SKP Engineering College

Tiruvannamalai – 606611

A Course Material

on

Antenna and Wave Propagation

By

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Electronics And Communication Engineering Department

Quality Certificate

This is to Certify that the Electronic Study Material

Subject Code:EC6602

Subject Name:Antenna and Wave Propagation

Year/Sem:III/VI

Being prepared by me and it meets the knowledge requirement of the University curriculum.

Signature of the Author

Name: A.Parimala

Designation: Assistant Professor

This is to certify that the course material being prepared by Ms.A.Parimala is of the adequate quality. She has referred more than five books and one among them is from abroad author.

Signature of HD	Signature of the Principal
Name:	Name: Dr.V.Subramania Bharathi
Seal:	Seal:

EC6602 ANTENNA AND WAVE PROPAGATION L T P C 3 0 0 3

OBJECTIVES: The student should be made to:

- To give a thorough understanding of the radiation characteristics of different types of antennas
- To give insight of the radiation phenomena.
- To create awareness about the different types of propagation of radio waves at different frequencies.

UNIT I FUNDAMENTALS OF RADIATION 9

Definition of antenna parameters – Gain, Directivity, Effective aperture, Radiation Resistance, Band width, Beam width, Input Impedance. Matching – Baluns, Polarization mismatch, Antenna noise temperature, Radiation from oscillating dipole, Half wave dipole. Folded dipole, Yagi array.

UNIT II APERTURE AND SLOT ANTENNAS 9

Radiation from rectangular apertures, Uniform and Tapered aperture, Horn antenna , Reflector antenna , Aperture blockage , Feeding structures , Slot antennas ,Microstrip antennas – Radiation mechanism – Application ,Numerical tool for antenna analysis.

UNIT III ANTENNA ARRAYS 9

N element linear array, Pattern multiplication, Broadside and End fire array – Concept of Phased arrays, Adaptive array, Basic principle of antenna Synthesis-Binomial array.

UNIT IV SPECIAL ANTENNAS 9

Principle of frequency independent antennas –Spiral antenna, Helical antenna, Log periodic. Modern antennas- Reconfigurable antenna, Active antenna, Dielectric antennas, Electronic band gap structure and applications, Antenna Measurements-Test Ranges, Measurement of Gain, Radiation pattern, Polarization, VSWR

UNIT V PROPAGATION OF RADIO WAVES 9

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

TOTAL : 45 PERIODS

TEXT BOOKS:

1. John D Kraus," Antennas for all Applications", 3rd Edition, Mc Graw Hill, 2005.

REFERENCES:

1. Edward C.Jordan and Keith G.Balmain" Electromagnetic Waves and Radiating Systems" Prentice Hall of India, 2006.
2. R.E.Collin,"Antennas and Radiowave Propagation", Mc Graw Hill 1985.
3. Constantine.A.Balanis "Antenna Theory Analysis and Design", Wiley Student Edition, 2006.
4. Rajeswari Chatterjee, "Antenna Theory and Practice" Revised Second Edition New Age International Publishers, 2006.
5. S. Drabowitch, "Modern Antennas" Second Edition Springer Publications, 2007.
6. Robert S.Elliott "Antenna Theory and Design" Wiley Student Edition, 2006.
7. H.Sizun "Radio Wave Propagation for Telecommunication Applications", First Indian Reprint, Springer Publications, 2007.

CONTENTS

Unit – I

Fundamentals of Radiation

Part – A

1 What is elementary dipole? [CO1-L1]

A short dipole that has uniform current is known as elementary dipole.

2. What are θ and φ patterns in antenna radiation pattern? [CO1-L1]

The patterns obtained as variation of field or power as a function of the spherical coordinates θ and φ are called as θ and φ patterns of radiation pattern.

3. What is radiation resistance? [CO1-L1- April/May 2011]

The antenna is a radiating device in which power is radiated into space in the form of electromagnetic wave. W' = I2 R Rr = W'/ I2 Where Rr is a fictitious resistance called called as radiation resistance.

4. What is meant by isotropic radiator? [CO1-L1]

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

5. Define gain. [CO1-L1- Nov/Dec 2011]

The ratio of maximum radiation intensity in given direction to the maximum radiation intensity from a reference antenna produced in the same direction with same input power.

6. Define Aperture efficiency. [CO1-L1]

The ratio of the effective aperture to the physical aperture is the aperture efficiency. i.e Aperture efficiency = hap = Ae / Ap(Dimensionless).

7. Define directivity. [CO1-L1- May/June 2012]

The directivity of an antenna is equal to the ratio of the maximum power density P(q,f)max to its average value over a sphere as observed in the far field of an antenna. D = P(q,f)max / P(q,f)av. Directivity from Pattern. D = 4p / WA. . Directivity from beam area(WA).

8. What is meant by reciprocity theorem? [CO1-L1- May/June 2012]

If an e.m.f is applied to the terminals of an antenna no.1 and the current measured at the terminals of the another antenna no.2, then an equal current both in amplitude and phase will be obtained at the terminal of the antenna no.1 if the same emf is applied to the terminals of antenna no.2.

9. Define axial ratio. [CO1-L1- April/May2014]

The ratio of the major to the minor axes of the polarization ellipse is called the Axial Ratio. (AR).

10. Define self impedance. [CO1-L1]

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

11. What is meant by mutual impedance. [CO1-L1]

The presence of near by antenna no.2 induces a current in the antenna no.1 indicates that presence of antenna no.2 changes the impedance of the antenna no.1.This effect is called mutual coupling and results in mutual impedance.

12. What is meant by antenna beamwidth? [CO1-L1]

Antenna beamwidth is a measure of directivity of an antenna. Antenna beam width is an angular width in degrees, measured on the radiation pattern (major lobe) between points where the radiated power has fallen to half its maximum value .This is called as “beam width” between half power points or half power beam width.(HPBW).

13. What is meant by polarization? [CO1-L1- April/May2013]

The polarization of the radio wave can be defined by direction in which the electric vector E is aligned during the passage of atleast one full cycle.Also polarization can also be defined the physical orientation of the radiated electromagnetic waves in space.The polarization are three types. They are Elliptical polarization ,circular polarization and linear polarization.

14. Define radiation intensity. [CO1-L1- April/May2013]

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

15. Define beam efficiency. [CO1-L1]

The total beam area (ΩA) consists of the main beam area (ΩM) plus the minor lobe area (Ωm) . Thus $\Omega A = \Omega M + \Omega m$. The ratio of the main beam area to the total beam area is called beam efficiency. Beam efficiency = $\Sigma M = \Omega M / \Omega A$.

16. What are different types of apertures? [CO1-L1]

i) Effective aperture.
ii). Scattering aperture.
iii) Loss aperture.
iv) collecting aperture.

17. What is meant by front to back ratio? [CO1-L1- April/May2015]

It is defined as the ratio of the power radiated in desired direction to the power radiated in the opposite direction. i.e FBR = Power radiated in desired direction / power radiated in the opposite direction.

18. What are field zone? [CO1-L1]

The fields around an antenna ay be divided into two principal regions.

i. Near field zone (Fresnel zone)
ii. Far field zone (Fraunhofer zone).

19. What are dB_i and dB_d? [CO1-L1]

dB_i indicates decibels over isotropic antenna.
dB_d indicates decibels over directional antenna.

20. Define half wave dipole. [CO1-L1- Nov/Dec2014]

Half wave dipole antenna is one of the simplest antenns and is frequently employes as an element of a more complex directional system.

21. Write down the important features of yagi uda antenna. [CO1-L2- April/May2015]

- If three element array is used then such type of antenna is called as beam antenna.
- It has unidirectional beam of moderate directivity, light weight and low cost.
- It provides gain of the order of 8dB.

Part-B

1. Define the following parameters and explain the terms (1) Radiation pattern (2) Input Impedance (3) Polarization (4) Antenna temperature (5) Reciprocity theorem.(16M)
[CO1-L2-April/May 2011, Nov/Dec2011,May/June 2012]

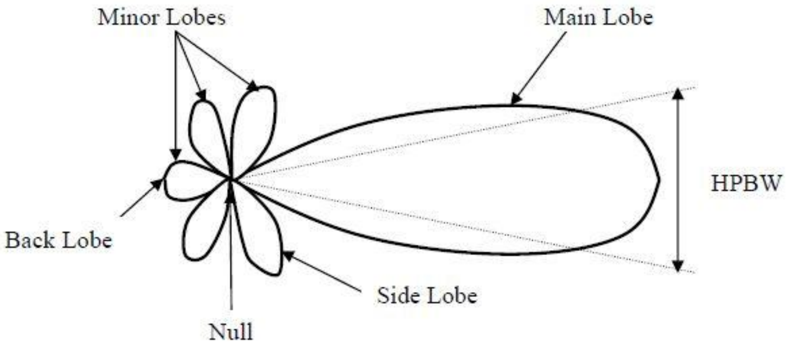
Radiation pattern is one of the important characteristic of an antenna as tells the spatial relative distribution of the electromagnetic wave generated by the antenna. The radiation pattern is a plot of the magnitude of the radiation field as a function of direction (θ,ɳ). The radiation pattern is essentially a 3-D surface.

In the field of antenna design the term radiation pattern (or antenna pattern or far-field pattern) refers to the directional (angular) dependence of the strength of the radio waves from the antenna or other source.

Particularly in the fields of fiber optics, lasers, and integrated optics, the term radiation pattern, or near-field radiation pattern, may also be used as a synonym for the near-field pattern or Fresnel pattern.
This refers to the positional dependence of the electromagnetic field in the near-field, or Fresnel region of the source. The near-field pattern is most commonly defined over a plane placed in front of the source, or over a cylindrical or spherical surface enclosing it.

The far-field pattern of an antenna may be determined experimentally at an antenna range, or alternatively, the near-field pattern may be found using a near-field scanner, and the radiation pattern deduced from it by computation. The far-field radiation pattern can also be calculated from the antenna shape by computer programs such as NEC. Other software, like HFSS can also compute the near field.

The far field radiation pattern may be represented graphically as a plot of one of a number of related variables, including; the field strength at a constant (large) radius (an amplitude pattern or field pattern), the power per unit solid angle (power pattern) and the directive gain. Very often, only the relative amplitude is plotted, normalized either to the amplitude on the antenna boresight, or to the total radiated power. The plotted quantity may be shown on a linear scale, or in dB. The plot is typically represented as a three dimensional graph (as at right), or as separate graphs in the vertical plane and horizontal plane. This is often known as a polar diagram



There are various parts of radiation pattern:

(a) HPBW: The half power beamwidth (HPBW) can be defined as the angle subtended by the half power points of the main lobe.

(b) Main Lobe: This is the radiation lobe containing the direction of maximum radiation.

(c) Minor Lobe: All the lobes other then the main lobe are called the minor lobes. These lobes represent the radiation in undesired directions. The level of minor lobes is usually expressed as a ratio of the power density in the lobe in question to that of the major lobe. This ratio is called as the side lobe level (expressed in decibels).

(d) Back Lobe: This is the minor lobe diametrically opposite the main lobe.

(e) Side Lobes: These are the minor lobes adjacent to the main lobe and are separated by various nulls. Side lobes are generally the largest among the minor lobes.

(2) Input Impedance

The impedance of antenna measured at the terminals where transmission line carrying R.F. power connected is called antenna input impedance. These terminals are nothing but feed points of the antenna, the impedance is also called feed point impedance or terminal impedance. As the R.F. power carried by the transmission line from the transmitter, excites or drives the antenna, the antenna input impedance can be alternatively called driving point impedance of antenna.

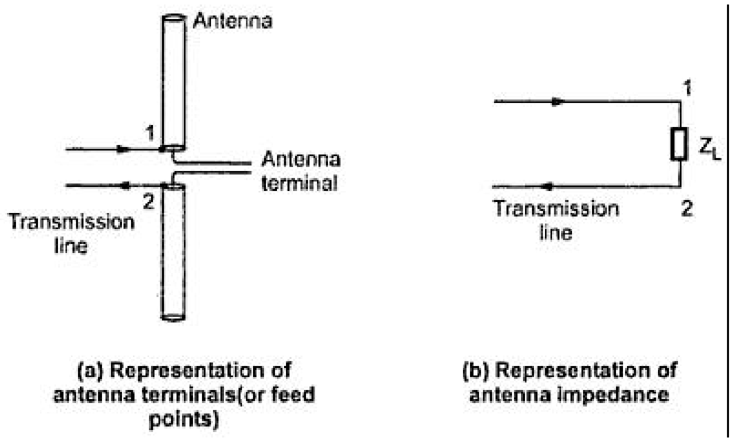


Fig 6 Representation of antenna impedance

When the antenna is lossless and isolated from ground and other objects, the impedance offered by antenna to the transmission line is represented by two terminal networks with impedance Z_L as shown in the Fig. 6(b). Note that the notation Z_L represents that the antenna impedance acts as load to the transmission line driving antenna. With a lossless and isolated antenna, the antenna terminal impedance is same as the self impedance of the antenna, which is represented by Z_{11}. The self impedance of the antenna is a complex quantity given by,

$$Z_{11} = R_{11} + j X_{11} \quad \text{...(1)}$$

The real part of Z_{11} i.e. R_{11} is called self resistance or radiation resistance of antenna, while the imaginary part of Z_{11} i.e. X_{11} is called self reactance of antenna.
For half wave dipole, the self impedance is typically given by,

$$Z_{11} = R_{11} + j X_{11} = 73 + j 42.45 \quad \text{... (2)}$$

Note: The self impedance of antenna is always positive. The value of self impedance is same for antenna used either as transmitting antenna or receiving antenna. The self impedance of the The impedance of antenna measured at the terminals where transmission line carrying R.F. power connected is called antenna input impedance. These terminals are nothing but feed points of the antenna, the impedance is also called feed point impedance or terminal impedance. As the R.F. power carried by the transmission line from the transmitter, excites or drives the antenna, the antenna input impedance can be alternatively called driving point impedance of antenna.

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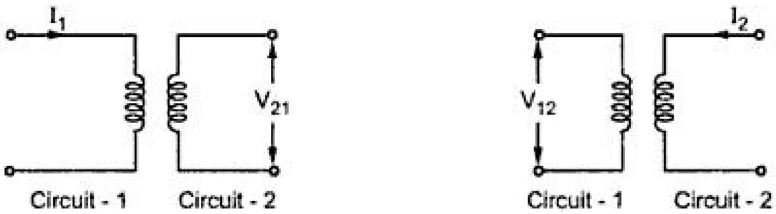


Thus the mutual impedance of the coupled circuit is defined as negative ratio of the voltage induced at the open terminals of once circuit to the current in other circuit. Mathematically we can write,

$$Z_{21} = -\frac{V_{21}}{I_1} \quad \text{... (3)}$$

$$Z_{12} = -\frac{V_{12}}{I_2} \quad \text{... (4)}$$

Exactly on the similar lines to the coupled circuits, the mutual impedance of the antenna is given by,

$$Z_{21} = -\frac{V_{21}}{I_1} \quad \text{... (5)}$$

$$Z_{12} = -\frac{V_{12}}{I_2} \quad \text{... (6)}$$

But according to reciprocity theorem, we can write mutual impedance of antenna as,

$$\boxed{Z_m = \frac{V_{21}}{I_1} = \frac{V_{12}}{I_2}} \quad \text{... (7)}$$

The mutual impedance depends on,

i) Magnitude of induced voltage,

ii) Phase difference between induced voltage and input current,

iii) Tuning conditions of coupled antennas.

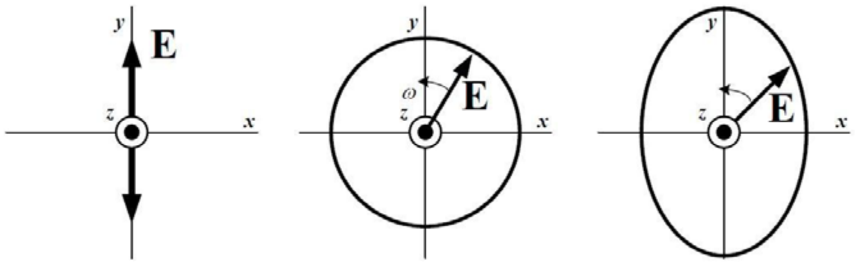
(3) Polarization

The polarization of the EM field describes the orientation of its vectors at a given point and how it varies with time. In other words, it describes the way the direction and magnitude of the field
vectors (usually **E**) change in time. Polarization is associated with TEM time-harmonic waves where the **H** vector relates to the **E** vector simply by

$$\mathbf{H} = \hat{\mathbf{r}} \times \mathbf{E} / \eta$$

In antenna theory, we are concerned with the polarization of the field in the plane orthogonal to the direction of propagation—this is the plane defined by the vectors of the far field. Remember that the far field is a quasi-TEM field.
Hence the polarization is the locus traced by the extremity of the time-varying field vector at a fixed observation point.

According to the shape of the trace, three types of polarization exist for harmonic fields: linear, circular and elliptical. Any polarization can be represented by two orthogonal linear polarizations, (E_x , E_y) or (E_H, E_V), whose fields are out of phase by an angle of δ_L.



(a) linear polarization (b) circular polarization (c) elliptical polarization

(4) Antenna temperature

Antenna Temperature (T_A) is a parameter that describes how much noise an antenna produces in a given environment. This temperature is not the physical temperature of the antenna. Moreover, an antenna does not have an intrinsic "antenna temperature" associated with it; rather the temperature depends on its gain pattern and the thermal environment that it is placed in. Antenna temperature is also sometimes referred to as Antenna Noise Temperature $T(\theta,\phi)$. Hence, an antenna's temperature will vary depending on whether it is directional and pointed into space or staring into the sun.

For an antenna with a radiation pattern given by $R(\theta,\phi)$, the noise temperature is
mathematically defined as:

$$T_A = \frac{1}{4\pi}\int_0^{2\pi}\int_0^{\pi} R(\theta,\phi)T(\theta,\phi)\sin\theta d\theta d\phi$$

This states that the temperature surrounding the antenna is integrated over the entire sphere, and weighted by the antenna's radiation pattern. Hence, an isotropic antenna would have a noise temperature that is the average of all temperatures around the antenna; for a perfectly directional antenna (with a pencil beam), the antenna temperature will only depend on the temperature in which the antenna is "looking".

The noise power received from an antenna at temperature T_A can be expressed in terms of the bandwidth (*B*) the antenna (and its receiver) are operating over:

$$P_{TA} = KT_A B$$

In the above, *K* is Boltzmann's constant (1.38 x 10^{-23} [Joules/Kelvin = J/K]). The receiver also has a temperature associated with it (T_R), and the total system temperature (antenna plus receiver) has a combined temperature given by $T_{sys} = T_A + T_R$. This temperature can be used in the above equation to find the total noise power of the system. These concepts begin to illustrate how antenna engineers must understand receivers and the associated electronics, because the resulting systems very much depend on each other.

A parameter often encountered in specification sheets for antennas that operate in certain environments is the ratio of gain of the antenna divided by the antenna temperature (or system temperature if a receiver is specified). This parameter is written as G/T, and has units of dB/Kelvin [dB/K].

(5) Reciprocity theorem

Statement: "In any linear and bilateral network consisting the linear and bilateral impedance the ratio of voltage V applied between any two terminals to the current I measured in any branch is same as the ratio V to I obtained by interchanging the positions of voltage source and the ammeter used for current measurement."

The ratio V to I is generally called transfer impedance. Here both the voltage source and ammeter are assumed to have zero impedance. This theorem holds good if both, voltage source and ammeter have same internal impedances.

This theorem is equally useful in the circuit theory as well as the field theory. Let us consider that the antenna system is represented as a 4-terminal network with pair of terminals at input and another pair of terminals at the output. It is also called two port network as pair of terminals is defined as port. The 4-terminal representation of the antenna system is as shown in the Fig. 10

(a). Note that the pair of terminals or ports are nothing but the terminals of the dipoles as shown in the Fig. 10 (b).

$$V_1 = z_{11} I_1 + z_{12} I_2$$
$$V_2 = z_{21} I_1 + z_{22} I_2$$

Or

$$I_1 = y_{11} V_1 + y_{12} V_2$$
$$I_2 = y_{21} V_1 + y_{22} V_2$$

Thus according to the reciprocity theorem for the linear and bilateral networks, the conditions of the reciprocity of the network are,

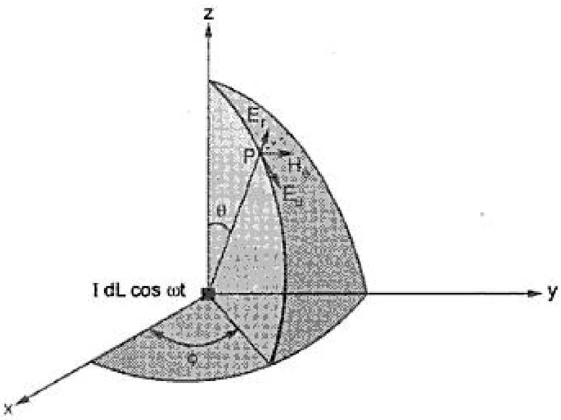
The impedances z_{12} and z_{21} are called mutual impedances which are individually ratio of open circuit voltage at one port to the current at other port. Similarly admittances y_{12} and y_{21} are called transfer admittances which are individually the ratio of a short circuited current at one port to the voltage at other port. Finally the impedances $z\eta_{12}$ and $z\eta_{21}$ are called transfer impedances which are individually the ratio of an open circuit voltage at one port to a short circuit current at other port.

2. Derive the magnetic field components and vector potential of the current element hertizian dipole. (16M)

[CO1-H1-April/May 2011, Nov/Dec2011, May/June 2012, April/May 2014]

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 1. This length of a thin wire is assumed to be very short, so that the filamentary current can be considered as constant along the length of an element. The important usage of this approximation is observed in case of current carrying antenna. In such cases, an antenna can be considered as made up of large numbers of such elements connected end to end. Hence if the electromagnetic field of such small element is known, then the electromagnetic field of any long antenna can be easily calculated.
Let us study how to calculate the electromagnetic field due to an alternating current element. Consider spherical co-ordinate system. Consider that an alternating current element IdLcos ωt is located at the centre as shown in the Fig



The aim is to calculate electromagnetic field at point P placed at a distance R from the origin. The current element IdLcos ωt is placed along the z-axis.

Let us write the expression for vector potential Ā at point P, using previous knowledge. The vector potential Ā is given by,

$$\bar{A}(r) = \frac{\mu}{4\pi}\int_v \frac{\bar{J}\left(t-\frac{r}{v}\right)}{R} dv' \quad \text{... (1)}$$

Here the vector potential is retarded in time by *r/v* sec, where *v* is the velocity of propagation. As the current clement is placed along the z-axis, the vector potential will also have only one component in positive z-direction. Hence we can write,

$$A_z = \frac{\mu}{4\pi}\int_v \frac{\bar{J}\left(t-\frac{r}{v}\right)}{R} dv' \quad \text{... (2)}$$

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

$$\int_v \bar{J}\left(t-\frac{r}{v}\right) dv' = I\, dL \cos \omega\left(t-\frac{r}{v}\right) \quad \ldots (3)$$

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

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

Now the magnetic field is given by

$$\mu \bar{H} = \nabla \times \bar{A} \quad \ldots (5)$$

As we are using spherical co-ordinate system, to find the curl of Ā, we must find the component of Ā in r, θ and η directions. From the Fig. 18, it is clear that,

$$A_r = A_z \cos\theta$$
$$A_\theta = -A_z \sin\theta \quad \ldots (6)$$
$$A_\phi = 0$$

Hence is given by,

$$\nabla \times \bar{A} = \frac{1}{r\sin\theta}\left[\frac{\partial}{\partial\theta}(A_\phi \sin\theta) - \frac{\partial A_\theta}{\partial\phi}\right]\bar{a}_r + \frac{1}{r}\left[\frac{1}{\sin\theta}\frac{\partial}{\partial\phi}A_r - \frac{\partial}{\partial r}(r A_\phi)\right]\bar{a}_\theta + \frac{1}{r}\left[\frac{\partial}{\partial r}(r A_\theta) - \frac{\partial}{\partial\theta}(A_r)\right]\bar{a}_\phi \quad \ldots (7)$$

Now note that A_η=0 and because of symmetry ∂/∂_η=0 as no variation along η direction. Thus first two terms in equation (7) can be neglected being zero.
Putting values of A_θ and A_η, from equation (5), we get,

$$\nabla \times \bar{A} = \frac{1}{r}\left[\frac{\partial}{\partial r}\{r(-A_z \sin\theta)\} - \frac{\partial}{\partial\theta}\{A_z \cos\theta\}\right]\bar{a}_\phi \quad \ldots (8)$$

Substituting value of A_Z,

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

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

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

$$\nabla\times\bar{A} = \frac{\mu I\,dL\sin\theta}{4\pi}\left[\frac{-\omega\sin\omega\left(t-\frac{r}{v}\right)}{rv}+\frac{\cos\omega\left(t-\frac{r}{v}\right)}{r^2}\right]\bar{a}_\phi \quad \text{... (9)}$$

Hence the magnetic field $\bar{H}$ is given by,

$$\bar{H} = \frac{1}{\mu}[\nabla\times\bar{A}]$$

Putting value of (∆ x Ā) from equation (9) we get,

$$\bar{H} = \frac{I\,dL\sin\theta}{4\pi}\left[\frac{-\omega\sin\omega\left(t-\frac{r}{v}\right)}{rv}+\frac{\cos\omega\left(t-\frac{r}{v}\right)}{r^2}\right]\bar{a}_\phi \quad \text{... (10)}$$

Equation (10) indicates that the magnetic field $\bar{H}$ exists only in ɳ direction.

$$\therefore \quad H_\phi = \frac{I\,dL\sin\theta}{4\pi}\left[\frac{-\omega\sin\omega\left(t-\frac{r}{v}\right)}{rv}+\frac{\cos\omega\left(t-\frac{r}{v}\right)}{r^2}\right] \quad \text{... (11)}$$

Let (t- *r/v*) = t ɳ, substituting the value in equation (11), we get,

$$H_\phi = \frac{I\,dL\sin\theta}{4\pi}\left[\frac{-\omega\sin\omega t'}{rv}+\frac{\cos\omega t'}{r^2}\right] \quad \text{... (12)}$$

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

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

$$\therefore \quad \partial \overline{E} = \frac{1}{\varepsilon}(\nabla \times \overline{H})\ dt$$

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 $\overline{a}_r$ direction is given by

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

But $\partial/\partial_\eta = 0$

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

Substituting value of H_η from equation (12),

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

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

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

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

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

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

But again $\partial/\partial_\eta = 0$

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

Substituting value of H_η from equation (11),

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

$$\therefore \quad (\nabla \times \overline{H})_\theta = -\frac{I\,dL\sin\theta}{4\pi r}\left\{\left[\frac{-\omega\cos\omega\left(t-\frac{r}{v}\right)}{v}\right]\left[-\frac{\omega}{v}\right]+\frac{1}{r^2}\left[(r)\sin\omega\left(t-\frac{r}{v}\right)\left(\frac{\omega}{v}\right)-\cos\omega\left(t-\frac{r}{v}\right)\right]\right\}$$

$$\therefore \quad (\nabla \times \overline{H})_\theta = -\frac{I\,dL\sin\theta}{4\pi}\left[\frac{\omega^2\cos\omega\left(t-\frac{r}{v}\right)}{v^2 r}+\frac{\omega\sin\omega\left(t-\frac{r}{v}\right)}{v r^2}-\frac{\cos\omega\left(t-\frac{r}{v}\right)}{r^3}\right] \quad \ldots (15)$$

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

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

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

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

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

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

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

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

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

$$E_\theta = \frac{1}{\varepsilon}\int (\nabla\times\bar{H})_\theta\, dt$$

Substituting the value of $(\nabla\times\bar{H})_\theta$ from equation (15),

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

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

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

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

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

Significance of field components:

In this section the significance of each term in the expressions for the field components are describe. Let us rewrite the expressions for the field components. There is only one component for the magnetic field, in $\bar{a}_\phi$ direction given by,

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

There are two components for the electric field, in $\bar{a}_r$ and $\bar{a}_\theta$ direction, given by,

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

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

3. **Derive the expression for the field quantities radiated from half wave dipole and provide its radiation resistance. (16M)**
[CO1-H1-April/May2011, Nov/Dec 2012 April/May 2013, April/May 2015]

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

$$H_r = 0 \quad\text{.....................................}(7.37d)$$

$$H_\theta = -\frac{E_\phi}{\eta} = \frac{j\omega}{\eta}A_\phi \quad\text{.....................................}(7.37e)$$

$$H_\phi = \frac{E_\theta}{\eta} = -\frac{j\omega}{\eta}A_\theta \quad\text{.....................................}(7.37f)$$

The relationship stated above equation (7.37a) - (7.37f) may be verified for a Herzian dipole using equations (7.22), (7.24a) and (7.24b).

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

This distribution assumes that the current vanishes at the two end points i.e., $z' = \pm l/2$. The plot of current distribution are shown in the figure 7.7 for different 'l'.

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

$$I = I_0 \cos k_0 z' \qquad -\frac{\lambda_0}{4} \le z' \le \frac{\lambda_0}{4} \quad ....(7.39)$$

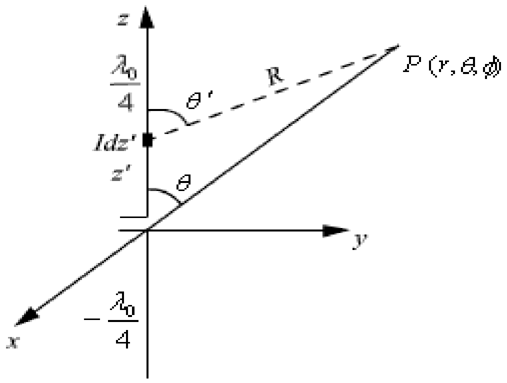
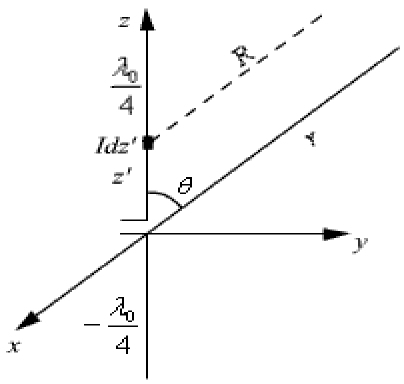


Fig7.8(a): Half wave dipole



7.8(b):Farfieldapproximation for half wave dipole

From equation (7.21) we can write

$$d\vec{A} = \hat{a}_z \frac{\mu_0 I(z')\,dz'}{4\pi} \frac{e^{-jk_0 R}}{R}$$ ....(7.40)

From Fig 7.8(b), for the far field calculation, $R \cong r - z'\cos\theta$ for the phase variation and $R \cong r$ for amplitude term.

$$\therefore\ d\vec{A} = \hat{a}_z \frac{\mu_0 I(z')\,dz'}{4\pi} \frac{e^{-jk_0 R}}{r} e^{jk_0 z'\cos\theta}$$ ....(7.41)

Substituting $I(z') = I_0 \cos k_0 z'$ from (7.39) to (7.41) we get

$$d\vec{A} = \hat{a}_z \frac{\mu_0 I_0}{4\pi} \frac{e^{-jk_0 r}}{r} \cos k_0 z' e^{jk_0 z'\cos\theta} dz'$$ ....(7.42)

Therefore the vector potential for the halfwave dipole can be written as:

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

$$\vec{A} = \hat{a}_z \frac{\mu_0 I_0}{4\pi} \frac{e^{-jk_0 r}}{r} \frac{2\cos\left(\frac{\pi}{2}\cos\theta\right)}{k_0 \sin^2\theta}$$ ....(7.43)

From (7.37b),

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

$$= jI_0 \frac{\omega k_0}{2\pi k_0} \frac{e^{-jk_0 r}}{r} \frac{\cos(\pi/2\cos\theta)}{k_0 \sin^2\theta}$$

$$= j\frac{I_0 \eta_0}{2\pi} \frac{e^{-jk_0 r}}{r} \cos\frac{(\pi/2\cos\theta)}{\sin\theta}$$ ....(7.44)

Similarly from (7.37c)

$$E_\phi = 0 \quad \text{...(7.45)}$$

and from (7.37e) and (7.37f)

$$H_\phi = \frac{jI_0}{2\pi}\frac{e^{-jk_0 r}}{r}\frac{\cos\left(\pi/2\cos\theta\right)}{\sin\theta} \quad \text{................................(7.46)}$$

and $H_\theta = 0$ ....(7.47)

The radiated power can be computed as

$$P_r = \int_0^{2\pi}\int_0^{\pi}\frac{1}{2}\left(\vec{E}\times\vec{H^*}\right)r^2\sin\theta\, d\theta\, d\phi$$

$$= 36.565\left|I_0\right|^2 \quad \text{...(7.48)}$$

Therefore the radiation resistance of the half wave dipole antenna is $36.565\times 2\ \Omega$= $73.13\ \Omega$

Further, using Eqn(7.27) the directivity function for the dipole antenna can be written as

$$D\left(\theta,\phi\right) = 1.64\left[\frac{\cos\left(\pi/2\cos\theta\right)}{\sin\theta}\right]^2 \quad \text{...........................(7.49)}$$

Thus directivity of such dipole antenna is 1.64 as compared to 1.5 for an elementary dipole. The half power beam width in the E-plane can be found to be 78^0 as compared to 90^0 for a Hertzian dipole.

4. Explain the construction and working principle of Yagi Uda Antenna and derive its impedance.(16M)
[CO1-L2-April/May 2011, Nov/Dec2011, April/May 2013, April/May 2015]

A **Yagi–Uda antenna**, commonly known as a **Yagi antenna**, is a directional antenna consisting of multiple parallel elements in a line, usually half-wave dipoles made of metal rods. Yagi–Uda antennas consist of a single driven element connected to the transmitter or receiver with a transmission line, and additional parasitic elements: a so-called *reflector* and one or more *directors*.

The reflector element is slightly longer than the driven dipole, whereas the directors are a little shorter. This design achieves a very substantial increase in the antenna's directionality and gain compared to a simple dipole.

Also called a "beam antenna", the Yagi is very widely used as a high-gain antenna on the HF, VHF and UHF bands It has moderate gain which depends on the number of elements used, typically limited to about 17 dBi, linear polarization,unidirectional (end-fire) beam pattern with high front-to-back ratio of up to 20 db. and is lightweight, inexpensive and simple to construct.

