L/\lambda) \qquad \text{-------- (3.55)}$$

Topic 5

Principle of Pattern Multiplication

- The field pattern of an array of non-isotropic but similar sources is the product of the pattern of the individual sources and the pattern of isotropic point sources having the same locations, relative amplitudes, and phase as the non-isotropic point sources.
- This is referred to as pattern multiplication for arrays of identical elements.

$$\text{Total field } (E) = \underbrace{\{E_i(\theta,\phi) \times E_a(\theta,\phi)\}}_{\substack{\textit{Multiplication of} \\ \textit{Field pattern}}} \times \underbrace{\{E_{pi}(\theta,\phi) + E_{pa}(\theta,\phi)\}}_{\substack{\textit{Addition of Phase} \\ \textit{Pattern}}} \qquad \text{-------- (3.57)}$$

where ;

$E_i(\theta,\phi)$ = Field pattern of individual source

$E_a(\theta,\phi)$ = Field pattern of array of isotropic source

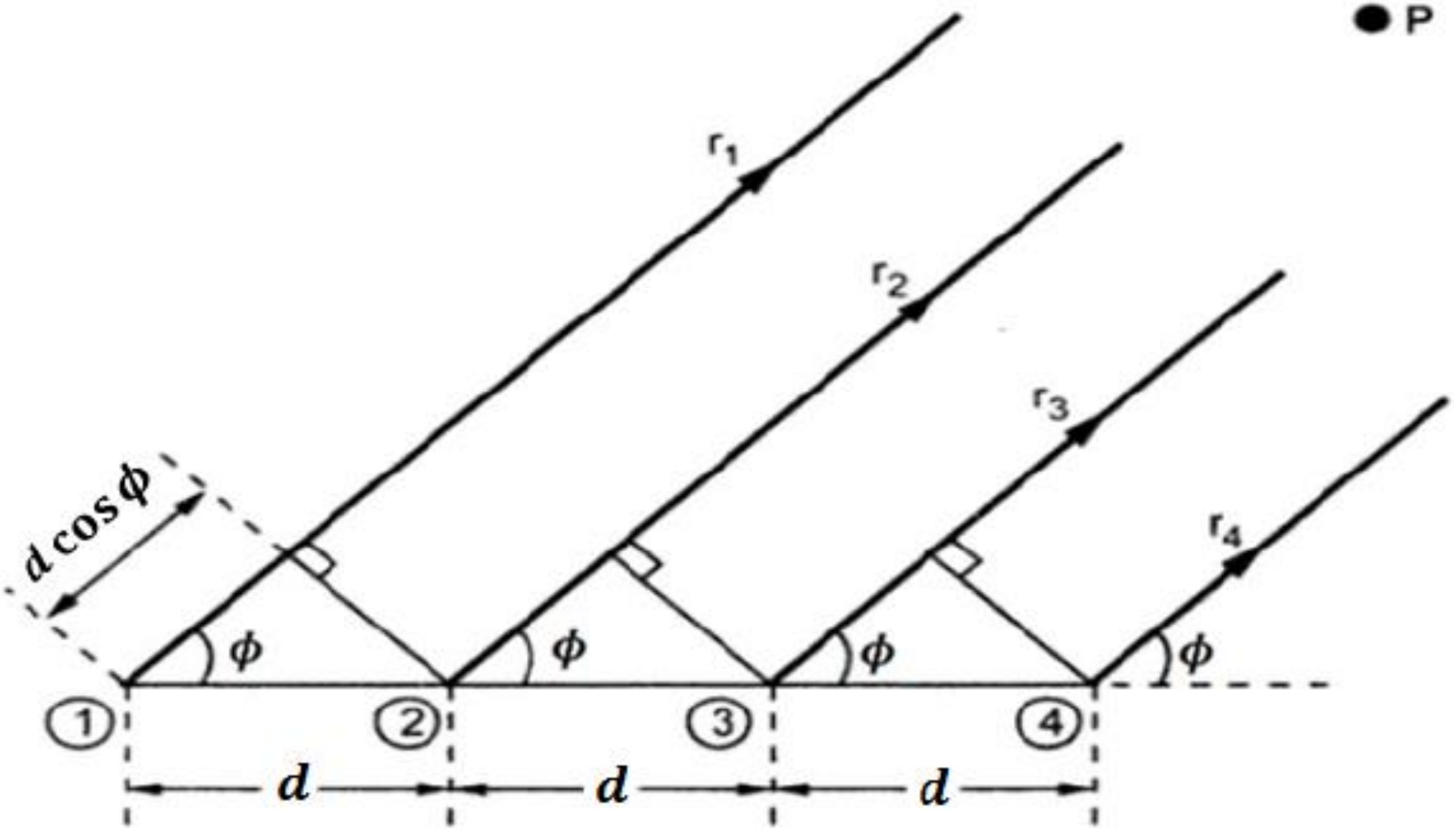
$E_{pi}(\theta,\phi)$ = Phase pattern of individual source

$E_{pa}(\theta,\phi)$ = Phase pattern of array of isotropic source

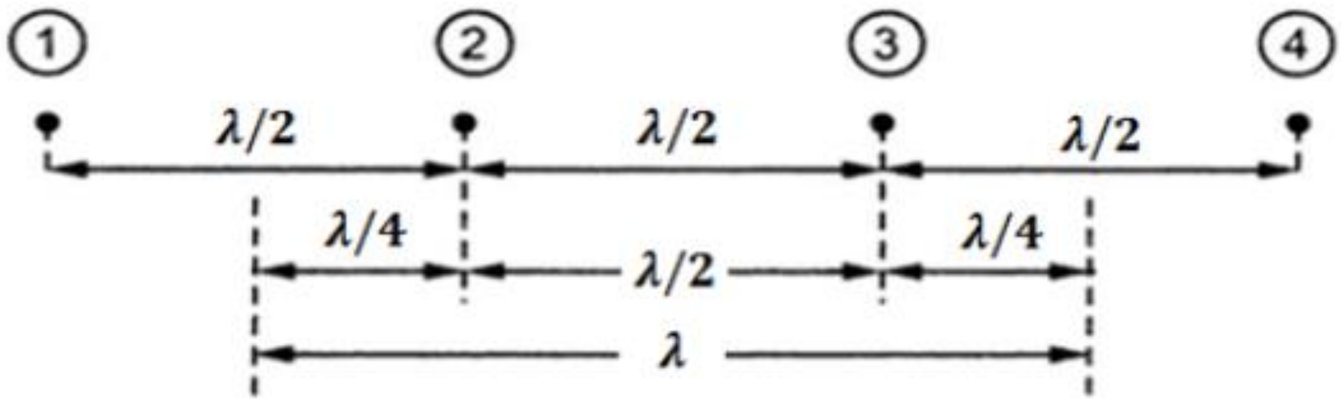
RADIATION PATTERN OF 4-ISOTROPIC ELEMENTS FED IN PHASE & SPACED $\boldsymbol{\lambda/2}$ APART

- Consider a 4-element array of antennas as shown in Fig. 1, in which the spacing between the elements is $\boldsymbol{\lambda/2}$ and the currents are in-phase (α = 0).
- The pattern can be obtained directly by adding the four electric fields due to the 4 antennas.
- However the same radiation pattern can be obtained by pattern multiplication in the following manner

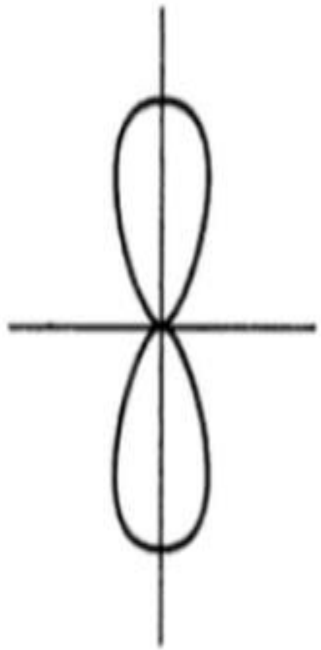
Fig 1 / Linear Array of 4 Isotropic elements



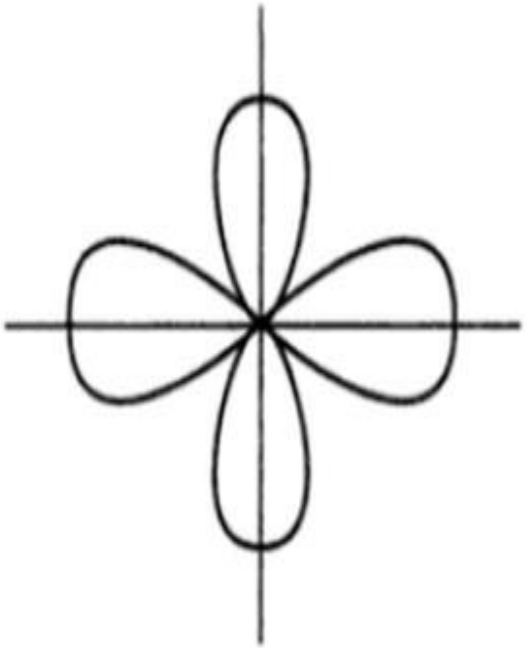
- Two isotropic point sources spaced $\lambda/2$ apart fed in-phase provides a bidirectional pattern as in Fig. 2 (b).
- Now the elements 1 and 2 are considered as one unit and this new unit is considered to be placed between the midway of elements 1, 2 and similarly the elements 3,4 as shown in Fig. 2 (a).



(a) Antenna ① and ② and ③ and ④ replaced by single antenna separately



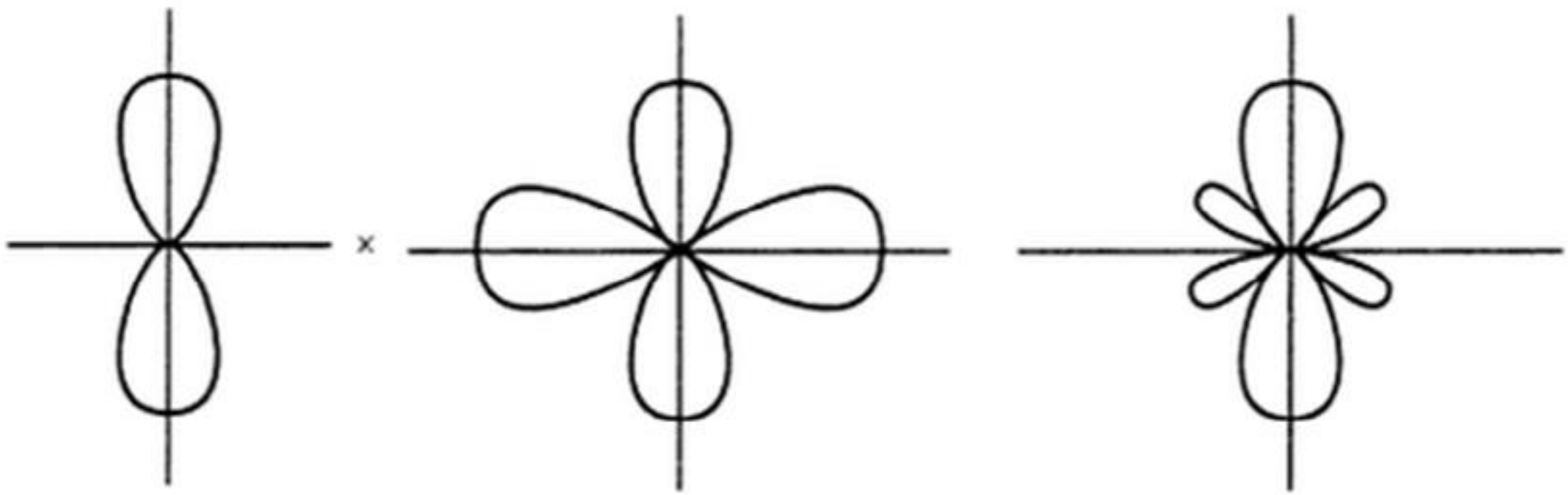
(b) Radiation pattern of 2 antennas spaced at distance $\lambda/2$ and fed with equal currents in phase



(c) Radiation pattern of 2 antennas spaced at distance λ and fed with equal currents in phase

4 elements spaced $\lambda/2$ have been replaced by 2 units spaced λ and therefore the problem of determining radiation of 4 elements has been reduced to find out the radiation pattern of 2 antennas spaced λ apart as in Fig. 3-9 (a).

$$\left\{\begin{matrix} \textit{Resultant radiation} \\ \textit{pattern of 4 elements} \end{matrix}\right\} = \left\{\begin{matrix} \textit{Resultant radiation} \\ \textit{pattern of individual elements} \end{matrix}\right\} \times \left\{\begin{matrix} \textit{Array of 2 units} \\ \textit{spaced } \lambda \end{matrix}\right\}$$



Individual (unit pattern) pattern due to 2 individual elements

Group pattern due to array of two isotropic separated by λ

Resultant pattern of 4 isotropic elements

- Here the width of the principal lobe is the same as the width of the corresponding lobe of the group pattern.
- The number of secondary lobes can be determined from the nulls in the resultant pattern, which is sum of the nulls in the unit and group pattern

Topic 6

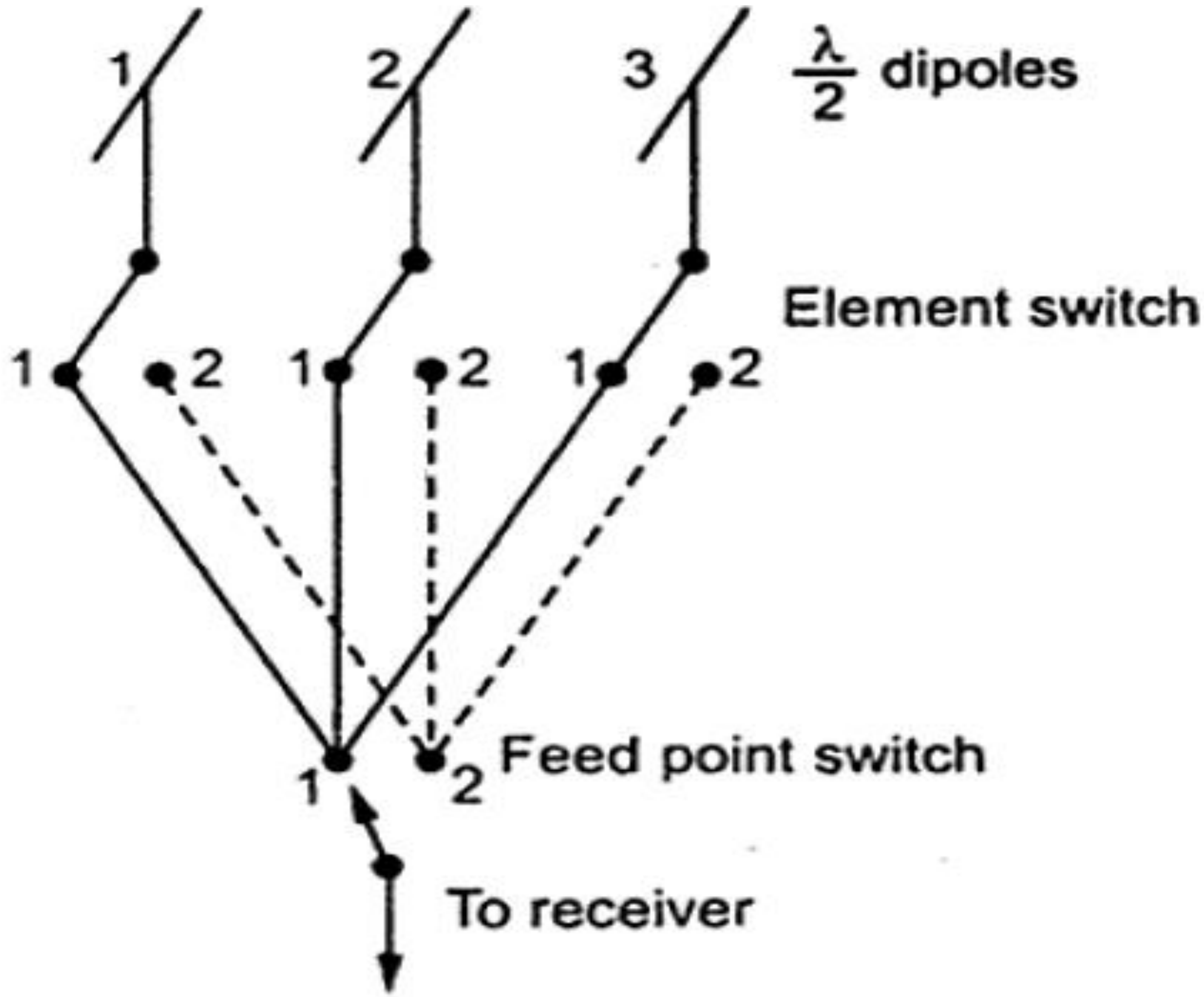
Concept of Phased Array / Adaptive Array

- In case of the broadside array and the end fire array, the maximum radiation obtained by adjusting the phase excitation between elements in the direction normal and along the axis of array respectively.
- In other words elements of antenna array can be phased in particular way.
- So we can obtain an array which gives maximum radiation in any direction by controlling phase excitation in each element.
- Such an array is commonly called phased array.
- The array in which the phase and the amplitude of most of the elements is variable, provided that the direction of maximum radiation and pattern shape along with the side lobes controlled, is called as phased array.

- Suppose the array gives maximum radiation in direction $\phi = \phi 0$ where $0 \leq \phi 0 \leq 180°$, then the phase shift that must be controlled can be obtained as follows.
  - $\psi = kd \cos \phi + \alpha|_{\phi=\phi 0} = 0$
- Thus from above Equation, that the maximum radiation can be achieved in any direction if the progressive phase difference between the elements is controlled.
- The electronic phased array operates on the same principle.
- Consider a three element array as shown in the Fig. 1.

- The element of array is considered as $\lambda/2$ dipole. All the cables used are of same length.
- All the three cables are brought together at common feed point. Here mechanical switches are used.
- Such switch is installed one at each antenna and one at a common feed point.
- All the switches are ganged together. Thus by operating switch, the beam can be shifted to any phase shift.
- To make operation reliable and simple, the ganged mechanical switch is replaced by PIN dipole which acts as electronic switch.
- But for precision in results, the number of cables should be minimized.

Fig 1 / Phased Array with mechanical Switches



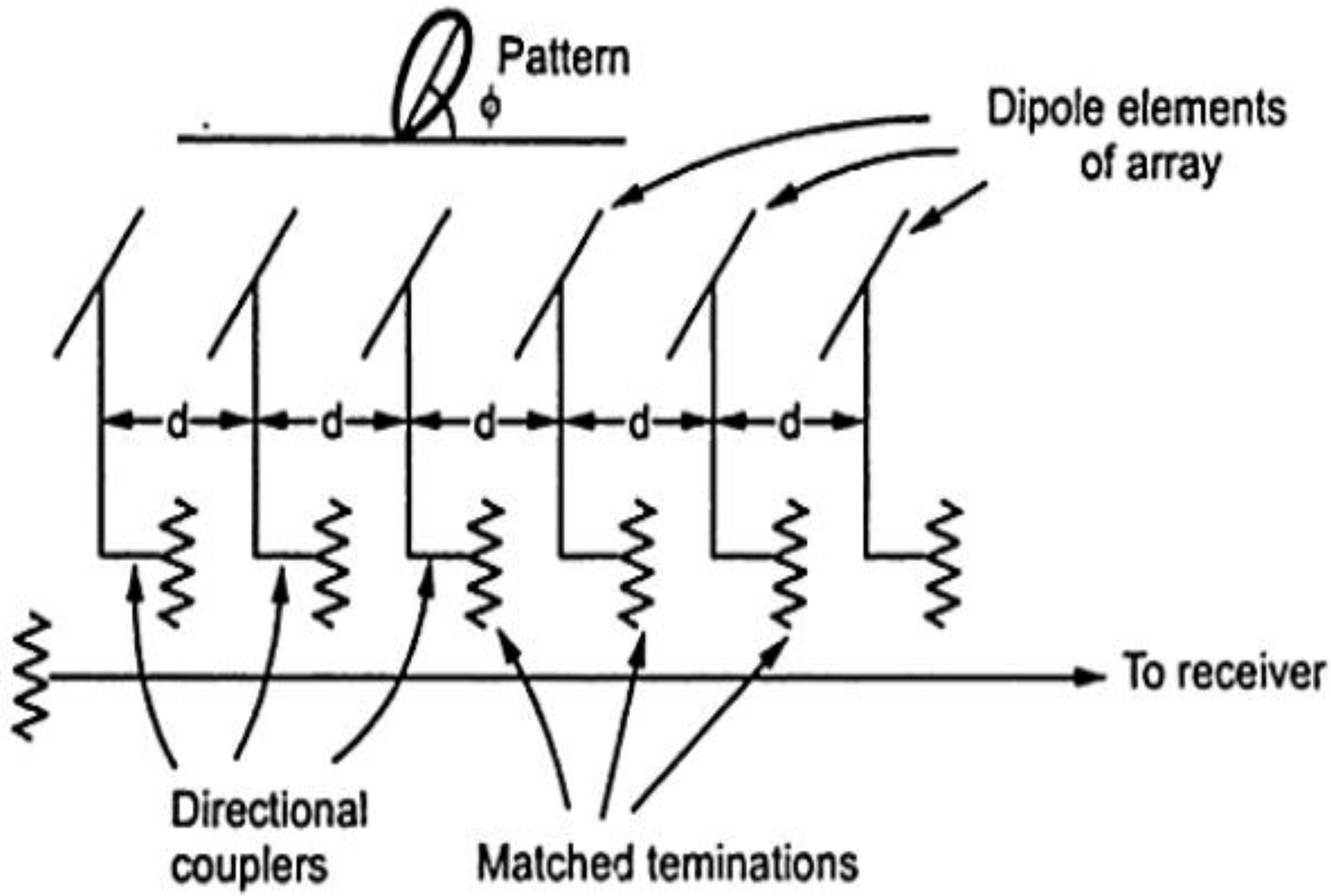
Types of Phased Arrays

- In many applications phase shifter is used instead of controlling phase by switching cables.
- It can be achieved by using ferrite device. The conducting wires are wrapped around the phase shifter.
- The current flowing through these wires controls the magnetic field within ferrite and then the magnetic field in the ferrite controls the phase shift.
- The phased array for specialized functional utility are recognized by different names such as frequency scanning array, retroarray and adoptive array.

Frequency Scanning Array

- The array in which the phase change controlled by varying the frequency is called frequency scanning array.
- This is found to be the simplest phased array as at each element separate phase control is not necessary.
- A simple transmission line fed frequency scanning array as shown in the Figure 2.
- Each element of the scanning array is fed by a transmission line via directional coupler.
- The directional couplers are fixed in position, while beam scanning is done with a frequency change.

Fig 2 / Frequency Scanning Array



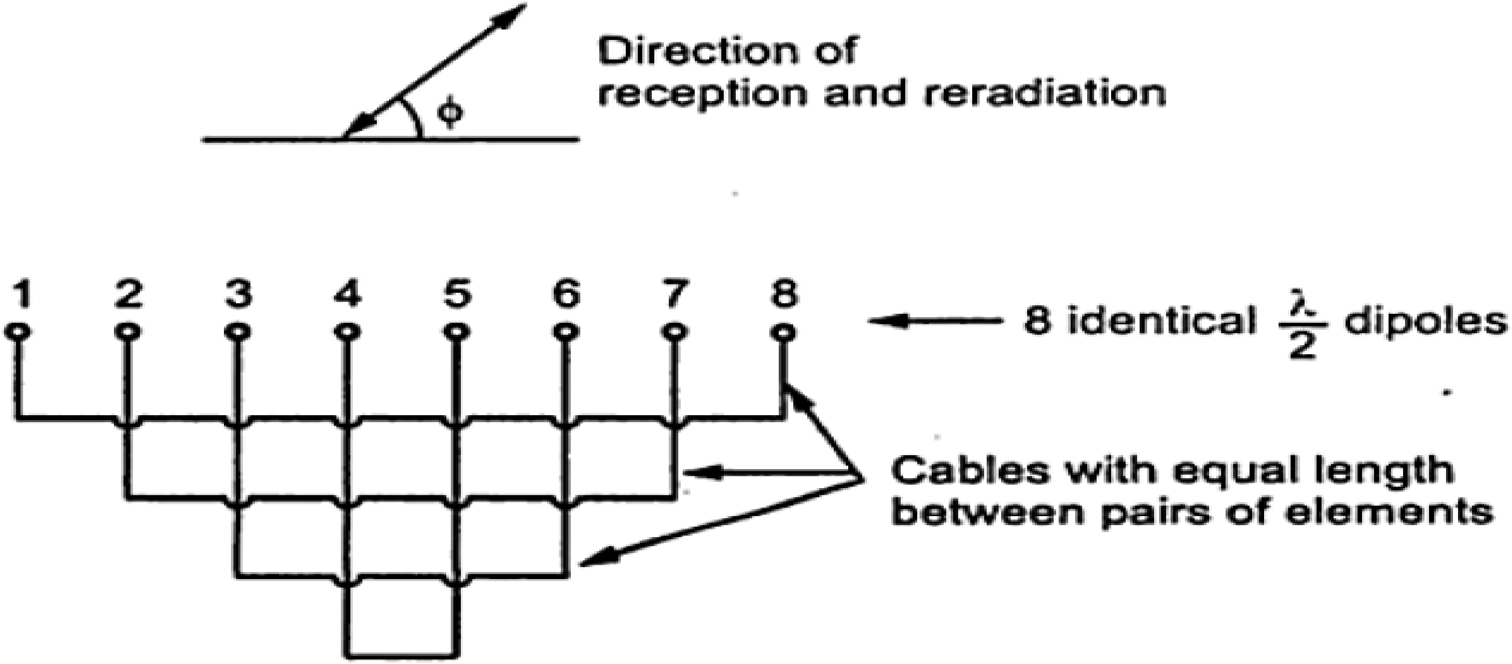
Advantages of Frequency Scanning Array

- The transmission line is properly terminated of the load to avoid reflections.
- There are no moving parts and no switches and phase shifters are required.

Retro-array

- The array which automatically reflects an incoming signal back to the source is called retro-array.
- It acts as a retro-reflector similar to the passive square corner reflector.
- That means the wave incident on the array is received and transmitted back in the same direction.
- In other words, each element of the retro-array reradiates signal which is actually the conjugate of the received one.

- Simplest form of the retroarray is the Van Atta array as shown in the Fig. 3 in which 8 identical dipole elements are used, with pairs formed between elements l and 8, 2 and 7, 3 and 6, 4 and 5 using cables of equal length.
- If the wave arrives at angle say ϕ , then it gets transmitted in the same direction.



Adaptive Arrays

- An array which automatically turn the maximum beam in the desired direction while turn the null in the undesired direction is called adoptive array.
- The adoptive array adjust itself in the desired direction with awareness of its environment.
- In modem adoptive arrays, the output of each element in the array is sampled, digitized and then processed using computers.
- Such arrays are commonly called smart antennas.

Topic 7

Principle of Antenna Synthesis / Binomial Array

- In case of uniform linear array, to increase the directivity, the array length has to be increased.
- But when the array length increases, the secondary or side lobes appear in the pattern.
- In some of the special applications, it is desired to have single main lobe with no minor lobes.
- That means the minor lobes should be eliminated completely or reduced to minimum level as compared to main lobe.
- To achieve such pattern, the array is arranged in such away that the broad side array radiate more strongly at the centre than that from edges.

Binomial Series

- To reduce the side lobe level, John Stone proposed that sources have amplitudes proportional to the coefficients of a binomial series of the form -

$$(1+x)^{m-1} = 1 + (m-1)x + \frac{(m-1)(m-2)}{2!}x^2 + \frac{(m-1)(m-2)(m-3)}{3!}x^3 + \dots$$

- where is ‘m’ the number of radiating sources in the array.
- Binomial array is an array whose elements are excited according to the current levels determined by the binomial coefficient.

The positive cooefficients of the series expansion for different values of m are ;

$m=1$										1										
$m=2$									1		1									
$m=3$								1		2		1								
$m=4$							1		3		3		1							
$m=5$						1		4		6		4		1						
$m=6$					1		5		10		10		5		1					
$m=7$				1		6		15		20		15		6		1				
$m=8$			1		7		21		35		35		21		7		1			
$m=9$		1		8		28		56		70		56		28		8		1		
$m=10$	1		9		36		84		126		126		84		36		9		1	

Pascal's Triangle

- The above represents Pascal's triangle.
- If the values of 'm' are used to represent the number of elements of the array, then the coefficients of the expansion represent the relative amplitudes of the elements.
- Since the coefficients are determined from a binomial series expansion, the array is known as binomial array.

Non-uniform Amplitude Arrays with Odd & Even Elements

Assuming that the amplitude excitation is symmetrical about the origin, the array factor for a nonuniform amplitude array can be written as ;

$$(AF)_{2M}(\text{even}) = \sum_{n=1}^{M} a_n \cos[(2n-1)u]$$

$$(AF)_{2M+1}(\text{odd}) = \sum_{n=1}^{M+1} a_n \cos[2(n-1)u] \qquad \text{-------- (3.61)}$$

where

$$u = \frac{\pi d}{\lambda} \cos \phi$$

where a_n's are the excitation coefficients of the array elements.

- From the above equations, the amplitude co-efficients of the following arrays are –

1. Two elements ($2M = 2$)

   $a_1 = 1$

2. Three elements ($2M + 1 = 3$)

   $2a_1 = 2 \Rightarrow a_1 = 1$

   $a_2 = 1$

3. Four elements ($2M = 4$)

   $a_1 = 3$

   $a_2 = 1$

4. Five elements ($2M + 1 = 5$)

   $2a_1 = 6 \Rightarrow a_1 = 3$

   $a_2 = 4$

   $a_3 = 1$

- Binomial array's do not exhibit any minor lobes provided the spacing between the elements equal or less than one-half of a wavelength.
- The design using a $\lambda/2$ spacing leads to a pattern with no minor lobes, the half-power beam width and maximum directivity for $d = \lambda/2$ spacing in terms of the numbers of elements or the length of the array are given by ;

$$HPBW\ (d = \lambda/2) \approx \frac{0.75}{\sqrt{L/\lambda}}$$

$$D_{max} = 1.77\sqrt{1 + 2L/\lambda}$$

Advantages & Disadvantages of Binomial Arrays

- The advantages of binomial array is that there are no side lobes in the resultant pattern.
- The disadvantages are -
- i. Beam width of the main lobe is large which is undesirable
- ii. Directivity is small and high excitation levels are required for the center elements of large arrays.

UNIT – III

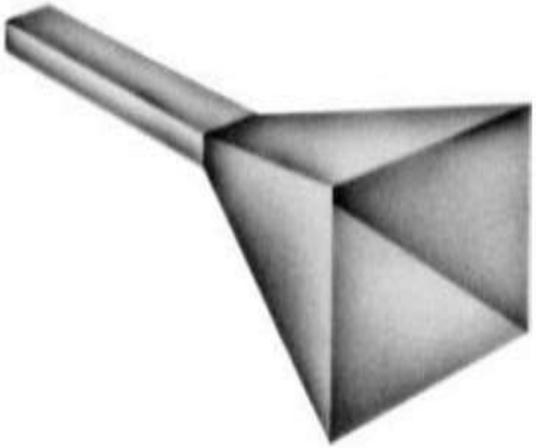
APERTURE AND SLOT ANTENNAS

Introduction to Aperture Antennas

Introduction to Aperture Antennas

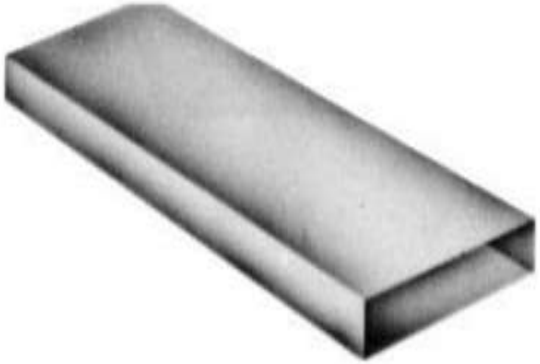
- Aperture antennas are most common at microwave frequencies.
- In general aperture means opening. Aperture in antenna, means opening in a closed surface.
- There are different geometrical configurations of an aperture antenna shown in Fig. 1.
- They may take the form of a waveguide or a horn whose aperture may be square, rectangular, circular, elliptical, or any other configuration.

Aperture Antenna Configuration



(a) Pyramidal horn

(b) Conical horn



(c) Rectangular wave guide

- Aperture antennas are very practical for space applications, because flush mounted on the surface of the spacecraft or aircraft.
- Their opening can be covered with a dielectric material to protect them from environmental conditions.
- This type of mounting does not disturb the aerodynamic profile of the craft for high-speed critical applications .
- The radiation characteristics of wire antennas can be determined once the current distribution on the wire known.
- For many configurations, current distribution is not known exactly & only physical experimental measurements can provide a reasonable approximation.

- This is even more evident in aperture antennas.
- Analysis of aperture type antennas is the conversion of original antenna geometry into an equivalent geometry which can be looked at as radiation through an aperture in a closed surface.
- This equivalence is obtained by the principle known as field equivalence principle.
- Along with this principle, the duality, uniqueness theorem and image principles are also useful in the aperture type antenna analysis.

Topic 1

Field Equivalence Principle

(Huygen’s Principle)

Huygen’s Principle

- Huygens’ principle states that “each point on a primary wave-front can be considered to be a new source of a secondary spherical wave and that a secondary wave-front can be constructed as the envelope of these secondary spherical waves.”
- The field equivalence principle is a principle by which actual sources, such as antenna and transmitter, are replaced by equivalent sources.
- The fictitious sources are said to be equivalent within a region because they produce the same fields within that region.

- By the equivalence principle, the fields outside an imaginary closed surface obtained by placing over the closed surface suitable electric and magnetic-current densities which satisfy the boundary conditions.
- The current densities are selected so that the fields inside the closed surface are zero and outside they are equal to the radiation produced by the actual sources.
- The equivalence principle is developed by considering an actual radiating source, which electrically is represented by current densities $\boldsymbol{J}_1$ and $\boldsymbol{M}_1$, as shown in Fig. 2.
- The source radiates fields $\boldsymbol{E}_1$and $\boldsymbol{H}_1$everywhere.

- It is desired to develop a method that will yield the fields outside a closed surface.
- A closed surface S is chosen, shown dashed in Fig. 2, which encloses the current densities $\boldsymbol{J}_1$ and $\boldsymbol{M}_1$. The volume within S is denoted by V_1 and outside S by V_2.
- The primary task will be to replace the original problem shown in Fig. 2, by an equivalent one which yields the same fields $\boldsymbol{E}_1$and $\boldsymbol{H}_1$ outside S (within V_2).
- An equivalent problem of Fig. 2 is shown in Fig. 2 (b). The original sources $\boldsymbol{J}_1$ and $\boldsymbol{M}_1$ are removed, and we assume that there exist fields $\boldsymbol{E}$ and $\boldsymbol{H}$ inside S and fields $\boldsymbol{E}_1$and $\boldsymbol{H}_1$ outside of S.

- For these fields to exist within and outside S, they must satisfy the boundary conditions on the tangential electric and magnetic field components.
- Thus on the imaginary surface S there must exist the equivalent sources and they radiate into an unbounded space (same medium everywhere).

$$\boldsymbol{J_s} = \boldsymbol{\hat{n}} \times [\boldsymbol{H_1} - \boldsymbol{H}]$$

$$\boldsymbol{M_s} = -\boldsymbol{\hat{n}} \times [\boldsymbol{E_1} - \boldsymbol{E}]$$

- The current densities of above equations are said to be equivalent only within V_2, because they produce the original fields ($\boldsymbol{E_1}$, $\boldsymbol{H_1}$) only outside S.

Actual & Equivalent Models

- Fields $\boldsymbol{E}$, $\boldsymbol{H}$ different from the originals ($\boldsymbol{E}_1$, $\boldsymbol{H}_1$), result within V_1.

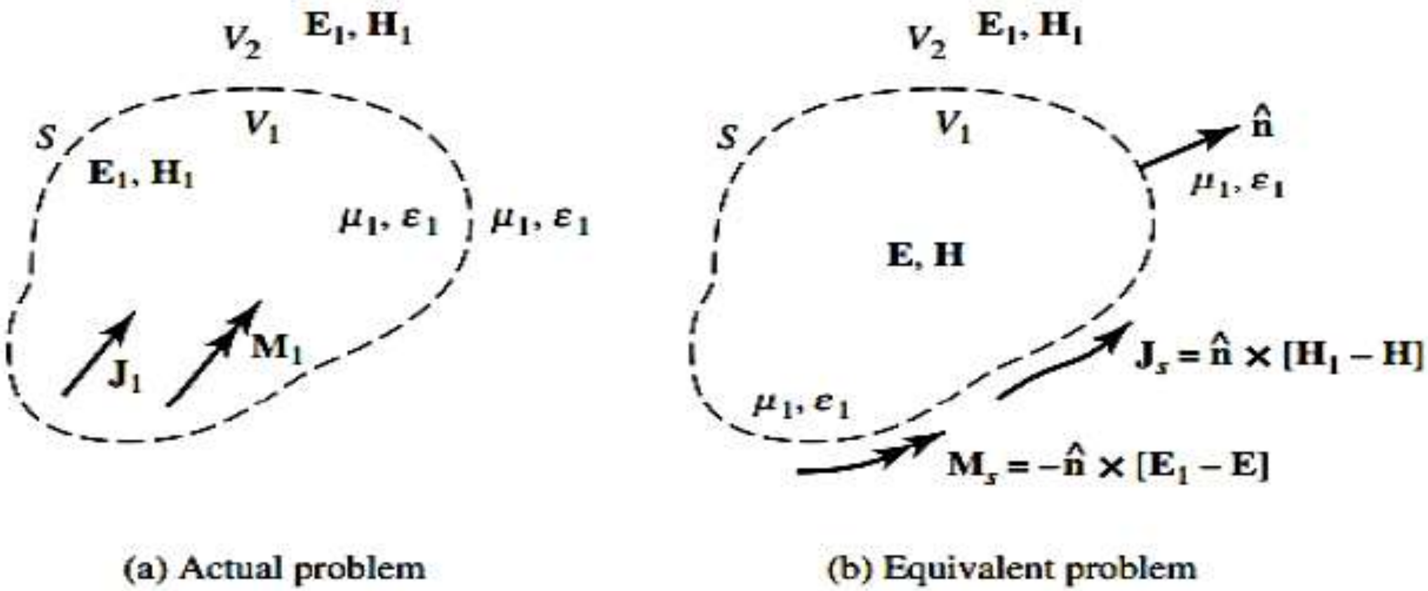


Fig. 2-2. Actual and equivalent models.

- From electromagnetic uniqueness concepts, it is known that the tangential components of only $\boldsymbol{E}$ or $\boldsymbol{H}$ are needed to determine the fields.
- Since the fields $\boldsymbol{E}$, $\boldsymbol{H}$ within S can be anything, it can be assumed that they are zero.
- In that case the equivalent problem of Fig. 2-2 (b) reduces to that of Fig. 2-2 (a) with the equivalent current densities being equal to -

$$\boldsymbol{J_s} = \hat{\boldsymbol{n}} \times [\boldsymbol{H_1} - \boldsymbol{H}]|_{\boldsymbol{H}=0} = \hat{\boldsymbol{n}} \times \boldsymbol{H_1}$$

$$\boldsymbol{M_s} = -\hat{\boldsymbol{n}} \times [\boldsymbol{E_1} - \boldsymbol{E}]|_{\boldsymbol{E}=0} = -\hat{\boldsymbol{n}} \times \boldsymbol{E_1}$$

Equivalence Principle Models

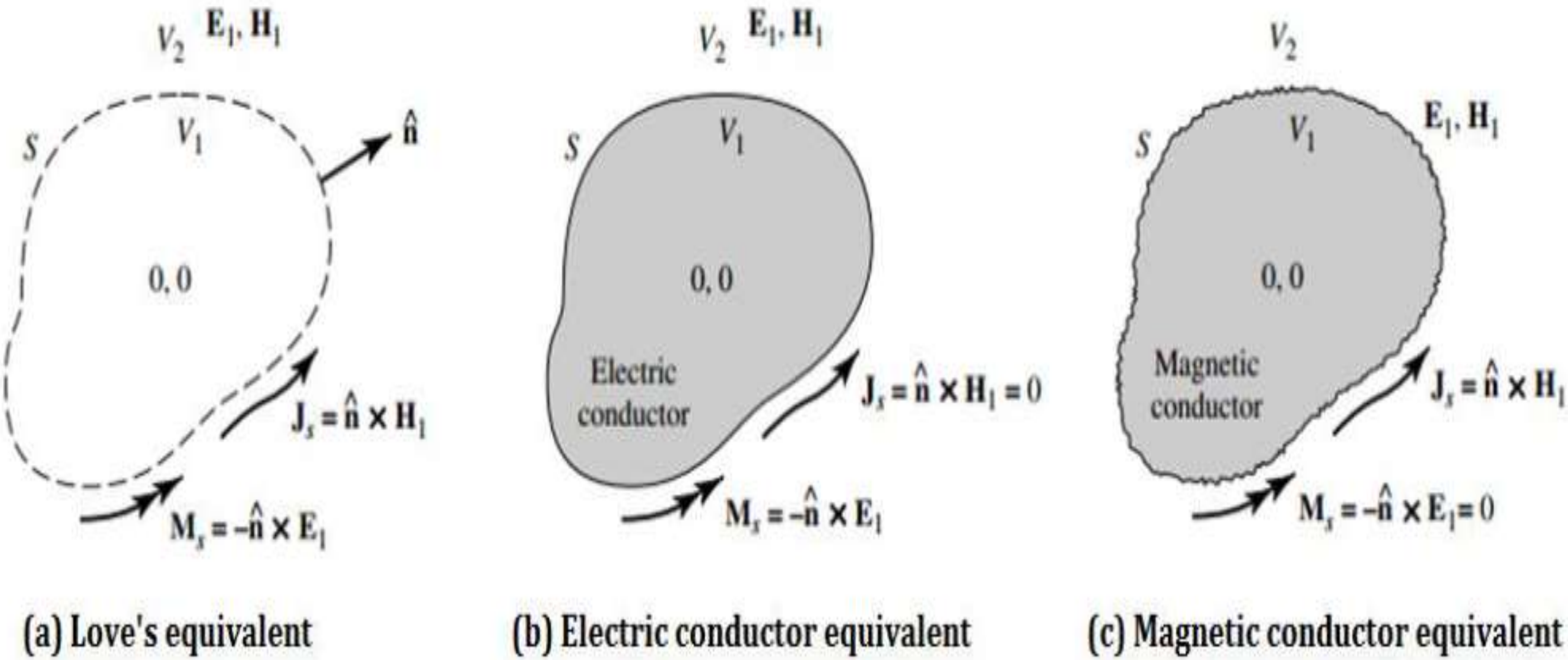


Fig. 2-2 Equivalence principle models

- Love's Equivalence Principle of Fig. 2-3 (a) produces a null field within the imaginary surface S.
- Since the value of the $\boldsymbol{E} = \boldsymbol{H} = \mathbf{0}$ within S cannot be disturbed if the properties of the medium within it are changed, let us assume that it is replaced by a perfect electric conductor ($\sigma = \infty$) .
- The introduction of the perfect conductor will have an effect on the equivalent source $\boldsymbol{J_s}$. As the electric conductor takes its place, as shown in Fig. 2-3 (b), the electric current density $\boldsymbol{J_s}$, which is tangent to the surface S, is short-circuited by the electric conductor.
- Thus the equivalent problem of Fig. 2-3 (a) reduces to that of Fig. 2-3 (b).

- There exists only a magnetic current density $\boldsymbol{M_s}$ over S, and it radiates in the presence of the electric conductor producing outside S the original fields $\boldsymbol{E_1}, \boldsymbol{H_1}$.
- Let us assume that instead of placing a perfect electrical conductor within S we introduce a perfect magnetic conductor which will short out the magnetic current density and reduce the equivalent problem to that shown in Fig. 2-3 (c).

Topic 2

Radiation from Aperture Fields

- The general coordinate system for aperture antenna analysis is shown in Fig. 2-5. The radiating fields are determined by first finding the vector potentials $\boldsymbol{A}$ and F, from the surface current densities $\boldsymbol{J_s}$ and $\boldsymbol{M_s}$ respectively.

Coordinate System for Aperture Antenna Analysis

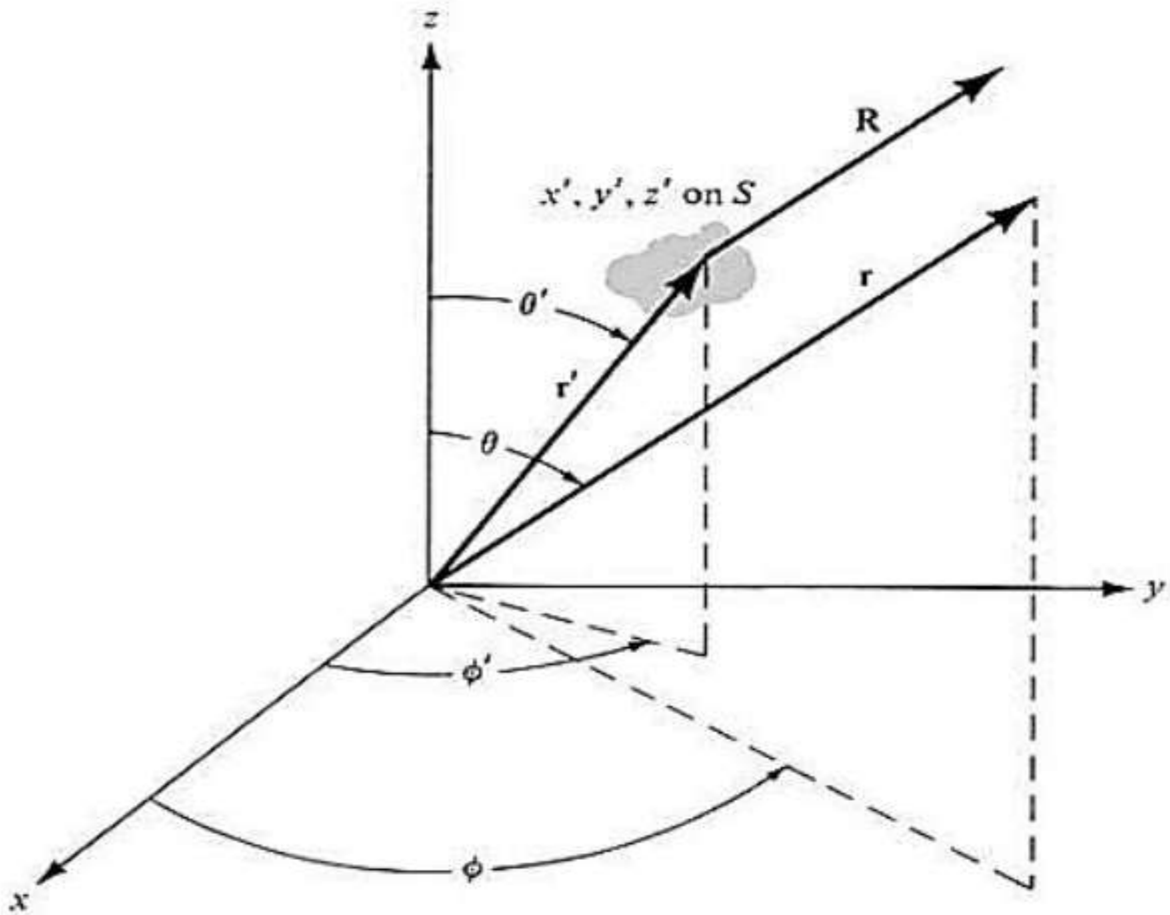


Fig. 2-5 Coordinate system for aperture antenna analysis.

