MODULE 1

MODULE STRUCTURE

1.1 MICROWAVE TUBES: INTRODUCTION

For extremely high-frequency applications (above 1 GHz), the inter electrode capacitances and transit-time delays of standard electron tube construction become prohibitive. However, there seems to be no end to the creative ways in which tubes may be constructed, and several high-frequency electron tube designs have been made to overcome these challenges. It was discovered in 1939 that a toroid cavity made of conductive material called a cavity resonator surrounding an electron beam of oscillating intensity could extract power from the beam without actually intercepting the beam itself. The oscillating electric and magnetic fields associated with the beam “echoed” inside the cavity, in a manner similar to the sounds of traveling automobiles echoing in a roadside canyon, allowing radio-frequency energy to be transferred from the beam to a waveguide or coaxial cable connected to the resonator with a coupling loop.

1.2 OBJECTIVE

This module enables students Describe the microwave properties and its transmission media.

1.3REFLEX KLYSTRON OSCILLATOR

The electron gun emits the electron beam, which passes through the gap in the anode cavity. These electrons travel towards the Repeller electrode, which is at high negative potential. Due to the high negative field, the electrons repel back to the anode cavity. In their return journey, the electrons give more energy to the gap and these oscillations are sustained. The constructional details of this reflex klystron are as shown in the following Figure 1.1. It is assumed that oscillations already exist in the tube and they are sustained by its operation. The electrons while passing through the anode cavity, gain some velocity.

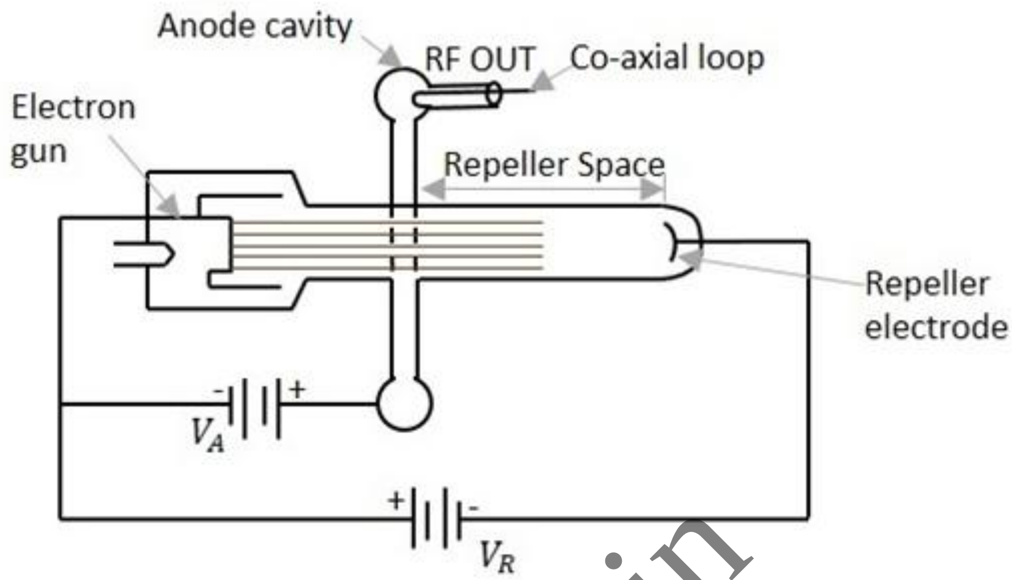


Figure 1.1: Construction details of Reflex Klystron

1.4 Mechanism of Oscillation

The mechanism of oscillation is explained in below figure 1.2. The operation of Reflex Klystron is understood by some assumptions. The electron beam is accelerated towards the anode cavity. Let us assume that a reference electron er crosses the anode cavity but has no extra velocity and it repels back after reaching the Repeller electrode, with the same velocity. Another electron, let's say ee which has started earlier than this reference electron, reaches the Repeller first, but returns slowly, reaching at the same time as the reference electron. We have another electron, the late electron el, which starts later than both erand ee, however, it moves with greater velocity while returning back, reaching at the same time as er and ee. Now, these three electrons, namely er, ee and el reach the gap at the same time, forming an electron bunch. This travel time is called as transit time, which should have an optimum value. The following figure illustrates this. The anode cavity accelerates the electrons while going and gains their energy by retarding them during the return journey. When the gap voltage is at maximum positive, this lets the maximum negative electrons to retard.

The optimum transit time is represented as

T=n+34wherenisanintegerT=n+34wherenisaninteger

This transit time depends upon the Repeller and anode voltages.

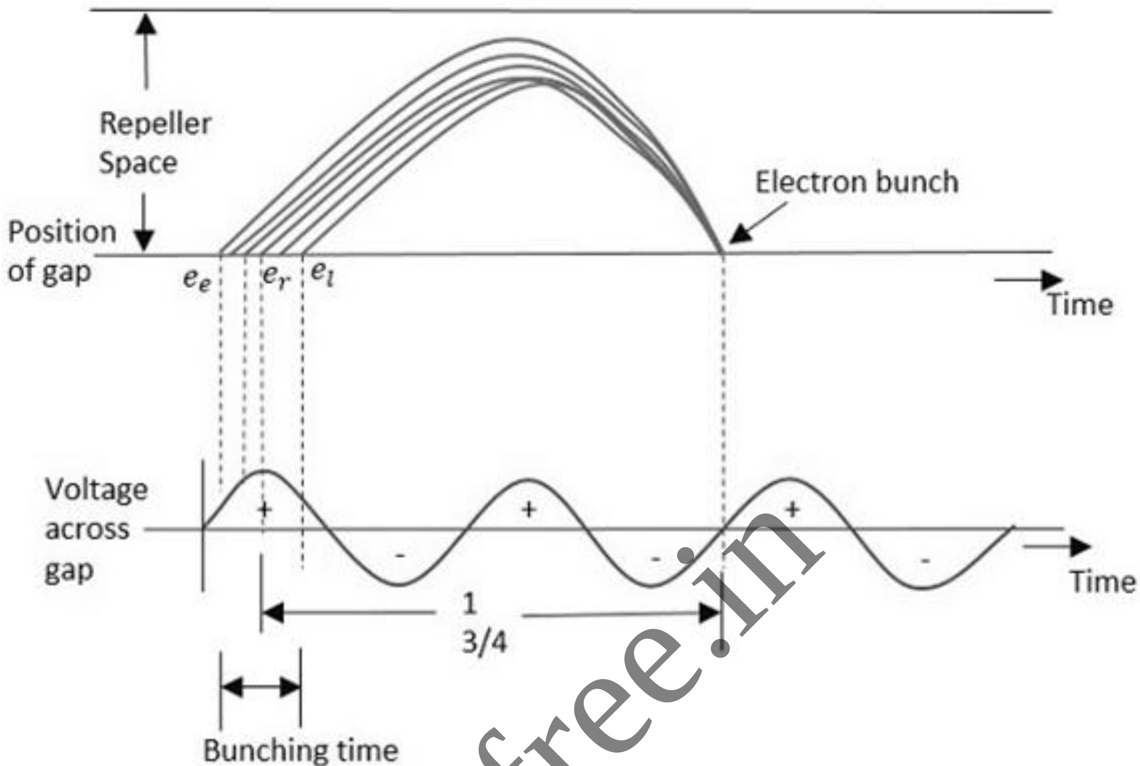


Figure1.2: Mechanism of Oscillation

1.5 Modes of Oscillation

The electrons should return after 1¾, 2 ¾ or 3 ¾ cycles – most optimum departure time. If T is the time period at the resonant frequency, to is the time taken by the reference electron to travel in the repelled space between entering the repelled space and returning to the cavity at positive peak voltage on formation of the bunch

Then, to = (n + ¾)T = NT

Where N = n + ¾, n = 0,1,2,3…….

N – mode of oscillation.

The mode of oscillation is named as N = ¾, 1 ¾ 2 ¾ etc for modes n = 0,1,2……resp.

1.6 Mode Curve (Qualitative Analysis only)

Since the output power and Frequency can be Electronically Controlled by varying the repeller voltage, expansions for those parameters in terms of repeller voltage are important to draw mode curves. Figure 1.3 shows mode curve. PRF= 0.3986V0 I0(V0+VR)sqrt(e/2mV0)/2*f*L

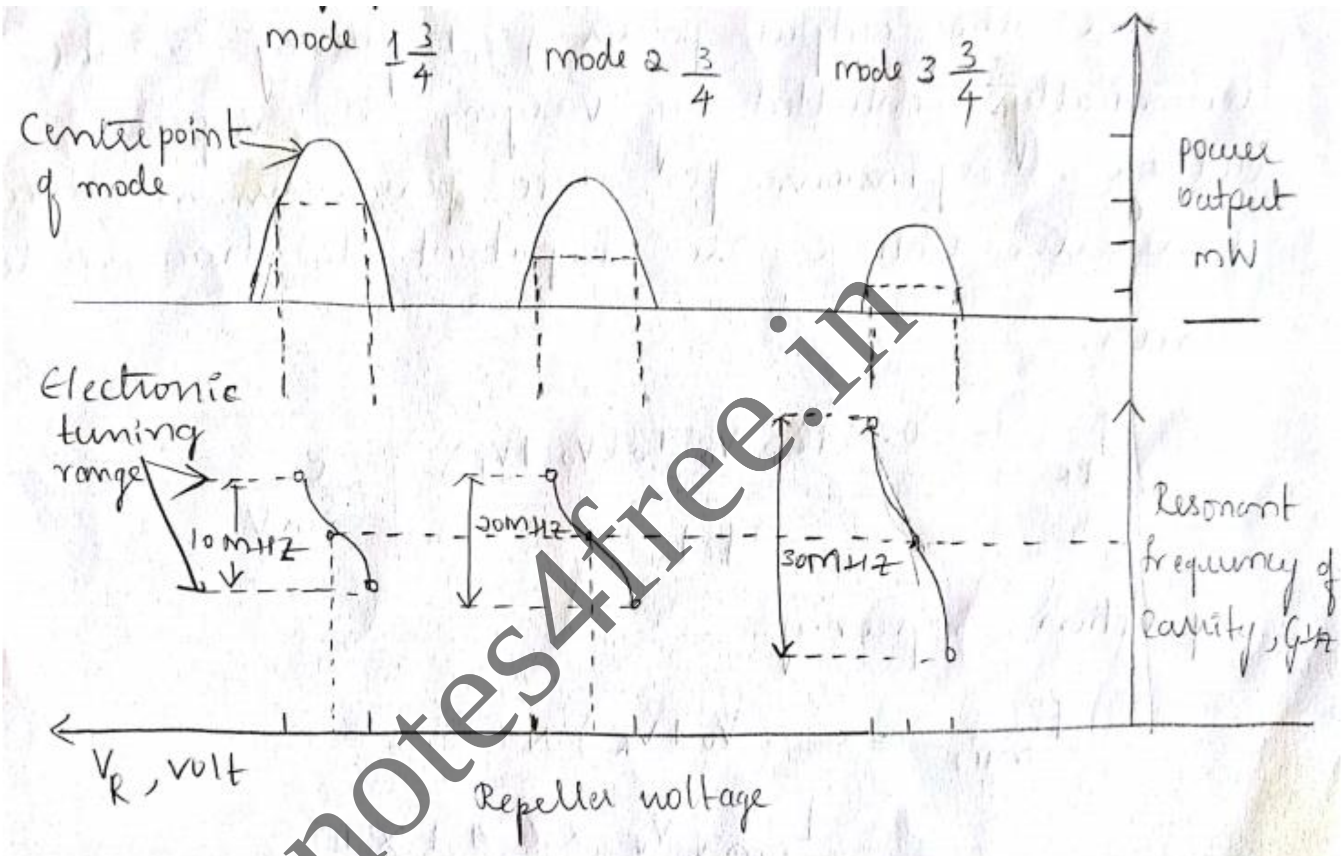


Figure 1.3: Electronic tuning and output mode power of a reflex Klystron

1.7 MICROWAVE FREQUENCIES

The term microwave frequencies are generally used for those wavelengths measured in centimetres, roughly from 30 cm to 1 mm (1 to 300 GHz). However, microwave really indicates the wavelengths in the micron ranges. This means microwave frequencies are up to infrared and visible-light regions. In this revision, microwave frequencies refer to those from 1 GHz up to 106 GHz.

Designation	Frequency range in gigahertz
HF	0.003– 0.030
VHF	0.030– 0.300
UHF	0.300– 1.000
L band	1.000– 2.000
S band	2.000– 4.000
C band	4.000– 8.000
X band	8.000– 12.000
Ku band	12.000– 18.000
K band	18.000– 27.000
Ka band	27.000– 40.000
Millimeter	40.000–300.000
Submillimeter	>300.000

1.8 MICROWAVE DEVICES

In the late 1930s it became evident that as the wavelength approached the physical dimensions of the vacuum tubes, the electron transit angle, interelectrode capacitance, and lead inductance appeared to limit the operation of vacuum tubes in microwave frequencies. In 1935 A. A. Heil and 0. Heil suggested that microwave voltages be generated by using transit-time effects together with lumped tuned cirSec. 0.3 Microwave Systems circuits. In 1939 W. C. Hahn and G. F. Metcalf proposed a theory of velocity modulation for microwave tubes. Four months later R. H. Varian and S. F. Varian described a two-cavity klystron amplifier and oscillator by using velocity modulation. In 1944 R. Kompfner invented the helix-type traveling-wave tube (TWT). Ever since then the concept of microwave tubes has deviated from that of conventional vacuum tubes as a result of the application of new principles in the amplification and generation of microwave energy. Historically microwave generation and amplification were accomplished by means of velocity-modulation theory. In the past two decades, however, microwave solid-state devices-such as tunnel diodes, Gunn diodes, transferred electron devices (TEDs), and avalanche transit-time devices have been developed to perform these functions. The conception and subsequent development of TEDs and avalanche transit-time devices were among the outstanding technical achievements. B. K. Ridley and T. B. Watkins in 1961 and C. Hilsum in 1962 independently predicted that the transferred electron effect would occur in GaAs (gallium arsenide). In 1963 J. B.

Gunn reported his "Gunn effect." The common characteristic of all microwave solid-state devices is the negative resistance that can be used for microwave oscillation and amplification. The progress of TEDs and avalanche transit-time devices has been so swift that today they are firmly established as one of the most important classes of microwave solid-state devices.

1.9 MICROWAVE SYSTEMS

A microwave system normally consists of a transmitter subsystem, including a microwave oscillator, waveguides, and a transmitting antenna, and a receiver subsystem that includes a receiving antenna, transmission line or waveguide, a microwave amplifier, and a receiver. Figure 1.4 shows a typical microwave system. In order to design a microwave system and conduct a proper test of it, an adequate knowledge of the components involved is essential. Besides microwave devices, the text therefore describes microwave components, such as resonators, cavities, microstrip lines, hybrids, and microwave integrated circuits.

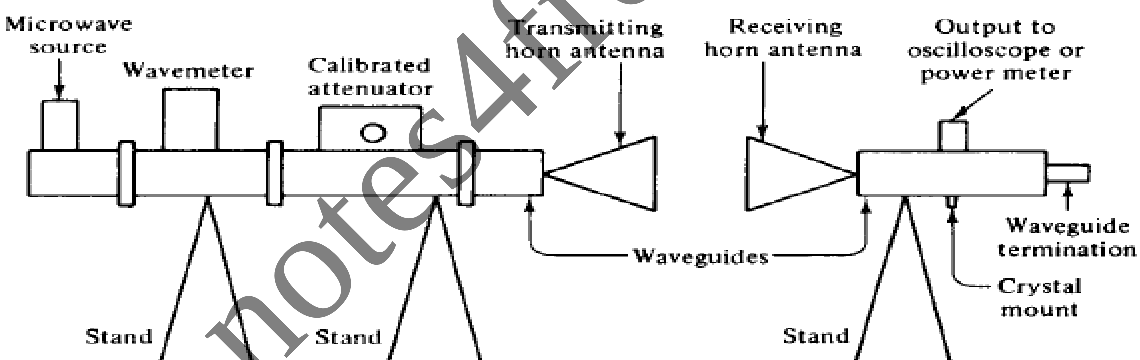


Figure 1.4: Electronic tuning and output mode power of a reflex Klystron

1.10 TRANSMISSION LINE EQUATIONS AND SOLUTIONS

A transmission line can be analysed either by the solution of Maxwell's field equations or by the methods of distributed-circuit theory. The solution of Maxwell's equations involves three space variables in addition to the time variable. The distributed-circuit method, however, involves only one space variable in addition to the time variable. Here latter method is used to analyse a transmission line in terms of the voltage, current, impedance, and power along the line. Based on uniformly distributed-circuit theory, the schematic circuit of a conventional two-conductor transmission line with constant parameters R, L, G, and C. The parameters are expressed in their respective names per unit length, and the wave propagation is assumed in the positive z direction.

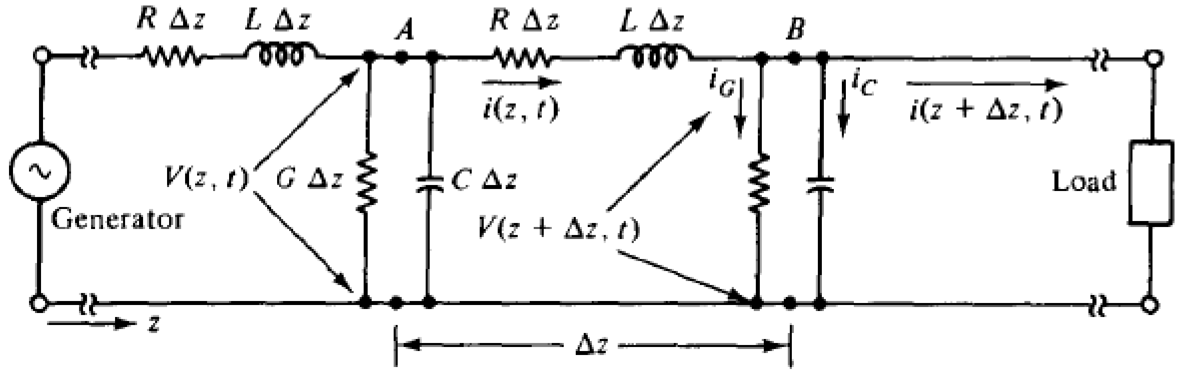


Figure 1.5: Elementary Section of Transmission Line

By Kirchhoff's voltage law, the summation of the voltage drops around the central loop is given by

$$v(z,\ t) = i(z,\ t)R\ \Delta z + L\,\Delta z\frac{\partial i(z,\ t)}{\partial t} + v(z,\ t) + \frac{\partial v(z,\ t)}{dz}\,\Delta z$$

Rearranging this equation, dividing it by Az, and then omitting the argument (z, *t),* which is understood, we obtain

$$-\frac{\partial v}{\partial z} = Ri + L\frac{\partial i}{\partial t}$$

Using Kirchhoff's current law, the summation of the currents at point *B, we obtain*

$$-\frac{\partial i}{\partial z} = Gv + C\frac{\partial v}{\partial t}$$

All these transmission-line equations are applicable to the general transient solution. The voltage and current on the line are the functions of both position *z* and time *t.* The instantaneous line voltage and current can be expressed as

$$v(z,\ t) = \text{Re}\ \mathbf{V}(z)e^{j\omega t}$$

$$i(z,\ t) = \text{Re}\ \mathbf{I}(z)e^{j\omega t}$$

$$\mathbf{V}(z) = \mathbf{V}_+ e^{-\gamma z} + \mathbf{V}_- e^{\gamma z}$$

$$\mathbf{I}(z) = \mathbf{I}_+ e^{-\gamma z} + \mathbf{I}_- e^{\gamma z}$$

$$\gamma = \alpha + j\beta \qquad \text{(propagation constant)}$$

Where V+ and I+ indicate complex amplitudes in the positive *z* direction, V- and **L** signify complex amplitudes in the negative *z* direction, *a* is the attenuation constant in nepers per unit length, and *f3* is the phase constant in radians per unit length. If we substitute *jw* for *a/at* and divide each equation by *jw*, the transmission-line equations in phasor form of the frequency domain become.

$$\frac{d\mathbf{V}}{dz} = -\mathbf{ZI}$$

$$\frac{d\mathbf{I}}{dz} = -\mathbf{YV}$$

$$\frac{d^2\mathbf{V}}{dz^2} = \gamma^2 \mathbf{V}$$

$$\frac{d^2\mathbf{I}}{dz^2} = \gamma^2 \mathbf{I}$$

in which the following substitutions have been made:

$$\mathbf{Z} = R + j\omega L \qquad \text{(ohms per unit length)}$$

$$\mathbf{Y} = G + j\omega C \qquad \text{(mhos per unit length)}$$

$$\gamma = \sqrt{ZY} = \alpha + j\beta \qquad \text{(propagation constant)}$$

For a lossless line, R = G = 0, and the transmission-line equations are expressed as

$$\frac{d\mathbf{V}}{dz} = -j\omega L\mathbf{I} \longrightarrow \quad 1.1$$

$$\frac{d\mathbf{I}}{dz} = -j\omega C\mathbf{V} \longrightarrow \quad 1.2$$

$$\frac{d^2\mathbf{V}}{dz^2} = -\omega^2 LC\mathbf{V} \longrightarrow \quad 1.3$$

$$\frac{d^2\mathbf{I}}{dz^2} = -\omega^2 LC\mathbf{I} \longrightarrow \quad 1.4$$

Solutions to Transmission Line: The one possible solution for Eq. 1.3

$$\mathbf{V} = \mathbf{V}_+ e^{-\gamma z} + \mathbf{V}_- e^{\gamma z} = \mathbf{V}_+ e^{-\alpha z} e^{-j\beta z} + \mathbf{V}_- e^{\alpha z} e^{j\beta z}$$

The one possible solution for Eq. 1.4

$$\mathbf{I} = \mathbf{Y}_0(\mathbf{V}_+ e^{-\gamma z} - \mathbf{V}_- e^{\gamma z}) = \mathbf{Y}_0(\mathbf{V}_+ e^{-\alpha z} e^{-j\beta z} - \mathbf{V}_- e^{\alpha z} e^{j\beta z})$$

At microwave frequencies it can be seen that $R << wL$ and $G << wC$

By using the binomial expansion, the propagation constant can be expressed as

$$\gamma = \sqrt{(R + j\omega L)(G + j\omega C)}$$

$$= \sqrt{(j\omega)^2 LC}\sqrt{\left(1 + \frac{R}{j\omega L}\right)\left(1 + \frac{G}{j\omega C}\right)}$$

$$\approx j\omega\sqrt{LC}\left[\left(1 + \frac{1}{2}\frac{R}{j\omega L}\right)\left(1 + \frac{1}{2}\frac{G}{j\omega C}\right)\right]$$

$$\approx j\omega\sqrt{LC}\left[1 + \frac{1}{2}\left(\frac{R}{j\omega L} + \frac{G}{j\omega C}\right)\right]$$

$$= \frac{1}{2}\left(R\sqrt{\frac{C}{L}} + G\sqrt{\frac{L}{C}}\right) + j\omega\sqrt{LC}$$

$$\alpha = \frac{1}{2}\left(R\sqrt{\frac{C}{L}} + G\sqrt{\frac{L}{C}}\right)$$

$$\beta = \omega\sqrt{LC}$$

$$\mathbf{Z}_0 = \sqrt{\frac{R + j\omega L}{G + j\omega C}}$$
$$= \sqrt{\frac{L}{C}}\left(1 + \frac{R}{j\omega L}\right)^{1/2}\left(1 + \frac{G}{j\omega C}\right)^{-1/2}$$
$$\approx \sqrt{\frac{L}{C}}\left(1 + \frac{1}{2}\frac{R}{j\omega L}\right)\left(1 - \frac{1}{2}\frac{G}{j\omega C}\right)$$
$$\approx \sqrt{\frac{L}{C}}\left[1 + \frac{1}{2}\left(\frac{R}{j\omega L} - \frac{G}{j\omega C}\right)\right]$$
$$\approx \sqrt{\frac{L}{C}}$$

$$v_p = \frac{\omega}{\beta} = \frac{1}{\sqrt{LC}}$$

Problem on Line Characteristic Impedance:

A transmission line has the following parameters:

$R = 2\ \Omega/\mathrm{m}$ $\quad G = 0.5$ mmho/m $\quad f = 1$ GHz

$L = 8$ nH/m $\quad C = 0.23$ pF

Calculate: **(a)** the characteristic impedance; **(b)** the propagation constant.

