EC8701 ANTENNAS AND MICROWAVE ENGINEERING

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OBJECTIVES:

- To enable the student to understand the basic principles in antenna and microwave system design
- To enhance the student knowledge in the area of various antenna designs.
- To enhance the student knowledge in the area of microwave components and antenna for practical applications.

UNIT I INTRODUCTION TO MICROWAVE SYSTEMS AND ANTENNAS 9

Microwave frequency bands, Physical concept of radiation, Near- and far-field regions, Fields and Power Radiated by an Antenna, Antenna Pattern Characteristics, Antenna Gain and Efficiency, Aperture Efficiency and Effective Area, Antenna Noise Temperature and G/T, Impedance matching, Friis transmission equation, Link budget and link margin, Noise Characterization of a microwave receiver.

UNIT II RADIATION MECHANISMS AND DESIGN ASPECTS 9

Radiation Mechanisms of Linear Wire and Loop antennas, Aperture antennas, Reflector antennas, Microstrip antennas and Frequency independent antennas, Design considerations and applications.

UNIT III ANTENNA ARRAYS AND APPLICATIONS 9

Two-element array, Array factor, Pattern multiplication, Uniformly spaced arrays with uniform and non-uniform excitation amplitudes, Smart antennas.

UNIT IV PASSIVE AND ACTIVE MICROWAVE DEVICES 9

Microwave Passive components: Directional Coupler, Power Divider, Magic Tee, attenuator, resonator, Principles of Microwave Semiconductor Devices: Gunn Diodes, IMPATT diodes, Schottky Barrier diodes, PIN diodes, Microwave tubes: Klystron, TWT, Magnetron.

UNIT V MICROWAVE DESIGN PRINCIPLES 9

Impedance transformation, Impedance Matching, Microwave Filter Design, RF and Microwave Amplifier Design, Microwave Power amplifier Design, Low Noise Amplifier Design, Microwave Mixer Design, Microwave Oscillator Design

TOTAL: 45 PERIODS

OUTCOMES:

The student should be able to:

- Apply the basic principles and evaluate antenna parameters and link power budgets
- Design and assess the performance of various antennas
- Design a microwave system given the application specifications

TEXTBOOKS:

1. John D Krauss, Ronald J Marhefka and Ahmad S. Khan, "Antennas and Wave Propagation: Fourth Edition, Tata McGraw-Hill, 2006. (UNIT I, II, III)
2. David M. Pozar, "Microwave Engineering", Fourth Edition, Wiley India, 2012.(UNIT I,IV,V)

REFERENCES:

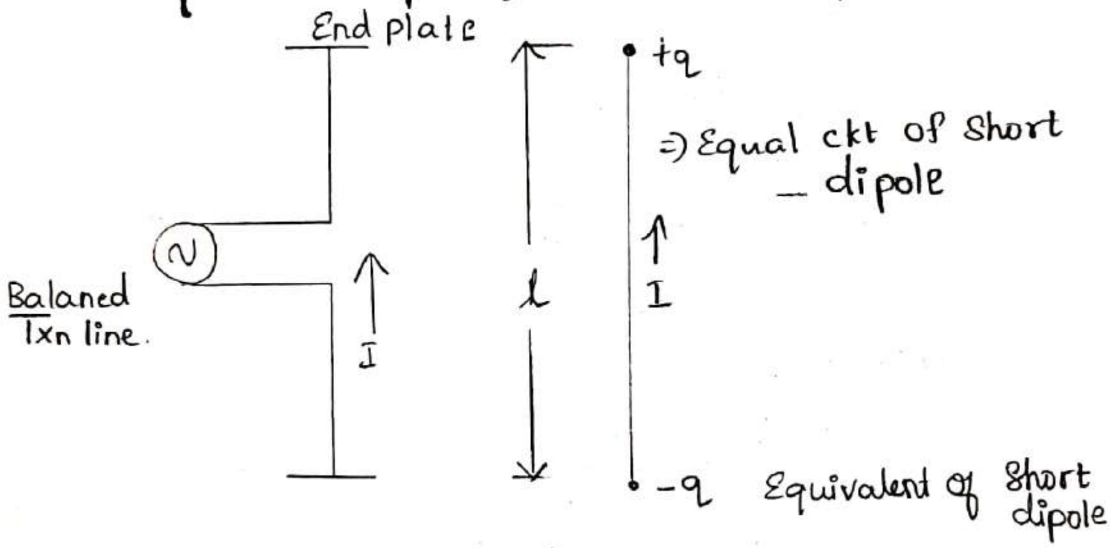
1. Constantine A.Balanis, “Antenna Theory Analysis and Design”, Third edition, John Wiley India Pvt Ltd., 2005.
2. R.E.Collin, "Foundations for Microwave Engineering", Second edition, IEEE Press, 2001