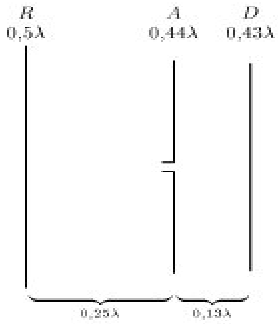
Description:

The Yagi–Uda antenna consists of a number of parallel thin rod elements in a line, usually half-wave long, typically supported on a perpendicular crossbar or "boom" along their centers. There is a single driven element driven in the center (consisting of two rods each connected to one side of the transmission line), and a variable number of parasitic elements, a single *reflector* on one side and optionally one or more *directors* on the other side. The parasitic elements are not electrically connected to the transmitter or receiver, and serve as passive radiators, reradiating the radio waves to modify the radiation pattern. Typical spacings between elements vary from about 1/10 to 1/4 of a wavelength, depending on the specific design. The directors are slightly shorter than the driven element, while the reflector(s) are slightly longer. The radiation pattern is unidirectional, with the main lobe along the axis perpendicular to the elements in the plane of the elements, off the end with the directors.

Conveniently, the parasitic elements have a node (point of zero RF voltage) at their centre, so they can be attached to a conductive metal support at that point without need of insulation, without disturbing their electrical operation. They are usually bolted or welded to the antenna's central support boom. The driven element is fed at centre so its two halves must be insulated where the boom supports them.

The gain increases with the number of parasitic elements used. Only one reflector is used since the improvement of gain with additional reflectors is negligible, but Yagis have been built with up to 30–40 directors.

The bandwidth of the antenna is the frequency range between the frequencies at which the gain drops 3 dB (one-half the power) below its maximum. The Yagi–Uda array in its basic form has very narrow bandwidth, 2–3 percent of the centre frequency.There is a tradeoff between gain and bandwidth, with the bandwidth narrowing as more elements are used.For applications that require wider bandwidths, such as terrestrial television, Yagi–Uda antennas commonly feature trigonal reflectors, traps (described below), and larger diameter conductors, in order to cover the relevant portions of the VHF and UHF bands.



Yagi–Uda antennas used for amateur radio are sometimes designed to operate on multiple bands. These elaborate designs create electrical breaks along each element (both sides) at which point a parallel LC (inductor and capacitor) circuit is inserted. This so-called *trap* has the effect of truncating the element at the higher frequency band, making it approximately a half wavelength in length. At the lower frequency, the entire element (including the remaining inductance due to the trap) is close to half-wave resonance, implementing a *different* Yagi–Uda antenna. Using a second set of traps, a "triband" antenna can be resonant at three different bands. Given the associated costs of erecting an antenna and rotor system above a tower, the combination of antennas for three amateur bands in one unit is a very practical solution. The use of traps is not without disadvantages, however, as they reduce the bandwidth of the antenna on the individual bands and reduce the antenna's electrical efficiency and subject the antenna to additional mechanical considerations (wind loading, water and insect ingress).

Theory of Operation:

Consider a Yagi–Uda consisting of a reflector, driven element and a single director as shown here. The driven element is typically a λ/2 dipole or folded dipole and is the only member of the structure that is directly excited (electrically connected to the feedline). All the other elements are considered *parasitic*. That is, they reradiate power which they receive from from the driven element (they also interact with each other).

One way of thinking about the operation of such an antenna is to consider a parasitic element to be a normal dipole element fed at its centre, with a short circuit across its feed point. As is well known in transmission linetheory, a short circuit reflects all of the incident power 180 degrees out of phase. So one could as well model the operation of the parasitic element as the superposition of a dipole element receiving power and sending it down a transmission line to a matched load, and a transmitter sending the same amount of power up the transmission line back toward the antenna element.

If the transmitted voltage wave were 180 degrees out of phase with the received wave at that point, the superposition of the two voltage waves would give zero voltage, equivalent to shorting out the dipole at the feedpoint (making it a solid element, as it is). Thus a half-wave parasitic element radiates a wave 180° out of phase with the incident wave.

The fact that the parasitic element involved is not exactly resonant but is somewhat shorter (or longer) than λ/2 modifies the phase of the element's current with respect to its excitation from the driven element.

The so-called *reflector* element, being longer than λ/2, has an inductive reactance which means the phase of its current lags the phase of the open-circuit voltage that would be induced by the received field. The *director* element, on the other hand, being shorter than λ/2, has a capacitive reactance with the voltage phase lagging that of the current.

The elements are given the correct lengths and spacings so that the radio waves radiated by the driven element and those reradiated by the parasitic elements all arrive at the front of the antenna in phase, so they superpose and add, increasing signal strength in the forward direction. In other words, the crest of the forward wave from the reflector element reaches the driven element just as the crest of the wave is emitted from that element. These waves reach the first director element just as the crest of the wave is emitted from that element, and so on. The waves in the reverse direction interfere destructively, cancelling out, so the signal strength radiated in the reverse direction is small. Thus the antenna radiates a unidirectional beam of radio waves from the front (director end) of the antenna.

While the above qualitative explanation is useful for understanding how parasitic elements can enhance the driven elements' radiation in one direction at the expense of the other, the assumptions used are quite inaccurate. Since the so-called reflector, the longer parasitic element, has a current whose phase lags that of the driven element, one would expect the directivity to be in the direction of the reflector, opposite of the actual directional pattern of the Yagi–Uda antenna. In fact, that would be the case were we to construct a phased array with rather closely spaced elements all driven by voltages in phase, as we posited.

However these elements are not driven as such but receive their energy from the field created by the driven element, so we will find almost the opposite to be true. For now, consider that the parasitic element is also of length λ/2. Again looking at the parasitic element as a dipole which has been shorted at the feedpoint, we can see that if the parasitic element were to respond to the driven element with an open-circuit feedpoint voltage in phase with that applied to the driven element (which we'll assume for now) then the *reflected* wave from the short circuit would induce a current 180° out of phase with the current in the driven element. This would tend to cancel the radiation of the driven element. However, due to the reactance caused by the length difference, the phase lag of the current in the reflector, added to this 180° lag, results in a phase *advance*, and vice versa for the director. Thus the directivity of the array indeed is in the direction towards the director.

One must take into account an additional phase delay due to the finite distance between the elements which further delays the phase of the currents in both the directors and reflector(s). The case of a Yagi–Uda array using just a driven element and a director is illustrated in the accompanying diagram taking all of these effects into account. The wave generated by the driven element (green) propagates in both the forward and reverse directions (as well as other directions, not shown). The director receives that wave slightly delayed in time (amounting to a phase delay of about 35° which will be important for the reverse direction calculations later), and generating a current that would be out of phase with the driven element (thus an additional 180° phase shift), but which is further *advanced* in phase (by about 70°) due to the director's shorter length. In the forward direction the net effect is a wave emitted by the director (blue) which is about 110° (180°–70°) retarded with respect to that from the driven element (green), in this particular design. These waves combine to produce the net forward wave (bottom, right) with an amplitude slightly larger than the individual waves.

In the reverse direction, on the other hand, the additional delay of the wave from the director (blue) due to the spacing between the two elements (about 35° of phase delay traversed twice) causes it to be about 180° (110° + 2 × 35°) out of phase with the wave from the driven element (green). The net effect of these two waves, when added (bottom, left), is almost complete cancellation. The combination of the director's position and shorter length has thus obtained a unidirectional rather than the bidirectional response of the driven (half-wave dipole) element alone.

A full analysis of such a system requires computing the *mutual impedances* between the dipole elements[11] which implicitly takes into account the propagation delay due to the finite spacing between elements. We model element number *j* as having a feedpoint at the centre with a voltage V_j and a current I_j flowing into it. Just considering two such elements we can write the voltage at each feedpoint in terms of the currents using the mutual impedances Z_{ij}:

$$V_1 = Z_{11} I_1 + Z_{12} I_2$$

$$V_2 = Z_{21} I_1 + Z_{22} I_2$$

Z_{11} and Z_{22} are simply the ordinary driving point impedances of a dipole, thus 73 + j43 ohms for a half-wave element (or purely resistive for one slightly shorter, as is usually desired for the driven element). Due to the differences in the elements' lengths Z_{11} and Z_{22} have a substantially different reactive component. Due to reciprocity we know that $Z_{21} = Z_{12}$.

Now the difficult computation is in determining that mutual impedance Z_{21}which requires a numerical solution. This has been computed for two exact half-wave dipole elements at various spacings in the accompanying graph.

$$0 = V_2 = Z_{21} I_1 + Z_{22} I_2$$

and so

$$I_2 = -\frac{Z_{21}}{Z_{22}} I_1 .$$

This is the current induced in the parasitic element due to the current I_1 in the driven element. We can also solve for the voltage V_1 at the feedpoint of the driven element using the earlier equation:

$$V_1 = Z_{11} I_1 + Z_{12} I_2 = Z_{11} I_1 - Z_{12} \frac{Z_{21}}{Z_{22}} I_1$$
$$= \left(Z_{11} - \frac{Z_{21}^2}{Z_{22}} \right) I_1$$

$$V_1 = Z_{11} I_1 + Z_{12} I_2 = Z_{11} I_1 - Z_{12} \frac{Z_{21}}{Z_{22}} I_1$$
$$= \left(Z_{11} - \frac{Z_{21}^2}{Z_{22}} \right) I_1$$

where we have substituted $Z_{12} = Z_{21}$. The ratio of voltage to current at this point is the *driving point impedance* Z_{dp} of the 2-element Yagi:

$$Z_{dp} = V_1 / I_1 = Z_{11} - \frac{Z_{21}^2}{Z_{22}}$$

Unit- II

Aperture and slot Antennas

Part – A

1. State Huygens Principle? [CO2-L2- April/May 2012, April/May 2014]

The Huygen's principle states that each point of an advancing wavefront is in fact the centre of a fresh disturbance and the source of a new train of waves.

2. What are the different types of horn antennas? [CO2-L1]

i)E plane sectorial horn.
ii) H plane sectorial horn
iii)Pyramidal horn.
iv)Conical horn

3. List the applications of parabolic reflector. [CO2-L2]

i)Radio astronomy.ii)Microwave communication. Iii)Satellite tracking

4. What is Slot Antenna?
[CO2-L1- April/May 2012, April/May 2013, May/June 2016]

The slot antenna consists of a radiator formed by cutting a narrow slot in a large metal surface.The shape and size of the slot,as well as driving frequency determine the radiation pattern.

5. What is parabolic reflector? [CO2-L1]

It is a parabola shaped reflective devices used to distribute energy entering the reflector at a particular angle.

6.State Uniqueness theorem. [CO2-L2- April/May 2013]

The uniqueness theorem can be stated in several different forms but it essentially states that for a given set of sources and boundary conditions in a lossy medium, the solution to Maxwell's equations is unique.

7.State field equivalence principle. [CO2-L2]

According to the field equivalence principle, 'the fields in V2 due to the sources in volume V 1 can also be generated by an equivalent set of virtual sources on surface S, given by J s = n x H and M, = E x n, where E and H are the fields on the surface S produced by the original set of sources in volume V 1'

8.Which antenna is complementary to the slot antenna? [CO2-L2]

The dipole antenna is complementary to the slot antenna. The metal and air regions of the slot are interchanged for the dipole.

9. What is the relationship between the terminal impedances of slot and dipole antennas? [CO2-L2- April/May 2012, April/May 2013]

$Z_s Z_d = \eta_0^2 / 4$

Where ZS = terminal impedance of slot antenna

Zd = Terminal impedance if dipole antenna

η_0 = Intrinsic impedance of free space = 377 ohms

10. What is the difference between slot antenna and its complementary antenna? [CO2-L1- Nov/Dec 2013, April/May 2014]

i. Polarization are different ie. The electric fields associated with the slot antenna are identical with the magnetic field of the complementary dipole antenna.

ii. The electric field be vertically polarized for the slot and horizontally polarized for the dipole.

iii.Radiation from the back side of the conducting plane of the slot antenna has opposite polarity from that of complementary antenna.

11. What are the methods of feeding slot antenna? [CO2-L1]

i) Coaxial line feed

ii) Waveguide feed

12. What are the different methods of feeding slot antennas? [CO2-L1-April/May 2015]

Slot antenna can be

a. Waveguide fed slot

b. Boxed in slot

c. Coaxial transmission line

13.Define spill over. [CO2-L1]

Some of the desired rays are not captured by the reflector antenna and this constitute spill over.

14.What is back lobe radiation? [CO2-L1]

Some radiation from the primary radiator occurs in the forward direction in addition to the desired parallel beam. This is known as back lobe radiation.

15.What are various feeds used in reflector? [CO2-L1]

a. Dipole antenna

b. Horn feed

c. End fire feed

d. Cassegrain feed

16.What are the advantages of cassegrain feed antenna? [CO2-L1]

1. Reduction in spill over
2. Simple in construction
3. Quite inexpensive
4. Widely used in fixed point to point microwave communication
5. Satellite reception and tracking
6. Ability to place feed in a convenient location

17.What are the different types of horn antenna? [CO2-L2]

a. Sectoral horn

b. Pyramidal horn

c. Conical horn

d. Biconical horn antenna

18.What are the uses of horn antenna? [CO2-L1]

Horn antenna are extensively used at microwave frequencies under the condition that power gain needed is moderate. For high power gain, since the horn dimensions become large, so the other antennas like lens or parabolic reflector etc. are preferred rather than horn.

19. What are Microstrip patch antennas? [CO2-L1- April/May 2015]

Microstrip antennas are popular for low profile applications at frequencies above 100 MHz. they usually consists of a rectangular metal patch on a dielectric coated ground plane (circuit board). Hence, a microstrip patch antenna is also called as printed antenna.

20.What are the advantages of Microstrip patch antenna? [CO2-L1]

i. These antennas can be flush mounted to metal or other existing surfaces

ii. Small size and weight

iii.Linear and circular polarization are possible.

iv.They only require space for the feed line which is normally placed behind the ground plane.

21.What are the disadvantages of Microstrip patch antenna? [CO2-L1]

i. Inefficiency

ii. Very narrow frequency bandwidth

22. State Babinet's principle.
[CO2-L2-April/May 2012, April/May 2014, Nov/Dec 2015]

The babinet's principal states that, "The field at any point behind a plane having a screen, if added to the field at the same point when the complementary field is substituted, in the same as the field at the point when there is no screen.

23. Define corner reflector. [CO2-L1]

A corner reflector is a retro reflector consisting of two or three mutually perpendicular, intersecting flat surfaces which reflects EM waves back towards the source.

24. List the advantage of cassegrain feed system. [CO2-L2]

i. It reduces the spillover and thus minor lobe radiations.

ii.The system has ability to place a feed at convenient place.

iii.Using this system, beam can be broadend by adjusting one of the reflector surfaces.

iv.With this system greater focal length greater than the physical focal length can be achieved.

25. Mention the applications of microstrip antennas. [CO2-L3]

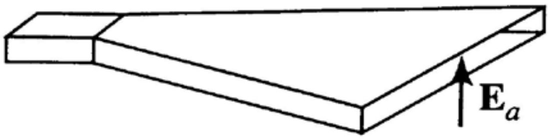
i. High velocity aircrafts, space crafts, missiles and rockets.

ii. Missile guiding, fuzing, telemetry, Satellite communication, radars etc.

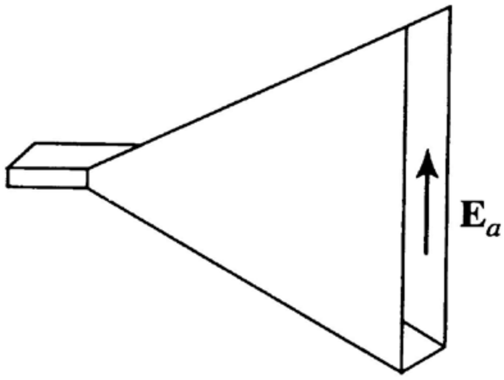
Part-B

1. What is Horn Antenna? Sketch the various types of Horn Antenna and explain its operation. (16M) [CO2-L2- Nov/Dec 2011, April/May 2014]

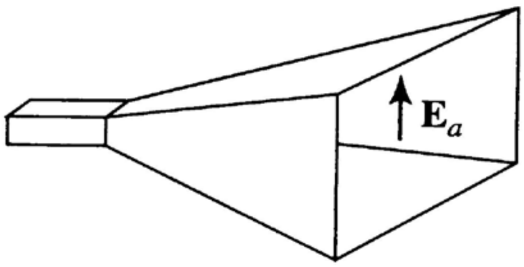
Rectangular Horn Antennas Horn antennas are popular in the microwave band (above 1 GHz). Horns provide high gain, low VSWR (with waveguide feeds), relatively wide bandwidth, and they are not difficult to make. There are three basic types of rectangular horns.



(*a*) *H*-plane sectoral horn.



(*b*) *E*-plane sectoral horn.

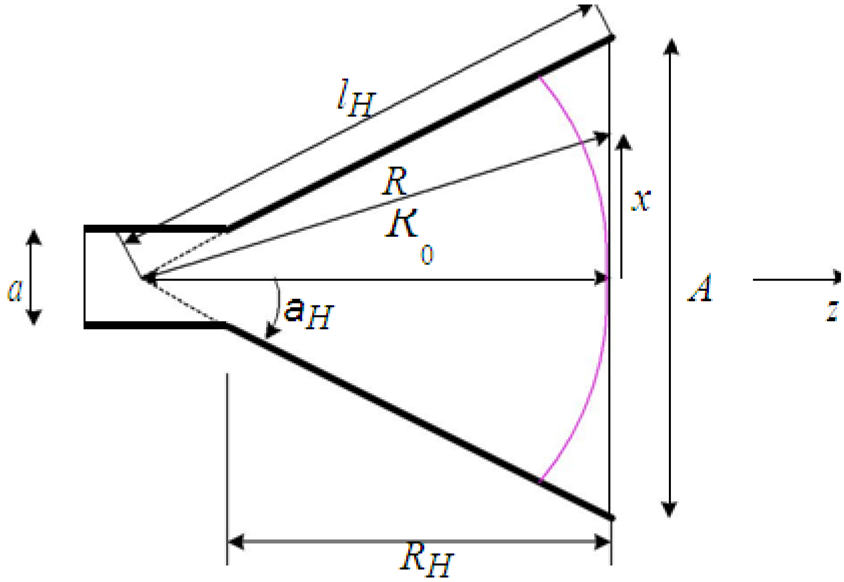


(*c*) Pyramidal horn.

The rectangular horns are ideally suited for rectangular waveguide feeders. The horn acts as a gradual transition from a waveguide mode to a free-space mode of the EM wave. When the feeder is a cylindrical waveguide, the antenna is usually a ***conical horn***.

The *H*-plane sectoral horn

The geometry and the respective parameters shown in the figure below are used often in the subsequent analysis.



H-plane (*x*-*z*) cut of an *H*-plane sectoral horn

The tangential field arriving at the input of the horn is composed of the transverse field components of the waveguide dominant mode TE_{10}:

$$E_y(x) = E_0 \cos\left(\frac{\pi}{a}x\right) e^{-j\beta_g z} \quad (18.4)$$

Here, $\beta_0 = \omega\sqrt{\mu\varepsilon} = 2\pi/\lambda$, and λ is the free-space wavelength. The field that is illuminating the aperture of the horn is essentially a spatially expanded version of the waveguide field. Note that the wave impedance of the flared waveguide (the horn) gradually approaches the intrinsic impedance of open space η, as *A* (the *H*-plane width) increases.

The complication in the analysis arises from the fact that the waves arriving at the horn aperture are ***not in phase*** due to the different path lengths from the horn apex. The aperture phase variation is given by

$$e^{-j\beta(R-R_0)}. \quad (18.5)$$

Since the aperture is not flared in the *y*-direction, the phase is uniform in this direction. We first approximate the path of the wave in the horn:

The last approximation holds if $\sqrt{\ } x\ 0\ R_0$, or $A / 2\ 0\ R_0$. Then, we can assume that

$$R - R_0 \approx \frac{1}{2}\frac{x^2}{R_0} . \qquad (18.7)$$

Using (18.7), the field at the aperture is approximated as

$$E_{ay}(x) \approx E_0 \cos \frac{\pi}{A} x\, e^{-j\frac{\beta}{2R_0}x^2} . \qquad (18.8)$$

The field at the aperture plane outside the aperture is assumed equal to zero. The field expression (18.8) is substituted in the integral I_y^E :

$$I_y^E = \iint_{S_A} E_{a_y}(x', y') e^{j\beta(x'\sin\theta\cos\varphi + y'\sin\theta\sin\varphi)} dx' dy' , \qquad (18.9)$$

$$I_y^E = E_0 \int_{-A/2}^{+A/2} \cos\frac{\pi}{A} x' e^{-j\frac{b}{2R_0}x'^2} e^{j b x'\sin\theta\cos\varphi} dx' \times \int_{-b/2}^{+b/2} e^{j b y'\sin\theta\sin\varphi} dy'. \qquad (18.10)$$

$-I(\theta,\varphi)$

C (*x*) and *S* (*x*) are Fresnel integrals, which are defined as

$$C(x) = \int_0^x \cos\frac{\pi}{2}\tau^2\, d\tau; \qquad C(-x) = -C(x),$$

$$S(x) \int_0^x \sin\frac{\pi}{2}\tau^2 \qquad S(-x) = -S(x).$$

More accurate evaluation of I_y^E can be obtained if the approximation in (18.6) is not made, and E_{ay} is substituted in (18.9) as

$$E_{ay}(x) = E_0 \cos\frac{\pi}{A}x\, e^{-j\beta\left(\sqrt{R_0^2 + x^2} - R_0\right)} = E_0\, e^{+j\beta R_0} \cos\frac{\pi}{A}x$$

$e^{-j\beta\sqrt{R_0^2 + x^2}}$. The far field can be now calculated as:

$$E_\theta = j\beta \frac{e^{-j\beta r}}{4\pi r}(1 + \cos\theta)\sin j \cdot I_y^E,$$

$$E_j = j\beta \frac{e^{-j\beta r}}{4\pi r}(1 + \cos\theta)\cos j \cdot I_y^E,$$

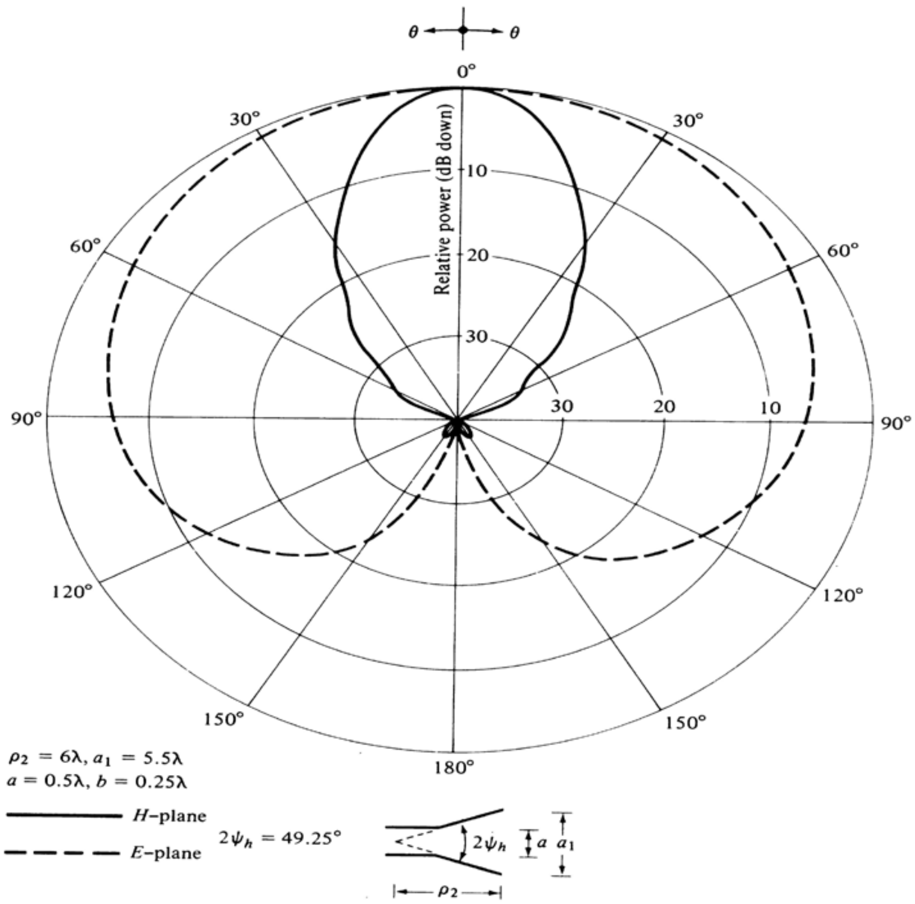
The amplitude pattern of the *H*-plane sectoral horn is obtained as

$$\bar{E} = \frac{1 + \cos\theta}{2} \cdot \frac{\sin\left(\frac{bb}{2}\sin\theta\sin\phi\right)}{bb} \cdot I(\theta, \phi). \qquad (18.17)$$

Principal-plane patterns

***H*-plane** ($\phi = 0°$):

$$F_H(\theta) = \frac{1 + \cos\theta}{2} \cdot \frac{I(\theta, \phi = 0°)}{I(\theta = 0°, \phi = 0°)} \qquad (18.18)$$



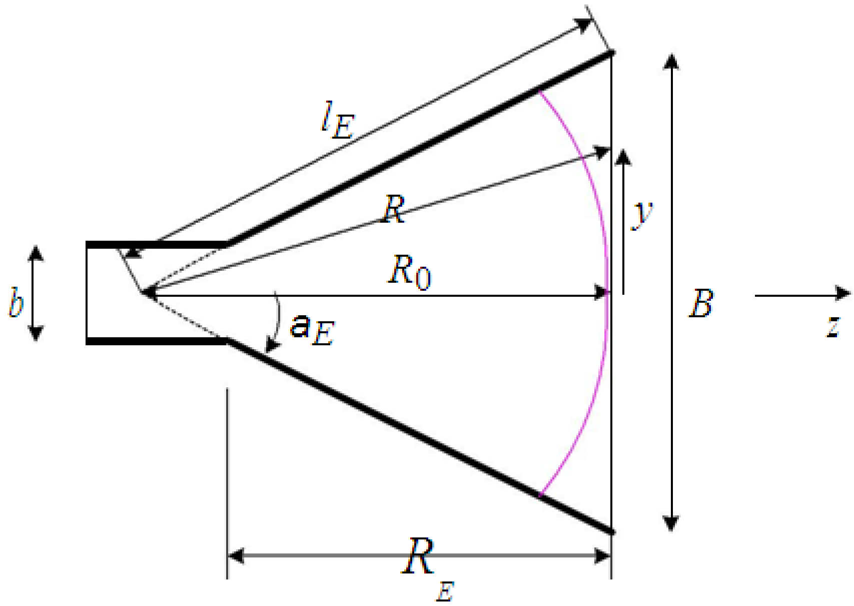
The directivity of the *H*-plane sectoral horn is calculated by the general directivity expression for apertures

$$D_0 = \frac{4\pi}{\lambda^2} \cdot \frac{\left| \iint_{S_A} \mathbf{E}_a ds' \right|^2}{\iint_{S_A} |\mathbf{E}_a|^2 ds'} . \tag{18.19}$$

It can be shown that the optimal directivity is obtained if the relation between *A* and R_0 is

$$A = \sqrt{3\lambda R_0} ,$$

The *E*-plane sectoral horn



E-plane (*y*-*z*) cut of an *E*-plane sectoral horn

The geometry of the *E*-plane sectoral horn in the *E*-plane (*y*-*z* plane) is analogous to that of the *H*-plane sectoral horn in the *H*-plane. The analysis is following the same steps as in the previous section. The field at the aperture is approximated by [compare with (18.8)]

$$E_{ay} = E_0 \cos\left(\frac{\pi}{a} x\right) e^{-j\frac{\beta}{2R_0} y^2} .$$

Here, the approximations

$$R = \sqrt{R_0^2 + y^2} = R_0 \sqrt{1 + \left(\frac{y}{R_0}\right)^2} \approx R_0 \left[1 + \frac{1}{2}\left(\frac{y}{R_0}\right)^2\right]$$

and

$$R - R_0 \approx \frac{1}{2}\frac{y^2}{R_0}$$

are made, which are analogous to (18.6) and (18.7).

The arguments of the Fresnel integrals used in (18.29) are

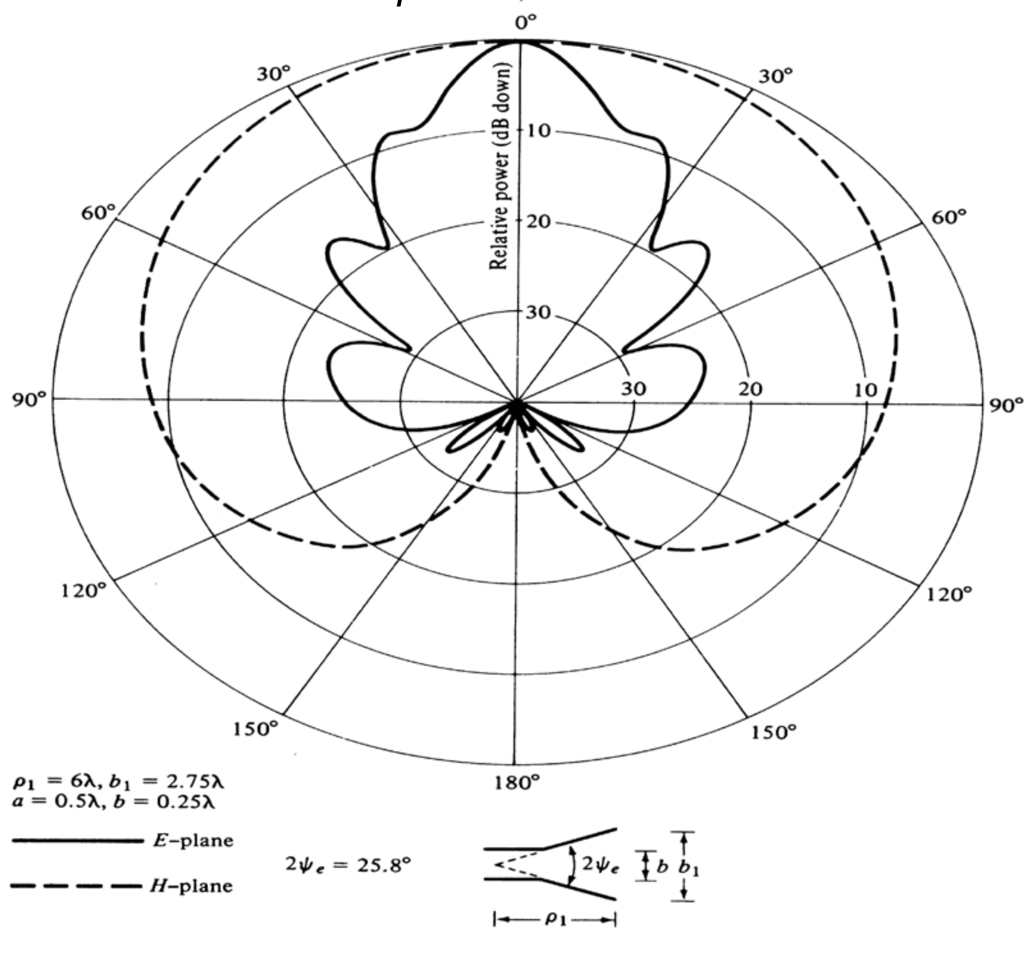
$$r_1 = \sqrt{\frac{\beta}{\pi R_0}}\left(-\frac{B}{2} - R_0 \frac{\beta B}{2} \sin\theta \sin\phi\right),$$

$$r_2 = \sqrt{\frac{\beta}{\pi R_0}}\left(+\frac{B}{2} - R_0 \frac{\beta B}{2} \sin\theta \sin\phi\right).$$

Directivity

The directivity of the *E*-plane sectoral horn is found in a manner analogous to the *H*-plane sectoral horn:

$$D_E = \frac{a}{\lambda}\frac{32}{\pi}\frac{B}{\lambda}\varepsilon_{ph}^{E} = \frac{4\pi}{\lambda^2}\varepsilon_t \varepsilon_{ph}^{E} aB, \qquad (18.36)$$



The optimal relation between the flared height B and the horn apex length R_0 that produces the maximum possible directivity is

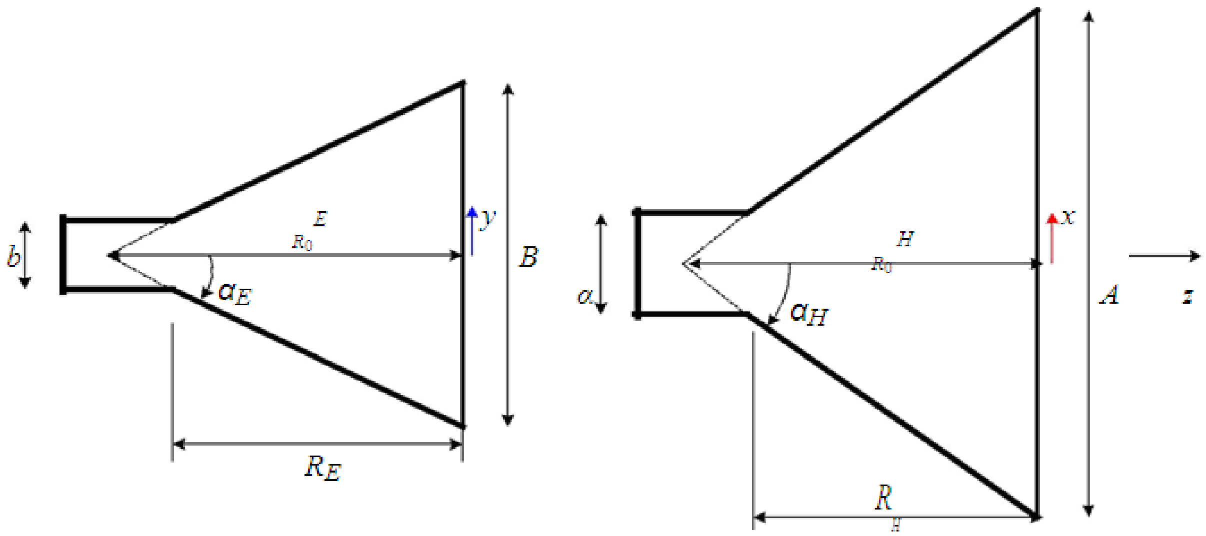
$$B = 2\lambda R_0$$

Optimum Pyramidal horn design

Usually, the optimum (from the point of view of maximum gain) design of a horn is desired because it results in the shortest axial length. The whole design can be actually reduced to the solution of a single fourth-order equation. For a horn to be realizable, the following must be true:

$$R_E = R_H = R_P \,. \quad (18.42)$$

The figures below summarize the notations used in describing the horn's geometry.