For far-field approximation:

$$R \approx r - r' \cos\psi \qquad \text{for phase variations} \qquad \text{-------- (2.19)}$$

$$R \approx r \qquad \text{for amplitude variations} \qquad \text{-------- (2.20)}$$

The auxiliary potential functions $\boldsymbol{A}$ and $\boldsymbol{F}$ generated by the current densities $\boldsymbol{J_s}$ and $\boldsymbol{M_s}$ is given by ;

$$\mathbf{A} = \frac{\mu}{4\pi} \iint_S \mathbf{J}_s \frac{e^{-jkR}}{R} ds' \simeq \frac{\mu e^{-jkr}}{4\pi r} \mathbf{N} \qquad \text{-------- (2.21)}$$

$$\mathbf{N} = \iint_S \mathbf{J}_s e^{jkr' \cos\psi} ds' \qquad \text{-------- (2.22)}$$

$$\mathbf{F} = \frac{\epsilon}{4\pi} \iint_S \mathbf{M}_s \frac{e^{-jkR}}{R} ds' \simeq \frac{\epsilon e^{-jkr}}{4\pi r} \mathbf{L} \qquad \text{-------- (2.23)}$$

$$\mathbf{L} = \iint_S \mathbf{M}_s e^{jkr' \cos\psi} ds' \qquad \text{-------- (2.24)}$$

In the far-field only the θ and ϕ components of the ***E*** and ***H*** fields are dominant.

$$\begin{aligned} (E_A)_\theta &= -j\omega A_\theta \\ (E_A)_\phi &= -j\omega A_\phi \\ (E_F)_\theta &= -j\omega\eta F_\phi \\ (E_F)_\phi &= j\omega \eta F_\theta \end{aligned} \qquad \text{-------- (2.25)}$$

$$\begin{aligned} (H_A)_\theta &= j\omega \frac{A_\phi}{\eta} \\ (H_A)_\phi &= -j\omega \frac{A_\theta}{\eta} \\ (H_F)_\theta &= -j\omega F_\theta \\ (H_F)_\phi &= j\omega F_\phi \end{aligned} \qquad \text{-------- (2.26)}$$

Combining Eqn's. (2.25) to (2.26), yields :

$$E_r \approx 0$$

$$E_\theta \approx -j\frac{\kappa}{4\pi r}[L_\phi + nN_\theta]e^{-j\kappa r}$$

$$E_\phi \approx j\frac{\kappa}{4\pi r}[L_\theta - \eta N_\phi]e^{-j\kappa r}$$

$$H_r \approx 0 \qquad \text{-------- } (2.27)$$

$$H_\theta \approx j\frac{\kappa}{4\pi r}[N_\phi - \frac{L_\theta}{\eta}]e^{-j\kappa r}$$

$$H_\phi \approx -j\frac{\kappa}{4\pi r}[N_\theta + \frac{L_\phi}{\eta}]e^{-j\kappa r}$$

where ;

$$N_\theta = \iint_S \left[J_x \cos\theta \cos\varphi + J_y \cos\theta \sin\varphi - J_z \sin\theta \right] e^{+jkr'\cos\psi} ds'$$

$$N_\varphi = \iint_S \left[-J_x \sin\varphi + J_y \cos\varphi \right] e^{+jkr'\cos\psi} ds'$$

-------- (2.28)

$$L_\theta = \iint_S \left[M_x \cos\theta \cos\varphi + M_y \cos\theta \sin\varphi - M_z \sin\theta \right] e^{+jkr'\cos\psi} ds'$$

$$L_\varphi = \iint_S \left[-M_x \sin\varphi + M_y \cos\varphi \right] e^{+jkr'\cos\psi} ds'$$

Topic 3

Radiation from Rectangular Apertures

Radiation from Uniform Apertures

Consider that a rectangular aperture is mounted on an infinite ground plane (Fig. 2-6) assuming field over the opening to be constant and is given by :

$$\mathbf{E}_a = \hat{\mathbf{a}}_y E_0 \quad -a/2 \le x' \le a/2, \quad -b/2 \le y' \le b/2 \qquad \text{-------- (2.28)}$$

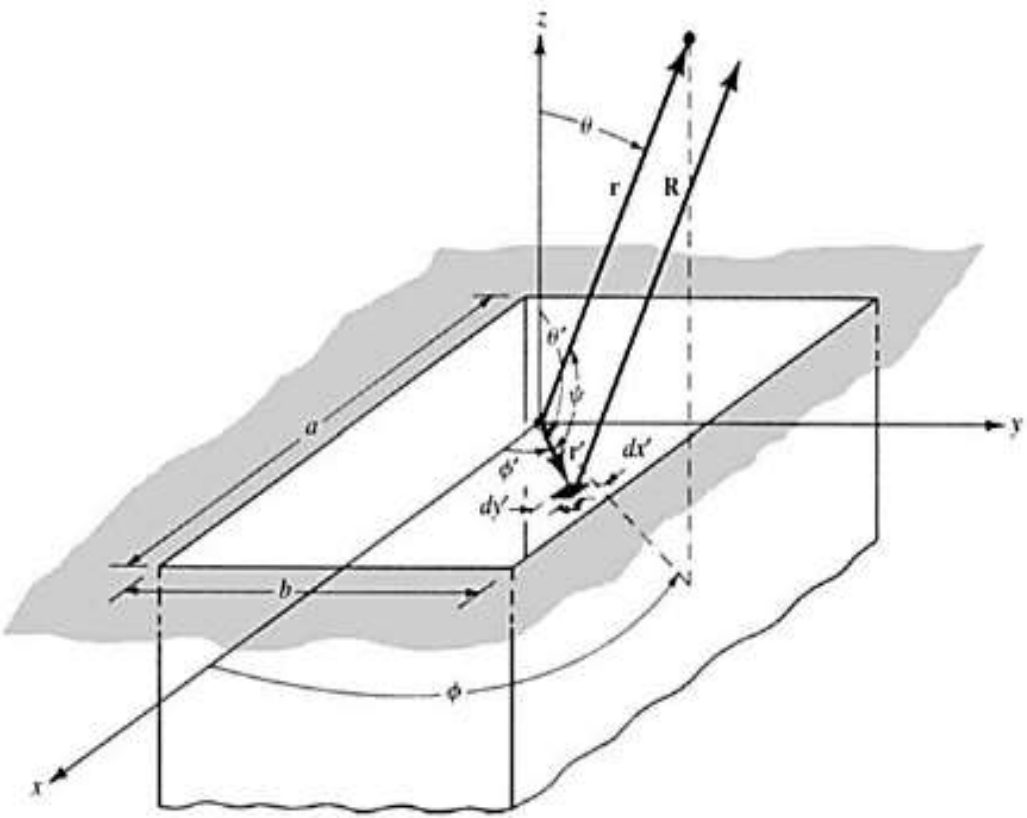


Fig. 2-6 Rectangular aperture on an infinite electric ground plane.

To obtain equivalent of this situation, assuming a closed surface extending from $-\infty$ to ∞ on the xy plane. Then we can write:

$$\mathbf{M}_s = \begin{cases} -2\hat{\mathbf{n}} \times \mathbf{E}_a = -2\hat{\mathbf{a}}_z \times \hat{\mathbf{a}}_y E_0 = +\hat{\mathbf{a}}_x 2E_0 & \begin{array}{l} -a/2 \le x' \le a/2 \\ -b/2 \le y' \le b/2 \end{array} \\ 0 & \text{elsewhere} \end{cases} \qquad \text{-------- (2.29)}$$

$$\mathbf{J}_s = \quad 0 \qquad\qquad \text{everywhere}$$

The far-zone fields radiated by the aperture of Fig. 2-6 can be found by using Eqn. (2.27), (2.28) and (2.9). Thus,

$$N_\theta = N_\phi = 0$$

$$L_\theta = \int_{-b/2}^{+b/2} \int_{-a/2}^{+a/2} [M_x \cos\theta \cos\phi] e^{jk(x' \sin\theta \cos\phi + y' \sin\theta \sin\phi)} dx'\, dy' \qquad \text{-------- (2.30)}$$

$$L_\theta = \cos\theta\ \cos\phi \left[\int_{-b/2}^{+b/2}\int_{-a/2}^{+a/2} M_x e^{jk(x'\sin\theta\cos\phi + y'\sin\theta\sin\phi)}\, dx'\, dy'\right] \quad \text{-------- (2.31)}$$

Using the integral

$$\int_{-c/2}^{+c/2} e^{j\alpha z}\, dz = c\left[\frac{\sin\left(\frac{\alpha}{2}c\right)}{\frac{\alpha}{2}c}\right]$$

- Eqn. (2.31) reduces to ;

$$L_\theta = 2abE_0\left[\cos\theta\cos\phi\left(\frac{\sin X}{X}\right)\left(\frac{\sin Y}{Y}\right)\right]$$

where

$$X = \frac{ka}{2}\sin\theta\cos\phi \quad \text{-------- (2.32)}$$

$$Y = \frac{kb}{2}\sin\theta\sin\phi$$

Similarly ;

$$L_\phi = -2abE_0\left[\sin\phi\left(\frac{\sin X}{X}\right)\left(\frac{\sin Y}{Y}\right)\right] \quad \text{-------- (2.33)}$$

From the above Eqn's. the fields radiated by the aperture can be written as

$$E_r = 0$$

$$E_\theta = j\frac{abkE_0e^{-jkr}}{2\pi r}\left[\sin\phi\left(\frac{\sin X}{X}\right)\left(\frac{\sin Y}{Y}\right)\right]$$

$$E_\phi = j\frac{abkE_0e^{-jkr}}{2\pi r}\left[\cos\theta\cos\phi\left(\frac{\sin X}{X}\right)\left(\frac{\sin Y}{Y}\right)\right]$$

$$H_r = 0$$

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

$$H_\phi = +\frac{E_\theta}{\eta} \quad \text{-------- (2.34)}$$

E −Plane ($\phi = \pi/2$)

$$E_r = E_\phi = 0$$

$$E_\theta = j\frac{abkE_0e^{-jkr}}{2\pi r}\left[\frac{\sin\left(\frac{kb}{2}\sin\theta\right)}{\frac{kb}{2}\sin\theta}\right]$$

H −Plane ($\phi = 0$)

$$E_r = E_\theta = 0$$

$$E_\phi = j\frac{abkE_0e^{-jkr}}{2\pi r}\left\{\cos\theta\left[\frac{\sin\left(\frac{ka}{2}\sin\theta\right)}{\frac{ka}{2}\sin\theta}\right]\right\}$$

Radiation from Tapered Apertures

One practical aperture of tapered source distribution is the open rectangular waveguide. The dominant TE_{10} mode has the following distribution:

$$\mathbf{E}_a = \hat{\mathbf{a}}_y E_0 \cos\left(\frac{\pi}{a}x'\right) \left.\begin{matrix} -a/2 \le x' \le a/2 \\ -b/2 \le y' \le b/2 \end{matrix}\right\} \quad \text{-------- (2.35)}$$

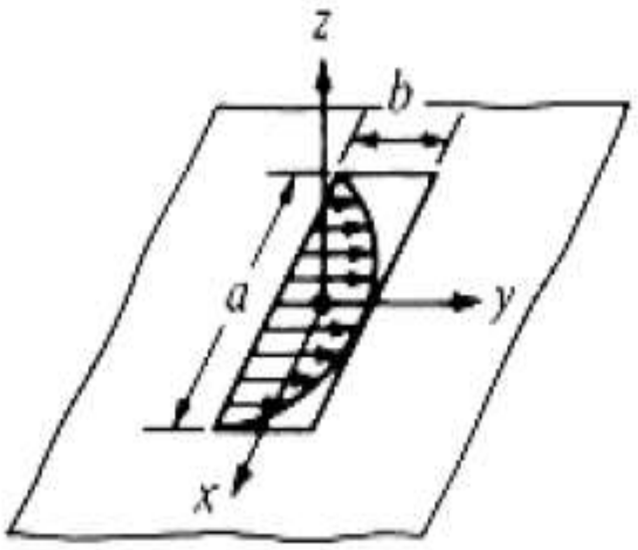


Fig. 2-7 TE_{10} -Mode Distribution Aperture on Ground Plane.

To obtain equivalent of this situation, assuming a closed surface extending from $-\infty$ to ∞ on the xy plane. Then we can write:

$$\mathbf{M}_s = \begin{cases} -2\hat{\mathbf{n}} \times \mathbf{E}_a & \left.\begin{matrix} -a/2 \le x' \le a/2 \\ -b/2 \le y' \le b/2 \end{matrix}\right\} \\ 0 & \text{elsewhere} \end{cases}$$
$$\mathbf{J}_s = 0 \qquad \text{everywhere} \qquad \text{-------- (2.36)}$$

Similarly the fields are computed as ;

$$E_r = H_r = 0$$
$$E_\theta = -\frac{\pi}{2} C \sin\phi \frac{\cos X}{(X)^2 - \left(\frac{\pi}{2}\right)^2} \frac{\sin Y}{Y}$$
$$E_\phi = -\frac{\pi}{2} C \cos\theta \cos\phi \frac{\cos X}{(X)^2 - \left(\frac{\pi}{2}\right)^2} \frac{\sin Y}{Y} \qquad \text{-------- (2.37)}$$
$$H_\theta = -E_\phi/\eta$$
$$H_\phi = E_\theta/\eta$$

Topic 4

Slot Antennas

Introduction to Slot Antennas

- Slot antenna is a radiating element formed by a slot in a metallic surface. An opening cut in a conducting sheet or in one of the walls of the waveguide acts as the antenna.
- It is excited either by a co-axial cable or through the waveguide. Slot antenna is the best suitable radiator at frequencies above 200 MHz.
- Consider an infinite conducting sheet as in Fig. 2-8 (a). Now consider that an aperture of any size or shape is made leaving a slot on a sheet.
- The flat strip obtained can be treated as short dipole as shown in Fig. 2-8 (b).

Metallic Conducting Sheet

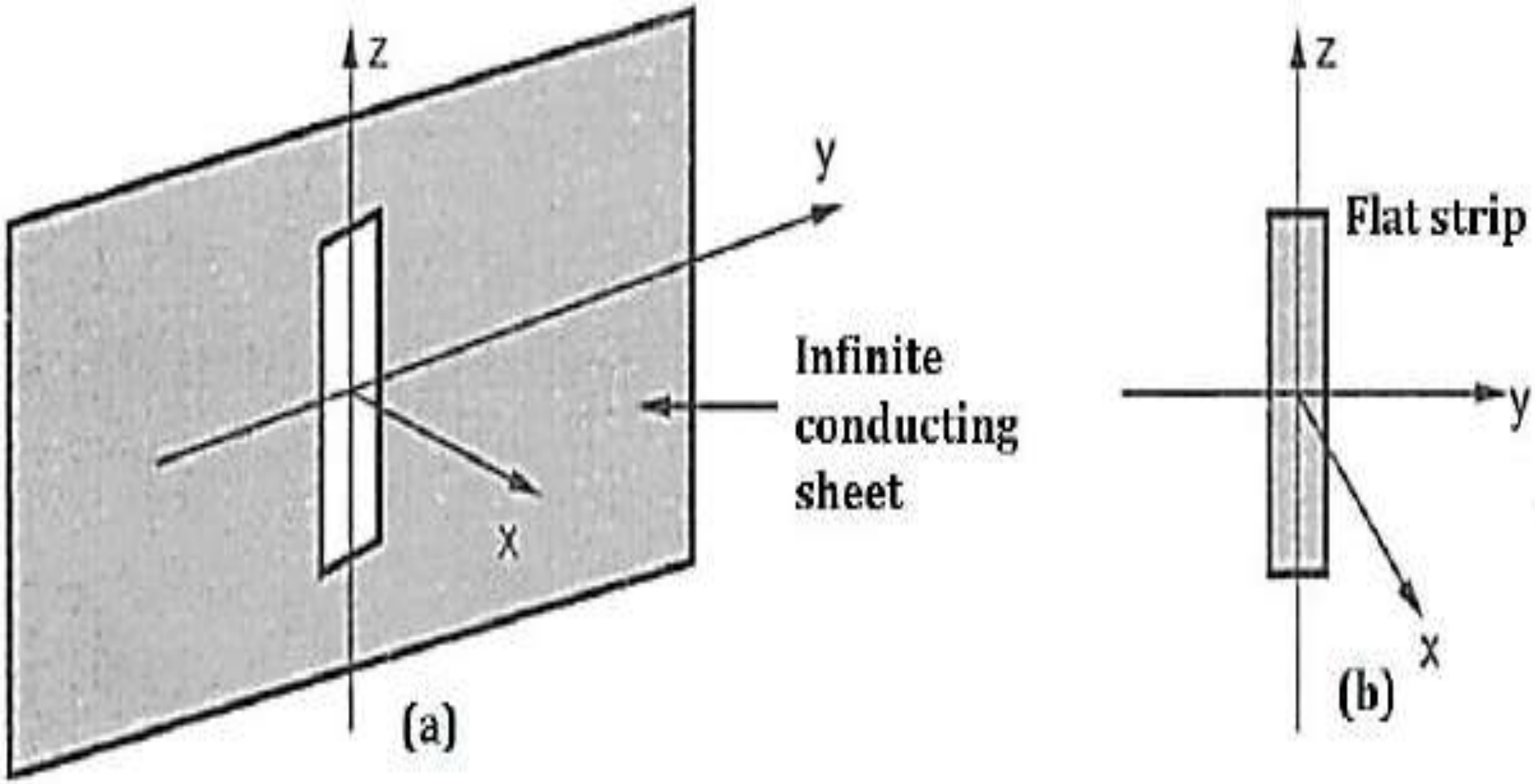


Fig. 2-8 Metallic conducting sheet and complementary flat strip

- When the two are combined together we get the complete original infinite conducting sheet.
- Hence the infinite conducting sheet with slot and the flat strip of the dimension same as of the slot are said to be complementary.
- Now consider that a slot of $\lambda/2$ is cut in a large conducting sheet, we get complementary dipole antenna.
- In general, the slot antenna is fed by either a generator (or) transmission line connected across it.

Working Principle of Slot Antenna

- Whenever a high frequency field exists across a very narrow slot in an infinite conducting sheet, the energy is radiated through slot. This is the working principle of the slot antenna.
- In case of the waveguides, the slot antenna is fed with the guided waves incident on slot. Consider that the slot antenna is fed with a transmission line connected across points A & B as in Fig. 2-9 (a).
- As the antenna is fed with a transmission line, the slot will radiate due to the currents in the conducting sheet. The complementary of the slot antenna is the dipole as shown in Fig. 2-9 (b).

- For the complementary dipole antenna, the regions with conducting sheet and air are interchanged.
- According to the G. Booker's theory, the field pattern of the slot is exactly identical in shape as that of the half dipole with $\boldsymbol{E}$ and $\boldsymbol{H}$ interchanged. That means for the slot, the electric field $\boldsymbol{E}$ will be horizontally polarized, while for the dipole, it is vertically polarized.
- A single half wavelength slot in a conducting sheet is analogous to the half wave dipole in terms of gain and directivity with only difference in the polarization.
- The horizontal slot produces vertical polarization in the direction normal to the slot, while the vertical slot produces horizontal polarization.

Slot & Dipole Antennas

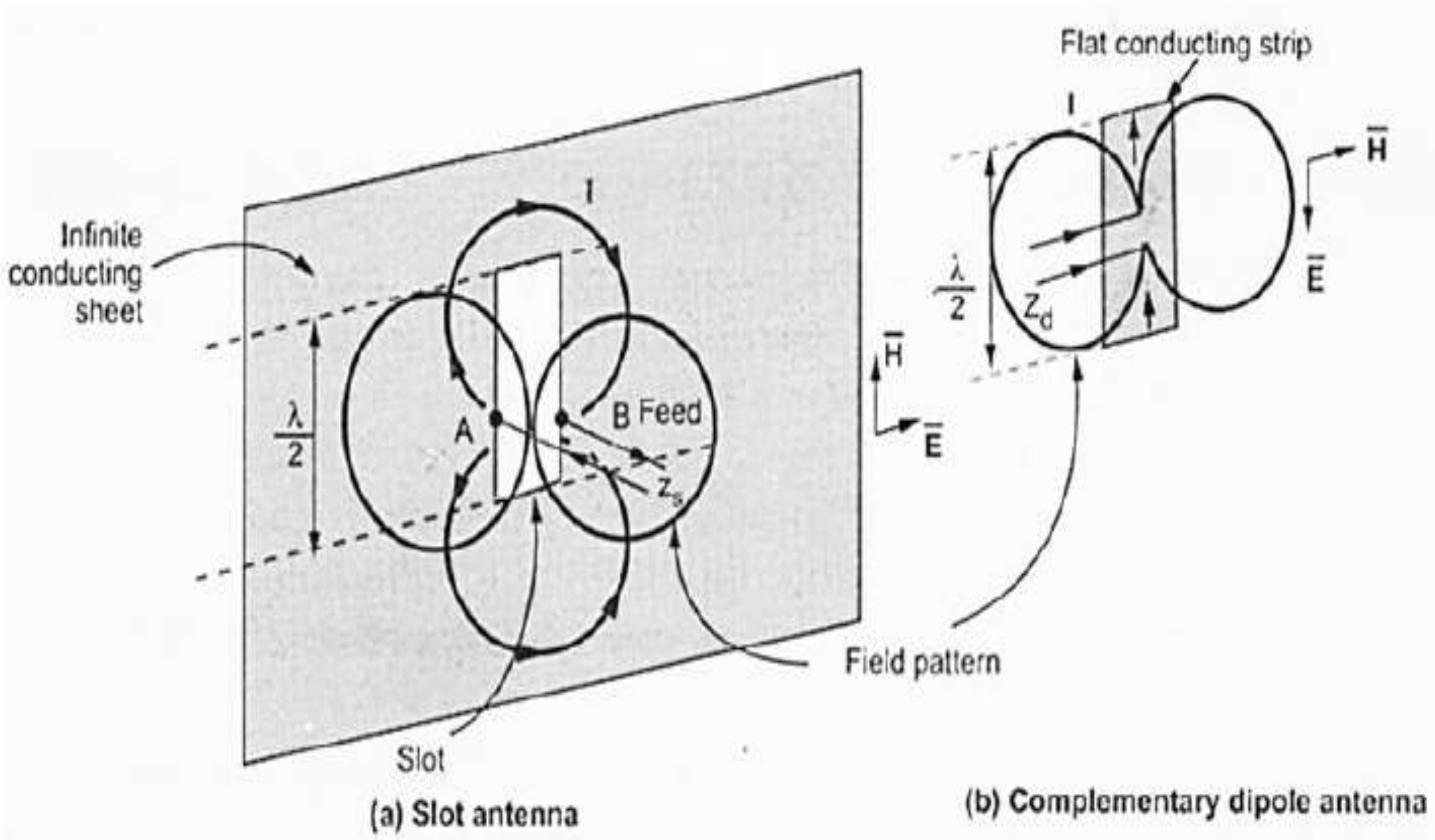


Fig. 2-9 Slot and complementary dipole antenna.

Although the width of the slot is small ($\omega \ll \lambda$), the currents are not confined to the edges of the slot but spread out over the sheet.

The terminal impedance Z_s of the slot is related to the terminal impedance of dipole Z_d by intrinsic impedance η of free space by the relation,

$$Z_s Z_d = \frac{\eta^2}{4} = \frac{(277)^2}{4} \qquad \text{-------- (2.38)}$$

$$Z_s \simeq \frac{25{,}476}{Z_d} \qquad \text{-------- (2.39)}$$

Suppose that the terminal impedance of dipole antenna is $Z_d = 72 + j\,42.5\,\Omega$, then the terminal impedance of the complementary slot will be,

$$Z_s = \frac{25476}{72 + j\,42.5} = \frac{25476 \times (72 - j\,42.5)}{(72 + j\,42.5)\,(72 - j\,42.5)}$$

$$Z_s \approx 262 - j\,211\,\Omega \qquad \text{-------- (2.40)}$$

Slot Antenna Vs Complimentary Dipole

- Polarization is different in both the antennas.
- That means if the polarization is horizontal in slot antenna, then it is vertical in the complementary antenna.
- The radiations from the backside of the slot antenna and the complementary antenna are of opposite polarity.

Methods of Feeding Slot Antennas

Feeding 1 using Co-axial Line

- Practically the slot antenna is fed with a co-axial transmission line.
- The outer conductor is bonded to the metal sheet as shown in Fig. 2-10 (a). In general, the terminal impedance of $\lambda/2$.
- Slot in a conducting large sheet is very large (approximately 500 Ω), while the characteristic impedance of the transmission line is much smaller.
- Thus under such conditions, to provide proper impedance matching, the off-center feed as shown in Fig. 2-10 (b) is used.

- In practical applications, generally a slot antenna fed with a transmission line is not observed.
- Instead of that the slot is boxed in any one of the sides of the metallic cavity so that by properly selecting dimensions of the cavity, the outward radiation from the opening of the cavity is not affected, while the backward radiation is virtually eliminated.

Feeding 1 Using Co-axial Line

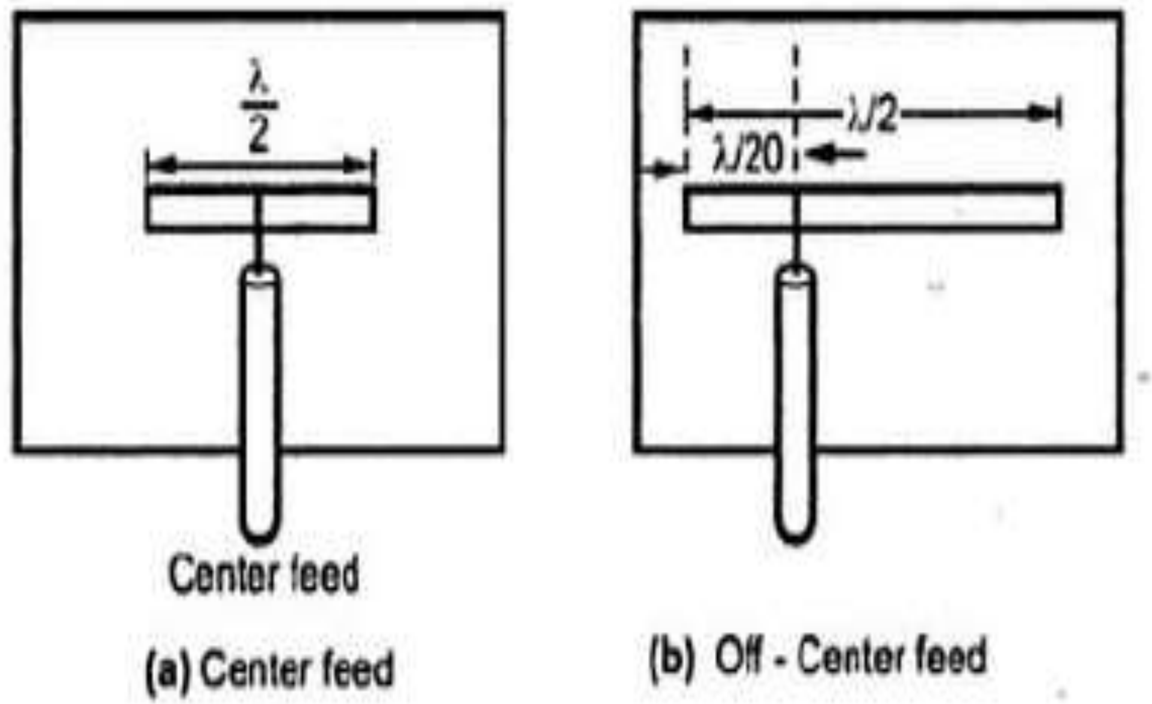


Fig. 2-10 Feeding of slot antenna using co-axial cable.

Feeding 2 - Cylindrical Slot Antenna

- For the applications at very high frequencies, a slot antenna with a slot cut in a conducting cylinder is most widely used.
- A longitudinal slot in infinitely long cylinder as shown in Fig. 2-11 (a), produces circular radiation that diameter of the cylinder is very small.
- The gain and directivity properties of a basic slot antenna can be improved by using array of slots placed half guide wavelength apart and placed on opposite side of central line as shown in Fig. 2-11 (b).
- Actually are the slots radiates in same phase, but there is a reversal of polarity of the field inside the guide.

Cylindrical Slot Antenna

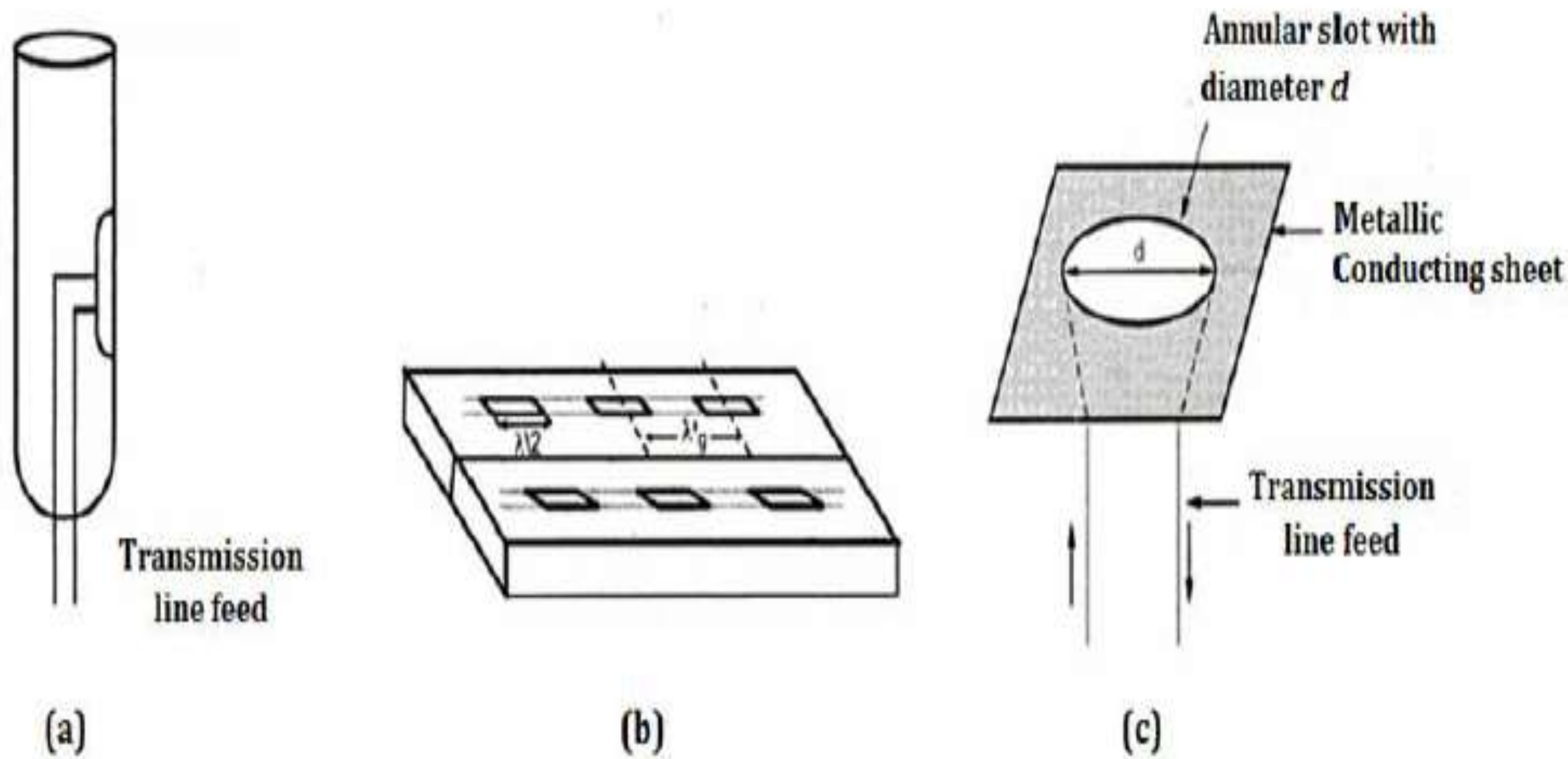


Fig. 2-11 a) Cylindrical slot antenna b) Planar array slot antenna
c) Annular slot antenna

- The shape of the slot may be either rectangular or circular.
- The slot with circular or annular shape is called annular slot antenna.
- The annular slot antenna is shown in 2-11 (c).
- When the diameter of the annular slot is less than half wavelength the resulting radiations are identical to those produced by short, vertical antenna.

Feeding 3 - Boxed Slot Antenna

- A flat sheet with a $\lambda/2$ slot radiates equally on both sides of the sheet.
- However, if the sheet is very large (ideally infinite) and boxed in as in Fig. 2-12, radiation occurs only from one side.
- The depth d of the box is approximate ($d \sim \lambda/4$) for a thin slot.

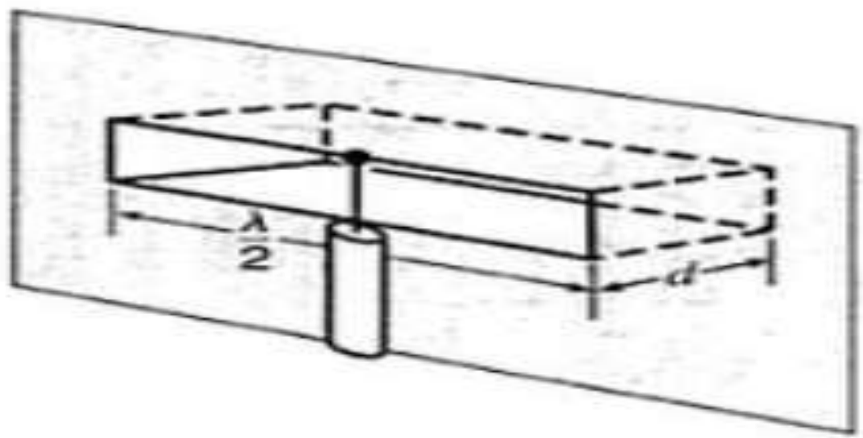


Fig. 2-12 Boxed-in slot antenna

Topic 5

Babinet's Principle

Statement of Babinet's Principle

- Babinet's principle states that when the field behind a screen with an opening is added to the field of a complementary structure, the sum is equal to the field when there is no screen.

Babinet's Principle

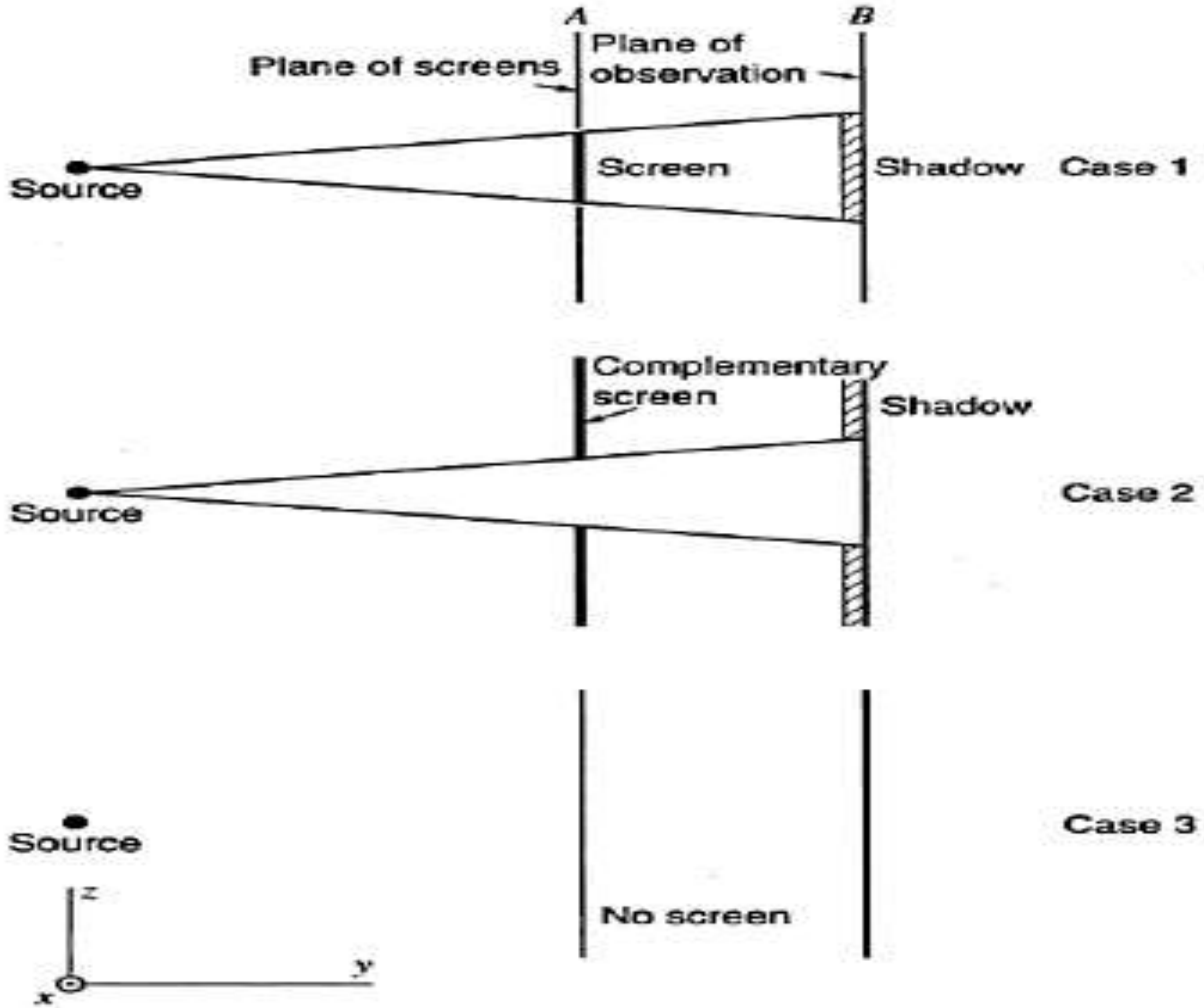


Fig. 2-13 Illustartion of Babinet's principle.

- Consider case 1 where a perfectly absorbing screen is placed in plane A which has a region of shadow in the observation plane B as shown in Fig. 2-13 (a).
- Let the field behind the screen be some function f_1 of x, y and z is given by $F_s = f_1(x, y, z)$.
- Now case 2 where the first screen is replaced by its complementary screen as shown in Fig. 2-13 (b).
- Let the field behind complementary screen is given by $F_{cs} = f_2(x, y, z)$.
- Finally in case 3, no screen is present and the field is given by $F_0 = f_3(x, y, z)$.
- Now according to Babinet's principle, at same point, the total field is given as $F_0 = F_s + F_{cs}$.

Topic 6

Horn Antenna

Introduction to Horn Antenna

- One of the simplest and probably the most widely used microwave antenna is the horn.
- The horn is widely used as a feed element for large radio astronomy, communication dishes and satellite tracking through out the world.
- The horn antenna can be considered as a waveguide with hollow pipe of different cross-sections which is flared or tapered into a large opening.
- When one end of the waveguide is excited while other end is kept open, it radiates in open space in all directions.

- As compared with the radiation through transmission line, the radiation through the waveguide is larger.
- In waveguide, a small portion of the incident wave is radiated and large portion is reflected back due to the open circuit.
- As one end of the waveguide is open circuited, the impedance matching with the free space is not perfect.
- To minimize reflections of the guided wave , the mouth of the waveguide is flared or opened out such that it assumes shape like horn.
- A horn antenna is nothing but a flared out or opened out waveguide. The main function of the horn antenna is to produce an uniform phase front with an aperture larger than waveguide to give higher directivity.

Types of Horn Antennas

- Basically the horn antennas are classified as rectangular horn antennas and circular horn antennas.
- The rectangular horn antennas are fed with rectangular waveguide, while the circular horn antennas are fed with circular waveguide.
- Depending upon the direction of flaring, the rectangular horns are further classified as Sectoral horn and Pyramidal horn.
- A sectoral horn is obtained if the flaring (tapering) is done in one direction only. A sectoral horn is further classified as E-plane sectoral horn and H-plane sectoral horn.

- The E-plane sectorial horn is obtained when the flaring is done in the direction of the electric field vector. The H-plane sectorial horn is obtained if the flaring is done in the direction of the magnetic field vector.
- When the flaring is done along both the walls of the rectangular waveguide in the direction of both the electric and magnetic field vectors, the horn obtained is called pyramidal horn.
- Similar to the rectangular horns, the circular horn antennas can be obtained by flaring the walls of the circular waveguide. The circular horn antennas are of two types namely conical horn antenna and biconical horn antenna.

- Many times, the transition region between the throat of the waveguide and the aperture is tapered with a gradual exponential taper. This minimizes the reflections of the guided waves. Such horns are called exponentially tapered horn antennas.
- Fig. 2-14 shows the horn antennas such as the E-plane sectoral horn, H-plane sectoral horn, pyramidal horn and conical horn.
- When the aperture size is large compared to the wavelength the wave impedance approaches the free space impedance. Thus, a pyramidal horn provides a slow transition from the waveguide impedance to the free space impedance, provided that the length of the transition is large compared to the wavelength.

Typical Horn Antennas

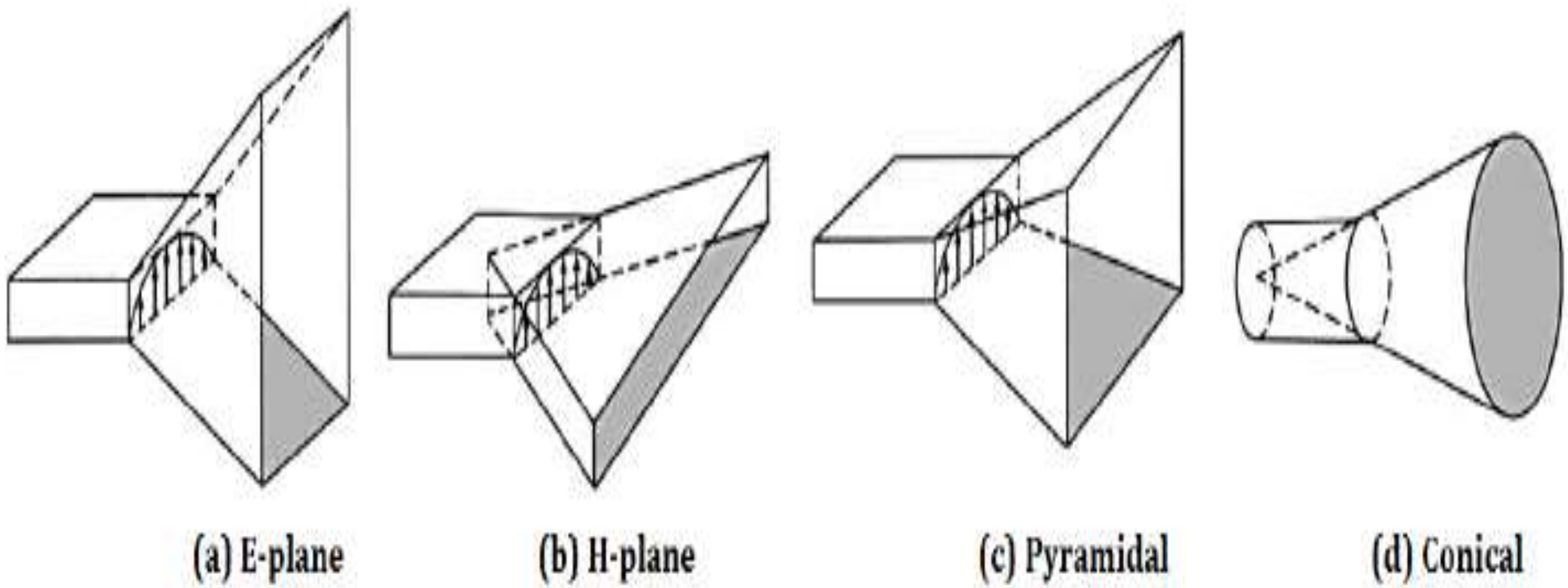


Fig. 2-14 Typical horn antennas

Design of Horn Antenna

The cross sectional view of rectangular horn antenna is shown in Fig. 2-15.