HERTIZIAN DIPOLE / SHORT ELECTRIC DIPOLE

[p.no → 414 (K.D prasad)]

Any linear antenna may be considered as a loop number of very short conductors connected in series (ie) End to End and hence it is important first to consider radiation properties of such short conductor

A Short linear conductor is so short that Current may be assume to be constant throughout its length. Such short linear conductor is called as short dipole / Hertizian dipole.



Hertizian dipole is a hypothetical antenna and it is defined as a short isolated conductor carrying Uniform alternating Current.

The Current Carrying element

$$I = \frac{dq}{dt}$$

A short dipole that does not have a Uniform Current. so it is called as a Elemental dipole.

Such a dipole will generally, be considerably shorter than 1 tenth Wavelength maximum specified for a short dipole. The other term called as Elementary doublet [1/10th shorter d]

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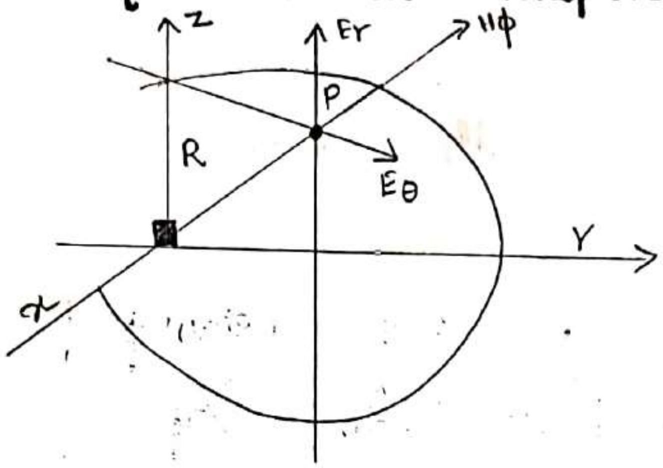
Hertizian dipole in which

$\boxed{L \ll \lambda}$

Basic elements of Antenna :-

5 types

1) Hertizian dipole
2) Short dipole
3) Short Monopole
4) Half wave dipole
5) quarter Wave Monopole.



The current distribution is constant for short dipole.

Triangular current distribution for short monopole.

Sinusoidal current distribution for $\lambda/2$ and $\lambda/4$ antenna.

The current distribution can be determined by Vector potential.

$$I = I_m \cos \omega t \qquad \therefore I_m = Idl$$

$$I = Idl \cos \omega t$$

$$A(r) = \frac{\mu}{4\pi} \int \frac{J(t - r/v)}{R} dv'$$

$$\therefore A_z = \frac{\mu}{4\pi} \frac{Idl \cos \omega (t - r/v)}{R}$$

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$$A = A_x \hat{a}_x^{\nearrow =0} + A_y \hat{a}_y^{\nearrow =0} + A_z \hat{a}_z$$

$$A = A_z \hat{a}_z$$

where A = magnetic Vector quantity.