It can be shown that

$$\frac{R_0^H}{R_H} = \frac{A/2}{A/2 - a/2} = \frac{A}{A-a} \,, \quad (18.43)$$

$$\frac{R_0^E}{R_E} = \frac{B/2}{B/2 - b/2} = \frac{B}{B-b} \,. \quad (18.44)$$

The optimum-gain condition in the *E*-plane (18.37) is substituted in (18.44) to produce

$$B^2 - bB - 2\lambda R_E = 0\,. \qquad (18.45)$$

There is only one physically meaningful solution to (18.45):

$$B = \frac{1}{2}\left(b + \sqrt{b^2 + 8\lambda R_E}\right). \qquad (18.46)$$

Similarly, the maximum-gain condition for the *H*-plane of (18.24) together with (18.43) yields

$$R_H = \frac{A - a}{A}\,\frac{A^2}{3\lambda} = A\,\frac{(A - a)}{3\lambda}\,. \qquad (18.47)$$

Since $R_E = R_H$ must be fulfilled, (18.47) is substituted in (18.46), which gives

$$B = \frac{1}{2}\left(b + \sqrt{b^2 + \frac{8A(A-a)}{3}}\right). \qquad (18.48)$$

Substituting in the expression for the horn's gain,

$$G = \frac{4\pi}{\lambda^2}\varepsilon_{ap} AB\,, \qquad (18.49)$$

gives the relation between *A*, the gain *G*, and the aperture efficiency ε_{ap} :

$$G = \frac{4\pi}{\lambda^2}\varepsilon_{ap} A \frac{1}{2}\left(b + \sqrt{b^2 + \frac{8A(a-a)}{3}}\right), \qquad (18.50)$$

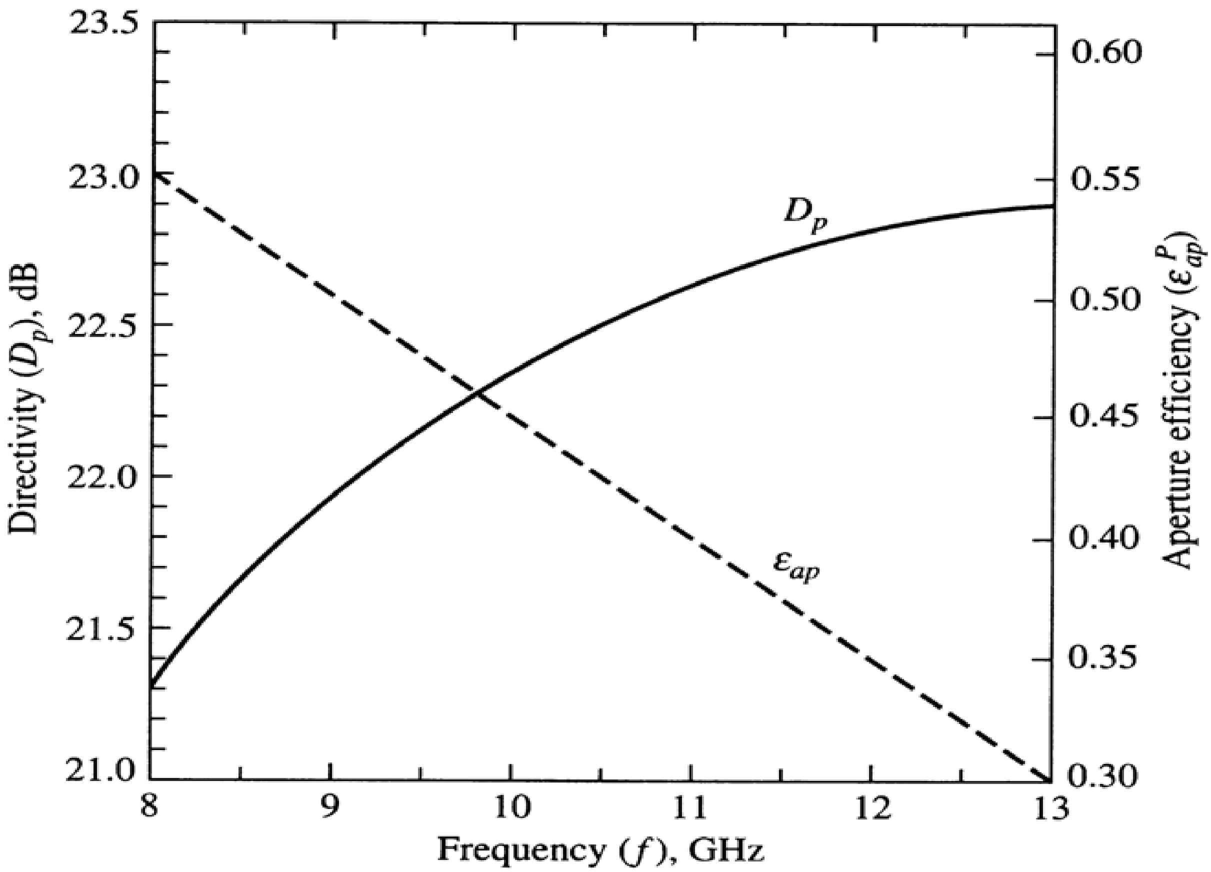
$$\Rightarrow A^4 - aA^3 + \frac{3bG\lambda^2}{8\pi\varepsilon_{ap}}A - \frac{3G^2\lambda^4}{32\pi^2\varepsilon_{ap}^2} = 0\,. \qquad (18.51)$$

Equation (18.51) is the optimum pyramidal horn design equation. The optimum-gain value of $\varepsilon_{ap} = 0.51$ is usually used, which makes the equation a fourth-order polynomial equation in *A*. Its roots can be found analytically (which is not particularly easy) and numerically. In a numerical solution, the first guess is usually set at $A^{(0)} = 0.45\lambda\sqrt{G}$. Once *A* is found, *B* can be computed from (18.48) and $R_E = R_H$ is computed from (18.47).

Sometimes, an optimal horn is desired for a known axial length R_0. In this case, there is no need for nonlinear-equation solution. The design procedure follows the steps: (a) find *A* from (18.24), (b) find *B* from (18.37), and (c) calculate the gain *G* using (18.49) where $\varepsilon_{ap} = 0.51$.

Horn antennas operate well over a bandwidth of 50%. However, gain performance is optimal only at a given frequency. To understand better the frequency dependence of the directivity and the aperture efficiency, the plot of these curves for an X-band (8.2 GHz to 12.4 GHz) horn fed by WR90 waveguide is given below (*a* = 0.9 in. = 2.286 cm and *b* = 0.4 in. = 1.016 cm).

The gain increases with frequency, which is typical for aperture antennas. However, the curve shows saturation at higher frequencies. This is due to the decrease of the aperture efficiency, which is a result of an increased phase difference in the field distribution at the aperture.



2. **Paraphrase the working of slot antenna. What is the terminal impedance of slot antenna and obtain its radiation pattern. (16M)**

[CO2-L2-Nov/Dec 2011, May/June 2012, April/May 2014]

The slot antenna consists of a radiator formed by cutting a narrow slot in a large metal surface. Suchan antenna is shown in figure 3-18. The slot length is a half wavelength at the desired frequency and thewidth is a small fraction of a wavelength. The antenna is frequently compared to a conventional half-wave dipole consisting of two flat metal strips. The physical dimensions of the metal strips are such thatthey would just fit into the slot cut out of the large metal sheet.

A slot is a narrow-width opening in a conductive sheet. When excited by a voltage across the narrow dimension it appears to radiate from an equivalent magnetic current flowing along the long dimension that replaces the voltage (or electric field) across it. Most slots, similar to dipoles, have a finite length with either short or open circuits at both ends. The voltage along the slot forms a standing wave. Of course, magnetic currents are fictitious, and real electric currents flow in the conductive sheet around the slot.

The characteristic impedance increases as the wave approaches the open-circuited ends. The slot is the dual of a strip dipole. A voltage excited across the slot propagates along a slotline toward short-circuited ends.

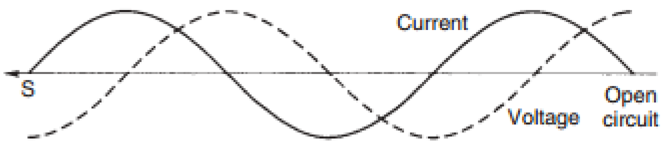
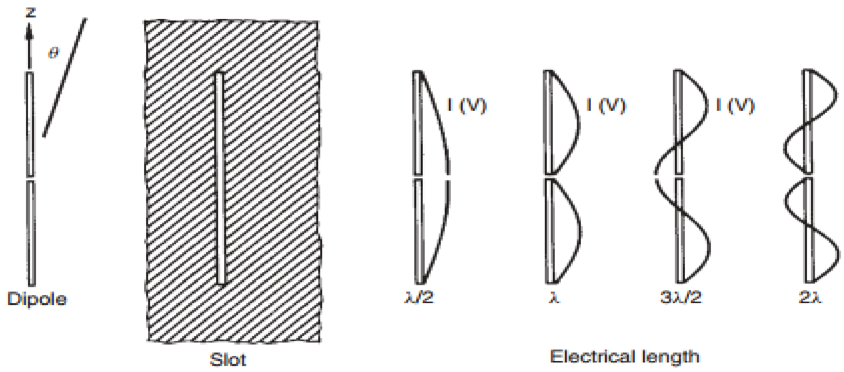


FIGURE 5-1 Standing wave.

Each type of transmission line reflects the incident wave from the terminations. The combination of two waves traveling in opposite directions creates a standing wave on the line. The current and voltage are 90◦ out of phase and 90◦ out of space phase (Figure 5-1). Current and voltage change places on the short-circuited termination of the slot. The dipole is not a uniform transmission line, but we can approximate the current as a standing wave with the current vanishing on the ends. The slot voltage is a standing wave also vanishing on the ends. The standing waves for a center-fed dipole or slot are expressed as follows:

Dipole	Slot	
$I = I_0 \sin k\left(\frac{L}{2} - z\right)$	$V = V_0 \sin k\left(\frac{L}{2} - z\right)$	$z \ge 0$
$I = I_0 \sin k\left(\frac{L}{2} + z\right)$	$V = V_0 \sin k\left(\frac{L}{2} + z\right)$	$z \le 0$

The voltage distribution on the slot is equivalent to a magnetic current. We calculate radiation from the linear sinusoidal current distributions by the vector potentials: electric (slot) (Section 2-1.2) and magnetic (dipole) (Section 2-1.1). Figure 5-2 gives typical sinusoidal distributions for various lengths. The currents match at the feed point and vanish on the ends. Consider the pattern of the 2λ dipole at θ = 90◦ . We can assume that it is a continuous array and sum the fields from each portion along the axis. The equal positive and negative portions of the standing-wave current sum to zero and produce a pattern null normal to the axis.



By integrating Eqs. (2-5) and (2-10), we compute far fields for radiators centered on the z-axis through the far-field conversion

$$E_\theta = j\eta \frac{I_0}{2\pi r} e^{-jkr} \frac{\cos(kL/2\cos\theta) - \cos(kL/2)}{\sin\theta} \qquad \text{dipole}$$

where L is the total dipole length. Using the Y = 0 plane as the slot ground plane, the far-field magnetic field is found as

$$H_\theta = \frac{\pm jV_0}{\eta 2\pi r} e^{-jkr} \frac{\cos(kL/2\cos\theta) - \cos(kL/2)}{\sin\theta} \qquad \text{slot}$$

where L is the total slot length. We apply the upper sign for Y > 0 and the lower sign for Y < 0. The electric field of the slot is found from $E_\varphi = -\eta H_\theta$. Equations (5-2) and (5-3) have the same pattern shape and directivity. We integrate the magnitude squared of Eqs. (5-2) and (5-3) to determine the average radiation intensity.

RADIATION RESISTANCE (CONDUCTANCE)

The far-field power densities, Poynting vectors, are given by

$$S_r = \begin{cases} \dfrac{|E_\theta|^2}{\eta} & \text{dipole} \\ |H_\theta|^2\eta & \text{slot} \end{cases}$$

where η is the impedance of free space (376.7 □). When these are integrated over the radiation sphere to compute the power radiated, the results contain either |I0| 2 (dipole) or |V0| 2 (slot), the maximum sinusoidal current (voltage). We define the radiation resistance (conductance) as

$$R_r = \frac{P_r}{|I_0|^2} \quad \text{dipole}$$
$$G_r = \frac{P_r}{|V_0|^2} \quad \text{slot} \tag{5-4}$$

Figure 5-4 is a plot of the radiation resistance of each versus length [2, p. 157]. The input resistance differs from the radiation resistance because it is the ratio of the input current (voltage) to the power radiated:

$$I_i = I_0 \sin\frac{kL}{2} \quad \text{dipole}$$
$$V_i = V_0 \sin\frac{kL}{2} \quad \text{slot} \tag{5-5}$$

Combining Eqs. (5-4) and (5-5), we find that

$$R_i = \frac{R_r}{\sin^2(kL/2)} \quad \text{dipole}$$
$$G_i = \frac{G_r}{\sin^2(kL/2)} \quad \text{slot} \tag{5-6}$$

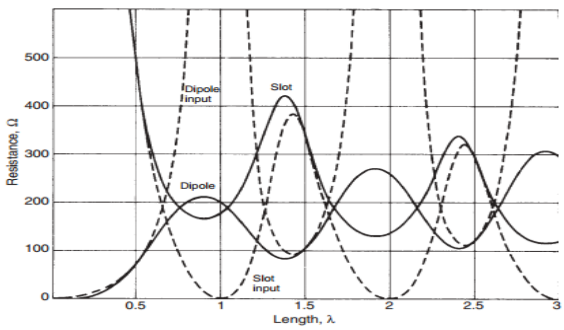


FIGURE 5-4 Dipole and slot radiation and center-fed input resistances.

The input resistances (Figure 5-4) differ from the radiation resistances by Eq. (5-6). The input resistance of a one-wavelength dipole is large but not infinite, as shown; it depends greatly on the diameter and input region. If we take the product of the radiation or input resistances, we determine that

$$R_{\text{dipole}} R_{\text{slot}} = \frac{\eta^2}{4}$$

The input resistance depends on the current at the input [Eq. (5-6)]. When the standing-wave current is high and the voltage is low, the input resistance is moderate. A center-fed half-wavelength dipole has the same input resistance as radiation resistance, since the current maximum occurs as the input. On the other hand, a center-fed halfwavelength slot has a current minimum (voltage maximum) at its input, which gives it high input resistance. When both are a full wavelength long, the dipole standingwave current is at a minimum and the slot standing-wave current is at a maximum (Figure 5-2). The dipole has a high input resistance and the slot has a low input resistance. We can lower the input resistance by feeding at a high current point, but we may excite a distribution different from that expected. A short dipole looks like a capacitor at the input. As the length increases, the radiation resistance grows and the capacitance decreases. Just before the length reaches λ/2, the capacitance becomes zero. The exact length at which the antenna resonates (zero reactance) depends on the diameter of the elements and the input gap. A good starting point is 95% of a half wavelength. Beyond the resonant length, the dipole becomes inductive. The impedance of a thin half-wavelength dipole is 73 + j42.2 □, whereas the resonant-length dipole resistance is about 67 □. The slot looks like an inductor when short.

The inductance increases as its length increases and the slot resonates like the dipole, just short of λ/2. Additional resonances occur at longer lengths. Increasing the frequency is equivalent to increasing the length for the thin dipole.

BABINET–BOOKER PRINCIPLE:

A strip dipole and a slot are complementary antennas. The solution for the slot can be found from the solution to an equivalent dipole by an interchange of the electric and magnetic fields. Not only the pattern but also the input impedance can be found. Figure 5-5 shows two such complementary structures.

Babinet's principle of optical screens (scalar fields) states that given the solutions to the diffraction patterns of a screen, Fi, and the screen's complement, Fc, the sum equals the pattern without the screen.

Booker extended Babinet's principle to vector electromagnetic fields. Strict complementation of an electric conductor requires a nonexistent magnetic conductor. Booker solved this problem by using only perfectly conducting infinitesimally thin screens and by interchanging the electric and magnetic fields between the screen and its complement. If we take two such complementary screens and perform line integrals over identical paths to compute the impedance of each, we obtain the result

$$Z_1 Z_c = \frac{\eta^2}{4} \qquad (5\text{-}8)$$

where Z1 is the input impedance of the structure, Zc the input impedance of the complementary structure, and η the impedance of free space (376.7 □). Equation (5-8) extends Eq. (5-7) to the total impedance and includes mutual impedances as well as self-impedances. Certain antennas, such as flat spirals, are self-complementary—an exchange of the spaces and conductors leaves the structure unchanged except for rotation. For a twoarm structure,

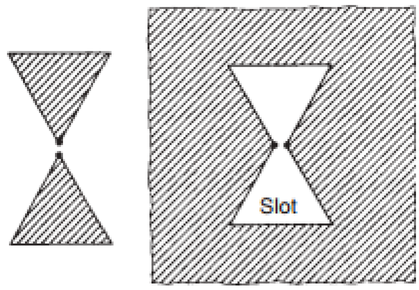


FIGURE 5-5 Complementary screens.

$$Z_0^2 = \frac{\eta^2}{4} \quad \text{or} \quad Z_0 = 188\ \Omega$$

Rumsey [5, p. 28] extended these ideas to antennas with more than two conductors to determine the input impedances in various feeding modes. We must relate flat-strip dipoles to normal round-rod dipoles to use the available results for round dipoles. The diameter of an equivalent round rod equals one-half the strip width of the flat structure. Consider a thin dipole with its near λ/2 resonance of 67 □. We calculate equivalent slot impedance from Eq. (5-8):

$$Z_{slot} = \frac{376.7^2}{4(67)} = 530\,\Omega$$

A half-wavelength slot impedance is

$$Z_{slot} = \frac{376.7^2}{4(73 + j42.5)} = 363 - j211\,\Omega$$

The λ/2 dipole is inductive when it is longer than a resonant length, whereas the slot is capacitive.

3. **Explain the special features of parabolic reflector antenna and discuss on the types of feed used with neat diagram. (16M)**

[CO2-L2- April/May 2013, May/June 2016]

A paraboloidal reflector transforms a spherical wave radiated by the feed located at its focus into a plane wave. Although the feed wave spreads from the focus, which reduces its amplitude, geometric optics predicts a plane wave reflection that remains constant. The reflected wave does not remain a plane wave but spreads because the fields must be continuous across the reflection boundary of the beam plane wave column because fields can be discontinuous only across physical boundaries. Nevertheless, we will use the aperture theory on the projected diameter to predict its performance. Since the reflected rays are parallel, we can place the aperture plane anywhere along the axis, but somewhat close in front of the reflector. The equations for the reflector surface are

$$r^2 = 4f(f + z) \quad \text{rectangular coordinates} \qquad \rho = \frac{f}{\cos^2(\psi/2)} \quad \text{polar coordinates}$$

where f is the focal length, D the diameter, ρ the distance from the focus to the reflector, and ψ the feed angle from the negative z-axis. The reflector depth from the rim to the center is z0 = D2/16f .

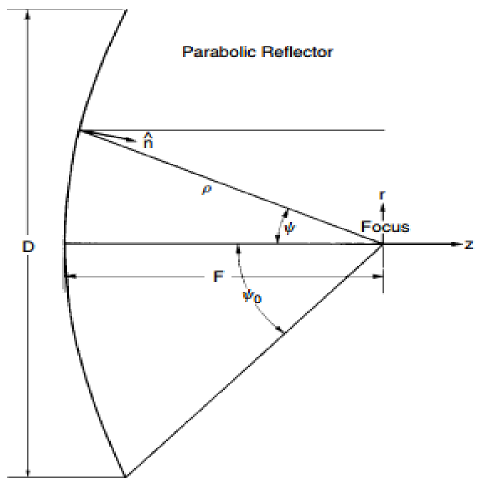


FIGURE 8-1 Geometry of a parabolic reflector.

We eliminate the dimensions of the reflector by using the ratio f /D. The half subtended angle of the reflector, ψ0, relates to f/D by

$$\psi_0 = 2\tan^{-1}\frac{1}{4f/D}$$

Scale 8-1 computes the total feed subtended angle from reflector f/D. When we place the aperture plane at the focus, the ray path distance becomes

$$\rho + \rho\cos\psi = 2\rho\cos^2\frac{\psi}{2} = 2f$$

The normal unit vector at a point on the reflector (r, z) is found from the feed angle:

$$\hat{\mathbf{n}} = -\sin\frac{\psi}{2}\hat{\mathbf{r}} + \cos\frac{\psi}{2}\hat{\mathbf{z}}$$

At this point we need the radius of curvatures in the principal planes to apply Eq. (2-77) reflection from a curved surface: R1 in the r –z plane and R2 in the φ–z plane:

$$R_1 = \frac{2f}{\cos^3(\psi/2)} \quad \text{and} \quad R_2 = \frac{2f}{\cos(\psi/2)}$$

The spherical wave spreads from the feed as 1/ρ. At the surface of the reflector the wave curvature changes to a plane wave and propagates to the aperture plane at a constant amplitude. The spherical wave spreading multiplies the feed distribution by [Eq. (8-1)] cos2(ψ/2) in the aperture. Then

$$\text{added edge taper} = \cos^2 \frac{\psi_0}{2} \qquad \text{voltage}$$

Deeper reflectors (smaller f/D) have greater edge tapers than shallow reflectors (larger f/D). Scale 8-2 provides a quick calculation of the added edge taper due to spherical wave spreading. The reflector beamwidth is given by

$$\text{HPBW} = 67.3^\circ \frac{\lambda}{D} \quad \text{and} \quad \text{HPBW} = 67^\circ \frac{\lambda}{D}$$

HPBW = 70◦ λ/D for a parabolic reflector. An integration of the aperture distribution for the far-field pattern gives the following results:

$$\text{HPBW} = 67.46^\circ \frac{\lambda}{D} \qquad \text{sidelobe level} = 27\,\text{dB}$$

FEED MISMATCH DUE TO THE REFLECTOR

The feed receives some of its transmitted power because it reflects from the parabola and returns as a mismatch at the feed terminals. We calculate the reflected field at the feed by using surface currents and the magnetic vector potential. The only significant contribution comes from areas near where the normal of the reflector points at the feed. Around every other point, the phase of the reflection varies rapidly and cancels and we need to consider only points of stationary phase. We calculate the reflection from each point of stationary phase

$$\Gamma = -j\frac{G_f(\rho_0)}{4k\rho_0}\sqrt{\frac{\rho_1\rho_2}{(\rho_1+\rho_0)(\rho_2+\rho_0)}}e^{-j2k\rho_0}$$

where is the reflection coefficient, ρ0 the distance to the stationary phase point, Gf (ρ0) the feed gain in the direction of ρ0, and ρ1 and ρ2 the radiuses of curvature of the reflector at ρ0. The vertex is the only point of stationary phase on a paraboloidal reflector: ρ1 = ρ2 = −2f and ρ0 = f . Equation above reduces to

$$\Gamma = -j\frac{G_f(0)}{2kf}e^{-j2kf}$$

We can express the reflector reflection of a paraboloidal reflector as

$$|\Gamma| = V\frac{\lambda}{D}$$

and calculate V versus f /D for feeds with 10-dB beamwidths equal to the reflector subtended angle. Higher reflector f /D values produce larger feed reflections, since the feed gain increases faster than the reduced area of the reflector seen from the feed.

FRONT-TO-BACK RATIO

Figure 2-9 illustrates the pattern response of a paraboloidal reflector and shows that the pattern behind the reflector peaks along the axis. The diffractions from all points along the rim add in-phase along the axis and produce a pattern peak. We can reduce this rim diffraction by using a rolled, serrated, or castellated edge to reduce diffraction. An absorber-lined cylindrical shroud extending out to enclose the feed will greatly reduce back radiation, including spillover, and allows the close spacing of terrestrial microwave antennas with reduced crosstalk.

For a normal truncated circular reflector rim, the following equation estimates the front-to-back ratio given the reflector gain G, the feed taper T , and feed gain Gf

$$\text{F/B} = G + T + K - G_f \qquad \text{dB}$$

The constant K, given by Scale 8-9, is related to f /D:

$$K = 10 \log \left[1 + \frac{1}{(4f/D)^2} \right]$$

OFFSET-FED REFLECTOR

Moving the feed out of the aperture eliminates some of the problems with axisymmetrical reflectors. Blockage losses and diffraction-caused sidelobes and cross-polarization disappear. We can increase the size of the feed structure and include more if not all of the receiver with the feed. For example, the reflector may be deployed from a satellite, with the feed mounted on the main satellite body. Figure 8-7 shows the offset-fed reflector geometry. We form the reflector out of a piece of a larger paraboloid. Every piece of the paraboloidal reflector converts spherical waves from the focus into a plane wave moving parallel with its axis. We point the feed toward the center of the reflector to reduce the spillover, but we still locate the feed phase center at the focus of the reflector. The aperture plane projects to a circle, although the rim shape is an ellipse. ψ0 is the angle from the axis of the parabola to

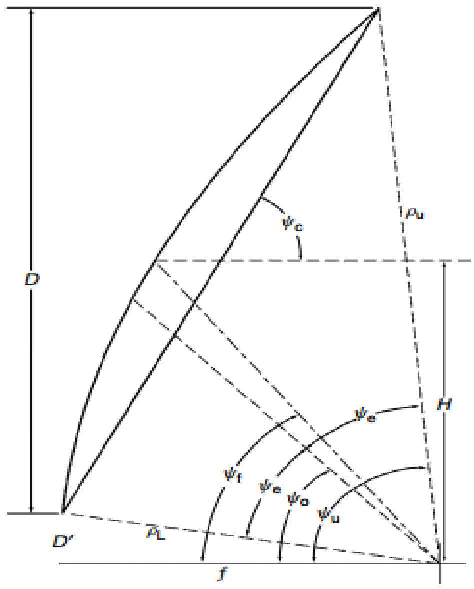


FIGURE 8-7 Parameters of an offset-fed parabolic reflector.

the center of the cone of the reflector, and the reflector subtends an angle 2ψe about this centerline. Given the aperture plane diameter D and the height H of the center, we find the lower rim offset D = H − D/2. From these parameters we determine the angle of the center of the rim cone from the z-axis:

$$\psi_0 = \tan^{-1}\frac{16fH}{16f^2 + D^2 - 4H^2} = \tan^{-1}\frac{2f(D+2D')}{4f^2 - D'(D+D')}$$

The half cone angle defines the rim:

$$\psi_e = \tan^{-1}\frac{8fD}{16f^2 + 4H^2 - D^2} = \tan^{-1}\frac{2fD}{4f^2 + D'(D+D')}$$

We direct the feed an angle ψf from the z-axis to the center of the projected diameter different from the angle ψ0 of the rim cone axis:

$$\psi_f = 2\ \tan^{-1}\frac{H}{2f} = 2\ \tan^{-1}\frac{2D' + D}{4f}$$

The rim lies in a plane at an angle ψc with respect to the z-axis:

$$\psi_c = \tan^{-1}\frac{2f}{H} = \tan^{-1}\frac{4f}{2D' + D}$$

The rim is an ellipse in this plane with major and minor axes given by

$$a_e = \frac{D}{2 \sin \psi_c} \quad \text{and} \quad b_e = \frac{D}{2}$$

The offset angle modifies the f /D of the reflector:

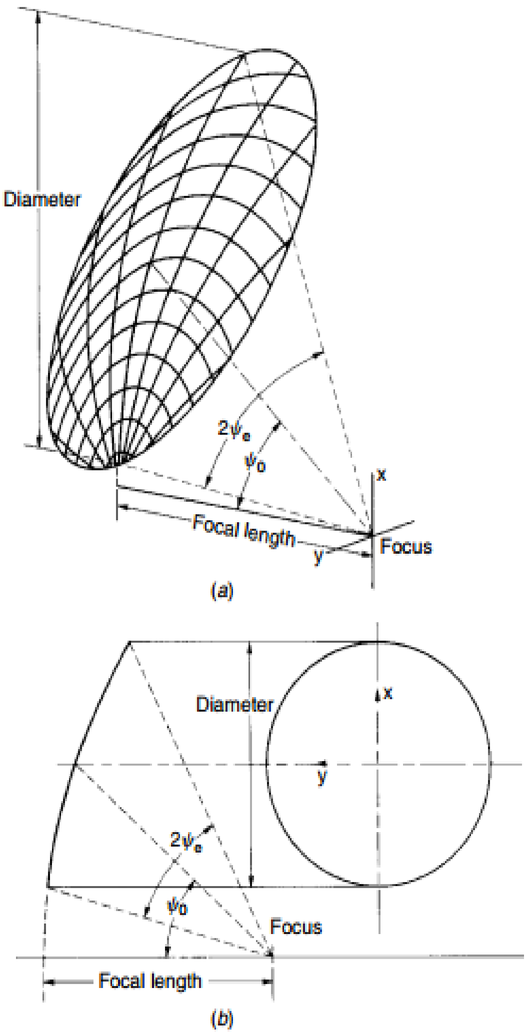
$$\frac{f}{D} = \frac{\cos \psi_e + \cos \psi_0}{4 \sin \psi_e}$$

We calculate the rim offset from the cone angles:

$$D' = 2f \tan \frac{\psi_0 - \psi_e}{2}$$

To align the reflector, we use the angle of the reflector rim major axis ψc = sin−1 (D/L) with respect to the z-axis and the radial distances from the lower and upper edges of the reflector in the offset plane, since the center offset H is not a distinguishable point:

We analyze the offset reflector with the same tools as those used with the axisymmetric reflector: aperture field, physical optics, and GTD. The asymmetry of the reflector to feed geometry introduces anomalies. Huygens sources no longer eliminate crosspolarization, because the source must be tilted. Symmetry prevents cross-polarization in the plane containing the x-axis (Figure 8-8), but cross-polarization for linear polarization increases in the plane containing the y-axis (symmetry plane) as f /D decreases



Offset-fed paraboloidal reflector geometry: (a) perspective; (b) orthographic

CASSEGRAIN FEED SYSTEM:

i) Feed Blockage

The increased effective focal length requires feeds with narrow beamwidths, and we can no longer consider the feed as a point source. It projects a shadow into the center of the reflector (Figure 8-15) and causes a central blockage. The subreflector also blocks the center. As we reduce the subreflector diameter to reduce blockage, the feed antenna moves closer to the subreflector and its projected shadow increases. The optimum occurs when the projected feed blockage diameter equals the subreflector diameter.

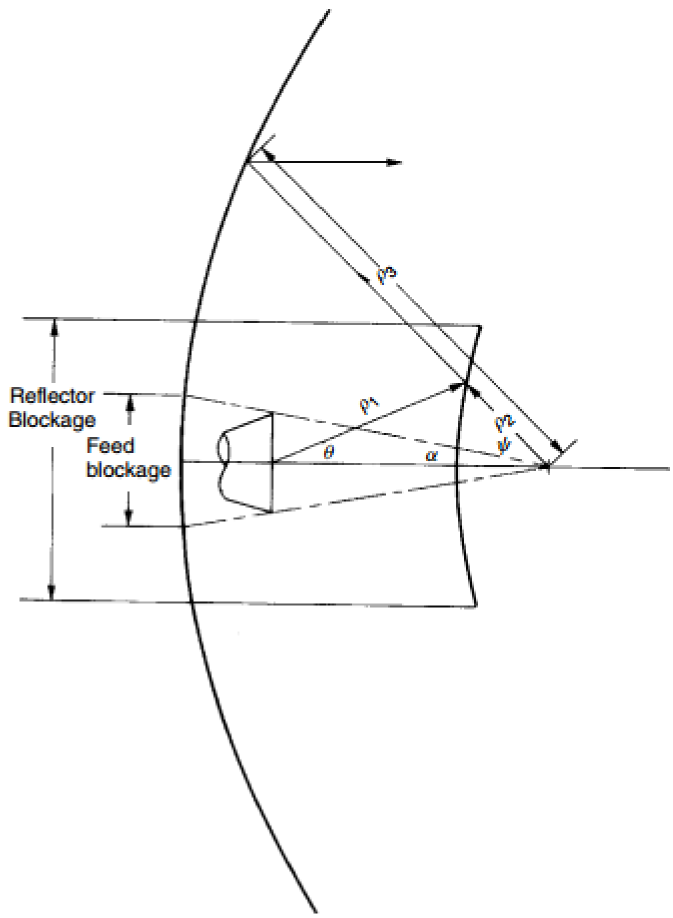


FIGURE 8-15 Cassegrain central blockage.

The feed size depends on the frequency of operation and the effective f/D value, whereas the subreflector diameter depends only on geometry. We cannot determine the optimum independent of frequency.

SPHERICAL REFLECTOR

When we feed-scan a paraboloidal reflector, the pattern sidelobes develop coma and the beam shape generally degrades. Feed scanning is limited. In a spherical reflector a feed moved in an arc from the center of the sphere and sees the same reflector geometry if we discount the edge effects. Greater scanning is possible, but the spherical reflector fails to focus an incident plane wave to a point and requires more elaborate feeds. We can design many types of feeds for the spherical reflector.

The reflector can be fed from a point source for large f /D by assuming that it is a distorted parabola [55,56]. It can be fed with a line source to follow the axis fields. Corrector subreflectors can be designed to correct the spherical aberrations [58]. Like the parabolic reflector, we can design arrays [24] to compensate for spherical aberrations and give multiple beams. Figure 8-23 shows the geometry and ray tracing of a spherical reflector illuminated by a plane wave. All rays intersect a radial line of the sphere (the axis) in the direction of the incident wave because the reflector has circular symmetry about all axes. The diagram traces rays hitting the outer portion of the reflector as passing through the axis closer to the vertex than do the rays reflected from areas closer to the axis. The reflector has a line focus. A distorted paraboloidal reflector with a line focus exhibits spherical aberration because the focal length depends on the radial distance from the axis of the reflection point. The spherical reflector has a cusplike caustic where GO

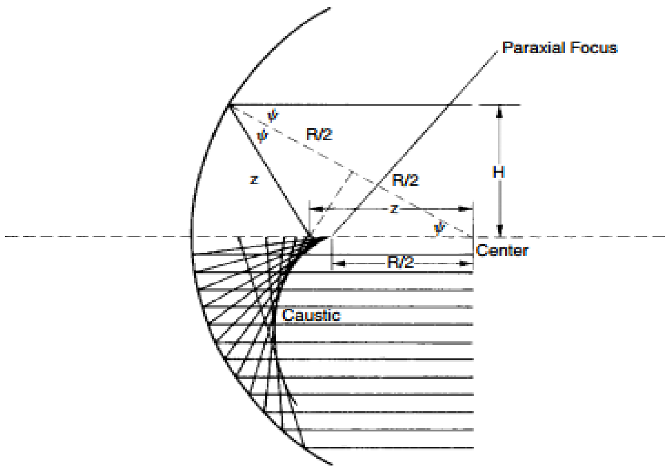


FIGURE 8-23 Ray tracing in a spherical reflector.

predicts infinite fields. The second side of Figure 8-23 traces a single ray. We can easily solve the isosceles triangle for the results:

$$z = \frac{R/2}{\sqrt{1 - H^2/R^2}}$$

$$H^2 = R^2\left(1 - \frac{R^2}{4z^2}\right)$$

where z is the location of the focus for a given ray.