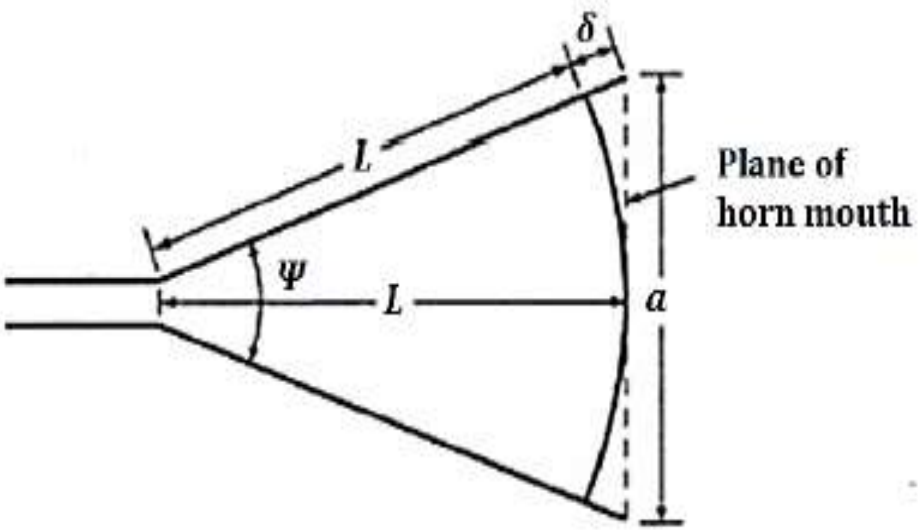


Fig. 2-15 Cross sectional view of rectangular horn antenna

From the geometry ;

$$\cos\frac{\Psi}{2} = \frac{L}{L+\delta}$$

$$\sin\frac{\Psi}{2} = \frac{a}{2(L+\delta)}$$

$$\tan\frac{\Psi}{2} = \frac{a}{2L}$$

where Ψ = flare angle (Ψ_E for E plane, Ψ_H for H plane), a = aperture (a_E for E plane, a_H for H plane), L = horn length and δ = path difference.

- From the geometry we have also that

$$L = \frac{a^2}{8\delta} \quad (\delta \ll L) \qquad \text{-------- (2.41)}$$

$$\Psi = 2\tan^{-1}\frac{a}{2L} = 2\cos^{-1}\frac{L}{L+\delta} \qquad \text{-------- (2.42)}$$

- When the flare angle is small, the aperture area for a specified length becomes small. Thus at the mouth of the horn, the uniform phase front is resulted, which increases directivity with decrease in the beam width. The angle represented in Eqn. (2.42) is known as optimum aperture angle.

- The directivity of maximum value can be obtained at the largest flare angle for which the value
- δ does not exceed typical value such as 0.25 λ for E-plane sectora horn, 0.22 λ for conical horn and 0.40 λ for H-plane sectoral horn.
- The directivity of the pyramidal horn and conical horn is highest as compared to other types of the horns because they have more than one flare angle.
- One more advantage of the horn antenna is that it can be operated over a wide range of high frequency as there is no resonant element in the antenna.

Applications of Horn Antenna

- It is used as a feed element in antennas such as parabolic reflectors
- It is the most wide used antenna for measurement of various antenna parameters in the laboratories.
- It is most suitable antenna for various application in microwave frequency range where moderate gains are sufficient.

Topic 7

Reflector Antennas – Plane Reflector, Corner Reflector, Parabolic Reflector

Introduction to Reflector Antennas

- The reflector antennas are most important in microwave radiation applications. At microwave frequencies the physical size of the high gain antenna becomes so small that practically any suitable shaped reflector can produce desired directivity.
- In reflector antenna, another antenna is required to excite it. Hence the antenna such as dipole, horn, slot which excites the reflector antenna is called primary antenna, while the reflector antenna is called secondary antenna.
- In general, reflector antenna can be represented in any geometrical configuration, but the most commonly used shapes are plane reflector, corner reflector and curved or parabolic reflectors.

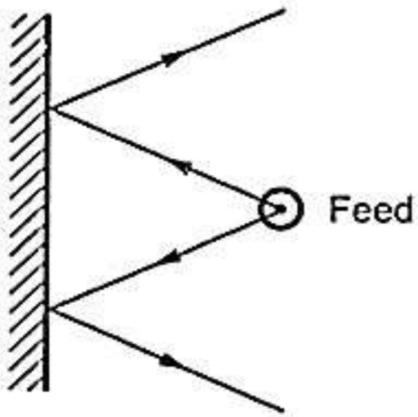
Introduction to Reflector Antennas 2

- Using reflectors, the radiation pattern of a radiating antenna can be modified.
- By using a large, metallic plane sheet as a reflector, the backward radiations from the antenna can be eliminated thus improving radiation pattern of an antenna.
- Thus for an antenna, desired radiation characteristics can be produced with the help of a large, suitably illuminated and suitably sized and shaped reflector surface.
- Some of the common reflectors are plane reflector, corner reflector and curved reflector.

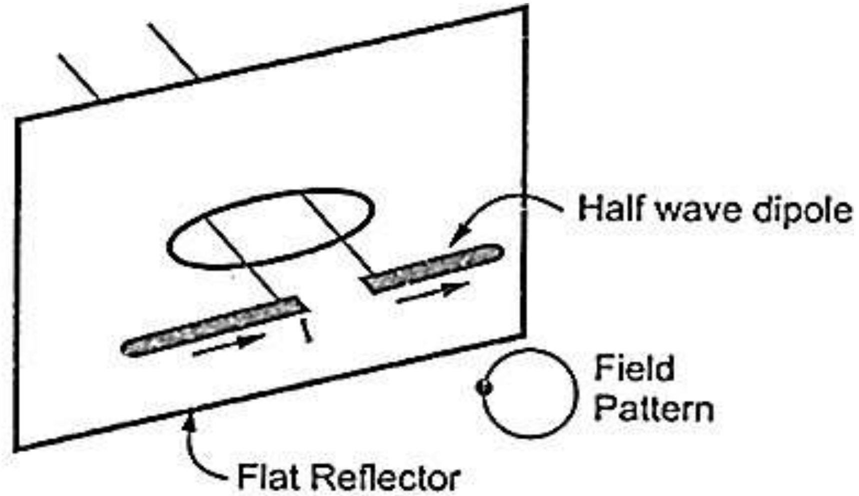
Type 1 / Flat or Plane Reflector Antennas

- The plane reflector is the simplest form of the reflector antenna. When the plane reflector is kept in front of the feed, the energy is radiated in the desired direction. The plane reflector is as shown in Fig. 2-16.

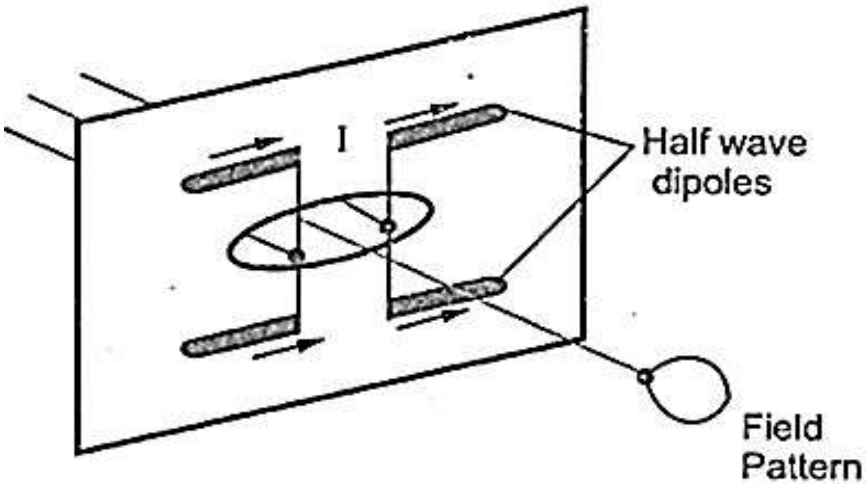
Plane Reflector Antennas



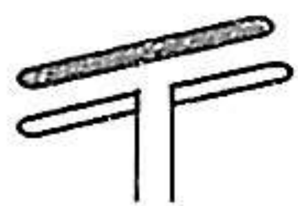
(a) Plane reflector



(b) Half wave dipole with plane reflector



(c) Half wave dipole array with plane reflector



(d) Half wave dipole with reflector element

- To increase the directivity of the antenna, a large flat sheet can be kept as plane reflector in front of a half dipole as shown in Fig. 2-16 (b).
- The main advantage of the plane reflector is that for the dipole backward radiations are reduced and the gain in the forward direction increases.
- To increase directivity further, we can use array of two half wave dipoles in front of a flat plane reflector as shown in Fig. 2-16 (c).
- It is observed that the flat sheet is less frequency sensitive than the thin element. Hence only a reflector element can be used to increase directivity. Such arrangement is shown in Fig. 2-16 (d).

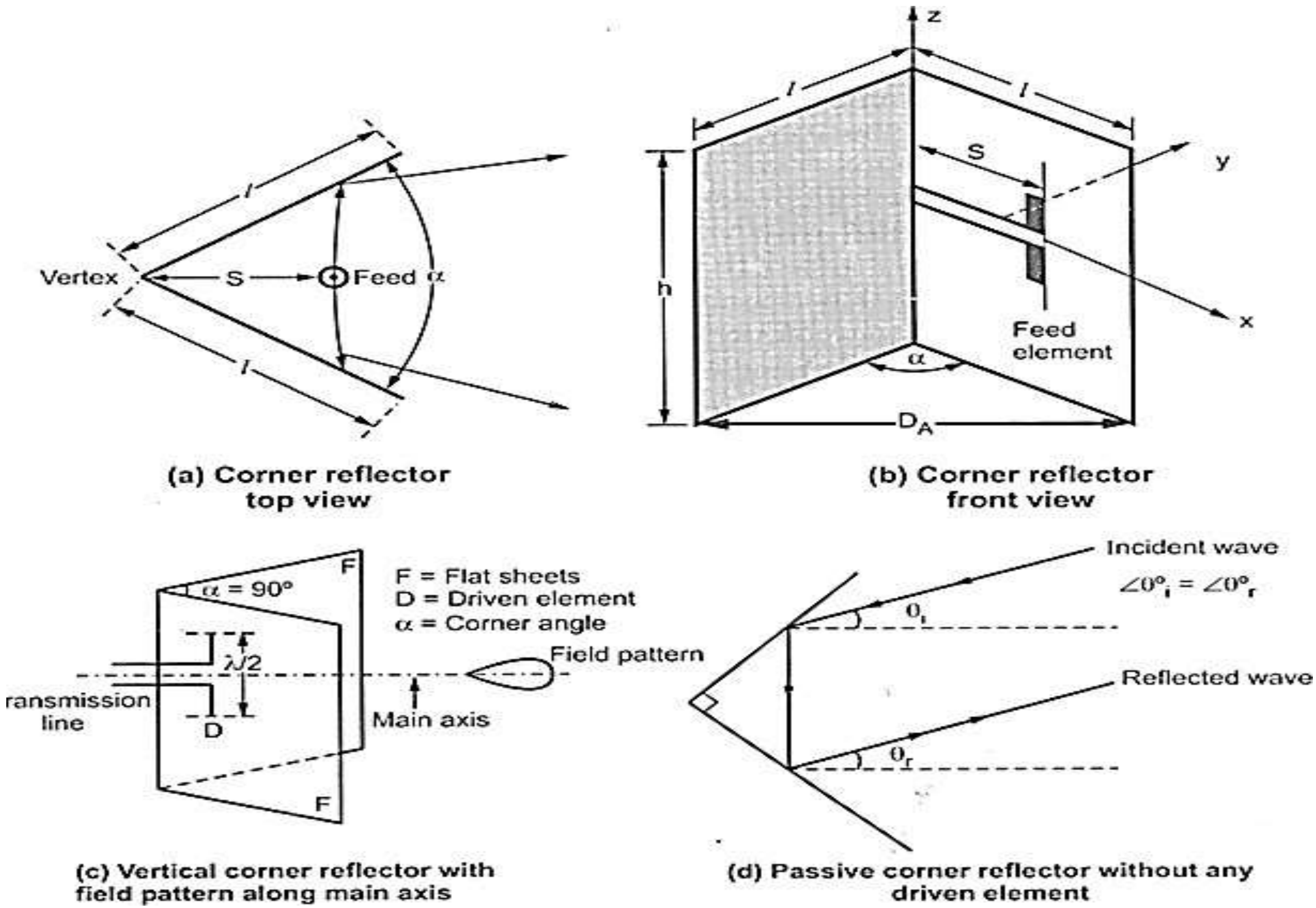
- In case of the plane reflectors, the polarization of the radiating source and its position with respect to the reflector both are important as one can control radiating properties of the overall antenna such as radiation pattern, directivity, impedance etc.
- Image theory has been used to analyze the radiation characteristics of such a antenna

Type 2 / Corner Reflector Antennas

- The disadvantage of the plane reflector is - radiation in back and side directions.
- In order to overcome this limitation, the shape of the plane reflector is modified so that the radiation is in forward direction only.
- The modified arrangement consists of two plane reflector joined to form a corner with some angle.
- The reflector is thus known as corner reflector.
- The angle at which two plane reflectors joined is called included angle (α).
- In most of the practical applications, the included angle is 90°.

- In some other applications angles other than 90° are also used.
- A typical corner reflector is shown in Fig. 2-17. The top view of the corner reflector is shown in Fig. 2-17 (a).
- The Fig. 2-17 (b) indicates the front view of the corner reflector.
- The vertical corner reflector with field pattern along main axis is shown in Fig. 2-17 (c).

Corner Reflector Antenna



Physical Arrangement of Corner Reflector

- When two flat reflecting sheets intersect each other at corner or at an angle, a directional antenna called corner reflector.
- When the corner angle is 90°, the corner reflector is called square corner reflector.
- Corner reflectors with included angle less than 90° are not advantageous.
- When included angle tends to 180° a flat sheet reflector obtained - the limiting condition of corner reflector.
- The analysis of the corner reflector is carried out when two intersecting planes are perfectly conducting & infinite.

Dimensions of Corner Reflector

- In most corner reflectors, the feed element is either a dipole or array of collinear dipoles placed parallel to the vertex at a distance S as shown in Fig. 2-17 (b).
- To increase the bandwidth, instead of thin wires as feed element, the biconical or cylindrical dipoles are preferred.
- For mathematical analysis, the dimensions specified are aperture of corner reflector (D_A) length of the reflector (l) and height (h).
- Generally the dimension of the aperture of the reflector (D_A) is selected between one and two wavelengths ($\lambda < D_A < 2\lambda$).

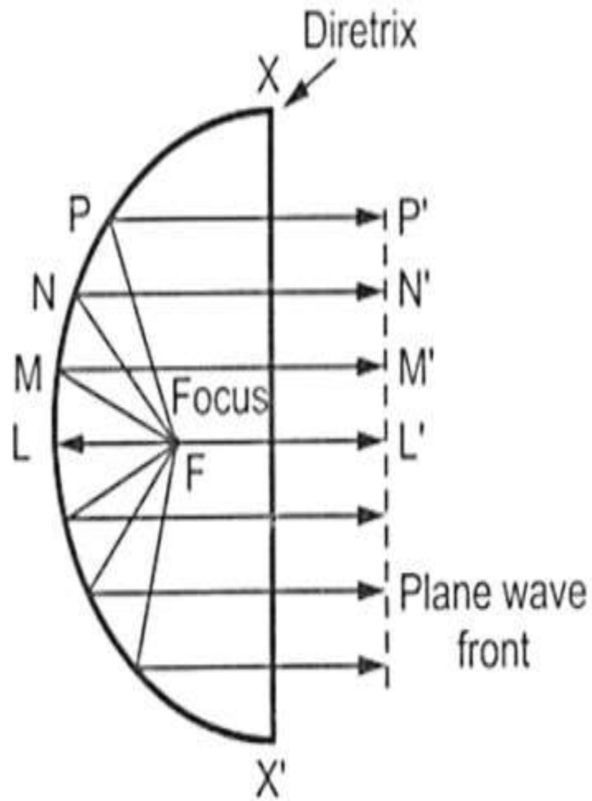
- The spacing between the vertex of the reflector and the feed element is selected as a fraction of wavelength ($\lambda/2 < S < 2\lambda$).
- The length of the reflectors is typically selected as twice the spacing between feed and vertex ($i.\ e.\ l \approx 2S$) for the included angle of 90°.
- The radiation resistance is the function of spacing between the feed and the reflector.
- If the spacing is too large, the unwanted multiple lobes are produced and hence the directivity is lost. If the spacing is very small, the radiation resistance decreases.

- The losses in the system increase as the decreased radiation resistance becomes comparable with the loss resistance of the antenna. Thus antenna is treated as inefficient antenna.
- The height of the reflector (h) is selected about 1.2 to 1.5 times greater than total length of the feed element.
- A corner reflector with two flat conducting sheets at a corner angle α and a driven antenna is called *active corner reflector antenna*.
- If the corner reflector antenna consists only two flat conducting sheet at a corner angle α without any driven element is called *passive corner reflector antenna*.

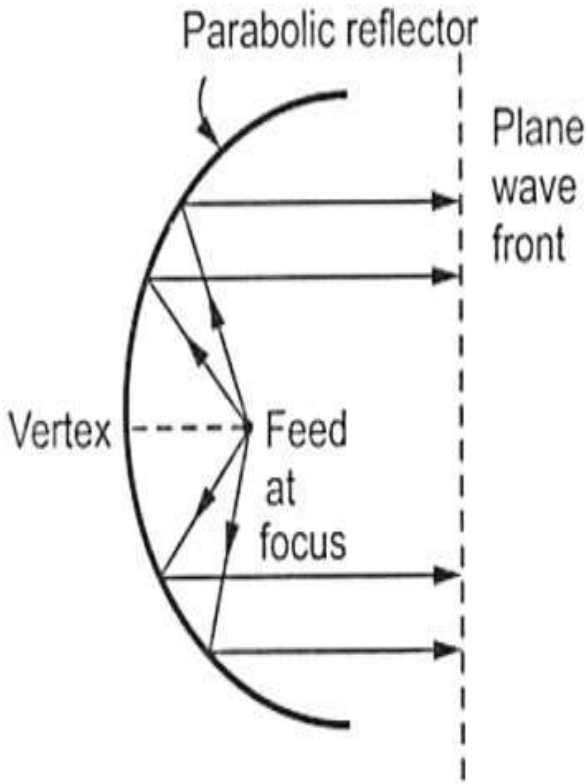
Type 3 / Parabolic Reflector

- To improve the overall radiation characteristics of the reflector antenna, the parabolic structure used.
- A parabola is a locus of a point which moves in such a way that the distance of the point from fixed point called focus plus the distance from the straight line called directrix is constant shown in Fig. 2-18.
- By definition ; $FM + MM' = FN + NN' = FP + PP' = constant$;
- When the point source is placed at the focal point, then the rays reflected by the parabolic reflector form parallel wave front as shown in Fig. 2-18 (b).
- This principle is used in the transmitting antenna.

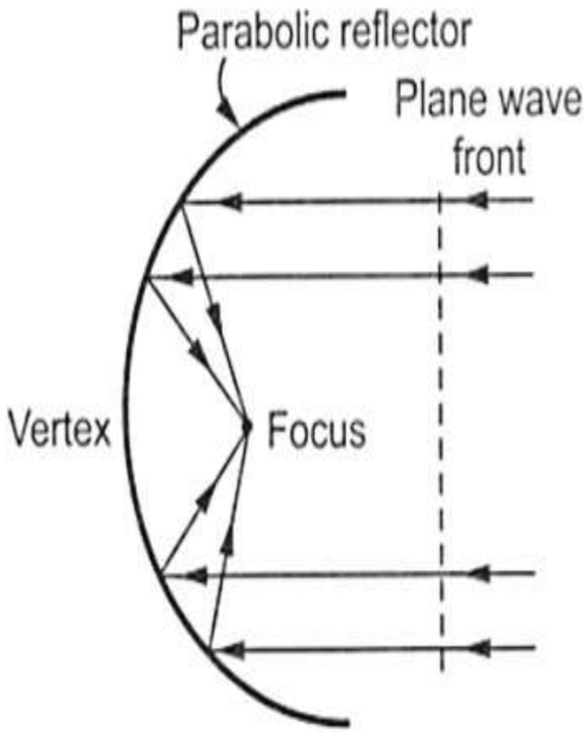
Parabolic Reflector



(a) Geometry of parabola



(b) Parabolic reflector at transmitting end



(c) Parabolic reflector at receiving end

Working of Parabolic Reflector

- When the beam of parallel rays incident on a parabolic reflector, then the radiations focus at a focal point shown in Fig. 2-18 (c). This principle is used in the receiving antenna.
- Consider a parabolic reflector shown in Fig. 2-18 (b).
- When point source is kept at the focal point of the parabola, the radiations striking the reflector are reflected parallel to the axis of parabola irrespective of the striking angle.
- That means the rays reflected by the parabolic reflector travel same distance to reach near the mouth of the reflector.

- The open end of the parabolic reflector is called aperture.
- The time taken by the reflected rays to travel a distance up to the directrix of the parabola is same. That means all the reflected rays are in phase.
- Thus the wave front at the aperture of the parabolic reflector is uniform phase front & very strong and concentrated beam obtained along the axis.
- Thus parabolic reflector is the most effective microwave antenna which produces concentrated radiation beam along the axis of parabola.
- The power gain of the paraboloid is a function of ratio between diameter of aperture.

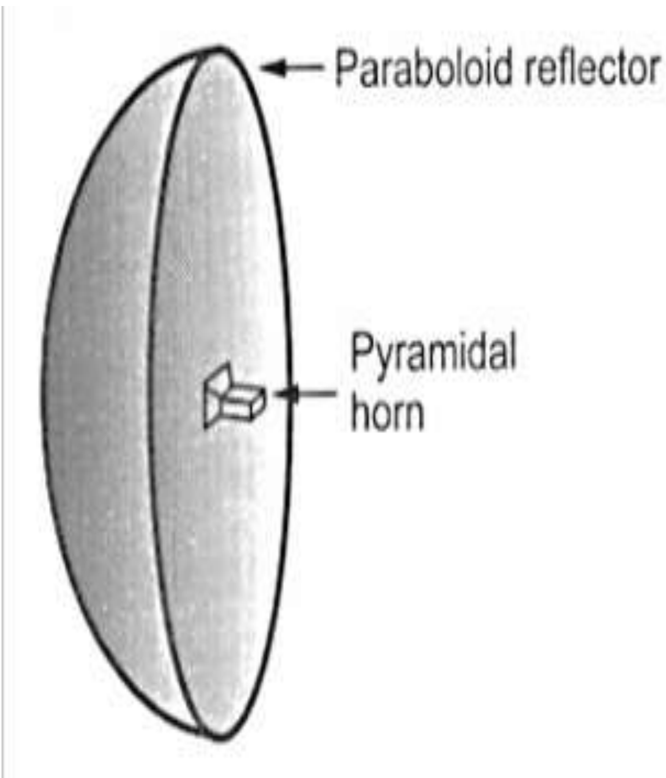
Topic 8

Reflector Antennas – Paraboloidal Reflector or Paraboloid or Microwave Dish Antenna

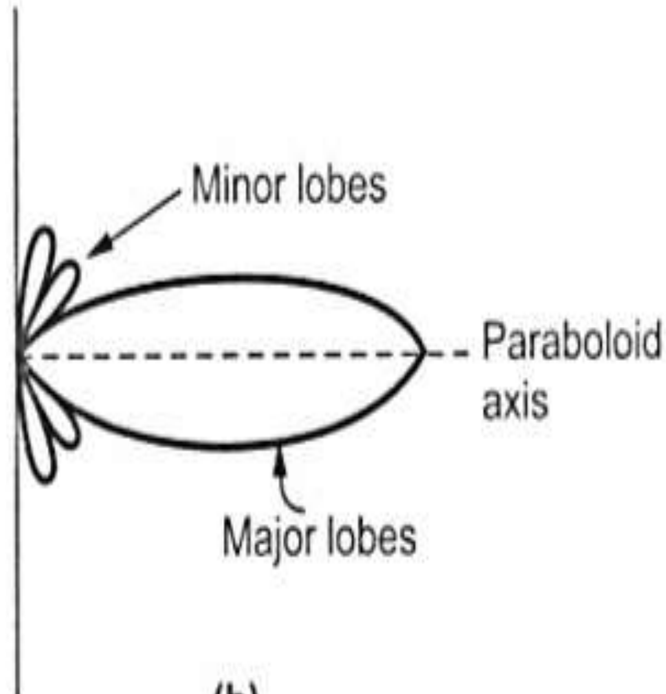
Introduction to Paraboloidal Reflector

- The parabolic reflector is a two dimensional structure. In practical applications, a three dimensional structure of the parabolic reflector is used.
- The three dimensional structure of the parabolic reflector can be obtained by rotating the parabola around its axis and it is called paraboloid.
- The paraboloid is as shown in Fig. 1 (a). The radiation pattern of the paraboloid is as shown in Fig. 1 (b). As the mouth of the paraboloid is circular in shape, the parallel, beam produced are of the circular cross-section.
- The radiation pattern consists very sharp major lobe and smaller minor lobes.

Fig 1 / Paraboloid with Pyramidal Horn as Feed



(a) Paraboloid



(b)

Power Gain of Paraboloid

- Consider that the power gain of the paraboloid, with circular mouth or aperture, with respect to half wave dipole is given by,

$$G = \frac{4\pi A_0}{\lambda^2}$$

- Here A_0 is the capture area which is less than the actual area A_e of the mouth and it is given by,

$$A_0 = k.A_e$$

- where k = constant dependent on feed antenna used. It is 0.65 for dipole.
- The actual area of circular aperture with diameter d is given by

$$A_e = \pi\left(\frac{d}{2}\right)^2 = \frac{\pi d^2}{4}$$

Power Gain of Paraboloid 2

- Hence the power gain given by –

$$G = \frac{4\pi(kA_e)}{\lambda^2} = \frac{4\pi \times 0.65\ A_e}{\lambda^2} = 6\left(\frac{d}{\lambda}\right)^2$$

- From the above equation the power gain of the paraboloid depends on the ratio or diameter d of the circular aperture to the wavelength in free space. The ratio d/λ is called aperture ratio of the paraboloid.
- Hence the effective radiated power (ERP) is the product of the input power fed and the power gain G.
- With small diameter of the paraboloid, the gain of the paraboloid is extremely large when λ is small in microwave frequency range.
- For lower frequencies, λ is large, the diameter of the circular aperture becomes too large – hence use of parabolic reflectors are avoided at lower frequency.

Spillover / Back Lobes

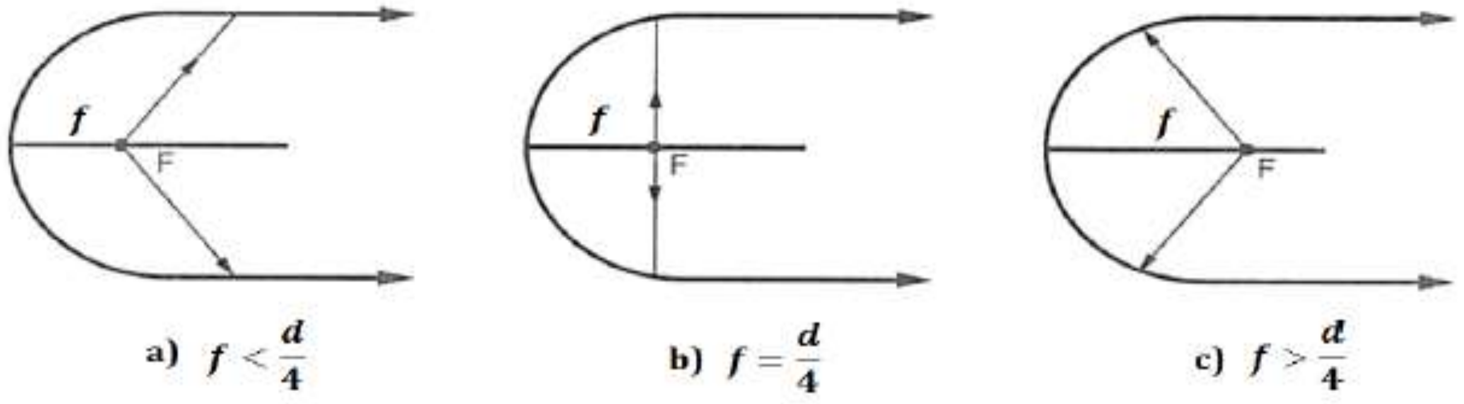
- In paraboloid reflector, the ratio of the focal length f to the diameter of aperture is another important design constraint.
- The paraboloid can be designed to obtain pencil shape radiation beam by keeping the diameter of the aperture fixed and changing the focal length f.
- The three possible cases are as follows ;
  - Focal point inside the aperture of paraboloid.
  - Focal point along the plane of open mouth of paraboloid.
  - Focal point beyond the open mouth of paraboloid
- CASE 1 : When the focal length is very small, the focal point lies inside the open mouth of paraboloid as shown in Fig. 2 (a). It is very difficult to obtain uniform illumination over a wide angle.

Spillover / Back Lobes 2

- Case 2 : When the focal point lies on the plane of the open mouth of the paraboloid by the geometry, the focal length f is one fourth of the open mouth diameter d. This condition gives maximum gain pencil shaped radiation equal in horizontal and vertical plane as shown in Fig. 2 (b).
- Case 3: When the focal length is too large, the focal point lies beyond the open mouth of the paraboloid as shown in Fig. 2 (c) - difficult to direct all the radiations from the source on the reflector.
  - For practical applications, the value of the focal length to diameter ratio lies between 0.25 to 0.5.

Spillover / Back Lobes 3

- Some of the desired rays are not fully captured by reflector, such non-captured rays form *spill over*. The noise pick up increases with spill-over.
- Few radiations originated from the primary radiators are observed in forward direction - such radiations get added with desired parallel beam. This is *called back lobe radiation. They* are unwanted as they affect the reflected beam.

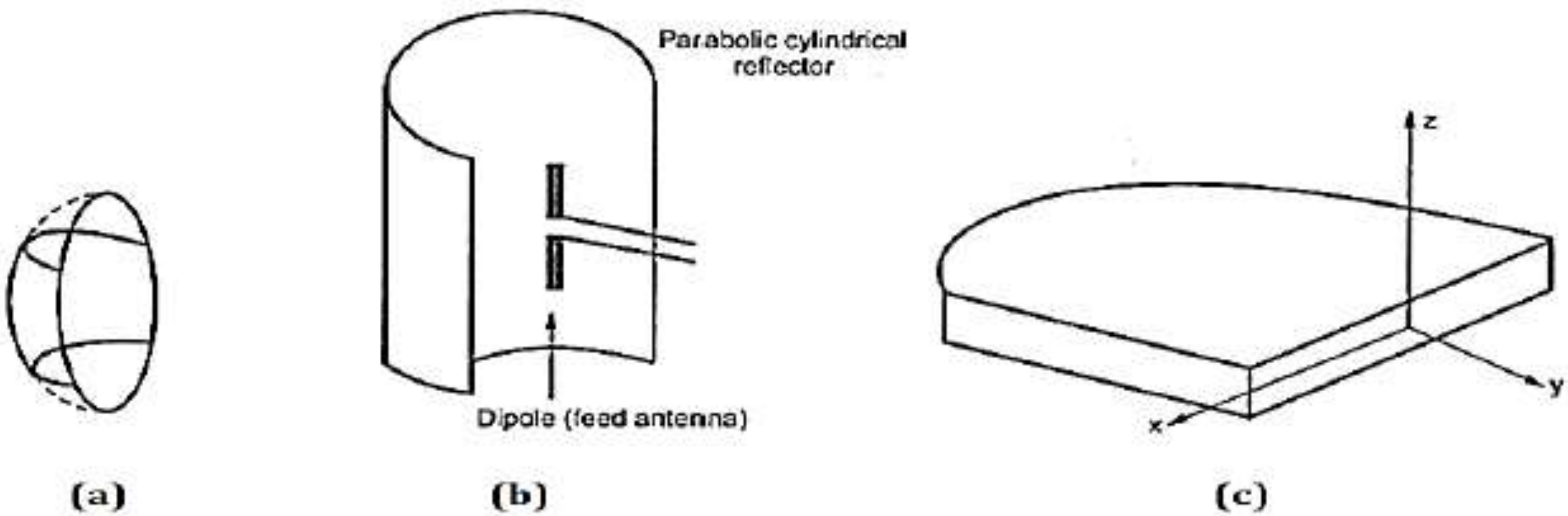


Types of Paraboloid Reflectors

- **Truncated paraboloid :** This type formed by cutting some of the portion of the paraboloid to meet the requirements. As the portion of the paraboloid is cut away or truncated as shown in Fig. 3 (a) , the paraboloid is called cut paraboloid or truncated paraboloid.
- **Parabolic right cylinder:** The right cylindrical structure of the parabolic reflector shown in Fig. 3 (b). This structure is obtained by moving the parabola side ways. This parabolic structure has focal line instead of a focal point and a vertex line instead of a vertex. In parabolic right cylinder reflector the energy is collimated at a line parallel to the axis through the focal point of the reflector.

Types of Paraboloid Reflectors 2

- **Pill box or cheese antenna:** The cheese antenna or pill box is a short parabolic right cylinder enclosed by parallel plates as shown in Fig. 3 (c). This antenna is useful in producing wide beam in one of the planes while a narrow in other.



Topic 9

Feed Structures for Paraboloidal Reflector (Reflector Antennas)

Introduction to Feed Systems

- A parabolic reflector antenna system consists of two basic parts - a source of radiation focus and a reflector. The source placed at the focus is called primary radiator, while the reflector is called secondary radiator. The primary radiator is commonly called feed radiator or simply feed.
- In case of a parabolic reflector a feed radiates entire energy towards the reflector illuminating the entire surface of reflector & no energy radiated in any unwanted direction.
- Practically there are number of possible feeds to the parabolic reflector antenna. The secondary radiator used is most of the times a paraboloid.

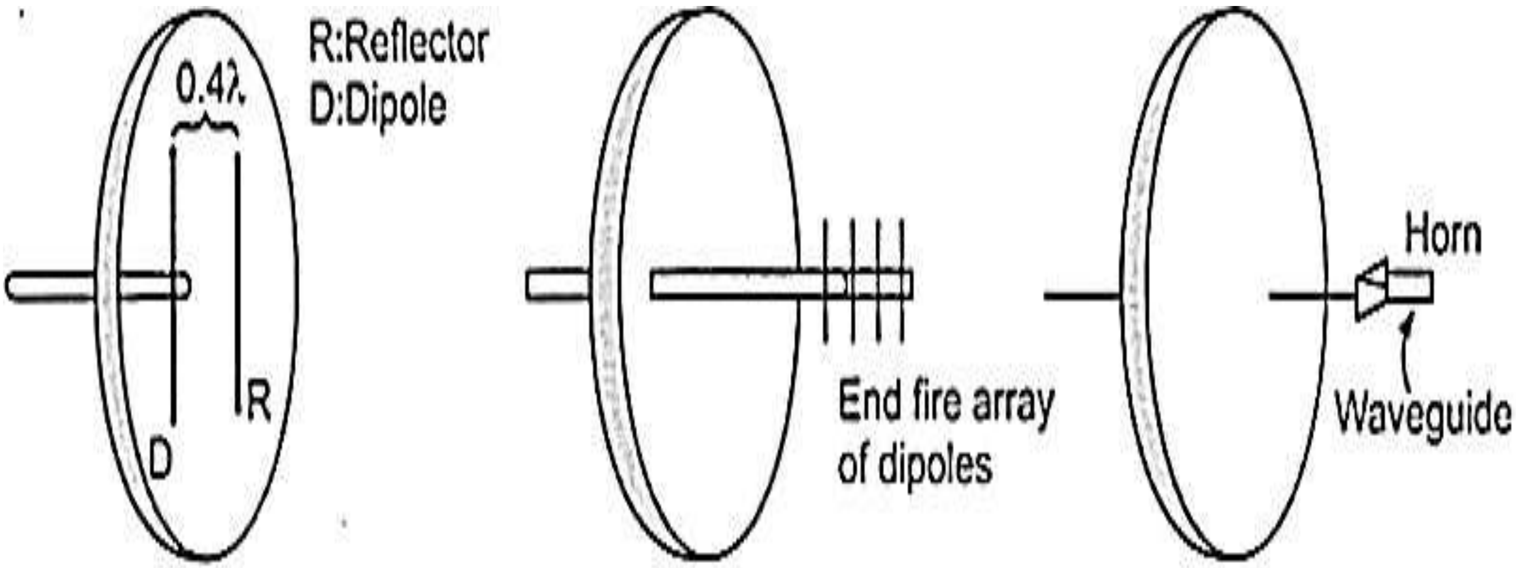
General Feed Systems

- Case 1 : The simplest type of feed used is a dipole antenna. But it is not a suitable feed for the parabolic reflector antenna. Instead of only dipole, a feed consisting dipole with parasitic reflectors can be used as a feed system.
- In such cases, the spacing between the dipole as a driven element and the parasitic reflector is 0.125 λ . In some cases a dipole along with a plane reflector spaced 0.4 λ apart from the dipole is used. It is shown in Fig. 1 (a).
- Case 2: An end fire array of dipoles is used as feed radiator as shown in Fig. 1 (b). The dipoles are spaced in such a way that the end fire pattern of an array illuminates reflector.

General Feed Systems 2

- Case 3 : The most widely used feed system in the parabolic reflector antenna is horn antenna as shown in Fig. 1 (c). The horn antenna is fed with a waveguide. If circular polarization is required, then in place of a rectangular horn, a conical horn or helix antenna is used at the focus.
- In all three cases, the feed is placed at the focus to obtain maximum beam pattern. If the feed is moved along a line perpendicular to the main axis then beam deteriorates. But if the feed is moved along the main axis, then the beam gets broadened.
- Hence focus is the important point on the main axis at which the feed is placed to obtain maximized beam pattern.

Fig 1 / General Feed Systems



(a) Dipole with plane reflector

(b) End fire array of dipole

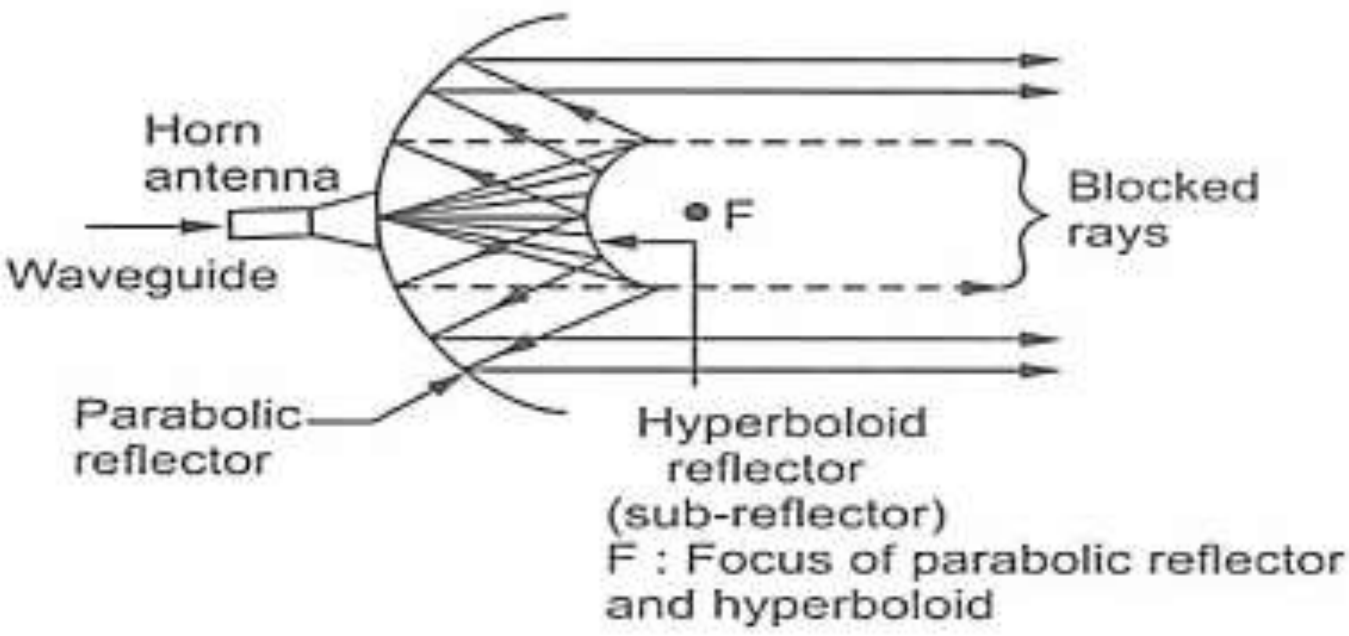
(c) Horn with waveguide

Case 4 - Cassegrain Feed

- This system of feeding paraboloid reflector named after a mathematician Prof. Cassegrain. In all the feed systems, the feed is located at the focus.
- But in Cassegrain feed system, the feed radiator is placed at the vertex of the parabolic reflector instead of placing it at the focus.
- This system uses a hyperboloid reflector, such that its one of the foci coincides with the focus of the parabolic reflector. This hyperboloid reflector is called Cassegrain secondary reflector.
- The primary radiator or feed radiator used is generally a horn antenna.

Cassegrain Feed 2

- The radiation emitted from primary feed radiator reach sub-reflector. The sub reflector reflects and illuminates the main parabolic reflector. The main reflector reflects the rays parallel to the axis. The geometry of the Cassegrain feed system is as shown in Fig. 2.

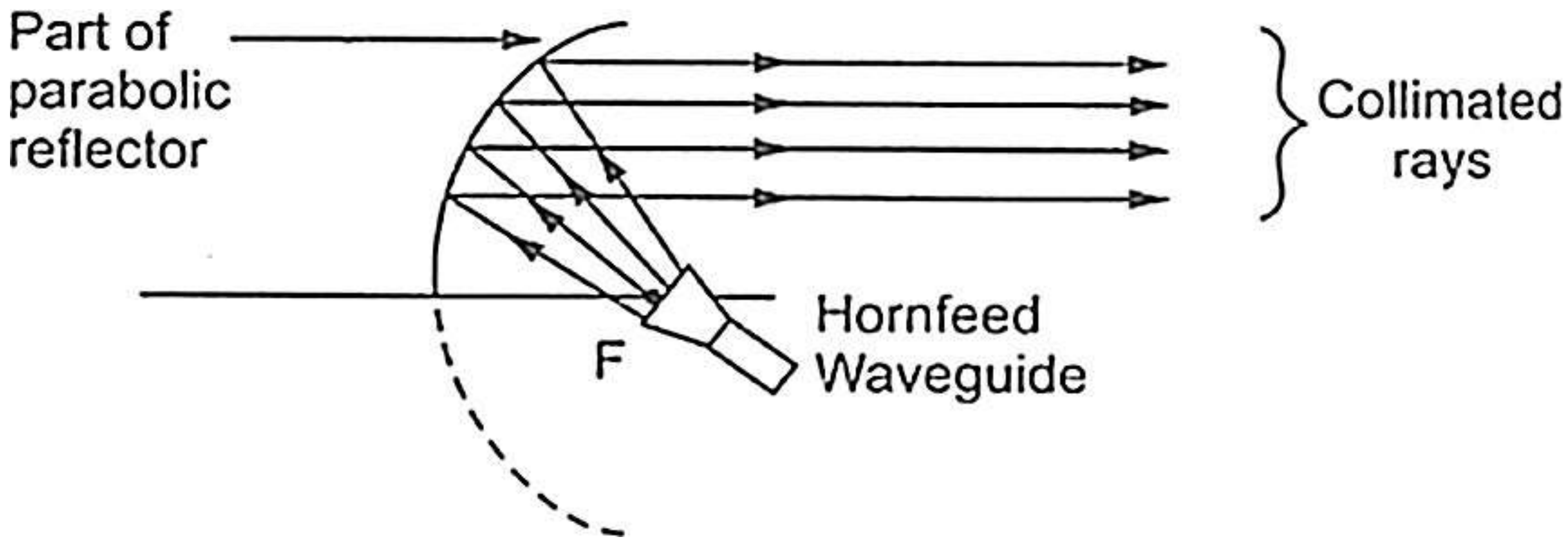


Advantages & Disadvantages of Cassegrain Feed

- It reduces the spill over & minor lobe radiations.
- Focal length greater than the physical focal length achieved.
- Ability to place a feed at convenient place.
- Beam can be broadened by adjusting one of the reflector surfaces.
- Some of the radiation from the parabolic reflector obstructed or blocked by the hyperboloid reflector creating region of blocked rays. It is called ***aperture blockage***.
- For small dimension parabolic reflector it is the main drawback of the Cassegrain feed system.

Case 5 – Offset Feed System

- To overcome the aperture blocking effect due to the dependence of the secondary reflector dimensions on the distance between feed and sub-reflector, the offset feed system as shown in Fig. 3 is used.
- Here feed radiator is placed at the focus.
- With this system all the rays are properly collimated without formation of the region of blocked rays.