$$\therefore A = \frac{\mu}{4\pi} \frac{I dl \cos\omega (t - r/v)}{R} a_z$$

Convert the Cartesian Coordinate System (x, y, z) to Spherical Co-ordinate System (r, θ, ϕ)

$$A_x = A_r = A_z \cos\theta$$

$$A_y = A_\theta = -A_z \sin\theta \qquad -A_z \sin\theta.$$

$$A_z = A_\phi = 0$$

normal formula:-

$$\nabla \times A = \frac{1}{r^2 \sin\theta} \begin{vmatrix} e_r & r e_\theta & r\sin\theta e_\phi \\ \partial/\partial r & \partial/\partial\theta & \partial/\partial\phi \\ A_r & rA_\theta & r\sin\theta A_\phi \end{vmatrix}$$

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W.k.T $A_\phi = 0$

∴ So the fields are Spherically Symmetric

$$\nabla \times A = \frac{e_r}{r^2\sin\theta}\left[\frac{\partial}{\partial\theta}\underbrace{(r\sin\theta A_\phi)}_{=0\ (\text{since } A_\phi = 0)} - \frac{\partial}{\partial\phi}(rA_\theta)\right] - \frac{r e_\theta}{r^2 \sin\theta}$$

$$\left[\frac{\partial}{\partial r}\underbrace{(r\sin\theta A_\phi)}_{=0\ (A_\phi = 0)} - \frac{\partial}{\partial\phi}(A_r)\right] + \frac{r\sin\theta e_\phi}{r^2\sin\theta}$$

$$\left[\frac{\partial}{\partial r}(rA_\theta) - \frac{\partial}{\partial\theta}(A_r)\right]$$

$$\therefore \nabla \times A = \frac{e_r}{r^2\sin\theta}\left[-\frac{\partial}{\partial\phi}(rA_\theta)\right] - \frac{e_\theta}{r\sin\theta}\left[-\frac{\partial}{\partial\phi}(A_r)\right]$$

$$+ \frac{e_\phi}{r}\left[\frac{\partial}{\partial r}(rA_\theta) - \frac{\partial}{\partial\theta}(A_r)\right]$$

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$$\nabla\times A = \frac{e_r}{r^2\sin\theta}\left[-\partial/_{\partial\phi}(A_\theta \cdot r)\right] + \frac{e_\theta}{r\sin\theta}\left[\partial/_{\partial\phi}(A_r)\right]$$
$$+ \frac{e_\phi}{r}\left[\partial/_{\partial r}(rA_\theta) - \partial/_{\partial\theta}(A_r)\right]$$

According to formulae, we obtain

$$\nabla\times A = \frac{1}{r^2\sin\theta}\left[e_r(0) - r e_\theta(0) + r\sin\theta\, e_\phi\left[-A_z\sin\theta + A_z\sin\theta\right]\right]$$

The field is spherically symmetry

So $\partial/_{\partial\phi} = 0$ and $A_\phi = 0$

$(\nabla\times A)_r = \mu H_r = 0$

$(\nabla\times A)_\theta = \mu H_\phi = 0$

$(\nabla\times A)_\phi = \mu H_\phi = \frac{1}{r}\left[\partial/_{\partial r}(rA_\theta) - \partial/_{\partial\theta}(A_r)\right]$

$A_r = A_z\cos\theta$; $H_r = 0$

$A_\theta = -A_z\sin\theta$; $H_\theta = 0$

$A_\phi = 0$

$$\therefore \mu H_\phi = \frac{1}{r}\left[\partial/_{\partial r}(-A_z\sin\theta\, r) - \partial/_{\partial\theta}(A_z\cos\theta)\right] \longrightarrow (1)$$

but

$$A_z = \frac{\mu}{4\pi r} I m\,dl\cos\omega(t - r/c)$$

multiply by $-r\sin\theta$ on both sides

$$\underbrace{-r\sin\theta A_z}_{A_\theta} = -\frac{\mu \cancel{r}}{4\pi\cancel{r}} I m\,dl\cos\omega(t - r/c)\sin\theta$$

$$rA_\theta = -\frac{\mu}{4\pi} I m\,dl\cos\omega(t - r/c)\sin\theta$$

$$\partial/_{\partial r}(rA_\theta) = -\frac{\partial}{\partial r}\left[\frac{\mu I m\,dl\cos\omega(t - r/c)\sin\theta}{4\pi}\right]$$