Solution

a. From Eq. (3-1-25) the line characteristic impedance is

$$\mathbf{Z}_0 = \sqrt{\frac{R + j\omega L}{G + j\omega C}} = \sqrt{\frac{2 + j2\pi \times 10^9 \times 8 \times 10^{-9}}{0.5 \times 10^{-3} + j2\pi \times 10^9 \times 0.23 \times 10^{-12}}}$$
$$= \sqrt{\frac{50.31\angle 87.72^\circ}{15.29 \times 10^{-4}\angle 70.91^\circ}} = 181.39\angle 8.40^\circ = 179.44 + j26.50$$

b. From Eq. (3-1-18) the propagation constant is

$$\gamma = \sqrt{(R + j\omega L)(G + j\omega C)} = \sqrt{(50.31\angle 87.72^\circ)(15.29 \times 10^{-4}\angle 70.91^\circ)}$$
$$= \sqrt{769.24 \times 10^{-4}\angle 158.63^\circ}$$
$$= 0.2774\angle 79.31^\circ = 0.051 + j0.273$$

1.11 REFLECTION COEFFICIENT

In the analysis of the solutions of transmission-line equations in 1.10, the traveling wave along the line contains two components: one traveling in the positive z direction and the other traveling the negative z direction. If the load impedance is equal to the line characteristic impedance, however, the reflected traveling wave does not exist. Figure 1.5 shows a transmission line terminated in impedance Ze. It is usually more convenient to start solving the transmission-line problem from the receiving rather than the sending end, since the voltage-to-current relationship at the load point is fixed by the load impedance. The incident voltage and current waves traveling along the transmission line are given by

The reflection coefficient, which is designated by Γ (gamma), is defined as,

$$\text{Reflection coefficient} \equiv \frac{\text{reflected voltage or current}}{\text{incident voltage or current}}$$

$$\Gamma \equiv \frac{V_{ref}}{V_{inc}} = \frac{-I_{ref}}{I_{inc}}$$

TRANSMISSION COEFFICIENT

A transmission line terminated in its characteristic impedance Zo is called a properly terminated line. Otherwise it is called an improperly terminated line. As described earlier, there is a reflection coefficient r at any point along an improperly terminated line. According to the principle of conservation of energy, the incident power minus the reflected power must be equal to the power transmitted to the load. This can be expressed as,

$$T \equiv \frac{\text{transmitted voltage or current}}{\text{incident voltage or current}} = \frac{V_{tr}}{V_{inc}} = \frac{I_{tr}}{I_{inc}}$$

A certain transmission line has a characteristic impedance of $75 + j0.01\ \Omega$ and is terminated in a load impedance of $70 + j50\ \Omega$. Compute **(a)** the reflection coefficient; **(b)** the transmission coefficient. Verify: **(c)** the relationship shown in Eq. (3-2-21); **(d)** the transmission coefficient equals the algebraic sum of 1 plus the reflection coefficient as shown in Eq. (2-3-18).

Solution

a. From Eq. (3-2-17) the reflection coefficient is

$$\mathbf{\Gamma} = \frac{\mathbf{Z}_\ell - \mathbf{Z}_0}{\mathbf{Z}_\ell + \mathbf{Z}_0} = \frac{70 + j50 - (75 + j0.01)}{70 + j50 + (75 + j0.01)}$$

$$= \frac{50.24\underline{/95.71^\circ}}{153.38\underline{/19.03^\circ}} = 0.33\underline{/76.68^\circ} = 0.08 + j0.32$$

b. From Eq. (3-2-18) the transmission coefficient is

$$\mathbf{T} = \frac{2\mathbf{Z}_\ell}{\mathbf{Z}_\ell + \mathbf{Z}_0} = \frac{2(70 + j50)}{70 + j50 + (75 + j0.01)}$$

$$= \frac{172.05\underline{/35.54^\circ}}{153.38\underline{/19.03^\circ}} = 1.12\underline{/16.51^\circ} = 1.08 + j0.32$$

c.

$$\mathbf{T}^2 = (1.12\underline{/16.51^\circ})^2 = 1.25\underline{/33.02^\circ}$$

$$\frac{\mathbf{Z}_\ell}{\mathbf{Z}_0}(1 - \mathbf{\Gamma}^2) = \frac{70 + j50}{75 + j0.01}[1 - (0.33)\underline{/76.68^\circ})^2]$$

$$= \frac{86\underline{/35.54^\circ}}{75\underline{/0^\circ}} \times 1.10\underline{/-2.6^\circ} = 1.25\underline{/33^\circ}$$

Thus Eq. (3-2-21) is verified.

d. From Eq. (2-3-18) we obtain

$$\mathbf{T} = 1.08 + j0.32 = 1 + 0.08 + j0.32 = 1 + \mathbf{\Gamma}$$

1.13 STANDING WAVE AND STANDING WAVE RATIO

The general solutions of the transmission-line equation consist of two waves traveling in opposite directions with unequal amplitude as shown in Eq 1.3 and 1.4.

$$\begin{aligned}\mathbf{V} &= \mathbf{V}_+ e^{-\alpha z} e^{-j\beta z} + \mathbf{V}_- e^{\alpha z} e^{j\beta z} \\ &= \mathbf{V}_+ e^{-\alpha z}[\cos(\beta z) - j\sin(\beta z)] + \mathbf{V}_- e^{\alpha z}[\cos(\beta z) + j\sin(\beta z)] \\ &= (\mathbf{V}_+ e^{-\alpha z} + \mathbf{V}_- e^{\alpha z})\cos(\beta z) - j(\mathbf{V}_+ e^{-\alpha z} - \mathbf{V}_- e^{\alpha z})\sin(\beta z)\end{aligned}$$

With no loss in generality it can be assumed that *V+e-az* and V-e''' are real. Then the voltage-wave equation can be expressed as

$$\mathbf{V}_s = \mathbf{V}_0 e^{-j\phi}$$

This is called the *equation of the voltage standing wave,* where

$$\mathbf{V}_0 = [(\mathbf{V}_+ e^{-\alpha z} + \mathbf{V}_- e^{\alpha z})^2 \cos^2(\beta z) + (\mathbf{V}_+ e^{-\alpha z} - \mathbf{V}_- e^{-\alpha z})^2 \sin^2(\beta z)]^{1/2}$$

$$\phi = \arctan\left(\frac{\mathbf{V}_+ e^{-\alpha z} - \mathbf{V}_- e^{\alpha z}}{\mathbf{V}_+ e^{-\alpha z} + \mathbf{V}_- e^{\alpha z}} \tan(\beta z)\right)$$

which is called the *phase pattern of the standing wave.*

By doing so and substituting the proper values of *f3z* in the equation, we find that,

1. The maximum amplitude is

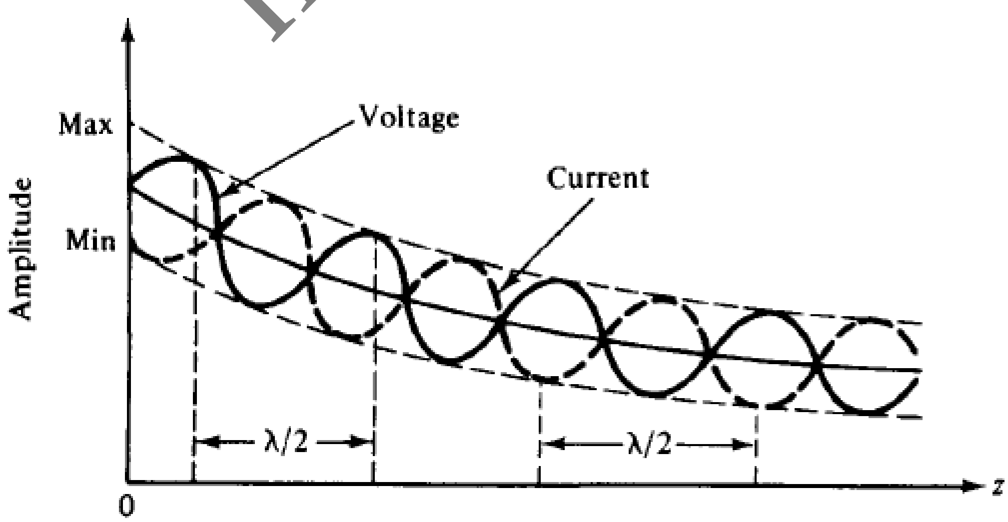
$$\mathbf{V}_{max} = \mathbf{V}_+ e^{-\alpha z} + \mathbf{V}_- e^{\alpha z} = \mathbf{V}_+ e^{-\alpha z}(1 + |\Gamma|)$$

and this occurs at $\beta z = n\pi$, where $n = 0, \pm 1, \pm 2, \ldots$

2. The minimum amplitude is

$$\mathbf{V}_{min} = \mathbf{V}_+ e^{-\alpha z} - V_- e^{\alpha z} = \mathbf{V}_+ e^{-\alpha z}(1 - |\Gamma|)$$

and this occurs at $\beta z = (2n - 1)\pi/2$, where $n = 0, \pm 1, \pm 2, \ldots$



Standing waves result from the simultaneous presence of waves traveling in opposite directions on a transmission line. The ratio of the maximum of the standing-wave pattern to the minimum is defined as the standing-wave ratio, designated by *p*. That is,

$$\text{Standing-wave ratio} \equiv \frac{\text{maximum voltage or current}}{\text{minimum voltage or current}}$$

$$\rho \equiv \frac{|\mathbf{V}_{max}|}{|\mathbf{V}_{min}|} = \frac{|\mathbf{I}_{max}|}{|\mathbf{I}_{min}|}$$

1.12 SMITH CHART

Many of the computations required to solve transmission-line problems involve the use of rather complicated equations. The solution of such problems is tedious and difficult because the accurate manipulation of numerous equations is necessary. To simplify their solution, we need a graphic method of arriving at a quick answer. A number of impedance charts have been designed to facilitate the graphic solution of transmission-line problems. Basically all the charts are derived from the fundamental relationships expressed in the transmission equations. The most popular chart is that developed by Phillip H. Smith [l]. The purpose of this section is to present the graphic solutions of transmission-line problems by using the Smith chart. The Smith chart consists of a plot of the normalized impedance or admittance with the angle and magnitude of a generalized complex reflection coefficient in a unity circle. The chart is applicable to the analysis of a lossless line as well as a lossy line. By simple rotation of the chart, the effect of the position on the line can be determined.

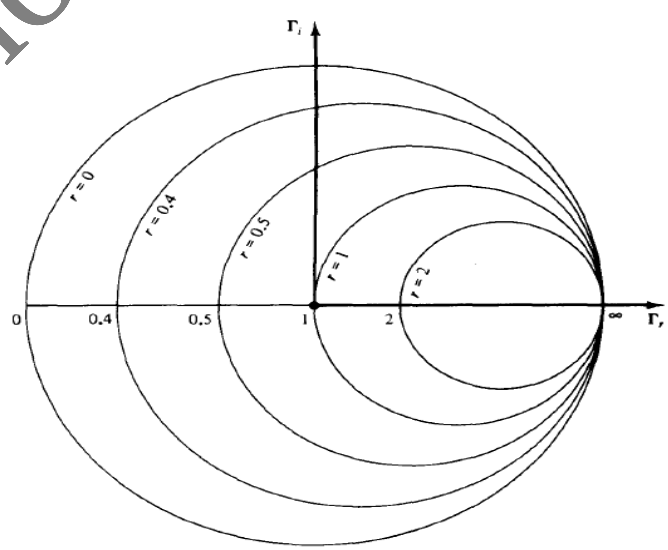


Figure: Constant r (Resistance) Circle

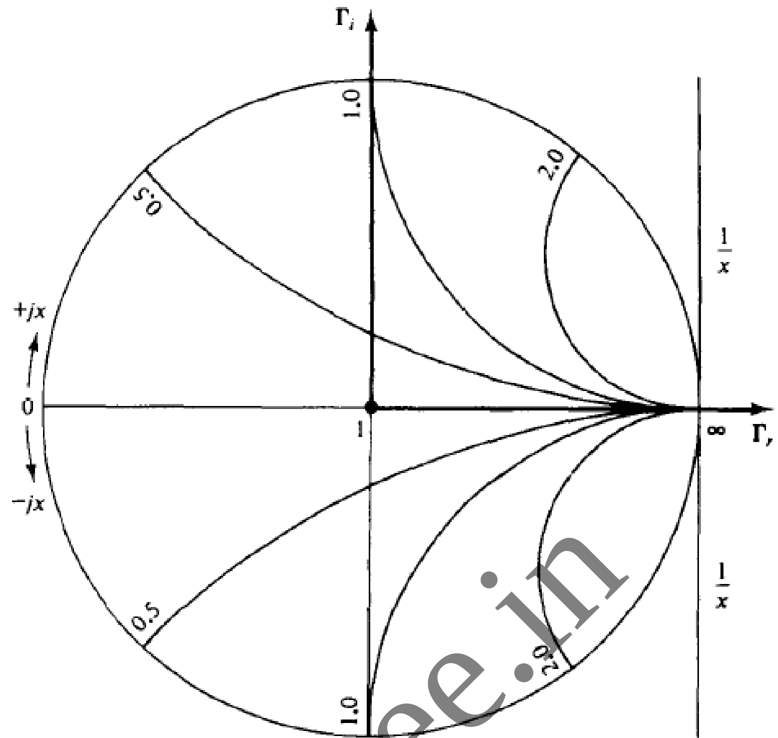


Figure: Constant x (Reactance) Circle

The characteristics of the Smith chart are summarized as follows:

1. The constant *r* and constant *x* loci form two families of orthogonal circles in the chart.
2. The constant *r* and constant *x* circles all pass through the point (L = 1,f; = O).
3. The upper half of the diagram represents + *jx*.
4. The lower half of the diagram represents - *jx*.
5. For admittance the constant *r* circles become constant *g* circles, and the constant *x* circles become constant susceptance *b* circles.
6. The distance around the Smith chart once is one-half wavelength (A/2).
7. At a point of Zmin = 1 / *p*, there is a Vmin on the line.
8. At a point of Zmax = *p'* there is a Vmax on the line.
9. The horizontal radius to the right of the chart center corresponds to Vmax, /min, Zmax, and *p* (SWR).
10. The horizontal radius to the left of the chart center corresponds to Vmin, *I* max, Zmin, and l/p.
11. Since the normalized admittance y is a reciprocal of the normalized impedance z, the corresponding quantities in the admittance chart are 180° out of phase with those in the impedance chart.
12. The normalized impedance or admittance is repeated for every half wavelength of distance.
13. The distances are given in wavelengths toward the generator and also toward the load.

1.14 SINGLE STUB MATCHING

Although single-lumped inductors or capacitors can match the transmission line, it is more common to use the susceptive properties of short-circuited sections of transmission lines. Short-circuited sections are preferable to open-circuited ones because a good short circuit is easier to obtain than a good open circuit. For a lossless line with $Y8$ = Yo, maximum power transfer requires Y11 = Yo, where Y11 is the total admittance of the line and stub looking to the right at point 1-1 (see Fig. 3-6-2). The stub must be located at that point on the line where the real part of the admittance, looking toward the load, is Yo. In a normalized unit $Y11$ must be in the form.

$$Y11 = Yd \pm Ys = 1$$

A lossless line of characteristic impedance Ro = 50 fl is to be matched to a load Ze = 50/[2 + j(2 + \/3)] fl by means of a lossless short-circuited stub. The characteristic impedance of the stub is 100 fl. Find the stub position (closest to the load) and length so that a match is obtained.

OUTCOME:

Student will be able to,

- Describe the use and advantages of microwave transmission and analyze various parameters related to microwave transmission lines and waveguides

RECOMMENEDED QUESTIONS:

1 What is Microwave system?

2 What are Microwave Frequencies?

3 What is Smith Chart and Single Stub Matching?

4 Derive equation for Reflection Coefficient.

5 Derive Equation for Transmission Coefficient.

6 Define SWR and derive the same.

MODULE 2

MODULE STRUCTURE

2.1Microwave Network theory: Symmetrical Z and Y-Parameters for Reciprocal Networks

2.2 Objective

2.3 S matrix representation of Multi-Port Networks.

2.4 Microwave Passive Devices: Coaxial Connectors and Adapters

2.5 Attenuators.

2.6 Phase Shifters

2.7 Waveguide Tees

2.8 Magic tees.

2.1Microwave Network theory: Symmetrical Z and Y-Parameters for Reciprocal Networks

A microwave network consists of coupling of various microwave components and devices such as attenuators, phase shifters, amplifiers, resonators etc., to sources through transmission lines or waveguides. Connection of two or more microwave devices and components to a single point results in a microwave junction. In a low frequency network, the input and output variables are voltage and current which can be related in terms of impedance Z-parameters, or admittance Y-parameters or hybrid h-parameters or ABCD parameters. These relationships for a two-port network of figure 2.1 can be represented by

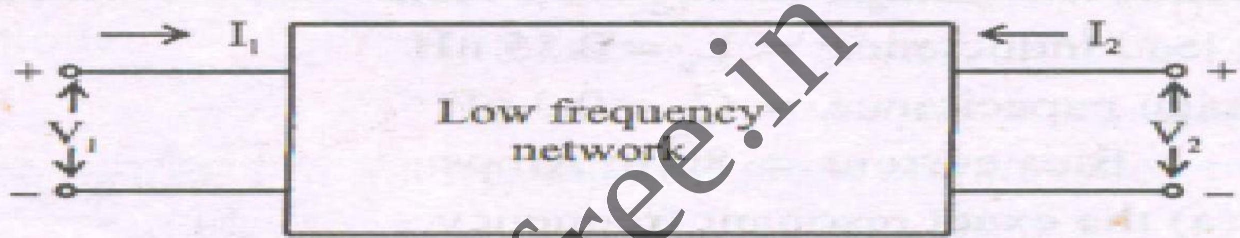


Figure 4.1': Low frequency two port network

$$\left.\begin{aligned} V_1 &= Z_{11}\ I_1 + Z_{12}\ I_2 \\ V_2 &= Z_{21}\ I_1 + Z_{22}\ I_2 \end{aligned}\right\} \quad ..... (4.1)$$

or $$\begin{bmatrix} V_1 \\ V_2 \end{bmatrix} = \begin{bmatrix} Z_{11} & Z_{12} \\ Z_{21} & Z_{22} \end{bmatrix} \begin{bmatrix} I_1 \\ I_2 \end{bmatrix} \quad ..... (4.2)$$

$$\left.\begin{aligned} I_1 &= Y_{11}\ V_1 + Y_{12}\ V_2 \\ I_2 &= Y_{21}\ V_1 + Y_{22}\ V_2 \end{aligned}\right\} \quad ..... (4.3)$$

or $$\begin{bmatrix} I_1 \\ I_2 \end{bmatrix} = \begin{bmatrix} Y_{11} & Y_{12} \\ Y_{21} & Y_{22} \end{bmatrix} \begin{bmatrix} V_1 \\ V_2 \end{bmatrix} \quad ..... (4.4)$$

$$\left.\begin{aligned} V_1 &= h_{11}\ I_1 + h_{12}\ V_2 \\ I_2 &= h_{21}\ I_1 + h_{22}\ V_2 \end{aligned}\right\} \quad ..... (4.5)$$

$$\text{or} \quad \begin{bmatrix} V_1 \\ I_2 \end{bmatrix} = \begin{bmatrix} h_{11} & h_{12} \\ h_{21} & h_{22} \end{bmatrix} \begin{bmatrix} I_1 \\ V_2 \end{bmatrix} \qquad ..... (4.6)$$

$$\left. \begin{aligned} V_1 &= AV_2 - BI_2 \\ I_1 &= CV_2 - DI_2 \end{aligned} \right\} \qquad ..... (4.7)$$

$$\text{or} \quad \begin{bmatrix} V_1 \\ I_1 \end{bmatrix} = \begin{bmatrix} A & -B \\ C & -D \end{bmatrix} \begin{bmatrix} V_2 \\ I_2 \end{bmatrix} \qquad ..... (4.8)$$

These parameters, Z, Y,h and ABeD parameters can be easily measured at low frequencies under short or open circuit conditions and can be used for analyzing the circuit. The physical length of the device or the line at microwave frequencies, is comparable to or much larger than the wavelength. Due to this, the voltage and current are difficult to measure as also the above mentioned parameters. The reasons for this are listed as below. (a) Equipment is not available to measure the total voltage and total current at any point.

(b) Over a wide range of frequencies, short and open circuits are difficult to realize.

(c) Active devices such as power transistors, tunnel diodes etc, will become unstable under short or open circuit conditions.

Therefore, a new representation is needed to overcome these problems at microwave frequencies. The logical variables are traveling waves rather than voltages and currents and these variables are labeled as "Scattering or S-parameters". These parameters for a two port network are represented as shown in Figure 4.2 .

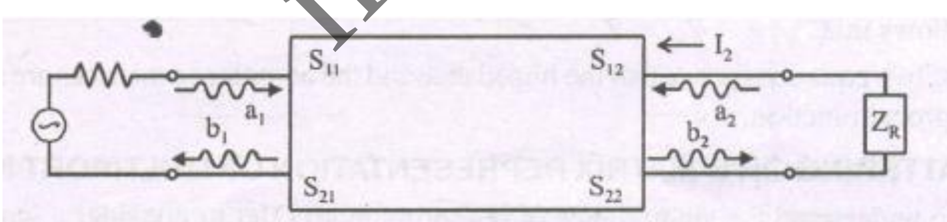


Figure 4.2 : S-parameters of a two-port network

These S-parameters can be represented in an equation form related to the traveling waves a1,a2 and b1 b2 through

$$\left. \begin{aligned} b_1 &= S_{11}\, a_1 + S_{12}\, a_2 \\ b_2 &= S_{21}\, a_1 + S_{22}\, a_2 \end{aligned} \right\} \qquad ..... (4.9)$$

2.2 Objective

This chapter enables student's to learn S- parameters and different types of connectors.