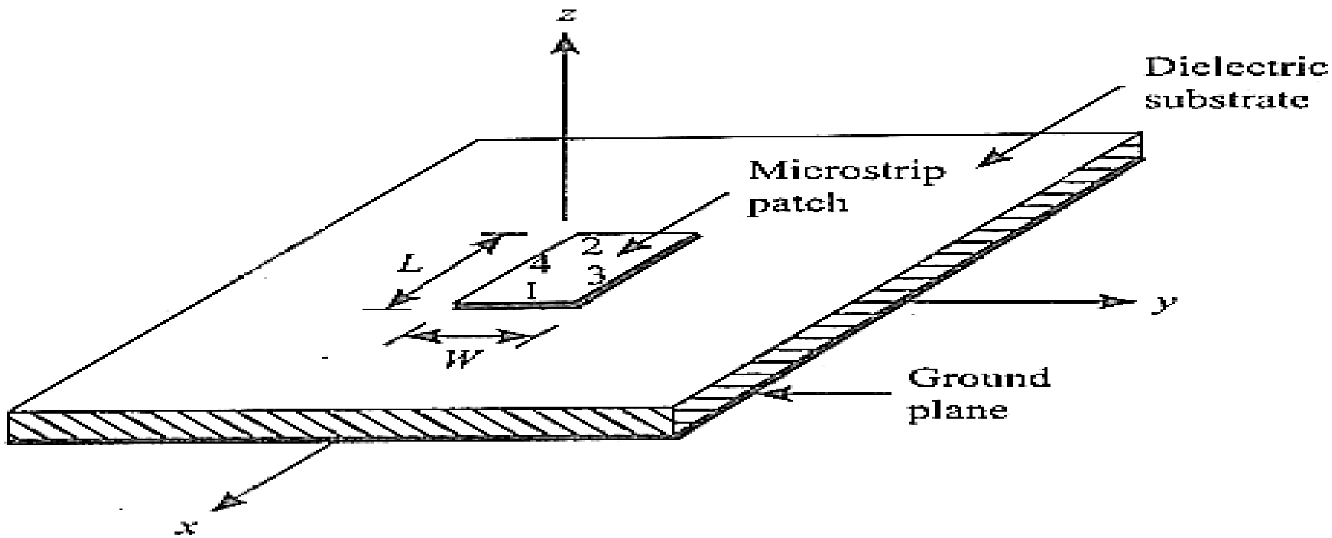
Topic 10

Microstrip Patch Antennas

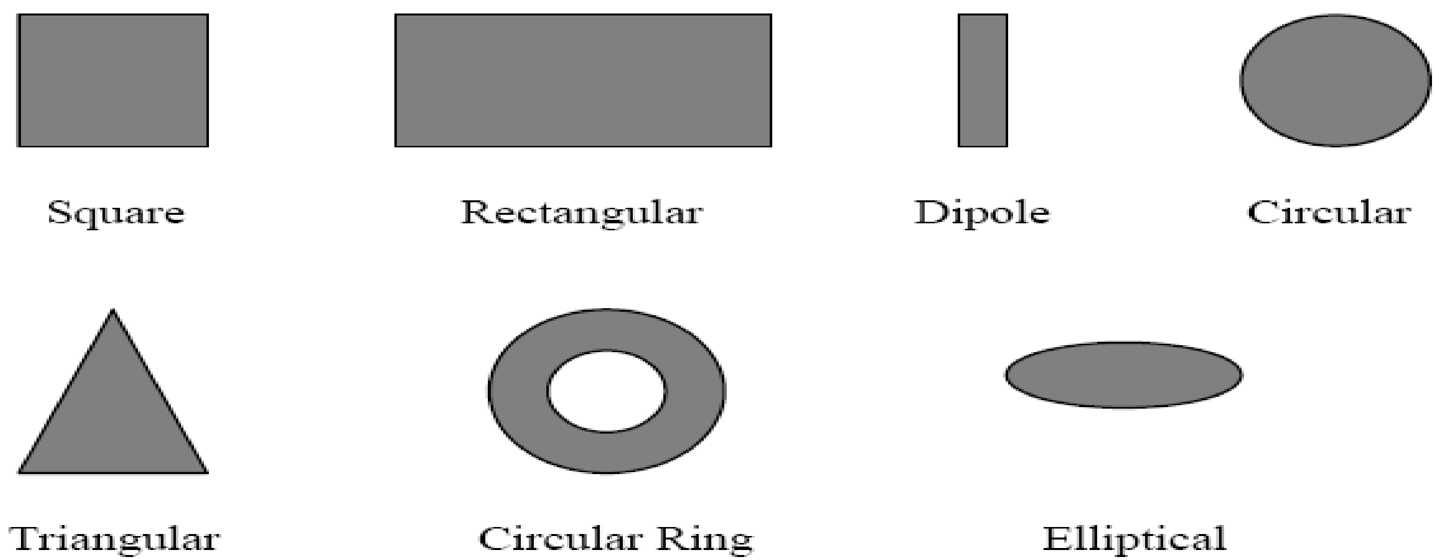
Introduction to Microstrip Patch Antenna

- Microstrip patch antenna consists of a radiating patch on one side of a dielectric substrate which has a ground plane on the other side as shown in Fig. 1.
- The patch is generally made of conducting material such as copper or gold and can take any possible shape.
- The radiating patch and the feed lines are usually photo etched on the dielectric substrate.
- In order to simplify analysis and performance prediction, the patch is generally square, rectangular, circular, triangular, elliptical or some other common shape as shown in Fig. 2.

Structures & Shapes of Microstrip



1 and 2: Radiating edges
3 and 4: Non-radiating edges
$L = \lambda_g/2$



Dimensions of Patch Antenna

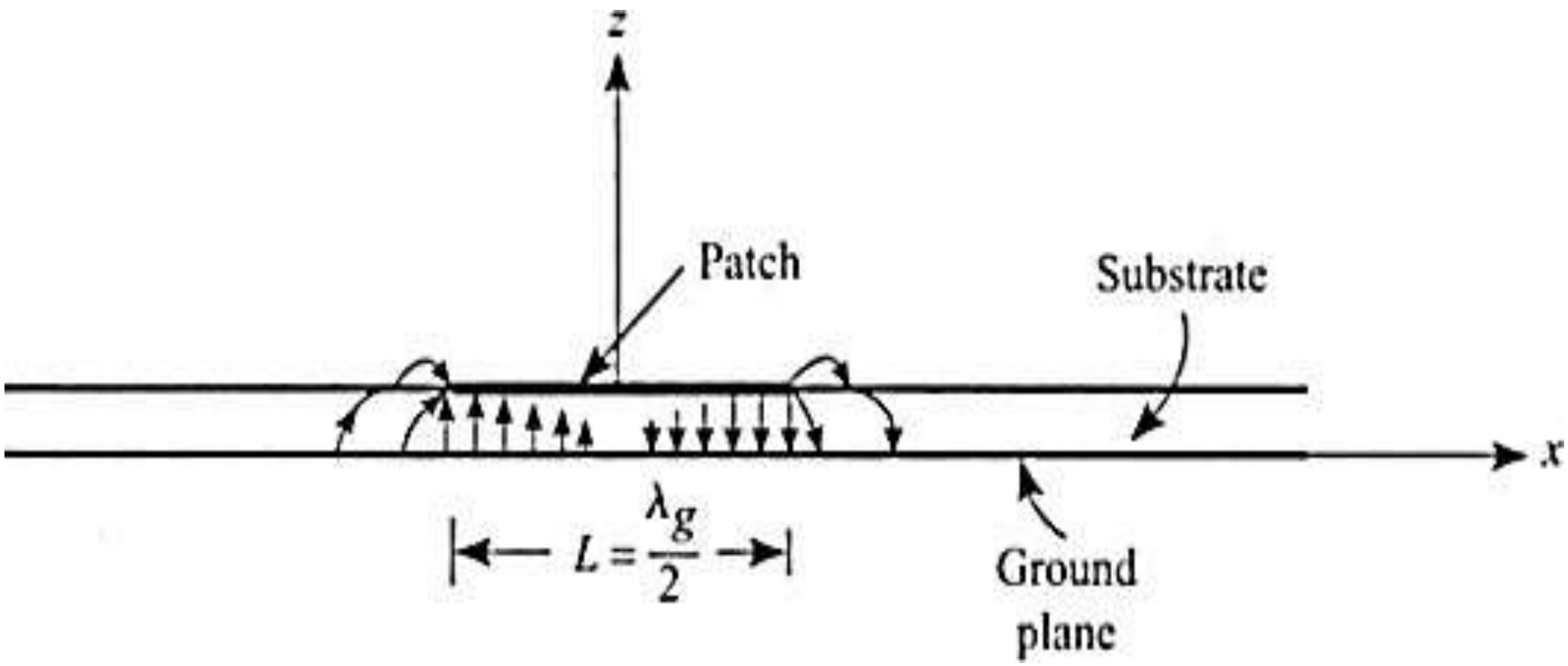
- For a rectangular patch, the length L of the patch is $0.333\lambda_0 \le L \le 0.5\lambda_0$, where λ_0 is free-space wavelength. The patch is selected to be very thin such that $t << \lambda_0$ (where t is the patch thickness). The height h of the dielectric substrate is $0.033\lambda_0 \le h \le 0.05\lambda_0$. The dielectric constant of the substrate (ε_r) is typically in the range $2.2 \le \varepsilon_r \le 12$.
- Microstrip patch antennas radiate primarily because of the fringing fields between the patch edge and the ground plane.
- For good antenna performance, a thick dielectric substrate having a low dielectric constant is desirable since this provides better efficiency, larger bandwidth and better radiation.

Radiation Mechanism of Patch Antenna

- Consider a rectangular patch of length L and width W printed on a dielectric substrate of height *h*. The length of the patch is chosen as λ_g /2, *where* λ_g is the guide wavelength of microstrip line of width W printed on the same dielectric substrate.
- The electric field along x-direction undergoes a 180° phase reversal as in Fig. 3 from one edge to the other.
- It can be shown that the fields near edges 1 and 2 add up producing the radiation with a maximum along the z-direction.

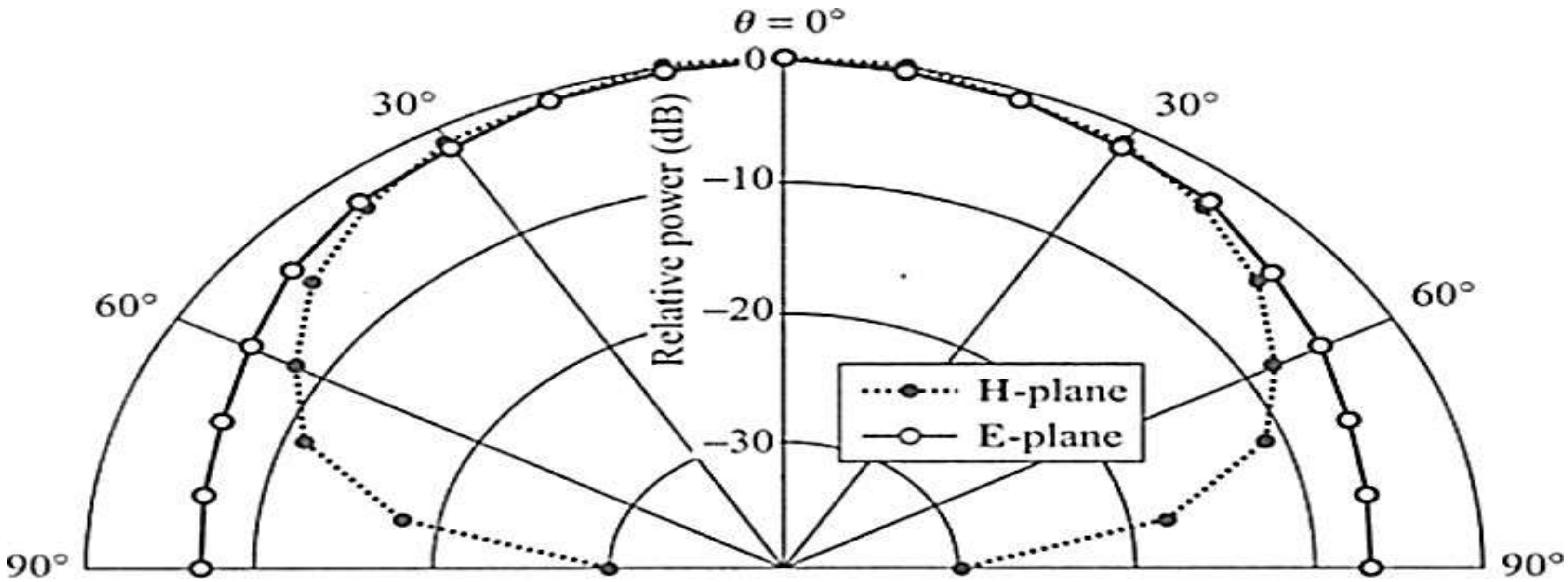
Radiation Mechanism of Patch Antenna 2

- Hence, edges 1 and 2 are known as the radiating edges.
- Further, the fields near edges 3 and 4 do not contribute to the radiation. The rectangular microstrip patch shown in Fig. 4 radiates linearly polarized waves, with the electric field oriented along the x-direction when looking in the direction of maximum radiation.



Radiation Mechanism of Patch Antenna 3

- The radiation patterns in the two principal planes, viz., the E-plane (x-z plane) and the H-plane (y-z plane) are shown in Fig. 4. The pattern is very broad and has nulls along the y-direction.
- For efficient transfer of power from a transmission line to the patch antenna, the input impedance of the antenna matched to the characteristic impedance of the transmission line.



Advantages of Patch Antenna

- Light weight and low volume.
- Low profile planar configuration easily made conformal to host surface.
- Low fabrication cost.
- Supports both linear as well as circular polarization.
- Can be easily integrated with microwave integrated circuits (MICs).
- Capable of dual and triple frequency operations.
- Mechanically robust when mounted on rigid surfaces.

Disadvantages / Applications of Patch Antenna

- Narrow bandwidth
- Low efficiency
- Low gain
- Extraneous radiation from feeds and junctions
- Poor end fire radiator except tapered slot antennas
- Low power handling capacity.
- Used for high-performance aircraft, spacecraft, satellite, and missile applications where size, weight, cost, performance, installation and aerodynamic profile are constraints.
- Other areas where microstrip antennas widely used are: GPS, Telemetry, Radars, Altimeters, etc.,

Topic 11

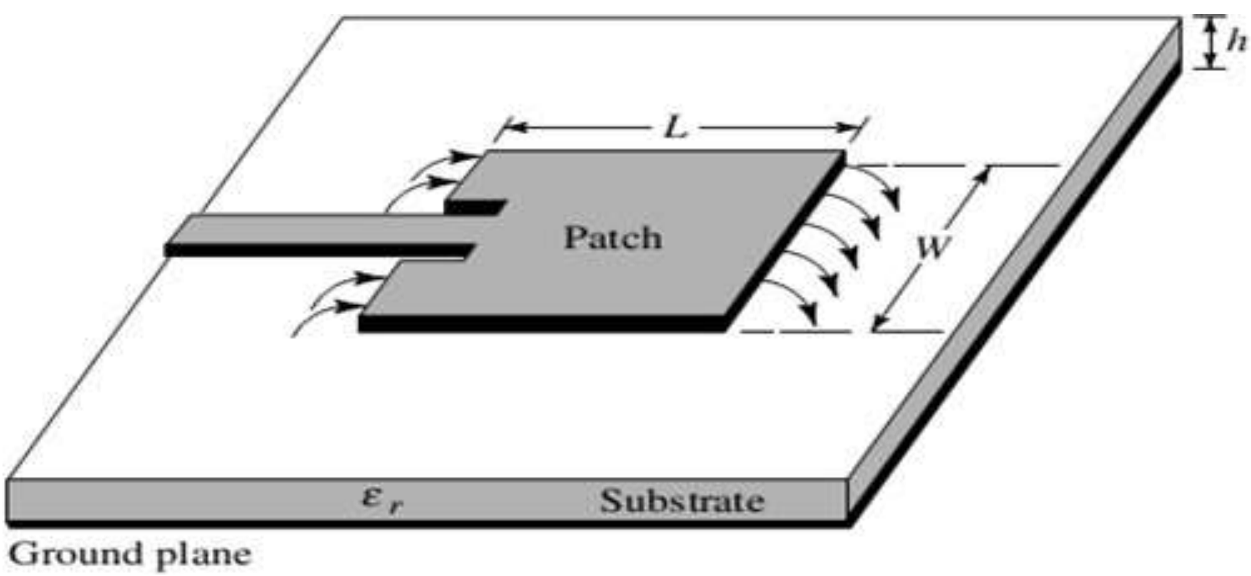
Feeding Structures for Microstrip Patch Antennas

Type 1 – Microstrip Line Feed

- In this type of feed technique, a conducting strip is connected directly to the edge of the microstrip patch as shown in Fig. 1.
- The conducting strip is smaller in width compared to the patch. It has the advantage that the feed can be etched on the same substrate to provide a planar structure.
- An inset cut can be incorporated into the patch to obtain good impedance matching without the need for any additional matching element.
- This is achieved by properly controlling the inset position. Hence this is an easy feeding technique & provides ease of fabrication and simplicity in modeling as well as impedance matching.

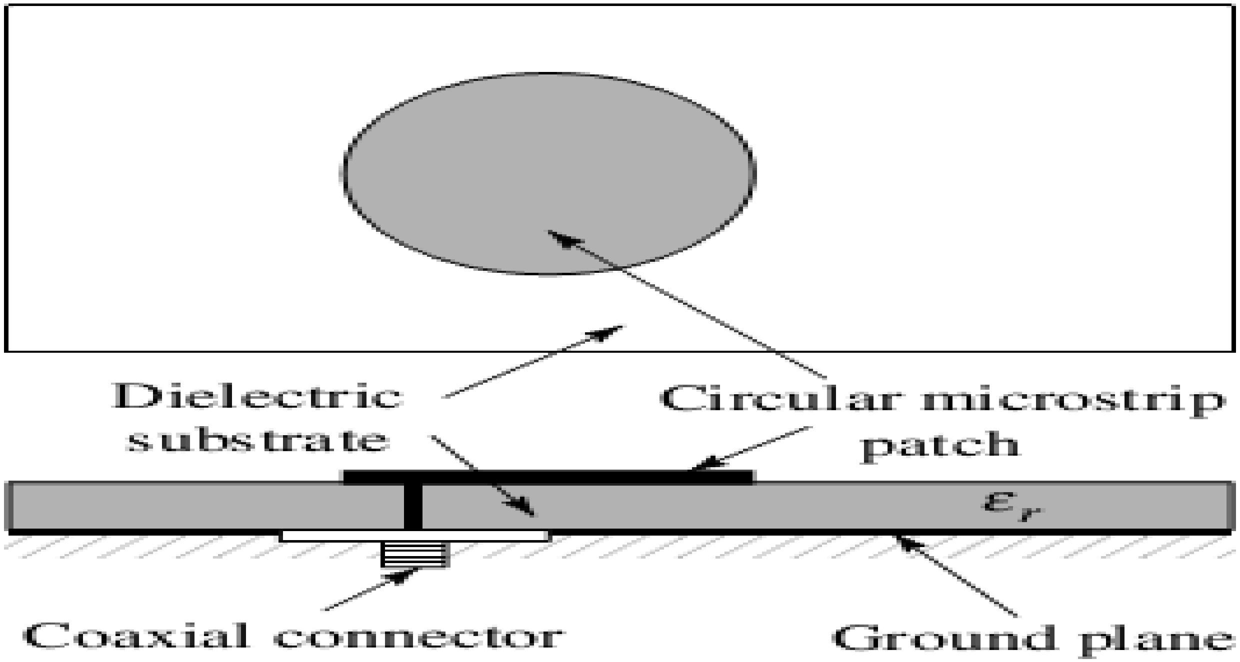
Type 1 – Microstrip Line Feed 2

- However as the thickness of the dielectric substrate increases, surface waves and spurious feed radiation also increases, which hampers the bandwidth of the antenna.
- This type of feeding technique results in undesirable cross polarization effects.



Type 2 – Coaxial Feed

- The Coaxial feed or probe feed is one of the most common techniques used for feeding microstrip patch antennas.
- From Fig. 2, the inner conductor of the coaxial connector extends through the dielectric and soldered to the radiating patch, while the outer conductor is connected to the ground plane.



Type 2 – Coaxial Feed 2

– The main advantage that feed can be placed at any desired position inside the patch in order to obtain impedance matching. This feed method is easy to fabricate and has low spurious radiation effects.

– The major disadvantage is that it provides narrow bandwidth and difficult to model since a hole has to be drilled into the substrate. The increased probe length makes the input impedance more inductive, leading to matching problems.

– By using a thick dielectric substrate to improve the bandwidth, the microstrip line feed and the coaxial feed suffer from numerous disadvantages such as spurious feed radiation and matching problem.

Type 3 – Aperture Coupled Feed

– In aperture coupling the radiating microstrip patch element is etched on the top of the antenna substrate & microstrip feed line is etched on the bottom of the feed substrate to obtain aperture coupling.

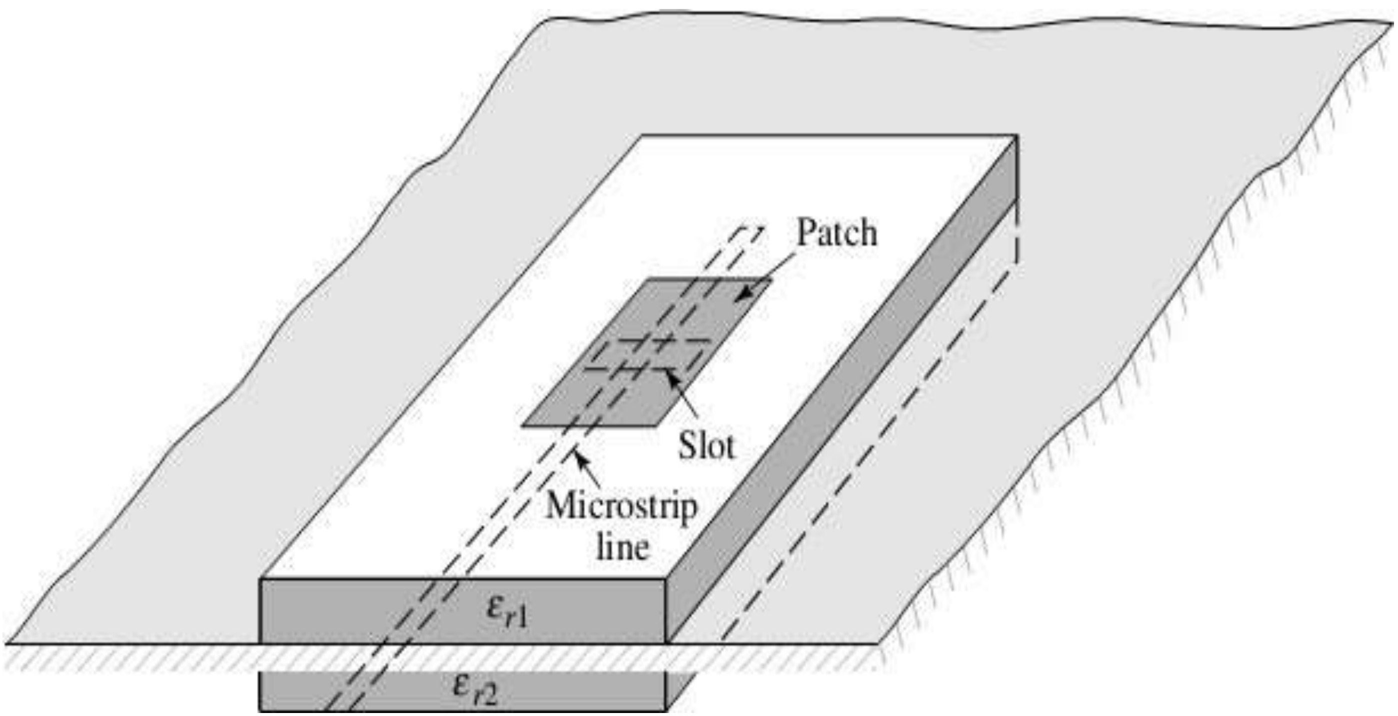
– The thickness and dielectric constants of these two substrates may thus be chosen independently to optimize the distinct electrical functions of radiation and circuitry.

– The coupling aperture is centered under the patch, leading to lower cross polarization due to symmetry of the configuration. The amount of coupling from the feed line to the patch is determined by the shape, size and location of the aperture.

– Since the ground plane separates the patch and the feed line, spurious radiation is minimized as shown in Fig. 3.

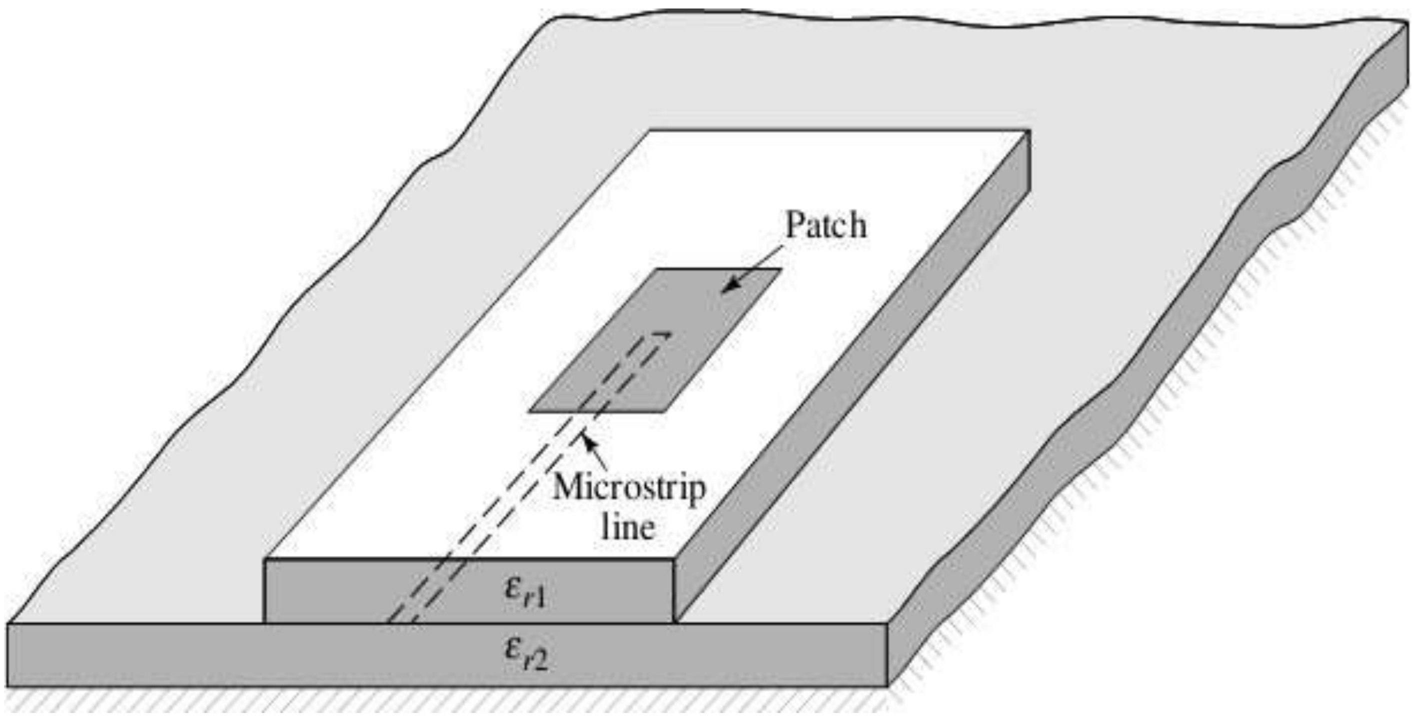
Type 3 – Aperture Coupled Feed 2

- This type of feeding technique can give very high bandwidth of about 21%. Also the effect of spurious radiation is very less as compared to other feed techniques. The major disadvantage is that difficult to fabricate due to multiple layers, which also increases the antenna thickness.



Type 4 - Proximity Coupled Feed

- This type of feed technique is also called as the electromagnetic coupling scheme. As shown in Fig. 4, two dielectric substrates are used such that the feed line is between the two substrates and the radiating patch is on top of the upper substrate.



Advantages & Disadvantages of Proximity Feed

- Eliminates spurious feed radiation and provides very high bandwidth of about 13%, due to increase in the electrical thickness of the microstrip patch antenna.
- Provides choices between two different dielectric media, one for the patch and one for the feed line to optimize the individual performances.
- Difficult to fabricate because of the two dielectric layers that need proper alignment. Also, there is an increase in the overall thickness of the antenna.

Comparison of Feeds – Patch Antenna

Characteristics	Microstrip Line Feed	Coaxial Feed	Aperture coupled Feed	Proximity coupled Feed
Spurious feed radiation	More	More	Less	Minimum
Reliability	Better	Poor due to soldering	Good	Good
Ease of fabrication	Easy	Soldering and drilling needed	Alignment required	Alignment required
Impedance Matching	Easy	Easy	Easy	Easy
Bandwidth (achieved with impedance matching)	2-5%	2-5%	2-5%	13%

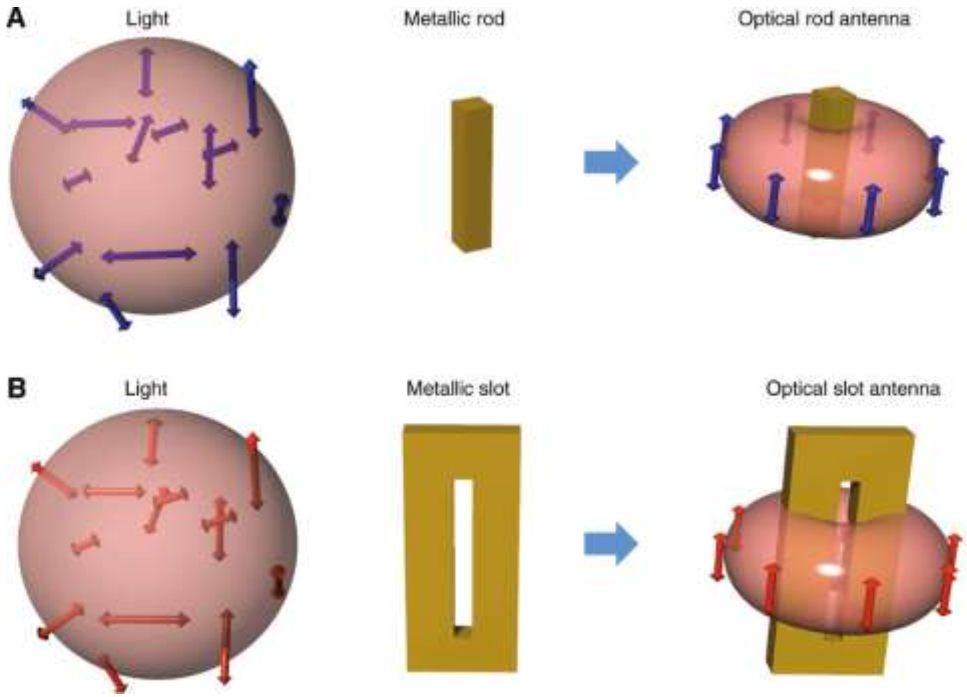
Topic 12

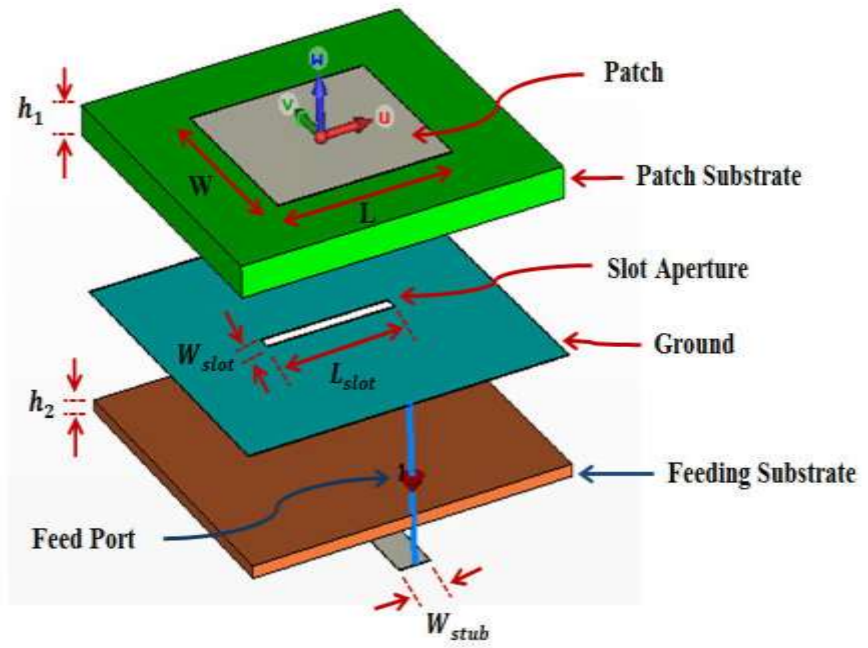
Numerical Tool for Antenna Analysis

Numerical Tool for Antenna Analysis

Software Name	Theoretical Model	Company
Ensemble (Designer)	Moment method	Ansoft
IE3D	Moment method	Zeland
Momentum	Moment method	HP
EM	Moment method	Sonnet
PiCasso	Moment method/genetic	EMAG
FEKO	Moment method	EMSS
PCAAD	Cavity model	Antenna Design Associates, Inc
Micropatch	Segmentation	Microstrip Designs, Inc.
Microwave Studio (MAFIA)	FDTD	CST
Fidelity	FDTD	Zeland
HFSS	Finite element	Ansoft

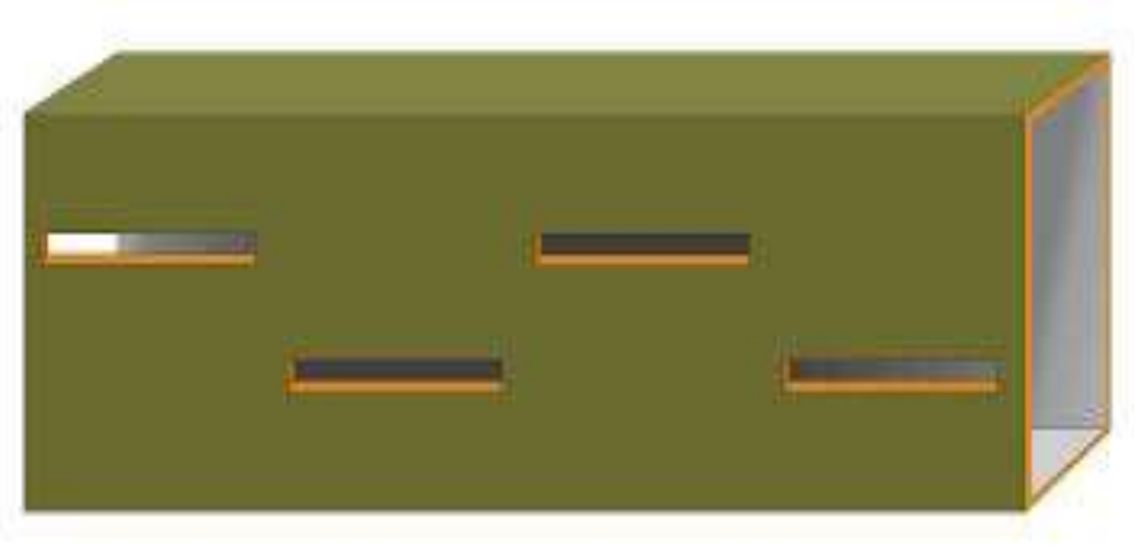
Applications





Applications

- **Slot Antenna**. **Slot** radiators or **slot antennas** are **antennas** that are used in the frequency range from about 300 MHz to 25 GHz. They are often used in navigation radar usually as an array fed by a waveguide. ... The polarization of a **slot antenna** is linear.



UNIT – IV

SPECIAL ANTENNAS AND ANTENNA MEASUREMENTS

Yagi-Uda Antenna

Introduction

- Yagi-Uda arrays or Yagi-Uda antennas are high gain antennas.
- The antenna was first invented by a Japanese Prof. S. Uda in early 1940's and described in English by Prof. H. Yagi.
- Hence the antenna name Yagi-Uda antenna was given after Prof. S. Uda and Prof. H. Yagi.
- A basic Yagi-Uda antenna consists a driven element, one reflector and one or more directors.
- Basically it is an array of one driven element and one of more parasitic elements.
- The driven element is a folded dipole made of a metallic rod which is excited.

- A Yagi-Uda antenna uses both the reflector (R) and the director (D) elements in same antenna.
- The element at the back side of the driven element is the reflector. It is of the larger length compared with remaining elements.
- The element in front of the driven element is the director which is of lowest length.
- Directors and reflector are called parasitic elements.
- All the elements are placed parallel and close to each other as shown in Fig. 1.
- The length of the folded dipole is about $\lambda/2$ and it is at resonance. Length of the director is less than $\lambda/2$ and length of the reflector is greater than $\lambda/2$.

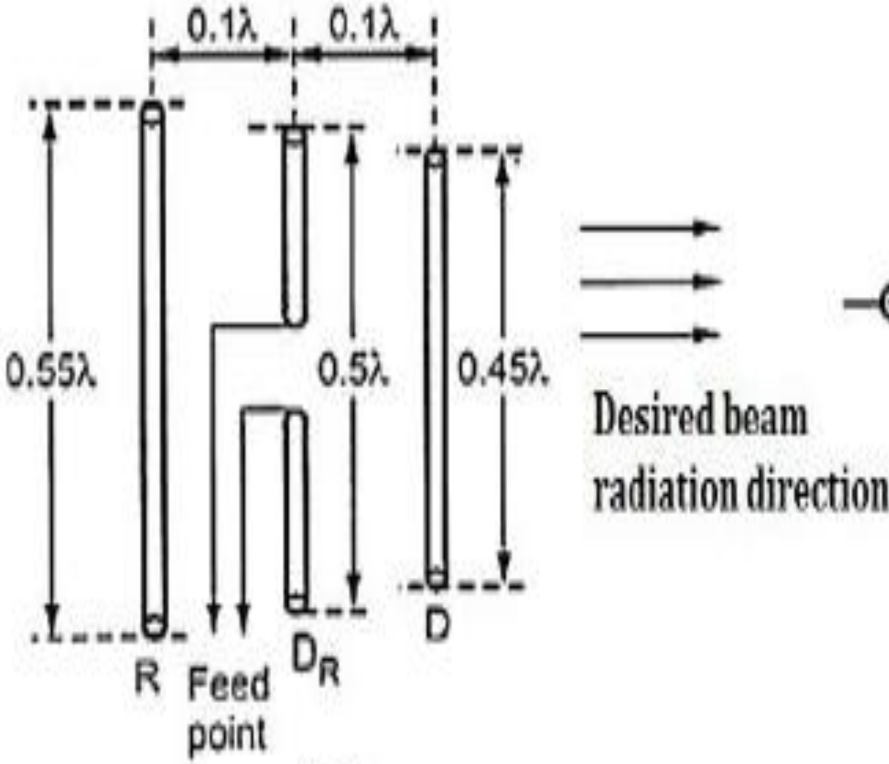
- The parasitic element receive excitation through the induced e.m.f. as current flows in the driven element.
- The phase and amplitude of the currents through the parasitic elements mainly depends on the length of the elements and spacing between the elements.
- To vary reactance of any element, the dimensions of the elements are readjusted.
- Generally the spacing between the driven and the parasitic elements is kept nearly 0.1 λ to 0.15 λ.

$$Reflector\ length = \frac{152}{f_{MHz}}\quad meter$$

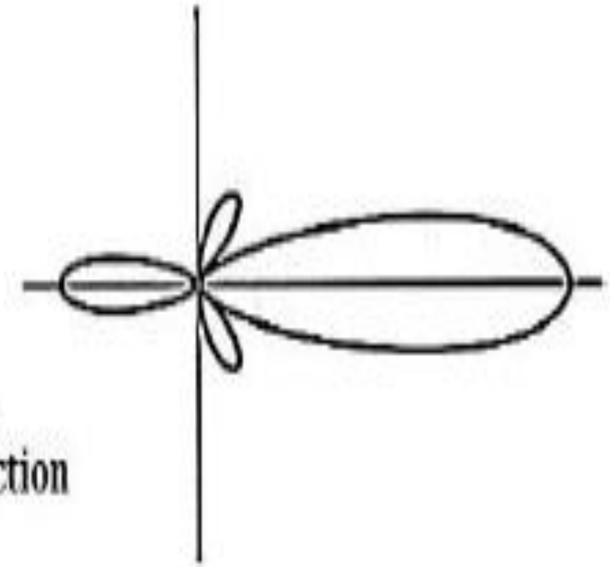
$$Driver\ element\ length = \frac{143}{f_{MHz}}\quad meter$$

$$Director\ length = \frac{137}{f_{MHz}}\quad meter$$

Fig 1 / Yagi Uda Antenna



(a)



(b) Radiation pattern

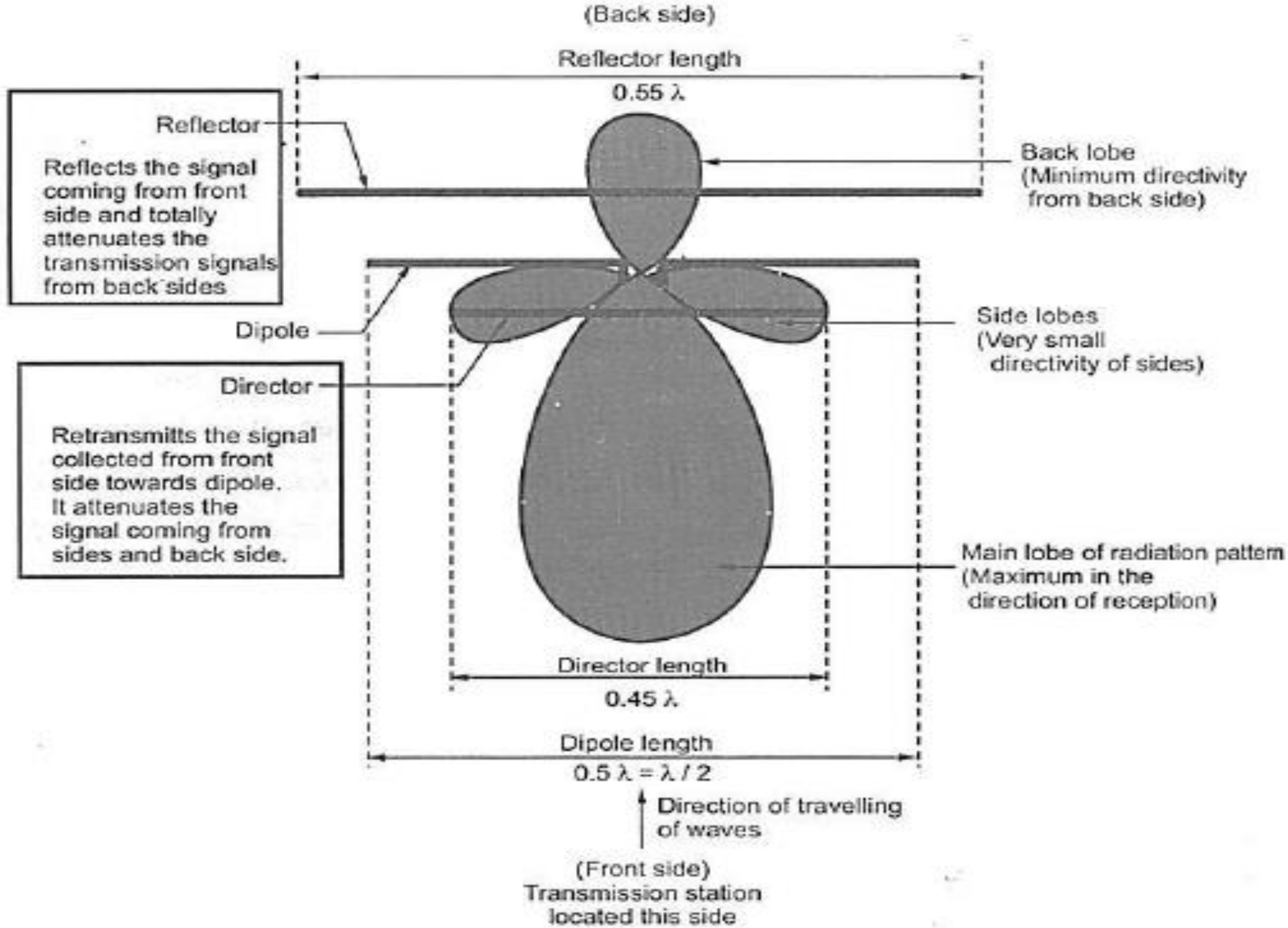
Working of Yagi Antenna

- The parasitic element is used either to direct or to reflect, the radiated energy forming compact directional antenna.
- If the parasitic element is greater than length $\lambda/2$, (i.e. reflector) then it is inductive in nature. Hence the phase of the current in such element lags the induced voltage.
- If the parasitic element is less than resonant length $\lambda/2$ (i.e. director), then it is capacitive in nature. Hence the current in director leads the induced voltage.
- The directors adds the fields of the driven element in the direction away from the driven element. If more than one directors are used, then each director will excite the next.
-

Working of Yagi Antenna 2

- To increase the gain of the Yagi-Uda antenna, the number of directors is increased in the beam direction.
- To get good excitation, the elements are closely spaced.
- The driven element radiates from front to rear (i.e., from reflector to director).
- Part of this radiation induces currents in the parasitic elements which reradiate almost all radiations.
- With the proper lengths of the parasitic elements and the spacing between the elements, the backward radiation is cancelled and the radiated energy is added in front.

Fig 2 / Yagi Antenna Radiation Pattern



Applications of Yagi Antenna

- Yagi-Uda array is the most popular antenna for the reception of terrestrial television signals in the VHF band (30 MHz-300 MHz).
- The array for this application is constructed using aluminium pipes.
- The driven element is usually a folded dipole, which gives four times the impedance of a standard dipole.
- Thus, a two-wire balanced transmission line having a characteristic impedance of 300 Ω can be directly connected to the input terminals of the Yagi-Uda array.
- Yagi-Uda arrays have been used in the HF, VHF, UHF, and microwave frequency bands

Principle of Frequency Independent Antennas

Frequency Independent Antennas

- The numerous applications of electro-magnetics need utilization of most of the electromagnetic spectrum. The invention of various broadband systems need the design of the broadband antennas.
- The antennas which are simple, small, light weight, economical and operating over the entire frequency band are most desirable.
- Such antennas are frequency independent antennas used in $\mathbf{10 - 10,000}\ \boldsymbol{MHz}$ region for applications such as TV, point to point communication, feeds for reflectors and lenses.
- According to the antenna scale model measurements, if the shape of the antenna is specified completely by angles, then its performance would be independent of frequency.

Frequency Independent Antennas 2

- To have such practical infinite structures, the current on the structure should decrease with distance away from the input terminals. After a certain point, the current becomes negligible and then the structure beyond that point to infinity can be removed.
- Such truncated antenna has lower cut-off frequency and beyond this cutoff frequency the radiation characteristics of the truncated antenna and infinite structure are identical. The lower cut-off frequency is that for which the current at the point of truncation becomes negligibly small.
- The biconical antenna can be completely specified by angles but the current along the structure does not reduce with distance away from the input terminals. Also its pattern does not have a form limiting with frequency.

Frequency Independent Antennas 3

Rumsey proposed a general shape equation which has frequency independent impedance, pattern and polarization characteristics and with this shape the current distribution reduces to zero rapidly.