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$$\frac{\partial}{\partial r}\left[-A_z \sin\theta\, r\right] = (-)\frac{\mu I m\, dl \sin\theta}{4\pi}\left\{\frac{\partial}{\partial r}\underbrace{\cos\omega\left(t-r/c\right)}_{\cos[\omega t \cdot \omega r/c)}\right.$$

$$\frac{\partial}{\partial r}\left[-A_z \sin\theta\, r\right] = (-)\frac{\mu I m\, dl \sin\theta}{4\pi}\left[\sin\omega\left(t-r/c\right)\left(\omega/c\right)\right]$$

$$\frac{\partial}{\partial r}\left[-A_z \sin\theta\, r\right] = (-)\frac{\mu I m\, dl \sin\theta}{4\pi}\left[\left[\sin\omega\left(t-r/c\right)\right]\right]\left(\omega/c\right) \rightarrow (2)$$

$$\frac{\partial}{\partial \theta}\left[-A_z \cos\theta\right] = \frac{\partial}{\partial\theta}\left[\frac{\mu I m\, dl \cos\omega\left(t-r/c\right)}{4\pi r}\cos\theta\right]$$

$$\frac{\partial}{\partial \theta}\left[-A_z \cos\theta\right] = \frac{\mu I m\, dl \cos\omega\left(t-r/c\right)}{4\pi r}\left[\frac{\partial}{\partial\theta}\cos\theta\right]$$

$$\frac{\partial}{\partial \theta}\left[-A_z \cos\theta\right] = \frac{\mu I m\, dl \cos\omega\left(t-r/c\right)}{4\pi r}\left[\sin\theta\right]$$

$$\frac{\partial}{\partial \theta}\left[-A_z \cos\theta\right] = \frac{\mu I m\, dl \cos\omega\left(t-r/c\right)}{4\pi r}(\sin\theta) \rightarrow (3)$$

$$(\nabla \times A)_\phi = \mu H_\phi$$

substitute equ (2) & equ (3) in equ (1)

$$\mu H_\phi = \frac{1}{r}(-)\left[\left[\frac{\mu I m\, dl \sin\theta}{4\pi}\left[\frac{\omega \sin\omega\left(t-r/c\right)}{c}\right]\right. - \frac{\mu I m\, dl \sin\theta}{4\pi r}\cos\omega\left(t-r/c\right)\right]$$

$$\mu H_\phi = \frac{1}{r}(-)\left[\frac{\mu I m\, dl \sin\theta}{4\pi}\left(\frac{\omega \sin\omega\left(t-r/c\right)}{c} - \frac{\cos\omega\left(t-r/c\right)}{r}\right.\right.$$

let $t - r/c = t_1$

$$\cancel{\mu} H_\phi = \frac{1}{r}(-)\left[\frac{\cancel{\mu} I m\, dl \sin\theta}{4\pi}\left[\frac{\omega \sin\omega t_1}{c} - \frac{\cos\omega t_1}{r}\right]\right]$$

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$$\therefore \boxed{H_\phi = \frac{I_m dl \sin\theta}{4\pi}\left[\frac{\cos\omega t_1}{r^2 c} - \frac{\omega \sin\omega t_1}{rc}\right]}$$

ELECTRIC FIELD COMPONENT [E] CAN BE DERIVED BY THIS ABOVE EQUATION:-

$$\nabla \times H = \frac{\partial}{\partial t} D = \varepsilon \frac{\partial}{\partial t} E$$ [D → displacement of current density]

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

1) $$\varepsilon \frac{\partial E_r}{\partial t} = (\nabla \times H)_r$$

2) $$\varepsilon \frac{\partial E_\theta}{\partial t} = (\nabla \times H)_\theta$$