As H approaches zero, with rays near the axis, the reflected ray passes through the paraxial focus (z = R/2). We use Eq. (8-89) to find the power distribution on the axis by using the conservation of power. The power in a differential area of the plane wave reflects into a differential length on the axis: dA = 2πH dH. We differentiate Eq. (8-89) implicitly:

$$2H\,dH = \frac{R^2}{2z^3}\,dz$$

The power distribution along the axis is

$$P_z = \frac{P_0 R^3}{8z^3}$$

where P0 is the power at the paraxial focus. The peak power occurs at the paraxial focus and drops by one-eighth (−9 dB) at the vertex. We determine the required length of the line source feed from the rotation angle ψ of the illuminated portion of the reflector:

$$\text{feed length} = \frac{R(1/\cos\psi - 1)}{2}$$

4. Explain the construction and working principle of microstip antenna. (8M). [CO2-L2- Nov/Dec 2012, April/May 2015]

Rectangular patch antennas can be designed by using a transmission-line model [9] suitable for moderate bandwidth antennas. Patches with bandwidths of less than 1% or greater than 4% require a cavity analysis for accurate results, but the transmissionline model covers most designs. The lowest-order mode, TM10, resonates when the effective length across the patch is a half-wavelength. Figure 6-9 demonstrates the patch fed below from a coax along the resonant length. Radiation occurs from the fringing fields. These fields extend the effective open circuit (magnetic wall) beyond the edge. The extension is given by

$$\frac{\Delta}{H} = 0.412\frac{\varepsilon_{\text{eff}} + 0.300}{\varepsilon_{\text{eff}} - 0.258}\frac{W/H + 0.262}{W/H + 0.813}$$

where H is the substrate thickness, W the patch nonresonant width, and εeff the effective dielectric constant of a microstrip transmission line the same width as the patch.

A suitable approximation for εeff is given by

$$\varepsilon_{\text{eff}} = \frac{\varepsilon_r + 1}{2} + \frac{\varepsilon_r - 1}{2}\left(1 + \frac{10H}{W}\right)^{-1/2}$$

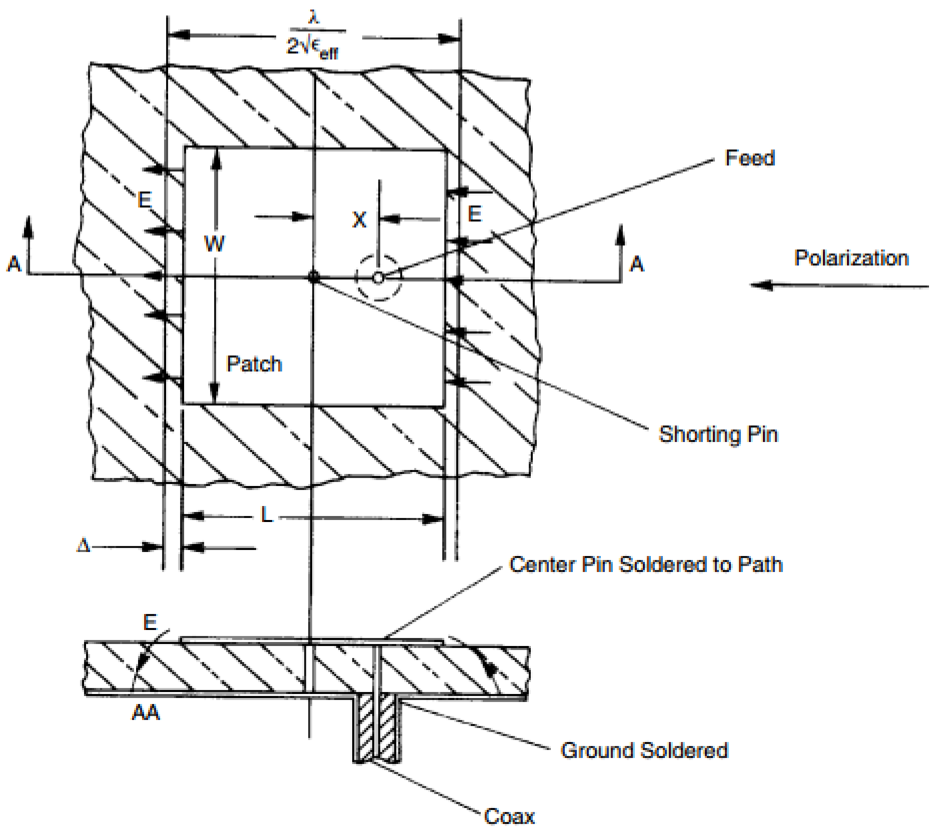


FIGURE 6-9 Coax-fed microstrip patch antenna.

where εr is the substrate dielectric constant. The transmission-line model represents the patch as a low-impedance microstrip line whose width determines the impedance and effective dielectric constant. A combination of parallel-plate radiation conductance and capacitive susceptance loads both radiating edges of the patch. Harrington [6, p. 183] gives the radiation conductance for a parallel-plate radiator as

$$G = \frac{\pi W}{\eta \lambda_0}\left[1 - \frac{(kH)^2}{24}\right]$$

where λ0 is the free-space wavelength. The capacitive susceptance relates to the effective strip extension:

$$B = 0.01668 \frac{\Delta}{H} \frac{W}{\lambda} \varepsilon_{\text{eff}}$$

The impedance varies from zero in the center to the edge resistance approximately as

$$R_i = R_e \sin^2 \frac{\pi x}{L} \qquad 0 \le x \le \frac{L}{2}$$

where Ri is the input resistance, Re the input resistance at the edge, and x the distance from the patch center. The feed location does not significantly affect the resonant frequency. By using Eq. (6-22), we locate the feed point given the desired input impedance:

$$x = \frac{L}{\pi} \sin^{-1} \sqrt{\frac{R_i}{R_e}}$$

The feed pin currents add to the pattern by radiating a monopole pattern. Figure 6-10 shows this radiation for a patch using a free-space substrate where the E-plane radiating edges are spaced λ/2. The pattern of Figure 6-10 has a null along the ground plane in the E-plane, but the monopole radiation increases the radiation along the ground plane. On one side the radiation adds and on the other it subtracts from the E-plane pattern to form a null tilted above the ground plane. The H-plane pattern now contains crosspolarization. We can reduce the monopole radiation by feeding the patch at a second port located an equal distance from the center on the opposite side. This requires an external feed network that divides the power equally between the two ports with a 180◦ phase difference. The problem with this feed arrangement is that significant power is coupled between the two feeds in the equivalent microwave circuit of the patch. The estimated value of −6 dB coupling between the ports causes a portion of the input power to be dissipated in the second port. At this level the patch efficiency drops 1.25 dB. We can reduce the monopole radiation by coupling to a second short-circuited probe to the patch instead of directly feeding it. The gap between the second probe and the patch is adjusted until the antenna radiates minimum cross-polarization in the H-plane.

This uses the microstrip patch as the feed network, and the second probe has no resistive load to dissipate power.

The feed probe across the microstrip patch substrate is a series inductor at the input. Higher-order modes excited in the patch by this feeding method add to the inductive component of the antenna.

Below resonance, the antenna is inductive and has nearzero resistance. As the frequency increases, the inductance and resistance grow as the parallel resonance is approached. Above the resonant frequency, the antenna is capacitive as the impedance sweeps clockwise around the Smith chart (Figure 6-11) and finally back to a slight inductive component near a short circuit.

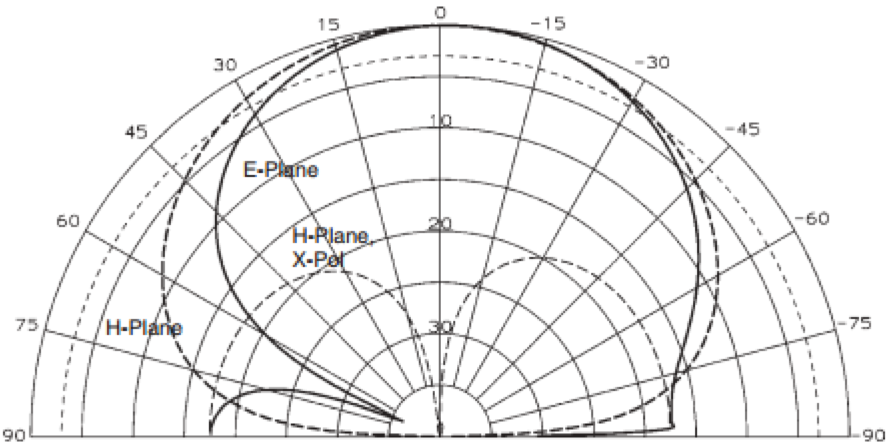


FIGURE 6-10 Pattern of coax-fed, microstrip patch including feed pin radiation for free-space substrate.

Unit-3

Antenna Arrays

Part-A

1. What is meant by uniform linear array.? [CO3-L1- Nov/Dec 2011]

An array is linear when the elements of the array are spaced equally along the straight line. If the elements are fed with currents of equal magnitude and having a uniform progressive phase shift along the line, then it is called uniform linear array .

2. What are the types of array? [CO3-L1]

a. Broad side array.
b. End fire array
c. Collinear array.
d. Parasitic array.

3. What is Broad side array? [CO3-L1- April/May 2011]

Broad side array is defined as an arrangement in which the principal direction of radiation is perpendicular to the array axis and also the plane containing the array element.

4. What is point source? [CO3-L1]

It is the waves originate at a fictitious volumeless emitter source at the center O of the observation circle.

5. What is meant by array? [CO3-L1]

An antenna is a system of similar antennas oriented similarly to get greater directivity in a desired direction

6. What is collinear array? [CO3-L1]

In this array the antenna elements are arranged coaxially by mounting the elements end to end in straight line or stacking them one over the other with radiation pattern circular symmetry. Eg. Omni directional antenna.

7. What is the need for the Binomial array? [CO3-L1- April/May 2015]

The need for a binomial array is

i). In uniform linear array as the array length is increased to increase the directivity, the secondary lobes also occurs.

ii) For certain applications, it is highly desirable that secondary lobes should be eliminated completely or reduced to minimum desirable level compared to main lobes.

8. Define power pattern. [CO3-L1]

Graphical representation of the radial component of the pointing vector Sr at a constant radius as a function of angle is called power density pattern or power pattern.

9. What is meant by similar Point sources.? [CO3-L1- April/May 2014]

Whenever the variation of the amplitude and the phase of the field with respect to the absolute angle for any two sources are same then they are called similar point sources. The maximum amplitudes of the individual sources may be unequal.

10. What is meant by identical Point sources? [CO3-L1]

Similar point sources with equal maximum amplitudes are called identical point sources.

11. State the principle of the pattern multiplication [CO3-L2- May/June 2012]

The total field pattern of an array of non isotropic but similar sources is the product of the

i) individual source pattern and

ii) The array pattern of isotropic point sources each located at the phase center of the individual source having the same amplitude and phase.

12. What is the advantage of pattern multiplication? [CO3-L1- April/May 2013]

i)Useful tool in designing antenna

ii) It approximates the pattern of a complicated array without making lengthy computations

13. What is tapering of arrays? [CO3-L1]

Tapering of array is a technique used for reduction of unwanted side lobes .The amplitude of currents in the linear array source is non-uniform; hence the central source radiates more energy than the ends. Tapering is done from center to end.

14. What is a binomial array? [CO3-L1]

It is an array in which the amplitudes of the antenna elements in the array are arranged according to the coefficients of the binomial series.

15. What are the advantages of binomial array? [CO3-L1]

Advantage:

No minor lobes

Disadvantages:

Increased beam width

Maintaining the large ratio of current amplitude in large arrays is difficult

16. What is the difference between isotropic and non- isotropic source. [CO3-L2- April/May 2015]

Isotropic source radiates energy in all directions but non-isotropic source radiates energy only in some desired directions.

Isotropic source is not physically realizable but non-isotropic source is physically realizable.

17. List the arrays used for array tapering. **[CO3-L2]**

Binomial Array:Tapering follows the coefficient of binomial series

Dolph Tchebycheff Array: Tapering follows the coefficient of Tchebycheff polynomial.

18. Define endfire array.

It is defined as an arrangement in which the principle direction of radiation coincides with the array axis and there exist a phase difference between adjacent elements.

[CO3-L1-May/June 016]

19. Define adaptive array.

The adaptive array can automatically steer its beam toward a desired signal while steering a null toward an undesired or interfering signal. Such an array may be called as smart antenna.

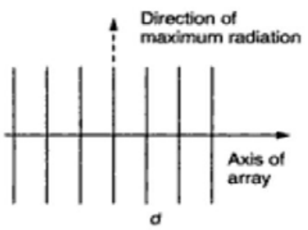
[CO3-L1-May/June2016]

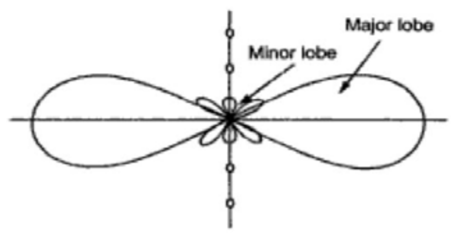
Part-B

1. Derive the expression for pattern maxima, minima and half power beam width for broadside array. (8M) [CO3-H1- Nov/Dec 2011, April/May 2012]

This is a type of array in which the number of identical elements is placed on a supporting line drawn perpendicular to their respective axes.

Elements are equally spaced and fed with a current of equal magnitude and all in same phase. The advantage of this feed technique is that array fires in broad side direction (i.e. perpendicular to the line of array axis, where there are maximum radiation and small radiation in other direction). Hence the radiation pattern of broadside array is bidirectional and the array radiates equally well in either direction of maximum radiation. In Fig. 1 the elements are arranged in horizontal plane with spacing between elements and radiation is perpendicular to the plane of array (i.e. normal to plane of paper.) They may also be arranged in vertical and in this case radiation will be horizontal. Thus, it can be said that broadside array is a geometrical arrangement of elements in which the direction of maximum radiation is perpendicular to the array axis and to the plane containing the array clement. Radiation pattern of a broad side array is shown in Fig. 2. The bidirectional pattern of broadside array can be converted into unidirectional by placing an identical array behind this array at distance of λ/4 fed by current leading in phase by 90^0.





Maxima direction

From equation (9), the total field is maximum when $\cos\left(\frac{\beta d \cos\phi}{2}\right)$ is maximum. As we know, the variation of cosine of a angle is ± 1. Hence the condition for maxima is given by,

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

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

If n = 0, then

$$\frac{\pi}{2}\cos\phi_{max} = 0$$

$$\text{i.e.}\quad \cos\phi_{max} = 0$$

$$\text{i.e.}\quad \boxed{\phi_{max} = 90° \text{ or } 270°} \qquad ...(11)$$

Minima direction

Again from equation (9), total field strength is minimum when $\cos\left(\frac{\beta d\cos\phi}{2}\right)$ is minimum i.e. 0 as cosine of angle has minimum value 0. Hence the condition for minima is given by,

$$\therefore\quad \boxed{\cos\left(\frac{\beta d\cos\phi}{2}\right) = 0} \qquad ...(12)$$

Again assuming d = λ/2 and β=2π/λ, we can write

$$\cos\left(\frac{\pi}{2}\cos\phi_{min}\right) = 0$$

$$\therefore\quad \frac{\pi}{2}\cos\phi_{min} = \cos^{-1} 0 = \pm(2n+1)\frac{\pi}{2}, \text{ where } n = 0, 1, 2, ...$$

If n = 0, then,

$$\frac{\pi}{2}\cos\phi_{min} = \pm\frac{\pi}{2}$$

$$\text{i.e. } \cos\phi_{min} = \pm 1$$

$$\text{i.e.}\quad \boxed{\phi_{min} = 0° \text{ or } 180°} \qquad ...(13)$$

Half power point direction:

When the power is half, the voltage or current is 1/√2 times the maximum value.

Hence the condition for half power point is given by,

Let d=λ/2 and β=2π/λ, then we can write,

$$\cos\left(\frac{\pi}{2}\cos\phi\right) = \pm\frac{1}{\sqrt{2}}$$

$$\text{i.e. } \frac{\pi}{2}\cos\phi = \cos^{-1}\left(\pm\frac{1}{\sqrt{2}}\right) = \pm(2n+1)\frac{\pi}{4}, \text{ where } n = 0, 1, 2,....$$

If n = 0, then

$$\frac{\pi}{2}\cos\phi_{HPPD} = \pm\frac{\pi}{4}$$

$$\text{i.e. } \cos\phi_{HPPD} = \pm\frac{1}{2}$$

$$\text{i.e. } \phi_{HPPD} = \cos^{-1}\left(\pm\frac{1}{2}\right)$$

$$\therefore \quad \boxed{\phi_{HPPD} = 60° \text{ or } 120°} \quad ...(15)$$

The field pattern drawn with E_T against υ for d=λ/2, then the pattern is bidirectional as shown in Fig 6. The field pattern obtained is bidirectional and it is a figure of eight.

If this pattern is rotated by 360^0 about axis, it will represent three dimensional doughnut shaped space pattern. This is the simplest type of broadside array of two point sources and it is called Broadside couplet as two radiations of point sources are in phase.

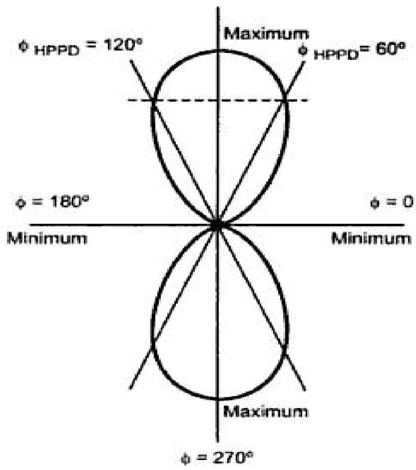


Fig. 6 Field pattern for two point source with spacing d=λ/2 and fed with currents equal in magnitude and phase.

2. Derive the expression for pattern maxima, minima and half power beam width for endfire array. (8M) [CO3-H1- April/May 2012]

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.

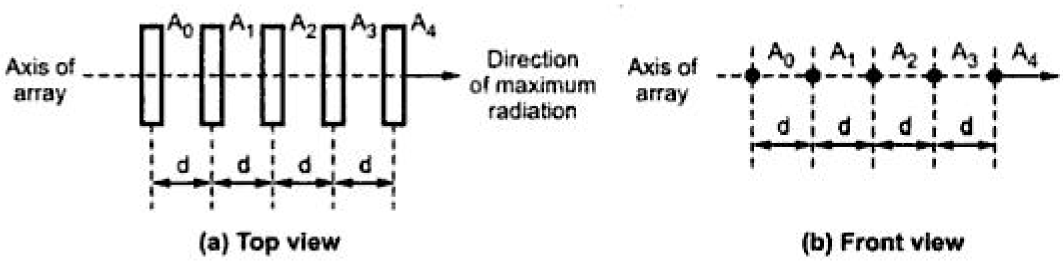


Fig. 3 End fire array

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

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

Consider two point sources separated by distance d and supplied with currents equal in magnitude but opposite in phase. Consider Fig. 5 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,

$$E_T = (-E_1) + (E_2) \quad \ldots(1)$$

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

$$E_1 = E_0 e^{-j\frac{\psi}{2}} \quad \ldots(2)$$

and $$E_2 = E_0 e^{j\frac{\psi}{2}} \quad \ldots(3)$$

Substituting values of E_1 and E_2 in equation (1), we get

$$E_T = -E_0 \cdot e^{-j\frac{\psi}{2}} + E_0 \cdot e^{j\frac{\psi}{2}}$$

$$\therefore \quad E_T = E_0\left(-e^{-j\frac{\psi}{2}} + e^{j\frac{\psi}{2}}\right)$$

Rearranging the terms in above equation, we get,

$$\therefore \quad E_T = (j2)\ E_0\left(\frac{e^{j\frac{\psi}{2}} - e^{-j\frac{\psi}{2}}}{j2}\right) \qquad \ldots(4)$$

By trigonometry identity,

$$\frac{e^{j\theta} - e^{-j\theta}}{2} = \sin\frac{\theta}{2}.$$

Equation (4) can be written as,

$$E_T = j2 E_0 \sin\left(\frac{\psi}{2}\right) \qquad \ldots(5)$$

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

Phase angle = βdcosυ ...(6) Substituting value of phase angle in equation (5), we get,

$$\boxed{E_T = j(2E_0)\sin\left(\frac{\beta d\cos\phi}{2}\right)} \qquad \ldots(7)$$

Maxima direction

From equation (7), the total field is maximum when $\sin\left(\frac{\beta d\cos\phi}{2}\right)$ is maximum i.e. ±1 as the maximum value of sine of angle is ±1. Hence condition for maxima is given by,

$$\boxed{\sin\left(\frac{\beta d\cos\phi}{2}\right) = \pm 1} \qquad \ldots(8)$$

Let the spacing between two isotropic point sources be equal to d=λ/2

Substituting d=λ/2 and β=2π/λ, in equation (8), we get,

$$\sin\left(\frac{\pi}{2}\cos\phi\right) = \pm 1$$

$$\text{i.e.} \quad \frac{\pi}{2}\cos\phi = \pm(2n+1)\frac{\pi}{2}, \text{ where } n = 0, 1, 2, \ldots\ldots$$

Minima direction

Again from equation (7), total field strength is minimum when $\sin\left(\frac{\beta d\cos\phi}{2}\right)$ is minimum i.e. 0.
Hence the condition for minima is given by,

$$\boxed{\sin\left(\frac{\beta d\cos\phi}{2}\right) = 0} \quad \ldots(10)$$

Assuming d=λ/2 and β=2π/λ in equation (10), we get,

$$\sin\left(\frac{\pi}{2}\cos\phi\right) = 0$$

$$\text{i.e.} \quad \frac{\pi}{2}\cos\phi = \pm n\pi, \text{ where } n = 0, 1, 2, \ldots\ldots$$

If n = 0, then

Half Power Point Direction (HPPD)

When the power is half of maximum value, the voltage or current equals to 1/√2 times the respective maximum value. Hence the condition for the half power point can be obtained from equation (7) as,

$$\boxed{\sin\left(\frac{\beta d\cos\phi}{2}\right) = \pm\frac{1}{\sqrt{2}}} \quad \ldots(12)$$

Let d=λ/2 and β=2π/λ, we can write,

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

$$\text{i.e. } \frac{\pi}{2}\cos\phi = \pm(2n+1)\frac{\pi}{4}, \quad \text{where } n = 0, 1, 2.$$

If n = 0, we can write,

$$\frac{\pi}{2}\cos\phi_{HPPD} = \pm\frac{\pi}{4}$$

$$\text{i.e. } \cos\phi_{HPPD} = \pm\frac{1}{2}$$

$$\therefore \quad \boxed{\phi_{HPPD} = 60° \text{ or } 120°} \quad \ldots(13)$$

Thus from the conditions of maxima, minima and half power points, the field pattern can be drawn as shown in the Fig. 7.

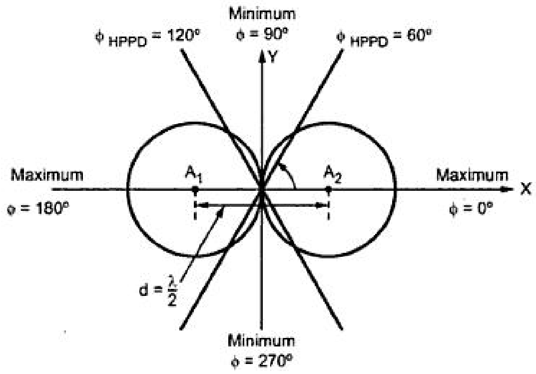


Fig. 7 Field pattern for two point sources with spacing d = d=λ/2 and fed with currents equal in magnitude but out of phase by 180^0.

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

3. Derive the expression for array factor of N-element linear array (8M)

[CO3-H1- Nov/Dec 2011, April/May 2014, May/June 2016]

At higher frequencies, for point to point communications it is necessary to have a pattern with single beam radiation. Such highly directive single beam pattern can be obtained by increasing the point sources in the arrow from 2 to n say. An array of n elements is said to be linear array if all the individual elements are spaced equally along a line. An array is said to be uniform array if the elements in the array are fed with currents with equal magnitudes and with uniform progressive phase shift along the line. Consider a general n element linear and uniform array with all the individual elements spaced equally at distance d from each other and all elements are fed with currents equal in magnitude and uniform progressive phase shift along line as shown in the Fig. 9.

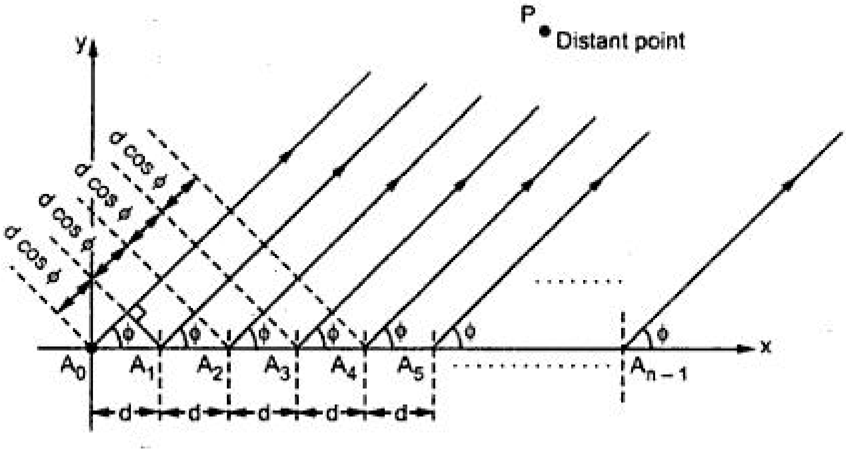


Fig. Uniform, linear array of n elements

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

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

$$\therefore \quad \boxed{E_T = E_0 \left[1 + e^{j\psi} + e^{j2\psi} + \ldots + e^{j(n-1)\psi}\right]} \quad \ldots (1)$$

Note that $\upsilon = (\beta d \cos\upsilon + \alpha)$ indicates the total phase difference of the fields from adjacent sources calculated at point P. Similarly α is the progressive phase shift between two adjacent point sources.

The value of α may lie between 0^0 and 180^0. If $\alpha = 0^0$ we get n element uniform linear broadside array. If $\alpha = 180^0$ we get n element uniform linear endfire array.

Multiplying equation (1) by $e^{j\upsilon}$, we get,

$$E_T e^{j\psi} = E_0 [e^{j\psi} + e^{j2\psi} + e^{j3\psi} + \dots + e^{jn\psi}] \qquad \dots(2)$$

Subtracting equation (2) from (1), we get,

$$E_T - E_T e^{j\psi} = E_0 \left\{[1 + e^{j\psi} + e^{j2\psi} + \dots + e^{j(n-1)\psi}] - [e^{j\psi} + e^{j2\psi} + \dots + e^{jn\psi}]\right\}$$

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

$$\therefore \quad \boxed{E_T = E_0 \left[\frac{1 - e^{jn\psi}}{1 - e^{j\psi}}\right]} \qquad \dots(3)$$

Simply mathematically, we get

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

According to trigonometric identity,

$$e^{-j\theta} - e^{j\theta} = -2j \sin\theta,$$

The resultant field is given by,

$$E_T = E_0 \left[\frac{\left(-j2\sin\frac{n\psi}{2}\right)e^{j\frac{n\psi}{2}}}{\left(-j2\sin\frac{\psi}{2}\right)e^{j\frac{\psi}{2}}}\right]$$

$$\therefore \quad \boxed{E_T = E_0 \left[\frac{\sin n\frac{\psi}{2}}{\sin\frac{\psi}{2}}\right] e^{j\left(\frac{n-1}{2}\right)\psi}} \qquad \dots(4)$$

This equation (4) indicates the resultant field due to n element array at distant point P. The magnitude of the resultant field is given by,

$$\therefore \qquad E_T = E_0 \left[\frac{\sin n \frac{\psi}{2}}{\sin \frac{\psi}{2}} \right] \qquad \ldots (5)$$

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

$$\therefore \qquad \theta = \frac{(n-1)}{2} \psi = \frac{(n-1)}{2} \beta d \cos \phi + \alpha \qquad \ldots (6)$$

4. Explain the working principle of BSA along with its properties. (16M)

[CO3-L2- April/May 2014, April/May 2015,May/June 2016]

Consider 'n' number of identical radiators carries currents which are equal in magnitude and in phase. The identical radiators are equispaced. Hence the maximum radiation occurs in the directions normal to the line of array. Hence such an array is known as Uniform broadside array. Consider a broadside array with n identical radiators as shown in the Fig. 10.

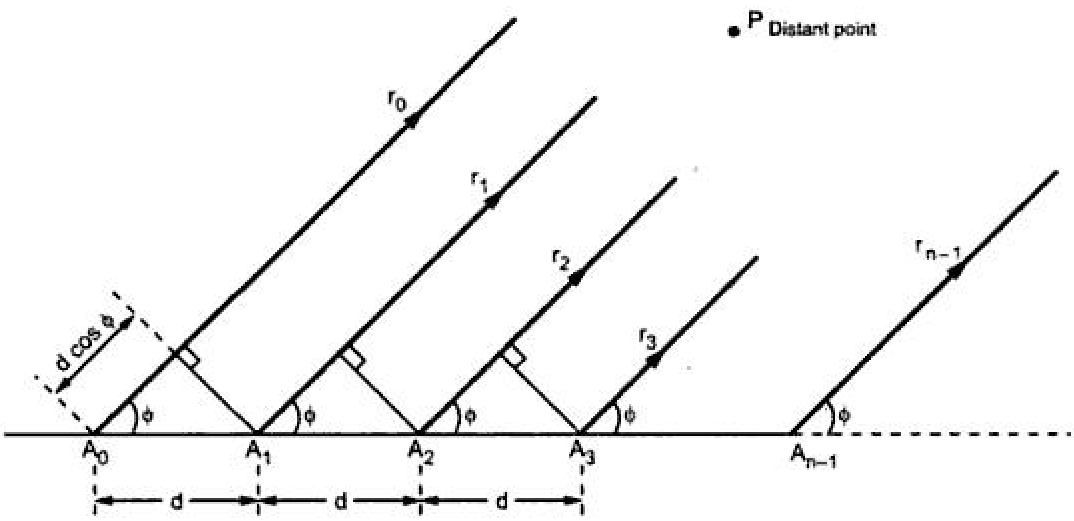


Fig Array of n elements with Equal Spacing

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

$$E_0 = \frac{I\, dL \sin\theta}{4\pi\omega\varepsilon_0}\left[j\frac{\beta^2}{r_0}\right] e^{-j\beta r_0} \qquad \text{... (1)}$$

As the distance of separation d between any two array elements is very small as compared to the radial distances of point P from A_0, A_1, ...A_{n-1}, we can assume r_0, r_1, ...r_{n-1} are approximately same.

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

$$\text{But} \qquad r_1 = r_0 - d\cos\phi$$

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

Exactly on the similar lines we can write the electric field produced at point P due to an element A_2 as,

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

$$\therefore \qquad E_2 = \left\{\frac{I\, dL \sin\theta}{4\pi\omega\varepsilon_0}\left[j\frac{\beta^2}{r_1}\right] e^{-j\beta r_1}\right\} e^{j\beta d\cos\phi}$$

$$\therefore \qquad E_2 = \frac{I\, dL \sin\theta}{4\pi\omega\varepsilon_0}\left[j\frac{\beta^2}{r_1}\right] e^{-j\beta(r_1 - d\cos\phi)} \qquad \text{... } r_2 = r_1 - d\cos\phi$$

But the term inside the bracket represent E_1

$$\therefore \qquad E_2 = E_1\, e^{j\beta d\cos\phi}$$

From equation (2), substituting the value of E_1, we get,

$$E_2 = \left[E_0\, e^{j\beta d\cos\phi}\right] e^{j\beta d\cos\phi}$$

$$\therefore \qquad E_2 = E_0 \cdot e^{j2\beta d\cos\phi} \qquad \text{... (3)}$$

Similarly, the electric field produced at point P due to element A_{n-1} is given by,

$$E_{n-1} = E_0 \cdot e^{j(n-1)\beta d\cos\phi} \quad \text{... (4)}$$

The total electric field at point P is given by,

$$E_T = E_0 + E_1 + E_2 + \dots + E_{n-1}$$

$$\therefore \quad E_T = E_0 + E_0\, e^{j\beta d\cos\phi} + E_0\, e^{j2\beta d\cos\phi} + \dots + E_0\, e^{j(n-1)\beta d\cos\phi}$$

Consider a series given by $s = 1 + r + r^2 + \dots + r^{n-1}$ where $r = e^{j\upsilon}$

Multiplying both the sides of the equation (i) by r, $s \cdot r = r + r^2 + \dots + r^n$

Subtracting equation (ii) from (i), we get. $s(1-r) = 1-r^n$

$$\therefore \quad s = \frac{1-r^n}{1-r} \quad \text{... (iii)}$$

Using equation (iii), equation (5) can be modified as,

From the trigonometric identities,

$$e^{-j\theta} = \cos\theta - j\sin\theta$$

$$e^{j\theta} = \cos\theta + j\sin\theta$$

$$\text{and} \quad e^{-j\theta} - e^{j\theta} = -j\,2\sin\theta$$

Equation (6) can be written as,

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

$$\therefore \quad \frac{E_T}{E_0} = e^{j\frac{(n-1)\psi}{2}}\left[\frac{\sin\left(\frac{n\psi}{2}\right)}{\sin\left(\frac{\psi}{2}\right)}\right] \quad \text{... (7)}$$

The exponential term in equation (7) represents the phase shift. Now considering magnitudes of the electric fields, we can write,

$$\left|\frac{E_T}{E_0}\right| = \frac{\sin\frac{n\psi}{2}}{\sin\frac{\psi}{2}} \qquad \text{... (8)}$$

Properties of Broadside Array

1. **Major lobe**

   In case of broadside array, the field is maximum in the direction normal to the axis of the array. Thus the condition for the maximum field at point P is given by,

$$\psi = 0 \quad \text{i.e.} \quad \beta d \cos\phi = 0 \qquad \text{...(9)}$$

$$\text{i.e.} \quad \cos\phi = 0$$

$$\text{i.e.} \quad \boxed{\phi = 90^\circ \text{ or } 270^\circ} \qquad \text{...(10)}$$

   Thus υ = 90^0 and 270^0 are called directions of principle maxima.