Topic 1

Frequency Independent Antennas

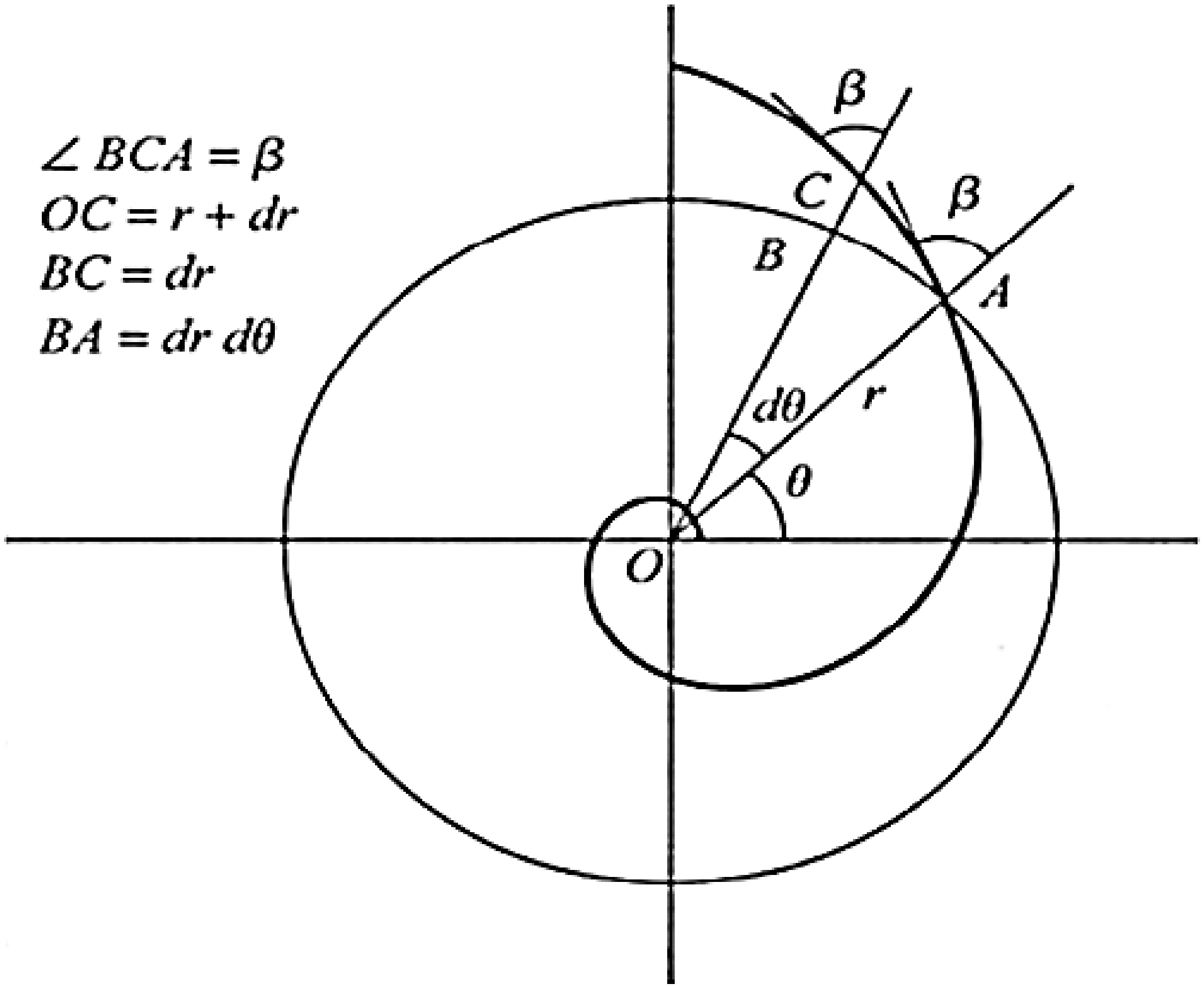
(Spiral Antenna)

Principle of Spiral Antenna

- Spiral is a geometrical shape found in nature. A spiral can be geometrically described using polar coordinates. Let (r, θ) be a point in the polar coordinate system. The equation
  - $r = r_0 e^{a\theta}$
- where, r_0 and a are positive constants, describes a curve known as a logarithmic spiral or an equiangular spiral. Taking natural logarithm on both sides of above equation.
  - $\ln r = \ln r_0 + a\theta$
- Differentiating with respect to ;

$$\frac{1}{r}\frac{dr}{d\theta} = a$$

Fig 1 / Logarithmic Spiral



Concept of Frequency Independence

- From $\Delta\, ABC$ in Figure 1 ; $$tan\beta = \frac{BA}{BC} = \frac{rd\theta}{dr} = \frac{1}{a}$$
-
- Therefore, the angle between the tangent at any point on the spiral and the radial line from the origin to that point (designated as β) is the same for all points on the spiral (a is a constant).
- Hence, the spiral represented by Equation 1 is also known as an equiangular spiral.
- Consider a spiral described by $$r_1 = r_0 e^{a\theta}$$
- The dimensions of an antenna designed to operate at a frequency, f_0. If the antenna is scaled by a factor K, it would have the same radiation and input properties at a frequency f_0/K.

Rumsey Principle

- Multiplying above Equation by a factor *K* - ; $r_2 = K r_0 e^{a\theta}$
- Expressing $K = e^{a\delta}$, above Equation reduces to ;

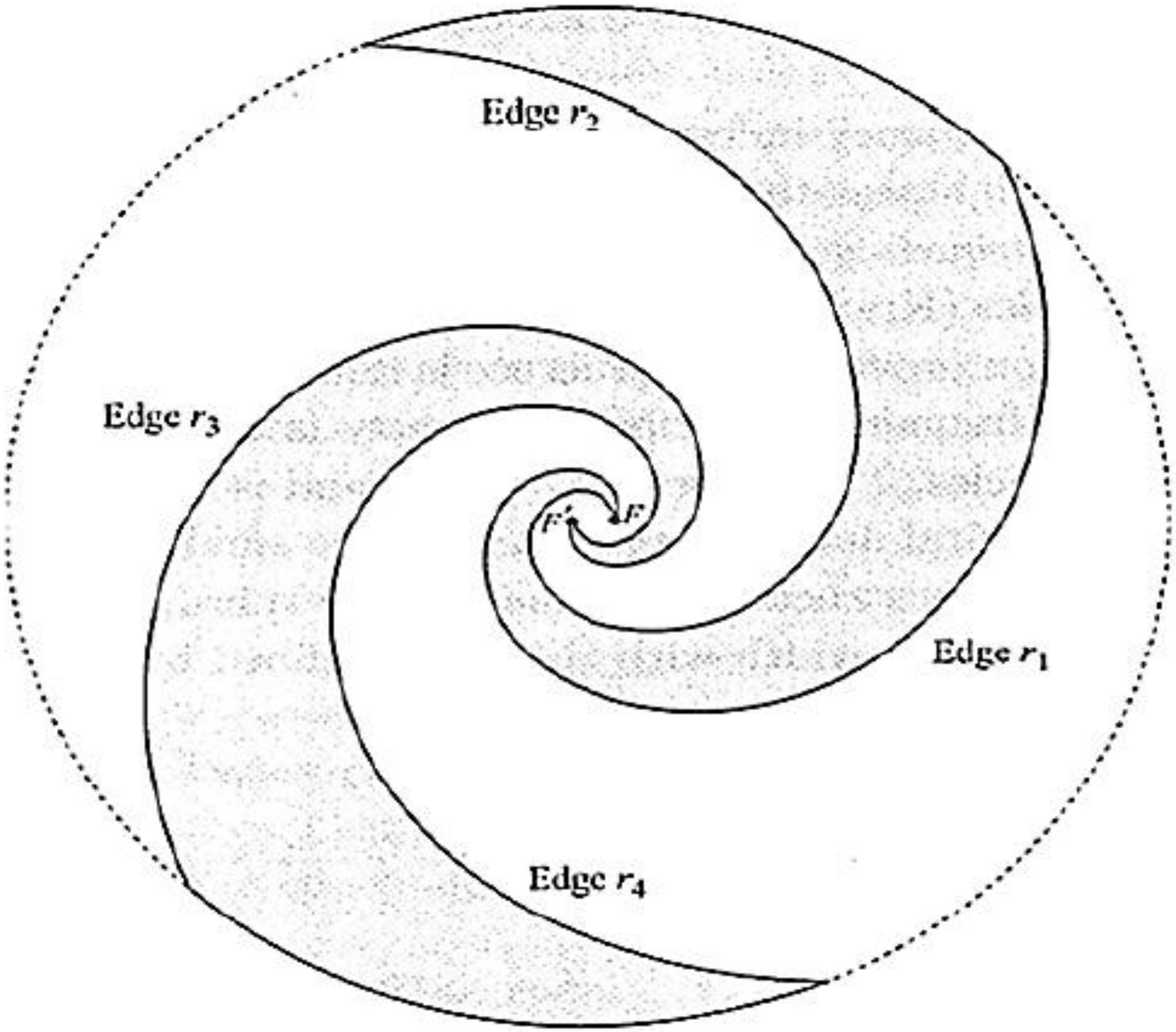
$$r_2 = e^{a\delta} r_0 e^{a\theta} = r_0 e^{a(\theta+\delta)}$$

- This shows that the scaled antenna is obtained by rotating the original antenna structure by an angle δ. The structure itself is unchanged.
- Hence, the radiation pattern alone rotates by an angle δ, keeping all the other properties the same. Such an antenna is known as a *frequency- independent antenna.*
- Frequency-independent antennas are governed by ***Rumsey's principle***, which states that the impedance and pattern properties of an antenna will be frequency independent if the antenna shape is specified only in terms of angles.

Rumsey Principle 2

- The antenna described by above Equation satisfies this criterion provided the structure is infinite.
- For structures that are finite in size, the frequency invariance property is exhibited over a limited range of frequencies. The lower end of this band is decided by the largest dimension of the spiral and the upper end by the smallest dimension.

Fig 2 / Logarithmic Spiral



Construction of Spiral Antenna

- Consider a thin conducting strip of variable width with the edges defined by the following two equations ;

$$r_1 = r_0 e^{a\theta} \qquad r_2 = r_0 e^{a(\theta-\delta)}$$

-

– These two edges are shown in Figure 2 for $0 \geq \theta \leq 2.25\pi$ *and* $\delta \geq \theta \leq (2.25\pi + \delta)$. A second conductor can be obtained by rotating the first spiral by 180°. The edges of the second spiral are given by ;

– $$r_3 = r_0 e^{a(\theta+\pi)} \qquad r_4 = r_0 e^{a(\theta+\pi-\delta)}$$

–These edges, edge r_3 and edge r_4 are shown in Figure 2 for $-\pi \geq \theta \leq 1.25\pi$ and $(-\pi + \delta) \geq \theta \leq (1.25\pi + \delta)$. These two conductors form a balanced structure with feed points *FF′*.

Construction of Spiral Antenna 2

- The parameters used to define this structure are ;

- δ : determines the width of the arm
- r_0 : determines the radius of the feed region
- a : rate of growth of the spiral, and
- θ_{max} : determines the maximum radius of the spiral.
- The spiral antenna has a bidirectional main lobe perpendicular to the plane of the antenna. The radiated field is right circularly polarized on one side and is left circularly polarized on the other side of the spiral.
- The axial ratio is used as one of the convenient parameters to define the acceptable bandwidth of the antenna.
- Outside the band of operation of the antenna, the radiation is elliptically polarized.

Topic 2

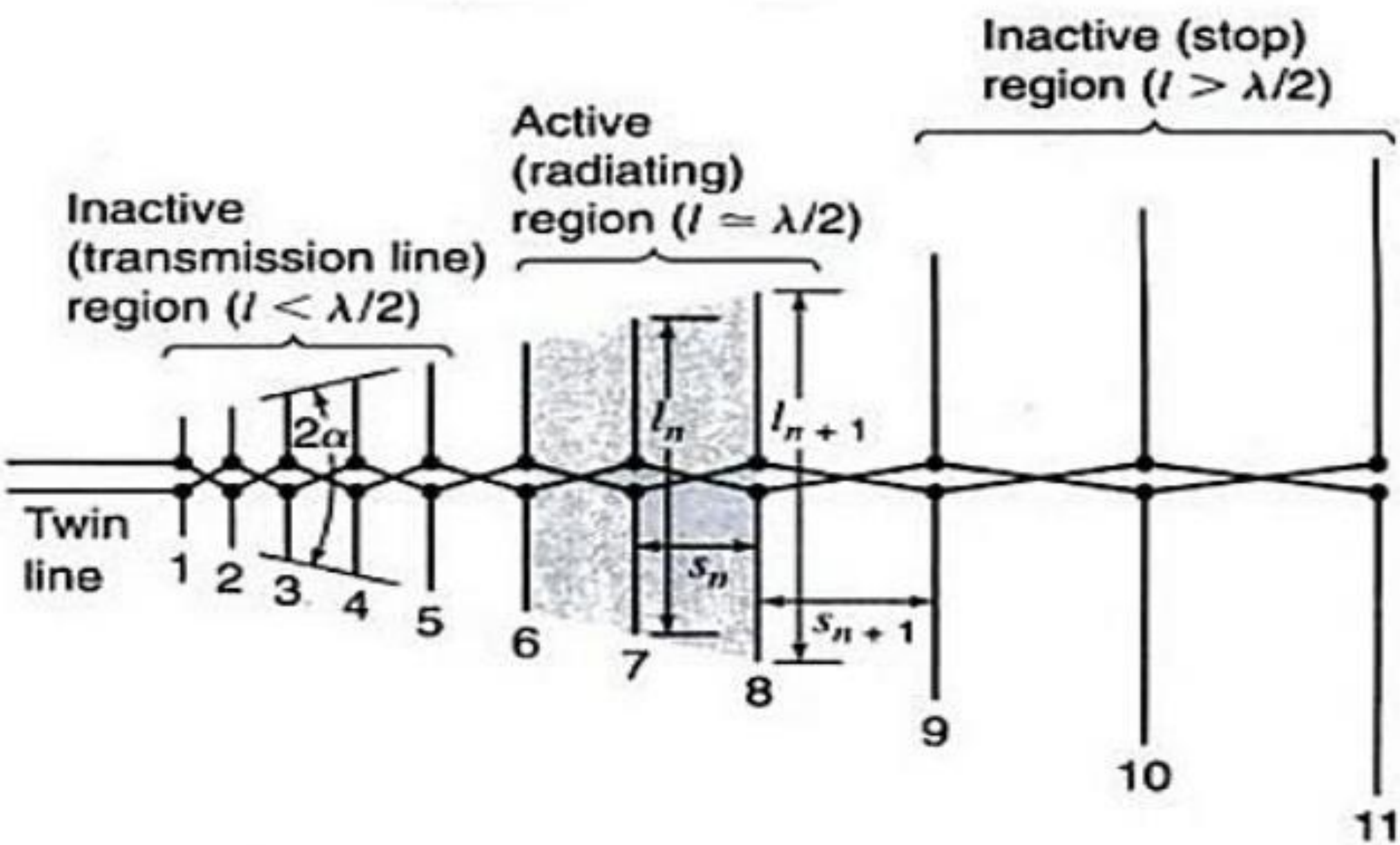
Frequency Independent Antennas

Log Periodic Antenna

Introduction to Log Periodic Antenna

- Any antenna defined in terms of angles, then it comes under the category of frequency independent antenna.
- In any frequency independent antenna, the impedance and the radiation pattern both are independent of frequency.
- In order to be frequency independent, the antenna should expand or contract in proportion to the wavelength. If the antenna structure is not mechanically adjustable, the size of the radiating region should be proportional to wavelength.
- It is a broadband antenna in which geometry of the antenna structure is adjusted such that all the electrical properties of the antenna are repeated periodically with the logarithm of the frequency.
- For every repetition, the structure size changes by a constant scale factor, with which structure can either expand or contract.

Fig 1 / Log Periodic Dipole Array (LPDA)



Construction of LPDA

- A typical log periodic dipole array (LPDA) consists number of dipoles of different lengths and spacings and fed by balanced two wire transmission line as shown in Figure 1.
- The feed line is connected at narrow end or apex of the array.
- The length of the dipoles increases from feed point towards other end such that the included angle α remains constant.
- The increase in the length of the dipole (l) and the spacing in wavelength between two dipoles (s) are adjusted such that the dimensions of the adjacent dipoles posses certain ratio with each other.
- The dipole lengths and the spacings between two adjacent dipoles are related through parameter called design ratio or scale factor which is denoted by τ.

Construction of LPDA 2

- Thus the relationship between s_n and s_{n+1} and l_n and l_{n+1} is given by –

$$\frac{l_{n+1}}{l_n} = \frac{s_{n+1}}{s_n} = k = \frac{1}{\tau}$$

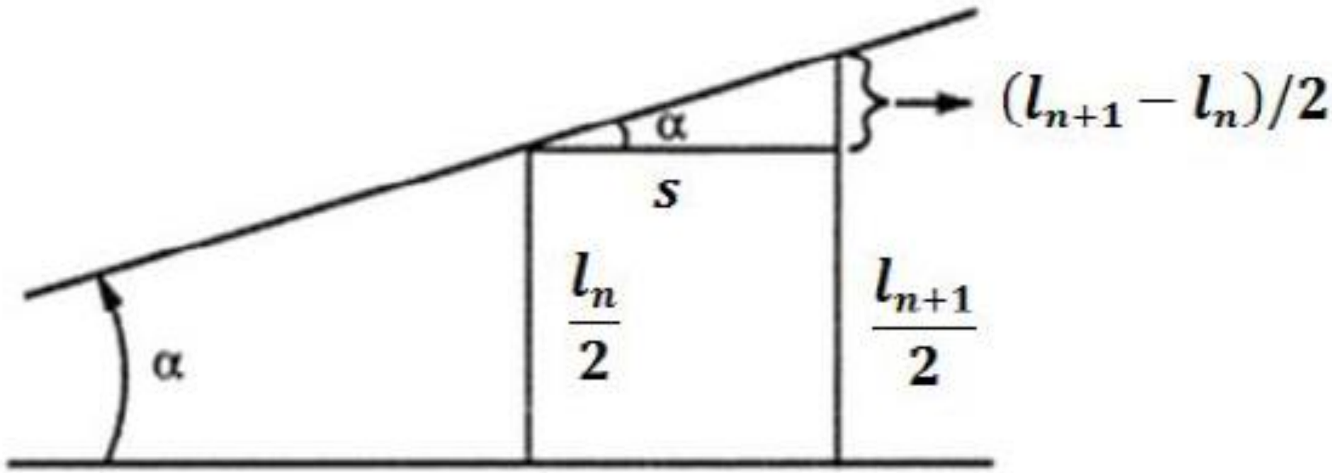
- The ends of the dipoles lie along straight lines on both the sides. These two straight lines meet at fixed point or apex giving angle 2α which is angle included by two straight line.
- Depending on the length of the dipoles, there are three regions in LPDA, namely inactive transmission line region, active region and inactive stop region.

Operating Regions of LPDA

- **Inactive transmission line region ($l < \lambda/2$):** The elements in this region provide capacitive impedance. The element spacing in this region is comparatively smaller. The currents in the region are very small hence it is considered as inactive region. These currents lead the voltage supplied by the transmission line.
- **Active region ($l \approx \lambda/2$): E**qual to resonant length. This is the central region of the array from where maximum radiation takes place. In this region, the dipoles offer resistive impedance. Thus the currents are large value and in phase with the base voltage.
- **Inactive stop region ($l > \lambda/2$): G**reater than resonant length. The dipoles offer inductive impedance. The currents are smaller in this region and also lags the base voltage.

Geometry of LPDA

- To find the relationship between the apex angle α, spacing s, and length l , consider part of the LPDA as shown in Figure 2.



- From the above figure –

$$\tan\alpha = \frac{(l_{n+1} - l_n)/2}{s} = \frac{l_{n+1} - l_n}{2s}$$

Geometry of LPDA 2

$$\tan\alpha = [l_{n+1}(1 - l_n/l_{n+1})]/2s$$

But ; $$\frac{l_{n+1}}{l_n} = k = \frac{1}{\tau}$$

$$\tan\alpha = \frac{[1-(1/k)](l_{n+1}/2)}{s}$$

For active region: $$l_{n+1} = \lambda/2$$

$$\tan\alpha = \frac{[1-(1/k)]}{4(s/\lambda)}$$

$$\tan\alpha = \frac{[1-(1/k)]}{4s_\lambda}$$

Geometry of LPDA 3

- where α = apex angle, k = scale factor , s_λ=spacing in wavelength short ward of $\lambda/2$ element.
- The length of any element say $n + 1^{th}$ element and length of first element is related as ;

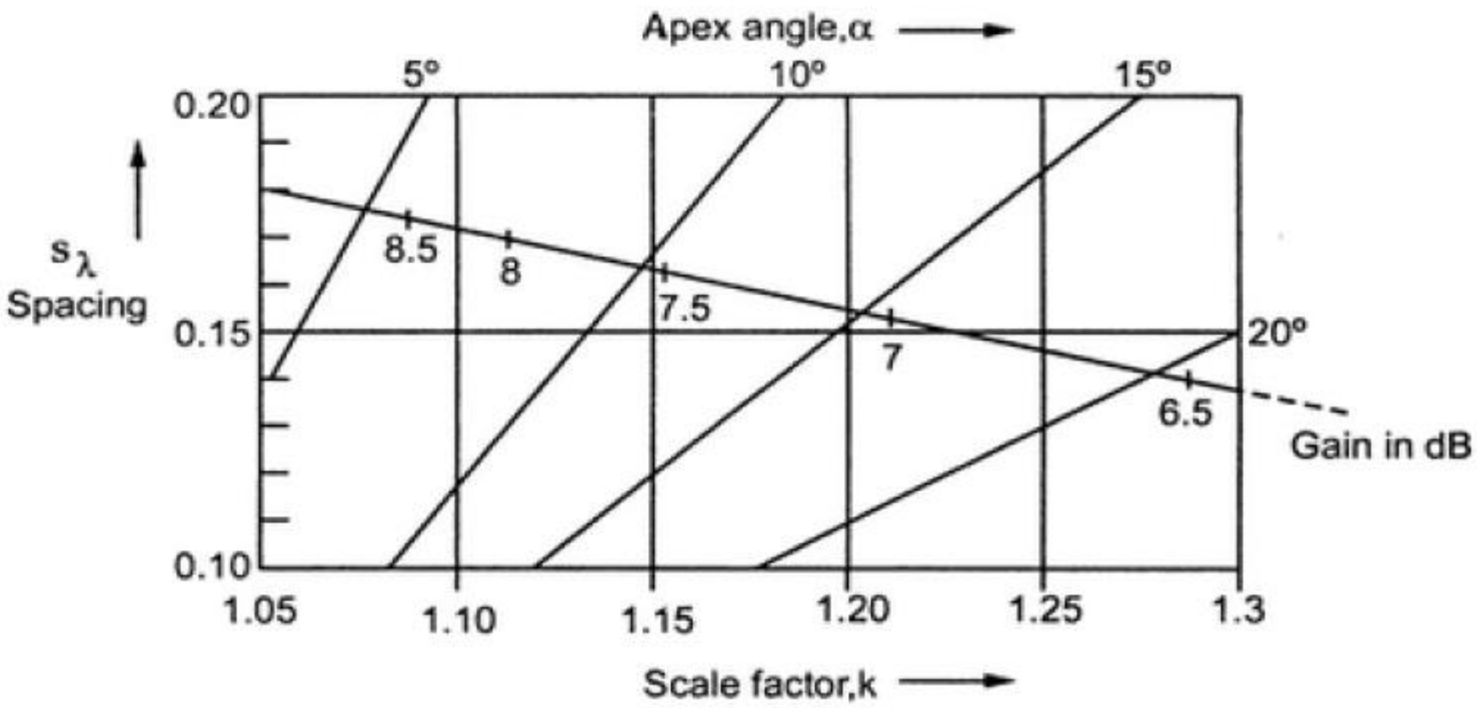
$$\frac{l_{n+1}}{l_1} = k^n = F$$

- When the length of the first element is l_1 then the length $n + 1^{th}$ element is k^n time greater than l_1. This ratio is also termed as frequency ratio F or it is called Bandwidth.
- The relation between the apex angle α, scale factor k and spacing s_λ with optimum design line and gain represented in Figure 3.

Design Curve of LPDA

- The number of elements in the array (n) can be obtained from upper frequency f_U and lower frequency f_L as ;

$$\log(f_U) - \log(f_L) = (n-1)\log\left(\frac{1}{\tau}\right)$$



Topic 3

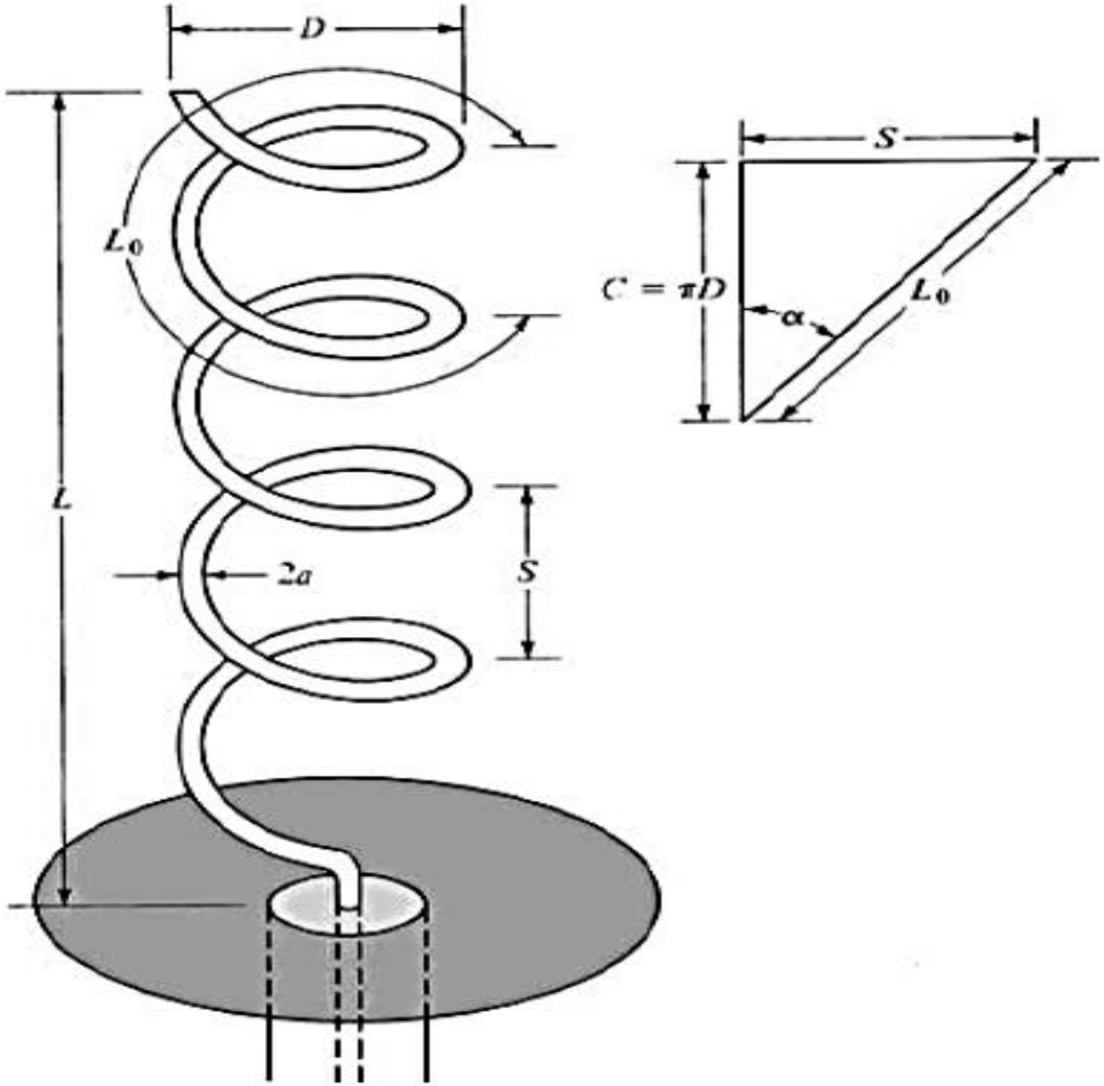
Frequency Independent Antennas

Helical Antenna

Introduction to Helical Antenna

- Helical antenna is a broadband VHF and UHF antenna used to provide circular polarization.
- It consists of a thick copper wire wound in the form of a screw thread forming a helix.
- In general , helix is used with a ground plane.
- There are different forms of ground plane such as flat ground plane, cylindrical cavity.
- The helix is usually connected to the center conductor of a coaxial transmission line at the feed point with the outer conductor of the line attached to the ground plane is as shown in Figure 1.

Fig 1 / Helical Antenna with Ground Plane

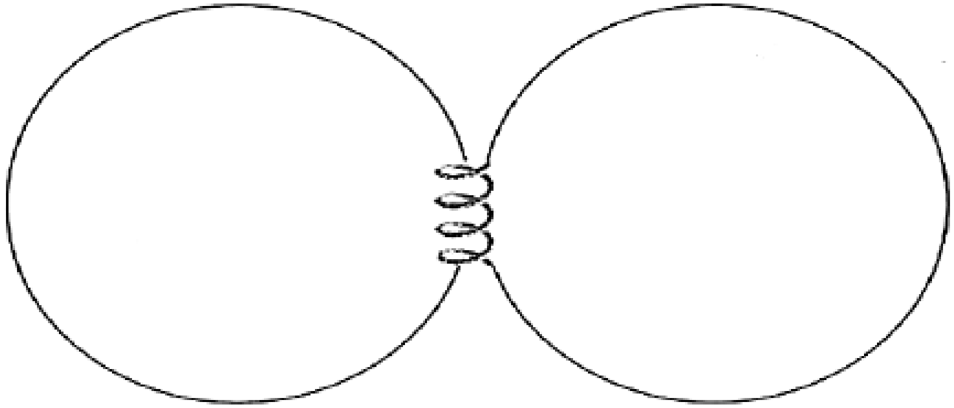


Construction of Helical Antenna

- The geometrical configuration of a helix consists of N turns, diameter D and spacing S between each turn.
- The total length of the antenna is $L = NS$ while the total length of the wire is $L_n = NL_0 = N\sqrt{S^2 + C^2}$, where $L_0 = \sqrt{S^2 + C^2}$ is the length of the wire between each turn and $C = \pi D$ is the circumference of the helix.
- The pitch angle α which is the angle formed by a line tangent to the helix wire and a plane perpendicular to the helix axis.
- The pitch angles defined by – $\alpha = \tan^{-1}\left(\frac{S}{\pi D}\right) = \tan^{-1}\left(\frac{S}{C}\right)$
- When $\alpha = 0°$, the winding is flattened and the helix reduces to a loop antenna of N turns; When $\alpha = 90°$ then the helix reduces to a linear wire; When $0° < \alpha < 90°$, a true helix is formed with a circumference greater than zero but less than the circumference when the helix is reduced to a loop ($\alpha = 0°$).

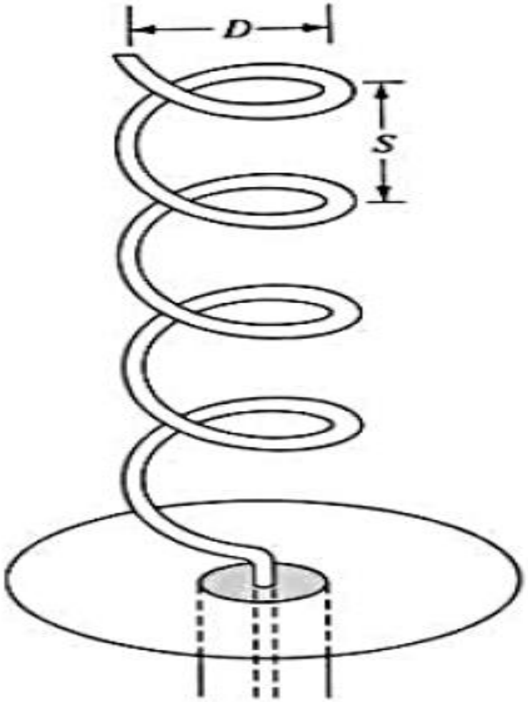
Normal Mode of Operation

- The helical antenna can operate in many modes but the two principal modes are the normal (broadside) and the axial (end-fire) modes.
- **Normal mode: T**he field radiated by the antenna is maximum in a plane normal to the helix axis and minimum along its axis, as shown sketched in Figure 2. To achieve the normal mode of operation, the dimensions of the helix are usually small compared to the wavelength (i.e., $NL_0 \ll \lambda_0$).

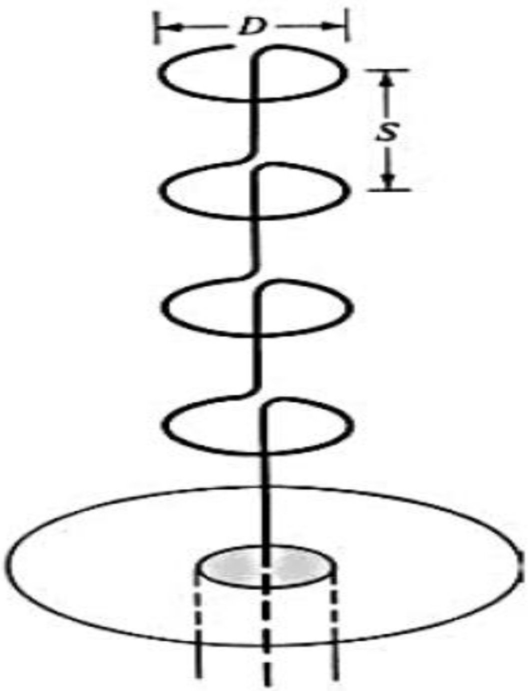


Normal Mode of Operation 2

- In the normal mode, the helix of Figure 3 (a) can be simulated approximately by N small loops and N short dipoles connected together in series as shown in Figure 3 (b). The fields are obtained by superposition of the fields from these elemental radiators. The planes of the loops are parallel to each other and perpendicular to the axes of the vertical dipoles.



(a) Normal mode



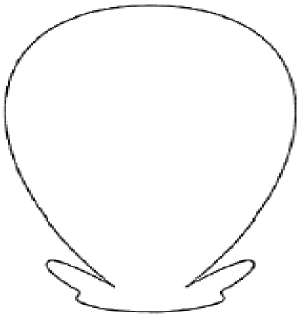
(b) Equivalent

Normal Mode of Operation 3

- Since in the normal mode the helix dimensions are small, the current throughout its length can be assumed to be constant and its relative far-field pattern to be independent of the number of loops and short dipoles.
- Thus its operation can be described by the sum of the fields radiated by a small loop of radius D and a short dipole of length S, with its axis perpendicular to the plane of the loop, and each with the same constant current distribution.
- Practically this mode of operation is limited and it is hardly used because its bandwidth and radiation efficiency is very small.

Axial Mode of Operation

- In this mode of operation, there is only one major lobe and its maximum radiation intensity is along the axis of the helix, as shown in Figure 4.
- The minor lobes are at oblique angles to the axis. To excite this mode, the diameter D and spacing S must be large fractions of the wavelength. The antenna is used in conjunction with a ground plane whose diameter is at least $\lambda_0/2$, and fed by a coaxial line.
- However, other types of feeds are possible, especially at microwave frequencies. The dimensions of the helix for this mode of operation are not as critical, thus resulting in a greater bandwidth



Design Procedure

- The terminal impedance of a helix radiating in the axial mode is nearly resistive with values between 100 and 200 ohms.

The input impedance (purely resistive) is obtained by ;

$$R \approx 140\left(\frac{C}{\lambda_0}\right)$$

which is accurate to about ±20%, the half-power beamwidth by ;

$$HPBW\ (degrees) \approx \frac{52\ {\lambda_0}^{3/2}}{C\sqrt{NS}}$$

The beamwidth between nulls ;

$$FNBW\ (degrees) \approx \frac{115\ {\lambda_0}^{3/2}}{C\sqrt{NS}}$$

The directivity by

$$D_0\ (dimensionless) \approx 15\ N\frac{C^2 S}{{\lambda_0}^3}$$

Design Procedure 2

The axial ratio (for the condition of increased directivity) by

$$AR = \frac{2N+1}{2N}$$

The normalized far-field pattern is given by

$$E = \sin\left(\frac{\pi}{2N}\right)\cos\theta\,\frac{\sin[(N/2)\psi]}{\sin[\psi/2]}$$

where

$$\psi = 2\pi\left[\frac{S}{\lambda_0}(1-\cos\theta)+\frac{1}{2N}\right]$$

The above formula valid for , $12° < \alpha < 15°$, $N \geq 3$ and $\frac{3}{4} < \frac{C}{\lambda_0} < \frac{4}{3}$.

Topic 4

Modern Antennas

Reconfigurable Antenna

Introduction to Reconfigurable Antenna

- A reconfigurable antenna is an antenna capable of modifying dynamically its frequency and radiation properties in a controlled and reversible manner.
- In order to provide a dynamical response, reconfigurable antennas integrate an inner mechanism (such as RF switches, varactors, mechanical actuators or tunable materials) that enable the intentional redistribution of the RF currents over the antenna surface and produce reversible modifications over its properties.
- Reconfigurable antennas differ from smart antennas because the reconfiguration mechanism lies inside the antenna rather than in an external beam forming network.
- The reconfiguration capability of reconfigurable antennas is used to maximize the antenna performance in a changing scenario or to satisfy changing operating requirements.

Classification of Reconfig Antennas

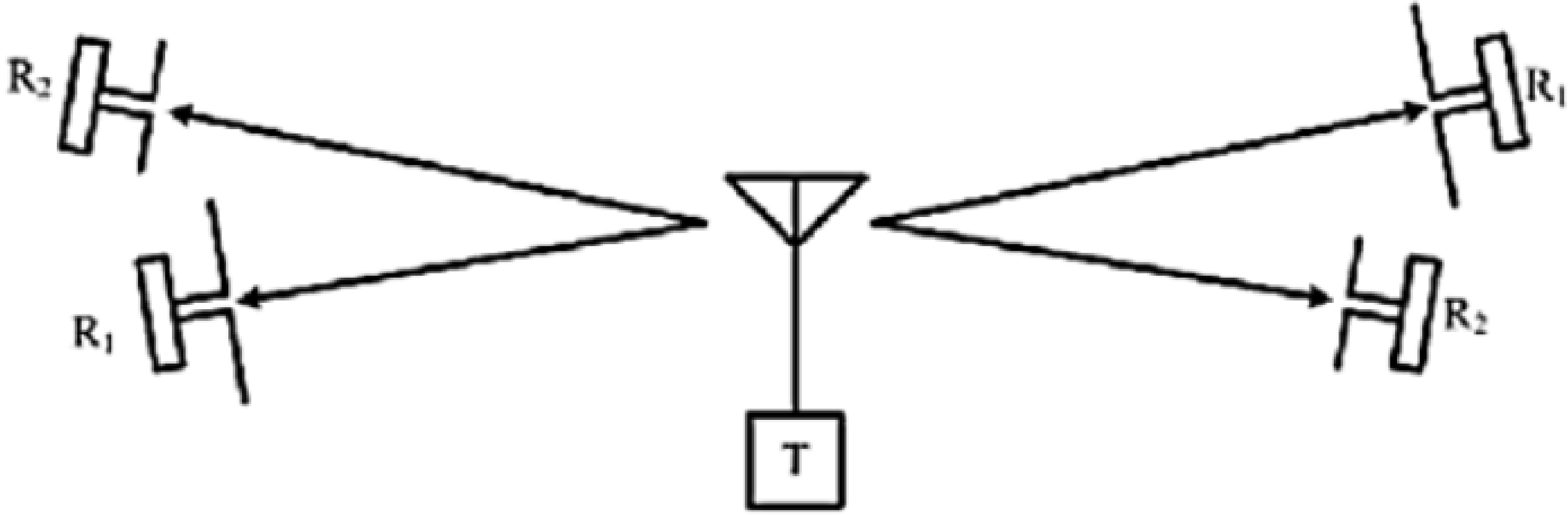
- Reconfigurable antennas come in a large variety of different shapes and forms. Their operation can largely be analyzed through existing design principles by utilizing well defined antennas as the base design and a point of reference for the desired operation.
- Reconfigurable antennas can be classified according to three categories that describe their operation:
- (1) the reconfigurable antenna parameters of interest, (2) the proximity of reconfiguration, and (3) the continuity of reconfiguration (e.g., having reconfigurable antenna parameters over a continuous range of values).
- Reconfigurable antennas are described by the first of these categories, including reconfigurable radiation (pattern or polarization) and reconfigurable impedance (frequency or bandwidth).

Reconfiguration of Antennas

- The proximity of reconfiguration describes physical properties inherent to the base antenna design—either direct (alteration of a driven element) or parasitic (alteration of a parasitic component).
- The continuity of the reconfiguration is defined by the nature and capabilities of the reconfiguration mechanism, either discrete (a finite number of reconfigured states) or continuous (reconfiguration within a range of states).

Frequency Reconfigurable Dipole

- A generic wireless communication link shown in Figure 1 illustrates a basic application of reconfigurable antenna.
- This scenario involves a transmitter $\boldsymbol{T}$ that broadcasts to two sets of wireless receivers $\boldsymbol{R_1}$ and $\boldsymbol{R_2}$. These receivers operate at two different frequency bands $\boldsymbol{B_1}$ and $\boldsymbol{B_2}$, centered at $\boldsymbol{f_1}$and $\boldsymbol{f_2}$, respectively (with$\boldsymbol{f_1} < \boldsymbol{f_2}$).



Frequency Reconfigurable Dipole 2

- The scenario assumes that all receiving antennas are coincidentally polarized, the transmitter broadcasts at the frequency bands centered at $\boldsymbol{f_1}$and $\boldsymbol{f_2}$, at times $\boldsymbol{t_1}$and $\boldsymbol{t_2}$,, respectively, and the radios require isolation between the bands such that a dual-band antenna is undesirable.
- Thus the reconfigurable antenna serves to allow communication with both sets of receivers using a single antenna.

Topic 5

Modern Antennas

Active Antenna

Introduction to Active Antenna

- The term *active antenna* implies an antenna integrated with an active circuit, including the DC bias circuit, and without an isolator or circulator between them.
- The absence of isolator/ circulator implies that neither the antenna nor the circuit needs to be designed in a 50-Ω environment.
- Figure 1 shows block diagrams of several types of active antennas, classified according to their functionality.

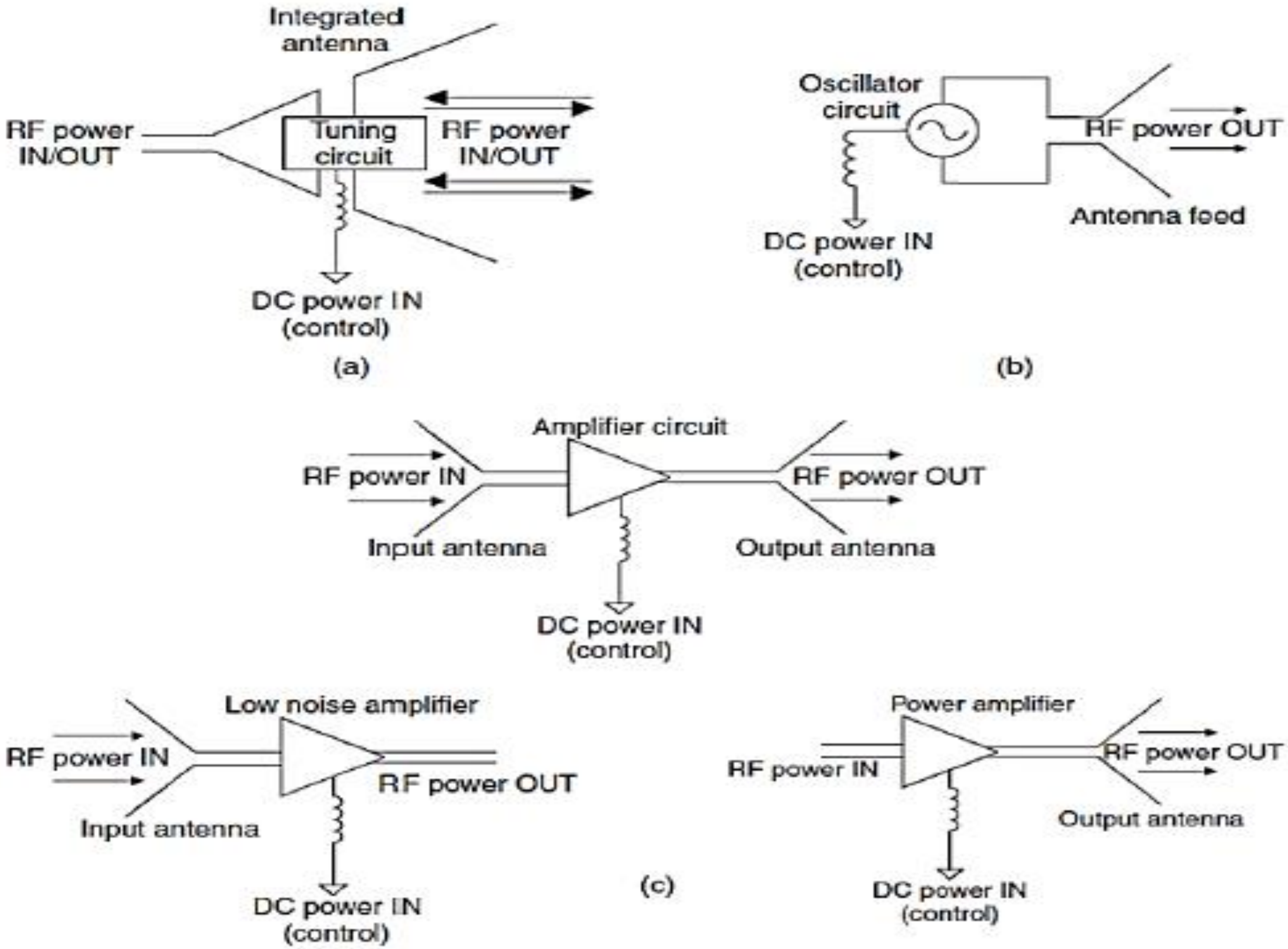
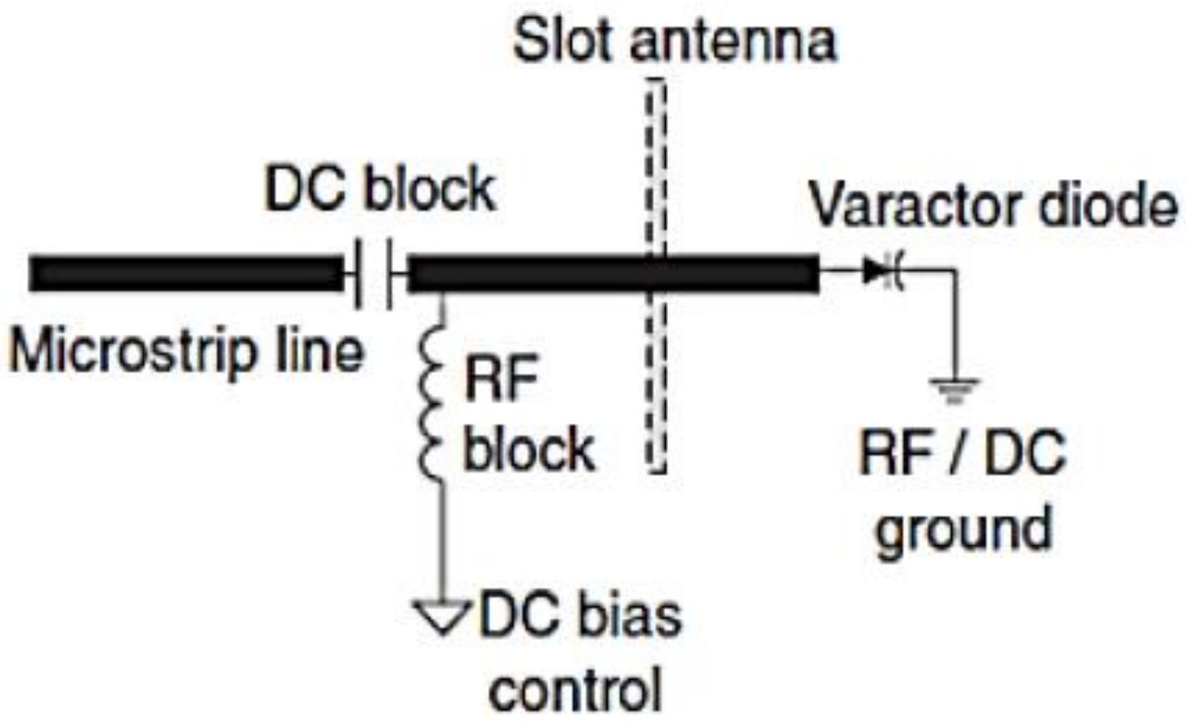


Fig. 4-11 Schematic diagrams of a few examples of active antennas: (*a*) actively tuned antenna, (*b*) oscillator antenna, and (*c*) amplifier antenna with repeater, receiver, and transmitter subclasses. An *active antenna* implies an antenna integrated intimately with an active circuit.