3) $$\varepsilon \frac{\partial E_\phi}{\partial t} = (\nabla \times H)_\phi$$

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

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

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

$$\nabla \times H = \frac{1}{r^2 \sin\theta}\begin{vmatrix} e_r & r e_\theta & r\sin\theta\, e_\phi \\ \partial/\partial r & \partial/\partial\theta & \partial/\partial\phi \\ H_r & rH_\theta & r\sin\theta H_\phi \end{vmatrix}$$

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

$$H_\phi = \frac{I_m dl \sin\theta}{4\pi}\left[\frac{-\omega}{rc}\sin\omega t_1 + \frac{\cos\omega t_1}{r^2 c}\right]$$

$$\varepsilon \frac{\partial E_r}{\partial t} = (\nabla \times H)_r \text{ (from ①)}$$

$$= \frac{1}{r\sin\theta}\left[\frac{\partial}{\partial\theta}(H_\phi \sin\theta)\right]$$ $H_\theta = 0$

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$$\epsilon \frac{\partial E_r}{\partial t} = \frac{1}{r\sin\theta}\left[\partial/_{\partial\theta}\left[\frac{Imdl\sin\theta}{4\pi}\left\{\frac{-\omega}{rc}\sin\omega t_1 + \frac{\cos\omega t_1}{r^2}\right\}\sin\theta\right]\right]$$

$\frac{\partial}{\partial\theta}\sin^2\theta$

$\frac{\partial}{\partial\theta}\sin^2\theta$

$\sin 2\alpha = 2\sin\alpha\cos\alpha$

$\sin 2\theta = 2\sin\theta\cos\theta$

$$\epsilon \frac{\partial E_r}{\partial t} = \frac{1}{r\sin\theta}\left[\frac{Imdl}{4\pi}\left[\partial/_{\partial\theta}\overset{2\sin\theta\cos\theta}{(\sin^2\theta)}\right]\left\{\frac{-\omega}{rc}\sin\omega t_1 + \frac{\cos\omega t_1}{r^2}\right\}\right.$$

$2\cos\theta\sin\theta$

$$\epsilon \frac{\partial E_r}{\partial t} = \frac{1}{r\sin\theta}\left[\frac{Imdl}{4\pi}\left[\frac{\cos\omega t_1}{r^2} - \frac{\omega}{rc}\sin\omega t_1\right] \times 2\cos\theta\sin\theta\right.$$

$$\frac{\partial E_r}{\partial t} = \frac{2Imdl\cos\theta\cancel{\sin\theta}}{4\pi r\epsilon\cancel{\sin\theta}}\left[\frac{\cos\omega t_1}{r^2} - \frac{\omega}{rc}\sin\omega t_1\right]$$

$$\frac{\partial E_r}{\partial t} = \frac{2Imdl\cos\theta}{4\pi\epsilon}\left[\frac{\cos\omega t_1}{r^3} - \frac{\omega}{r^2c}\sin\omega t_1\right]$$

$$\int\frac{\partial E_r}{\partial t} = \int\frac{2Imdl\cos\theta}{4\pi\epsilon}\left[\frac{\cos\omega t_1}{r^3} - \frac{\omega}{r^2c}\sin\omega t_1\right]dt$$

$$E_r = \frac{2Imdl\cos\theta}{4\pi\epsilon}\int\left[\frac{\cos\omega t_1}{r^3} - \frac{\omega}{r^2c}\sin\omega t_1\right]dt$$

$$E_r = \frac{2Imdl\cos\theta}{4\pi\epsilon}\left[\frac{\sin\omega t_1}{\omega r^3} - \frac{\omega}{r^2c}\frac{(-\cos\omega t_1)}{\omega}\right]$$

$$E_r = \frac{2Imdl\cos\theta}{4\pi\epsilon}\left[\frac{\cos\omega t_1}{r^2c} + \frac{\sin\omega t_1}{\omega r^3}\right]$$

$$\boxed{\therefore E_r = \frac{2Imdl\cos\theta}{4\pi\epsilon}\left[\frac{\cos\omega t_1}{r^2c} + \frac{\sin\omega t_1}{\omega r^3}\right]}$$

$$\epsilon\frac{\partial E_\theta}{\partial t} = (\nabla\times H)_\theta = \frac{1}{r}\left\{\frac{-\partial}{\partial r}(rH_\phi)\right\}$$