2. **Magnitude of major lobe**

   The maximum radiation occurs when υ=0. Hence we can write,

$$|\text{Major lobe}| = \left|\frac{E_T}{E_0}\right| = \lim_{\psi \to 0}\left\{\frac{\frac{d}{d\psi}\left(\sin n\frac{\psi}{2}\right)}{\frac{d}{d\psi}\left(\sin\frac{\psi}{2}\right)}\right\}$$

$$= \lim_{\psi \to 0}\left\{\frac{\left(\cos n\frac{\psi}{2}\right)\left(n\frac{\psi}{2}\right)}{\left(\cos\frac{\psi}{2}\right)\left(\frac{\psi}{2}\right)}\right\}$$

$$\therefore \quad \boxed{|\text{Major lobe}| = n} \qquad \text{...(11)}$$

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

   Thus from equation (10) and (11) it is clear that, all the field components add up together to give total field which is 'n' times the individual field when υ = 90^0 and 270^0.

3. Nulls

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

Equating ratio of magnitudes of the fields to zero,

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

The condition of minima is given by,

$$\therefore \quad \left|\frac{E_T}{E_0}\right| = \frac{\sin n\frac{\psi}{2}}{\sin\frac{\psi}{2}} = 0$$

$$\therefore \quad \boxed{\sin n\frac{\psi}{2} = 0; \quad \text{but } \sin\frac{\psi}{2} \neq 0} \quad \ldots(12)$$

Hence we can write,

$$\sin n\frac{\psi}{2} = 0$$

$$\text{i.e.} \quad n\frac{\psi}{2} = \sin^{-1}(0) = \pm m\pi, \text{ where } m = 1, 2, 3, \ldots\ldots$$

$$\text{Now } \psi = \beta d\cos\phi = \frac{2\pi}{\lambda}(d)\cos\phi$$

where, n= number of elements in the array d= spacing between elements in meter

λ= wavelength in meter

m= constant= 1, 2 , 3....

Thus equation (13) gives direction of nulls

4. Side Lobes Maxima

The directions of the subsidary maxima or side lobes maxima can be obtained if in equation (8), Hence sin(nʋ/2), is not considered. Because if nʋ/2=π/2 then sin nʋ/2 =1 which is the direction of principle maxima.

Hence we can skip sin nʋ/2 = ±π/2 value Thus, we get

$$\psi = \pm\frac{3\pi}{n}, \pm\frac{5\pi}{n}, \pm\frac{7\pi}{n}, \ldots\ldots\ldots\ldots$$

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

Now equation for ʋ can be written as,

$$\frac{2\pi}{\lambda}d\cos\phi = \pm\frac{3\pi}{n}, \pm\frac{5\pi}{n}, \pm\frac{7\pi}{n}, \ldots\ldots\ldots\ldots$$

$$\therefore \quad \cos\phi = \frac{\lambda}{2\pi d}\left[\pm\frac{(2m+1)}{n}\pi\right] \quad \text{where } m = 1, 2, 3, \ldots\ldots$$

$$\therefore \quad \boxed{\phi = \cos^{-1}\left[\pm\frac{\lambda(2m+1)}{2nd}\right]} \qquad \ldots(15)$$

The equation (15) represents directions of subsidary maxima or side lobes

maxima.

5. Beamwidth of Major Lobe

Beamwidth is defined as the angle between first nulls. Alternatively beamwidth is the angle equal to twice the angle between first null and the major lobe maximum direction. Hence beamwidth between first nulls is given by,

$$\therefore \quad \boxed{BWFN = 2\times\gamma, \ \text{where } \gamma = 90-\phi} \qquad \ldots(16)$$

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

Also $90-\phi_{min} = \gamma$ i.e. $90-\gamma = \theta_{min}$

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

Taking cosine of angle on both sides, we get

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

$$\therefore \quad \sin\gamma = \pm\frac{m\lambda}{n\,d} \quad \ldots(17)$$

If γ is very small, then sin γ ≈ γ. Substituting n above equation we get,

$$\gamma = \pm\frac{m\lambda}{n\,d} \quad \ldots(18)$$

For first null i.e. m=1,

$$\gamma = +\frac{\lambda}{n\,d}$$

$$\therefore \quad BWFN = 2\gamma = \frac{2\lambda}{n\,d}$$

But nd≈ (n-1)d if n is very large. This L= (nd) indicates total length of the array.

$$\therefore \quad \boxed{BWFN = \frac{2\lambda}{L}\ \text{rad} = \frac{2}{\left(\frac{L}{\lambda}\right)}\ \text{rad}} \quad \ldots(19)$$

BWFN in degree is written as,

$$\boxed{BWFN = \frac{114.6\lambda}{L} = \frac{114.6}{\left(\frac{L}{\lambda}\right)}\ \text{degrees}} \quad \ldots(20)$$

Now HPBW is given by,

$$\boxed{HPBW = \frac{BWFN}{2} = \frac{1}{\left(\frac{L}{\lambda}\right)}\ \text{rad}} \quad \ldots(21)$$

HPBW in degree is written as,

$$\therefore \quad \boxed{HPBW = \frac{57.3}{\left(\frac{L}{\lambda}\right)} \text{ degrees}} \qquad \ldots(22)$$

6. Directivity

$$G_{Dmax} = \frac{\text{Maximum radiation intensity}}{\text{Average radiation intensity}} = \frac{U_{max}}{U_{avg}} = \frac{U_{max}}{U_0} \qquad \ldots(23)$$

$$U_0 = \frac{P_{rad}}{4\pi} = \frac{1}{4\pi}\int_{\phi=0}^{2\pi}\int_{\theta=0}^{\pi}|E(\theta,\phi)|^2 \sin\theta \; d\theta \, d\phi \qquad \ldots(24)$$

or $|E_T| = n|E_0|$

For the normalized condition let us assume $E_0 = 1$, then

$$U_0 = \frac{P_{rad}}{4\pi} = \frac{1}{4\pi}\int_{\phi=0}^{2\pi}\int_{\theta=0}^{\pi}|E(\theta,\phi)|^2 \sin\theta \; d\theta \, d\phi \qquad \ldots(24)$$

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

$$|E_T| = n$$

Thus field from array is maximum in any direction θ when $\upsilon = 0$. Hence normalized field pattern is given by,

$$E_{Normalized} = \left|\frac{E_T}{E_{Tmax}}\right| = \frac{|E_0|}{n|E_0|} = \frac{1}{n}$$

Hence the field is given by,

$$\therefore \quad E_{Normalized} = \frac{\sin n\frac{\psi}{2}}{n\left(\sin\frac{\psi}{2}\right)} \quad \ldots(25)$$

Equation (23) indicated array factor, hence we can write electric field due to n array as

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

Assuming d is very small as compared to length of an array,

$$\sin\frac{\beta d\cos\phi}{2} \approx \frac{\beta d\cos\phi}{2}.$$

Then,

$$E = \frac{1}{n}\left[\frac{\sin\frac{n\beta d\cos\phi}{2}}{\sin\frac{\beta d\cos\phi}{2}}\right] \quad \ldots(26)$$

Substituting value of E in equation (24) we get

$$U_0 = \frac{1}{4\pi}\int_{\phi=0}^{2\pi}\int_{\theta=0}^{\pi}\left[\frac{\sin\frac{n\beta d\cos\phi}{2}}{\sin\frac{\beta d\cos\phi}{2}}\right]^2 \sin\theta \; d\theta \, d\phi$$

$$= \frac{1}{4\pi}\int_{\phi=0}^{2\pi} d\phi \cdot \int_{\theta=0}^{\pi}\left[\frac{\sin\frac{n}{2}\beta d\cos\phi}{\frac{n}{2}\beta d\cos\phi}\right]^2 \sin\theta \; d\theta$$

$$= \frac{1}{4\pi}[2\pi]\cdot\int_{\theta=0}^{\pi}\left[\frac{\sin z}{z}\right]^2 \sin\theta \; d\theta \quad \ldots(27)$$

Let

$$z = \frac{n}{2}\beta d \cos\theta$$

$$\therefore \qquad dz = -\frac{n}{2}\beta d \sin\theta \, d\theta$$

$$\therefore \qquad \sin\theta \; d\theta = -\frac{dz}{\frac{n}{2}\beta d}$$

Also when $\theta = \pi, \quad z = -\frac{n}{2}\beta d,$ and

when $\theta = 0, \quad z = +\frac{n}{2}\beta d$

Rewritting above equation we get,

For large array, n is large hence nβd is also very large (assuming tending to infinity). Hence rewriting above equation.

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

Interchanging limits of integration, we get

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

By integration formula,

$$\int_{-\infty}^{\infty}\left[\frac{\sin z}{z}\right]^2 dz = \pi.$$

Using above property in above equation we can write,

$$\boxed{U_0 = \frac{1}{n\beta d}[\pi] = \frac{\pi}{n\beta d}} \qquad \ldots(28)$$

From equation (23), the directivity is given by,

$$G_{Dmax} = \frac{U_{max}}{U_0}$$

But U_{max} = 1 at υ = 90° and substituting value of U_0 from equation (28), we get,

$$G_{Dmax} = \frac{1}{\left(\frac{\pi}{n\beta d}\right)} = \frac{n\beta d}{\pi} \quad \text{...(29)}$$

But β= 2π/λ

The total length of the array is given by, L = (n - 1) d ≈ nd, if n is very large. Hence the directivity can be expressed in terms of the total length of the array as,

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

5. Explain in detail about EFA with its properties. (16M)

[CO3-L2- April/May 2013, Nov/Dec 2014,May/June 2016]

Consider n number of identical radiators supplied with equal current which are not in phase as shown in the Fig. 11. Assume that there is progressive phase lag of βd radians in each radiator.

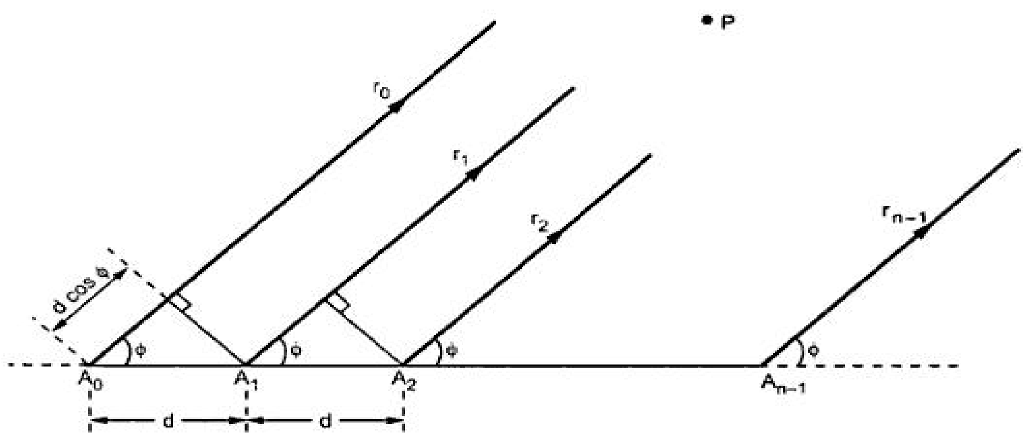


Fig. End fire array

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

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

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

Similarly the current supplied to A_2 is given by,

$$I_2 = I_1 \cdot e^{-j\beta d} = \left[I_0 \cdot e^{-j\beta d}\right] e^{-j\beta d} = I_0 \cdot e^{-j2\beta d}$$

Thus the current supplied to last element is

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

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

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

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

But $r_1 = r_0 - d\cos\upsilon$

$$\therefore \quad E_1 = \frac{I \, dL \sin\theta}{4\pi\omega\varepsilon_0}\left[j\frac{\beta^2}{r_0}\right] e^{-j\beta(r_0 - d\cos\phi)} \cdot e^{-j\beta d}$$

$$\therefore \quad E_1 = \left[\frac{I \, dL \sin\theta}{4\pi\omega\varepsilon_0}\left[j\frac{\beta^2}{r_0}\right] e^{-j\beta r_0}\right] e^{j\beta d\cos\phi} \cdot e^{-j\beta d}$$

$$\therefore \quad E_1 = E_0 \cdot e^{j\beta d(\cos\phi - 1)} \qquad \text{... (2)}$$

Let $\upsilon = \beta d(\cos\upsilon - 1)$

$$\therefore \quad E_1 = E_0 \, e^{j\psi} \qquad \text{... (3)}$$

The electric field produced at point P, due to A_2 is given by,

$$E_2 = E_0 \cdot e^{j2\psi} \qquad \text{... (4)}$$

Similarly electric field produced at point P, due to A_{n-1} is given by,

$$E_{n-1} = E_0 e^{j(n-1)\psi} \quad \text{... (5)}$$

The resultant field at point p is given by,

$$E_T = E_0 + E_1 + E_2 + ... + E_{n-1}$$

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

$$\therefore \quad E_T = E_0 \left[1 + e^{j\psi} + e^{j2\psi} + ... + e^{j(n-1)\psi}\right] \quad \text{... (6)}$$

$$E_T = E_0 \cdot \frac{1 - e^{jn\psi}}{1 - e^{j\psi}}$$

$$\therefore \quad \frac{E_T}{E_0} = \frac{\sin\frac{n\psi}{2}}{\sin\frac{\psi}{2}} \cdot e^{j\frac{(n-1)}{2}\psi} \quad \text{... (7)}$$

Considering only magnitude we get,

$$\therefore \quad \boxed{\left|\frac{E_T}{E_0}\right| = \frac{\sin\frac{n\psi}{2}}{\sin\frac{\psi}{2}}} \quad \text{...(8)}$$

Properties of Endfire Array:

1. Major lobe

For the end fire array where currents supplied to the antennas are equal in amplitude but the phase changes progressively through array, the phase angle is given by,

ʋ = βd(cosʋ -1) **...(9)**

In case of the end fire array, the condition of principle maxima is given by,

ʋ = = 0 i.e.

$$\boxed{\beta d(\cos\phi - 1) = 0} \qquad ...(10)$$

2. Magnitude of the major lobe

The maximum radiation occurs when υ= 0. Thus we can write,

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

$$\therefore \quad \boxed{|\text{Major lobe}| = n} \qquad ...(12)$$

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

3. Nulls

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

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

Equating ratio of magnitudes of the fields to zero,

$$\therefore \quad \left|\frac{E_T}{E_0}\right| = \frac{\sin n\frac{\psi}{2}}{\sin\frac{\psi}{2}} = 0$$

The condition of minima is given by,

$$\boxed{\sin n\frac{\psi}{2} = 0, \quad \text{but } \sin\frac{\psi}{2} \neq 0} \qquad ...(13)$$

Hence we can write,

$$\sin n\frac{\psi}{2} = 0$$

$$\text{i.e.} \quad n\frac{\psi}{2} = \sin^{-1}(0) = \pm m\pi, \text{ where } m = 1, 2, 3, ...$$

Substituting value of ʋ from equation (9), we get,

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

But β= 2π/λ

$$\therefore \frac{nd}{\lambda}(\cos\phi - 1) = \pm m \qquad \ldots(14)$$

Note that value of (cosʋ-1) is always less than 1. Hence it is always negative. Hence only considering -ve values, R.H.S., we get

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

$$\text{i.e. } \cos\phi - 1 = -\frac{m\lambda}{nd}$$

$$\boxed{\phi_{min} = \cos^{-1}\left[1 - \frac{m\lambda}{nd}\right]} \qquad \ldots(15)$$

where, n= number of elements in the array d= spacing between elements in meter

λ= wavelength in meter

m= constant= 1, 2 , 3....

Thus equation (15) gives direction of nulls

Consider equation(14),

$$\cos\phi_{min} - 1 = \pm\frac{m\lambda}{nd}$$

Expressing term on L.H.S. in terms of halfangles, we get,

$$2\sin^2\frac{\phi_{min}}{2} = \pm\frac{m\lambda}{nd} \qquad \ldots\left(\cos\theta - 1 = 2\sin^2\frac{\theta}{2}\right)$$

4. Side Lobes Maxima

The directions of the subsidary maxima or side lobes maxima can be obtained if in equation (8),

Hence sin(nυ/2), is not considered. Because if nυ/2=±π/2 then sin nυ/2 =1 which is the direction of principle maxima.

Hence we can skip sin nυ/2 = ±π/2 value Thus, we get

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

$$\therefore \quad \boxed{n\frac{\psi}{2} = \pm\frac{3\pi}{2}, \pm\frac{5\pi}{2}, \pm\frac{7\pi}{2}, \ldots\ldots} \qquad \ldots(17)$$

Putting

value of υ from equation (9) we get

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

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

$$\therefore \; n\beta d(\cos\phi - 1) = \pm(2m+1)\pi$$

Now equation for υ can be written as, But β = 2π/λ

$$n\left(\frac{2\pi}{\lambda}\right)d(\cos\phi - 1) = \pm(2m+1)\pi$$

$$\text{i.e. } \cos\phi - 1 = \pm(2m+1)\frac{\lambda}{2nd}$$

Note that value of (cosυ-1) is always less than 1. Hence it is always negative. Hence only considering -ve values, R.H.S., we get

5. Beamwidth of Major Lobe

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

From equation (16) we get,But nd≈ (n-1)d if n is very large. This L= (nd) indicates total length of the array. So equation (20) becomes,

BWFN is given by,

$$\phi_{min} = 2\sin^{-1}\left[\pm\sqrt{\frac{m\lambda}{2nd}}\right] \quad ...(19)$$

$$\therefore \quad \sin\frac{\phi_{min}}{2} = \pm\sqrt{\frac{m\lambda}{2nd}}$$

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

$$\phi_{min} = \pm\sqrt{\frac{4m\lambda}{2nd}} = \pm\sqrt{\frac{2m\lambda}{nd}} \quad ...(20)$$

$$\phi_{min} = \pm\sqrt{\frac{2m\lambda}{L}} = \pm\sqrt{\frac{2m}{L/\lambda}} \quad ...(21)$$

$$\boxed{BWFN = 2\phi_{min} = \pm 2\sqrt{\frac{2m}{L/\lambda}}} \quad ...(22)$$

BWFN in degree is expressed as

$$BWFN = \pm 2\sqrt{\frac{2m}{L/\lambda}} \times 57.3 = \pm 114.6\sqrt{\frac{2m}{L/\lambda}} \text{ degree}$$

6. Directivity

The directivity in case of endfire array is defined as,

$$G_{Dmax} = \frac{\text{Maximum radiation intensity}}{\text{Average radiation intensity}} = \frac{U_{max}}{U_{avg}} = \frac{U_{max}}{U_0} \quad \ldots(23)$$

where, U_0 is average radiation intensity which is given by, For endfire array, U_{max} = 1and $U_0 = \frac{\pi}{2n\beta d}$

$$\therefore \quad G_{Dmax} = \frac{1}{\frac{\pi}{2n\beta d}} = \frac{2n\beta d}{\pi}$$

$$\therefore \quad G_{Dmax} = 2n\left(\frac{2\pi}{\lambda}\right)\cdot\frac{d}{\pi}$$

$$\therefore \quad \boxed{G_{Dmax} = 4\left(\frac{nd}{\lambda}\right)} \quad \ldots(24)$$

The total length of the array is given by, L = (n - 1) d ≈ nd, if n is very large. Hence the directivity can be expressed in terms of the total length of the array as,

$$\therefore \quad \boxed{G_{Dmax} = 4\left(\frac{L}{\lambda}\right)} \quad \ldots(25)$$

6. Explain in detail about binomial array. (8M)

[CO3-L2- April/May 2014, April/May 2015]

In order to increase the directivity of an array its total length need to be increased. In this approach, number of minor lobes appears which are undesired for narrow beam applications. In has been found that number of minor lobes in the resultant pattern increases whenever spacing between elements is greater than λ/2.

As per the demand of modern communication where narrow beam (no minor lobes) is preferred, it is the greatest need to design an array of only main lobes. The ratio of power density of main lobe to power density of the longest minor lobe is termed side lobe ratio. A particular technique used to reduce side lobe level is called tapering. Since currents/amplitude in the sources of a linear array is non-uniform, it is found that minor lobes can be eliminated if the centre element radiates more strongly than the other sources. Therefore tapering need to be done from centre to end radiators of same specifications. The principle of tapering are primarily intended to broadside array but it is also applicable to end-fire array. Binomial array is a common example of tapering scheme and it is an array of n-isotropic sources of non-equal amplitudes. Using principle of pattern multiplication, John Stone first proposed the binomial array in 1929, where amplitude of the radiating sources arc arranged according to the binomial expansion. That is. if minor lobes

appearing in the array need to be eliminated, the radiating sources must have current amplitudes proportional to the coefficient of binomial series, i.e. proportional to the coefficient of binomial series, i.e.

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

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

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

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

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

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

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

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

and directivity

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

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

The far-field pattern can be found by substituting n = 3 and 4 in the above expression and they take shape as shown in Fig. 14(a) and (b).

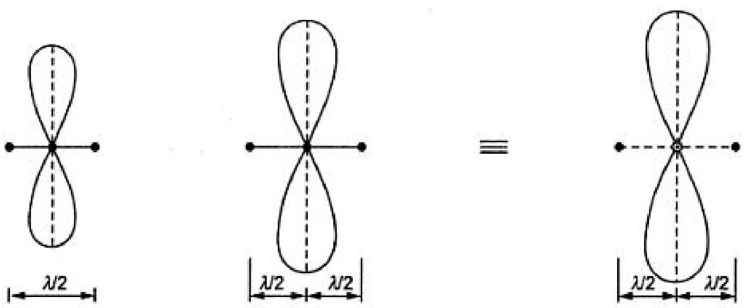


Fig. 14(a) Radiation pattern of 2-element array with amplitude ratio 1:2:1.

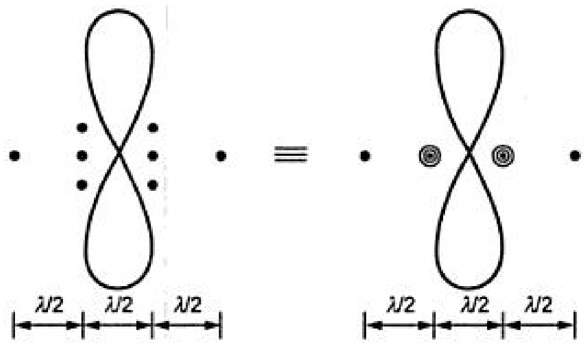


Fig 14(b) Radiation pattern of 3-element array with amplitude ratio 1:3:3:1.

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

Disadvantages of Binomial Array

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

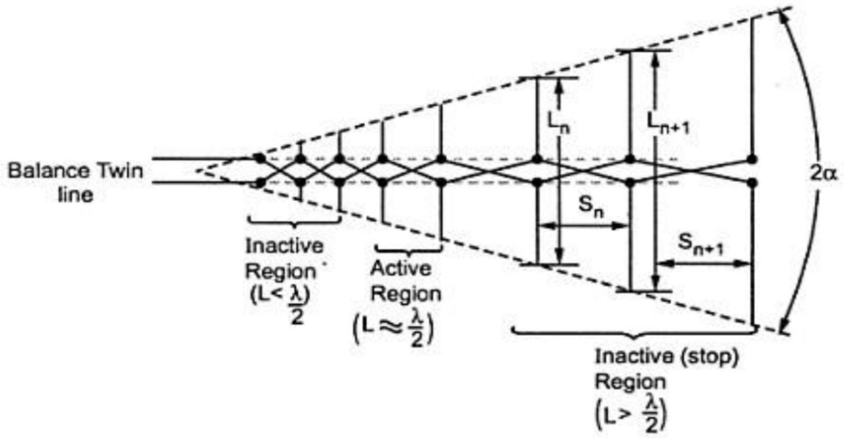
(b) As the length of array increases, larger current amplitude ratios are required.

Unit – IV

Special Antennas

Part – A

1.Draw the log periodic dipole antenna structures at UHF & VHF ranges. (or) Name & Draw a frequency independent antenna. [CO4-L1]



2.What is LPDA? Why it is called so? [CO4-L1- Nov/Dec 2011]

LPDA means log periodic dipole array. It is defined as an antenna whose electrical properties repeat periodically with logarithm of the frequency. The geometry of log periodic antenna is so chosen that electrical properties must repeat periodically with logarithm of frequency.

3.What is the need for transposing the lines in log periodic antenna? [CO4-L1]

Transposing lines introduce 1800 phase shift b/w adjacent dipoles.

4. What is meant by resonant antenna? [CO4-L1]

The resonant antenna consists of standing wave exist along its length. The radiation pattern is bidirectional.

5. Write the features of log periodic dipole array. [CO4-L2- April/May 2014]

- LPDA excited from the short length dipole side (or) high frequency side.
- For unidirectional LPDA the structure produces backward direction and forward direction is very small.

6.What is the effect of decreasing included angle? [CO4-L1]

The directivity of the antenna increases by means of decreasing the included angle.

7. What is the design ratio & frequency ratio of log periodic antenna? [CO4-H3]

Design ratio or scale factor is given by

t = Rn / Rn+1 = Ln/Ln+1

Frequency ratio or bandwidth:

F = Ln+1/ Ln

8.Differentiate between resonant and non resonant antenna. [CO4-L2]

Resonant antenna	**Non resonant antenna**
Standing wave exist along its length	No standing waves exist along its length.
Radiation pattern is bidirectional	Radiation pattern is unidirectional.

9.What is the effect of α =0^o and α =90^o in helical antenna? [CO4-L1]

If α =0 ^{o} then helix becomes loop.

If α =90^o then helix becomes linear dipole antenna.

10. State the Rumsey's principle. [CO4-L2- May/June 2016]

It states that the impedance and pattern properties of antenna will be frequency independent if the antenna shape is specified only in terms of angles.

11. Differentiate between normal and axial mode of operation in helical antenna. [CO4-L2- April/May 2015]

Normal mode	**Axial mode**
It is operated at broadside array direction	It is operated at end fire mode.
Bandwidth is low	Bandwidth is wider.

12. What are the applications of helical antenna. [CO4-L3- Nov/Dec 2012,May/June 2016]

<u>Helical antenna</u>:

- VHF transmission such as satellite communication.
- Space telemetry link with ballistic missiles, satellites, etc.
- Axial mode helical antennas are used to achieve circularly polarized waves over extremely wide bandwidths.

- They are used in space communications systems such as transmitting telemetry data from moon to the earth.
- A single helical antenna or an array of helical antennas are useful in transmitting or receiving VHF signals through the ionosphere.

13. What are the types of gain measurement? [CO4-L2]

- Direct comparison method
- Absolute gain of identical antennas

14. What is meant by pitch angle in helical antenna? [CO4-L1- Nov/Dec 2014]

The pitch angle is the angle between a line tangent to the helix wire and the plane normal to the helix axis. $\alpha = \tan^{-1}(S/\pi D)$

15.What is meant by frequency independent antenna. Give example. [CO4-L1-April/May 2014 &2015]

An antenna in which the impedance, radiation pattern and directivity remain constant as a function of frequency is called as frequency impendent antenna. E.g. :log periodic antenna. Eg. Log peridic antenna

16. For a 20 turn helical antenna operating at 3GHz with circumference C=10cm and the spacing between the turns 0.3. Calculate directivity and half power beam width. [CO4-H3- April/May 2015]

Gn:N=20 ,C=10 cm S=0.3 λ , f=3GHz

λ =c/f =3x108/3x109 =0.1 ,D=15NSC2 / λ3 =90

HPBW=52/C sqrt(λ3/NS) =21.22 deg

17.What are the two procedures for radiation pattern measurement? [CO4-L2]

- The primary is kept stationary whereas secondary is moved around along a circular path at a constant radius.
- Both the antennas are kept in fixed positions having a suitable spacing between them and secondary antenna beam aimed are primary antenna.

18.What are the two methods for measuring impedance? [CO4-L2]

- Bridge method for low frequencies
- Slotted line or standing wave method for high frequencies

19.What are the advantages of anechoic chamber? [CO4-L1- Nov/Dec 2012]

It stimulates a reflection less free space and allows all-weather antenna measurements in a controlled environment.

- The test area is isolated from interfering signals much better than at outdoor ranges.

20. What are the requirement of reflectors used in Yagi uda array?
[CO4-L1- April/May 2012, April/May 2014]

Reflector drive its power from a driver, it reduces the signal strength in it's own direction thus reflects the radiation towards the driver and directors.

21. Name the errors Possible in VSWR measurements?

[CO4-L2- May/June 2013]

The possible sources in VSWR measurement are

1) Vmax and Vmin may not be measured in the square – Law region of the crystal detector.
2) The Probe thickness and depth of penetration may produce reflections in the Line and also distortion in the field to be measured
3) When VSWR<1.05, the associated VSWR of connector produces significant error in VSWR measurement. Very Good low VSWR (,1.01) connectors should be used for very low VSWR measurements.
4) Any Harmonics and spurious signals from the source may be tuned by the probe to cause measurement error.
5) A residual VSWR of slotted line arises due to mismatch impedance between the slotted line and the main line

22. What is the significance of VSWR measurement? [CO4-L1- Nov/Dec 2014]

Double minima method is used where the Probe depth does not introduce an error.

23. What is called as Transverse electromagnetic cell(TEM Cell)? [CO4-L1]

It is a rectangular coaxial transmission line resembling a stripline and it allows a transmission of TEM waves only.

24. What is called as Anechoic chamber? [CO4-L1]

It is indoor type of antenna range in which chamber walls, celing, floor are covered with energy absorbers. It is best suited for small antennas.

25. What is called as Near field range? [CO4-L1- April/May 2011]

It is a small indoor type of range in which only near field measurements are made. Then using numerical methods, far field measurements can be determined from near field measurements.

26. What is called as Reflected range? [CO4-L1]

In this the heights of transmitting antenna and AUT are selected such that a constructive interference at AUT.

27. What is called as Ground range? [CO4-L1]

Ground range: This is a range in which tall towers are not needed as the transmitting antenna is placed above surface and acts like mirror.

28. What is called as frequency reuse? [CO4-L1- April/May 2013]

The design process of selecting and allocating the same bands of frequencies to different cells of cellular base stations within a system is referred to as frequency reuse.

29. What is called as reconfigurable antenna? [CO4-L1]

- A reconfigurable antenna is an antenna capable of modifying dynamically its frequency and radiation properties in a controlled and reversible manner.
- Reconfigurable antennas differ from smart antennas because the reconfiguration mechanism lies inside the antenna rather than in an external beamforming network.

30. What do you mean by travelling wave antenna? [CO4-L1]

- The antenna in which the standing wave does not exist along the length of the antenna is called as travelling wave antenna.
- In general the standing waves are produced when the line is not properly terminated which causes reflections at the output or loadside.
- The standing waves travel due to reflections in the resonant antennas. But in the travelling wave antenna, the standing waves do not exist.
- That means the travelling wave antenna is non-resonant type antenna or aperiodic antenna.

31. What is the significance of helical antenna? [CO4-L2- April/May 2014]

- For the space communications applications, the helical antennas are most suitable as they have wide bandwidth, higher directivity and circular polarization.
- To transmit or receive VHF signals through ionosphere generally an array of helical antennas is used. The helical antenna is widely used for space and satellite communications.

32.Write the basic concept of antenna measurement? [CO4-L1]

- The antenna under test (AUT) is considered to be located at the origin of the coordinate system. The source antenna is placed at different locations with respect to the AUT.
- The source antenna may be transmitting or receiving. To achieve different locations, the number of samples of the pattern are obtained.
- To achieve different locations, the number of samples of the pattern are obtained. To achieve different locations, generally AUT is rotated. To achieve sharp sample of pattern, it is necessary that there exists single direct signal path between the AUT and source antenna.

33. Write about near and far field measurement. [CO4-L1- April/May 2013]

- There are three main regions of the radiated field of the antenna. The region very close to antenna is called as reactive near field region(radiansphere).
- The region next to reactive near field region which is called as radiating near field region (Fresnal region).
- The region located far away from the antenna is called as far field region.(fraunhofer region).

34. Mention sources of errors in antenna measurement. [CO4-L1]

- Errors due to finite measurement distance between antennas.
- Reflections from surroundings.
- Errors due to coupling in the reactive near field.
- Errors due to misalignment of antennas.
- Errors due to manmade interface.
- Errors due to cables.
- Errors due to impedance mismatch.
- Errors due to imperfections of instruments.

35. Compare indoor ranges with outdoor ranges. [CO4-L2- April/May 2012]

Indoor ranges	**outdoor ranges**
1.They have controlled environment	They have controlled environment
2.They show all weather capability	They do not show all weather capability.
3. The measurements in indoor ranges are protected from external EMI.	The measurements in outdoor ranges are not protected from external EMI.

Part-B

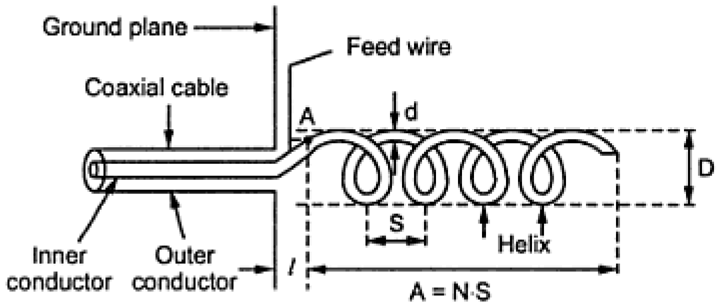
1. Draw the construction of Helical Antenna & principle of helical antenna with different modes of operation. How its differ from other antenna. (16M)

[CO4-L1&L2- May/June 2013 & 2014]

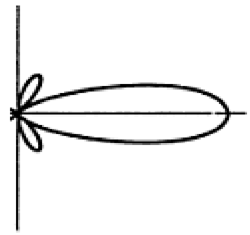
HELICAL ANTENNA:

Helical Antenna is a broadband VHF & UHF Antenna used to provide circular polarization.

Construction:



Radiation Pattern:



- Helical antenna consists of a helix of thick copper wire in a shape of screw thread and used with a flat metal called a ground plane.
- It is fed by a coaxial cable & is connected between Helix & ground plane ie, One end of the helix is connected to the centre conductor of the cable & outer conductor is connected to the ground plane.
- Radiation depends on the diameter of the helix D & turn spacing S.
- Dimensions of helix are,

  C → Circumference of helix (πD)

  D → Diameter of helix conductor.

  A → Axial length = NS

  N → No.of turns.

L → Length of one turn.

l → Spacing of helix from ground plane.