Frequency Agile Antennas

- A two- or three-terminal active device can be designed into the antenna to enable the antenna impedance to tune with frequency.
- Figure 2 shows an example of this type of active antenna: a slot microstrip feedline contains a varactor diode tuning element.
- When the capacitance of the diode changes, the electrical length of the antenna, which in turn depends on the antenna reactance, changes and the antenna becomes resonant at a different frequency.
- In this case, DC power is used to provide increased bandwidth of an antenna element.
- These antennas find applications in multifunctional systems with multiple non-simultaneous carriers.

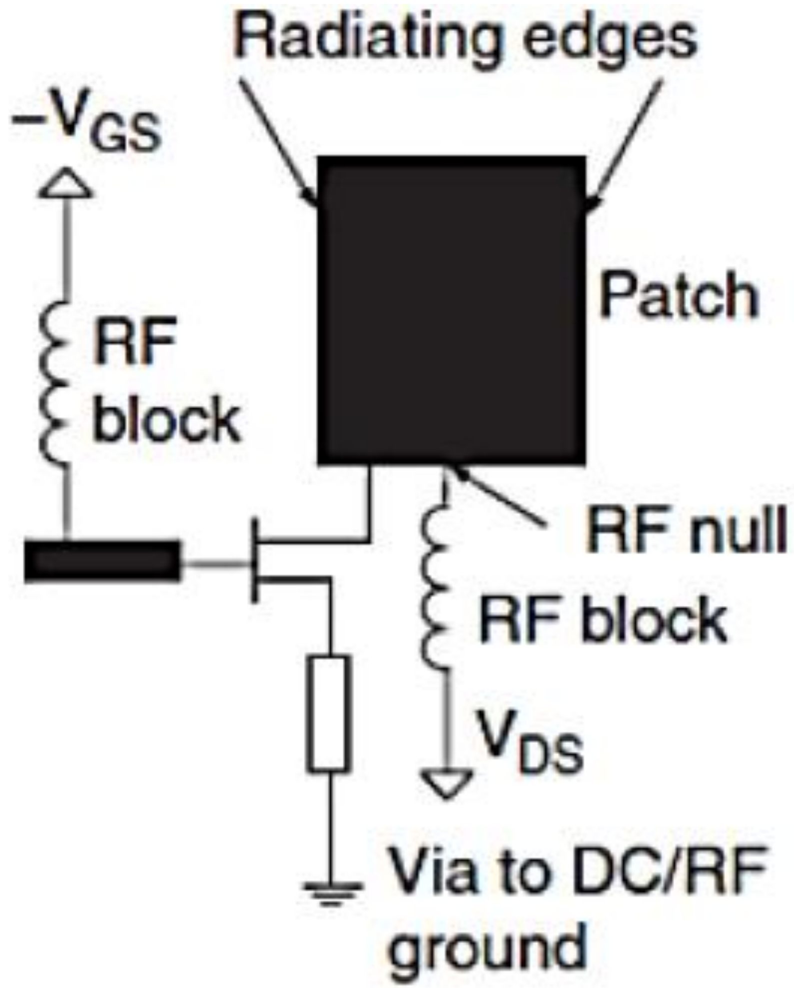
Fig 2 / Frequency Agile Antenna



Oscillator Antennas

- A two- or three-terminal negative-resistance device can be connected directly to the terminals of a single antenna element or an array of elements.
- In this case, DC power is converted to radiated RF power.
- An example of a patch antenna in the feedback loop of a transistor, shown in Figure 3.
- Oscillator antennas have been discussed for applications such as low-cost sensors, power combining and synchronized scanning antenna arrays.

Fig 3 / Oscillator Antenna

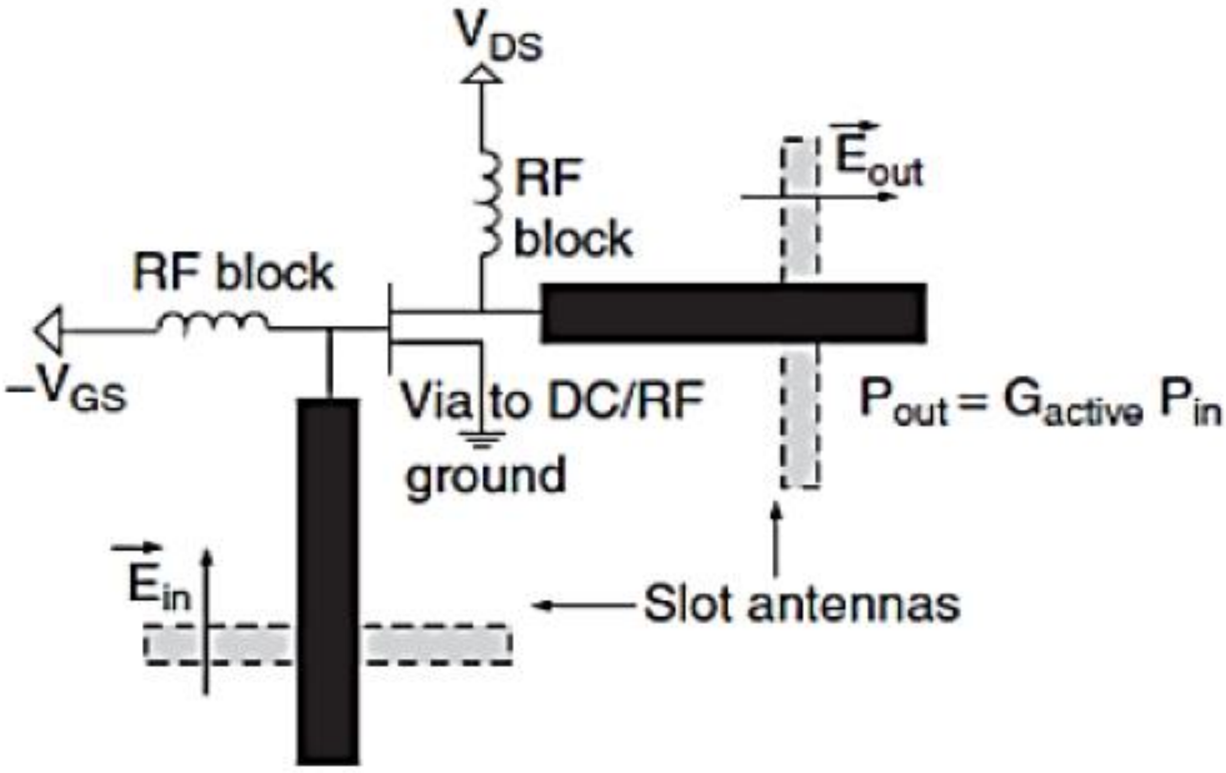


Amplifier Antennas

- An active device is connected to the terminals of an antenna element to provide amplification in receive mode or transmit mode.
- In the former case, the matching between the antenna and active element usually optimizes noise, while in the latter case, the matching optimizes power and/or efficiency.
- Figure 4 shows an example of a repeater element with two slot antennas and a pre-matched amplifier chip.
- In this case, increase in gain is enabled by adding DC power to the antenna and it becomes difficult to separate antenna gain from circuit gain.

Fig 4 / Amplifier Antennas 2

- They find applications in transmitters where spatial power combining can be achieved with an array & in receivers where the feed line loss which contributes to the total noise figure, can be eliminated by directly connecting an LNA to the receiving antenna.



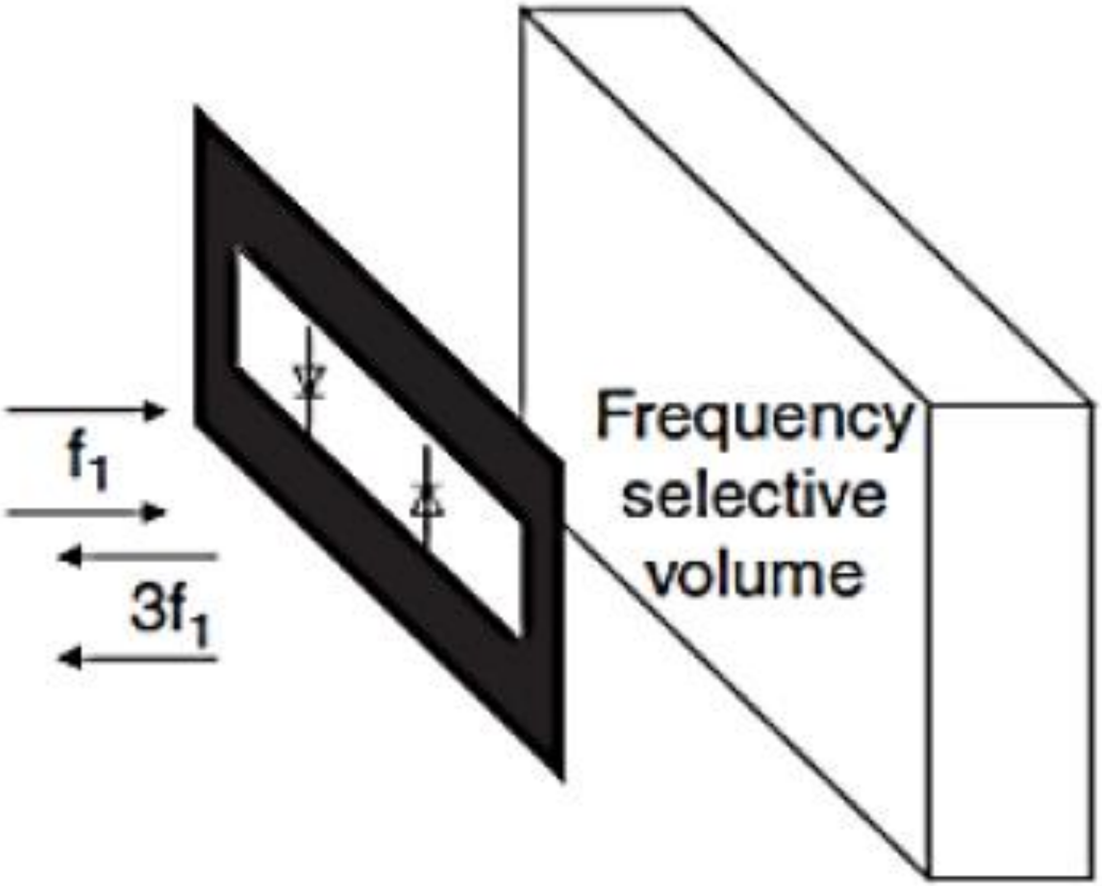
Frequency Conversion Antennas

- A two- or three-terminal active device integrated with an antenna can provide direct down or up conversion of a radiated signal, at frequencies that are direct harmonics or subharmonics of a fundamental frequency.
- Figure 5 shows an example of a slot antenna with a Schottky diode which can be used for frequency doubling since the slot is matched to the diode impedance at both the input frequency and its harmonic.
- Such antennas have applications in receivers, mixers with high dynamic range, detectors for millimeter-wave and THz receivers, phase conjugating RFID type antennas and high-frequency generation.

Frequency Conversion Antennas 2

- A special case of frequency-conversion antennas is when a two- or three-terminal rectifying device is connected directly to the terminals of a receiving antenna in such a way that the received RF power is converted with optimal efficiency to DC power, while harmonic production and re-radiation is minimized.
- This type of active antenna is referred to as a rectenna.
- Such antennas have been applied to RFID tags, sensor powering for cases when there is no solar power and where it is difficult to replace batteries, directed narrow-beam array power beaming, and for energy recycling and/or scavenging.

Fig 5 / Frequency Conversion Antennas



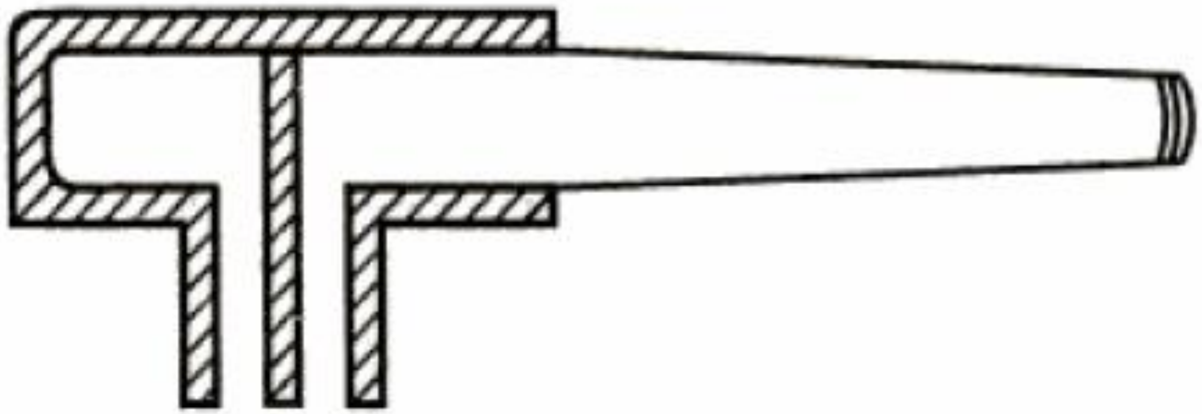
Topic 6

Modern Antennas

Dielectric Antenna

Introduction to Dielectric Antenna

- The travelling wave antenna in which the travelling wave is guided by a dielectric is called dielectric antenna as shown in Figure 1.
- In dielectric antenna, near cut-off, the phase velocity equals the velocity of light. The fields produced extend outside a dielectric guide.
- These outward fields excite the desired radiation in free-space. Such travelling wave antennas are useful for broad band signals.



Construction of DRA

- **A dielectric resonator antenna** (DRA) is a radio antenna mostly used at microwave frequencies and higher, that consists of a block of ceramic material of various shapes, the dielectric resonator mounted on a metal surface, a ground plane.
- Radio waves are introduced into the inside of the resonator material from the transmitter circuit and bounce back and forth between the resonator walls, forming standing waves.
- The walls of the resonator are partially transparent to radio waves, allowing the radio power to radiate into space.

Applications of DRA

- An advantage of dielectric resonator antennas is they lack metal parts, which become lossy at high frequencies, dissipating energy.
- Hence these antennas can have lower losses and be more efficient than metal antennas at high microwave and millimeter wave frequencies.
- Dielectric waveguide antennas are used in some compact portable wireless devices and military millimeter-wave radar equipment.

History of DRA

- The antenna was first proposed by Robert Richtmyer in 1939.
- In 1982, Long et al. did the first design and test of dielectric resonator antennas considering a leaky waveguide model assuming magnetic conductor model of the dielectric surface.
- Thus, they argued that the dielectric antenna behaved like a magnetic dipole antenna.
- The magnetic conductor model does not explain how current in the dielectric medium is transformed into electromagnetic waves which results in radiation.
- The electric field from oscillation of polarized dipole fall off inversely with cube of distance and cannot be responsible for far field radiation.

Topic 7

Electronic Band Gap Structure (EBG) and Applications

Definition for EBG

- Periodic structures are abundant in nature, which have fascinated artists and scientists alike.
- When they interact with electromagnetic waves, exciting phenomena appear and amazing features result.
- In particular, characteristics such as frequency stop bands, pass bands, and band gaps could be identified.
- These applications are seen in filter designs, gratings, frequency selective surfaces (FSS), photonic crystals and photonic band gaps (PBG), etc.
- Electromagnetic band gap structures are defined as artificial periodic objects that prevent/assist the propagation of electromagnetic waves in a specified band of frequency for all incident angles and all polarization states.

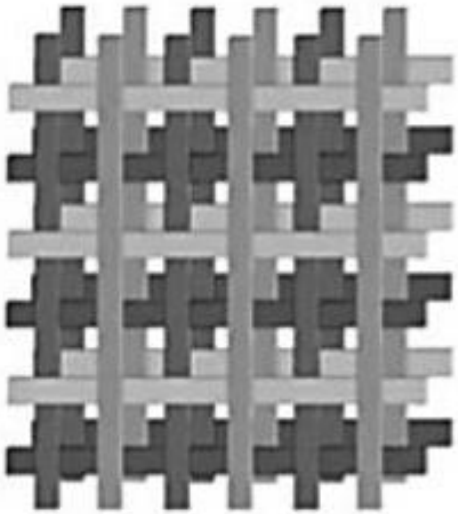
Classification of EBG Structures

- EBG structures are realized by periodic arrangement of dielectric materials and metallic conductors.
- They can be categorized into three groups according to their geometric configuration: (1) three-dimensional volumetric structures, (2) two-dimensional planar surfaces, and (3) one-dimensional transmission lines.
- Different EBG structures : 3-D EBG structures (a woodpile structure consisting of square dielectric bars and a multi-layer metallic tripod array) , 2-D EBG surfaces (mushroom-like surface and a uni-planar design without vertical vias), one-dimensional EBG transmission line designs as in Figure 1.

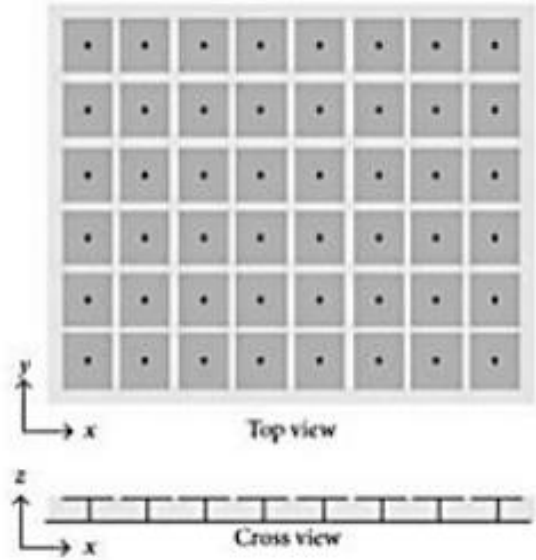
Advantages of EBG

- 2-D EBG surfaces has the advantages of low profile, light weight, and low fabrication cost, and are widely considered in antenna engineering.
- The planar electromagnetic band gap (EBG) surfaces exhibit distinctive electromagnetic properties with respect to incident electromagnetic waves.

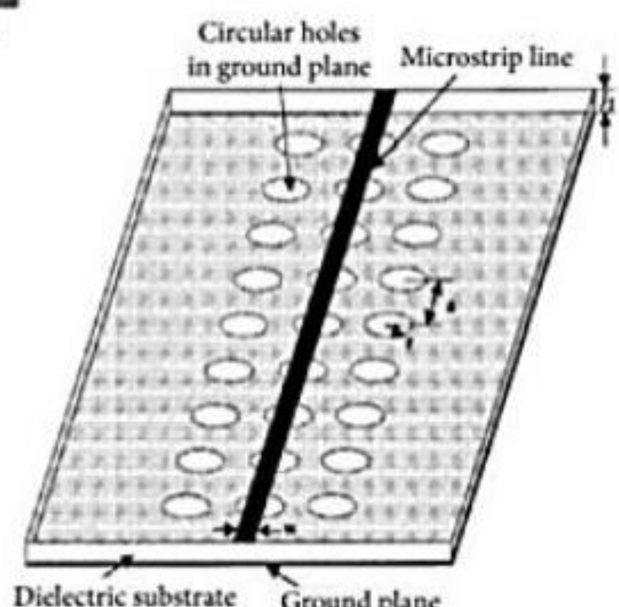
Fig 1 / Different EBG Structures



(a) 3-D EBG: a woodpile dielectric structure



2-D EBG: a mushroom EBG.



1-D EBG: a Micro strip line with periodic holes on the ground plane.

Applications of EBG Structures

- A multitude of basic EBG applications exists especially within the microwave and low millimeter wave region.
- For example, In electronically scanned phased arrays, high-precision GPS, Bluetooth, mobile telephony, waveguides, antennas, low loss- coplanar lines, and compact integrated filters.

Topic 8

Antenna Measurements – Test Ranges

Introduction to Test Ranges

- There are two methods of antenna measurements: indoor and outdoor.
- Both the methods have their own limitations; the outdoor measurements are not protected from the environmental conditions whereas indoor measurements suffer space restrictions.
- In general, for the accurate measurements uniform plane waves should incident on the antenna and this is possible only if measurements are carried out in far-field region.
- The region in which antenna measurements are performed effectively is termed antenna ranges and it is basically of two types: reflection ranges and free-space ranges.

Types of Test Ranges

- **Reflection ranges**:
  - Outdoor type test range, where ground is a reflecting surface.
  - Reflection ranges create constructive interference in the region surrounding AUT.
  - Height of Tx antenna adjusted while Rx antenna is maintained constant.
  - Suitable for the antenna systems operating in frequency range from UHF to 16 GHz.
- **Free-space ranges**:
  - Indoor type test range designed to minimize environmental effects.
  - Most popular test range where antennas mounted over tall towers.
  - Main problem of this method is reflection from the ground.

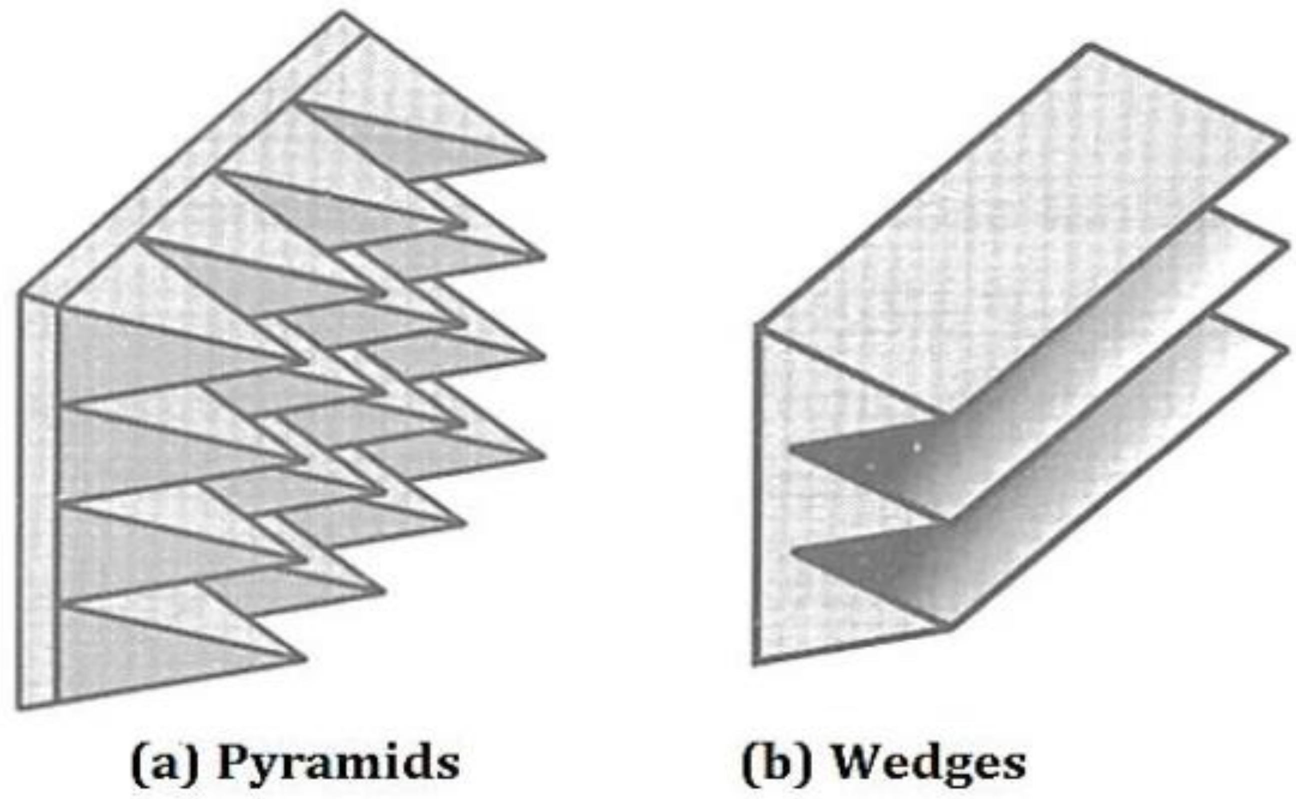
Anechoic Chamber

- Anechoic chamber is an indoor chamber. The chamber walls, celing and floor are filled with absorbing material except at the location of transmitting antenna and antenna under test (AUT).
- It simulates a reflection-less free space and allows all-weather antenna measurements in a controlled environment.
- In anechoic chamber, the area where test antenna situated is isolated from all types of interfering signals in better way.
- To improve isolation of test area, many times shielding is done which also allows Electromagnetic Compatibility (EMC) measurements.
- Small antennas - far field measurements possible using anechoic chamber with outdoor range; Large antennas - compact antenna test ranges and rear field ranges are installed in anechoic chambers itself.

Anechoic Chamber / Absorbing Materials

- The absorbing materials are not only the integral part of measurement ranges but important components of antennas used to reduce the side lobe and back lobe radiations.
- Typical broadband absorber used is carbon-loaded polyurethane foam. An ideal absorber can provide an impedance match for the incoming waves at all frequencies and angles of incidence.
- By shaping absorber or by gradually varying resistivity of material, a tapered transition in impedance from free space to back of the absorber can be achieved.
- The most widely used geometrical shapers are pyramids and wedges as shown in Figure 1 (a) and (b) respectively.

Fig 1 / Absorbing Materials



Anechoic Chamber / Absorbing Materials 2

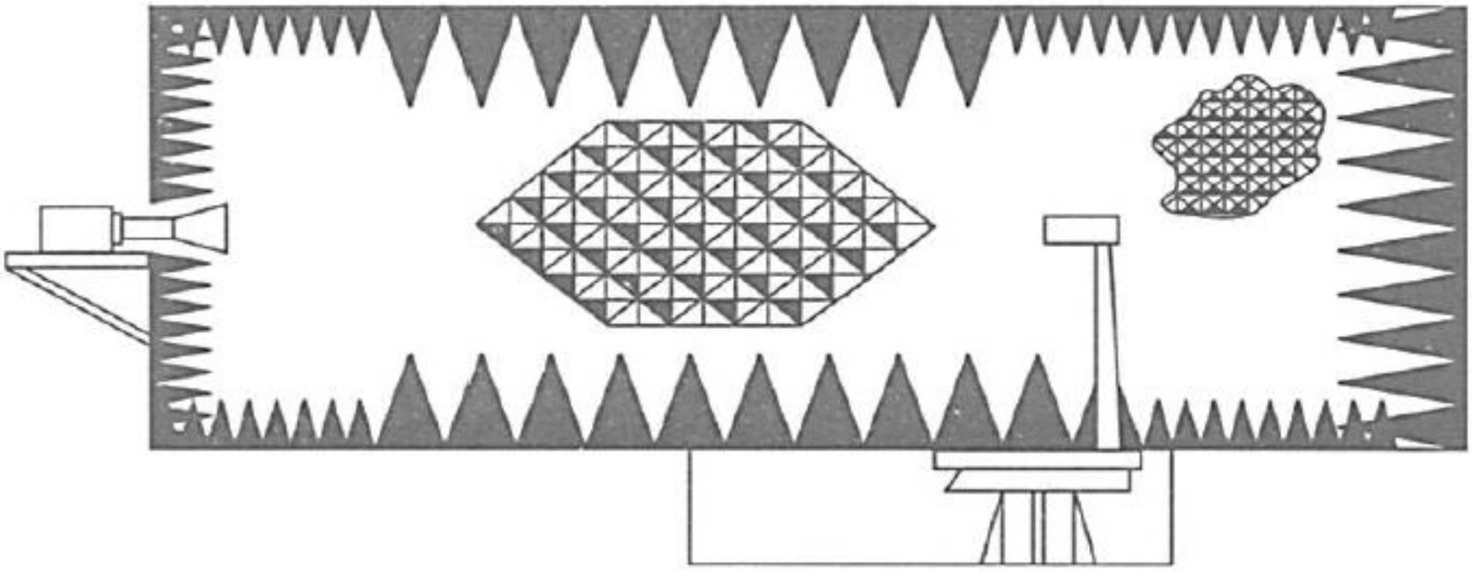
- For normal incidence the pyramid type absorber is the best option as they scatter as a random rough surface when large compared with wavelength.
- At higher frequencies, the reflection coefficient is larger and at lower frequencies, the thickness of the absorber should be larger.
- While the wedge shaped absorbers, with wedge direction along the plane of incidence, work perfectly at large angles of incidences but for normal incidence they cannot work satisfactorily compared with pyramidal absorbers.

Rectangular Chambers

- Figure 2 shows a longitudinal sectional view of a rectangular chamber in which the source antenna is located at the centre of one of the end walls.
- The location of the test antenna is at a point approximately equidistant from the side and back walls along the centre line of the chamber at the other end of the chamber with respect to the source antenna.
- The chamber is completely lined with microwave absorbing material. Still there will be reflections from the walls, floor and ceiling and the specular reflections reaching .the test antenna are the cause of concern.
- These arise from the regions midway between the source and test antennas on the side walls, floor, ceiling and also from the centre region of the back wall.

Rectangular Chambers 2

- For good absorption by the lining materials, the chamber width and height is designed such that the angle of incidence $\theta_i < 60°$.
- However, this requirement puts the restriction that the length to width ratio of the chamber be about 2 : 1 which is extended to 3 : 1 sometimes at the expense of higher levels of reflections.

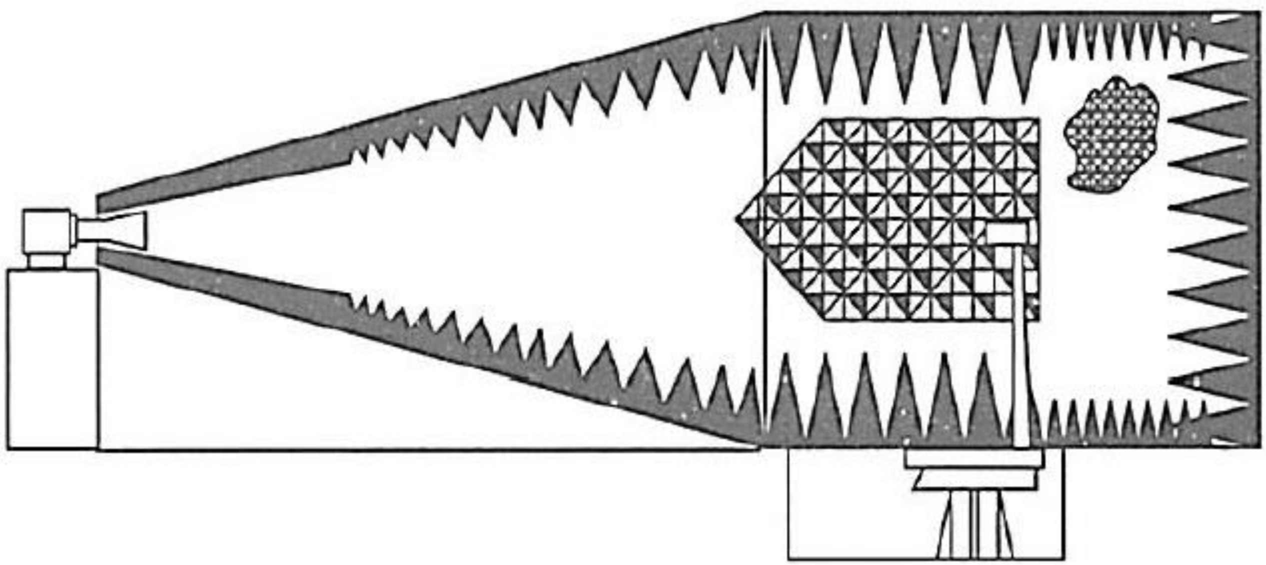


Rectangular Chambers 3

- The space in which the test antenna is located is termed the quiet zone.
- The volume of the quiet zone for a given chamber depends on the specified or allowable deviation of the incident field from a uniform plane wave.
- Rectangular chambers need bigger absorbing materials for frequencies below 1 GHz.
- It is difficult to obtain accurate measurements in these chambers mainly because it is usually not possible to obtain a source antenna with a sufficiently narrow beam width of these frequencies, to avoid illumination of the walls, floor and ceiling with the main beam.

Tapered Chambers

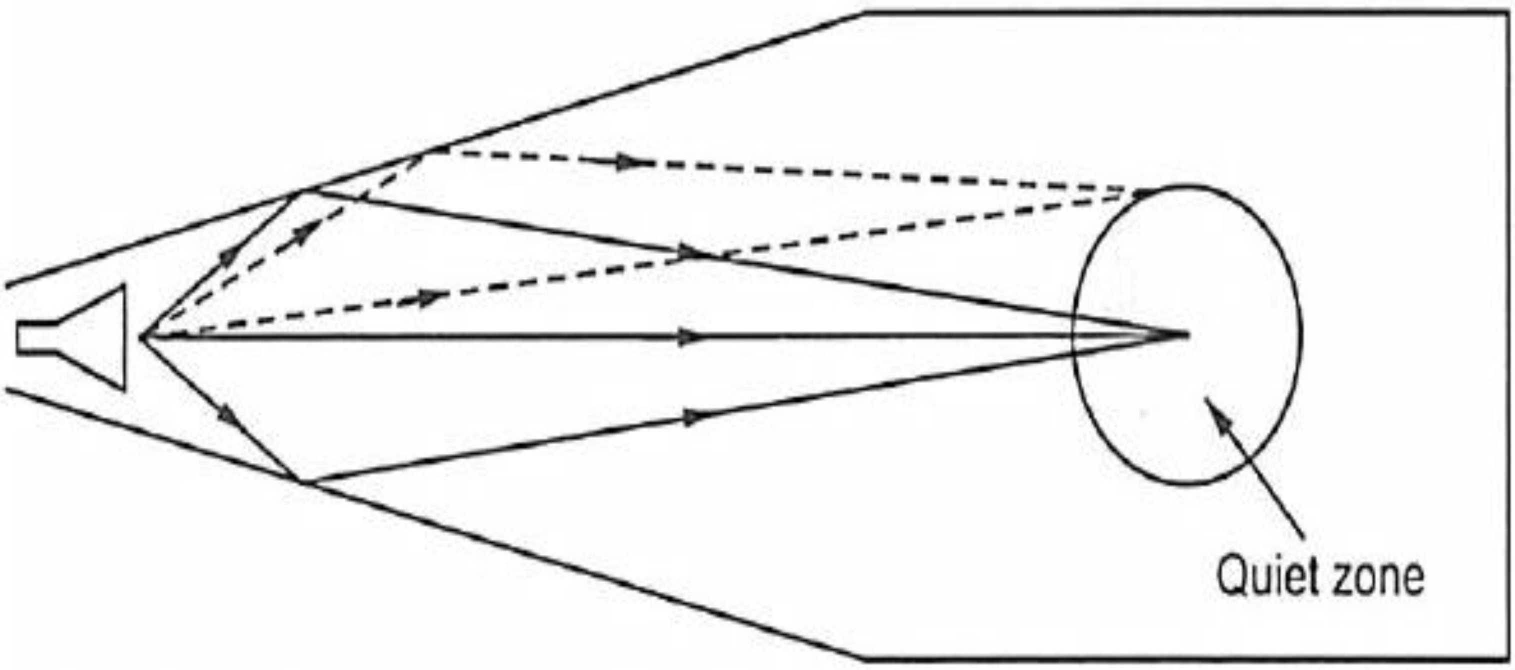
- The tapered anechoic chamber got introduced to overcome some of the limitations of the rectangular chamber, mentioned above.
- It consists of a tapered section opening into a rectangular section. The taper is shaped like a pyramidal horn that tapers from a small source end to a large rectangular test region. This construction is shown in Figure 3.



Tapered Chambers 2

- The rectangular section is approximately which and the tapered section is usually twice as long as the rectangular section.
- This geometry inherently requires less absorbing material which helps in substantially reducing the cost. In the tapered chamber, the specular reflections that reach the test region occur close to the source antenna as shown in Figure 3.
- However the path lengths of the reflected signals are not very different, electrically, from that of the direct signal which produces a slowly varying amplitude pattern which is beneficial since a constructive interference results.
- Also this allows use of thinner absorbing materials over the walls. The concept is illustrated in Figure 4.

Fig 4 / Tapered Chambers



Specular reflections that reach the quiet zone of a tapered anechoic chamber

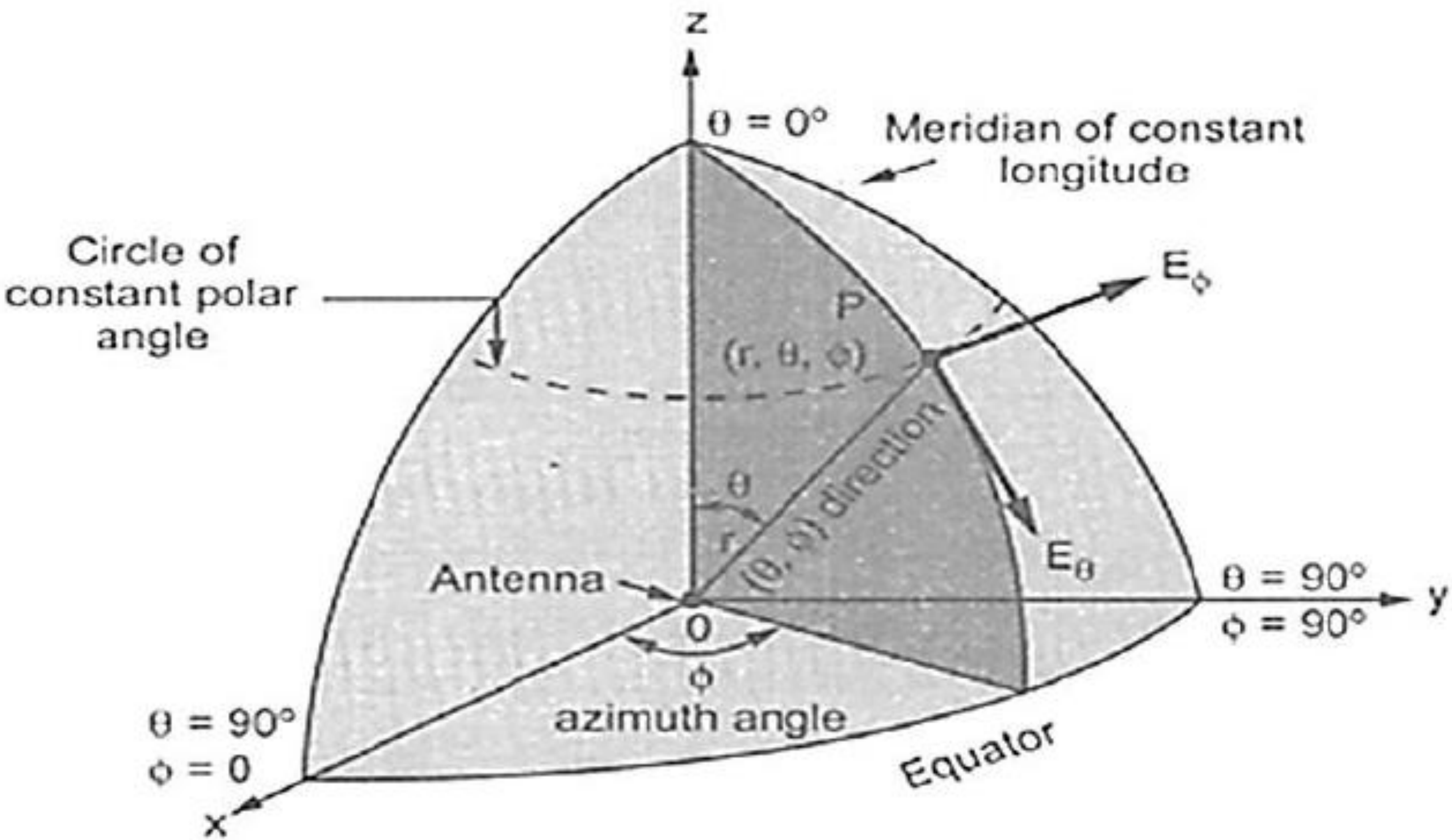
Topic 9

Antenna Measurements – Radiation Pattern

Introduction

- The radiation patterns, polarization, and gain of an antenna, which are used to characterize its radiation capabilities, are measured on the surface of a constant radius sphere.
- All these quantities are measured on the surface of a sphere with constant radius. Any point $'P'$ on such sphere can be described using spherical co-ordinate system as shown in Figure 1.
- The radiation characteristics of the antenna as a function of $\boldsymbol{\theta}$ and $\boldsymbol{\phi}$ for constant radius and frequency is called radiation pattern of an antenna.
- The minimum number of patterns required to construct a three dimensional pattern is 2 and they are selected as principle $\boldsymbol{E}$-plane and $\boldsymbol{H}$-plane patterns.
- The two dimensional pattern is generally called pattern cut.

Fig 1 / Radiation Pattern Measurement



Spherical co-ordinate system representation for radiation pattern measurement.

Introduction 2

- The pattern cuts can be obtained for one of the angles ($\boldsymbol{\theta}$ or $\boldsymbol{\phi}$) constant and varying the other. In most of the cases, the required patterns are horizontal plane i.e. x-y plane and vertical pattern in x-z plane.
- The radiation pattern of an antenna can be measured either in transmitting mode or receiving mode. For reciprocal antennas, even any mode is sufficient, receiving mode is selected.

Basic Procedure for Measurement

- For the measurement of radiation pattern of antenna, two antennas are required.
- One of the antennas in the system is the antenna under test, while the other illuminates the antenna under test and it is located away from the antenna under test.
- Thus one antenna is used in the transmitting mode, while other in the receiving mode.
- But according to the reciprocity principle, the radiation pattern will be same irrespective of the mode in which antenna is used.
- The antenna under test is usually referred as primary antenna, while the other one as secondary antenna.

First Procedure

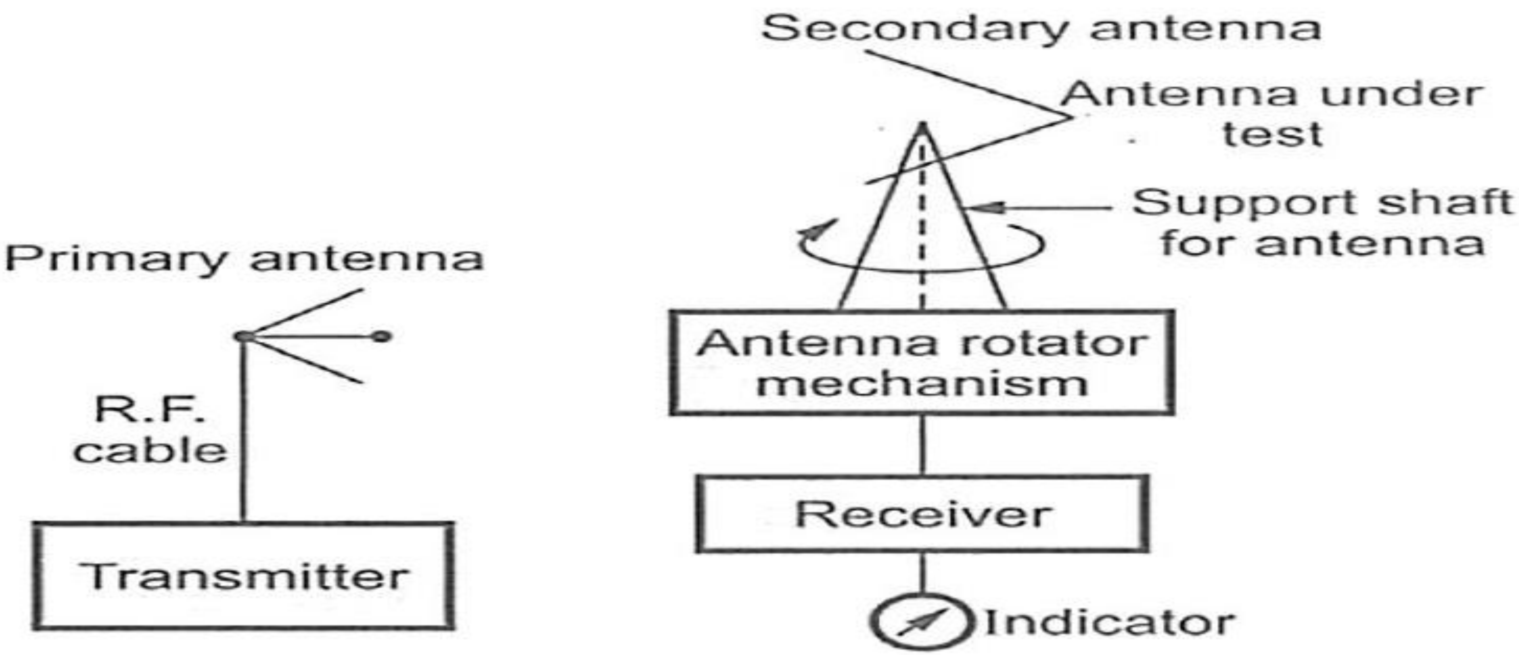
- The antenna under test i.e. primary antenna is kept stationary, while the secondary antenna is moved around the primary antenna along a circular path with uniform radius.
- If the secondary antenna is directional one, it is always aimed at the primary antenna.
- In this procedure, usually the primary antenna is transmitting.
- At different points, along the circular path, the readings of the field strength and direction with respect to the primary antenna are recorded.
- A plot of the radiation pattern of a primary antenna is plotted either as rectangular plot or polar plot.