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$$= (-)\frac{1}{r}\frac{\partial}{\partial r}\left\{\frac{I_m dl \sin\theta}{4\pi}\left\{-\frac{\omega \sin\omega t_1}{rc}+\frac{\cos\omega t_1}{r^2}\right\}r\right\}$$

$$\because t_1 = t - r/c$$

$$= \frac{1}{r}\frac{I_m dl \sin\theta}{4\pi}\left\{\frac{\partial}{\partial r}\left\{\frac{\omega}{rc}\sin\omega t_1 (r) - \frac{\cos\omega t_1}{r^2}r\right\}\right.$$

$$\epsilon\frac{\partial E_\theta}{\partial t} = \frac{I_m dl \sin\theta}{4\pi}\left\{\frac{\omega}{c}\cos\omega t_1(-\omega/c) - r\frac{\partial}{\partial r}\left[\frac{\cos\omega t_1 - \cos\omega t_1}{r^2}\right]\right\}$$

$$\frac{\partial E_\theta}{\partial t} = \frac{I_m dl \sin\theta}{4\pi\epsilon r}\left[-\frac{\omega^2}{c^2}\cos\omega t_1 - \left[\frac{r\sin\omega t_1(-\omega/c) - \cos\omega t_1}{r^2}\right]\right]$$

$$\frac{\partial E_\theta}{\partial t} = \frac{I_m dl \sin\theta}{4\pi r\epsilon}\left[-\frac{\omega^2 \cos\omega t_1}{c^2} - \frac{\omega \sin\omega t_1}{rc} + \frac{\cos\omega t_1}{r^2}\right]$$

$$\frac{\partial E_\theta}{\partial t} = \frac{I_m dl \sin\theta}{4\pi\epsilon}\left[-\frac{\omega^2 \cos\omega t_1}{rc^2} - \frac{\omega \sin\omega t_1}{r^2 c} + \frac{\cos\omega t_1}{r^3}\right]$$

$$\int\frac{\partial E_\theta}{\partial t} = \int\frac{I_m dl \sin\theta}{4\pi\epsilon}\left[-\frac{\omega^2 \cos\omega t_1}{rc^2} - \frac{\omega \sin\omega t_1}{r^2 c} + \frac{\cos\omega t_1}{r^3}\right]dt$$

$$E_\theta = \frac{I_m dl \sin\theta}{4\pi\epsilon}\int\left(\frac{\cos\omega t_1}{r^3} - \frac{\omega \sin\omega t_1}{r^2 c} - \frac{\omega^2 \cos\omega t_1}{rc^2}\right)dt$$

$$E_\theta = \frac{I_m dl \sin\theta}{4\pi\epsilon}\left[\frac{\sin\omega t_1}{\omega r^3} + \frac{\cancel{\omega}\cos\omega t_1}{\cancel{\omega} r^2 c} - \frac{\omega^2 \sin\omega t_1}{\omega r c^2}\right]$$

$$E_\theta = \frac{I_m dl \sin\theta}{4\pi\epsilon}\left[\frac{\sin\omega t_1}{\omega r^3} + \frac{\cos\omega t_1}{r^2 c} - \frac{\omega \sin\omega t_1}{rc^2}\right]$$

$$\boxed{E_\theta = \frac{I_m dl \sin\theta}{4\pi\epsilon}\left[\frac{\sin\omega t_1}{\omega r^3} + \frac{\cos\omega t_1}{r^2 c} - \frac{\omega \sin\omega t_1}{rc^2}\right]}$$