Relation between S, C and α :

Total length, $L = \sqrt{S^2 + C^2}$

$= \sqrt{S^2 + (\pi D)^2}$

Pitch angle:

It is the angle between a line tangent to the helix wire and the plane normal to the helix axis.

$$\tan \alpha = \frac{S}{C} = \frac{S}{\pi D}$$

$$\alpha = \tan^{-1}\left(\frac{S}{\pi D}\right)$$

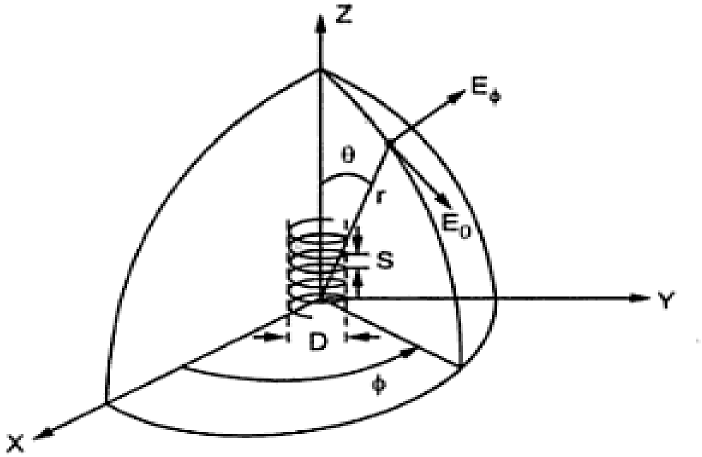
The different radiation characteristics are obtained by changing the above parameters in relation to wavelength.

MODES OF RADIATION / MODES OF OPERATION:

- Normal mode (or) Perpendicular mode
- Axial (or) Beam mode of radiation.
-

Normal mode of operation:

- It is operated in 'broad side array'.
- Maximum radiation is perpendicular to the helix axis.
- Bandwidth is narrow so efficiency is low.



- Dimensions of helix is small compared with wavelength NL << λ.

- Radiation pattern is a combination of the equivalent radiation from a short dipole & small loop.
- When α = 0, helix corresponds to a loop when α = 90˚, helix corresponds to a linear dipole.
- Helix antenna is having number of small loops & short dipole connected in series in which loop diameter is same as helix diameter & helix spacing is same as dipole length.
- Far field of the small loop is given by,

$$E\phi = \frac{120\pi^2[I]\sin\theta A}{r\lambda^2}$$

Where I → Retarded current

r → Distance

A → Area of loop = $\frac{\pi D^2}{4}$

Far field of the short dipole is given by,

$$E\phi = \frac{j60\pi[I]\sin\theta S}{r\lambda}$$

Where S = L = Length of dipole

Axial ratio:

$$AR = \frac{|E\theta|}{|E\phi|}$$

$$= \frac{60\pi[I]\sin\theta S}{r\lambda} \times \frac{r\lambda^2}{120\pi^2[I]\sin\theta}$$

$$= \frac{S\lambda}{2\pi A}$$

Sub A = $\frac{\pi D^2}{4}$

$$AR = \frac{S\lambda}{2\pi\left(\frac{\pi D^2}{4}\right)} \Rightarrow \frac{2S\lambda}{\pi^2 D^2} \text{ Axial ratio}$$

When axial ratio is 0 → linear horizontal polarization

Axial ratio is α → Linear vertical polarization

Axial ratio is 1 → Circular polarization

∴ For circular polarization

$$AR = 1$$

$$\frac{2S\lambda}{\pi^2 D^2} = 1$$

$$S = \frac{\pi^2 D^2}{2\lambda}$$

Sub in pitch angle (α),

$$\alpha = \tan^{-1}\left(\frac{S}{\pi D}\right)$$

$$= \tan^{-1}\left(\frac{\pi^2 D^2}{2\lambda} \times \frac{1}{\pi D}\right)$$

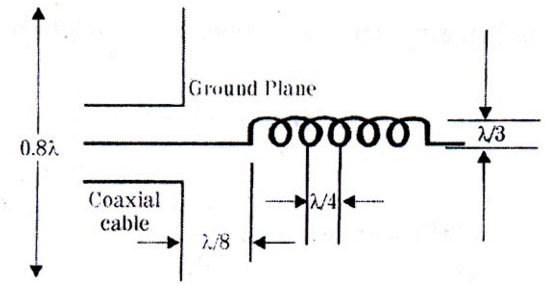
$$= \tan^{-1}\left(\frac{\pi D}{2\lambda}\right)$$

$$\alpha = \tan^{-1}\left(\frac{C}{2\lambda}\right) \qquad (\because C = \pi D)$$

This is the condition for pitch angle to get circular polarization.

Axial or Beam mode of radiation:

- It is operated in 'End Fire Array'.
- Maximum radiation along the axis.
- Bandwidth is broad so efficiency is high.
- It is otherwise called as Monofilar axial mode helical antenna.
- Monofilar helix is an antenna constructed by a single conductor.



- It is the easiest of all antenna to built.
- Design parameters are,

1. Beamwidth.
2. Gain.
3. Impedance.
4. Axial ratio.

- Terminal impedance is given by,

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

- Half power beam width is given by,

$$HPBW = \frac{52}{C} * \frac{\sqrt{\lambda^3}}{NS} \text{ degree}$$

- Beam width between first null is given by,

$$BWFN = \frac{115}{C} * \frac{\sqrt{\lambda^3}}{NS} \text{ degree}$$

- Directive gain is given by,

$$D = \frac{15NSC^2}{\lambda^3}$$

- Axial ratio is given by,

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

Advantages:

- Simple antenna.
- Higher directivity.
- Mostly circular polarization is obtained.
- Broadband bandwidth.

Applications:

- Radio astronomy.
- Telemetry.
- Satellite & space communication.

2. Draw the structure of LPDA & derive the design of LPDA (or) explain the construction and characteristics features of frequency independent antenna.(16M)

[CO4-H3- NovDec 2013, Nov/Dec 2014, May/June 2016]

- Log periodic dipole antenna (LPDA) (or) High frequency antenna (or) Frequency independent antenna.
- A log periodic antenna is a broadband narrow beam antenna.
- It is a frequency independent antenna.

Frequency independent antenna:

- It is defined as the antenna for which the impedance & radiation pattern remains constant as a function of the frequency.
- In order to be frequency independent, the antenna should expand or contract in proportion to the wavelength.

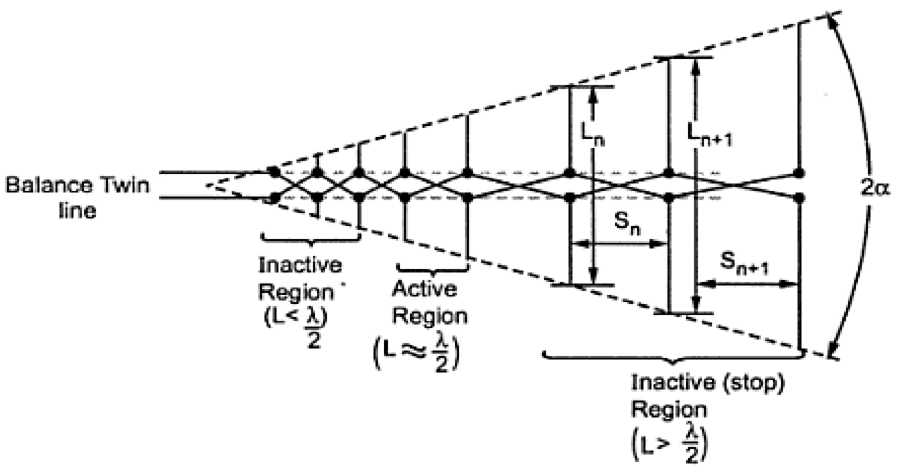
Log periodic dipole array:

The electrical properties of the antenna varies periodically as a function of logarithm of frequency.

Need for transposing lines:

Transposing lines is a twisted transmission line provides 180˚ phase shift between adjacent dipoles.

Construction:



- It consists of number of dipoles of different length & spacing.
- The array is fed using a balanced transmission line which is connected at apex of the array.
- Transmission line is transposed between adjacent pairs of terminals of dipoles.
- Length of the dipole increases from feed point towards the end.
- Dipole length & spacing between two adjacent dipoles are related through parameter called Design ratio (or) Scale factor.

- Design ratio is denoted by τ.
- Relationship between S, R, L

$$\frac{Rn}{Rn+1} = \frac{Sn}{Sn+1} = \frac{Ln}{Ln+1} = \tau$$

τ → Periodicity factor

$$K = \frac{1}{\tau}$$

Working principle:

Analysis of a log periodic dipole array can be done by considering 3 regions of antenna.

1. Inactive transmission region (L < λ/2)
2. Active region (L = λ/2)
3. Inactive reflective region (L > λ/2)

Inactive transmission region:

- Dipole length less than resonant length.
- Dipole in this region offers capacitive impedance.
- Current in region are small & hence small radiation in backward direction.
- Spacing between the element is smaller.

Active region:

- Dipole length is equal to resonant length.
- Dipole in this region offers Resistive impedance.
- Current in this region are large.
- More radiation produced in backward direction & little radiation in forward direction.

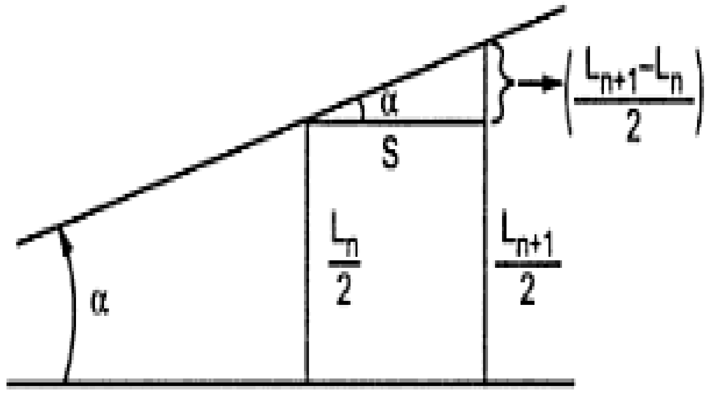
Inactive reflective region:

- Dipole length are larger than resonant length.
- Dipole in this region offers Inductive impedance.
- Current are smaller and also lag the base voltage.
- Small amount of incident wave is reflected back towards backward direction.

DESIGN OF LPDA:

Design parameters are,

1. Apex angle (α)
2. Design ratio (τ)
3. Spacing factor (σ)



$$tan\frac{\alpha}{2} = \frac{opp}{adj}$$

$$= \frac{\frac{Ln+1 - Ln}{2}}{S}$$

$$tan\frac{\alpha}{2} = \frac{\frac{Ln+1}{2}\left[1 - \frac{Ln}{Ln+1}\right]}{S}$$

$$= \frac{L_{n+1}\left(1 - \frac{Ln}{Ln+1}\right)}{2S}$$

Taking $L_{n+1} = \frac{\lambda}{2}$ (when active)

$$\frac{Ln}{Ln+1} = \frac{1}{K}$$

$$\therefore tan\ \frac{\alpha}{2} = \frac{\lambda/2\left(1 - \frac{1}{K}\right)}{2S}$$

$$= \frac{\lambda\left(1-\frac{1}{K}\right)}{4S}$$

$$\tan\frac{\alpha}{2} = \frac{\left(1-\frac{1}{K}\right)}{4\sigma}$$

Where $\sigma = \frac{S}{\lambda}$ called spacing factor

$\tau = \frac{1}{K}$

$$tan\frac{\alpha}{2} = \frac{1-\tau}{4\alpha}$$

$$\frac{\alpha}{2} = \tan^{-1}\left(\frac{1-\tau}{4\sigma}\right)$$

$$\alpha = 2\tan^{-1}\left(\frac{1-\tau}{4\sigma}\right)$$

Characteristics of LPDA:

Features:

- LPDA excited from the short length dipole side (or) high frequency side.
- For unidirectional LPDA the structure produces backward direction and forward direction is very small.

Applications:

- Used in field of HF communication.
- Used in TV reception.
- Best suited for all round monitoring.

3. Explain in detail about spiral antenna? (8M) [CO4-L2- April/May 2012]

SPIRAL ANTENNA:

- It is a frequency independent antenna.
- It is based on Rumsey's principle.

- It states that impedance & pattern properties will be frequency independent if the antenna shapes is specified only in terms of angles.
- It produces a circular polarized waves within the band of operation.
- It produces elliptical polarized waves outside the band of operation.
- A spiral is a geometrical shape found in nature & it can be geometrically described using polar co-ordinates (r,ϴ).
- 2 geometrical shapes of spiral antenna are,
    - Logarithmic / Equiangular spiral antenna.
    - Conical Equiangular spiral antenna.

Logarithmic spiral antenna:

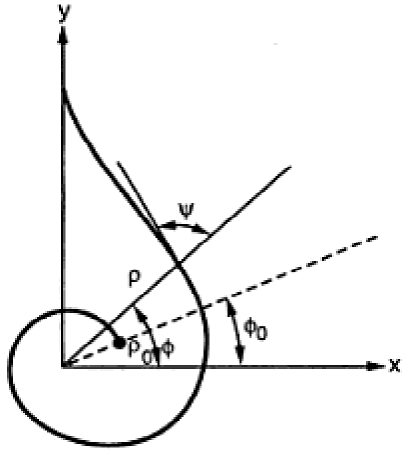
Equation for a log spiral is given by,

$$r = r_0 a^{\theta} \quad \rightarrow \quad (1)$$

where r radial distance to point P.

r_0 radius at ϴ = 0

Angle w.r.t x-axis.



Taking ln on both sides of eqn(1),

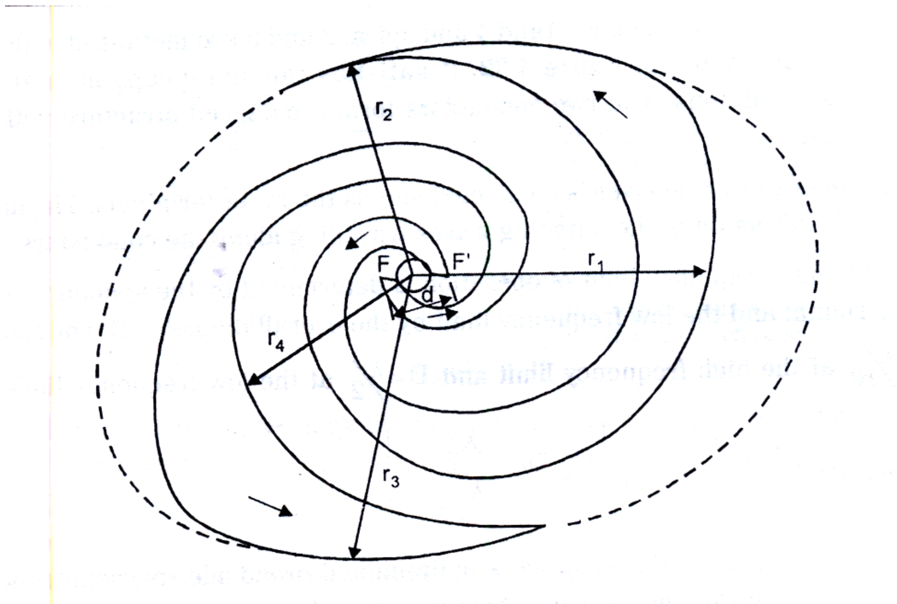
$$\ln r = \ln r_0 + \theta \ln a \quad \rightarrow \quad (2)$$

Diff w.r.t ϴ,

$$\frac{1}{r}.\frac{dr}{d\theta} = \ln a. \quad \rightarrow \quad (3)$$

$$\ln a = \frac{1}{tan\beta} \quad \rightarrow \quad (4)$$

$$\theta = \frac{\ln r}{\ln a} \quad \Rightarrow \quad tan\beta \ln r. \quad \rightarrow \quad (5)$$



Consider a spiral described by,

$$r_1 = r_0\, a^{\theta} \quad \rightarrow \quad (6)$$

Multiply above equation by K,

$$r_2 = K r_0\, a^{\theta} \qquad (\because\ K = a^{-\delta})$$

$$r_2 = r_0\, a^{\theta - \delta} \quad \rightarrow \quad (7)$$

Second antenna is obtained by rotating the original antenna by angle δ.

Third antenna is obtained by rotating 1st spiral by 180˚.

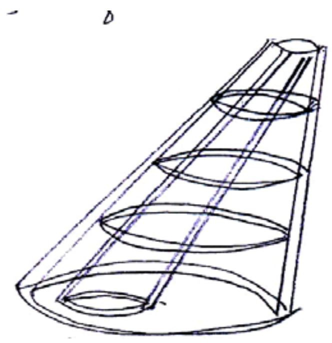
$$r_3 = a^{\theta - \pi} \quad \rightarrow \quad (8)$$

Fourth antenna is obtained by rotating 2nd spiral by 180˚

$$r_4 = a^{\theta - \pi - \delta} \quad \rightarrow \quad (9)$$

Conical equiangular spiral antenna:

- Conical spiral antenna is a balanced structure which may be fed at the apex by means of a balanced transmission line.
- Spiral arms are wrapped on the surface of the cone.
- Conical angle less than 45˚ is chosen.
- It produce unidirectional radiation pattern.



Conical equiangular spiral antenna:

4. Explain about radiation pattern measurement. (8M) [CO4-L2-April/May 2013]

RADIATION PATTERN MEASUREMENT:

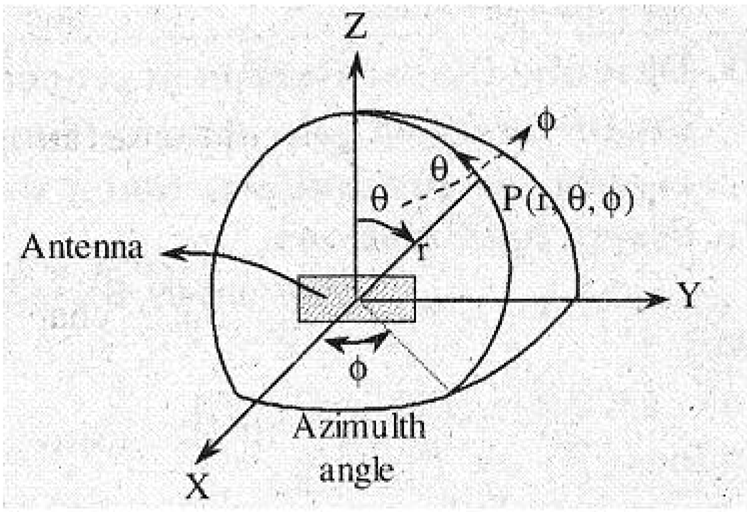


fig 5.1 Antenna Patterns

- Radiation pattern is described by field strength.
- Radiation characteristics of the antenna as a function of Θ, ϕ for constant radius & frequency is called radiation pattern.
- Test antenna is placed at the origin of spherical coordinates.

XY → Horizontal plane.

XZ → Vertical plane.

- There are 2 antennas,
  - Horizontal and Vertical

- Horizontal antenna has 2 pattern

  E ɵ (ɵ =90˚ , Φ) → $E - \text{plane pattern}$

  E ɵ (ɵ, Φ = 0˚) → $H - \text{plane pattern}$
- Vertical antenna has 2 pattern

  E ɵ (ɵ =90˚, Φ) → $H - \text{plane pattern}$

  E ɵ (ɵ, Φ = 0˚) → $E - \text{plane pattern}$
- 2 procedure for radiation pattern measurement,

  (i) 1˚ antenna kept stationary.

  2˚ antenna is moved around along a circular path.

  (ii) 2˚ antenna are kept stationary.

<u>Measurement set up for radiation pattern:</u>

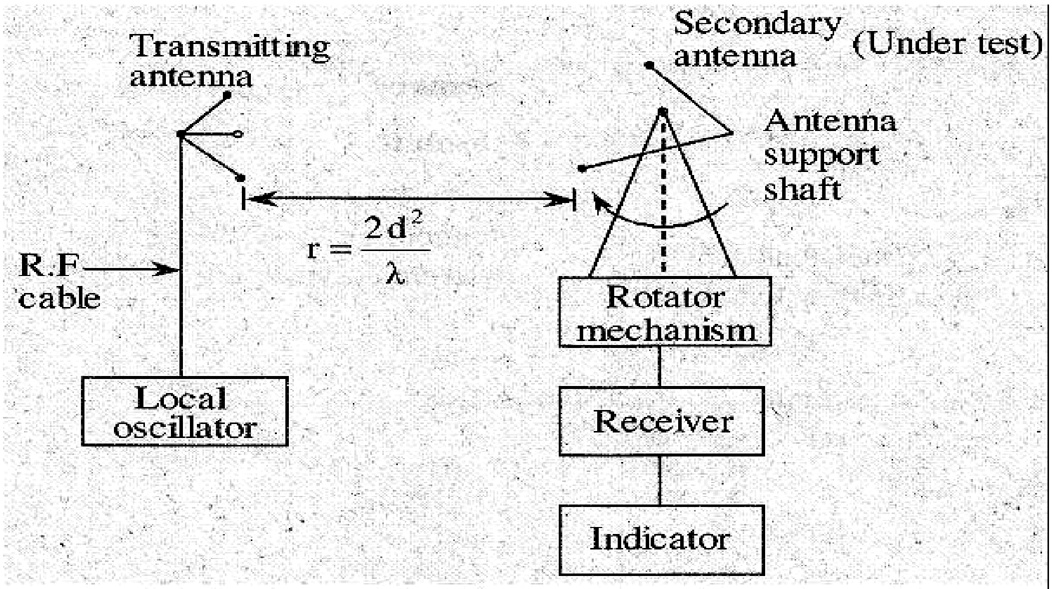


fig 6.1 Measurment of field pattern

-
- 1˚ antenna is a transmitting antenna.
- 2˚ antenna is a receiving antenna. It is rotated using antenna rotator mechanics.
- For H-plane pattern measurement, shaft is rotated with both the antennas vertical.
- For E-plane pattern measurement, shaft is rotated with both the antennas horizontal. Indicator is used to measure amplitude.

Uniform distance requirement:

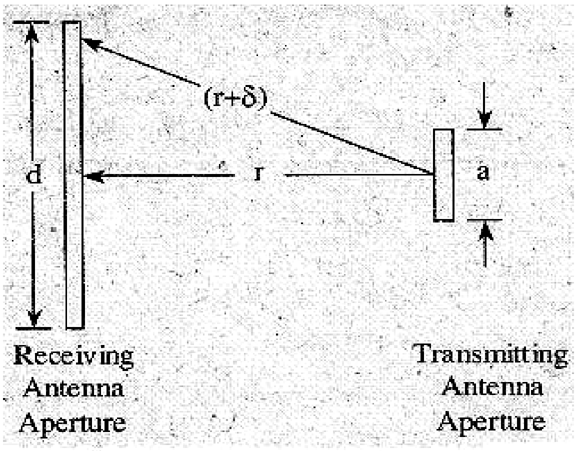
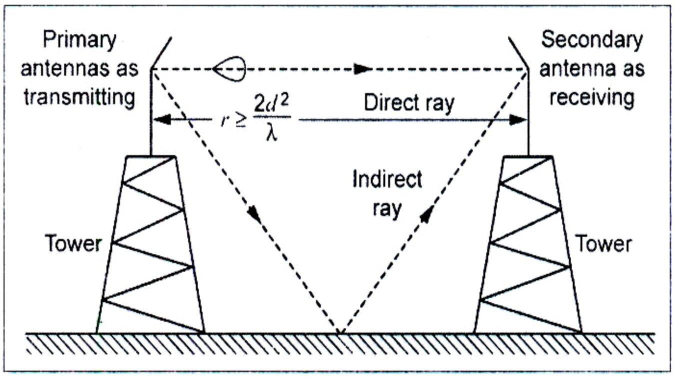


fig 9.1 Minimum distance requirement

- For far field radiation pattern, distance between 1˚ & 2˚ antenna is very large.
- For near field radiation pattern, distance between 1˚ & 2˚ antenna is very smaller.

Uniform amplitude requirement:



- Distance between 1˚ antenna & 2˚ antenna $r \geq 2d^2/\lambda$
- 1˚ antenna should produce wave with uniform amplitude & phase.
- Direct rays & indirect rays interferences should be minimized.
- Reflection from tall buildings, trees should be minimized.

Measurement of phase:

There are 2 types,

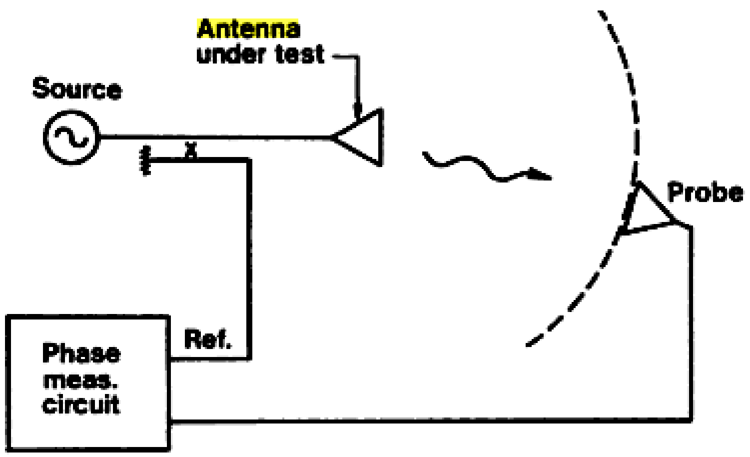
- Near field phase pattern measurement.
- Far field phase pattern measurement.

1.Near field phase pattern measurement:

Direct comparison of phase of the received signal with reference signal is possible.

2.Far field phase pattern measurement:

In this direct comparison is not possible.



5. Explain about measurement of gain. (8M) [CO4-L2- April/May 2014]

GAIN MEASUREMENT:

- The performance of any antenna can be described in terms of gain of a antenna.ie, gain refers to figure of merit.

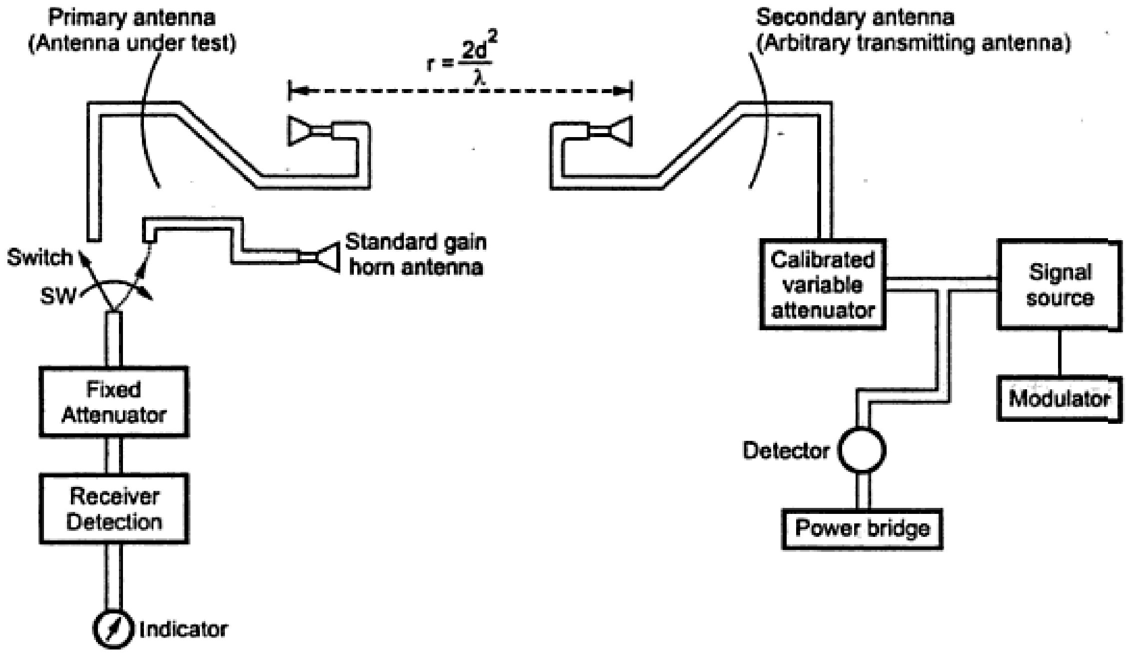
2 standard methods,

- Direct comparison method
- Absolute gain method

$$\text{Gain} = \frac{\text{Max radiation intensity of test antenna}}{\text{Max radiation intensity of reference antenna}}$$

- Gain is directly proportional to directivity.ie, G α D.
- Gain is compared between 2 antennas.
- This method is used at high frequencies.
- It compares signal strength between transmitter & receiver with unknown gain antenna & standard gain antenna.
- Standard antenna is a one in which gain is accurately known.
  Horn antenna is a standard antenna.

Gain measurement by direct comparison method:



Set up of gain measurement by gain comparison method

1˚ antenna – Antenna under test.
2˚ antenna – Receiver antenna.

Procedure:

- Antenna is adjusted to the direction of 2° antenna to have maximum signal intensity.
- AUT is connected to receiver by changing the position of switch.
- Finally Attenuator & power bridge reading are recorded.

Case (i):

$P_1 = P_2$ → No correction to be applied.

Power gain, $G_\phi = A_2/A_1$ where A_1 & A_2 – Power levels

$\log G_p = \log(A_2/A_1) \Rightarrow \log A_2 - \log A_1$

Case (ii):

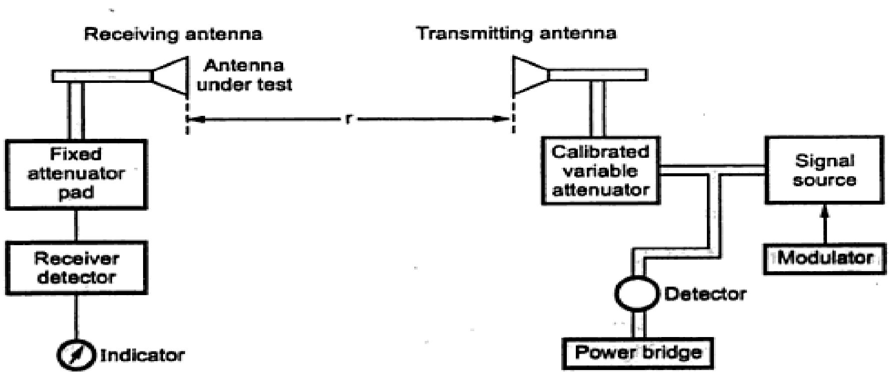
$P_1 \neq P_2$ → Correction need to be applied.

$$\frac{P1}{P2} = P$$

$$\log \frac{P1}{P2} = P$$

$$G = G_p \times \frac{P1}{P2} \Rightarrow \frac{A2}{A1} \times \frac{P1}{P2}$$

Measurement of absolute gain:



Transmitting and receiving antennas for absolute gain measurement

<u>Procedure</u>:

- Input to transmitting antenna is adjusted to a level & corresponding receiver reading level is recorded.
- Transmitter is disconnected & connected to receiver attenuation is adjusted & corresponding reading is recorded.

$$G_D = \frac{4\pi r}{\lambda}\sqrt{\frac{Pr}{Pt}}$$

Where P_t - Transmitting power.

P_r - Receiving power.

6.Explain about different types of Polarization measurement? (8M)

[CO4-L2-April/May 2014, May/June 2015]

<u>Measurement of Polarization</u>:

- The polarization of electromagnetic field may be measured at one frequency as a function of the space angles. (i.e.) polar angle θ and azimuth angle ϕ or at one angular position as a function of frequency.
- The polarization of the radiated field from an antenna is measured by measuring the received signal voltage with a linearly polarized receiving antenna, since its polarization is rotated in direction through 2π or 360'.
- The maxima and minima both will be 180' apart from each other or nulls will be 90' apart from the maxima. The direction of nulls are measured accurately than maxima.
- If instead of nulls, maxima or minima are obtained the fields is said to be elliptically polarized.
- The ratio of maximum to the minimum field intensity is known as polarization ratio or ellipticity.
- When the polarization ratio is measured in the axis of the main beam of test antenna then it is called as axial ratio (AR).
- If the field is constant as the polarization of secondary antenna is rotated, then the field is said to be circularly polarized. It is a special case of elliptical polarization having AR=1.
- Linearly polarized antennas such as half wave dipole or dipole array or dipole with reflectors may be used for polarization measurement.

There are four methods used for measuring the polarization characteristics of a electromagnetic wave,

1. **Polarization- pattern method:** A linearly polarized antenna is used to measure a polarization pattern and two circularly polarized antennas are used to determine the hand of rotation.

2. **Linear- component method**: Two perpendicular linearly polarized antennas are used to measure the linearly polarized components of the wave and also their phase difference.

3. **Circular-component method:** Two circularly polarized antennas are used to measure the circularly polarized components of the wave of opposite hand and the phase angle between them.

4. **Power measurement (without phase) method**: Some waves may consist of the superposition of a large number of statistically independent waves of various polarization.
- In ordinary communication the waves are usually completely polarized but in radio astronomy the waves from celestial sources, in general partially polarized and in many cases completely unpolarized.
- The polarization characteristic of wave may be completely determined without any phase measurements by noting the power response of six antennas: 1. Vertically polarized (VP), 1 horizontally polarized (HP), 1 linearly polarized (LP) at a slant angle of + 45' and 2 circularly polarized (CP) antennas, one right- circularly polarized (RCP) and the other left- circularly polarized(LCP).

1. Polarization- pattern method:

- In this method a rotatable linearly polarized antenna such as the λ/2 dipole antenna in figure, is connected to a receiver calibrated to read relative voltage.
- Let the wave be approaching and then as the antenna is rotated in the plane of the page, the voltage observed at each position is proportional to maximum component of E in the direction of the antenna.
- Thus the tip of the electric vector E describes the polarization ellipse as shown in figure b. (dashed curve) and the variation measured with a linearly polarized receiving antenna is given by the polarization pattern in figure b. (solid line).