Second Procedure

- Both the antennas are kept stationary with a suitable spacing between them. The secondary antenna is aimed at the primary antenna.
- The primary antenna is rotated about a vertical axis.
- In this procedure, the secondary antenna is used in the transmitting mode, so that the field strength reading and direction of the primary antenna with respect to the secondary antenna is made.
- The continuous readings at different points during rotation can be made using pattern recorder.
- Generally at low frequency, first procedure is used while at high frequency second one is preferred.

Set Up for Measurement

- The simple arrangement for the radiation pattern measurement consists primary antenna is transmitting mode, secondary antenna as antenna under test.
- The secondary antenna is coupled with the rotating shaft and it is rotated using antenna rotator mechanism. To measure the relative amplitude of the received field an indicator is used along with the receiver as shown in Figure 2.



Set Up for Measurement 2

- Usually the antenna under test is used in the receiving mode.
- It is properly illuminated by the stationary primary antenna.
- The secondary antenna is rotated about vertical axis.
- For E- plane pattern measurement, the antenna support shaft is rotated with both the antennas horizontal.
- While for H-plane pattern measurement, the shaft is rotated with both the antennas vertical.

Topic 10

Antenna Measurements – Gain

Introduction

- The gain and the directivity are usually measured in the direction of the pattern maximum. Their values in any other direction can be calculated from the radiation pattern.
- There are two techniques used for measuring the gain of an antenna-absolute gain measurement and gain transfer measurement.
- For the absolute gain measurement it is not necessary to have a prior knowledge of the gains of the antennas used in the measurement.
- The gain transfer method, requires the use of a gain standard with which the gain of the antenna under test is compared.

Absolute Gain Method

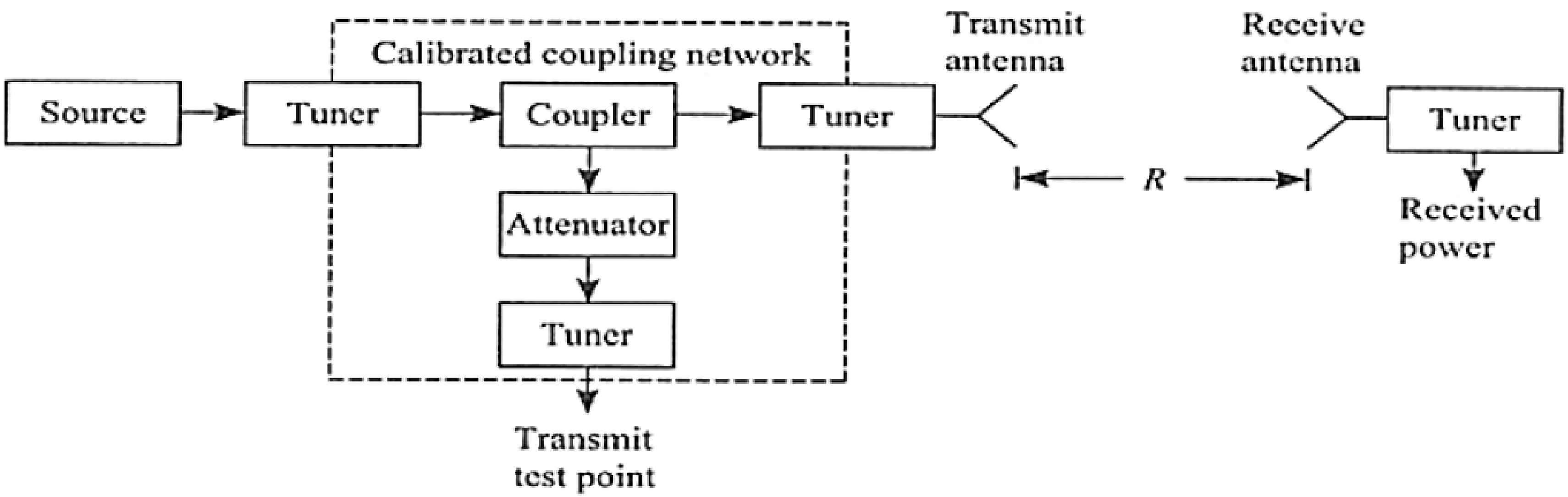
- Friis transmission formula forms the basis for absolute gain measurement. The Friis transmission formula expressed in decibels is ;

$$P_{rdBm} = P_{tdBm} + G_{tdB} + G_{r\,dB} + 20\log_{10}\left(\frac{\lambda}{4\pi R}\right)$$

- Consider two identical antennas placed in an elevated range or inside a rectangular anechoic chamber which are properly oriented and aligned such that (i) they are polarization matched and (ii) main beams of the two antennas are aligned with each other.
- With this arrangement, the gain in the direction of the maximum can be measured. The gain in any other direction can be computed from the radiation pattern.

Absolute Gain Method 2

- Let R be the separation between the two antennas chosen such that the antennas operate in the far-field region.
- Let λ be the wavelength corresponding to the operating frequency. A calibrated coupling network and a matched receiver unit, as shown in Figure 1, are used to measure the transmit and the receive powers P_{tdBm} and P_{rdBm} respectively. All the components are impedance matched using tuners.



Absolute Gain Method 3

- If the two antennas are identical, their gains are identical and above equation can be written as –

$$G_{tdB} = G_{r\,dB} = \frac{1}{2}\left[P_{rdBm} - P_{tdBm} - 20\log_{10}\left(\frac{\lambda}{4\pi R}\right)\right]$$

- and hence the gain of the antennas can be calculated.
- Since this method uses two antennas, it is known as *two-antenna method* for gain measurement.
- In the absence of two identical antennas, a third antenna is required to measure the gain. This is known as a *three-antenna method* of gain measurement

Gain Transfer Method

- The gain of the test antenna is measured by comparing with a standard gain antenna, of which the gain is known accurately.
- The test antenna is illuminated by a plane wave with its polarization matched to the transmitting antenna. The received power into a matched load, $Pr^T dB$ is then measured. Let G^T dB be the gain of the test antenna.
- From Friis formula ;

$$G_{tdB} + G_{dB}^{T} = P_{rdBm}^{T} - P_{tdBm} - 20\log_{10}\left(\frac{\lambda}{4\pi R}\right)$$

Now the test antenna is replaced by a standard gain antenna having a gain of G_{dB}^{S} and the received power P_{rdBm}^{S} is measured. Again from Friis transmission formula

$$G_{tdB} + G_{dB}^{S} = P_{rdBm}^{S} - P_{tdBm} - 20\log_{10}\left(\frac{\lambda}{4\pi R}\right)$$

Gain Transfer Method 2

- The gain of the test antenna can be calculated by subtracting the above two equations – $G_{dB}^{T} = G_{dB}^{S} + P_{rdBm}^{T} - P_{rdBm}^{S}$
- It is important that the polarization of the test antenna and the standard gain antenna need to be identical to each other and this should be matched with the polarization of the transmitter.
- Both antennas should be impedance matched to the receiver. This method is used to measure the gain of a linearly polarized antenna.

Gain Transfer Method 3

- The gain of a circularly polarized antenna can also be measured using a linearly polarized standard gain antenna.
- Since a circularly polarized wave can be decomposed into two orthogonal linear components, we can use a linearly polarized antenna to measure the gains of each of these components and then the total gain is obtained by combining the two.

Topic 11

Antenna Measurements – Directivity

Procedure for Measurement

- Sometimes it is found that the directivity of the antenna cannot be calculated using the analytical techniques alone.
- So the directivity can be obtained from the radiation pattern of the antenna. The following procedure is considered ;
  - Measure the two principal E- and H-plane patterns of the test antenna.
  - Determine the half-power beamwidths (in degrees) of the E- and H-plane patterns.
  - Compute the directivity using the formula

$$D_0 = \frac{4\pi}{\Omega_A}$$

$$\Omega_A = \theta_{HP}\phi_{HP}$$

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

Computation

Directivity can also be computed using the formula ;

$$D = \frac{4\pi U}{P_{rad}}$$

where ; P_{rad} = Total power radiated in the direction (θ, ϕ)

$U(\theta, \phi)$ = Radiation intensity in the direction (θ, ϕ)

The total power radiated by the antenna is obtained by integrating the radiation intensity over the entire solid angle of 4π. Thus,

$$P_{rad} = \oiint_{\Omega} U(\theta, \phi)\, d\Omega = \int_0^{2\pi} \int_0^{\pi} U(\theta, \phi) \sin\theta \, d\theta \, d\varphi$$

Topic 12

Antenna Measurements – Polarization

Polarization pattern method

- **Polarization Pattern Method :** This method can be used to measure the AR and the tilt angle τ of the polarization ellipse but not the sense of polarization as shown in Figure 1.
- The test antenna is connected as the source antenna while a linearly polarized antenna such as a dipole antenna is used to receive the power at different rotation angles.
- The square root of the received power plotted against the rotation angle ψ indicate the AR and title τ.

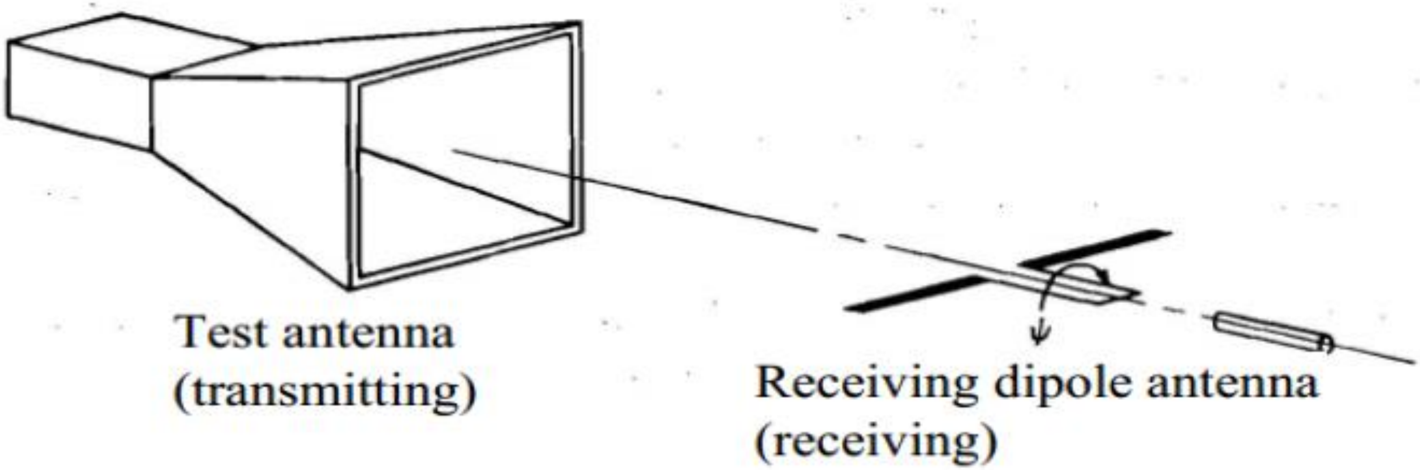
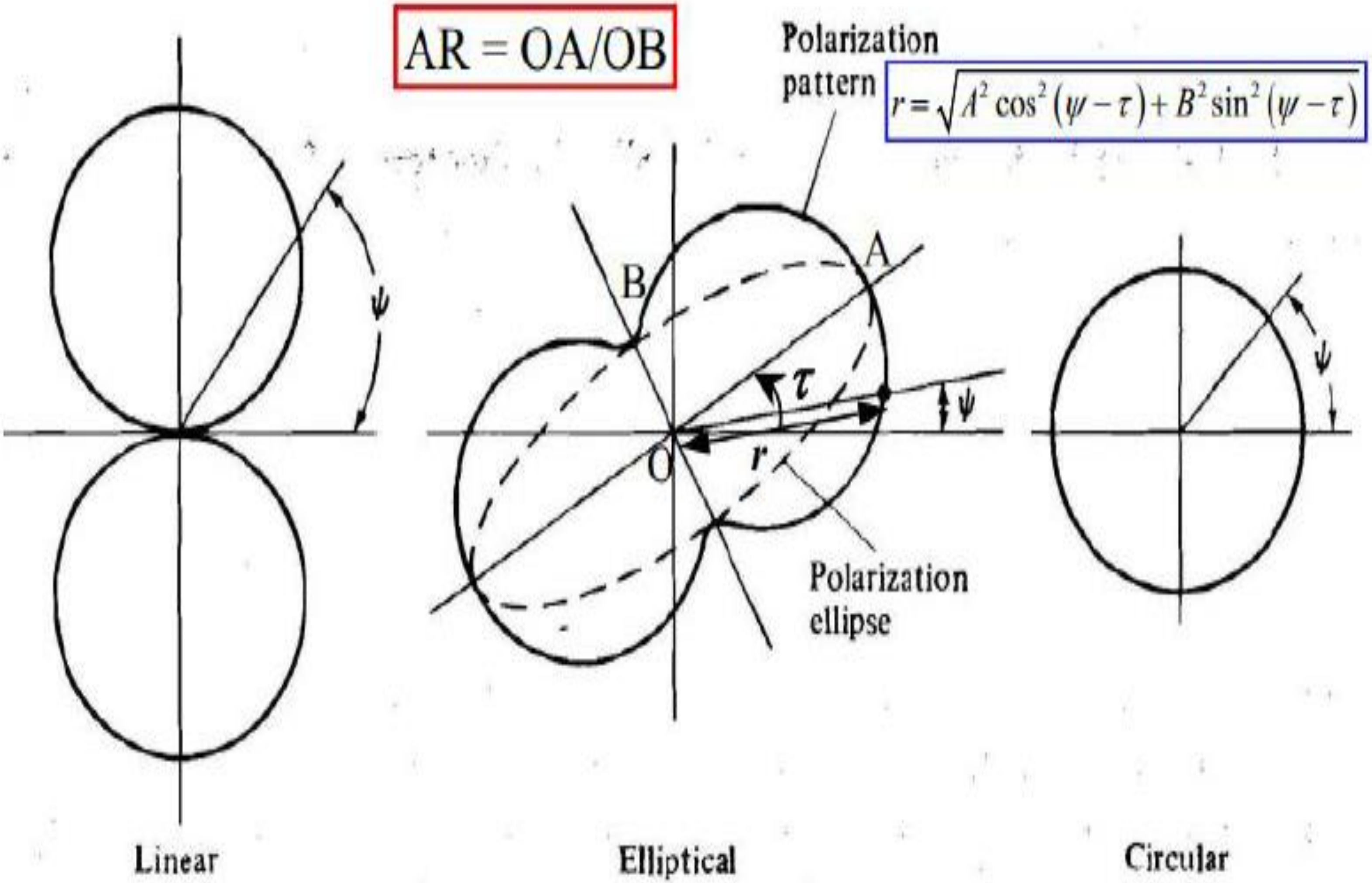


Fig 2 / Polarization Pattern Method



Topic 13

Antenna Measurements – VSWR

VSWR

- The input characteristics of an antenna such as the input impedance Z_A can be measured by a network analyzer.
- The advantage of a network analyzer is its ability to measure both the magnitude and the phase of the power received.
- Reflection Coefficient Measurement: The reflection coefficient ρ(or S_{11}) of an antenna can be obtained from its input impedance measurement.

$$\rho \text{ or } S_{11} = \frac{Z_A - Z_0}{Z_A + Z_0} \qquad \text{(dimensionless)}$$

- VSWR Measurement: The VSWR of an antenna can be obtained from its reflection coefficient measurement.

$$\text{VSWR} = \frac{1+|\rho|}{1-|\rho|} \qquad \text{(dimensionless)}$$

UNIT – V

PROPAGATION OF RADIO WAVES

Topic 1

Modes of Propagation

- Electromagnetic Waves generated by the radiated power from the current carrying conductor.
- In conductors, a part of the generated power escapes and propagates into free space in the form of **Electromagnetic wave**, which has a time-varying electrical field, magnetic field, and direction of propagation orthogonal to each other.
- Radiated from an **isotropic transmitter,** these wave travels through different paths to reach the receiver.
- The path taken by the wave to travel from the transmitter and reach the receiver is known as **Wave Propagation.**
- The mode of propagation of electromagnetic waves in the atmosphere and in free space may be divided in to the following three categories –
- Line of sight (LOS) propagation
- Ground wave propagation
- Sky wave propagation

- In ELF (Extremely low frequency) and VLF (Very low frequency) frequency bands, the Earth, and the ionosphere act as a wave guide for electromagnetic wave propagation.
- In these frequency ranges, communication signals practically propagate around the world.
- The channel band widths are small.
- Therefore, the information is transmitted through these channels has slow speed and confined to digital transmission.

FACTORS INVOLVED IN RADIO WAVE PROPAGATION

Radio wave or electromagnetic wave when travels from transmitter to receiver, many factors influence the propagation of wave.

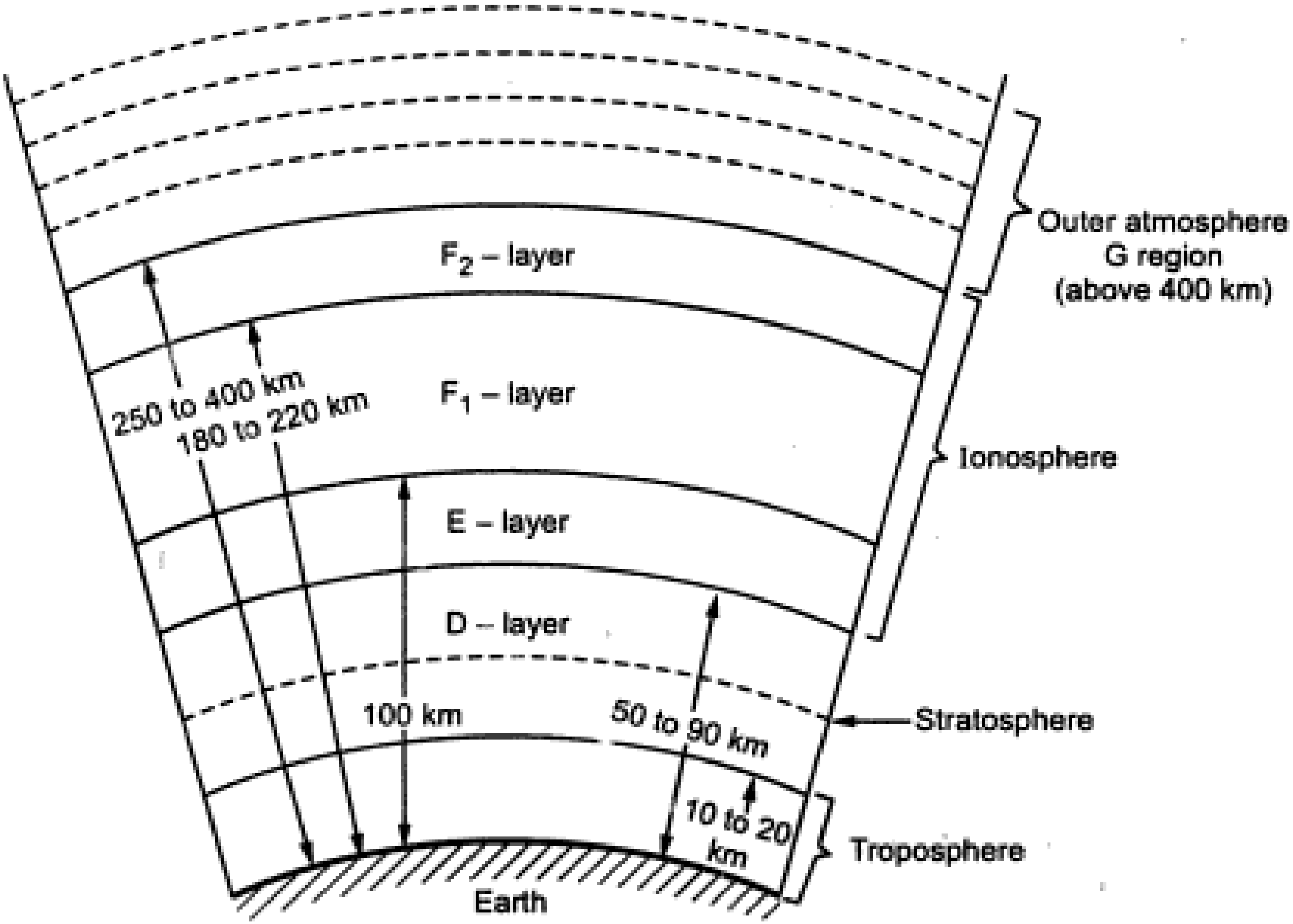
Some of the important factors are as follows:

- Characteristics of earth such as conductivity, permittivity, permeability.
- Curvature of the earth, magnetic field of the earth, roughness of the earth.
- Frequency of operation.
- Height and polarization of transmitting antenna and transmitter power.
- Characteristics of ionospheric regions.
- Distance between transmitter and receiver.
- Refractive index and permittivity of troposphere and ionospheric regions.

Topic 2

Structure of Atmosphere

Structure of Atmosphere



- In the radio wave propagation, the earth's environment between the transmitting and receiving antennas play very important role.
- The atmosphere of the earth mainly consists of three regions namely
  - Troposphere
  - Stratosphere
  - Ionosphere
- In 1925, Sir Edward Appleton showed that propagation of the radio waves at high frequencies is greatly supported by the upper part of the atmosphere of the earth.

Structure of Troposphere

- This is the nearest region in the atmosphere from the earth's surface and is around 10 km to 20 km above the earth's surface.
- But the height of the troposphere region slightly varies at the poles and the equators. Its height is least at the poles while maximum at the equators.
- The gas components in the troposphere remain almost constant in percentage with increase in height. But the water vapour components drastically decrease with increasing height.

- The significant property of the tropospheric region is that temperature decreases with increase in the height.
- The troposphere is also called region of change. At a certain height called critical height above troposphere, the temperature remains constant for narrow region and then increases afterwards.
- This region between the top of troposphere and the beginning of the stratosphere is called tropopause.

Structure of Stratosphere

- The region between 20 km to 50 km above the earth's surface is called region of calm or stratosphere.
- It is dense part of the atmosphere. It absorbs UV rays because of the presence of Ozone layer.
- The stratosphere has relatively little effect on radio waves because it is calm region with little or no temperature changes.

Structure of Ionosphere

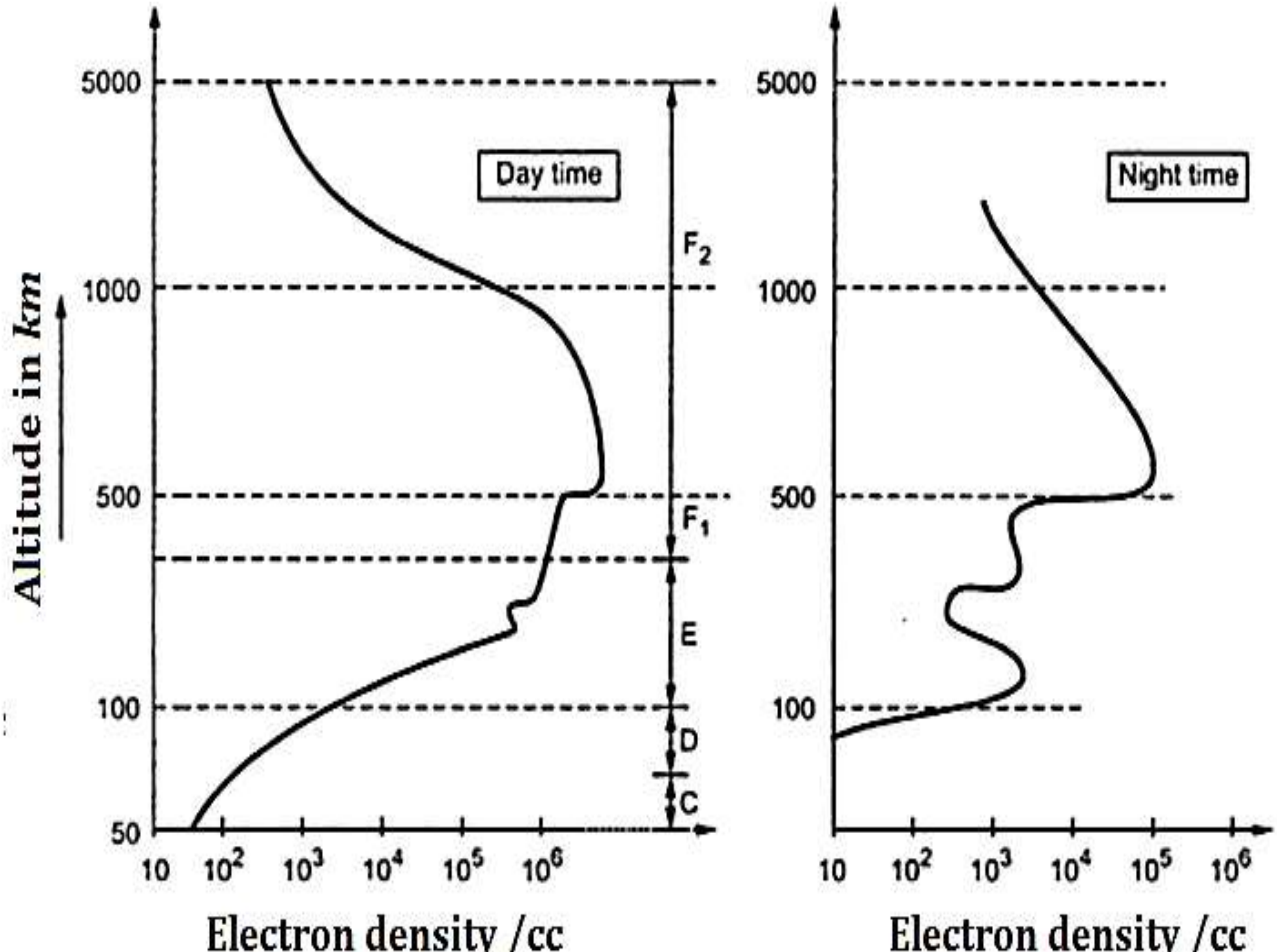
- The radiation from the space, in particular that from the sun, ionizes the gas molecules present in the atmosphere.
- The ionized layer that extends from about 50 km above the surface of the earth to several thousand kilometers is known as the ionosphere.
- At great heights from the surface of the earth the intensity of the ionizing radiation is very high, but there are very few molecules to be ionized.

- Therefore, in this region the ionization density (number of electrons or ions per unit volume) is low.
- As the height is decreased, atmospheric pressure increases, which implies that more molecules are present in the atmosphere.
- Therefore, the ionization density increases closer to the surface of the earth.

- With further reduction in height, though the number of molecules keeps increasing, the ionization density reduces because the energy in the ionizing radiation has been used up or absorbed to create ions.
- Therefore, the ionization density has a maximum that exists neither at the surface of the earth nor at the outer periphery of the ionosphere, but somewhere in the middle i.e., between 50 km to 400 km.
- It has been observed that the electron density profile (electron density versus height), has regions of maxima as well as regions of constant density (Fig.2). These regions are known as layers of the ionosphere.

- There are mainly three layers in the ionosphere designated by the letters D, E, and F. The F layer splits into separate layers $F1$ and $F2$ during day time.
- The F layer is also called as Appleton layer and it is ionized during day time as well as night time.
- The E layer is also called as Kennelly-Heaviside layer.
- The D layer, which is present only during the day time, does not reflect high frequency electromagnetic waves ($2 - 30\ MHz$), but attenuates the waves passing through it.

- Even though the D layer reflects lower frequency waves ($< 1\ MHz$), due to the high absorption of the electromagnetic energy by the D layer, the utility of the reflected waves is limited.
- The E and F layers, which are present during both day and night times, make long distance communication possible by reflecting radio waves in the frequency range of $2–30\ MHz$. Radio waves above $30\ MHz$ pass through the ionosphere.



Altitude in km
5000
1000
500
100
50
Day time
F2
F1
E
D
C
10
10^2
10^3
10^4
10^5
10^6
Electron density /cc
Night time
5000
1000
500
100
10
10^2
10^3
10^4
10^5
10^6
Electron density /cc

- Along with the *E layer, there exists the* Es *layer which has very high ionization density.*
- *This is known as sporadic E layer and it exists during night time also.*
- *It is not important layer from the point of view of long distance communication. But it provides sometime better reception during night.*
- The region lower to D region, where peak of the electron density is called *C layer.*
- The region at height 400 km above the earth's surface is called G region.

- Even though the upper limit of the ionosphere is not exactly known, the outer atmosphere is nothing but G region which consists of the charge particles trapped by the terrestrial magnetic field having shape similar to that of the magnetic lines of force.
- This region is occupied by the radiation belts girdling the earth.

- The number of layers in the ionosphere, their heights and the amount of sky wave that can bend by them will vary from day to day, month to month and year to year.
- For each layer there is a critical frequency, above which if radio wave is sent vertically upward, will not return back to the earth, but will penetrate it

Topic 3

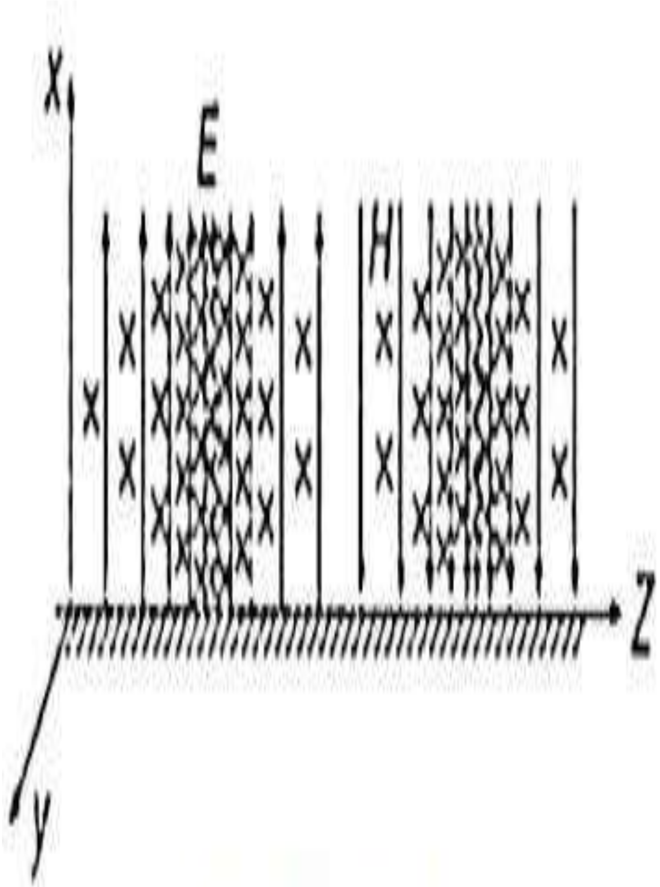
Ground Wave Propagation

GROUND WAVE PROPAGATION

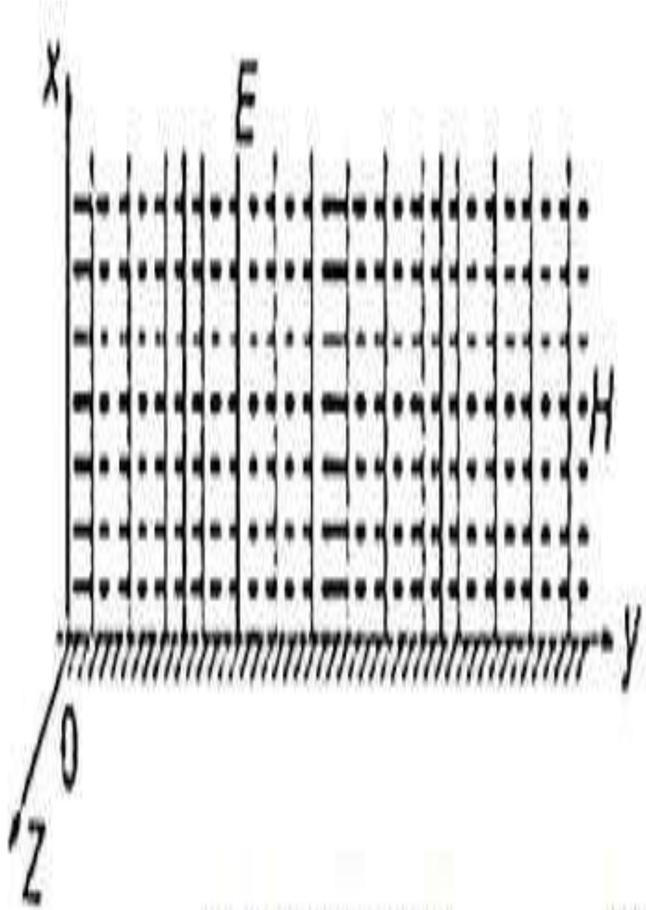
- The waves, which while traveling, glide over the earth's surface are called *ground waves.*
- *The ground wave is also called surface wave as the wave passes over the surface of the earth.*
- *Ground waves are always vertically polarized (produced by vertical antennas) .*
- The vertical antennas are the antennas in which the electromagnetic waves are vertically polarized i.e., electric field vectors of electromagnetic waves are vertical with respect to ground.

- Any horizontal component of the electric field vectors in contact with the ground gets short circuited.
- When the ground waves propagate along the surface of the earth, the charges are induced on the earth's surface.
- The number and polarity of these charges keep on changing with the intensity and location of the wave field.
- This variation causes the constitution of a current.

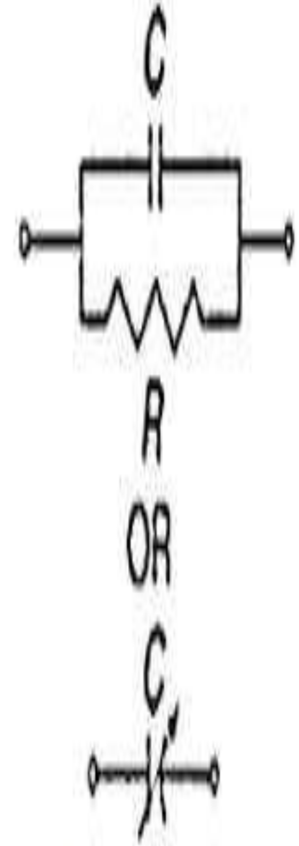
- In carrying this current, the earth behaves like a leaky capacitor.
- As the wave travels over the surface, it gets weakened due to absorption of some of its energy.
- This absorption, in fact, is the power loss in the earth's resistance due to the flow of current.
- This energy loss is partly replenished by the diffraction of energy, downward, from the portion of the wave present some what above the immediate surface of the earth. This process is shown in Fig.



(a) Side view



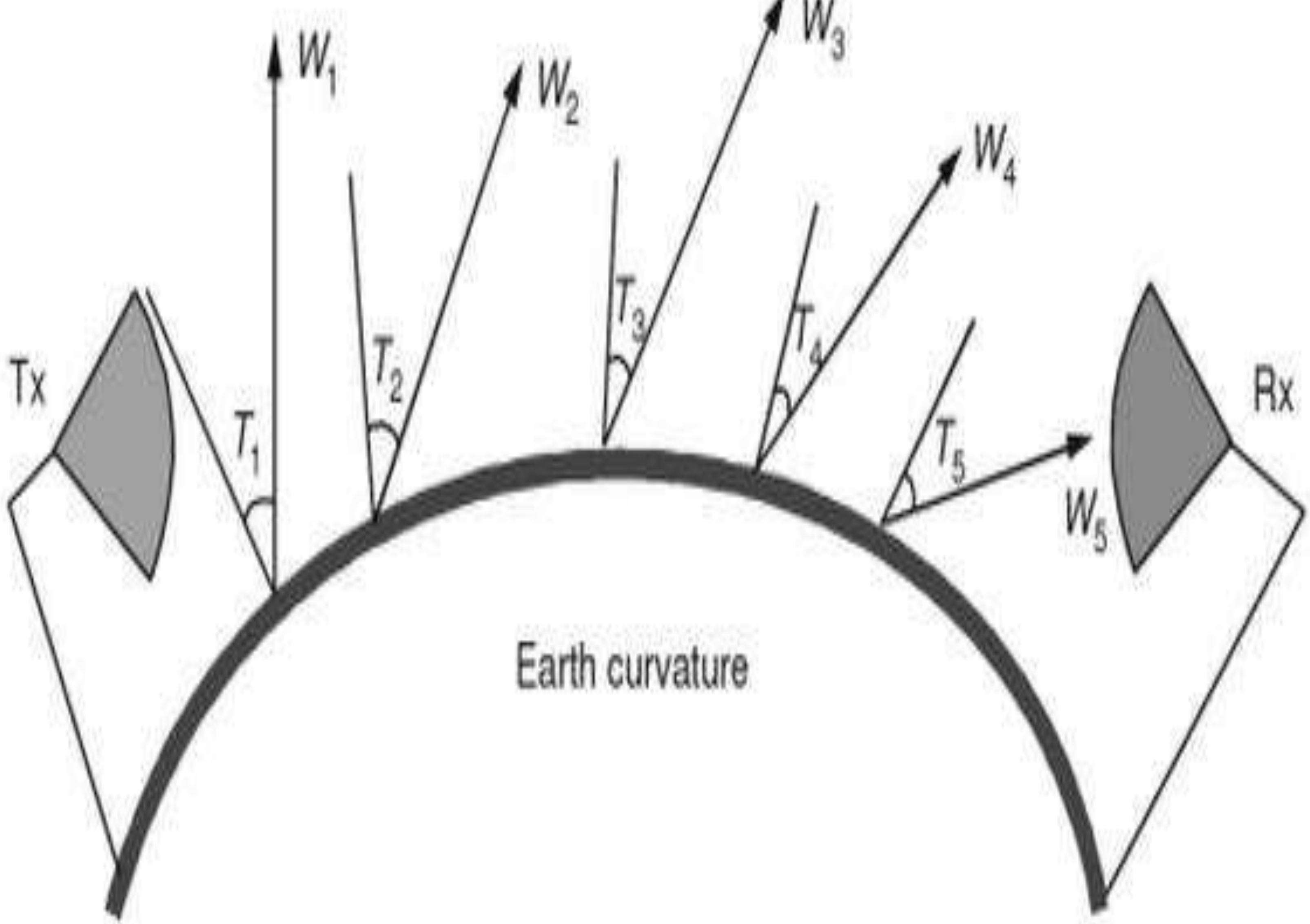
(b) Front view



(c) Earth represented by a leaky capacitor

- The earth's attenuation increases as frequency increases. So this mode of propagation is suitable for low and medium frequency i.e., upto 2 MHz only.
- It is also called as medium wave propagation. All the broadcast signals received during day time is due to ground wave propagation.
- Along with the ground attenuation, the ground waves or surface waves are suffered due to the diffraction and tilt in the wave front.
- As the ground wave propagates over a surface of the earth, the wave front gradually tilts more and more. As the wave front tilts more and more, the more electric field component gets short circuited.

- Hence the strength of the signal gradually decreases with increase in the tilt.
- At a particular distance from the transmitter, the ground wave completely dies due to the attenuation as a result of more and more tilt of the wave front.
- The phenomenon of wave tilting in successive wave front is shown in below Fig. in which $T1$, $T2$, $T3$, $T4$ and $T5$ are the tilting angles in increasing order and $W1$, $W2$, $W3$, $W4$ and $W5$ wave fronts.



W1
W2
W3
W4
W5
T1
T2
T3
T4
T5
Tx
Rx
Earth curvature

In general, surface of the earth is considered to be a plane if the distance between the transmitters and the receiver is less than the minimum barrier distance d given by expression ;

$$d = \frac{50}{(f_{MHz})^{\frac{1}{3}}} \quad in\ miles$$

Salient Features of Ground Wave Propagation

- The ground waves propagate along the surface of the earth.
- When the ground waves propagate along the surface of the earth, the charges are induced on the surface of the earth. These charges travel along the wave and hence the current gets induced.
- While carrying induced current, the earth acts as a leaky capacitor.
- The ground waves are produced in vertically polarized antennas which are placed very close to surface of the earth.

- The ground waves are important at broadcast and lower frequencies. These can be used up to 2 MHz.
- According to the characteristics of the earth, the strength of ground wave varies. These waves are not affected by the changes in the atmospheric conditions.
- The variations in surface or type of the earth affect propagation losses considerably.
- The maximum range of ground wave propagation depends -on the frequency and power of the transmitter.

Topic 4

Sky Wave Propagation

SKY WAVE PROPAGATION

- The sky wave propagation is very important from the point of view of long distance radio communication.
- In this mode of propagation, the electromagnetic waves reaching the destination point first get reflected by the region of ionized gases in the upper atmosphere region which is situated between 50 km to 400 km above earth's surface.
- Hence, this mode of wave propagation is called ionospheric propagation.

- The ionosphere acts like a reflecting surface and reflect back the electromagnetic waves of frequencies between 2 MHz to 30 MHz.
- Electromagnetic waves of frequency more than 30 MHz are not reflected back from ionosphere but they penetrate it.
- This mode is most effective from the frequencies between 2 MHz to 30 MHz , hence this mode is also commonly called short wave propagation.

- With the sky wave propagation, a long distance point to point communication is possible.
- The main advantage of the sky wave propagation is that the long distance communication is possible with the help of multiple reflections of the sky waves.
- But these signals are affected by fading in which the strength of the signal varies with time

Characteristics of Ionosphere

1. Characteristics of $\boldsymbol{D}$ layer

- It is the lowest layer of the ionosphere at a height of 50 km to 90 km.
- Its thickness is about 10 km.
- It exists only in day-time and disappears in night time.
- Its ionization properties depend on the altitude of the sun above the horizon.
- It is not useful layer for HF communication.
- It reflects some VLF and LF waves.
- It absorbs MF and HF waves to some extent.
- Its electron density is 400 $electrons/cc$.
- Critical frequency of the layer is 100 kHz.

2. Characteristics of $\boldsymbol{E}$ layer

- It exists next to D layer at an average height of 100 km.
- Its thickness is about 25 km. ⍰ It reflects some HF waves in day-time.
- Its electron density is $5 \times 105 electrons/cc$.
- Its critical frequency is 4 MHz.

3. Characteristics of Es layer

- It is a sporadic E-Layer.
- Its appearance is sporadic in nature.
- It exists in both day and night.
- It is a thin layer and its ionization density is high.
- It appears close to E-Layer.
- If it appears, it provides good reception.
- It is not a dependable layer for communication.

4. Characteristics of $\boldsymbol{F1}$ layer

- It exists at a height of about 180 km in day-time.
- Its thickness is about 20 km.
- It combines with $F2$ layer during nights.
- HF waves are reflected to some extent.
- It absorbs HF to a considerable extent.
- It passes on some HF waves towards $F2$ layer.
- Its critical frequency is 5 MHz.
-

5. Characteristics of $\boldsymbol{F2}$ layer

- It is the most import layer for HF communication.
- Its average height is about 325 km in day-time.
- Its thickness is about 200 km.
- It falls to a height of 300 km at nights as it combines with the $F1$ layer.
- It is the topmost layer of the ionosphere.
- It is highly ionized and offers better HF reflection.
- Electron density is 2×10^6 $electrons/cc$.
- Its critical frequency is 8 MHz in day time and 6 MHz at nights

Topic 5

Space Wave Propagation

SPACE WAVE PROPAGATION

- Space waves are useful in the frequency range of 30 MHz to 300 MHz.
- It is used in FM, TV and radar applications.
- In this propagation, wave propagates within the troposphere.
- It is the lowest portion of the atmosphere.

- Space wave consists of two components i.e. direct wave (line-of-sight, LOS) and indirect wave.
- Even though, both the wave namely direct wave and indirect wave are transmitted at the same time, with same phase,
- at the receiving end they may reach in phase or out of phase depending on the different path lengths

- Thus at the receiving end, the signal strength is the vector addition of the strengths of the direct and indirect waves.
- When the two waves are in phase, the strength of the signal at the receiver will be stronger.
- Similarly if the two waves are out of phase, the strength of the signal at the receiver will be weaker.
- The space wave propagation is mainly used in VHF (Very High Frequency) band as both previous modes namely ground wave propagation and sky wave propagation both fail at very high frequencies.

Topic 6

Tropospheric Propagation

TROPOSPHERIC PROPAGATION

- The tropospheric region extends from the surface of the earth to a height of about 10 km at the poles and 18 km at the equator.
- The temperature of this region decreases with height at the rate of about 6.5°C per km and falls a minimum value about −52°C at its upper boundary.
- In this region, the clouds are formed.
- Next to the troposphere, stratosphere exists.

- The propagation through the troposphere takes place due to mechanisms such as diffraction, normal refraction, abnormal reflection and refraction and tropospheric scattering.
- In troposphere, slight bending of radio waves occurs and causes signals to return to earth beyond the geometric horizon.
- Troposphere bending is evident over a wide range of frequencies, although it is most useful in the VHF and UHF regions.
- Radio signals can be trapped in the troposphere, travelling a longer distance than normal before coming back to the earth surface.

- Instead of gradual changes in the atmospheric conditions, sometimes distinct regions are formed and regions that have significantly different densities try to bend radio waves passing between regions.
- However, in a non-homogeneous atmosphere whose index of refraction decreases with height, rays of sufficiently small initial elevation angle are refracted downward with a curvature proportional to the rate of decrease of the index of refraction with height.

- Out of different mechanism of troposphere wave propagation diffraction, abnormal reflection and refraction, and troposphere scattering, the normal refraction is the main mechanism for most of troposphere propagation phenomenon.
- The dielectric constant (hence refractive index) of the atmosphere which varies above the earth and set mostly by the moisture contains is a primary factor in the troposphere refraction.
- When the wave passes between mediums of different densities, its path bends by an amount proportional to the difference in densities.
- Especially, at UHF and microwaves two cases of tropospheric propagation are observed.