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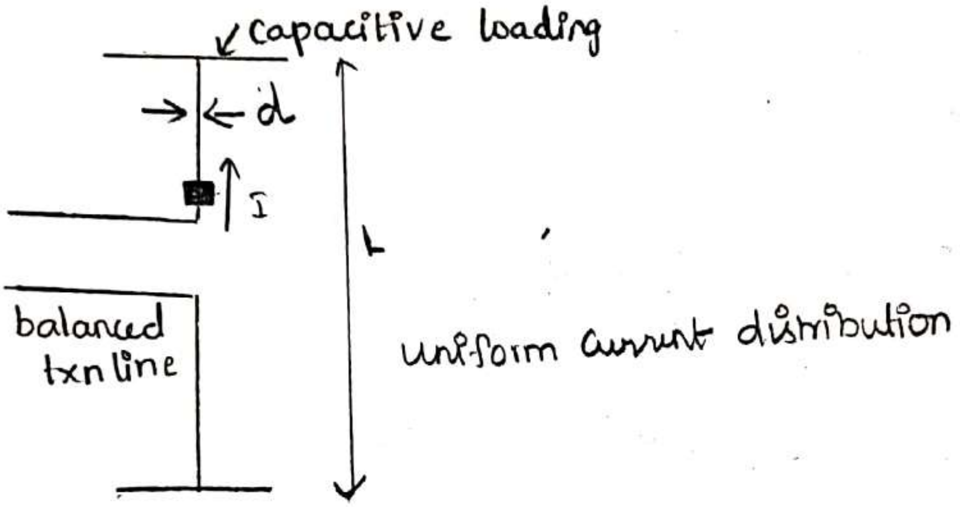
$$\therefore E_\theta = \frac{I_m dl \sin\theta}{4\pi\epsilon}\left[\frac{\sin\omega t_1}{\omega r^3} + \frac{\cos\omega t_1}{r^2 c} - \frac{\omega \sin\omega t_1}{r c^2}\right]$$

$$E_r = \frac{2 I_m dl \cos\theta}{4\pi\epsilon}\left[\frac{\cos\omega t_1}{r^2 c} + \frac{\sin\omega t_1}{\omega r^3}\right]$$

$$H_\phi = \frac{I_m dl \sin\theta}{4\pi}\left[\frac{\cos\omega t_1}{r^2} - \frac{\omega \sin\omega t_1}{r c}\right]$$

Retardation Effect:-



Definition:- If current is flowing in the short dipole, the effect of this current is not felt instantaneously at the point, but only after an interval is equal to time required for disturbance to propagate over 'r'. This is called as retardation effect.

Retardation Current : $I = I_m e^{j\omega(t - r/c)}$ A

Retardation density : $J = J_m e^{j\omega(t - r/c)}$ A/m²

Retardation Vector potential : $A = \frac{\mu}{4\pi}\int \frac{I}{r}\, dl$

$$A = \frac{\mu}{4\pi}\int \frac{I_m e^{j\omega(t - r/c)}}{r}\, dl$$

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$I = I_m \cos\omega t$

The current Excited in short dipole

$$[A] = \frac{\mu}{4\pi} \frac{I_m \cos\omega(t - r/c)}{r} dl$$

(near field) (far field)

Induction of radiation field:-

$$H\phi = \frac{I_m dl \sin\theta}{4\pi} \left[\underbrace{\frac{\cos\omega t_1}{r^2}}_{\Rightarrow \text{near field component}} - \underbrace{\frac{\omega \sin\omega t_1}{rc}}_{\Rightarrow \text{far field component}} \right]$$

(The antenna musto only Excite with current source)

r = distance from source to destination.

The first term Varies inversely square of the distance, when r is small $\frac{1}{r^2}$ is a predominant at points flows to the current Element.

The first term is responsible for Energy stored in the magnetic field and it cannot be trusted for reception.