2.3 S-MATRIX REPRESENTATION OF MULTIPORT NETWORK

In a reciprocal network, the junction media are characterized by scalar electrical parameters namely absolute permeability and absolute permittivity E . In such a network, the impedance and the admittance matrices became symmetrical. This property can be proved by considering an N-port network. Let Ej and Hi be the respective electric and magnetic field intensities at the ith port and let the total voltage Vo = 0 at all ports for n = 0, 1,2 .... Except at i1hport. Similarly if E. and H. are considered for the ith port with V = 0 at other ports, then from reciprocity theorem. Let us now consider a junction of "n" number of rectangular waveguides as shown in figure 2.1. In this case, all "a' s" represent the incident waves at respective ports and all "b's" the reflected waves from the microwave junction coming out of the respective ports. In this case also, are still valid where S.. and S. have the following meanings: Sii= Scattering coefficient corresponding to the input power applied at IJ the i1hport and output power coming out of jth port and Sjj = Scattering coefficient corresponding to the power applied at the ith port " and output taken out of i1hport itself. This coefficient is a measure of amount of mismatch between the ith port and the junction.

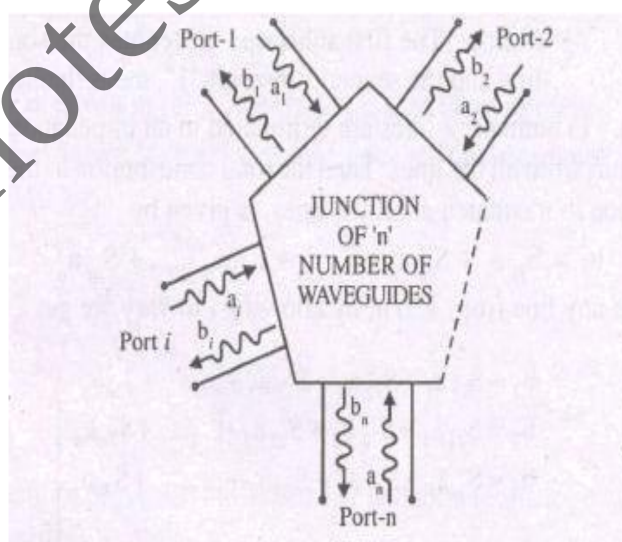


Figure 4.4 : Junction of "n" number of waveguides.

Figure 2.1: Rectangular Waveguide

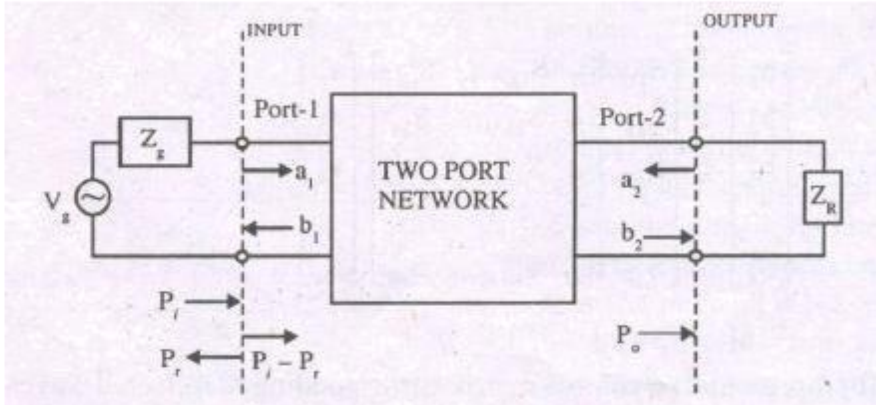


Figure 4.5 : Illustrating two-port network

The relationship. between the incident and reflected waves in terms of scattering coefficients can be written as.

$$b_1 = S_{11}\, a_1 + S_{12}\, a_2 \quad \text{..... (4.20)}$$

$$b_2 = S_{21}\, a_1 + S_{22}\, a_2 \quad \text{..... (4.21)}$$

From these equations, the scattering coefficients are found as

$S_{11} = \left.\frac{b_1}{a_1}\right|_{a_2=0}$ = reflection coefficient at port-1 when port-2 is terminated with a matched load ($a_2 = 0$)

$S_{22} = \left.\frac{b_2}{a_2}\right|_{a_1=0}$ = reflection coefficient at port-2 when port-1 is terminated with a matched load ($a_1 = 0$)

$S_{12} = \left.\frac{b_1}{a_2}\right|_{a_1=0}$ = attenuation of the wave travelling from port-2 to port-1.

$S_{21} = \left.\frac{b_2}{a_1}\right|_{a_2=0}$ = attenuation of the wave travelling from port-1 to port-2.

In figure 4.5, we have

P_i = incident power at port-1

P_r = power reflected by the network coming out of port-1 itself.

P_o = output power at port-2.

The various losses can be expressed in terms of S-parameters as given below:

$$\text{Insertion loss in dB} = 10 \log_{10} \frac{P_i}{P_o} \quad \text{..... (4.22)}$$

But $P_i \propto |a_1|^2$

$P_o \propto |b_2|^2$

$$\therefore \quad \frac{P_i}{P_o} = \frac{|a_1|^2}{|b_2|^2} = \frac{1}{\left|b_2/a_1\right|^2} = \frac{1}{|S_{21}|^2} = \frac{1}{|S_{12}|^2} \quad \text{..... (4.23)}$$

2.4 CO-AXIAL CABLES, CONNECTORS AND ADAPTERS

Coaxial Cables Microwave components and devices are interconnected using these co-axial cables of suitable length and operated at microwave frequencies. In this section let us consider some practical aspects of these co-axial cables. TEM mode is propagated through the co-axial line and the outer conductor guides these signals in the dielectric space between itself and inner conductor.

The outer conductor also acts as a shield to prevent the external signals to interfere with the internal signal. It also prevents the internal signal leakage. The co-axial cables usually possess characteristic impedance of either 50 ohms or 75 ohms Based on the structure of shielding, coaxial cables are classified into three basic types.

(i) Flexible co-axial Cable: Figure below shows the structure of flexible-type of co-axial cable consisting of low loss solid or foam type polyethylene dielectric. Electromagnetic shielding is provided for outer single braid or double braid of the flexible cable as shown, by using knitted metal wire mesh. The centre conductor usually consists of multi strand wire.

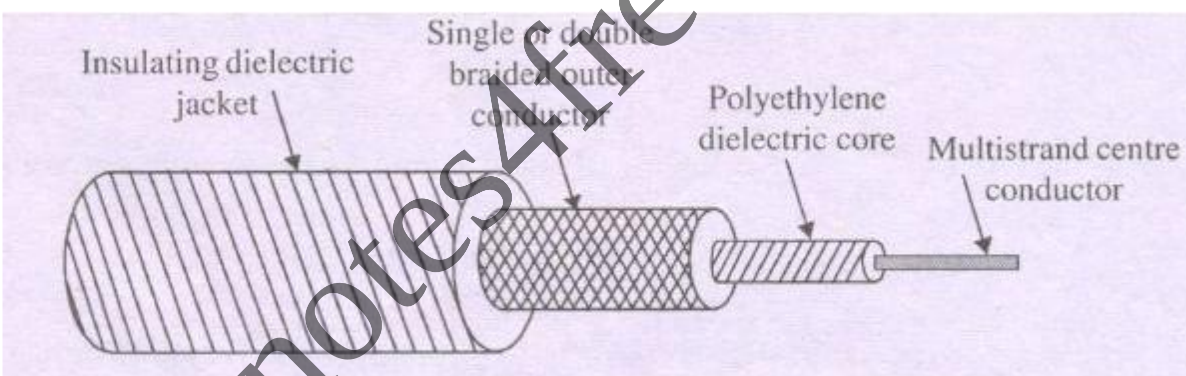


Figure 5.1 : Structure of flexible co-axial cable

1. Semi-rigid co-axial cable: Figure 5.2 shows the cross-sectional view of semi-rigid co-axial cable. Semi rigid co-axial cables make use of thin outer conductor made of copper and a strong inner conductor also made of copper. The region between the inner and outer conductor contains a solid dielectric. These cables can bent for convenient routing and are not as flexible as the first type.

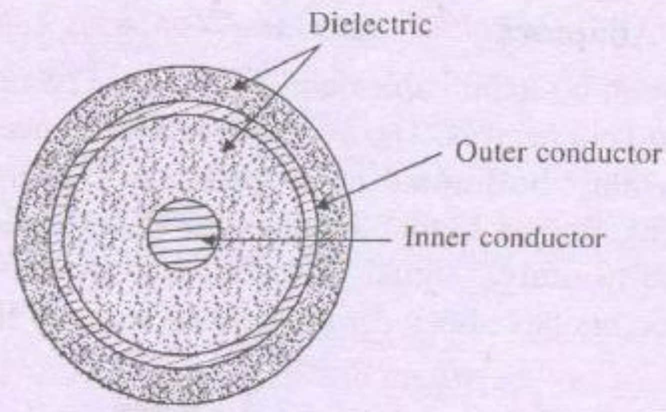


Figure 5.2 : Cross-section of semi-rigid co-axial cable

Rigid co-axial cable: Figure below shows the structure of a rigid co-axial cable consisting of inner and outer conductor with air as dielectric. To support the inner conductor at the centre dielectric spacers are introduced at regular intervals as shown. The thickness of these dielectric spacers is made small so that they do not produce significant discontinuities to the wave propagation.

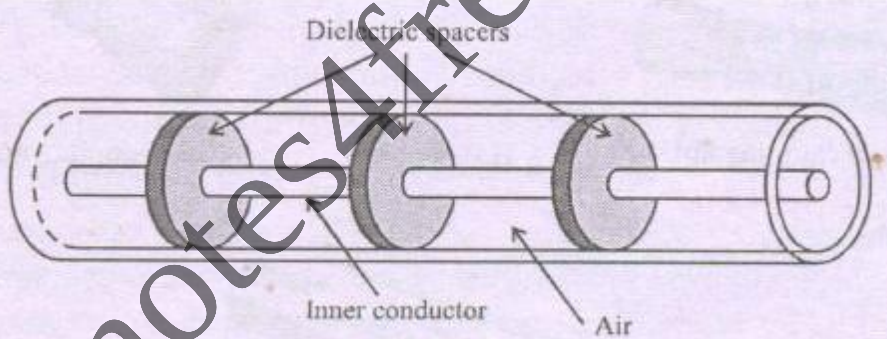


Figure 5.3 : Illustration of rigid co-axial cable

Co-axial cables can be used up to microwave -range of frequencies. Beyond these frequencies attenuation becomes very large (since attenuation increases with frequency) which makes co-axial cables unsuitable at higher frequencies. Some characteristics of standard coaxial cables with their radio guide (RG) and universal (U) numbers along with conductor (inner and outer) dimensions.

Interconnection between co-axial cables and microwave components is achieved with the help of shielded standard connectors. The average circumference of the co-axial cable, for mar high frequency operation must be limited to about one wavelength. This requirement is a VI necessary to reduce propagation at higher modes and also to eliminate erratic reflection coefficients (VSWR close to unity), signal distortion and power losses. Several types of co-axial connectors have been developed and some of them are described below.

(a) APC 3.5 (Amphenol Precision Connector - 3.5 mm)

HP (Hewlett - Packard) originally developed this connector, but it is now being manufactured by Amphenol. This connector can operate up to a frequency of 34 GHz and has a very low voltage standing wave ratio (VSWR). This connector provides repeatable connections and has 50 Q characteristic impedance. The male connector. it introduces higher ordermodes and hence not used above 24 GHz.

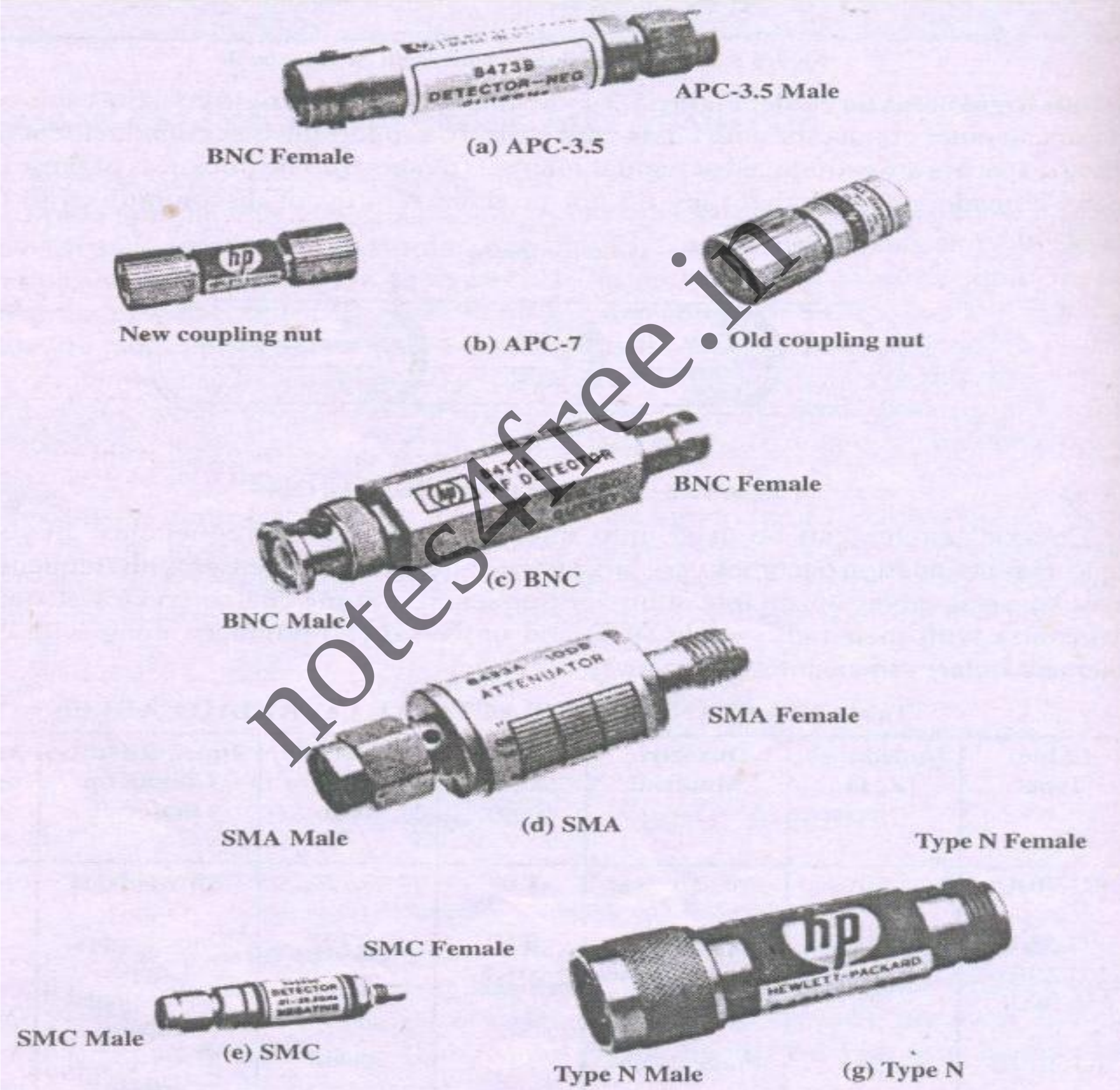


Figure 5.4 : Co-axial microwave connectors

2.5 ATTENUATORS

In order to control power levels in a microwave system by partially absorbing the transmitted microwave signal, attenuators are employed. Resistive films (dielectric glass slab coated with aquatic) are used in the design of both fixed and variable attenuators. A co-axial fixed attenuator uses the dielectric loss material inside the center conductor of the co-axial line to absorb some of the center conductor microwave power propagating through it dielectric rod decides the amount of attenuation introduced. The microwave power absorbed by the lossy material is dissipated as heat.

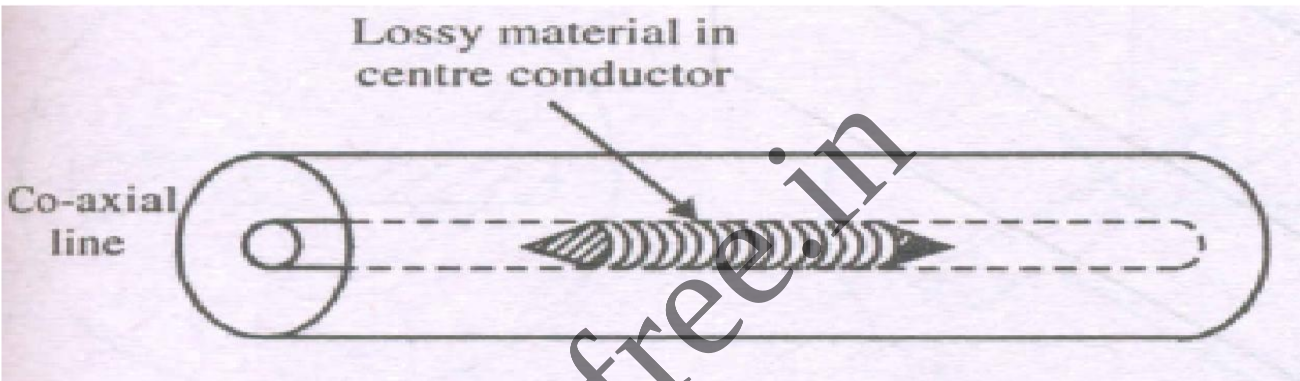


Figure 5.6 Co-axial line fixed attenuator

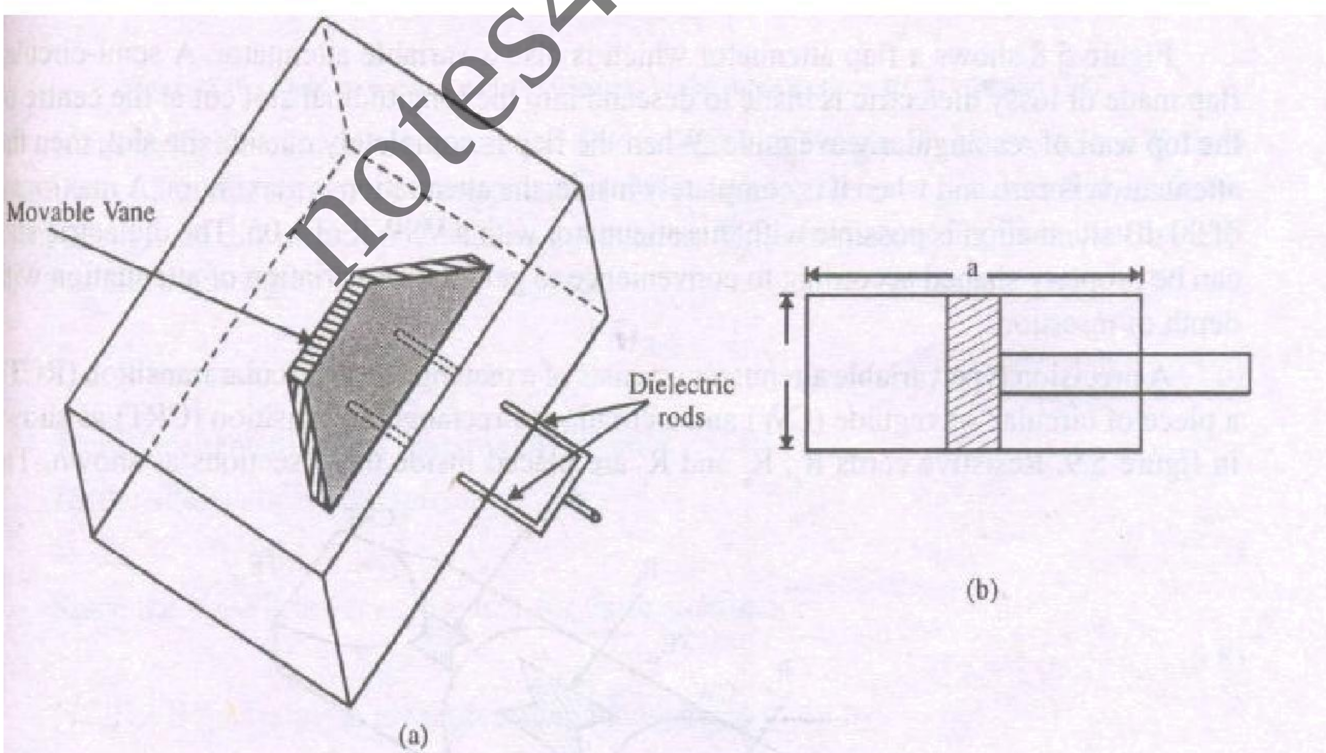


Figure 5.7 : Movable vane variable attenuator

In waveguides, the dielectric slab coated with aduadag is placed at the centre of the waveguide parallel to the maximum E-field for dominant TEIO mode. Induced current on the lossy material due to incoming microwave signal, results in power dissipation, leading to attenuation of the signal. The dielectric slab is tapered at both ends upto a length of more than half wavelength to reduce reflections as shown in figure 5.7. The dielectric slab may be made movable along the breadth of the waveguide by supporting it with two dielectric rods separated by an odd multiple of quarter guide wavelength and perpendicular to electric field. When the slab is at the centre, then the attenuation is maximum (since the electric field is concentrated at the centre for TEIO mode) and when it is moved towards one side-wall, the attenuation goes on decreasing thereby controlling the microwave power corning out of the other port.

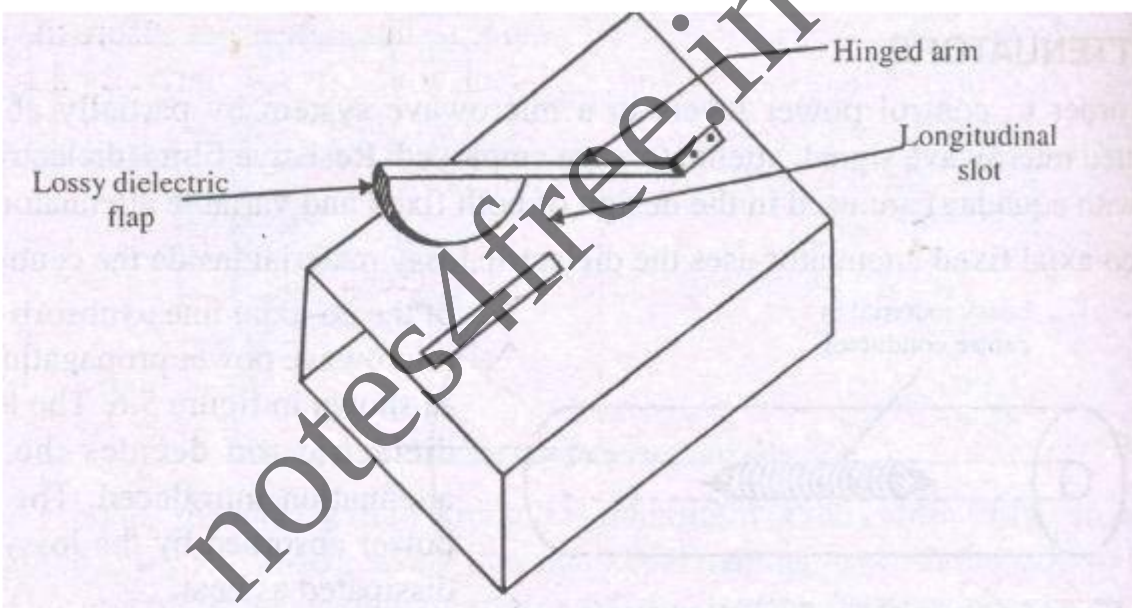


Figure 5.8 : Flap attenuator

Figure 5.8 shows a flap attenuator which is also a variable attenuator. A semi-circular flap made of lossy dielectric is made to descend into the longitudinal slot cut at the centre of the top wall of rectangular waveguide. When the flap is completely outside the slot, then the attenuation is zero and when it is completely inside, the attenuation is maximum. A maximum direction of 90 dB attenuation is possible with this attenuator with a VSWR of 1.05. The dielectric slab can be properly shaped according to convenience to get a linear variation of attenuation within the depth of insertion. A precision type variable attenuator consists of a rectangular to circular transition (ReT), a piece of circular waveguide (CW) and a circular-to-rectangular transition (CRT) as shown in figure 5.9. Resistive cards R , Rand R are placed inside these sections as shown. The centre circular section containing the resistive card Rb can be precisely rotated by 3600 with respect to the two fixed resistive cards. The induced current on the resistive card R due to the incident signal is dissipated as heat producing attenuation of the transmitted signal. TE mode in RCT is converted into TE in circular waveguide. The resistive cards R and R a kept perpendicular to the electric field of TEIO mode so that it does not absorb the energy. But any component parallel to its plane will be readily absorbed. Hence, pure TE mode is excited in circular waveguide section. II If the resistive card in the center section is kept at an angle 8 relative to the E-field direction of the TEll mode, the component E cos8 parallel to the card get absorbed while the component E sin 8 is transmitted without attenuation. This component finally comes out as E sin2θ as shown in figure 5.10.