Topic 7

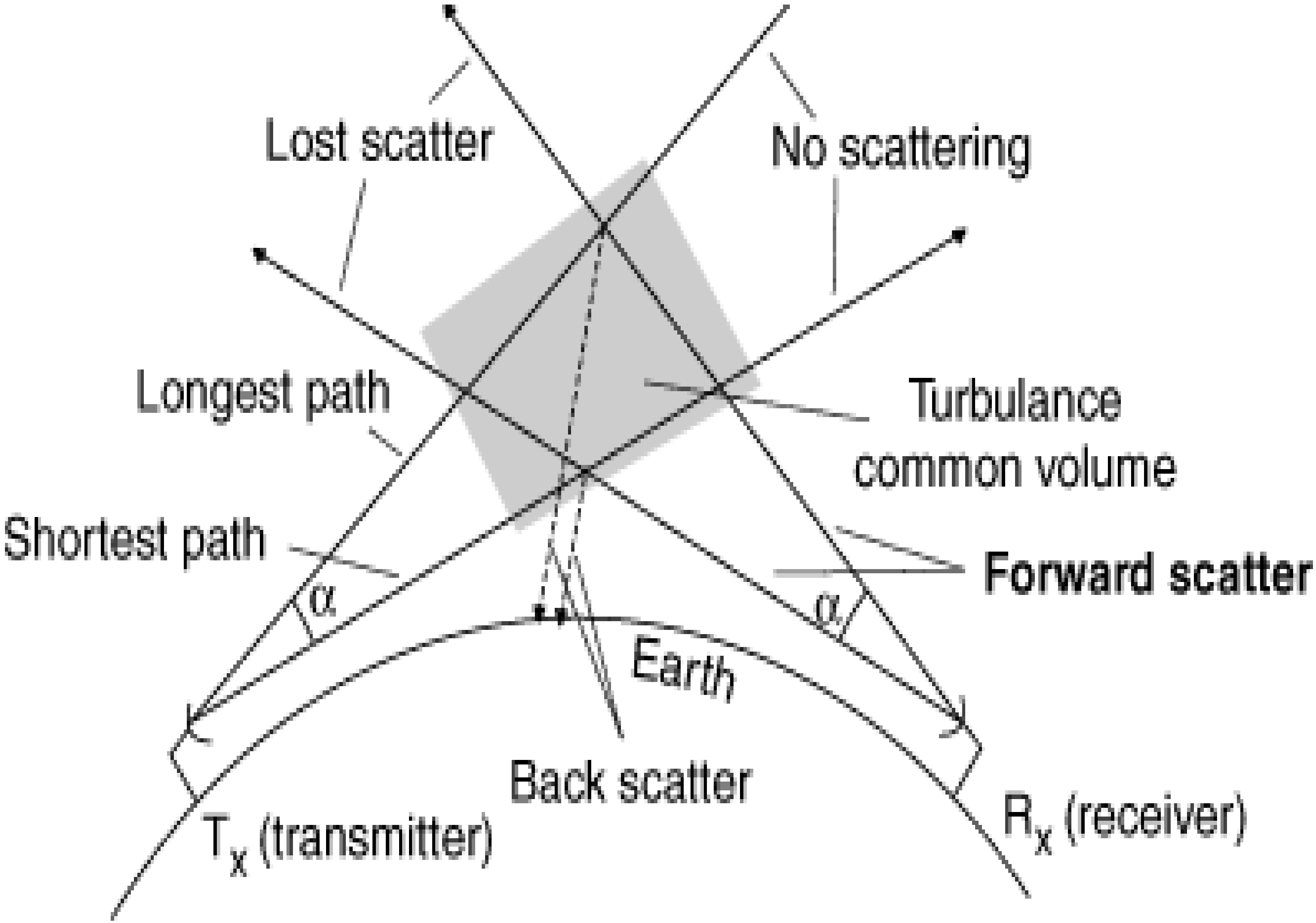
Tropo-scatter Propagation

TROPOSCATTER PROPAGATION

- Troposcatter is a mechanism by which propagation is possible by the scatter and diffracted rays.
- The scattering takes place in the tropospheric region.
- This mode of propagation occurs in VHF, UHF and microwave band.
- UHF and microwaves signals were found to be propagated much beyond the line of sight propagation through the forward scattering in the tropospheric irregularities.
- This mechanism helps to get unexpectedly large field strengths at the receivers even when they are is shadow zone.
- It is possible to achieve a very reliable communication over a range of 160 km to 1600 km by using high power transmitter and high gain antennas.

- The tropospheric scattering phenomenon can be used to establish a communication link over a distance much beyond the radio horizon.
- The troposphere can scatter electromagnetic waves due to its in-homogenous nature.
- The tropospheric scattering has been attributed to the blobs of refractive index changes and turbulence.
- These could be due to sudden changes in the temperature or humidity or the presence of dust particles.

- Waves passing through such turbulent regions get scattered. When λ is large compared to the size of the turbulent eddies, waves scatter in all the directions.
- When λ is small compared to these irregularities then most of the scattering takes place within a narrow cone surrounding the forward direction of propagation of the incident radiation.



Lost scatter
No scattering
Longest path
Turbulance
common volume
Shortest path
Forward scatter
α
α
Earth
Back scatter
T_X (transmitter)
R_X (receiver)

- When the wavelength is small (frequency is high) than the eddies, forward scattering dominates into the cone of angle α. The angle α should be very small.
- To receive scattered signal at a point well beyond the horizon, the transmitting and receiving antennas must be of high gain and must be so oriented that their beams overlap in a region where forward scattering is taking place.

- The scattering angle should also be as small as possible. This process is shown in Fig.
- Since the scattering process is of random nature, the scattered signals continuously fluctuate in amplitude and phase over a wide range.
- Troposcatter can be used to establish communication links in the *UHF and microwave frequency bands.*
- *These links typically have a range of up to a thousand kilometers and can have bandwidths of a few MHz.*
- *Troposcatter links can be used in multi-channel telephony and television applications.*

Features Of Troposcatter Propagation

- It is useful for propagation in the range of 100 MHz to 10 GHz.
- It produces undesirable noise and fading which may be minimized to certain extent by diversity reception.
- The field strength received is usually on the order of d1/7 or d1/8 where d is the distance between the transmitter and receiver.
- Since the signal strength is very weak, high gain antennas are required for reception.
- The propagation exhibits seasonal variation.
- The forward scatter propagation is useful for point to point communications and radio or television relay links.

Topic 8

Duct Propagation

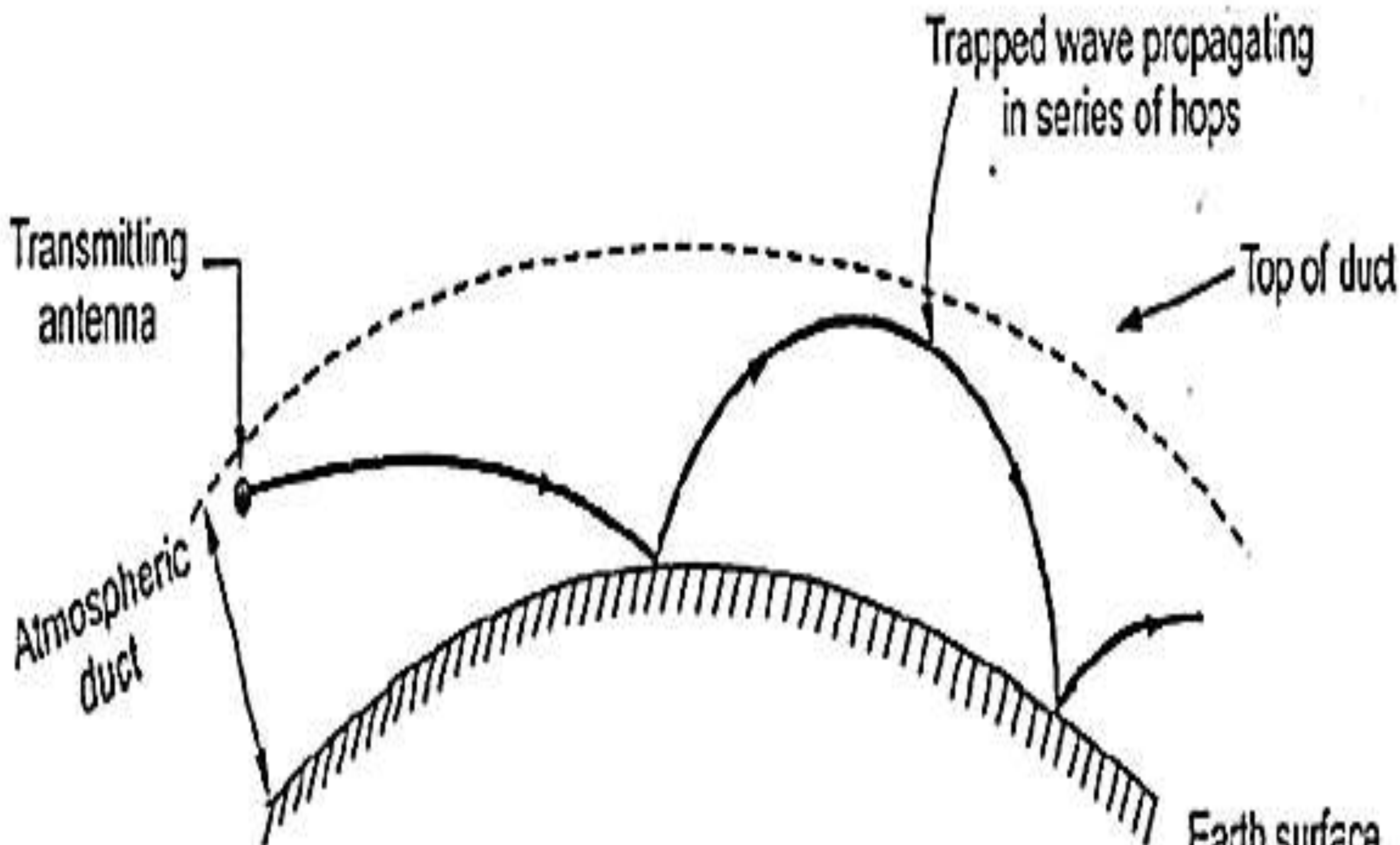
DUCT PROPAGATION

- The VHF, UHF and microwave frequencies are the frequencies which are neither propagated along the surface of the earth nor reflected by ionosphere.
- But in the troposphere region, the high frequency waves are refracted and transmission takes far beyond line-of-sight (LOS) distance.

- An atmosphere where the dielectric constant is assumed to decrease uniformly with height to value equal to unity at which air density is supposed to be zero is commonly called normal atmosphere or standard atmosphere.
- There are different air regions or layers one above other with different temperatures and water vapour contents.
- In one of the regions, there is a region where dN/dh is negative. In this region, the curvature along which the radio waves pass is slightly greater than that of the earth.

- Due to this, the wave originally directed almost parallel to the surface of the earth gets trapped in such regions.
- The energy originating in this region propagates around curved surfaces in the form of series of hops with successive reflections from the earth as shown in the below Fig.

- This phenomenon is called super refraction or duct propagation. Two boundaries of surfaces between two air layers form a duct which guide the radio waves between walls i.e. boundaries.
- The concepts like line of sight and diffraction cannot be applied when the wave propagates through duct and it is found that the energy travels high distances round the earth without much attenuation.



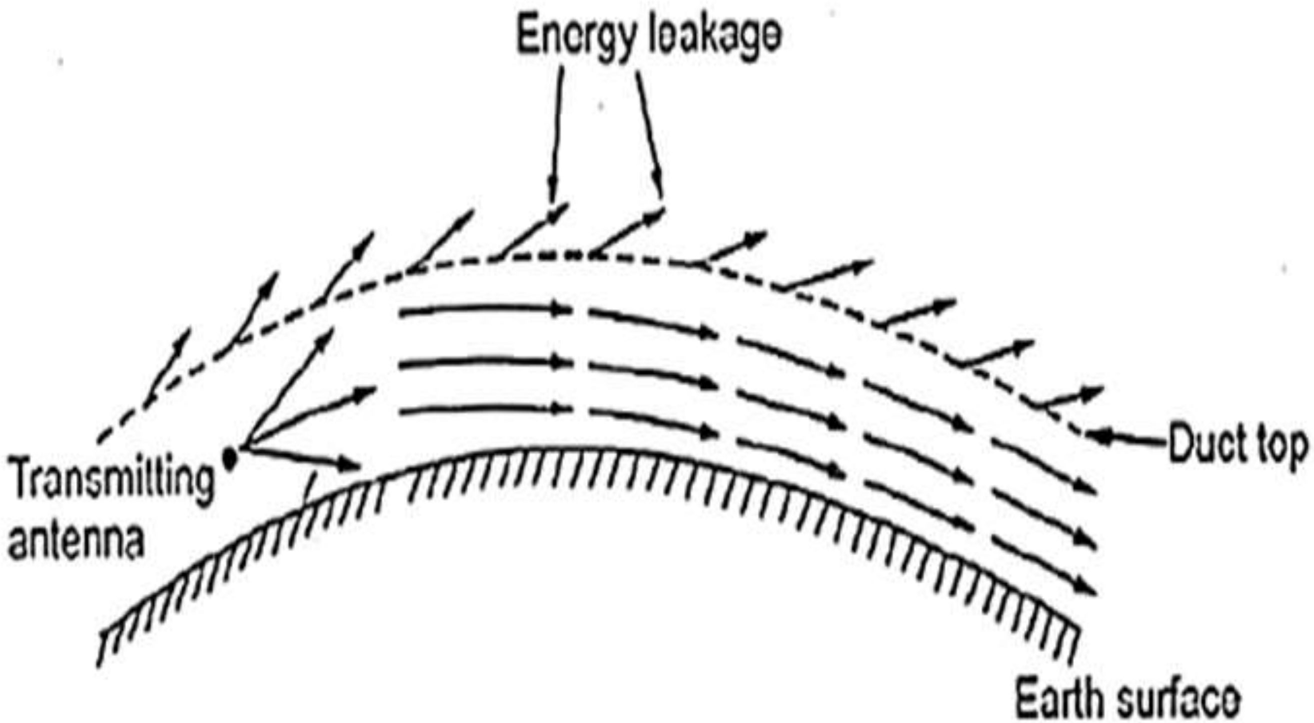
Trapped wave propagating
in series of hops
Transmitting
antenna
Top of duct
Atmospheric
duct
Earth surface

- The concept of wave trapping can be considered as a phenomenon similar to wave guide.
- But the main difference between waveguide and duct propagation is that in wave guide all the modes are confined within guide only.
- But in case of duct propagation, part of energy within duct may escape to the space as shown in the below Fig.
- There is a limit on the wavelength of the signal of maximum value λmax to be trapped in duct.
- It is the maximum wavelength for which the duct propagation holds good. If the wavelength of the signal exceeds the value λmax, then duct effect vanishes almost completely. The value of λmax is given by,

$$\lambda_{max} = 2.5\, h_d \sqrt{\Delta N \times 10^{-6}}$$

where ΔN = change in N value across height of duct

h_d = height of duct



Duct propagation as leaky wave guide

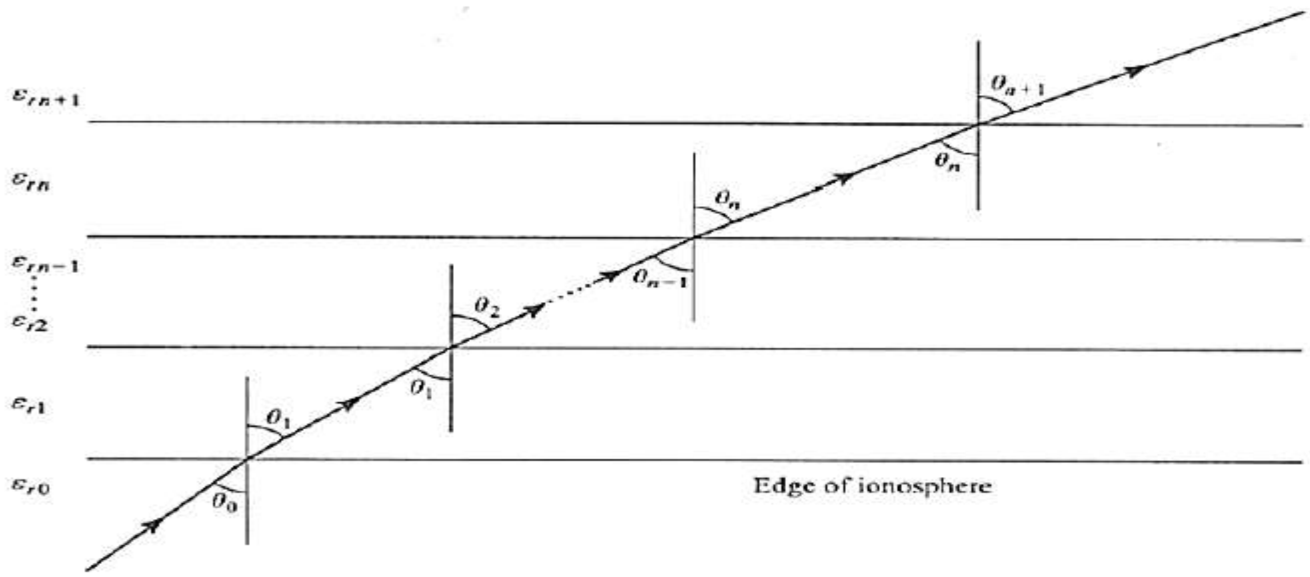
- In general, the duct height hd ranges from 10 to hundreds of meters.
- While the ΔN value is typically 50 units. So considering these values, the phenomenon of duct propagation is found mostly in UHF (ultra high frequency) and microwave frequency regions.
- Moreover the duct propagation is possible only if height of transmitting antenna is less than that of duct height.
- If the transmitting antenna exists considerably above duct, there is comparatively less effect of presence of duct on the signal either inside or above duct.

Characteristic Parameters of Ionospheric Propagation

Topic 9

Critical Frequency

- The critical frequency for the ionized layer of the ionosphere is defined as the highest frequency that can be reflected back to the earth by a particular layer for a vertical incidence. It is denoted by f_{cr}. Note that the critical frequency is different for different layers.
- Let us assume that the ionosphere is lossless, has a relative permeability of unity and can be modeled as plane stratified media in the Figure.



- The wave path can be predicted using Snell's law
- $\sqrt{\epsilon_{r0}} \sin \theta_0 = \sqrt{\epsilon_{r1}} \sin \theta_1 = \cdots = \sqrt{\epsilon_{rn}} \sin \theta_n$ - ---- (5.9)
- where θ_0 is the angle of incidence with respect to the normal and $\theta_1$is the angle of refraction at the lower edge of the ionosphere.
- At the next interface between layers having dielectric constants ϵ_{r1} and ϵ_{r2}, the angle of incidence is $\theta_1$and the angle of refraction is θ_2. At the lower edge of the ionosphere the electron density is zero and hence ϵ_{r0}=1. Therefore, the equation representing Snell's law reduces to ;
- $\sin \theta_0 = \sqrt{\epsilon_{rn}} \sin \theta_n$ ------ (5.10)

- The relative dielectric constant is a function of the electron density N. As the electromagnetic wave propagates deeper into the ionosphere, it passes through a region of higher N into a region of lower
- For a given angle of incidence $\theta_0 = \theta_i$, if N increases to a level such that the angle of refraction, $\theta_n = 90°$, the wave becomes horizontal.
- Under this condition Eqn. (5.10) reduces to
- $\sin \theta_i = \sqrt{\epsilon_{rn}}$ -------- (5.11)

- Let the dielectric constant of the nth layer be $\epsilon_{rn} = \epsilon_r$. Therefore , the refractive index of ionosphere can be defined by

$$n = \sqrt{\epsilon_r} = \left[1 - \left(\frac{81N}{f^2}\right)\right]^{\frac{1}{2}} \quad \text{-------- (5.12)}$$

Substituting the value of ϵ_r from Eqn.(5.12)

$$\sin\theta_i = \sqrt{1 - \frac{81N}{f^2}} \quad \text{-------- (5.13)}$$

Topic 10

Virtual Height

Virtual Height

Definition:

When a **wave** is refracted, it is bent down gradually, but not sharply. However, the path of incident **wave** and reflected **wave** are same if it is reflected from a surface located at a greater **height** of this layer. Such a greater **height** is termed as **virtual height**.

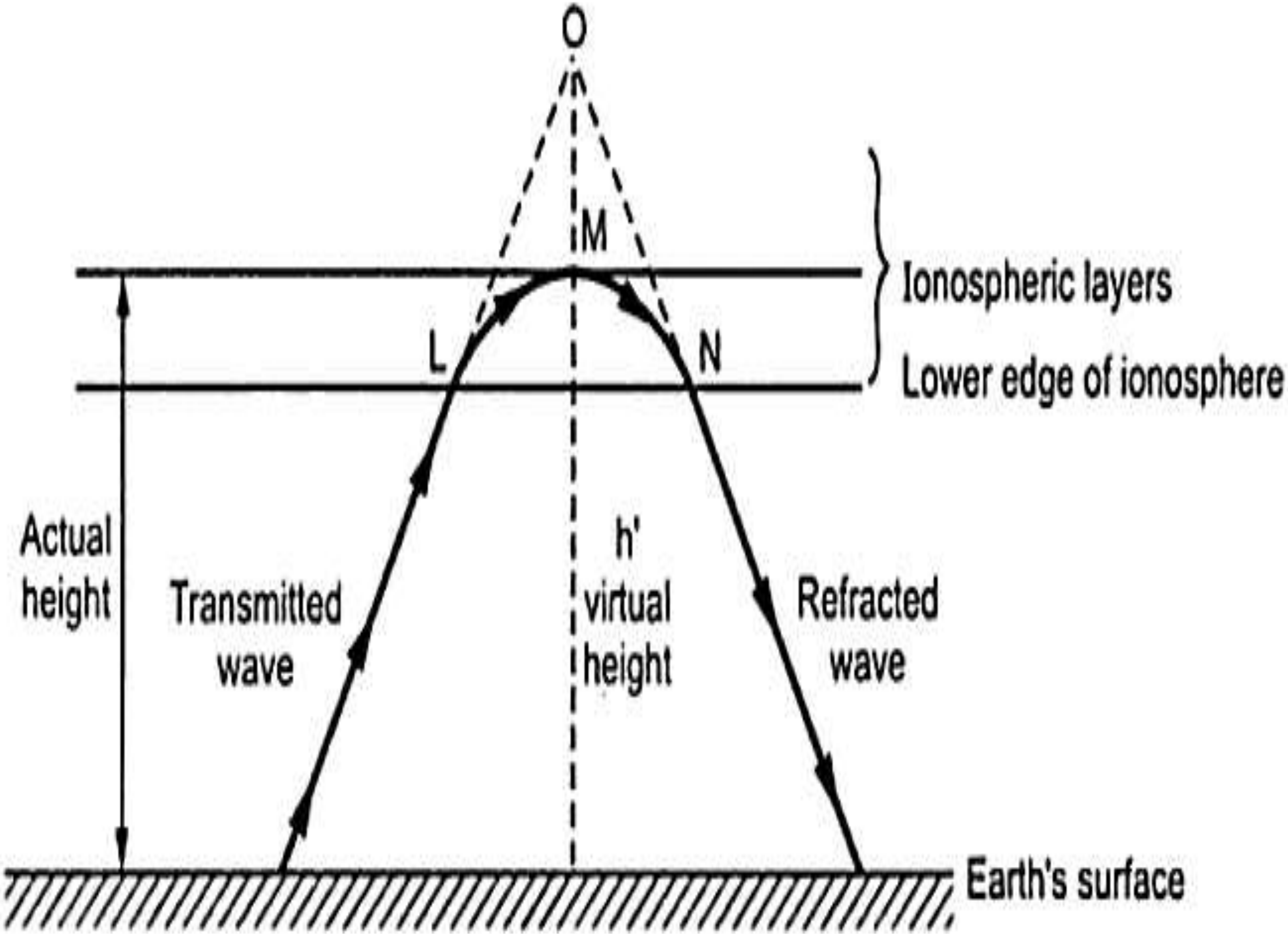
- Consider an electromagnetic wave from a transmitter reaching the receiver after being reflected by the ionosphere as shown in below Fig.
- Let the wave enter the ionosphere at L, and take a curved path LMN before it emerges out of the ionosphere.

- The height at a point above the surface at which the wave bends down to the earth is called *actual height or true height.*
- If the incident and the reflected rays are extended, they meet at point O as shown in Fig. i.e., it is more convenient to think of the wave being reflected rather than refracted. So the path can be assumed to be straight lines.

- The vertical height from the ground to the point O is known as the virtual height of the ionized layer and it is not true height.
- An ionosonde is the instrument used to measure the virtual height of the ionosphere.
- This instrument transmits an RF pulse vertically into the ionosphere from the ground. This pulse is reflected from the ionosphere and is received by the ionosonde.

- The time delay between the transmit and the receive pulse is measured and plotted as a function frequency of the electromagnetic wave.
- The time T duration required for the round trip is noted and then virtual height is determined by using ;
- $h = c\,T\,/2$

where ; h = virtual height , c = velocity of light (m/s) and T = time period (s).


O
M
Ionospheric layers
L
N
Lower edge of ionosphere
Actual height
Transmitted wave
h' virtual height
Refracted wave
Earth's surface

- As the frequency of the electromagnetic wave increases, the virtual height also increases slightly, indicating that the waves of higher frequencies are returned from higher levels within the layer.
- As the frequency approaches the critical frequency (5 MHz for the F1layer), the virtual height steeply increases.
- Once the critical frequency is crossed, the virtual height drops back to a steady value (350 km for 5.5 MHz) which is higher than that for a lower frequency (200 km for *4 MHz).*

- If the incidence angle is greater than θ_i, the wave returns to the earth. For a given angle of incidence, higher frequency electromagnetic waves are reflected from the region having a higher value of N.
- Consider an electromagnetic wave launched vertically into the ionosphere having a maximum electron density N_{max}. Substituting θ_i = 0 in Eqn. (5.13), the highest frequency that gets reflected is given by
- $f_{cr} = \sqrt{81 N_{max}}$ -------- (5.14)
- which is known as the critical frequency.

Topic 11

Maximum Usable Frequency (MUF)

- The critical frequency is the maximum frequency reflected back to the earth by the ionosphere for the vertical incidence. If the frequency of the radio wave exceeds the critical frequency f_{cr} , then the path of propagation in the ionosphere layer depends on the angle of incidence.
- Maximum usable frequency is defined as the limiting maximum frequency that can be reflected back to the earth by the ionospheric layer for a specific angle of incidence other than the angle of incidence for vertical incidence. It is denoted by f_{MUF}.
- The maximum usable frequency f_{MUF} can also be defined as the maximum frequency that can be used for the sky wave propagation for specific distance between two points on the earth.
- Thus f_{MUF} is the highest frequency used for the sky wave communication and for each pair of points on the globe, the value of f_{MUF} will be different. Generally the value of f_{MUF} ranges between 8 MHz to 35 MHz.

For any other angle of incidence, the highest frequency that can be reflected from the ionosphere will be greater than the critical frequency. The highest frequency that gets reflected by the ionosphere for a given value of angle of critical incidence (say θ_m), is known as the maximum usable frequency, f_{MUF}.

Substituting $\theta_i = \theta_m$ and $f = f_{MUF}$ in Eqn. (5.13) ;

$$\sin\theta_m = \sqrt{1 - \frac{81N_{max}}{f_{MUF}^2}} \qquad \text{-------- (5.16)}$$

From Eqn. (5.14), $81N_{max} = f_{cr}^2$. Substituting this in Eqn. (5.16);

$$\sin\theta_m = \sqrt{1 - \frac{f_{cr}^2}{f_{MUF}^2}} \qquad \text{-------- (5.17)}$$

which can be written as ;

$$1 - \sin^2\theta_m = \frac{f_{cr}^2}{f_{MUF}^2} \qquad \text{-------- (5.18)}$$

$$\cos^2\theta_m = \frac{f_{cr}^2}{f_{MUF}^2} \qquad \text{-------- (5.19)}$$

The expression that relates the critical frequency and the angle of incidence to the maximum usable frequency ;

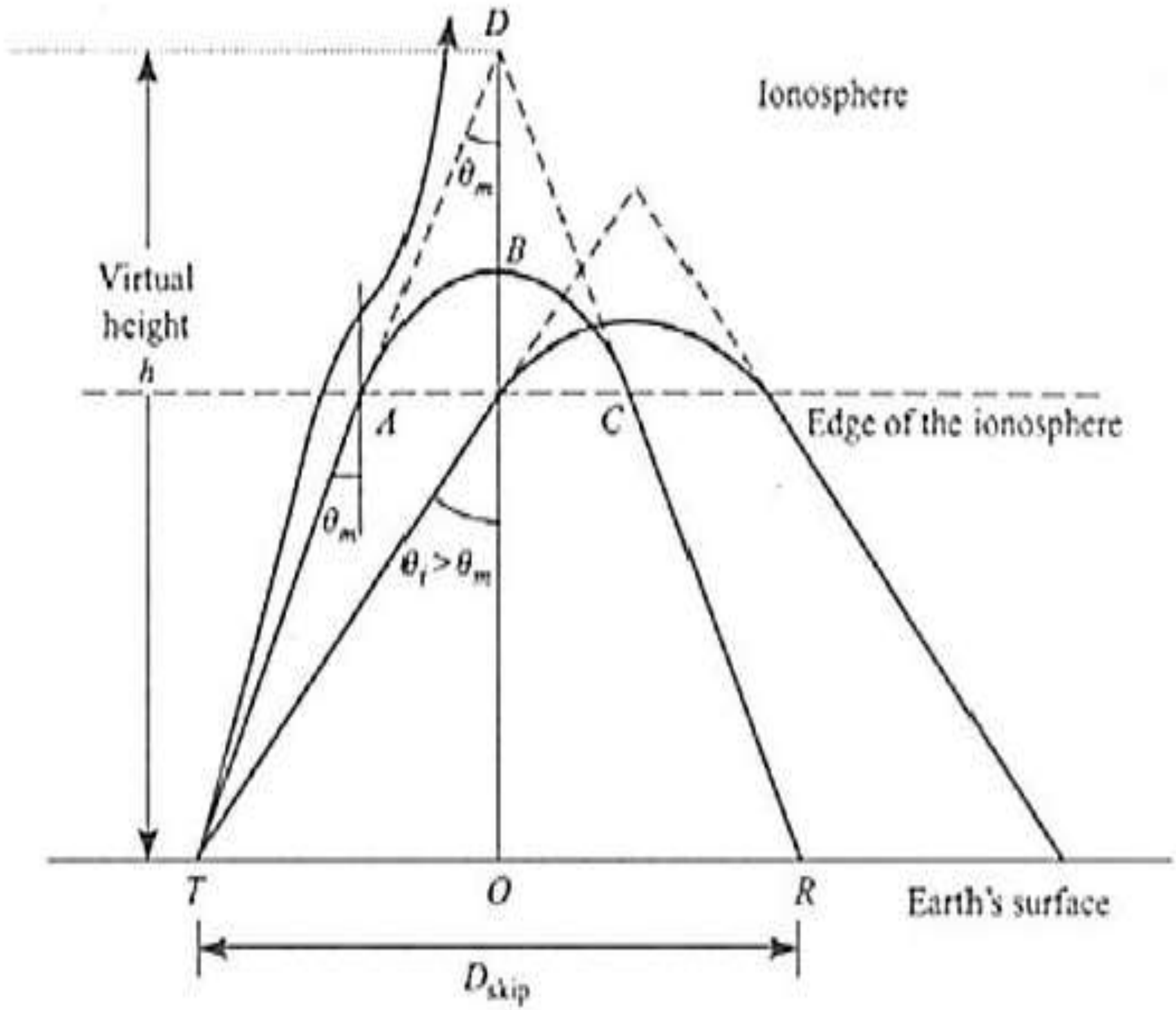
- $f_{MUF} = f_{cr} \sec \theta_m$ -------- (5.20)

- For example, if the critical frequency is 9 MHz ,the maximum usable frequency corresponding to an angle of incidence of 45° is 12.73 MHz.

Topic 12

Skip Distance

- The skip distance is the shortest distance from the transmitter, measured along the surface of the earth, at which a sky wave of fixed frequency will return back to the earth.
- The angle of incidence for which the wave returns back to the earth at minimum distance from the transmitter , i.e., at the skip distance is called angle of critical incidence.
- Assume that the ionosphere can be modeled as a flat reflecting surface at a height h (virtual height) from the surface of the flat earth.



Ray paths for different amgles of incidence, illustrating skip distance

- Consider the frequency of the transmitted wave is kept constant and the angle of critical incidence, θ_m ,.
- For launch angles, $\theta_i < \theta_m$, the waves are received beyond point R. For $\theta_i > \theta_m$, the ionosphere cannot reflect the waves back .
- Let the wave launched at $\theta_i = \theta_m$ reach the surface of the earth at R, at a distance of D_{skip} from the transmitter. The distance D_{skip} is known as the skip distance.
- In the region of radius less than D_{skip}, it is not possible to establish a communication link by the waves reflected from the ionosphere.

- To derive an expression for the skip distance in terms of the critical frequency and the maximum usable frequency by considering the ΔDOT in Fig. 5.15.

From ΔDOT,

$$\cos^2\theta_m = \left(\frac{DO}{DT}\right)^2 = \left[\frac{h}{\sqrt{h^2 + \left(\frac{D_{skip}}{2}\right)^2}}\right]^2 = \frac{1}{1 + \left(\frac{D_{skip}}{2h}\right)^2} \quad \text{-------- (5.21)}$$

From Eqn. (5.19) and Eqn. (5.21) ;

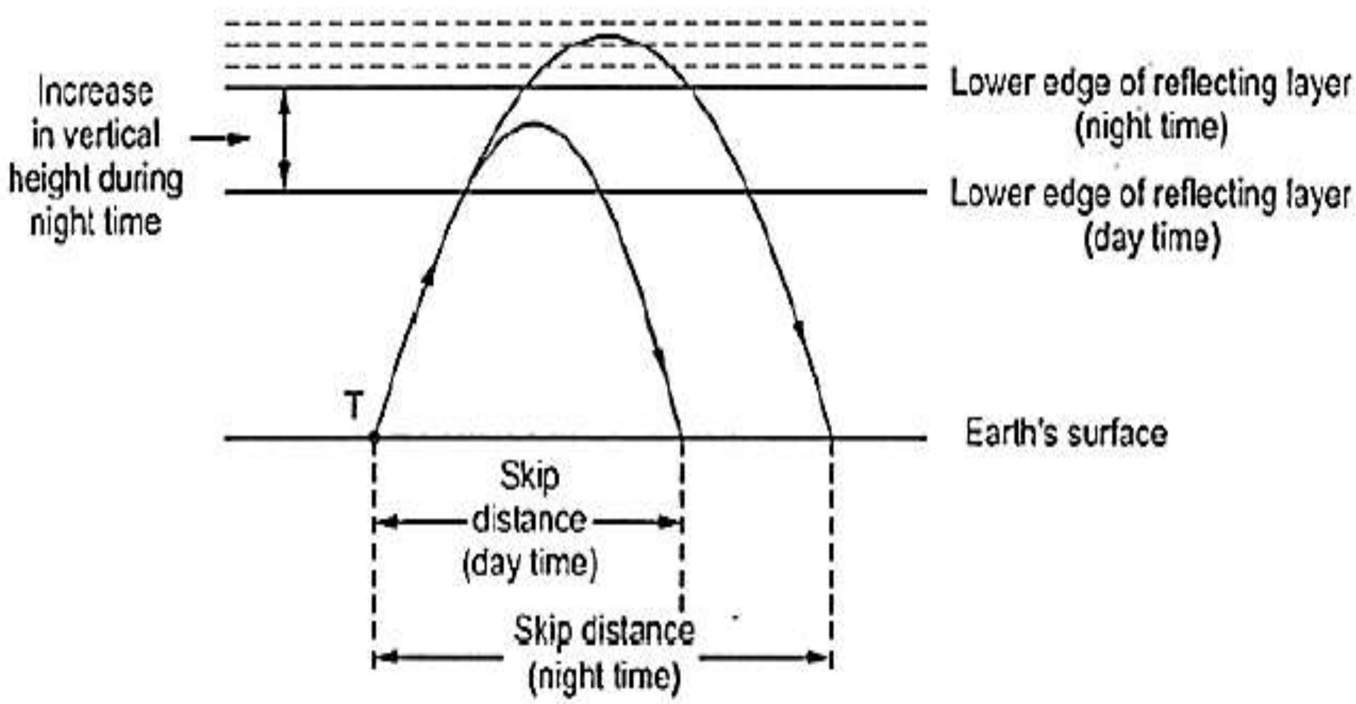
$$D_{skip} = 2h\sqrt{\left(\frac{f_{MUF}}{f_{cr}}\right)^2 - 1} \quad \text{-------- (5.22)}$$

Topic 13

Optimum Working Frequency (OWF)

- For the ionospheric propagation, it is desirable to use as high a frequency as possible. This clearly points out that the frequency used for the ionospheric transmission should be the maximum usable frequency i.e. MUF.
- But MUF depends upon the distance between the transmitter and the receiver and also upon the state of ionosphere. It is observed that due to the daily continuous changes and irregularities in the ionosphere, the MUF varies about 15% of its maximum value. Hence practically the frequency used should be 15% less than the value of MUF.
- Thus the frequency normally used for the ionospheric propagation is known as optimum working frequency. The optimum working frequency between the transmitter and the receiver for the ionospheric transmission is defined as the frequency laying between 50% to 85 % of the predicted MUF between the transmission and the reception points.

- It is observed that the maximum usable frequency at a particular location varies considerably with time of the day, from season to season and from months to months.
- As the optimum working frequency is selected as the fraction of the maximum usable frequency, the OWF also varies in the similar way as the maximum usable frequency varies.



- Practically it is not at all possible to change the frequency of the signal propagated from hour to hour. Hence for the propagation of wave, two frequencies are used namely one for the day time, while other for the night time.
- Sometimes it is preferred to have a third frequency even during the transition period from the day time to night time. It is observed that in the night time vertical height of the ionospheric layer increases as compared to that during the day time. Thus the skip distance also increases. It is illustrated in the Fig. 5-16.
- As we have studied that, the wave with lower frequency is bent more quickly as compared to the wave with higher frequency. Hence the increase in the skip distance during night time is cancelled by using lower frequency during night time.

Topic 14

Fading

FADING

- Fading is basically the undesirable variation in the intensity of the signal received at the receiver.
- Hence the fading is defined as the fluctuations in the received signal strength caused due to variations in height and density of the ionization in different layers.
- Basically the fading is the common characteristic of the high frequency short wave propagation i.e. sky wave propagation.
- At receiver, the strength of the signal received is the vector sum of the waves received.
- Because the waves leave from transmitter at same time but reach at the receiver through different paths.
- So the fading is caused due to interference between two waves of different path lengths.

Various types of fading are as follows.

1. Selective Fading

- It is more dominant at high frequencies for which sky propagation is used.
- The selective fading produces serious distortion of modulated signal.
- Due to the selective fading, the amplitude modulated signals are seriously affected.
- The AM signal are more distorted due to the selective fading rather than SSB signals.
- Hence to reduce the selective fading Exalted carrier reception and single side band system can be used

2. Interference Fading

- As name indicates, it is the fading produced because of upper and lower rays of the sky wave interfering with each other. This is the most serious fading.
- It is also produced due to the interference between a ground wave and a sky wave or between sky waves reaching receiver by different paths or different number of hops.
- For a single sky wave frequency, interference fading takes place due to the fluctuations in the height of the ionospheric layer or due to the variation in the ionic density of the layer.

- As ionosphere is subjected to the continuous small variations. Because of this, the length of the path that the reflected wave follows also undergoes small variations.
- Thus the relative phase of the wave reaching receiver varies randomly.
- Because of these conditions, the amplitude of the resultant also varies continuously which is nothing but the interference fading.
- This can be minimized by using space diversity or frequency diversity reception.

3. Absorption Fading

- This type of fading occurs due to the variations of single strength with the different amount of absorption of waves absorbed by the transmitting medium.

4. Polarization Fading

- When the sky wave reaches after the reflection, the state of polarization is constantly changing.
- The polarization of the sky wave coming down changes because of the superposition of the ordinary and extra ordinary waves (which are having random amplitudes and phases) which are oppositely polarized.
- Thus the polarization of the wave changes continuously with respect to antenna, which gives rise to the variations in the amplitude at the receiver. Such type of fading is called polarization fading.

5. Skip Fading

- At distances near the skip range or skip zone, the fading occurs which is called skip fading.
- Due to the variations in the height and the density of the ionized layer, the point at which the wave can be received moves in or out of the skip zone.
- Thus due to this the amplitude at the receiver also varies producing skip fading near the skip range.
- The minimize the fading, the most common method is to use automatic volume control (AVC or AGC), in the receiver.

Topic 15

Multi-hop Propagation

- Let us now consider the transmit and receive antennas located on the surface of the spherical earth. The ionosphere is modeled as a spherical reflecting surface at a virtual height h from the surface of the earth [Fig. 5.17 (a)].
- A wave launched at a grazing angle ψ from point *A* gets reflected by the ionosphere (provided $\theta_i > \theta_m$) and reaches the surface of the earth at *C*.
- If the earth is a good reflector, the wave can undergo multi-hops and thus can establish communication between the points *A* and *E* in addition to that between *A* and *C*.

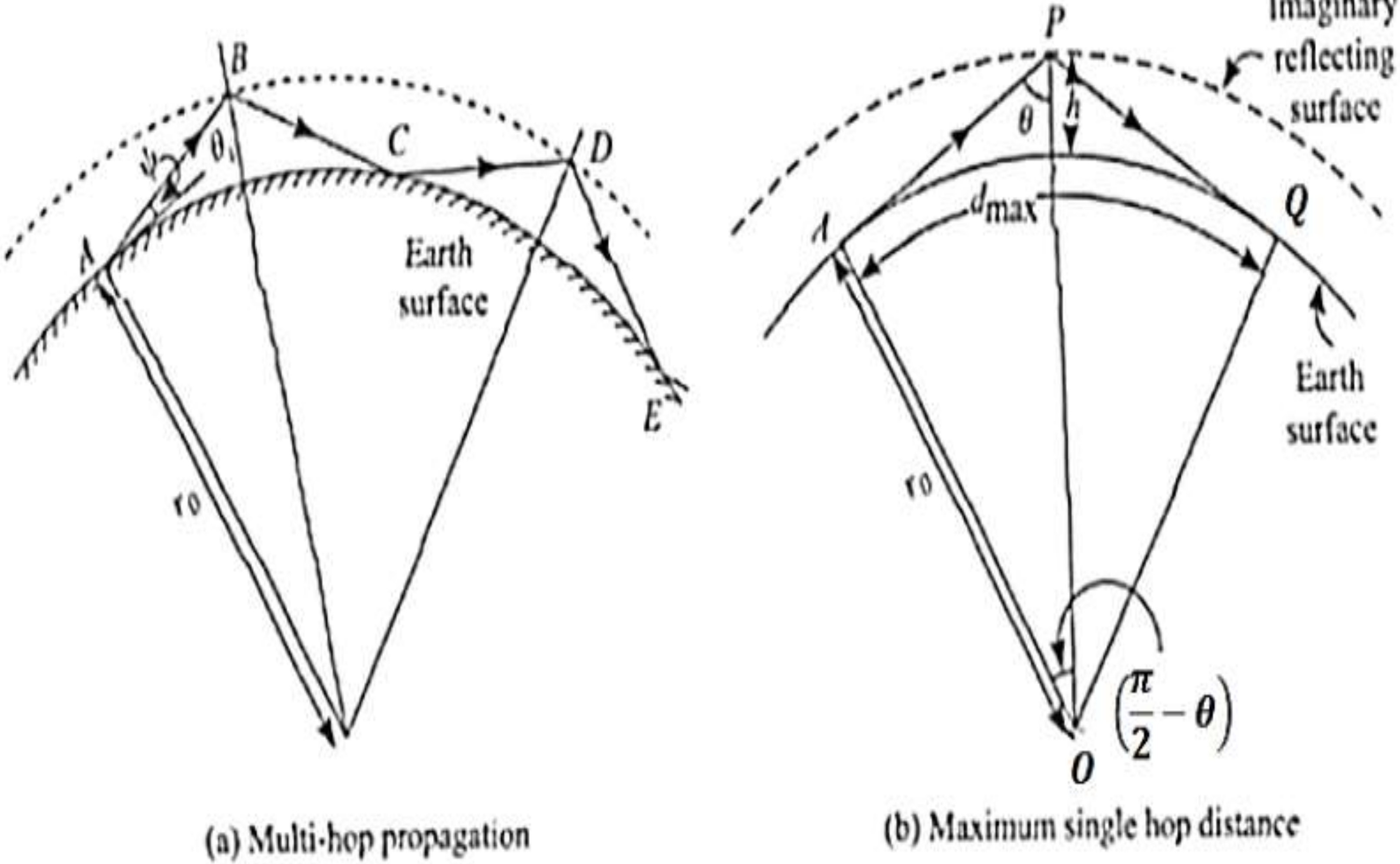


Fig. 5-17 Propagation of sky waves above the spherical earth

The single-hop distance *AC* is a function of the grazing angle, ψ. The maximum value of the single hop distance occurs for $\psi = 0$ or horizontal launch as shown in Fig. 5.17 (b). The incidence angle at P is given by ;

$$\theta = \sin^{-1}\left(\frac{r_0}{r_0 + h}\right) \qquad \text{-------- (5.23)}$$

Therefore the maximum single hop distance is

$$d_{max} = 2r_0\left(\frac{\pi}{2} - \theta\right) \qquad \text{-------- (5.24)}$$

For example, the reflection from the E layer with $h = 100\ km$, the angle of incidence is

$$\theta = \sin^{-1}\left(\frac{6370}{6470}\right) = 79.91^\circ = 1.395\ rad$$

The maximum single-hop distance is

$$d_{max} = 2(6370)\left(\frac{\pi}{2} - 1.395\right) = 2240\ km$$

Similarly, for the F layer, with a virtual height of 300 km, the maximum angle of incidence is 72.75° and the maximum single-hop distance is 3836 km.

Applications

- In fact **waves** have **applications** in almost every field of everyday life – from wireless communications to detecting over-speeding vehicles, from the music of guitar to laser – almost every aspect of our everyday life in some way involves **wave**.