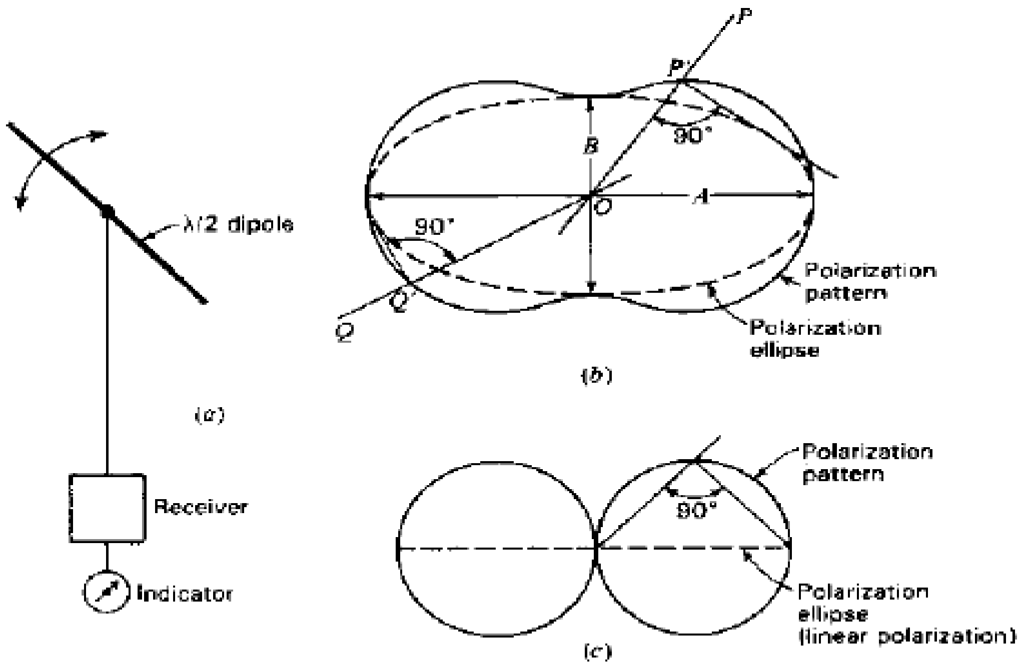
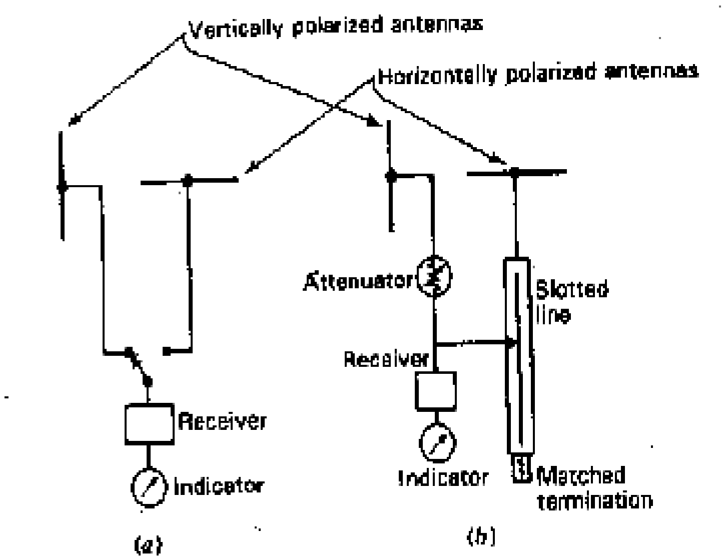


Figure (*a*) Schematic arrangement for measuring wave polarization by the *polarization-pattern method*. (*b*) Measured polarization pattern and polarization ellipse for elliptical polarization. (*c*) Measured polarization pattern for linear polarization with polarization ellipse collapsed to a line.

- In figure this is the length OP'. If the linearly polarized antenna orientation is OQ, the response is proportional to the length OQ'.
- For the case of linear polarization, the polarization ellipse degenerates to a straight line and the corresponding polarization pattern is a figure-of-eight as indicated in figure c.
- Thus by this method the polarization ellipse can be drawn and the rotation direction indicated.
- Although such a diagram completely describes the polarization characteristics of a wave, it is simpler to measure the maximum amplitude A/2 and the minimum amplitude B/2 and take the ratio of two amplitudes which is the axial ratio of the polarization ellipse or simply the axial ratio (AR).
- The axial ratio is expressed so that it is equal to or greater than unity.
- The axial ratio of the polarization ellipse of figure b is given by,
  AR= A/B



➢ **Figure .** Schematic arrangement for measuring polarization by the *linear-component method* with vertical and horizontal components given by (*a*) and phase by (*b*).

<u>B) Linear- Component method:</u>

- ➢ In this method two fixed linearly polarized antennas can be mounted at right angles, like the two λ/2 antennas as shown in figure a.
- ➢ The wave is normally out of the page. By connecting the receiver first to the terminals of one antenna and the other, the ratio E_2/E_1 can be measured.
- ➢ Then by connecting both antennas to a phase comparator, the angle δ can be measured.
- ➢ This may be done as in figure b, using a matched slotted line. From a knowledge of E_1,E_2 and δ the polarization ellipse can be calculated and the direction of rotation E is determined.

<u>C) Circular-component method</u>:

- ➢ In this method 2 circularly polarized antennas of opposite hand are connected successively to the receiver and the amplitudes E_L and E_R of circularly polarized component are measured.
- ➢ The antennas can consist of two long monofilar axial-mode helical antennas, one wound left-handed and the other wound right- handed as in figure.
- ➢ The left- circular component E_L of the wave is measured with the switch to the left as in figure so that the receiver is connected to the left-handed helix.
- ➢ The right- circular component E_R of the wave is measured with the switch thrown to the right so that the receiver is connected to the right-handed helix.

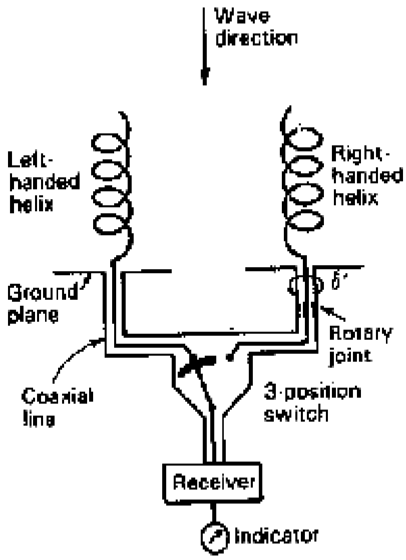


Figure Arrangement for measuring polarization by the *circular-component method*. Left- and right-handed components are measured by individual helices and phase angle by rotating one helix with both connected.

- The axial ratio of the received wave is then given by,

$$AR = \frac{E_R + E_L}{E_R - E_L}$$

- For positive values of AR the wave is right- elliptical and for negative values it is left-elliptical.
- Thus three measurements E_L , E_R and δ' with the helical antennas determine the polarization characteristics of the received wave completely.
- The accuracy depends on the circularity of polarization of the helices. This is improved (AR nearer unity) by making the helices long since

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

Where n= number of turns of the helix.

7. Explain in detail about the antenna test ranges in antenna measurement?
[CO4-L2- April/May 2013, May/June 2016]

<u>ANTENNA RANGES:</u>

- The measurements on antenna are carried out in the antenna ranges that are well equipped for the testing and evaluation of antenna systems.
- For measurement of certain antenna, the choice of the best suitable range types depends on physical size of antenna and the frequency of antenna.

- A typical antenna range includes a source antenna to illuminate the AUT, posititoners to support and orient antennas, the electronic instrumentation along with interface and physical space between and around the antennas.
- For far field measurements, the antenna to be tested (AUT) is operated in the receiving mode and the antenna whose radiation pattern and polarization is known may be used as source antenna for transmission.

The antenna measurements following types of antenna ranges are practically used:

Transverse electromagnetic cell(TEM Cell)
It is a rectangular coaxial transmission line resembling a stripline and it allows a transmission of TEM waves only.

Anechoic chamber:
It is indoor type of antenna range in which chamber walls, celing, floor are covered with energy absorbers. It is best suited for small antennas.

GTEM Cell (Gega Hertz TEM Cell): This type of antenna range is hybrid between TEM cell and anechoic chamber and can be used for wide range of frequencies.

Near field range: It is a small indoor type of range in which only near field measurements are made. Then using numerical methods, far field measurements can be determined from near field measurements.

Compact range: In this range, the transmitting antenna feeds paraboloid and the paraboloid changes the spherical waves into plane waves.

Reflected range: In this the heights of transmitting antenna and AUT are selected such that a constructive interference at AUT.

Slant range: In this type AUT is placed on a tower at fixed heights while the transmitting antenna on ground.

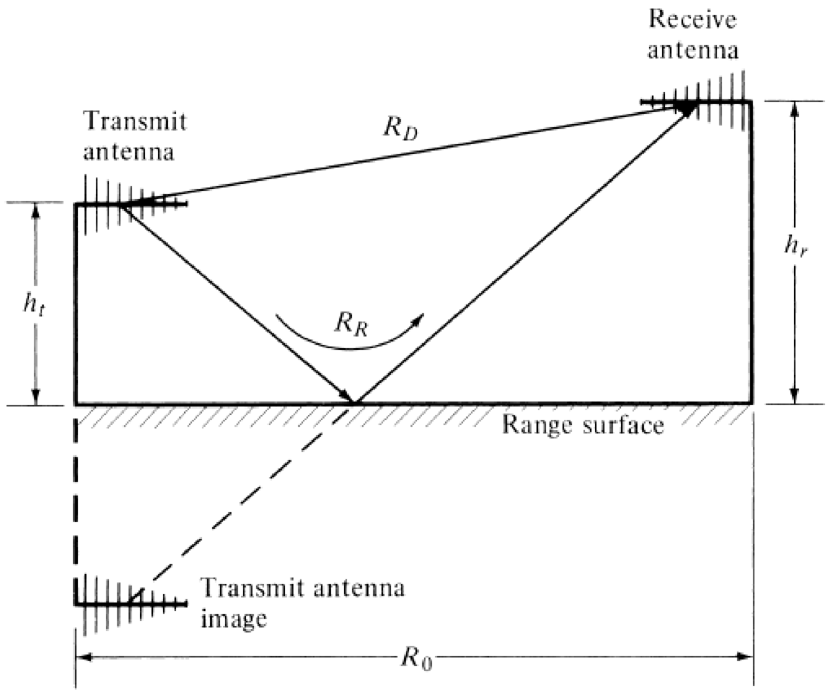
Ground range: This is a range in which tall towers are not needed as the transmitting antenna is placed above surface and acts like mirror.

OUT DOOR RANGES:

There are three basic types of outdoor ranges as follows:

- Reflection ranges
- Elevated or free space ranges
- Slant ranges

1. **Reflection ranges:**
   - These ranges are of the outdoor type, where the ground acts as a reflecting surface.
   - Typically the reflection ranges are used in UHF region for the measurement of patterns of moderately broad antennas.
   - The reflection range arrangement is shown below:



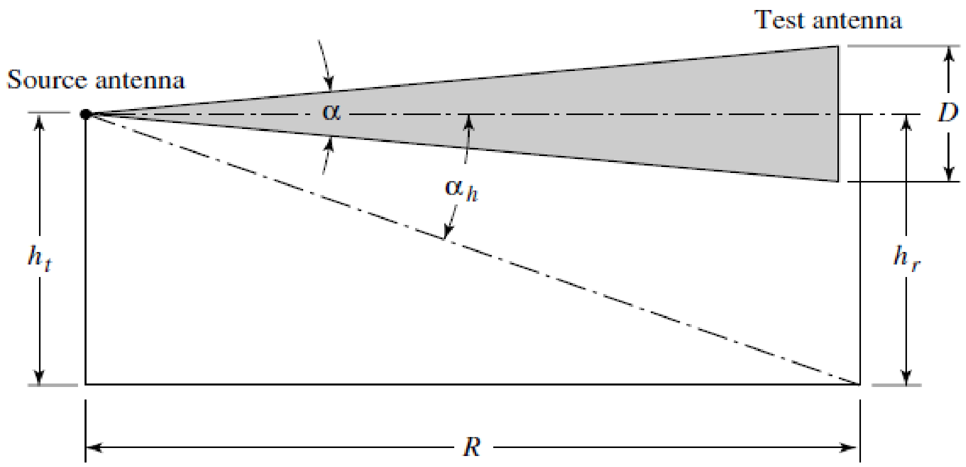
Reflection range geometrical arrangement

   - The reflection range is designed such that a constructive interference is created in the region of AUT which is called quite zone.
   - As shown in figure, the interference pattern is achieved by designing the range in such a way that the specular reflections from the ground combine constructively with the direct waves.

- The illuminating field may be given a small and symmetric amplitude taper by constructively with the direct waves.
- The reflection range is useful in measurement of antenna in receiving mode and used for antenna gain measurement.

2.Elevated or free space ranges:

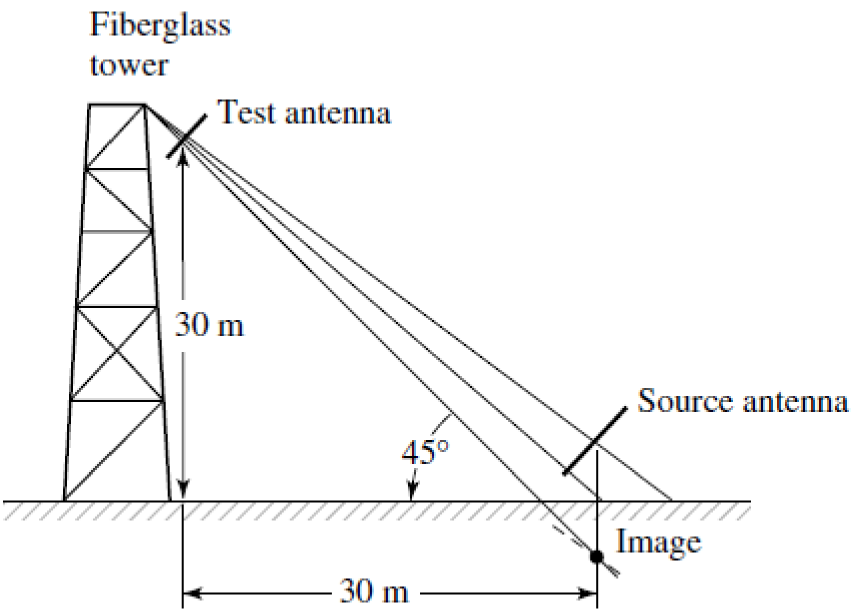
- Elevated or free space ranges are designed to stimulate near-free space conditions for antenna measurements. Such ranges suppress the effects of surroundings.
- The free space conditions are attempted on the choice of the range length, heights of transmitting and receiving antennas and the radiation characteristics of receiving antennas.
- The geometry of elevated range is shown below:
- In the elevated range, the range length R is determined by the maximum allowable phase taper of incident field over the aperture of test antenna.
- Typically the dimension chosen lies between $2D^2/\lambda$ to $12D^2/\lambda$ where D is maximum dimension of antennas.
- The AUT should be located at an appropriate height so that the main beam of the source antenna does not illuminate the range surface.
- Thus the height of AUT h_r should be atleast 4 times greater than the maximum dimension D (i.e) $h_r \geq 4D$.



- If the range surface is irregular or there are objects which cannot be removed from the site, then the Fresnel zones are achieved over a topographic map of the range surface and then the specular regions are identified and located such that from these regions significant reflections occur.
- If the range surface is flat, then the diffraction fences are employed to redirect the reflected signals which are moved away from the test regions.
- Along with this the edges of fence should not be straight but serrated. The serrated fence reduces the effect of diffraction causing increase in the indirect signal.

3.Slant range:

- It is a type of range in which the source antenna is placed close to the ground and AUT is placed at top of tower along with its positioned.
- The orientation of two antennas is such that the beam axis of the source antenna points towards the centre of test antenna.
- The arrangement of slant range is shown below:



- For the measurements on highly directive antenna only the main beam and few side lobes of antenna are required.
- This arrangement is useful for some antennas such as satellite antennas which are required to be protected by a redome housing from contaminations.

b) INDOOR RANGES:

- Due to the environmental changes, adverse weather, electromagnetic interference there is a practical limitation on the measurement made in the outdoors.
- To overcome these limitations, the indoor ranges are used effectively. Such indoor ranges provide relatively better environment than outdoor ranges.

- The indoor ranges make use of microwave absorbing material to avoid unwanted reflections. Such an absorbing material keep the enclosures free from any echoes.

- The room completely lined with such absorbing materials is called anechoic chamber (no echo chamber).

1) ANECHONIC CHAMBER MEASUREMENT:

- An anechoic chamber is a reflection less room specially designed to perform all whether antenna measurement in a controlled laboratory environment.

- It is an indoor chamber.

- Chamber walls, ceiling & floor are filled with RF energy absorbs except at the location of transmitting antenna & antenna under test.

- Anechoic chamber can be used for far field measurement of small antenna.

- Absorbs are used for reducing side lobe & back lobe radiation.

- Absorbers used are,
  Pyramids: Works best at normal incident angle.

  Widely used.

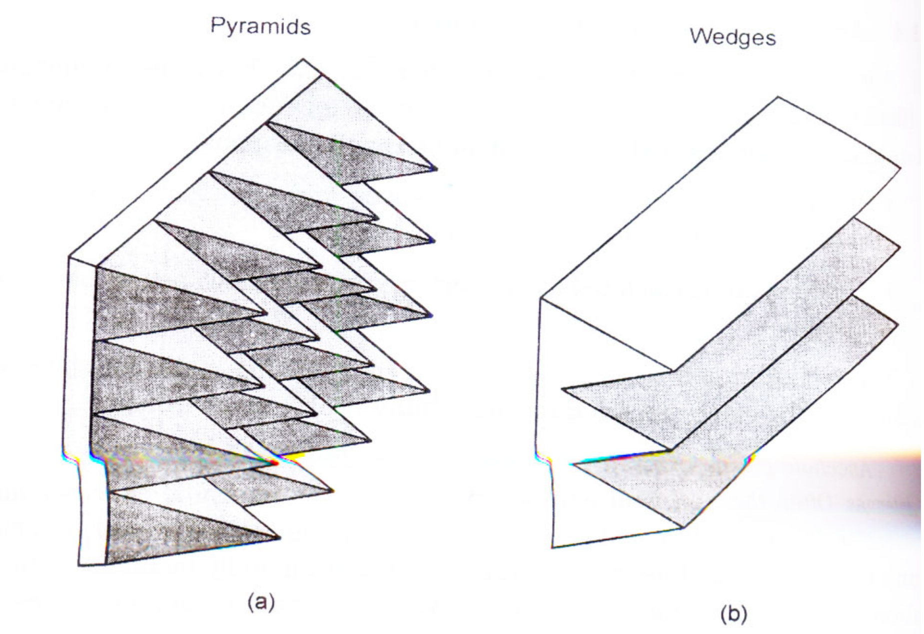
  Wedges: Works best at large incident angle.

ANECHOIC CHAMBER TYPES:

Anechoic chamber are usually two types,

(i)Rectangular anechoic chamber.

(ii)Tapered anechoic chamber.



(i) Rectangular anechoic chamber:

- Fig.(a)shows the rectangular chamber. Here the end walls and the center parts of the sidewalls, floor and ceiling are covered with pyramids.
- The other parts are covered with wedges. the antenna are placed on the middle line of the chamber; the source antenna close to one end wall, the AUT a little further away from the end wall.
- The test zone where the reflections are minimized is called the quiet zone.
- The dimensions of the chamber should be such that the angle of incidence on side walls is less than 60˚.At larger angles, the reflections would be large.

- Typically the length to width ratio is 2:1.The source antenna should be chosen so that its main beam doesn't illuminate the side walls, ceiling and floor.

(ii)Tapered anechoic chamber:

- At frequency below 1 GHz, the rectangular chambers having absorbers of reasonable size has a high level of reflections.
- Therefore a tapered is chosen at this frequency.
- The source antenna is close to apex of the tapered section and specular reflections occur close to the source.
- The phase difference of the direct wave and the specular reflections changes slowly in the quiet zone which result in a more planar wave front than in the case of rectangular chamber.
- At higher frequencies, the source is moved from apex closer to the rectangular section and the chamber is used as a normal rectangular chamber.

TEST EQUIPMENT OUTSIDE THE CHAMBER:

- Outside the chamber, a network analyser measures the power received by the AUT. The motor controls are used to set the Azimuth and roll of positioned . A PC workstation runs the software that collects and stores the data.

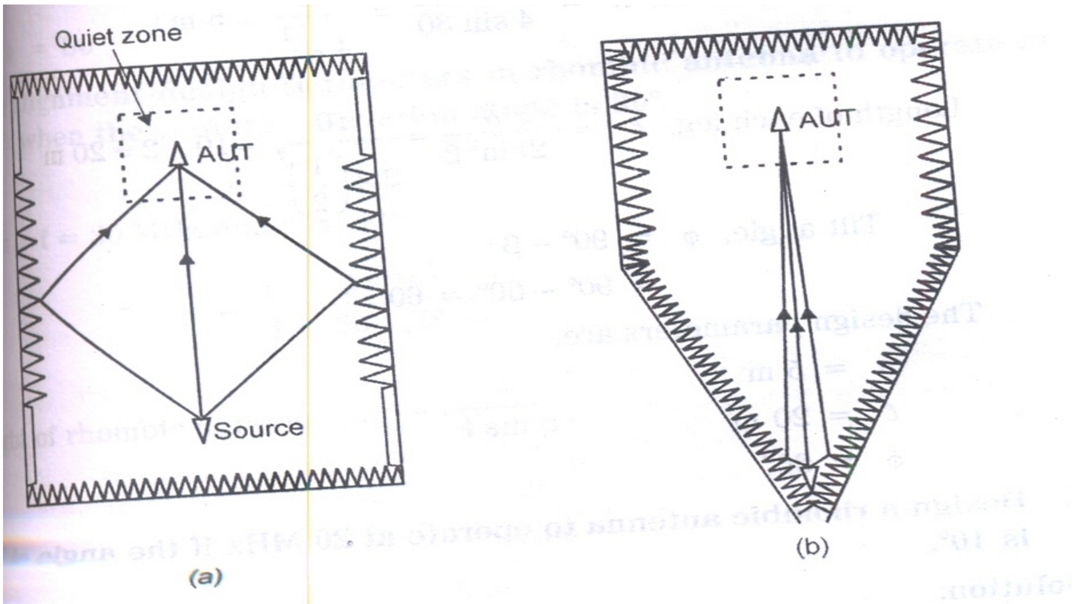
Advantages of anechoic chamber:

- It stimulates a reflection less free space and allows all-weather antenna measurements in a controlled environment.
- The test area is isolated from interfering signals much better than at outdoor ranges.

DRAWBACKS OF ANECHOIC CHAMBER:

- The size of the chamber is large since the antennas need to be several wavelength apart to simulate far field conditions.
- Cost is also high because of its large size.
- The RF absorbing material typically works well at and above UHF ranges. Therefore the anechoic chambers are most often used for frequencies above 300MHz.

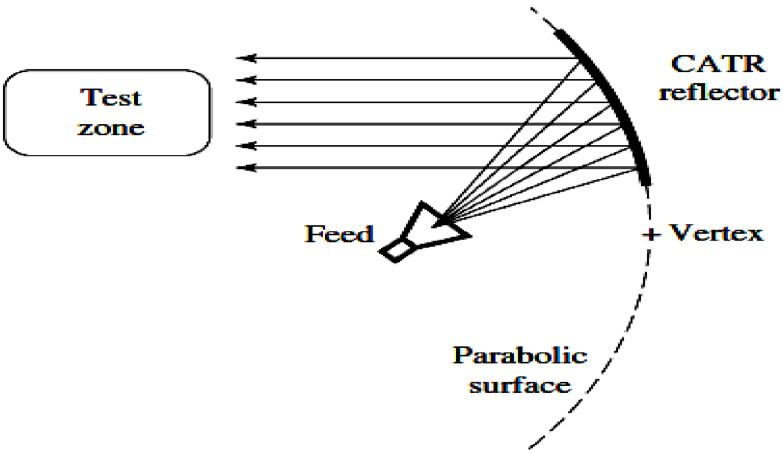
- If the chamber is large ,then the source antenna's mainlobe is not in view of the side walls , ceiling of the floor



(a) Rectangular (b) Tapered

2. COMPACT RANGES:

- Measurement on microwave antenna often require that the radiator under test be illuminated by a uniform plane wave which is usually achievable only in the far field regions and involves large distances.
- Here the source antenna is used as an offset feed that illuminates a paraboloid reflector.



A Compact Antenna Test Range (CATR)

- The paraboloidal reflector acts as a spherical to plane wave transformer and produces a plane wave in the test zone of an indoor range.
- The advantage is that the parabolic reflector moves the far field region in very close, making the compact range equivalent to a much larger conventional range.
- For proper coverage the linear dimension of the reflector are three to four times greater than those of test antenna.

8.Explain in detail about VSWR Measurement? (8M)

[CO4-L2- Nov/Dec 2014, April/May 2015]

<u>VSWR measurement:</u>

<u>SWR or slotted line method:</u>

This method is based on the characteristics of travelling wave via input impedance may be uniquely determined from the knowledge of voltage or current minimum and the reference point at which the impedance is measured.

The antenna impedance is given by,

$$Z1 = Z0\left(\frac{1+K}{1-K}\right)$$

$$VSWR = \left|\frac{V_{max}}{V_{min}}\right| = \frac{|V_i|+|V_r|}{|V_i|-|V_r|} = \frac{1+\left|\frac{V_r}{V_i}\right|}{1-\left|\frac{V_r}{V_i}\right|} = \frac{1+|k|}{1-|k|}$$

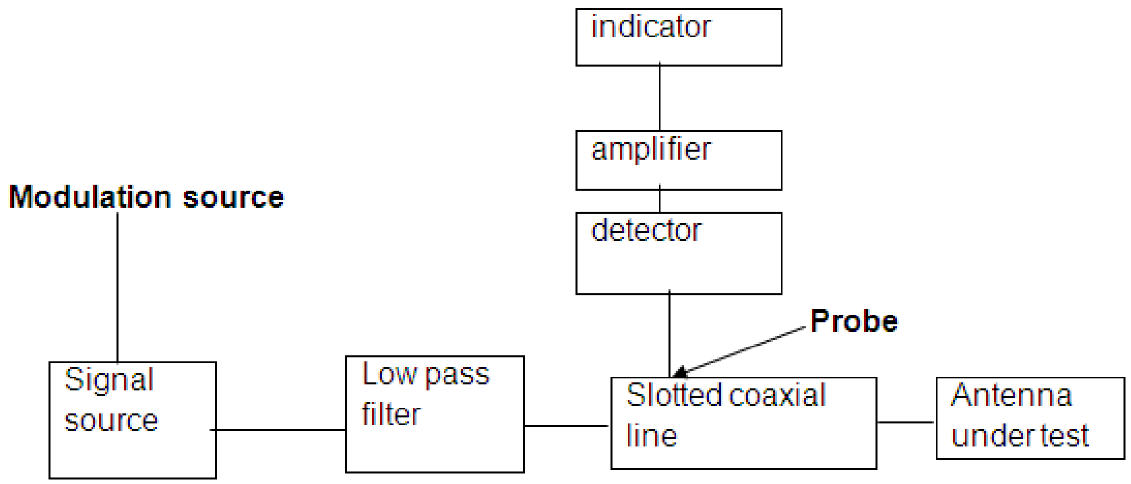
Reflection coefficient k= $\left|\frac{V_r}{V_i}\right|$

Alternatively k= $\frac{S-1}{S+1}$; S is VSWR

And Ɵ = π - 2β d = π $\left[1-\frac{4d}{\lambda}\right]$ as β= $\frac{2\pi}{\lambda}$

= 180 $\left[1-\frac{4d}{\lambda}\right]$ degrees.

Block diagram of slotted line method:



- The slotted line arrangement consists of length of transmission line or coaxial cable with an axial slot, along which moves a travelling carriage carrying a probe.
- The probe projects through the slot. The voltage measuring device may be a crystal detector and a micro ammeter.
- A signal source is connected to the left and the right end is connected to unknown impedance being measured.

Voltage Standing Wave Ratio:

It is the ratio between maximum voltage and a minimum voltage corresponding to the waves.

$$V_{SWR}=\frac{Vmax}{Vmin}$$

$$V_{SNR}=(1+|e|) / (1-|e|)$$

e-> reflection co efficient

Condition for perfect match:

The condition=1:1

VSWR value should always be low .in case of high VSWR, and inference will be produced.

Measurement of VSWR:

- reflected power.
- reflection co efficient

Reflected power:

- It is the property of incident power which is reflected back to the transmitter by the mismatch antenna.
- It is determined by the reflection co efficient. E

Reflection co efficient:

$E=(Z_L-Zo) / (Z_L+Zo)$

Zo-> characteristic impedance.

Z_L-> load impedance.

Ideal condition:

In equation 1, when $Z_L=Z_{0,}$ there is no reflected signal.

Under this condition, all the powers are accepted by the antenna.

VSWR Measurement:

Case(i):

VSWR(S<10)

- This VSWR measurement is one of the important measurement in Mw.

We can determine the impedance at any point in a waveguide.

- VSWR<=10 are very easily measured with this detector.
- Variable attenuator is tuned to get a maximum reading on the meter.
- Note down the reading ,that value is represented to be Vmax. The probe of the slotted line is adjusted to get the minimum reading in the meter. Hereby , we calculate the VSWR as

$$VSWR=\frac{Vmax}{Vmin}$$

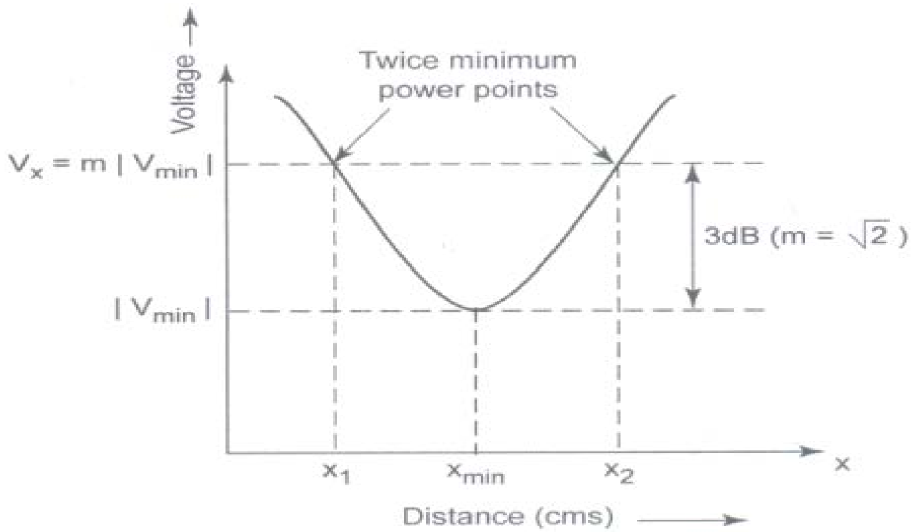
VSWR ranges from 1 to ∝

Limitations:

- signal source should generate the signal frequency.
- Reflections should not be there.
- Depth of penetration of probe into slotted waveguide introduces error.
- When VSWR is nearer to the value of 10,Vmin will be small.

Case 2:

VSWR(S>10)



To methods are used in this condition,

- Double minimum method
- Calibrated attenuator method.

Double minimum method:

Probe depth does not introduce an error,

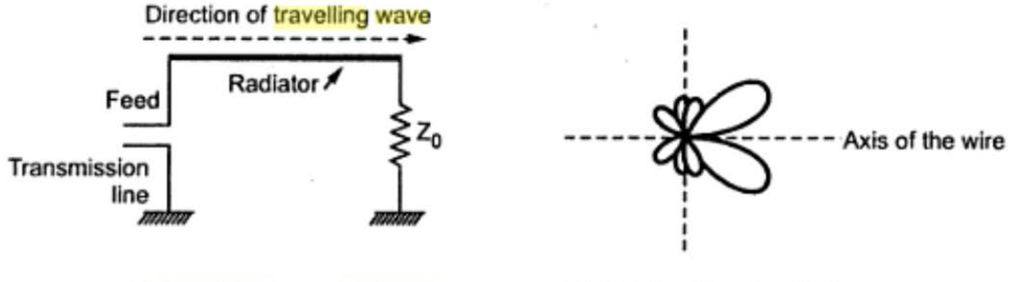
- When a probe depth is inserted to calculate a maximum value.
- The probe depth is inserted at a minimum point that can be represents in a graph as E_{min}.

- Insert the probe depth, where the power is twice the minimum point. that can be repress by dB in graph.
- The probe depth is then moved to twice power point on the other side of minimum that Point. That can be represented as d2 m the graph.
- And from the calculate d_1 &d_2 values, the VSWR is calculated.

9.Explain in detail about the Radiation from a travelling wave antenna? (8M)
[CO4-L2-April/May 2012]

Radiation from a travelling wave antenna

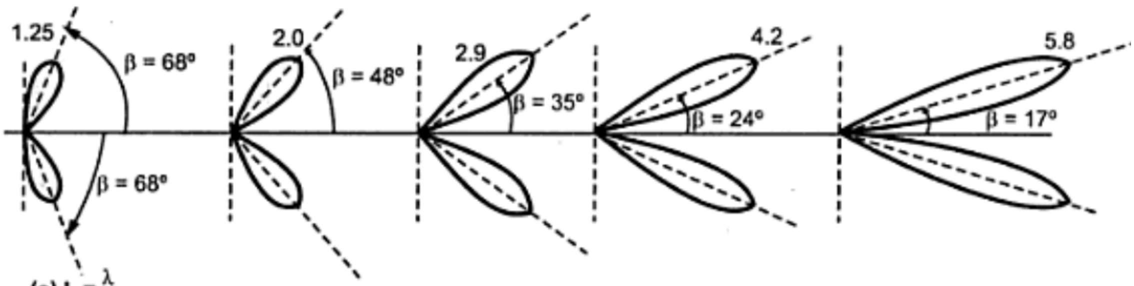
- The antenna in which the standing wave does not exist along the length of the antenna is called travelling wave antenna. In general the standing waves are produced, when the line is not properly terminated which causes reflections at the output or load side.
- The standing waves travel due to reflections in the resonant antenna. But in the travelling wave antenna, the standing waves do not exist. That means, the travelling wave antenna is non-resonant type antenna or aperiodic antenna.
- In case of the radio communication involving ionospheric reflections, the frequency spectrum required is wide. Thus the antenna used for the radio communication should have larger bandwidth
- In such antennas one of the ends is terminated into the characteristic impedance Z_o while other end is connected to the input signal.
- Due to the proper termination at the load, the reflections are avoided. Because of this the unidirectional radiation pattern is obtained as shown in the Fig.



(a) Travelling wave antenna arrangement (b) Unidirectional radiation pattern

- The angle of major lobe and the amplitude of the major lobe depends on the length of the wire. As the length of the wire increases, the angle of major lobe with respect to the axis of wire decreases.
- Hence the major lobe comes closer to the axis of wire as the length increases. Also with the increase in the length, the amplitude increases.
- The Table given below indicates different values of angle of major lobe and amplitude of the lobes for the increasing length.

Length of the travelling wave	Angle of Major Lobe (β)	Amplitude of Major lobe
$L = \frac{\lambda}{2}$	68°	1.25
$L = \lambda$	48°	2.0
$L = 2\lambda$	35°	2.9
$L = 4\lambda$	24°	4.2
$L = 8\lambda$	17°	5.8



(a) $L = \frac{\lambda}{2}$ (b) $L = \lambda$ (c) $L = 2\lambda$ (d) $L = 4\lambda$ (e) $L = 8\lambda$

Advantages:

- Standing waves does not exist.
- Bandwidth is more.
- With increasing length, the major lobes becomes narrower and closer.