2.6 PHASE SHIFTERS:

A microwave phase shifter is a two port device which produces a variable shift in phase of the incoming microwave signal. A lossless dielectric slab when placed inside the rectangular waveguide produces a phase shift. The rotary type of precision phase shifter is shown in figure 5.12 which consists of a circular waveguide containing a lossless dielectric plate of length 2l called "half-wave section", a section of rectangular-to-circular transition containing a lossless dielectric plate of length l, called "quarter-wave section", oriented at an angle of 45° to the broader wall of the rectangular waveguide and a circular-to-rectangular transition again containing a lossless dielectric plate of same length 1 (quarter wave section) oriented at an angle 45°. The incident TEIO mode becomes TEll mode in circular waveguide section. The half-wave section produces a phase shift equal to twice that produced by the quarter wave section. The dielectric plates are tapered at both ends to reduce reflections due to discontinuity.

2.7 WAVE GUIDE TEE JUNCTIONS:

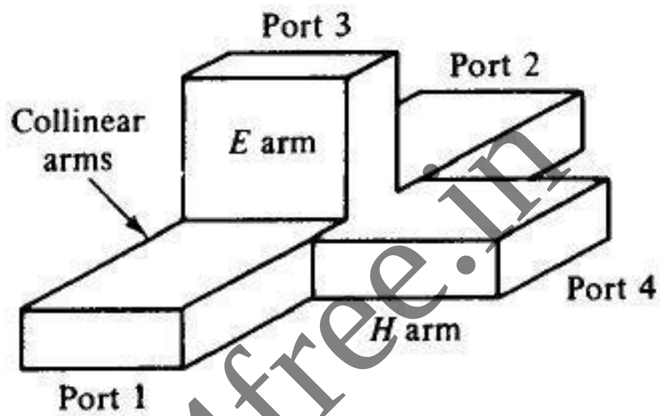
A waveguide Tee is formed when three waveguides are interconnected in the form of English alphabet T and thus waveguide tee is 3-port junction. The waveguide tees are used to connects a branch or section of waveguide in series or parallel with the main waveguide transmission line either for splitting or combining power in a waveguide system. There are basically 2 types of tees namely 1.H- plane Tee junction 2.E-plane Tee junction A combination of these two tee junctions is called a hybrid tee or " Magic Tee". E-plane Tee(series tee): An E-plane tee is a waveguide tee in which the axis of its side arm is parallel to the E field of the main guide . if the collinear arms are symmetric about the side arm.

If the E-plane tee is perfectly matched with the aid of screw tuners at the junction , the diagonal components of the scattering matrix are zero because there will be no reflection.

When the waves are fed into side arm, the waves appearing at port 1 and port 2 of the collinear arm will be in opposite phase and in same magnitude.

2.8 Magic Tee (Hybrid Tees)

A magic tee is a combination of E-plane and H-plane tee. The characteristics of magic tee are:



1. If two waves of equal magnitude and same phase are fed into port 1 and port 2 the output will be zero at port 3 and additive at port 4.

3. If a wave is fed into port 4 it will be divided equally between port 1 and port 2 of the collinear arms and will not appear at port 3.

4. If a wave is fed into port 3 , it will produce an output of equal magnitude and opposite phase at port 1 and port 2. the output at port 4 is zero.

5. If a wave is fed into one of the collinear arms at port 1 and port 2, it will not appear in the other collinear arm at port 2

OUTCOME:

At the end of the course student will be able to understand and explain the different microwave passive devices and S-parameters.

Recommended Questions

1. What is S-parameter?
2. What is the need for S-parameter
3. Explain Magic-Tee
4. Explain Phase shifter
5. Explain Attenuators.
6. Explain Microwave Passive devices

Module 3: ANTENNA BASICS

Structure

3.0 INTRODUCTION

An antenna is used to radiate electromagnetic energy efficiently and in desired directions. Antennas act as matching systems between sources of electromagnetic energy and space. Antenna is a source or radiator of Electromagnetic waves or a sensor of Electromagnetic waves. It is a transition device or transducer between a guided wave and a free space wave or vice versa. It is also an electrical conductor or system of conductors that radiates EM energy into or collects EM energy from free space. Antennas function by transmitting or receiving electromagnetic (EM) waves. Examples of these electromagnetic
waves include the light from the sun and the waves received by your cell phone or radio. Your eyes are basically "receiving antennas" that pick up electromagnetic waves that are of a particular frequency. The colors that you see (red, green, blue) are each waves of different frequencies that your eyes can detect. All electromagnetic waves propagate at the same speed in air or in space. This speed (the speed of light) is roughly 671 million miles per hour (1 billion kilometers per hour). This is roughly a million times faster than the speed of sound (which is about 761 miles per hour at sea level). The speed of light will be denoted as c in the equations that follow. We like to use "SI" units in science (length measured in
meters, time in seconds, mass in kilograms):

A rough outline of some major antennas and their discovery /fabrication dates are listed:

- Yagi-Uda Antenna,1920s
- Horn Antennas,1939
- Antenna Arrays 1940s

- Parabolic Reflectors late 1940s, early 1950s.
- Patch Antennas,1970s
- PIFAs 1980s.

3.1 OBJECTIVES:

1. Introduction about the antenna parameters in terms of antenna language
2. Antenna field zones utilization.
3. Link budget calculation for any communication link.

3.2 BASIC ANTENNA PARAMETERS

A radio antenna may be defined as the structure associated with the region of transition between a guided wave and a free space wave or vice versa.

Principle: Under time varying conditions, Maxwell's equations predict the radiation of EM energy from current source (or accelerated charge). This happens at all frequencies, but is insignificant as long as the size of the source region is not comparable to the wavelength. While transmission lines are designed minimize this radiation loss, radiation into free space becomes main purpose in case of Antennas. The basic principle of radiation is produced by accelerated charge. The basic equation of radiation is

$$\mathbf{I\ L = Q\ V \quad (Ams\text{-}1) \qquad (1)}$$

where, I = Time changing current in Amps/sec

L = Length of the current element in meters

Q = Charge in Coulombs

V = Time changing velocity

Thus time changing current radiates and accelerated charge radiates. For steady state harmonic variation, usually we focus on time changing current. For transients or pulses, we focus on accelerated charge. The radiation is perpendicular to the acceleration and the radiated power is proportional to the square of IL or QV. Transmission line opened out in a Tapered fashion as Antenna:

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

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

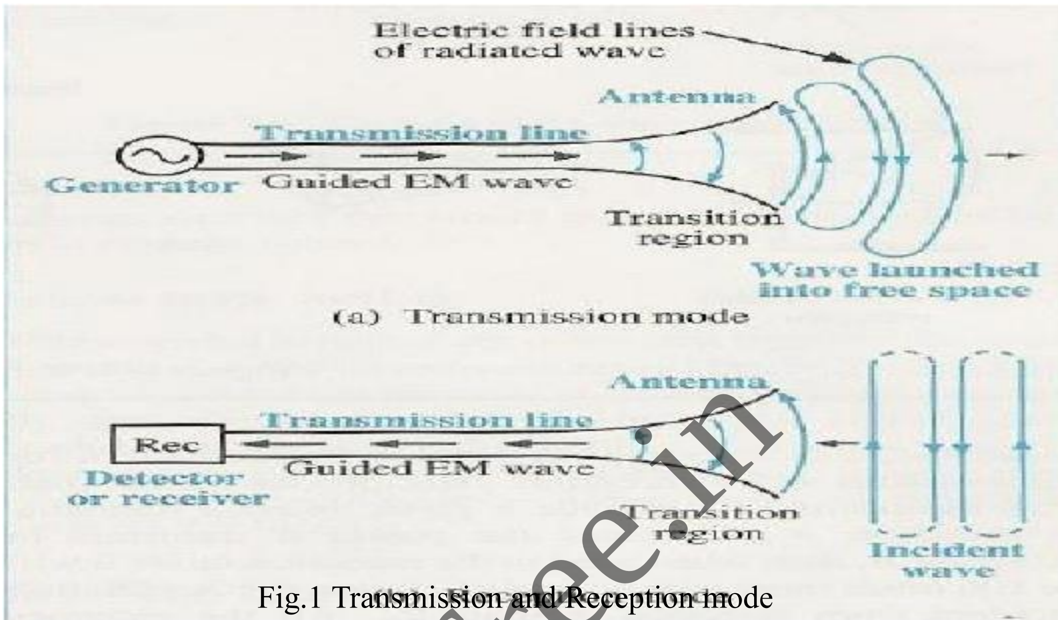


Fig.1 Transmission and Reception mode

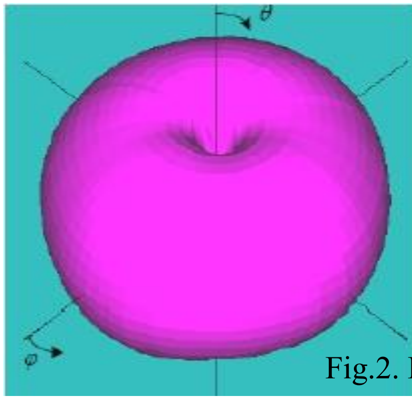
Thus, an antenna is a transition device, or transducer, between a guided wave and a free space wave or vice versa. The antenna is a device which interfaces a circuit and space.

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

3.3 PATTERNS

The radiation pattern or antenna pattern is the graphical representation of the radiation properties of the antenna as a function of space. That is, the antenna's pattern describes how the antenna radiates energy out into space (or how it receives energy). It is important to state that an antenna can radiate energy in all directions, so the antenna pattern is actually three-dimensional. It is common, however, to describe this 3D pattern with two planar patterns, called the principal plane patterns. These principal plane patterns can be obtained by making two slices through the 3D pattern, through the maximum value of the
pattern. It is these principal plane patterns that are commonly referred to as the antenna patterns.

Radiation pattern or Antenna pattern is defined as the spatial distribution of a ‘quantity’ that characterizes the EM field generated by an antenna. The ‘quantity’ may be Power, Radiation Intensity, Field amplitude, Relative Phase etc



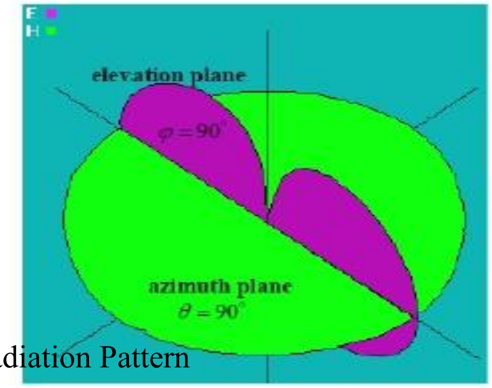


Fig.2. Radiation Pattern

Always the radiation has Main lobe through which radiation is maximum in the z direction and Minor lobe (side and back lobes) in the x and y direction. Any field pattern is presented by 3D spherical coordinates or by plane cuts through main lobe axis. Two plane cuts as right angles are called as principal plane pattern. To specify the radiation pattern with respect to field intensity and polarization requires three patterns:

i. The θ component of the electric field as a function of the angles θ and Φ or $E_\theta(\theta,\Phi)$ in Vm-1.
ii. The Φ component of the electric field as a function of the angles θ and Φ or $E_\Phi(\theta,\Phi)$ in Vm-1.
iii. The phases of these fields as a function of the angles θ and Φ or $\delta_\theta(\theta,\Phi)$ and $\delta_\Phi(\theta,\Phi)$ in radian or degree.

Normalized field pattern: It is obtained by dividing a field component by its maximum value. The normalized field pattern is a dimensionless number with maximum value of unity

$$E_\Phi(\theta,\Phi)\, n = E_\Phi(\theta,\Phi) / E_\Phi(\theta,\Phi)\, max \qquad (2)$$

Half power level occurs at those angles (θ,Φ)for which $E\theta(\theta,\Phi)n = 0.707$. At distance d>>λ and d>> size of the antenna, the shape of the field pattern is independent of the distance

Normalized power pattern: Pattern expressed in terms of power per unit area is called power pattern. Normalizing the power with respect to maximum value yields normalized power patterns as a function of angle which is dimensionless and maximum value is unity.

$$Pn(\theta,\Phi)n = S(\theta,\Phi)/ S(\theta,\Phi)max \qquad (3)$$

Where, $S(\theta,\Phi)$ is the Poynting vector = $[\, E_\theta^2(\theta,\Phi) + E_\Phi^2(\theta,\Phi)\,] / Z0$ Wm-2

$S(\theta,\Phi)max$ is the maximum value of $S(\theta,\Phi)$, Wm-2

Z0 is the intrinsic impedance of free space = 376.7Ω.

Decibel level is given by dB = 10 log10 Pn(θ,Φ)

Half power levels occurs at those angles (θ,Φ) for which P(θ,Φ)n =0.5.

Pattern Lobes and Beamwidths:

The radiation pattern characteristics involve three dimensional vector fields for full representation, but the scalar quantities can be used. They are:

1. Half power beam-width HPBW
2. Beam Area, ΩA
3. Bema Efficiency, εM
4. Directivity D, Gain G
5. Effective Aperture, Ae
6. Radiation Intensity

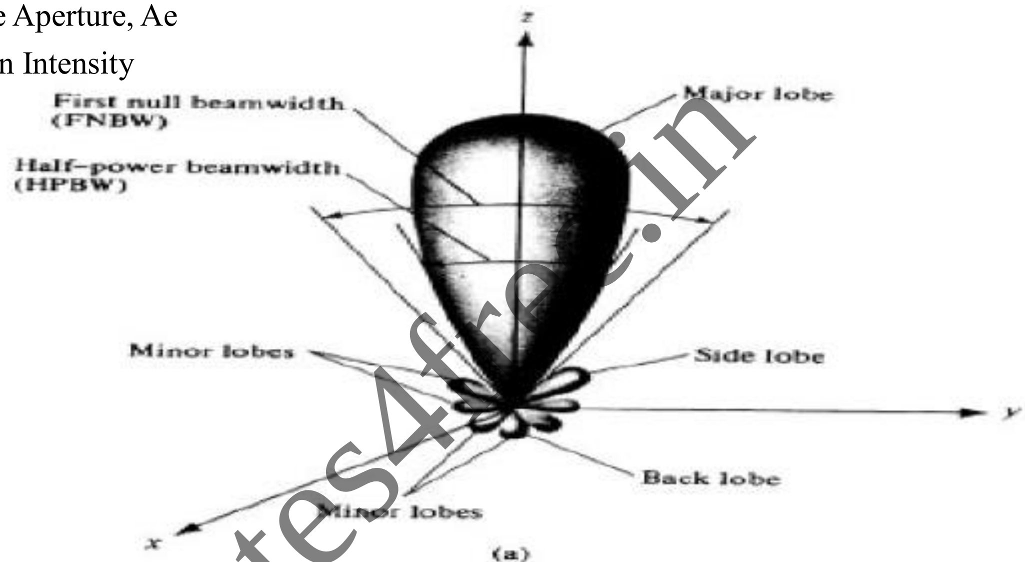


Fig.3 Pattern in spherical co-ordinate system

Beamwidth is associated with the lobes in the antenna pattern. It is defined as the angular separation between two identical points on the opposite sides of the main lobe. The most common type of beamwidth is the half-power (3 dB) beamwidth (HPBW). To find HPBW, in the equation, defining the radiation pattern, we set power equal to 0.5 and solve it for angles. Another frequently used measure of beamwidth is the first-null beamwidth (FNBW), which is the angular separation between the first nulls on either sides of the main lobe.

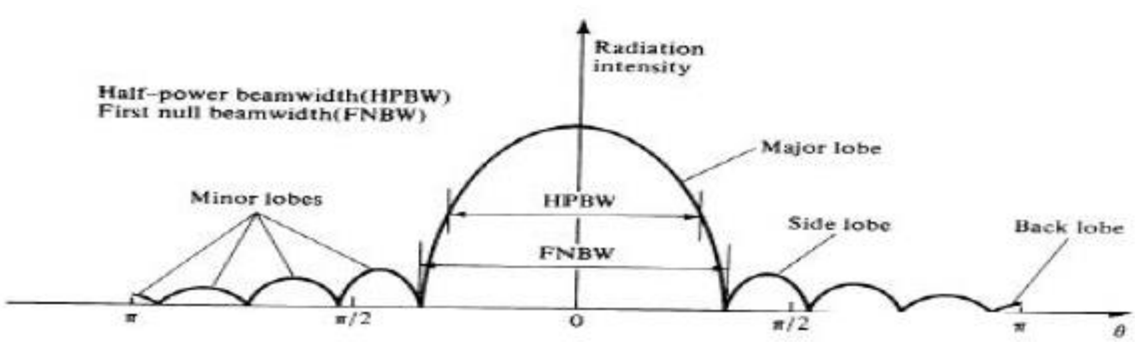


Fig.4 Pattern in Cartesian co-ordinate system

Beamwidth:

Antenna Beam-width is the measure of directivity of an antenna. The antenna beamwidth is the angular width expressed in degrees which is measured on the major lobe of the radiation pattern of an antenna.

HPBW:

The angular width on the major lobe of radiation pattern between two points where the power is half of the maximum radiated power is called Half Power Beam-width. Here the power decreases to half of its maximum value.

FNBW:

When the angular width is measured between the first nulls or first side lobes it is called First Null Beam Width.

The factors affecting beam width are:

1. Shape of the radiation pattern.
2. Dimensions of antenna.
3. Wavelength.

Beam width defines the resolution capability of the antenna, i.e., the ability of the system to separate two adjacent targets.

The beam solid angle of an antenna is given by the integral of the normalized power pattern over a sphere (4π steradians).

$$\Omega A = \int_{\Phi=0}^{\Phi=2\pi} \int_{\theta=0}^{\theta=2\pi} Pn(\theta, \Phi) \sin\theta \, d\theta \, d\Phi \qquad (4)$$

Beam area ΩA is the solid angle through which all of the power radiated by antenna would stream if P(θ, Φ) maintained its maximum value over ΩA and was zero.

$$\text{Total power radiated} = P(\theta, \Phi)\, \Omega A \text{ watts}$$

Beam area is the solid angle ΩA is often approximated in terms of the angles subtended by the Half Power points of the main lobe in the two principal planes(Minor lobes are neglected)

$$\Omega A = \theta HP \; \Phi HP$$

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

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

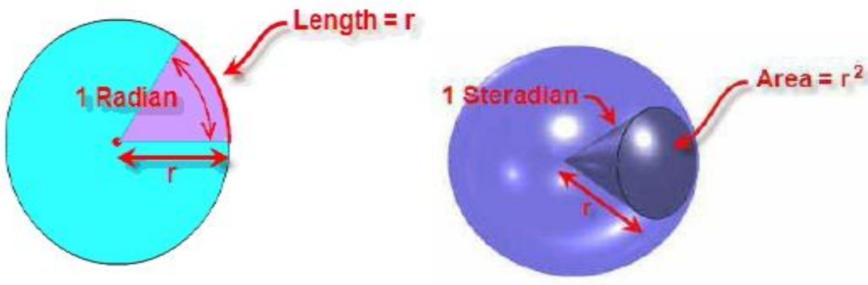


Fig.5 Beam Area

1steradian = $(1\text{radian})^2$

= $(180 / \pi)^2$

= 3282.8064 square degrees

4π steradians = 3282.8064 x 4π

= 41,253 square degree

The infinitesimal area ds on a surface of a sphere of radius r in spherical co-ordinates (with θ as vertical angle and Φ as azimuth angle) is

$$ds = r^2 \sin\theta \, d\theta \, d\Phi$$

By definition of solid angle: $ds = r^2 \, d\Omega$

Hence,

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

3.4 Radiation Intensity

Definition: The power radiated from an Antenna per unit solid angle is called the Radiation Intensity. "U" Units: Watts/steradians or Watts/ square degree

Poynting vector or power density is dependent on distance from the antenna while Radiation intensity is independent of the distance from the antenna. The normalized power pattern can also be expressed as the ratio of radiation intensity as a function of angle to its maximum value.

1.7 Beam Efficiency $Pn(\theta,\Phi)n = S(\theta,\Phi)/ S(\theta,\Phi)max$

3.5 Beam Efficiency

The total beam area ΩA consists of the main beam area ΩM plus the minor lobe area Ωm .

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

The ratio of main beam area to the total beam area is called the beam efficiency ε_M

$$\varepsilon_M = \Omega_M / \Omega_A$$

The ratio of minor lobe area to the total beam area is called stray factor εm

$$\varepsilon_m = \Omega_m / \Omega_A$$

3.8 Directivity D and Gain G

From the field point of view, the most important quantitative information on the antenna is the directivity, which is a measure of the concentration of radiated power in a particular direction. It is defined as the ratio of the radiation intensity in a given direction from the antenna to the radiation intensity averaged over all directions. The average radiation intensity is equal to the total radiated power divided by 4π. If the direction is not specified, the direction of maximum radiation is implied. Mathematically, the directivity (dimensionless) can be written as

$$D = U(\theta, \varphi)_{max} / U(\theta, \varphi)_{avg}$$

The directivity is a dimensionless quantity. The maximum directivity is always ≥ 1.

Directivity and Beam area

$$P(\theta, \phi)_{Av} = \frac{1}{4\Pi} \int_0^{2\Pi} \int_0^{\Pi} P(\theta, \phi) Sin\theta d\theta d\phi$$

$$= \frac{1}{4\Pi} \int_0^{2\Pi} \int_0^{\Pi} P(\theta, \phi) d\Omega$$

$$\therefore D = \frac{P(\theta, \phi)_{max}}{\frac{1}{4\Pi} \int_0^{2\Pi} \int_0^{\Pi} P(\theta, \phi) d\Omega}$$

$$D = \frac{1}{\frac{1}{4\Pi} \int_0^{2\Pi} \int_0^{\Pi} P_n(\theta, \phi) d\Omega}$$

$$i.e., D = \frac{4\Pi}{\Omega_A}$$

Directivity is the ratio of total solid angle of the sphere to beam solid angle. For antennas with rotationally symmetric lobes, the directivity D can be approximated as:

$$D = 4\pi / \theta \Phi$$

Directivity of isotropic antenna is equal to unity, for an isotropic antenna Beam area $\Omega A = 4\pi$

Directivity indicates how well an antenna radiates in a particular direction in comparison with an isotropic antenna radiating same amount of power Smaller the beam area, larger is the directivity

Gain: Any physical Antenna has losses associated with it. Depending on structure both ohmic and dielectric losses can be present. Input power Pin is the sum of the Radiated power Prad and losses Ploss

$$P_{in}=P_{rad} + P_{loss}$$

The Gain G of an Antenna is an actual or realized quantity which is less than Directivity D due to ohmic losses in the antenna. Mismatch in feeding the antenna also reduces gain. The ratio of Gain to Directivity is the Antenna efficiency factor k (dimensionless)

$$G = KD, where\ 0 \leq K \leq 1$$

In practice, the total input power to an antenna can be obtained easily, but the total radiated power by an antenna is actually hard to get. The gain of an antenna is introduced to solve this problem. This is defined as the ratio of the radiation intensity in a given direction from the antenna to the total input power accepted by the antenna divided by 4π. If the direction is not specified, the direction of maximum radiation is implied. Mathematically, the gain (dimensionless) can be written as

$$G=4\pi U/Pin$$

Directivity and Gain: Directivity and Gain of an antenna represent the ability to focus it's beam in a particular direction. Directivity is a parameter dependent only on the shape of radiation pattern while gain takes ohmic and other losses into account.

3.9 Effective Aperture

Aperture Concept: Aperture of an Antenna is the area through which the power is radiatedor received. Concept of Apertures is most simply introduced by considering a Receiving Antenna. Let receiving antenna be a rectangular Horn immersed in the field of uniform plane wave as shown

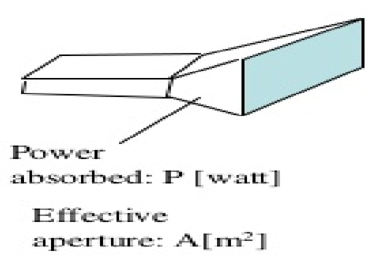


Fig.6 Aperture

Let the Poynting vector or power density of the plane wave be S watts/sq –m and let the area or physical aperture be Ap sq-m.

But the Field response of Horn is not uniform across Ap because E at sidewalls must equal zero. Thus effective Aperture Ae of the Horn is less than Ap.

Aperture Efficiency is as follows:

$$\varepsilon_{ap}=Ae/Ap$$

The effective antenna aperture is the ratio of the available power at the terminals of the antenna to the power flux density of a plane wave incident upon the antenna, which is matched to the antenna in terms of polarization. If no direction is specified, the direction of maximum radiation is implied. Effective Aperture (Ae) describes the effectiveness of an Antenna in receiving mode, It is the ratio of power delivered to receiver to incident power density.

It is the area that captures energy from a passing EM wave an Antenna with large aperture (Ae) has more gain than one with smaller aperture(Ae) since it captures more energy from a passing radio wave and can radiate more in that direction while transmitting

Effective Aperture and Beam area: Consider an Antenna with an effective Aperture Ae which radiates all of it's power in a conical pattern of beam area ΩA, assuming uniform field Ea over the aperture, power radiated is

$$P=\frac{E_a^2}{z_0}A_e$$

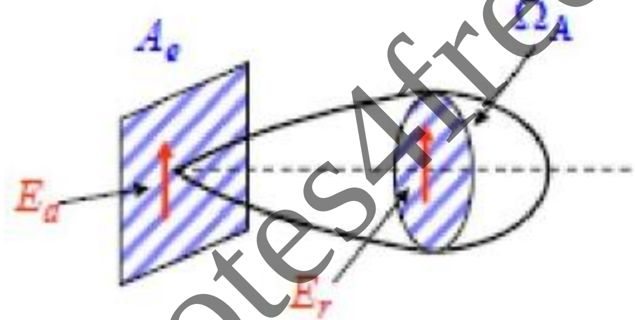


Fig.7 Effective Aperture

Assuming a uniform field Er in far field at a distance r. Power radiated is also given by

$$P=Er^2/Z0\ r^2\ \Omega A$$

Equating the two and noting that Er = Ea Ae/r λ we get Aperture Beam Area relation

$$\Lambda 2 = Ae\ \Omega A$$

At a Given wavelength if effective aperture is known, Beam Area can be determined or vice versa

1. 10 Effective height

The effective height is another parameter related to the apertures. Multiplying the effective height, he(meters), times the magnitude of the incident electric field E (V/m) yields the voltage V induced. Thus V=he E or he= V/ E (m). Effective height provides an indication as to how much of the antenna is involved in radiating (or receiving. To demonstrate this, consider the current distributions a dipole antenna for two different lengths.

If the current distribution of the dipole were uniform, it's effective height would be l Here the current distribution is nearly sinusoidal with average value $2/\pi=0.64$(of the maximum) so that it's effective height is 0.64l .It is assumed that antenna is oriented for maximum response.

If the same dipole is used at longer wavelength so that it is only 0.1λ long, the current tapers almost linearly from the central feed point to zero at the ends in a triangular distribution. The average current is now 0.5 & effective height is 0.5l

Fig.8 Effective Height

For an antenna of radiation resistance R_r matched to it'd load , power delivered to load is

$$P = V^2/4r \text{ and}$$

voltage is given by $V=h_eE$

Therefore, $P=(h_eE)^2/(4R_r)$

In terms of Effective aperture the same power is given by

$$P=SA_e=(E^2/z_0)A_e$$

Equating the two,

$$P=\frac{h_e^2E^2}{4R_r}=\frac{E^2}{Z_0}A_e \Rightarrow h_e=\sqrt{\frac{4R_rA_e}{Z_0}}\ (\text{m}) \text{ and } A_e=\frac{h_e^2Z_0}{4R_r}\ (\text{m}^2)$$

Bandwidth or frequency bandwidth:

This is the range of frequencies, within which the antenna characteristics (input impedance, pattern) conform to certain specifications, Antenna characteristics, which should conform to certain requirements, might be: input impedance, radiation pattern, beamwidth, polarization, side-lobe level, gain, beam direction and width, radiation efficiency. Separate bandwidths may be introduced: impedance bandwidth, pattern bandwidth, etc.

The FBW of broadband antennas is expressed as the ratio of the upper to the lower frequencies, where the antenna performance is acceptable. Based on Bandwidth antennas can be classified as

1. Broad band antennas-BW expressed as ratio of upper to lower frequencies of acceptable operation eg: 10:1 BW means fH is 10 times greater than fL

2. Narrow band antennas-BW is expressed as percentage of frequency difference over center frequency eg:5% means (fH –fL) /fo is .05. Bandwidth can be considered to be the range of frequencies on either sides of a center frequency (usually resonant freq. for a dipole)

The FBW of broadband antennas is expressed as the ratio of the upper to the lower frequencies, where the antenna performance is acceptable

$$FBW = \frac{f_{max}}{f_{min}}$$

Broadband antennas with FBW as large as 40:1 have been designed. Such antennas are referred to as frequency independent antennas. For narrowband antennas, the FBW is expressed as a percentage of the frequency difference over the center frequency

$$\text{FBW} = \frac{f_{\max} - f_{\min}}{f_0} \cdot 100\ \%.$$

Usually, $f_0 = (f_{\max} + f_{\min})/2$ or $f_0 = \sqrt{f_{\max} f_{\min}}$.

The characteristics such as Zi, G, Polarization etc of antenna does not necessarily vary in the same manner. Sometimes they are critically affected by frequency Usually there is a distinction made between pattern and input impedance variations. Accordingly, pattern bandwidth or impedance bandwidth are used. Pattern bandwidth is associated with characteristics such as Gain, Side lobe level, Polarization, Beam area. (large antennas) Impedance bandwidth is associated with characteristics such as input impedance, radiation efficiency (Short dipole) Intermediate length antennas BW may be limited either by pattern or impedance variations depending on application If BW is Very large (like 40:1 or greater), Antenna can be considered frequency independent.

3.11 Radiation Efficiency

Total antenna resistance is the sum of 5 components

Rr+Rg+Ri+Rc+Rw

Where, Rr is Radiation resistance

Rg is ground resistance

Ri is equivalent insulation loss

Rc is resistance of tuning inductance

Rw is resistance equivalent of conductor loss

Radiation efficiency=Rr/(Rr+Rg+Ri+Rc+Rw). It is the ratio of power radiated from th eantenna to the total power supplied to the antenna

Antenna temperature

The antenna noise can be divided into two types according to its physical source:

- noise due to the loss resistance of the antenna itself; and
- noise, which the antenna picks up from the surrounding environment The noise power per

unit bandwidth is proportional to the object's temperature and is given by Nyquist's relation

$$p_h = kT_P, \text{ W/Hz}$$

where

TP is the physical temperature of the object in K (Kelvin degrees); and k is

Boltzmann's constant (1.38×10^{-23} J/K

A resistor is a thermal noise source. The noise voltage (rms value) generated by a resistor

R, kept at a temperature T, is given by

$$V_n = \sqrt{4kTBR}$$

Where,

k is Boltzmann's constant (1.38×10^{-23} J/K). And

B is the bandwidth in Hz

Often, we assume that heat energy is evenly distributed in the frequency band Δf .

Then, the associated heat power in Δf is

$$P_h = kT_P \Delta f, \text{ W}.$$

For a temperature distribution $T(\theta, \Phi)$ and radiation pattern $R(\theta, \Phi)$ of the antenna ,

Then noise temperature T_A is given by

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

The noise power P_{TA} received from an antenna at temperature T_A can be expressed in terms of Bandwidth B over which the antenna(and it's Receiver) is operating as

$$P_{TA} = kT_A B$$

The receiver also has a temperature TR associated with it and the total system noise temperature (i.e., Antenna + Receiver) has combined temperature given b

$$T_{sys} = T_A + T_R$$

And total noise power in the system is

$$P_{Total} = kT_{sys} B$$

3.12 THE RADIO COMMUNICATION LINK

The usefulness of the aperture concept is well illustrated by using it to derive the important Friis transmission formula published in 1946 by Harald T. Friis (1) of the Bell Telephone Laboratory.

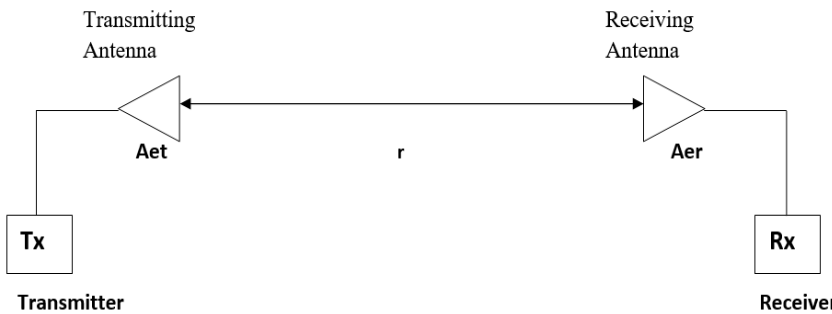


Fig: Radio communication link

Referring to Fig.9, the formula gives the power received over a radio communi- cation link. Assuming lossless, matched antennas, let the transmitter feed a power P_t to a transmitting antenna of effective aperture A_{et} . At a distance r a receiving antenna of effective aperture A_{er} intercepts some of the power radiated by the transmitting antenna and delivers it to the receiver R. Assuming for the moment that the transmitting antenna is isotropic, the power per unit area available at the receiving antenna is

$$S_r = \frac{P_t}{4\pi r^2}$$

If the antenna has gain G_t , the power per unit area available at the receiving antenna will be increased in proportion as given by

$$S_r = \frac{P_t G_t}{4\pi r^2}$$

Now the power collected by the lossless, matched receiving antenna of effective aperture

Aer is

$$P_t = S_r A_{er} = \frac{P_t G_t A_{er}}{4\pi r^2}.$$

The gain of the transmitting antenna can be expressed as

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

Substituting this in the previous equation yields the ***Friis transmission formula***

$$P_r = \frac{A_{et} A_{er} P_t}{r^2 \lambda^2}$$

Example: **Radio Communication Link**

A radio link has a 15-W transmitter connected to an antenna of 2.5 m^2 effective aperture at 5 GHz. The receiving antenna has an effective aperture of 0.5 m^2 and is located at a 15-km line-of-sight distance from the transmitting antenna. Assuming lossless, matched antennas, find the power delivered to the receiver. **Answer P_r=23uWatts**

3.13 ANTENNA FIELD ZONES

The fields around an antenna may be divided into two principal regions, one near the antenna called the near field or Fresnel zone and one at a large distance called the far field or Fraunhofer zone. Referring to Fig. 2–17, the boundary between the two may be arbitrarily taken to be at a radius

$R=2L^2/\lambda$

L = maximum dimension of the antenna, m

λ = wavelength, m

In the far or Fraunhofer region, the measurable field components are transverse to the ra- dial direction from the antenna and all power flow is directed radially outward. In the far field the shape of the field pattern is independent of the distance. In the near or Fresnel region, the longitudinal component of the electric field may be significant and power flow is not entirely radial. In the near field, the shape of the field pattern depends, in general, on the distance.

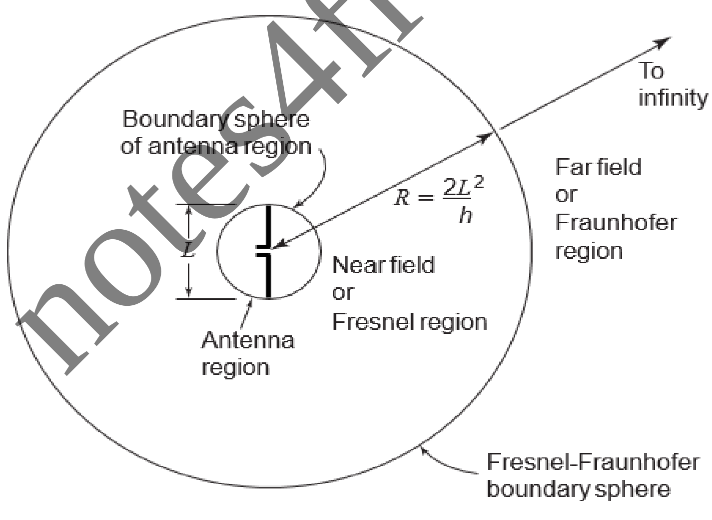


Fig 9. Antenna field zones

Enclosing the antenna in an imaginary boundary sphere as in Fig. 9 it is as though the region near the poles of the sphere acts as a reflector. On the other hand, the waves expanding perpendicular to the dipole in the equatorial region of the sphere result in power leakage through the sphere as if partially transparent in this region.

This results in reciprocating (oscillating) energy flow near the antenna accompanied by outward flow in the equatorial region. The outflow accounts for the power radiated from the antenna, while the reciprocating energy represents reactive power that is trapped near the antenna like in a resonator

Note that although the term *power flow* is sometimes used, it is actually *energy* which flows, power being the time rate of energy flow. A similar loose usage occurs when we say we pay a power bill, when, in fact, we are actually paying for electric energy.

Near Field (Fresnel's Region)

- Power flow is not entirely radiated.
- Shape of field pattern is dependent of radial distance.
- There is an energy reciprocating b/n antenna and space.
- Reactive energy.

Far field (Fraunhofer's region)

- Real power flow is directed radially outwork.
- Shape of the field pattern is independent of distance.
- The outward power flow represents radiated energy.

Measurable field components are transverse to the direction of propagation.

3.14 Recommended questions

1. With the help of Maxwell's equation, explain how radiation and reception of EM takes place?

2. Explain the following terms as related to antenna system:

(1) Directivity (2) HPBW (3) Effective length (4) Beam efficiency (5) Gain (6) Isotropic radiator (7) Beam area/Beam solid angle (8) Radiation resistance

3. Show that the directivity for unidirectional operation is 2(n+1) for an intensity variation of $u=u^m \cos^n\theta$.

4. Prove that maximum effective aperture for a λ/2 antenna is 0.13 λ2.

5. The effective aperture of transmitting and receiving antennas in a communication system are 8 λ2 and 12 λ2 respectively with a separation of 1.5 km between them. The E.M wave is travelling with a frequency of 6MHz and the total input power is 25KW. Find the power received by the receiving antenna.

6. Define the following with respect to antenna:

(1) Radiation pattern (power and field pattern) (2) field zones (3) Aperture

3.15 OUTCOMES

- Student will able to define the parameters and importance of all in communication systems
- Student able to solve link budget problems required for the applications
- Student able to describe the importance of Fraunhofer zone.

3.16 Further Readings

1. **Antenna Theory Analysis and Design** - C A Balanis, 3rd Edn, John Wiley India Pvt. Ltd, 2008

Module 4

POINT SOURCES AND ARRAYS

Structure

4.0 Introduction

As of now the antenna was treated aperture. in this chapter it is formally considered as point source and later the concept extended to the formation of arrays of point sources. The pattern of any antenna can be regarded as produced by an array of point sources.

Here we discuss the array arrays confined to isotropic point sources, which may represent different kind of antennas.

Point Sources:

***** Antenna that doesn't have any specified shape is called "point source".

Consider an antenna and observation circle as shown in fig.2.1 where the radiated fields of antenna transverse radially at a sufficient distance id far field whereas near fields have actual variation ignored.

Provided that observation made at the sufficient distance, any antenna regardless of size or complexity can be represented as single point source. Far field is considered because power flow and fields are radiated outwards at this region properly.

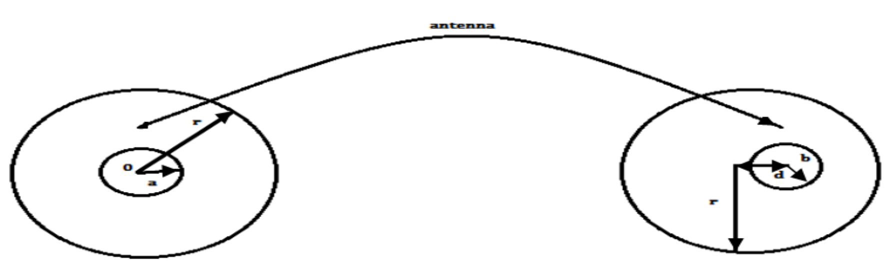


Fig 4.1. Antenna and the observation

Field measurements can be done either by fixing antenna or fixing observation point but both effects are approximately same.

If in case the center of the antenna is displaced by distance 'd' as shown in fig.1, the distance between two centers are negligible effect on the field pattern at observation circle provided that

R >> **d , R** >> **b** and **R** >> **λ**

As we discussed complete description of the far field of a source requires 3 components.

1. $\mathbf{E_\theta(\theta, \phi)}$
2. $\mathbf{E_\phi(\theta, \phi)}$
3. $\boldsymbol{\delta(\theta, \phi)}$

Power Patterns:

Let transmitting antenna in free space by point source radiation located at origin of the co-ordinates as shown in fig.4.2.

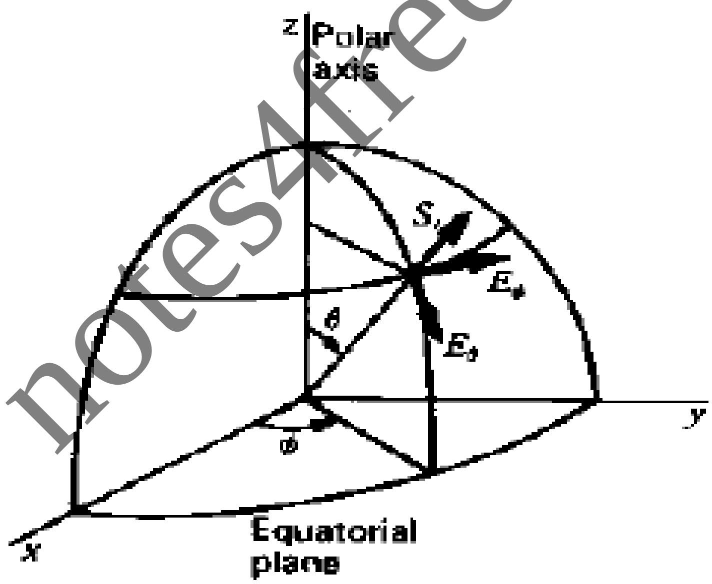


Fig 4.2 Point source at origin.

The radiated energy streams from the source radial lines.

The time rate of energy flow per unit area is "**Poynting vector or power flow density**". The magnitude of Poynting vector is equal t radial component ($|\mathbf{S}| = \mathbf{S_r}$)

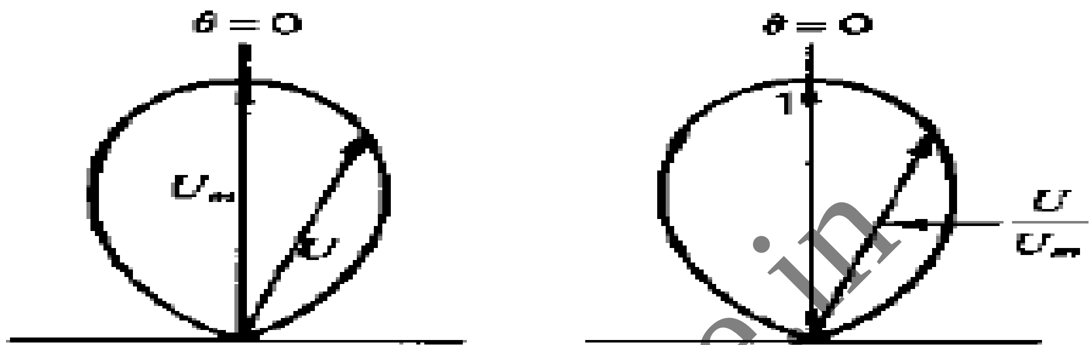
* A sources that radiates energy uniformly in all directions is called "**isotropic antenna**".

A graph of S_r at constant radius as a function of angle is Poynting vector, power density, pattern but usually called "**power pattern**".

Although the isotropic source is convenient in theory, it is not physically realizable type.

Even simplest antenna has bidirectional properties i.e., they radiate energy in some directions than others.

- In contrast with isotropic antennas, they might be called as "**anisotropic antennas**".
- If S_r is expressed in W/m^2 , the absolute power pattern. On the other hand if it express with its reference value then the graph is called "**relative power pattern**".



S_{rm} - maximum power in the direction.

A pattern with a maximum of unity is called "**normalized pattern**".

4.1 Objectives

- Apply the power theorem to solve the problems
- Analyse the arrays of point source and their patterns
- Analyse the different conditions and importance of various types of arrays

4.2 Power Theorem

If the Poynting vector is known at all points on a sphere of radius **r** from a point source in a lossless medium, the total power radiated by the source is the integral over the surface of the sphere of the radial components **S_{ref}** the average poynting vector". Thus

$$\mathbf{P = \oiint S \cdot ds}$$

Where , P - power radiated (**W**)

S_r - radial component of average poynting vector (Wm^{-2})

ds - infinitesimal element of area of sphere

$$ds = r^2 \, Sin\theta \, d\theta \, d\phi \, (m^2)$$

For an isotropic source, S_r is independent of **θ and ϕ** so

$$P = S_r \oiint ds$$

$$S_r = \frac{P}{4\pi r^2}$$

This equation indicates that the magnitude of Poynting vector varies inversely proportional to the square of the distance from point source radiator.