Disadvantages:

- Large space requirement and not useful at higher frequencies.

Unit-V

Propagation of radio waves

Part-A

1. **What is the sporadic E layer in Ionosphere? [CO5-L1]**

   The sporadic E layer is an anomalous ionization layer in ionosphere. It usually occurs in the form of clouds, varying in size from 1 km to several hundred km across.

2. **What is the critical frequency of radio wave for reflection at vertical incidence if the maximum value of electron density in $1.26\times10^{6}cm^{-3}$? [CO5-H3-April/May2013]**

   $Fc = 9 \sqrt{Nm}$
   $= 9 \sqrt{1.26*10^{6}\ cm^{-3}}$
   $= 10.1$ MHz

3. **A pulse of a given frequency transmitted upward is received back after a period of 5ms. Find virtual height? [CO5-L1]**

   h' = CT/2
   $= (3 \times 10^{8} \text{ x } 5\times10^{-3}) /2$
   = 750Km

4. **Define skip distance. [CO5-L1- May/June 2012]**

   The distance with in which a signal of given frequency fails to be reflected back is the skip distance for that frequency. The higher the frequency the greater the skip distance.

5. **What is duct propagation? [CO5-L1- May/June 2013]**

   Due to the presence of moisture, the EM rays are curved along the earth's surface having the same curvature as that of earth. Under such conditions, wave tends to get trapped or guided along a duct making the atmosphere to behave as a leaky wave guide. With the duct, it is possible for line of sight, sometimes up to 400 miles. This is known as duct propagation.

6. **What are the factors that affect the propagation of radio waves? [CO5-L2]**
   - Curvature of earth.
   - Earth's magnetic field.
   - Frequency of the signal.
   - Plane earth reflection.

7. A VHF communication link is to which the line of sight communication may be possible. If the heights of transmitting and receiving antennas are 40, 25 meters respectively. Find the maximum distance through which signal could be received? [CO5-H3- April/May 2014]

Solution:

h_t= 40m , h_r=25m

$d_{max} = 4.12 [\sqrt{ht} + \sqrt{hr}]$

$= 4.12 [\sqrt{40} + \sqrt{25}]$

= 46.6km

8. What are the effects of earth curvature on troposphere propagation? [CO5-L1]

For any layer, the highest frequency that will be reflected back for vertical incidence is $f_{cr} = 90^{\circ}$ max

9. Define gyro frequency. [CO5-L1]

Frequency whose period is equal to the period of an electron in its orbit under the influence of the earth's magnetic flux density B.

10. What is multihop propagation? [CO5-L1]

The coverage of transmission distance between transmitter and receiver in more than one hop (jump) is known as multihop propagation.

11. State Faraday's law. Or what is meant by faraday rotation? [CO5-L2- April/May 2011, Nov/Dec 2011]

Any linearly polarized wave may be considered as the vector sum of two anti rotating circularly polarized wave. If such a wave propagates in the direction of magnetic field, then the two circularly polarized components will travel at different phase velocities and thus the plane of polarization will rotate with distance. This phenomenon is known as faraday's rotation.

12. Write the limitations of ground wave propagation. [CO5-L2]

It does not travel over the long distance and also travel only low frequency range. If frequency increases attenuation increases.

13. State secant law. [CO5-L2]

The maximum usable frequency is equal to the product of critical frequency and secant of incidence angle.

$f_{muf} = f_{cr}$ seci

14. Define Fading. List various types of fading? [CO5-L1- April/May 2013]

It is defined as the fluctuations in the received signal strength caused due to variations in height and density of the ionization in different layers. Types of fading are,

(i) Selective fading (ii) interference fading (iii) Absorption fading (iv) polarization fading

15. Which layer of ionosphere is called as kernelly Heaviside layer & application layer? [CO5-L1]

E layer.

16. What is super refraction? [CO5-L1]

The higher frequencies or microwaves are continuously reflected act in the duct and reflected by the ground. So that they propagate around the curvature for beyond the line of sight. This special refraction of electromagnetic waves is called super refraction.

17. Define MUF. [CO5-L1- April/May 2012, May/June 2015]

The maximum frequency that can be reflected back for a given distance of transmission is called the maximum usable frequency for that distance. It seen that the MUF is related to the critical frequency and the angle of incident by the simple expression.

$MUF = f_{cr} \sec \phi_I$

18. Define actual height and virtual height.
[CO5-L1- April/May 2012, April May 2014]

The height at a point above the surface at which the wave bends down to the earth is called actual height.

Virtual height of an ionosphere layer may be defined as the height to which a short pulse of energy sent vertically upward and travelling with the speed of light would reach taking the same two ways travel time as done the actual pulse reflected from the layer.

$h' = CT/2$

19. Define critical frequency.
[CO5-L1- Nov/Dec 2011, April/May2012, April/may 2013]

The critical frequency of an ionized layer is defined as the highest frequency which can be reflected by a particular layer at vertical incidence.

It is different for different layers. It is otherwise called as plasma frequency.

Fc= 9 $\sqrt{N\ max}$

Nmax-maximum electron density

20. Calculate MUF for a critical frequency 10 MHz and an angle of incidence 45°

[CO5-H3- April/May 2015, May/June 2016]

Ans:

$f_{muf} = f_{cr}$ (sec i)

$= 10 \times 10^6 (\sec 45^\circ) = (1.4142)*(10 \times 10^6) = 14.142$ MHz

21. What is the critical frequency for reflection at vertical incidence if the maximum value of electron density is 1.24 *$10^6 cm^{-3}$?

[CO5-H3- Nov/Dec 2012, April/May 2013, April/May 2014]

Ans: N_{max} = 1.24 *$10^6 cm^{-3}$ **=** 1.24 *$10^6 * 10^{-6} m^{-3}$ = 1.24 m^{-3}

f_{cr} = $\sqrt{81\ Nmax} = \sqrt{81\ (1.24)}$ = 10.022 MHZ

22. Find the maximum distance that can be covered by a space wave, when the antenna heights are 60 m and 120 m **[CO5-H1- May/June- 2013]**

Ans: Let the heights of transmitting and receiving antenna are 60m and 120m respectively. Then the maximum distance in km covered by a space wave is given by

d_{max} = 4.12 [$\sqrt{ht} + \sqrt{hr}$] = 4.12 [$\sqrt{60} + \sqrt{120}$] = 77.045 km

23.What do you mean by magneto ionic spiltting? **[CO5-L1]**

The earth's magnetic field splits up the incident waves into 2 different components. They are ordinary wave and extra ordinary wave. This phenomenon of splitting of wave into 2 different components is called as magneto ionic splitting.

24. How can we minimize fading? **[CO5-L1- May/June- 2013]**

The most common method to minimize fading is diversity reception. In this method, some part of the signal is duplicated and even if one part experiences a deep fades, the other may not.

Part-B

1. Briefly describe the terms related to the sky wave propagation: virtual heights, critical frequency, maximum usable frequency, skip distance and fading? (16M) [CO5-L1- May/June 2012, April/May 2014, April/May 2015]

Sky wave propagation:

It is also called as Ionosphere wave propagation. The ionosphere acts like a reflecting surface and is able to reflect back the electromagnetic waves of frequencies between 2 MHz to 30MHz .Since, long distance point to point communication is possible with sky propagation, it is also called as point to point propagation. This mode of propagation is also known as short wave propagation.

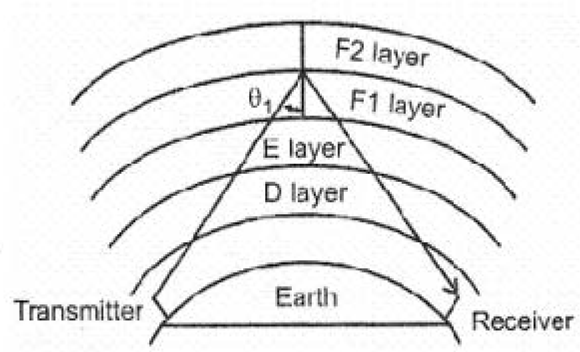


fig 1.1 Sky Wave Propagation

Virtual heights: The virtual height (h) has the great advantage of being easily measured, and it is very useful in transmission path calculations. For fiat earth approximation and assuming that ionosphere conditions are symmetrical for incident and refracted waves, The transmission path distance,

TR=2h/tan β

Where β=Angle of elevation h =Virtual height

Critical frequency: When the refractive index, n has decreased to the point where n = sin φi the angle of refraction φ will be 90° and wave will be travelling horizontally. The higher point reached by the wave is free. The electron density N at the that point satisfies the relation

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

(or)

$$N'=\frac{f^2\cos^2\phi_i}{81}$$

If the electron density at some level in a layer is sufficient great to satisfy the above condition,then the wave will be returned to earth from that level. If maximum electron density in a layer is less than n', the wave will penetrate the layer (Though it may be reflected back from a higher layer for which N is greater). The largest electron density required for reflection occurs when the angle of incident φi is zero, i.e., for vertical incidence. For any given layer the highest frequency that will be reflected back for vertical incidence will be

$$f_{Cr}=\sqrt{81N\max}$$

Where fcr = Critical frequency for the layer Nmax = Maximum ionization density (electrons per cubic meter). The characteristics of the ionospheric layers are usually described in terms of their virtual heights and critical frequencies, as these quantities can be readily measured. The virtual height is the height that would be reached by a short pulse of energy showing the same time delay as the actual pulse reflected from the layer travelling with the speed of light. The virtual height is always greater than the true height of reflection, because the interchange of energy taking place between the wave and electrons of the ionosphere causes the velocity of propagation to be reduced. The extent of this difference is influenced, by the electron distributions in the regions below the level of reflection. It is usually very small, but on occasions may be as large as 100 Kms or so.

The critical frequency is the highest frequency that is returned by a layer at vertical incidence. For regular layers,

$$f_c=\sqrt{\text{max electron density in the layer}}$$

i.e.

$$f_c=\sqrt{Ne}$$

The critical frequencies of the E and F1 layers primarily depend on the zenith angle of the sun. It, therefore, follows a regular diurnal cycle, being maximum at noon and tapering off an either side. The fc of the F2 layer, shows much larger seasonal variation and also changes more from day to day. It can be seen that the critical frequencies of the regular layers decrease greatly during night as a result of recombination in the absence of solar radiation. But the fc of sporadic E shows regular variation throughout the day and night suggesting that sporadic E is affected strongly by factors other than solar radiation. There is a long term variation in all ionospheric characteristics closely associated with the 11 year sunspot cycle. From the minimum to maximum of the cycle, fc of F2 layer varies from about 6 to 11 MHz (ratio of 1:1.8), fc of E layer varies from 3.1 to 3.8 MHz (a ratio of mere 1 to 1.2). Long term predictions of ionospheric characteristics are based on predictions of the sunspot number. Reliable estimates can be made, for as much as a year, in advance.

Maximum usable Frequency :

Although the critical frequency for any layer represents the highest frequency that will be reflected back from that layer at vertical incidence, it is not the highest frequency that can be reflected from the layer. The highest frequency that can be reflected depends also upon the angle of incidence, and hence, for a given layer height, upon the distance between the transmitting and receiving points. The maximum, frequency that can be reflected back for a given distance of transmission is called the maximum usable frequency (MUF) for that distance. It is seen that the MUF is related to the critical frequency and the angle of incidence by the simple expression

$$MUF = f_{cr} \sec\varphi_i$$

The MUF for a layer is greater than the critical frequency by the factor secφ_i the largest angle of incidence φ_i that can be obtained in F-layer reflection is of the order of 74°. This occurs for a ray that leaves the earth at the grazing angle. The geometry for this case is shown below

Where $\varphi_{imax} = \sin^{-1}(r/r+h)$

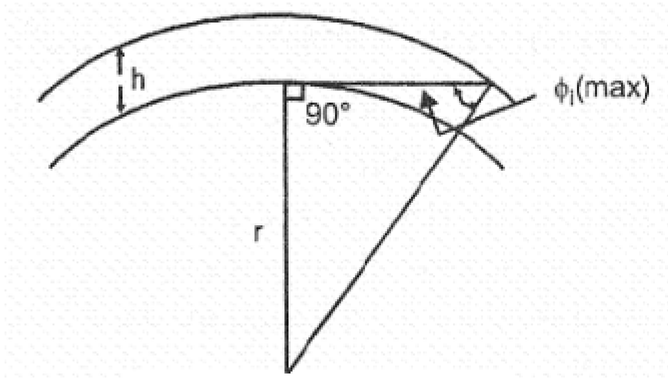


fig 1.2 Geometric of MUF

The MUF at this limiting angle is related to the critical frequency of the layer by

$$MUFmax = f_{cr}/\cos 74° = 3.6\, f_{cr}$$

2. Explain the structure of ionosphere on the surface of the earth ? (8M)

[CO5-L2- Nov/Dec 2011, April/May 2013,May/June 2016]

Structure of the ionosphere

As the medium between the transmitting and receiving antennas plays a significant role, it is essential to study the medium above the earth, through which the radio waves propagate. The various regions above the earth's surface are illustrated in Fig.2.1

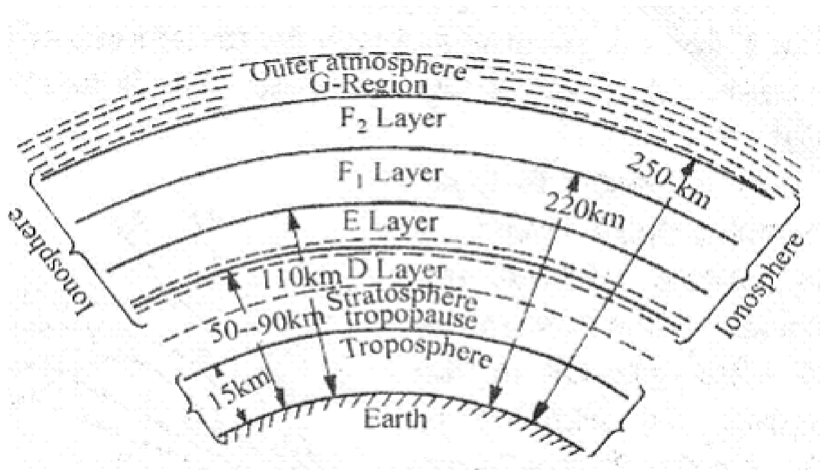


fig 2.1 Structure of ionosphere

The portion of the atmosphere, extending up to a height (average of 15 Km) of about 16 to 18 Kms from the earth's surface, at the equator is termed as troposphere or region of change. Tropopause starts at the top of the troposphere and ends at the beginning of or region of calm. Above the stratosphere, the upper stratosphere parts of the earth's atmosphere absorb large quantities of radiant energy from the sun. This not only heats up the atmosphere, but also produces some ionization in the form of free electrons, positive and negative ions. This part of the atmosphere where the ionization is appreciable, is known as the ionosphere. The most important ionizing agents are ultraviolet UV radiation, ά , β and cosmic rays and meteors. The ionization tends to be stratified due to the differences in the physical properties of the atmosphere at different heights and also because various kinds of radiation are involved.

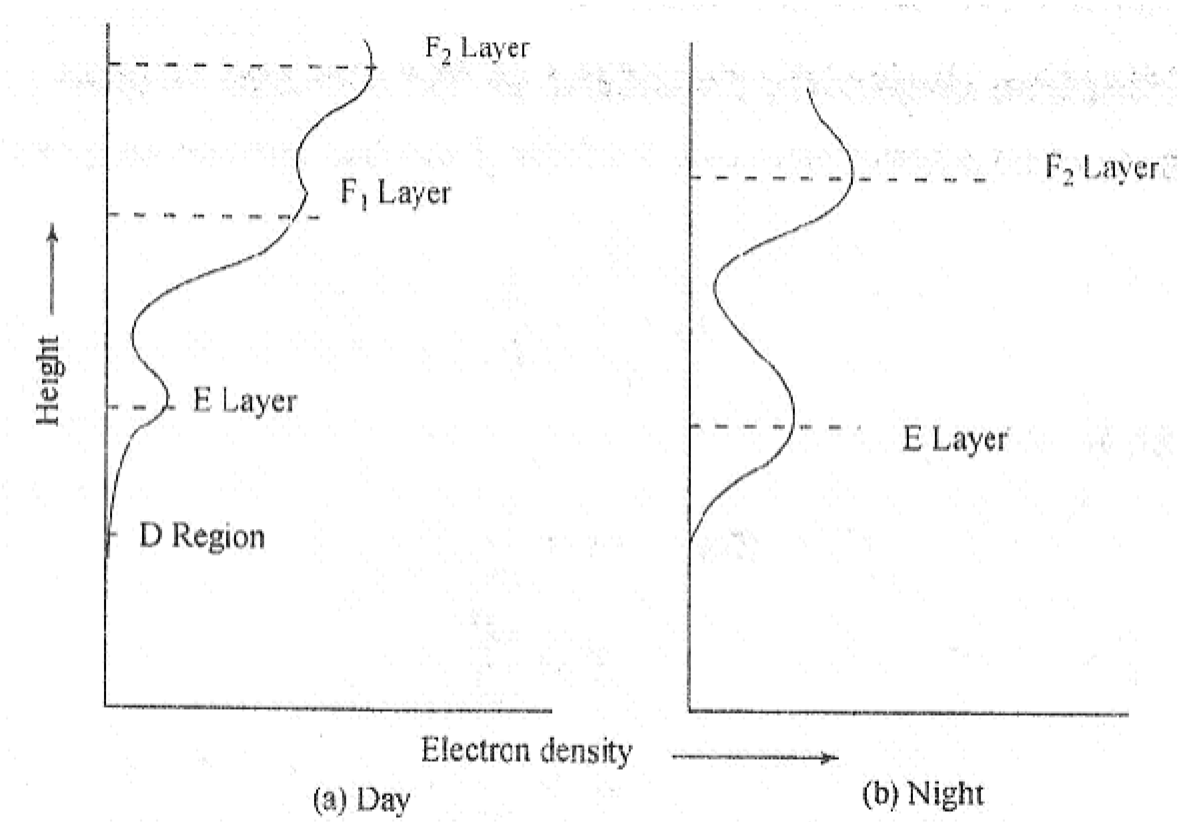


fig 2.2 Electron density of ionosphere layers

The levels, at which the electron density reaches maximum, are called as layers. The three principal day time maxima are called E, F1, and F2 layers.

In addition to these three regular layers, there is a region (below E) responsible for much of the day time attenuations of HF radio waves, called D region (ref. Fig. 4a). It lies between the heights of 50 and 90 Km (ref. Fig. 3). The heights of maximum density of regular layers E and F1are relatively constant at about 110 Km and 220Km respectively. These have little or no diurnal variation, whereas the F2 layer is more variable, with heights in the range of 250 to 350 Km.

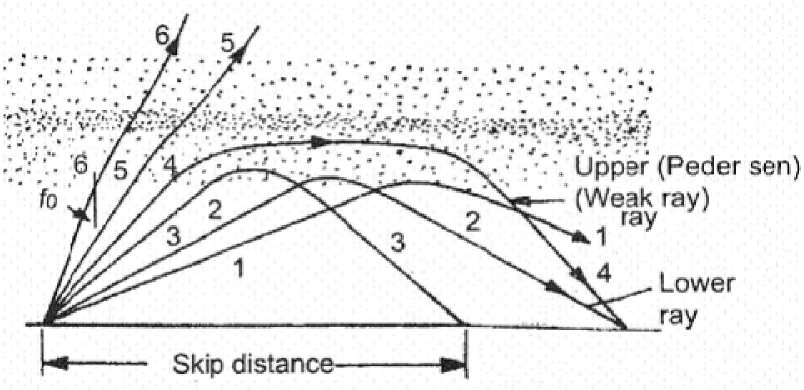


fig 2.3 Effect of ionosphere on rays

At night F1 and F2 layers combine to form a single night time F2 layer (Fig. 4b). The E layer is governed closely by the amount of UV light from the sun and at night tends to decay uniformly with time. The D layer ionization is largely absent during night A sporadic E layer is not a thick layer. It is formed without any cause. The ionization is often present in the region, in addition to the regular E ionization. Sporadic E exhibits the characteristics of a very thin layer appearing at a height of about 90 to 130 Kms. Often, it occurs in the form of clouds, varying in size from 1 Km to several 100 Kms across and its occurrence is quite unpredictable. It may be observed both day and night and its cause is still uncertain.

3 Explain the mechanism of refraction , under what circumstances do it occurs and what causes it ? (8M) [CO5-L2- May/June 2012]

We have mentioned earlier, that the path of the radio wave is bent by the ionosphere. Neglecting the effect of the earth's magnetic field and the effect of energy loss, the refractive index of the ionosphere is given by

$$n = \sqrt{\mu r * \epsilon r}$$

$$\mu_r = 1$$

$$n = \sqrt{\epsilon r} = \sqrt{1 - (81N/f\hat{}2)}.$$

This will always show the values of n < 1. Lower the frequency and higher the electron density, greater is the deviation of the Refractive Index from unity. When f 2 < 81N, n is imaginary, i.e. the ionized region is not able to transmit a wave freely at such a frequency. Instead, attenuation takes place, analogous to the action of a waveguide operating beyond cut off

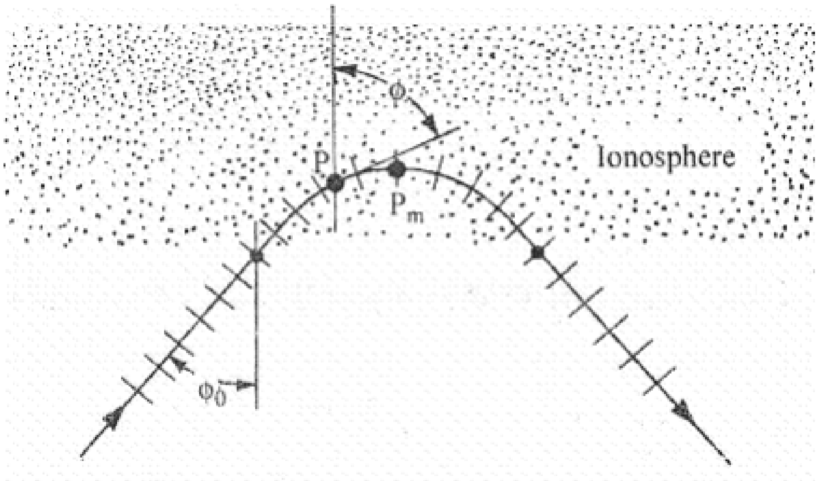


fig 3.1 Refraction of sky waves

The phase velocity of a wave travelling through the ionosphere behaves in the same way as the phase velocity of a wave on a transmission line, i.e. the velocity is inversely proportional to the square root of the dielectric constant.

$$\text{i.e. Phase velocity} = \frac{\text{Velocity of light}}{n} = \frac{c}{n}$$

since n < 1 for an ionized medium, the phase velocity in the ionosphere, is always greater than V by an amount that is greater, larger the quantity .

$$\frac{N}{f\text{^}2}$$

As a result, when a wave enters the ionosphere, the edge of the wave front in the region of the highest electron density will advance faster than the part of the waveforms encountering regions of lower electron density. Accordingly, the path of the wave is bent in the ionosphere as illustrated in Fig. 6. This bending of the wave follows ordinary optical laws. The direction, in which a wave travels at P, in the ionosphere, is given by Snell's Law.

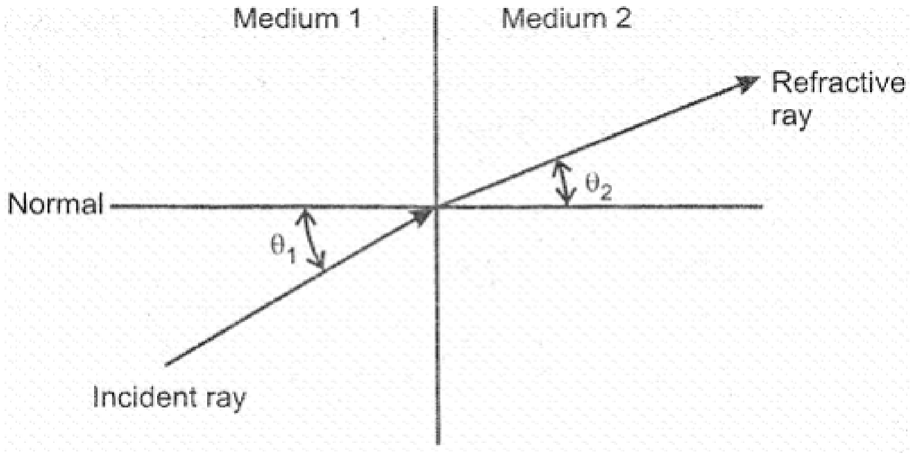


fig 3.2 Refraction

$n_1 \sin \theta_1 = n_2 \sin \theta_2$.

Where

θ1 = angle of incidence

θ2 = angle of refraction

n1 = refractive index of 1st medium

n2 =refractive index of 2nd medium

Here, it is assumed that below the ionosphere, where the direction of travel is given by φ0 ,n =1

$$\frac{\sin\theta 1}{\sin\theta 2} = \sqrt{\frac{€r1}{€r2}}$$

The top Pm of the path corresponds to φ= 90° and occurs at a point in the ionosphere where

$$n = \sin\varphi_0 .$$

Pm is commonly referred to as the point of reflection, though, actually, it is the point of refraction. Eq. (3) shows that smaller the φ0, smaller is the 'n' required to return the wave to the earth.

With vertical incidence, i.e. φ0 = 0, n must be reduced to 0 for reflection to take place. The wave then penetrates the ionized region until it reaches a point, where the electron density N and the frequency fv of the vertically incident wave are so related that

$$f^2_v = 81N.$$

4.Explain briefly about ground wave propagation with neat sketch ? (8M)

[CO5-L2- May/June 2012, April/May 2014, April/May 2015]

Ground Wave Propagation

The ground wave is a wave that is guided along the surface of the earth just as an electromagnetic wave is guided by a wave guide or transmission line. This ground wave propagation takes place around the curvature of the earth in the frequency bands up to 2 MHz This also called as surface wave propagation.

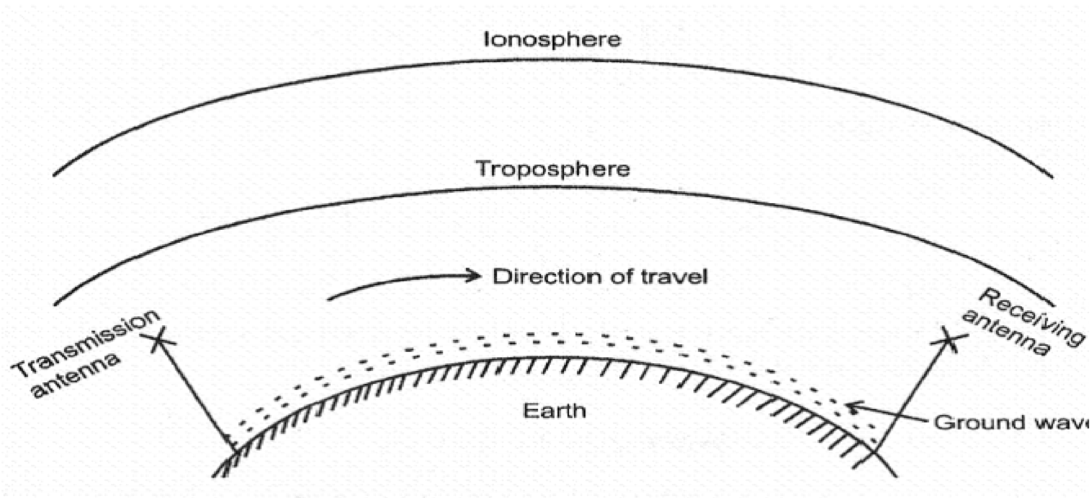


fig 4.1 Ground Wave Propagation

The ground wave is vertically polarized, as any horizontal component of the E field in contact with the earth is short-circuited by it. In this mode, the wave glides over the surface of the earth and induces charges in the earth which travel with the wave, thus constituting a current, (see Fig. 4.1). While carrying this current, the earth acts as a leaky capacitor. Hence it can be represented by a resistance or conductance shunted by a capacitive reactance.

Thus, the characteristics of the earth as a conductor can be described in terms of conductivity (a) and dielectric constant (e).

As the ground wave passes over the surface of the earth, it is weakened due to the absorption of its energy by the earth. The energy loss is due to the induced current flowing through the earth's resistance and is replenished partly, by the downward diffraction of additional energy, from the portions of the wave in the immediate vicinity of the earth's surface.

5. Discuss the characteristics of F1 and F2 layers ? (8M)

[CO5-L2- May/June 2012, April/May 2013, May/June 2016]

Characteristics of F1 Layer:

1. F1 layer is the lower end region of F-layer and which will be situated at an average height of 220 km. (generally, 140 km to 250 km).

2 . The behavior of F1 layer is similar to that of E-region (normal) and obeys the Chapman's law of variations.

3. Its critical frequency ranges from 5 MHz to 7 MHz at noon time.

4. The value of electron density varies from 2 x 105 to 4.5 x 105. www.jntuworld.com

5. F1 layer is formed by the ionization of oxygen atoms, due to an accepted view.

6. Maximum HF waves are penetrated through the F1 layer, even though some of them are reflected back.

7. The main function of F1 layer is to provide more absorption for HF waves.

8. The density of F1 layer is lowers in winter than summer, even though no great variations in height.

Characteristics of F2 Layer :

F2 layer is the upper end region of F-layer and which will be situated at a height range of 250 km to 400 km. Its critical frequency ranges from 5 MHz to 12 MHz (basically 10 MHz) and may be even more at low altitude stations. The electron density of F2 layer may varies from 3 x 105 to 2 x 106. Being the upper most regions, the air density is very low due to which ionization disappears very slowly. F2 layer is formed by ionization of UV, X-rays and corpuscular radiations.

The earth's magnetic field, atmospheric, ionosphere storms and other geomagnetic disturbances have large effect on the ionization in F2 layer. This layer does not follow Chapman's law of variations. This is the most important reflecting medium for high frequency radio waves.

6. Write a short note on, (a) Selective fading and interference fading (b) Lowest usable high frequency (c) Field strength calculation for radio AM Broadcast waves . (16M) [CO5-L1- April/May 2014, April/May 2015]

(a) Selective Fading

This type of fading produces serious distortion in modulated signal. Selective fading is important at higher frequencies. Selective fading generally occurs in amplitude modulated signals. SSB signals become less distorted compared to the AM signals due to selective fading.

Interference Fading

Interference fading occurs due to the variation in different layers of ionospheric region. This type of fading is very serious and produces interference between the upper and lower rays of sky wave propagation. Interference fading can be reduced with the help of frequency and space diversity reception.

(b) Lowest Usable High Frequency (LUHF)

The lowest usable frequency can be defined as the maximum value of frequency necessary to establish (or maintain) point to point communication. As the frequency decreases, the sensitivity and external noise increases. The lowest usable frequency (LUF) depends on the transmitted power. Lowest usable frequency is higher in day time compared to night time depending upon the noise level at the receiving side, lowest usable frequency is measured. Where, Lowest usable frequency for sky wave propagation is limited due to:

1. Sky wave absorption and

2. Atmospheric noise.

(c) Field strength calculation for radio AM Broadcast waves:

Ground wave propagation is very useful at lower frequencies between 1 -2 MHz this mode of propagation exists when the transmitting and receiving antennas ART very close to the surface of the earth. The genera expression for field strength of ground wave propagation is given as, Where

E = Field strength due to ground wave propagation

h_t = Height of transmitting antenna h_r = Height of receiving antenna

λ = Wavelength (meters)

d = Distance between the transmitting and receiving antenna

I_s = Current in antenna

The above expression is valid when distance (d) is very small. As the distance increases, ground attenuation and absorption increases. Field strength of ground wave propagation according to sommerfield is,

$E_g = (E_0 A)/d$ Where,

A = Attenuation factor

d = Distance between the transmitting and receiving antenna

E_0 = Ground field strength at the surface of earth

E_g = Ground field strength.

The value of ground field strength at the surface of earth (Eo) depends upon, (i) Directivity of planes which are vertical and horizontal. (ii) Power radiation of transmitting antenna. The field at unit distance (1Km) for a radiated power of 1 kW, can be calculated as,

$$E_0 = (300\sqrt{P})/d \text{ (V/m)}$$

Where,

d = Distance in kilometers (km)

P= Radiated power (1 kW) In case of vertical uni-pole antenna,

the field strength E_0 at a distance of d is,

$E_0 = \sqrt{90P}/d$ volts/meter

From above, field strength is directly proportional to the square root of the power radiated. E0= 300 mV/mt. at P = I kW, d = 1 km = 186.45 mV/m at J = 1 mile

7. What are the different mechanisms of propagation of electromagnetic waves ?Explain Modes of Propagation (8M) [CO5-L2- April/May 2015]

Electromagnetic waves may travel from transmitting antenna to the receiving antenna in a number of ways. Different propagations of electromagnetic waves are as follows,

(i) Ground wave propagation

(ii) Sky wave propagation

(iii) Space wave propagation

(iv) Tropospheric scatter propagation.

This classification is based upon the frequency range, distance and several other factors.

(i) Ground Wave Propagation

Ground wave propagation is also known as surface wave propagation. This propagation is practically important at frequencies up to 2 MHz. Ground wave propagation exists when transmitting and receiving antenna are very close to the earth's curvature. Ground wave propagation suffers attenuation while propagating along the surface of the earth. This propagation can be subdivided into two types which are space wave and surface wave propagation

Applications

Ground wave propagation is generally used in TV, radio broadcasting etc.

(ii) Sky Wave Propagation

Sky wave propagation is practically important at frequencies between 2 to 30 MHz Here the electromagnetic waves reach the receiving point after reflection from an atmospheric layer known as ionosphere. Hence, sky wave propagation is also known as 'ionospheric wave propagation'. It can provide communication over long distances. Hence, it is also known as point-to-point propagation or point-to-point communication.

Disadvantage

Sky wave propagation suffers, from fading due to reflections from earth surface, fading can be reduced with the help of diversity reception.

Applications

1. It can provide communication over long distances.

2. Global communication is possible.

(iii) Space Wave Propagation

Space wave propagation is practically important at frequencies above 30 MHz It is also known as tropospheric wave propagation because the waves reach the receiving point after reflections from tropospheric region. In space wave propagation, signal at the receiving point is a combination of direct and indirect rays. It provides communication over long distances with VHF .UHF and microwave frequencies. Space wave propagation is also known as "line of sight propagation".

Applications

1. Space wave propagation is used in satellite communication.

2. It controls radio traffic between a ground station and a satellite.

(iv) Troposcatter Propagation

Troposcatter propagation is also known as forward 1 scatter propagation, it is practically important at frequencies above 300 MHz.. This propagation covers long distances in the range of 160 to 1600 km.