Radiation Intensity:

It is defined as the power radiated by an antenna per unit solid angle. Denoted by **u** and unit is W/Sr.

Power Theorem for Radiation Intensity:

The total power radiated by an antenna is given by integral of radiation intensity over solid angle of the sphere.

i.e., $$P = \oiint u \, d\Omega$$

For an isotropic source radiation intensity remains same at any point or the surface of the sphere.

Let u_o be the radiation of isotropic source then

$$P = \oiint u_o \, d\Omega$$

$$P = \oiint u_o \sin\theta \, d\theta \, d\phi$$

$$P = u_o \int_0^{\pi} \sin\theta \, d\theta \int_0^{2\pi} d\phi$$

$$P = 2\pi u_o(-\cos\pi - (-\cos 0))$$

$$P = u_o (4\pi)$$

$$u_o = \frac{P}{4\pi}$$

Therefore, the relation between Poynting vector and radiation intensity as follows

W.K.T. $$S_r = \frac{P}{4\pi r^2}$$

$$S_r r^2 = \frac{P}{4\pi}$$

$$S_r r^2 = u_o$$

$$u_o = S_r r^2$$

Examples Of Power Patterns:

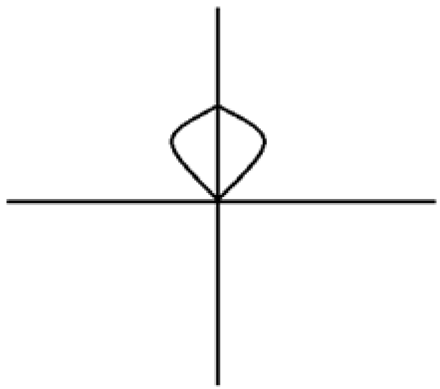
1. **Unidirectional Cosine Pattern**

The radiation intensity of unidirectional cosine pattern is given as

$$u = u_m \cos\theta$$

where u_m is the maximum radiation intensity and u is having value in upper hemisphere.

i.e.,
$$u = \begin{cases} u_m \cos\theta \,; & 0 < \theta < \frac{\pi}{2} \\ & 0 < \theta < \pi \\ 0; & Elsewhere \end{cases}$$



2. **Bidirectional Cosine Pattern**

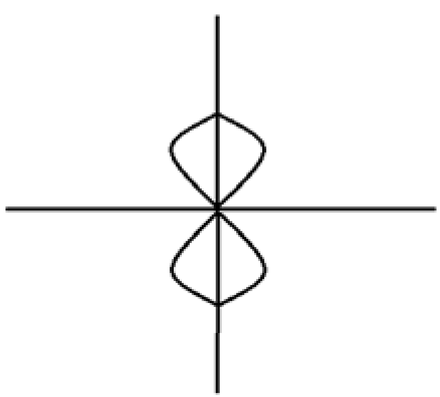
The radiation intensity of unidirectional cosnine pattern is given as

$$u = u_m \cos\theta$$

and has the value in both the hemisphere

i.e.,
$$u = \begin{cases} u_m \cos\theta \,; & 0 < \theta < \pi \\ & 0 < \theta < \pi \\ 0; & Elsewhere \end{cases}$$

It is also known as groundnut pattern because of its appearance as shown.

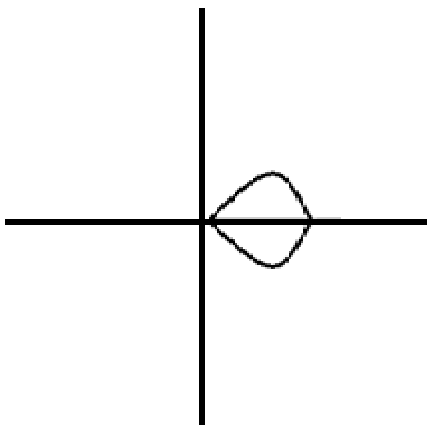


3.**Unidirectional Sine Pattern**

The radiation intensity is given as

i.e.,
$$u=\begin{cases} u_m \sin\theta \,;\; 0<\theta<\frac{\pi}{2} \\ \qquad\qquad 0<\theta<\pi \\ 0;\; Elsewhere \end{cases}$$

The maximum radiation intensity at $\theta = \frac{\pi}{2}$

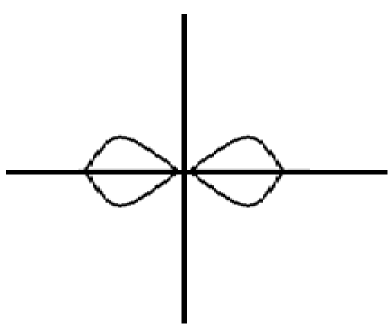


4.**Bidirectional Sine Pattern**

The radiation intensity is given as

i.e.,
$$u=\begin{cases} u_m \sin\theta \,;\; 0<\theta<\pi \\ \qquad\qquad 0<\theta<\pi \\ 0;\; Elsewhere \end{cases}$$

It is also known as doughnut pattern and pattern as shown.



2.3 Field pattern

A pattern showing variation of the electric field intensity at a constant radius r as a function of angle(θ , φ) is called **"field pattern"**

The power pattern and the field patterns are inter-related:

$$P(\theta, \varphi) = (1/\eta)*|E(\theta, \varphi)|2 = \eta*|H(\theta, \varphi)|2$$

P = power
E = electrical field component vector
H = magnetic field component vector
η = 377 ohm (free-space impedance)

The power pattern is the measured (calculated) and plotted received power: $|P(\theta, \varphi)|$ at a constant (large) distance from the antenna

The amplitude field pattern is the measured (calculated) and plotted electric (magnetic) field intensity, $|E(\theta, \varphi)|$ or $|H(\theta, \varphi)|$ at a constant (large) distance from the antennas

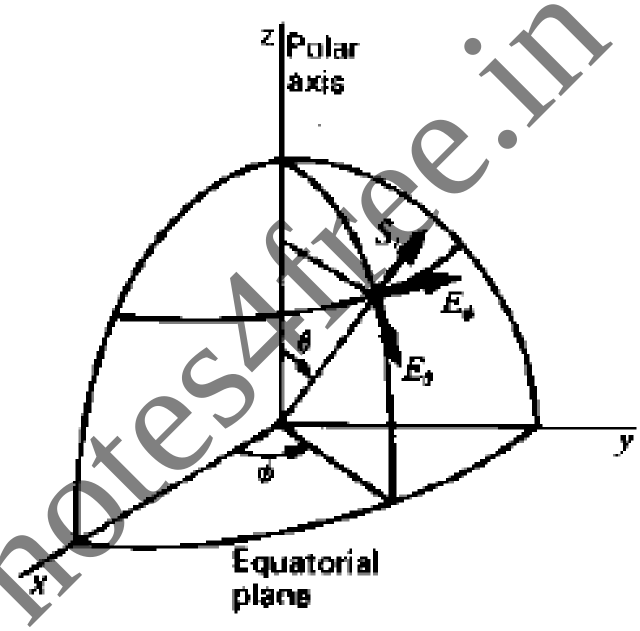


Fig 4.3: Relation of poynting vector s and 2 electric field components of a far field

4.4 ARRAY OF TWO POINT SOURCES

ARRAY is an assembly of antennas in an electrical and geometrical of such a nature that the radiation from each element add up to give a maximum field intensity in a particular direction& cancels in other directions. An important characteristic of an array is the change of its radiation pattern in response to different excitations of its antenna elements.

Here let us consider the different cases of two isotropic sources placed $\lambda/2$ apart with different scenarios.

1. **Obtain the field pattern for 2 isotropic point sources with equal amplitude and opposite phase. Assume distance between 2 sources is 'd'.**

Sol :

This case is identical with the previous but two sources are in opposite phase instead of same phase let the two sources 1 and 2 are located symmetrically with respect to origin of –ve coordinates consider a observation point p at distance 'r', the angle θ in measured clockwise from positive x-axis

if origin is considered as reference, the field from source 1 is related by $(dr/2)\cos\theta$ and field from source 2 is advanced by

$(dr/2)\cos\theta$ wr $dr = \beta d = 2\pi/2jEo * d$ ………… 1

then total electric feald in the direction at a large distance r is given

$$E=2Eo\,[\exp(j*\Psi/2 - \exp(-j*\Psi/2)]$$

From which $E=2JEo[(\exp(j*\Psi/2) - \exp(-j*\Psi/2))/2]$ …………..2

j indicates the phase reversal of one source and it is not in portent

$$E=2jEo\sin(\Psi/2)$$

$$E=2jEo\sin((dr/2)\cos\theta \text{…………….3}$$

Normalize eq3 $2jEo=1$, for $d=\pi$

$$E=\sin((\pi/2)\cos\theta) \text{…………….4}$$

Since $dr = \beta d = 2\pi/\lambda * (\lambda/2) = \pi$

In above eq

For $\theta=0$, $E=1$

$\theta=30$, $E=0.977$

$\theta=60$, $E=0.70$

$\theta=90$, $E=0$

$\theta=120$, $E=0.70$

$\theta=150$, $E=0.97$

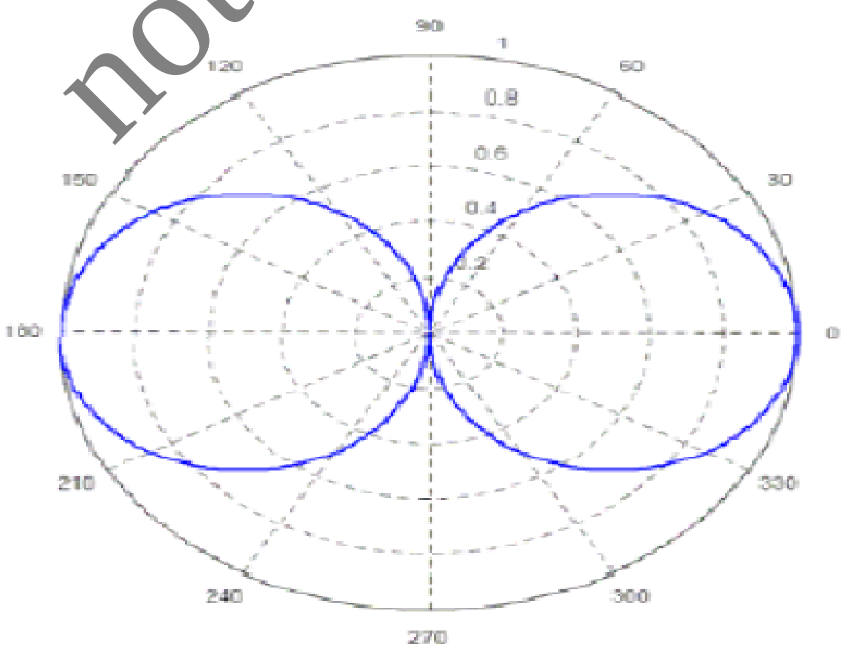
$\theta=180$, $E=-1$

$\theta=210$, $E=-0.97$

$\theta=240$, $E=-0.70$

$\theta=270$, $E=0$

$\theta=300$, $E=0.70$



θ=330, E=0.97, θ=360, E=1

4.5 ARRAY OF ‘n’ ISOTROPIC POINT SOURCES

Uniformly excited equally spaced linear arrays Linear arrays of N-isotropic point sources of equal amplitude and spacing: An array is said to be linear if the individual elements of the array are spaced equally along a line and uniform if the same are fed with currents of equal amplitude and having uniform phase shift along the line

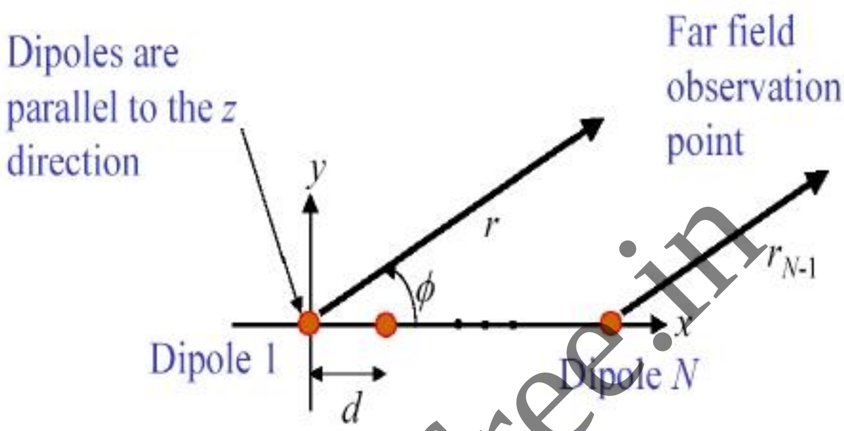


Fig.4.4 Linear arrays of N-isotropic point sources of equal amplitude and spacing:

The total field E at distance point in the direction of is given by

$$E=1+e^{j\psi}+e^{j2\psi}+e^{j2\psi}+\ldots\ldots\ldots + e^{j(n-1)\psi} \quad (1)$$

Where Ψ= total phase difference between adjacent source $\Psi = dr*\cos + \delta = 2\pi/\lambda * d*\cos\varphi + \delta$
δ = phase difference of adjacent source
multipliying equation (1) by $e^{j\varphi}$

$$E\, e^{j\psi} = e^{j\psi} + e^{j2\psi} + e^{j3\psi} + \ldots\ldots\ldots + e^{jn\psi} \quad (3)$$

Subtract (1)-(3) yields

$E(1- e^{j\psi}) = (1- e^{jn\psi})$ $E=1e^{jn\psi}/1- e^{j\psi}$

$E = e^{j(n-1)\psi/2} \{\sin(n\Psi/2)) /\sin(\Psi/2) \}$

If the phase is referred to the centre point of the array, then E redu ces to

$E=(\sin(n\Psi/2)) /\sin(\Psi/2)$

when Ψ=0 E=lim $(\sin(n\Psi/2)) /\sin(\Psi/2)$

ψ 0, E =n=Emax

Ψ=0 E=Emax= n ……….normalizing

$E_{norm} = E/E_{max} = (1/n)(\sin(n\Psi/2)) /\sin(\Psi/2)$

4.5 CASE 1: LINEAR BROAD SIDE ARRAY

An array is said to be broadside if the phase angle is such that it makes maximum radiation perpendicular to the line of array i.e. 90^0&270^0

For broad side array $\Psi=0$ & $\delta=0$

Therefore $\Psi = dr*\cos\Phi + \delta = \beta d\cos\Phi + 0 = 0$ $\Phi = \pm 90^0$

therefore $\Phi_{max} = 90^0$&270^0

Broadside array example for n=4 and d=λ/2

By previous results we have $\Phi_{max} = 90^0$&270^0

Direction of pattern maxima:

$$E=(1/n)(\sin(n\Psi/2))/\sin(\Psi/2)$$

This is maximum when numerator is maximum i.e. $\sin(n\Psi/2)=1$ $n\Psi/2= \pm(2k+1)\pi/2$
where k=0,1,2........
K=0 major lobe maxima
K=1 $n\Psi/2= \pm3\pi/2$ $\Psi= \pm3\pi/4$
Therefore $dr*\cos\Phi = 2\pi/\lambda*d*\cos\Phi = \pm3\pi/4$ $\cos\Phi = \pm3/4$

$\Phi = (\Phi_{max})_{minor\ lobe} = \cos^{-1}(\pm 3/4) = \pm41.40$ or ±138.60

At K=2, $\phi = \cos^{-1}(\pm 5/4)$ which is not possible

Direction of pattern minima or nulls

It occurs when numerator=0 i.e. $\sin(n\Psi/2)=0$ $n\Psi/2= \pm k\pi$
where k=1,2,3................................ now using condition $\delta=0$
$\Psi = \pm2k\pi/n = \pm k\pi/2$ $dr*\cos\Phi = 2\pi/\lambda*d/2*\cos\Phi$
Substituting for d and rearranging the above term $\pi\cos\Phi = \pm k\pi/2$ $\cos\Phi = \pm k/2$
Therefore $\Phi_{min} = \cos^{-1}(\pm k/2)$

k=1, $\Phi_{min} = \cos^{-1}(\pm1/2) = \pm60^0$ or $\pm120^0$

k=2, $\Phi_{min} = \cos^{-1}(\pm2/2) = \pm0^0$ or 180^0

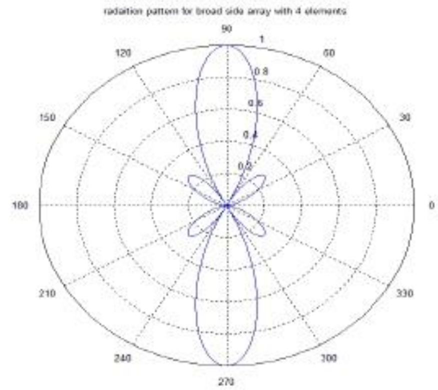


Fig.4.5 Radiation Array for Broadside Array with 4 Elements

From the pattern we see that

Beamwidth between first pair of nulls =BWFN=60 ^{0}

Half power beam width =BWFN/ 2=30^0

CASE2: END FIRE ARRAY

An array is said to be end fire if the phase angle is such that it makes maximum radiation in the line of array i.e. 0^0&180^0

For end fire array Ψ=0 & Φ =00 &1800

Therefore Ψ =dr*cos Φ +δ δ= -dr

The above result indicates that for an end fire array the phase difference b/w sources is retarded progressively by the same amount as spacing b/w the sources in radians.
If d= $\lambda/2$ δ= -dr = - $2\pi/\lambda$ x λ /2= $-\pi$
The above result indicates that source 2 lags behind source1 by π radians.

End fire array example for n=4 and d=λ/2
Direction of maxima
Maxima occurs when sin(nΨ/2)=1
i.e.$\Psi/2= \pm(2k+1)\pi/2$ where k=0,1,2........
$\Psi = \pm(2k+1)\pi/n$ dr*cos Φ +δ= $\pm(2k+1)\pi/n$
cos Φ = [$\pm(2k+1)\pi/n$ –δ]/dr

Therefore Φ_{max} =cos-1 {[$\pm(2k+1)\pi/n$ –δ]/dr}
By definition For end fire array : δ= -dr = -$2\pi/\lambda$*d
Therefore Φ_{max} =cos-1 {[$\pm(2k+1)\pi/n$ –δ]/ (-$2\pi/\lambda$*d) }
For n=4, d=λ/2 dr=π after substituting these values in above equation & solving we get
Φ_{max}=cos-1 {[$\pm$(2k+1)/4 +1} Where k=0,1,2......

For major lobe maxima,

$\Psi = 0 = dr^*\cos\Phi + \delta$
$= dr^*\cos\Phi - dr$
$= dr(\cos\Phi - 1)$ $\cos\Phi_{m=1}$ there fore $\Phi_{m=00}$ or 1800

Minor lobe maxima occurs when k=1,2,3......

K=1 $(\Phi\ max)minor1 = \cos^{-1}\{[\pm(3)/4 + 1\}$

$= \cos^{-1}(7/4 \text{ or } 1/4)$ Since $\cos^{-1}(7/4)$ is not possible

Therefore $(\Phi\ max)minor1 = \cos^{-1}(1/4) = 75.5$

K=2 $(\Phi\ max)minor2 = \cos^{-1}\{[\pm(5)/4 + 1\}$

$= \cos^{-1}(9/4 \text{ or } -1/4)$

Since $\cos^{-1}(9/4)$ is not possible

Therefore

$(\Phi\ max)minor1 = \cos^{-1}(-1/4) = 104.4$

Direction of nulls:

it occurs when numerator=0

i.e. $\sin(n\Psi/2) = 0$ $n\Psi/2 = \pm k\pi$

where k=1,2,3................................. Here $\Psi = dr^*\cos\Phi + \delta = dr(\cos\Phi - 1)$ $dr = 2\pi/\lambda^*\lambda/2 = \pi$

Substituting for d and n

$dr(\cos\Phi - 1) = \pm 2k\pi/n$ $\cos\Phi = \pm k/2 + 1$ therefore

$\Phi\ null = \cos^{-1}(\pm k/2 + 1)$

k=1 , $\Phi\ null1 = \cos^{-1}(\pm 1/2 + 1) = \cos^{-1}(3/2 \text{ or } 1/2)$

since $\cos^{-1}(3/2)$ not exist , $\Phi\ null1 = \cos^{-1}(1/2) = \pm 60$ there fore

$\Phi\ null1 = \pm 60$

k= 2,

$\Phi\ null2 = \cos^{-1}(\pm 2/2 + 1)$

$= \cos^{-1}(2 \text{ or } 0)$

since $\cos^{-1}(2)$ not exist ,

$\Phi\ null2 = \cos^{-1}(0) = \pm 90$ there fore $\Phi\ null2 = \pm 90^0$

k=3, $\Phi\ null3 = \cos^{-1}(\pm 3/2 + 1) = \cos^{-1}(5/2 \text{ or} -1/2)$

since $\cos^{-1}(5/2)$ not exist , $\Phi\ null3 = \cos^{-1}(-1/2) = \pm 120^0$ there fore, $\Phi\ null3 = \pm 120^0$

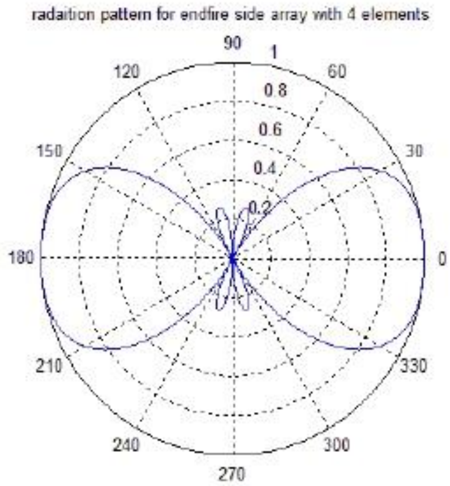


Fig.2.6 Radiation Array for End Fire Array with 5 Elements

4.6 OUTCOMES

- Able to calculate directivity for practical antennas by using the procedure
- Able to calculate major lobe minor lobe, HPBW, FNBW for two isotropic antennas, BSA, EFA different problems for given data

4.7 QUESTIONS

1. State and prove power theorem and its application.
2. Derive an expression for the power radiated from an isotropic point source with sine squared power pattern.
3. Eight point sources are spaced apart. They have a phase difference of π/3 between adjacent elements. Obtain the field pattern. Also find BWFN and HPBW.
4. Derive the expression for total field in case of two isotropic point sources with the same amplitude and equal phase. Plot the field pattern for two isotropic sources spaced apart.
5. Explain the principle of pattern multiplication.
6. Derive an expression and draw the field pattern for isotropic point sources of same amplitude and opposite phase. Also determine its maxima, minima and HPBW.
7. 4 isotropic point sources are placed apart. The power applied is with equal amplitude and a phase difference of π/3 between adjacent elements. Determine BWFN.
8. Derive the field equation for a linear array of n isotropic point sources of equal amplitude and spacing. Explain its operation as (a) broadside array (b) end fire array.

4.8 Further Readings

- **Antennas and Propagation for Wireless Communication Systems -** Sineon R Saunders, John Wiley, 2003.
- **Antennas and wave propagation -** G S N Raju: Pearson Education 2005.

Module 5

ANTENNA TYPES

5.1 INTRODUCTION

This unit describes about antenna types and their application. Types of antenna like horn antenna, helical antenna, Yagi-Uda array antenna, Log periodic antenna, reflector antennas, lens antenna are discussed. This unit also deals with the characteristics of each type of antenna and antenna application.

5.2 Objective

- To learn different types of antennas
- To learn the procedure to calculate different parameters of antennas.

5.3 Helical antenna

A helical antenna is a specialized antenna that emits and responds to electromagnetic fields with rotating (circular)polarization. These antennas are commonly used at earth-based stations in satellite communications systems. This type of antenna is designed for use with an unbalanced feed line such as coaxial cable. The center conductor of the cable is connected to the helical element, and the shield of the cable is connected to the reflector.

To the casual observer, a helical antenna appears as one or more "springs" or helixes mounted against a flat reflecting screen. The length of the helical element is one wavelength or greater. The reflector is a circular or square metal mesh or sheet whose cross dimension (diameter or edge) measures at least 3/4 wavelength. The helical element has a radius of 1/8 to ¼ wavelength, and a pitch of 1/4 to 1/2 wavelength. The minimum dimensions depend on the lowest frequency at which the antenna is to be used. If the helix or reflector is too small (the frequency is too low), the efficiency is severely degraded. Maximum radiation and response occur along the axis of the helix.

The most popular helical antenna (often called a 'helix') is a travelling wave antenna in the shape of a corkscrew that produces radiation along the axis of the helix. These helixes are referred to as axial-mode helical antennas. The benefits of this antenna is it has a wide bandwidth, is easily constructed, has a real input impedance, and can produce circularly polarized fields. The basic geometry is shown in Figure

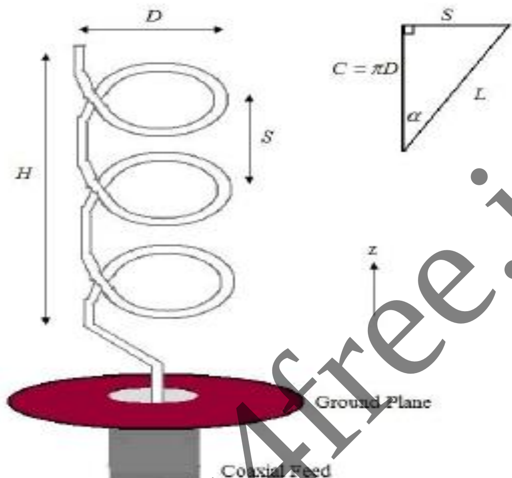


Fig 5.1 Geometry of Helical Antenna.

Fig 5.2 Helix Structure

The helix parameters are related by

$$(\pi D)^2 = L^2 - S^2$$

Let S = Spacing between each turns

N= No. of Turns

D= Diameter of the helix

L'=A=Ns=Total length of the antenna

L= Length of the wire between each turn = $\sqrt{(\pi D)^2 + s^2}$

L_n= LN = Total length of the wire

C =πD= Circumference of the helix

α = Pitch angle formed by a line tangent to the helix wire and a plane perpendicular to the helix axis.

$$\alpha = \tan^{-1}\frac{s}{c} = \tan^{-1}\frac{s}{\pi D}$$

The radiation characteristics of the antenna can be varied by controlling the size of its geometrical properties compared to the wavelength.

Mode of Operation

o Normal Mode

o Axial Mode

Normal Mode:

If the circumference, pitch and length of the helix are small compared to the wavelength, so that the current is approximately uniform in magnitude and phase in all parts of the helix, the normal mode of radiation is excited.

Fig 5.3 Normal Mode

In normal mode as shown in Fig 5.3 the radiation is maximum in the plane normal to the helix axis. The radiation may be elliptically or circularly polarized depending upon helix dimensions.

Disadvantages:

o Narrow Bandwidth

o Poor Efficiency

The radiation pattern in this mode is a combination of the equivalent radiation form a short dipole positioned along the axis of the helix and a small co-axial loop. The radiation pattern of these two equivalent radiators is the same with the polarization at right angles and the phase angle at a given point in space is at 900 apart. Therefore, the radiation is either elliptically polarized or circularly polarized depending upon the field strength ratio of the two components. This depends on the pitch angleα. When ‘α ’ is very small, the loop type of radiation predominates, when it becomes very large, the helix becomes essentially a short dipole. In these two limiting cases the polarization is linear. For intermediate value of the polarization is elliptical and at a particular value of ‘α ’ the polarization is circular

Analysis of normal mode:

Field due to short dipole is given by

$$E_\theta(\theta) = \frac{j60\pi Is\sin\theta}{\lambda r}$$

Field of a small loop

$$E_\phi(\theta) = \frac{j60\pi^2 IA\sin\theta}{\lambda^2 r}$$

Magnitude of $E_\theta(\theta)$ and $E_\phi(\theta)$ ratio defines axial ratio

$$\text{Axial ratio} = \frac{|E\theta|}{|E\phi|} = \frac{s\lambda}{2\pi A} = \frac{s}{\beta A}$$

The field is circularly polarized if $S = \beta A$

$$\therefore s = \frac{2\pi}{\lambda}\frac{\pi D^2}{4} = \frac{(\pi D)^2}{2\lambda}$$

$\frac{2s}{\lambda} = \left(\frac{\pi D}{\lambda}\right)^2$ From figure 6.1 $L^2 - s^2 = (\pi D)^2$

$$\therefore \left(\frac{L}{\lambda}\right)^2 - \left(\frac{s}{\lambda}\right)^2 = \left(\frac{\pi D}{\lambda}\right)^2 = \frac{2s}{\lambda}$$

$$1 + \left(\frac{L}{\lambda}\right)^2 = 1 + \frac{2s}{\lambda} + \left(\frac{s}{\lambda}\right)^2 = \left(1 + \frac{s}{\lambda}\right)^2$$

$$1+\frac{s}{\lambda}=\sqrt{1+\left(\frac{L}{\lambda}\right)^2}$$

$$\left(\frac{s}{\lambda}\right)=-1+\sqrt{1+\left(\frac{L}{\lambda}\right)^2}$$

This is the condition for circular polarization

The pitch angle is given by

$$\tan\alpha=\frac{s}{\pi D} \quad \text{but} \quad s=\frac{(\pi D)^2}{2\lambda}$$

$$\tan\alpha=\frac{(\pi D)^2}{2\lambda\pi D}=\frac{\pi D}{2\lambda}$$

Axial Mode:

If the dimensions of the helix are such that the circumference of one turn is approximately λ, the antenna radiates in the axial mode.

Advantages:

- Large Bandwidth and Good Efficiency
- The Radiation is circularly polarized and has a max value in the direction of helix axis.
- The directivity increase linearly with the length of the helix. It also referred as “helix beam antenna”.
- It acts like end fire array. The far field pattern of the helix can be developed by assuming that the helix consists of an array of N identical turns with an uniform spacing ‘s’ between them.

The 3db bandwidth is given by $f_{3db}=\frac{52}{C}\sqrt{\frac{\lambda^3}{NS}}$ deg

Directivity is given by $D_{max}=\frac{15N\,S\,C^2}{\lambda^3}$

N= Number of turns

C= Circumference

S=Spacing between turns

λ=Wavelength

Applications:

Used in space telemetry application at the ground end of the telemetry link for satellite and space probes at HF and VHF.

Low Frequency, Medium Frequency and High Frequency Antennas:

The choice of an antenna for a particular frequency depends on following factors.

- Radiation Efficiency to ensure proper utilization of power.
- Antenna gain and Radiation Pattern
- Knowledge of antenna impendence for efficient matching of the feeder.
- Frequency characteristics and Bandwidth
- Structural consideration

5.3 Yagi-Uda array

Yagi-Uda or Yagi is named after the inventors Prof. S.Uda and Prof. H.Yagi around 1928. The basic element used in a Yagi is $\lambda/2$ dipole placed horizontally known as driven element or active element. In order to convert bidirectional dipole into unidirectional system, the passive elements are used which include reflector and director. The passive or parasitic elements are placed parallel to driven element, collinearly placed close together as shown in Fig 5.4. The Parasitic element placed in front of driven element is called director whose length is 5% less than the drive element. The element placed at the back of driven element is called reflector whose length is 5% more than that of driver element. The space between the element ranges between 0.1λ to 0.3λ.

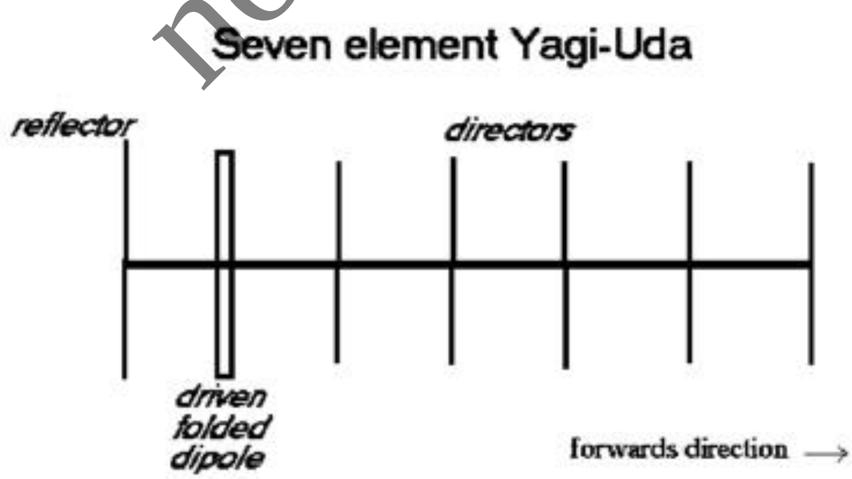


Fig 5.4 Yagi-Uda Antenna

For a three element system,

Reflector length = 500/f (MHz) feet

Driven element length = 475/f (MHz) feet

Director length = 455/f (MHz) feet.

The above relations are given for elements with length to diameter ratio between 200 to 400 and spacing between 0.1 λ to 0.2 λ. With parasitic elements the impedance reduces less than 73 Ω and may be even less than 25 Ω. A folded λ/2 dipole is used to increase the impedance. System may be constructed with more than one director. Addition of each director increases the gain by nearly 3 dB. Number of elements in a Yagi is limited to 11.

Basic Operation:

The phases of the current in the parasitic element depends upon the length and the distance between the elements. Parasitic antenna in the vicinity of radiating antenna is used either to reflect or to direct the radiated energy so that a compact directional system is obtained. A parasitic element of length greater than λ/2 is inductive which lags and of length less than λ/2 is capacitive which leads the current due to induced voltage. Properly spaced elements of length less than λ/2 act as director and add the fields of driven element. Each director will excite the next. The reflector adds the fields of driven element in the direction from reflector towards the driven element.

The greater the distance between driven and director elements, the greater the capacitive reactance needed to provide correct phasing of parasitic elements. Hence the length of element is tapered-off to achieve reactance.

A Yagi system has the following characteristics.

1.The three element array (reflector, active and director) is generally referred as "beam antenna"

2.It has unidirectional beam of moderate directivity with light weight, low cost and simplicity in design.

3. The band width increases between 2% when the space between elements ranges between 0.1λ to 0.15λ.

4. It provides a gain of 8 dB and a front-to-back ratio of 20dB.

5. Yagi is also known as super-directive or super gain antenna since the system results a high gain.

6. If greater directivity is to be obtained, more directors are used. Array up to 40 elements

can be used.

7. Arrays can be stacked to increase the directivity.

8. Yagi is essentially a fixed frequency device. Frequency sensitivity and bandwidth of about

3% is achievable.

9. To increase the directivity Yagi's can be staked one above the other or one by side of the other.

5.4 Corner reflector:

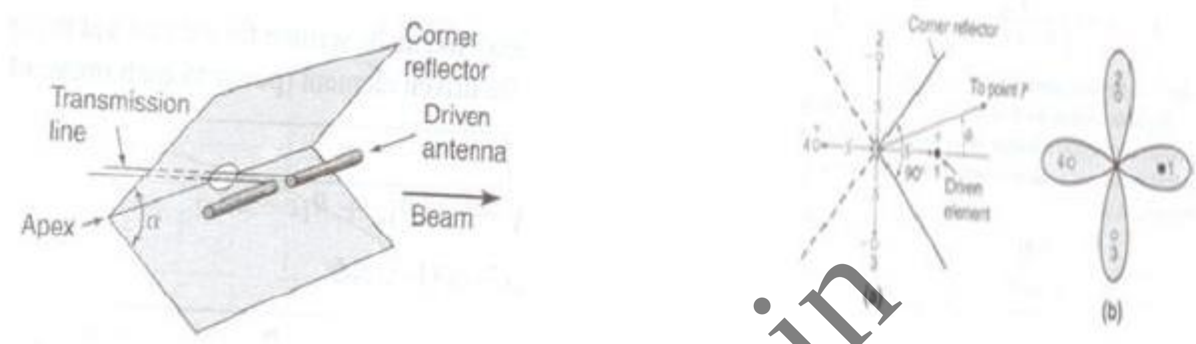


Fig. 5.5 Corner Reflector

Two flat reflecting sheets intersecting at an angle or corner as in Figure 5.5 form an effective directional antenna. When the corner angle $\alpha=90^0$, the sheets intersect at right angles, forming a square- corner reflector. Corner angles both greater or less than 90^0 can be used although there are practical disadvantages to angles much less than 90. A corner reflector with $\alpha=180^0$ is equivalent to a flat sheet reflector and may be considered as limiting case of the corner reflector. Assuming perfectly conducting reflecting sheets infinite extent, the method of images can be applied to analyze the corner reflector antenna for angle $\alpha = 180°/n$, where n is any positive integer. In the analysis of the 90° corner reflector there are 3 image elements, 2, 3 and 4, located shown in Figure. The driven antenna 1the 3 images have currents of equal magnitude. The phase of the currents in I and 4 is same. The phase of the currents in 2 and 3 is the same but 180° out of phase with respect the currents in 1and 4. All elements are assumed to be λ/2 long.

At the point P at a large distance D from the antenna. The field intensity is

$$E(\phi) = 2kI_1\left|\cos(S_r \cos\phi) - \cos(S_r \sin\phi)\right|$$

Where

I1= current in each element

Sr = spacing of each element from the corner, rad $=2\pi S/\lambda$

K=constant involving the distance D,

For arbitrary corner angles, analysis involves integrations of cylindrical functions. The emf Vt at the terminals at the center of the driven element is

$V_1 = I_1 Z_{11} + I_1 R_1 L + I_1 Z_{14} - 2 I_1 Z_{12}$

Where

Z_{11}= Self-Impedance of driven element R_1L=Equivalent loss resistance of driven element

Z_{12}=Mutual impedance of element 1 and 2

Z_{14}=Mutual impedance of element 1 and 4

If 'P' is the power delivered to the driven element, then from symmetry.

$$I_1 = \sqrt{\frac{P}{R_{11} + R_{1L} + R_{14} - 2R_{12}}}$$

$$E(\phi) = 2k\sqrt{\frac{P}{R_{11} + R_{1L} + R_{14} - 2R_{12}}}\left|\left[\cos(S_r \cos\phi) - \cos(S_r \sin\phi)\right]\right|$$

The Field Intensity at 'P' with reflector removed

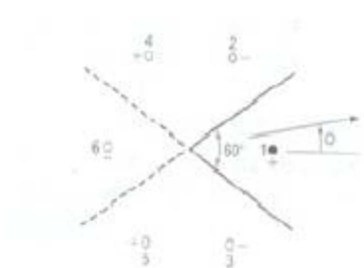
$$E_{HW}(\phi) = 2k\sqrt{\frac{P}{R_{11} + R_{1L}}}$$

The Gain in the field intensity of a square corner reflector antenna over a single λ/2 antenna

$$G_f(\phi) = \frac{E(\phi)}{E_{HW}(\phi)}$$

$$G_f(\phi) = 2\sqrt{\frac{R_{11} + R_{1L}}{R_{11} + R_{1L} + R_{14} - 2R_{12}}}\left|\left[\cos(S_r \cos\phi) - \cos(S_r \sin\phi)\right]\right|$$

Where the expression in brackets is the pattern factor and the expression included under the radical sign is the coupling factor. The pattern shape is a function of both the angle, and the antenna- to-corner spacing S. For the 60° corner the analysis requires a total of 6 elements, 1 actual antenna and 5 images as in Figure.



5.5 Parabolic reflectors

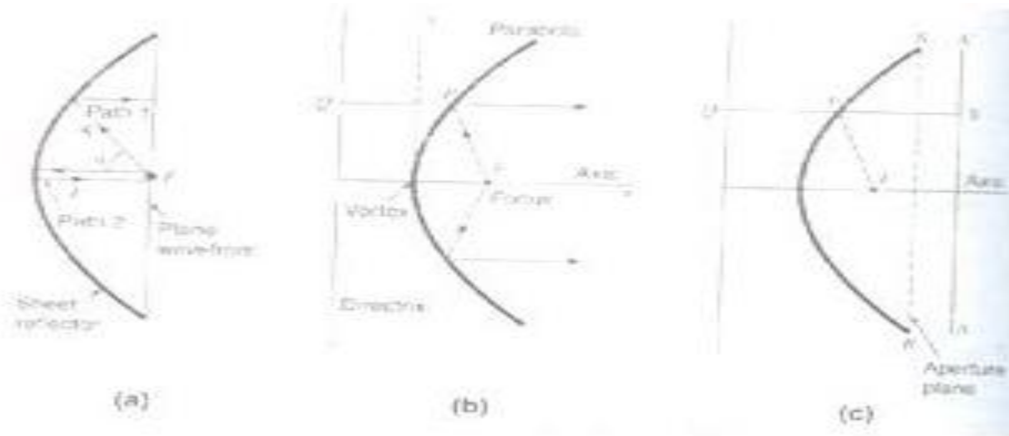


Fig 5.6 Parabolic Reflector

Suppose that we have a point source and that we wish to produce a plane-wave front over a large aperture by means of a sheet reflector. Referring to Fig(a), it is then required that the distance from the source to the plane-wave front via path 1 and 2 be equal or The parabola-general properties

$$2L = R(1 + \cos\theta)$$

$$R = \frac{2L}{1 + \cos\theta}$$

Referring to Fig. (b), the parabolic curve may be defined as follows. The distance from any point P on a parabolic curve to a fixed point F, called the focus, is equal to the perpendicular distance to a fixed line called the directrix. Thus, in Fig.(b), PF = PQ. Referring now to Fig.(c), let AA' be a line normal to the axis at an arbitrary distance QS from the directrix. Since PS = QS — PQ and PF =PQ, it follows that the distance from the focus to S is

PF+PS=PF+QS-PQ=QS

Thus, a property of a parabolic reflector is that waves from an isotropic source at the focus that are reflected from the parabola arrive at a line AA' with equal phase. The "image" of the focus is the directrix and the reflected field along the Jine A A' appears as though it originated at the directrix as a plane wave. The plane BB' (Fig. 5.6c) at which a reflector is cut off is called the aperture plane.

A cylindrical parabola converts a cylindrical wave radiated by an in-phase line source at the focus, as in Fig. 5.7a, into a plane wave at the aperture, or a paraboloid-of-revolution converts a spherical wave from an isotropic source at the focus, as in Fig. 5.7b, into a uniform plane wave at the aperture. Confining our attention to a single ray or wave path, the paraboloid has the property of directing or collimating radiation from the focus into a beam parallel to the axis. The presence of the primary antenna in the path of the reflected wave, as in the above examples, has two principle disadvantages. These are, first, that waves reflected from the parabola back to the primary antenna produce interaction and mismatching. Second, the primary antenna acts as an obstruction, blocking out the central portion of the aperture and increasing the minor lobes. To avoid both effects, a portion of the paraboloid can be used and the primary antenna displaced as in Fig below This is called an offset feed.

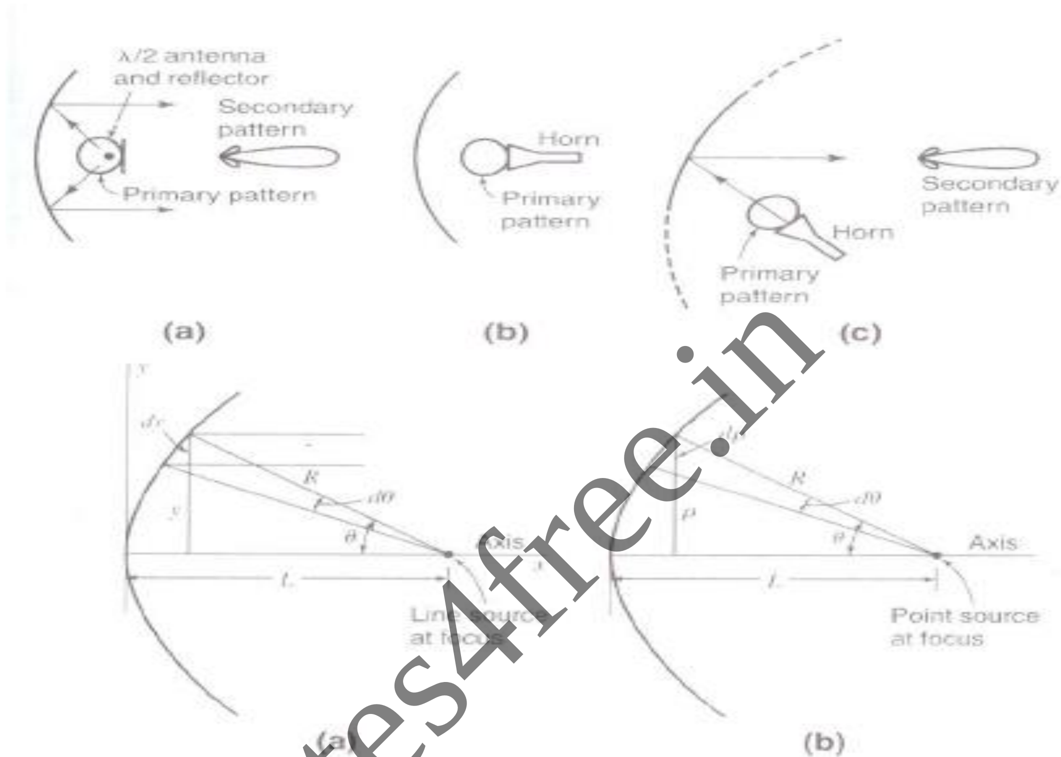


Fig .5.7 parabolic feed structure

Let us next develop an expression for the field distribution across the aperture of a parabolic reflector. Since the development is simpler for a cylindrical parabola, this case is treated first, as an introduction to the case for a paraboloid. Consider a cylindrical parabolic reflector with line source as in Fig.a. The line source is isotropic in a plane perpendicular to its axis (plane of page). For a unit distance in the z direction the power P in a strip of width dy is

$$P = dyS_y$$

Where S_y= the power density ay y, W m^{-2}

$$P = U'd\theta$$

U'=the power per unit angle per unit length in the direction

$$S_y dy = U'd\theta$$

$$\frac{S_y}{U'} = \frac{1}{(d/d\theta)(R\sin\theta)}$$

$$R = \frac{2L}{1+\cos\theta}$$

$$S_y = \frac{1+\cos\theta}{2L} U'$$

The ratio of power density

$$\frac{S_\theta}{S_0} = \frac{1+\cos\theta}{2}$$

The field intensity ratio in the aperture plane is equal to the square root of the power ratio

$$\frac{E_\theta}{E_0} = \sqrt{\frac{1+\cos\theta}{2}}$$

$$P = 2\pi\rho d\rho S\rho$$

$$P = 2\pi \sin\theta d\theta U$$

Equating the above two equations, we get,

$$\rho d\rho S\rho = \sin\theta d\theta U$$

$$\frac{S_\rho}{U} = \frac{\sin\theta}{\rho(d\rho/d\theta)}.$$

$$S_\rho = \frac{(1+\cos\theta)^2}{4L^2} U$$

$$\frac{S_\theta}{S_0} = \frac{(1+\cos\theta)^2}{4}$$

$$\frac{E_\theta}{E_0} = \frac{1+\cos\theta}{2}$$

5.6 LOG PERIODIC DIPOLE ARRAY

The log periodic dipole array (LPDA) is one antenna that almost everyone over 40 years old has seen. They were used for years as TV antennas. The chief advantage of an LPDA is that it is frequency-independent. Its input impedance and gain remain more or less constant over its operating bandwidth, which can be very large. Practical designs can have a bandwidth of an octave or more. Although an LPDA contains a large number of dipole elements, only 2 or 3 are active at any given frequency in the operating range. The electromagnetic fields produced by these active elements add up to produce a unidirectional radiation pattern, in which maximum radiation is off the small end of the array. The radiation in the opposite direction is typically 15 - 20 dB below the maximum. The ratio of maximum forward to minimum rearward radiation is called the Front-to-Back (FB) ratio and is normally measured in dB.

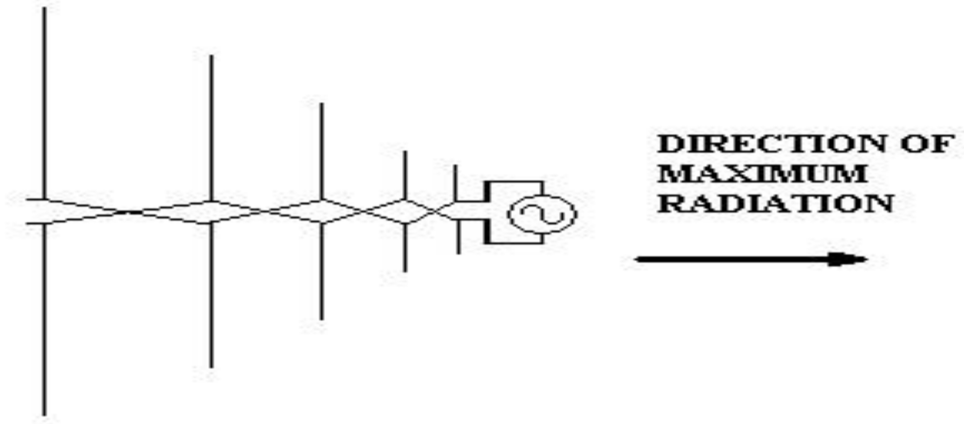


Fig .5.8 Log periodic Dipole Array

The log periodic antenna is characterized by three interrelated parameters, α, σ and τ as well as the minimum and maximum operating frequencies, f_{MIN} and f_{MAX}. The diagram below shows the relationship between these parameters.

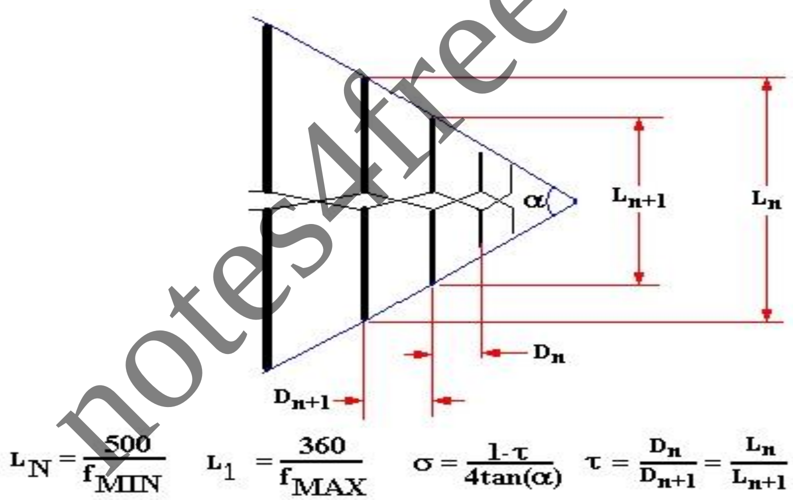


Fig .5.9 Relationship between different parameters

Unlike many antenna arrays, the design equations for the LPDA are relatively simple to work with. If you would like to experiment with LPDA designs, click on the link below. It will open an EXCEL spreadsheet that does LPDA design.

5.7 LENS ANTENNAS

With a LENS ANTENNA you can convert spherically radiated microwave energy into a plane wave (in a given direction) by using a point source (open end of the waveguide) with a COLLIMATING LENS. A collimating lens forces all radial segments of the spherical wavefront into parallel paths. The point source can be regarded as a gun which shoots the microwave energy toward the lens. The point source is often a horn radiator or a simple dipole antenna.

Waveguide type: the waveguide-type lens is sometimes referred to as a conducting- type. It consists of several parallel concave metallic strips which are placed parallel to the electric field of the radiated energy fed to the lens, as shown in Figure 3-10A and 3-10B. These strips act as waveguides in parallel for the incident (radiated) wave. The strips are placed slightly more than a half wavelength apart.

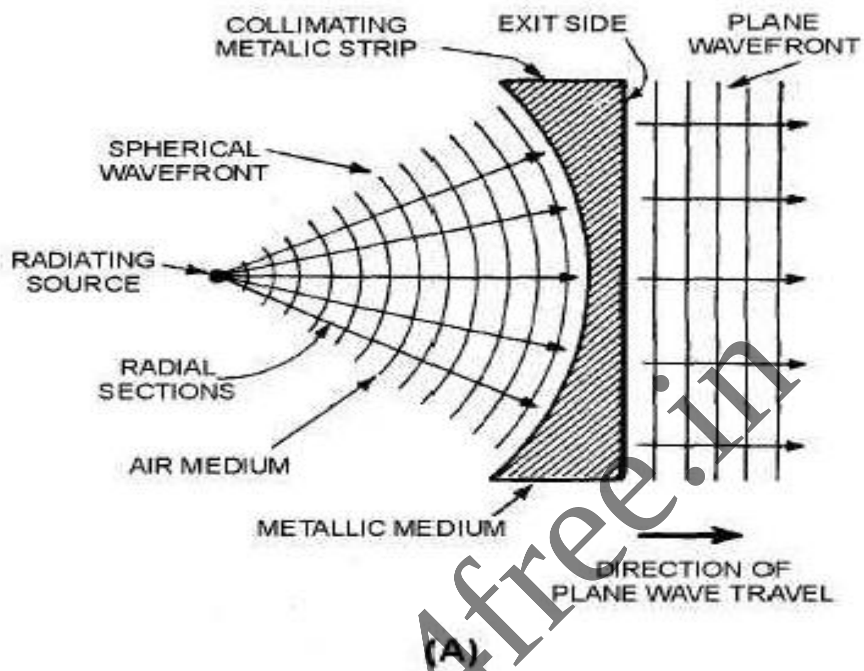


Fig 5.10 Lens Antenna

Advantages of Lens Antenna

- Can be used as wideband antenna since its shape is independent of frequency.
- Provides good collimation.
- Internal dissipation losses are low, with dielectric materials having low loss tangent.
- Easily accommodate large band width required by high data rate systems
- Quite in-expensive and have good fabrication tolerance

Disadvantages of Lens Antenna

- Bulky and Heavy
- Complicated Design
- Refraction at the boundaries of the lens

5.8 Antennas for Special applications

5.8.1 Sleeve antenna

Ground plane or sleeve type λ/4 long cylindrical system is called a sleeve antenna. The radiation is in a plane normal to the axis of this antenna. The second variety of sleeve is similar to stub with ground plane having the feed point at the centre of the stub. The lower end of the stub is a cylindrical sleeve of length λ/8.

A balanced-sleeve dipole antenna corresponding to the sleeve stub is shown in Fig. This is fed with a coaxial cable and balance to unbalance transformer or balun. For L ranging between λ/2 to λ, the operating frequency ranges through 2 to 1.

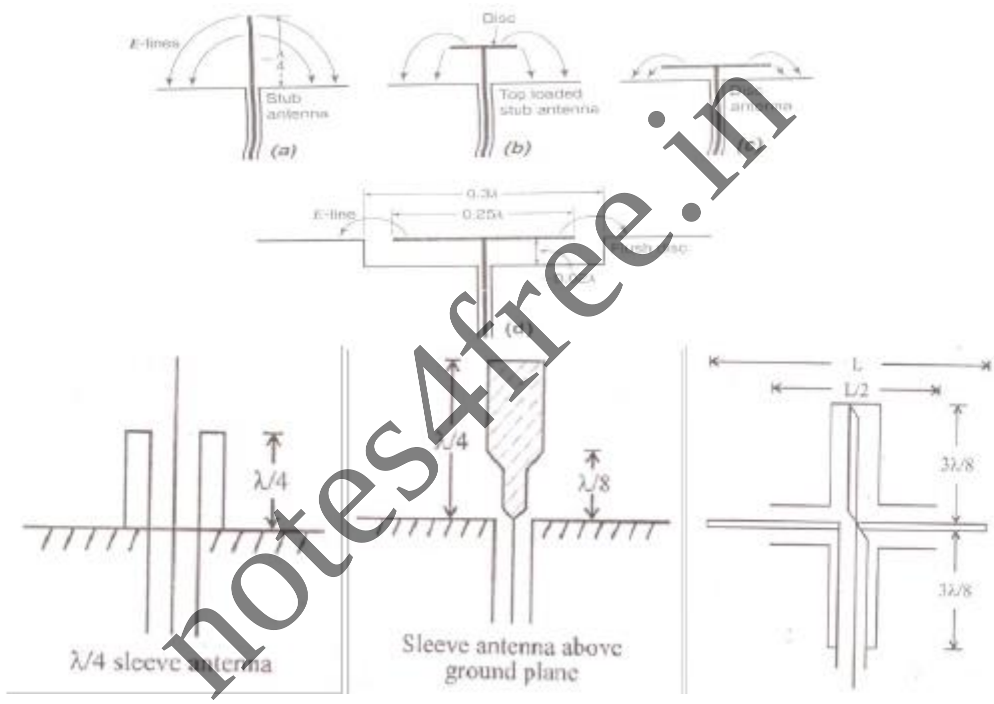


Fig 5.11 Sleeve Antenna

Evolution of flush-disk antenna from vertical λ/4 stub antenna

It is the modified ground plane antenna.

Here the ground plane has de-generated into a sleeve or cylinder λ/4 long.

Maximum radiation is normal to the axis.

5.8.2 Turn Stile Antenna

The Antenna is similar to stub antenna with ground plane but with a feed point moved to approximately the center of the stub. A basic turn stile consists of two horizontal short dipoles placed normal to each other as shown in Fig. The individual field patterns are 'Figure of eight' fitted by 900. The total field pattern is given by

$$E = \sin\theta\cos\omega t + \cos\theta\sin\omega t$$

$$E = \sin(\theta + \omega t)$$

$$\omega t = -\theta$$

$$|E| = \sqrt{\cos^2\omega t + \sin^2\omega t} = 1$$

$$E = \frac{\cos(90^0\cos\theta)}{\sin\theta}\cos\omega t + \frac{\cos(90^0\sin\theta)}{\cos\theta}\sin\omega t$$

$$I_1 = \frac{V}{70 + j70}$$

$$I_2 = \frac{V}{70 - j70}$$

Where

V = Impressed emf

I1 = current at terminals of dipole 1

I2 = current at terminals of dipole 2

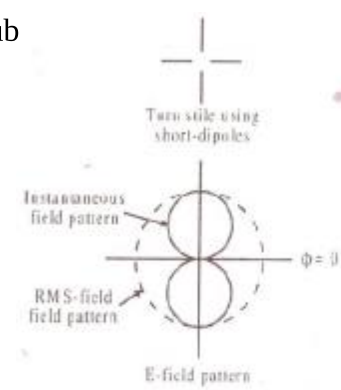
Thus

$$I_1 = \frac{V}{99}\angle -45^0$$

$$I_2 = \frac{V}{99}\angle +45^0$$

$$Z = \frac{1}{Y} = \frac{1}{\left[1/(70 + j70)\right] + \left[1/(70 - j70)\right]} = 70 + j0(\Omega)$$

The Antenna is similar to stub antenna with ground plane but with a feed point moved to approximately the center of the stub



Turn stile array with individual field pattern

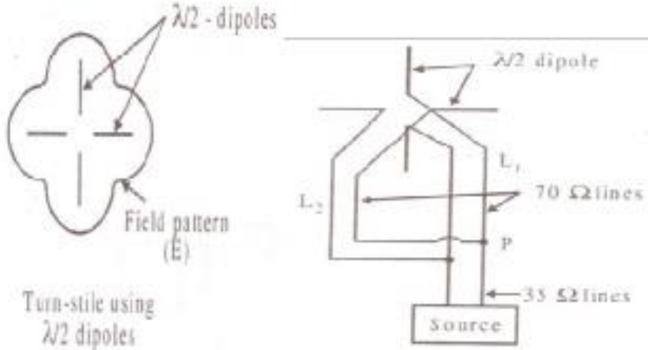


Fig.5.12. Turn stile array with resultant field pattern

The turn stile is most suited for TV transmission for frequency from 50 MHz. Directivity can be increased by stacking super turn stiles one above the other as shown in Figure.

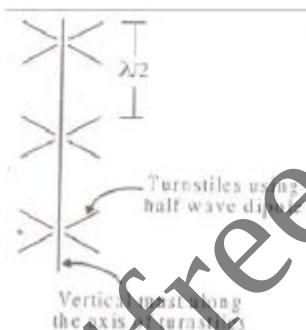


Fig 5.13 Stack of turn stile array

5.8.3 Omni-directional antennas

Slotted cylinder, and turnstile are almost omni-directional in horizontal plane. Clover-leaf is one more type of omni-directional whose directivity is much higher than that of turnstile. The system basically contains horizontal dipole which is bidirectional in vertical plane. A circular loop antenna as shown in Fig can be used to obtain omnidirectional radiation pattern.

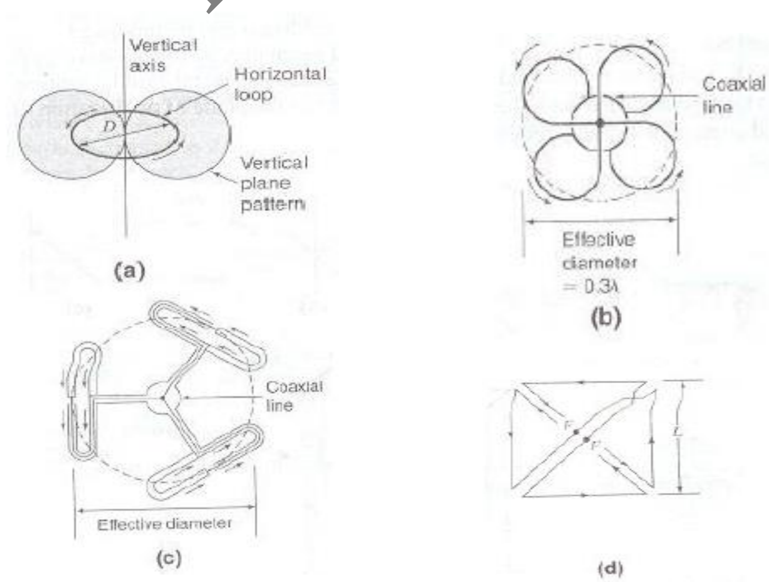


Fig.5.14 a) Circular Loop Antenna b) Approximately equivalent arrangements of "clover-leaf" type c) "triangular-loop" type Antenna d) Square or Alford loop

5.8.4 Antenna for Mobile Application

Switched Beam Antenna:

The base station antenna has several selectable beams of which each covers a part of the cell area as shown in the Figure 5.24. The switched beam antenna is constructed based on Butler matrix, which provides one beam per antenna element. The system operation is very simple but has limited adaptability.

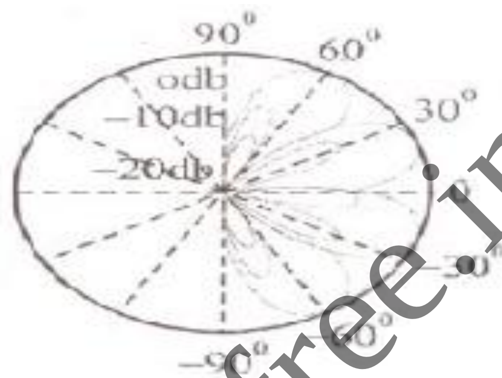


Fig 5.15 Switched Beam Pattern

Adaptive Antenna: Adaptive array is the most comprehensive and complex configuration. The system consists of several antennas where each antenna is connected to separate trans-receiver and Digital Signal Processor as shown in Fig. DSP controls the signal level to each element depending upon the requirements. Butler matrix can be adapted for the improvement of SNR during reception. Direction of arrival finding and optimization algorithms are used to select the complex weights for each mobile users. For frequency domain duplexing the transmission weights are estimated based on Direction of arrival information.

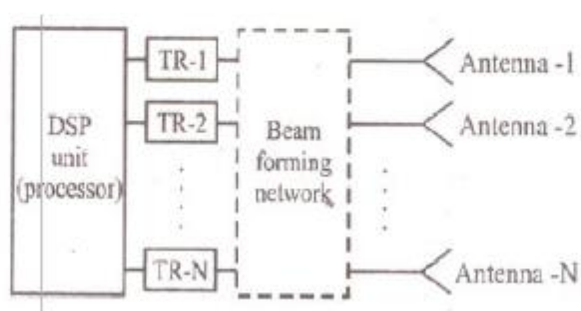


Fig 5.16. Adaptive Antenna

Antenna for satellite:

- High Frequency Transmitting Antenna
- Parabolic Reflector

6.8.5 Antennas for Ground Penetrating Radar (GPR)

- Like Earth Surface Radars, the radars can be used to detect underground anomalies both natural and Human Made.
- The anomalies include buried metallic or nonmetallic objects, earth abnormalities etc.,
- Pulse and its echo pulse are used for processing.
- Far field radar equation to be modified as distance travelled by wave is less.
- Power required is more since ground is lossy medium.
- Mismatch at air-ground interface.
- Pulse width should be less.

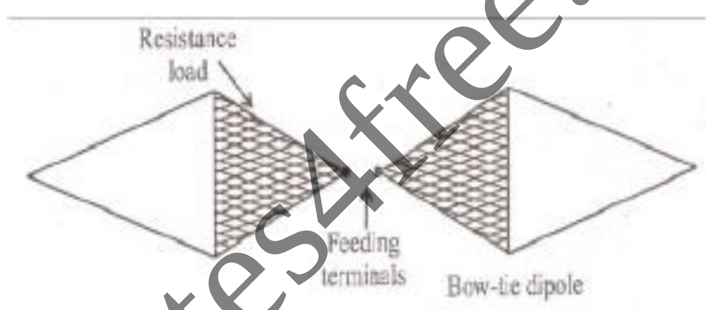


Fig 5.17 Ground Penetrating Radar (GPR) Antenna

5.8.6 Embedded Antennas

- If dipole is embedded in a dielectric medium of relative permitivity εr (>1), then its length can be reduced.
- A λ/2 dipole resonates at the same frequency when embedded in a dielectric medium having a length 0.5λ/sq root of εr
- If εr= 4, length required is half.
- Used in Bluetooth technology, interfacing RF Networks.

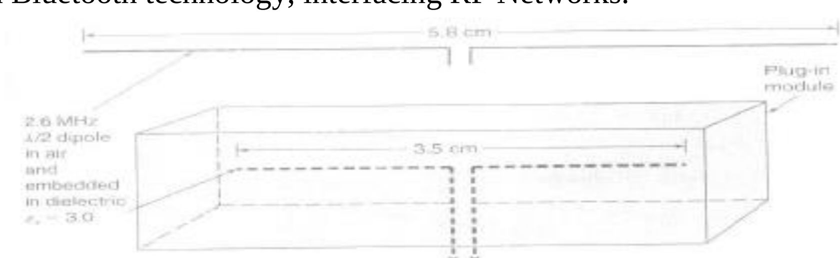


Fig 5.18..Half-wavelength dipole embedded in a dielectric for Bluetooth Application

5.8.7 Ultra Wide Band Antenna

- Used for digital Applications
- Pulse Transmission which results in Large bandwidth.
- Phase dispersion of pulse (transmitted at different instant of time)
- Degrading of signals

V Antenna used for Communication

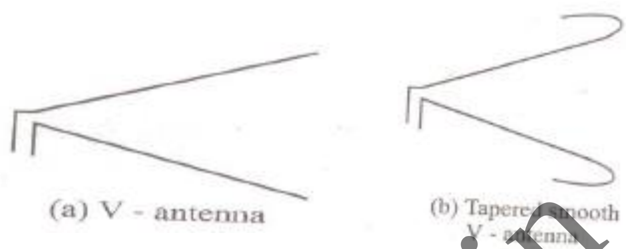


Fig.5.19 Ultra Wide Band Antenna

5.8.8 Plasma antenna

- A plasma surface wave can be excited along a column of low-pressure gas by adequate RF power coupled to the column in a glass tube.
- It is a system in which the radar cross section is only the thin wall glass tube when not transmitting. With a laser beam producing the plasma the radar cross section becomes zero when laser is off.

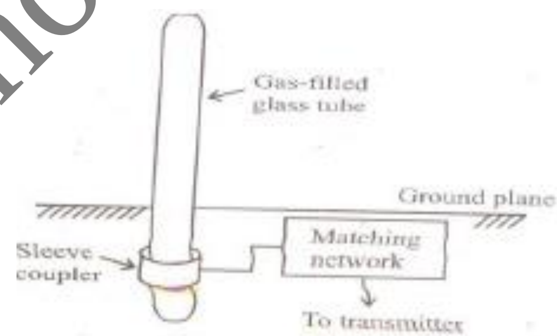


Fig.5.20 Plasma antenna

5.9 OUTCOMES

- To solve problems related to different types of antennas.
- Concept of Different antenna types is understood

5.10 Questions

1. With a neat figure explain the working of Yagi-Uda antenna. Write the design formulae for different components used in Yagi-Uda antenna. Also mention the applications of Yagi-Uda antenna.
2. Write a short note on log-periodic antenna.
3. Explain the features of a helical antenna. Explain the practical design considerations of helical antenna.
4. With a neat sketch explain the principle of lens antenna. Also list the merits and demerits of lens antenna.
5. Explain the corner reflector antenna.
6. What are parabolic reflectors? Where are these antennas used?
7. Write short notes on the following:
   i. Ultra-wide band antenna (4m)
   ii. Turnstile antenna (4m)
   iii. Patch antenna (4m)
   iv. Antenna for ground penetrating radar (4m)
   v. Plasma antenna (4m)

8. Draw the construction details of an embedded antenna.

5.11 Further Readings

1. **Antenna Theory Analysis and Design** - C A Balanis, 3rd Edn, John Wiley India Pvt. Ltd, 2008
2. **Antennas and Propagation for Wireless Communication Systems** - Sineon R Saunders, John Wiley, 2003