The Second term is inversely proportional to the distance and this term is trusted becoz it is in far field.

E_θ and E_r are the distance at which the radiation field = induction field.

$$\text{Induction field} = \frac{I_m dl \sin\theta}{4\pi} \left[\frac{\cos\omega t_1}{r^2} \right] \rightarrow (1)$$

$$\text{far field} = \frac{I_m dl \sin\theta}{4\pi} \left[\frac{-\omega \sin\omega t_1}{rc} \right] \rightarrow (2)$$

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Taking modulus and Equating the above equations.

$\neq$ JT becomes (+ve)

$$\left|\frac{Imdl\sin\theta}{4\pi}\left[\frac{\cos\omega t_1}{r^2}\right]\right| = \left|\frac{Imdl\sin\theta}{4\pi}\left[\frac{-\omega\sin\omega t_1}{rc}\right]\right|$$

$$\frac{\cos\omega t_1}{r^2} = \frac{\omega\sin\omega t_1}{rc}$$

$$\frac{\cos\omega t_1}{\omega\sin\omega t_1} = \frac{r\cancel{^2}}{\cancel{r}c}$$

$$\frac{\cos\omega t_1}{\omega\sin\omega t_1} = \frac{r}{c}$$

$$\frac{\cos\omega t_1}{\sin\omega t_1} = \frac{\omega r}{c}$$

W.K.T $\frac{\cos\theta}{\sin\theta} = \cot\theta$

$$\cot\omega t_1 = \frac{r\omega}{c}$$

$$\frac{c}{\omega}\cot\omega t_1 = r$$

$(\omega = 90)$

$\cot\omega t_1 = f(\omega + 90) = 1$

$$r = c/\omega$$

(or) $$r = \frac{c}{2\pi f}$$

Where

$\omega = 2\pi f$

$c/f = \lambda$

$$\boxed{\begin{aligned} r &= \frac{\lambda}{2\pi} \\ r &= \frac{\lambda}{6} \\ r &= 0.159\lambda \end{aligned}}$$

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Radiation pattern of Elemental dipole:-

1) field component $H_r = H_\theta = E_\phi = 0$

2) E_θ & H_ϕ are in time phase, in far field

$$\eta_0 = 120\pi = 377\,\Omega$$

3) E_θ & H_ϕ are proportional to $\sin\theta$

4) The Radiation pattern will be independ of ϕ

Power Radiated by Elemental dipole: [Hertizian dipole/ short dipole]

$$P = E \times H \quad \text{(power)}$$

$$P_r = E_\theta \times H_\phi \quad \text{(power radiated)}$$

$$P_r = \frac{I_m dl \sin\theta}{4\pi}\left[\frac{\sin\omega t_1}{\omega r^3} + \frac{\cos\omega t_1}{r^2 c} - \frac{\omega \sin\omega t_1}{r c^2}\right] \cdot$$

$$\frac{I_m dl \sin\theta}{4\pi}\left[\frac{\cos\omega t_1}{r^2} - \frac{\omega \sin\omega t_1}{r c}\right]$$

$$P_r = \frac{I_m^2 dl^2 \sin^2\theta}{16\pi^2}\left[\frac{\sin\omega t_1}{\omega r^3} \cdot \frac{\cos\omega t_1}{r^2} - \frac{\omega \sin^2\omega t_1}{\omega r^3 r c}\right.$$

$$+ \frac{\cos^2\omega t_1}{r^4 c^2} - \frac{\omega \sin\omega t_1 \cos\omega t_1}{r^3 c^2} - \frac{\omega \cos\omega t_1 \sin\omega t_1}{r^3 c^2}$$

$$\left. + \frac{\omega^2 \sin^2\omega t_1}{r^2 c^3}\right]$$

$$P_r = \frac{I_m^2 dl^2 \sin^2\theta}{16\pi^2}\left[\frac{\sin\omega t_1 \cos\omega t_1}{\omega r^5} - \frac{\sin^2\omega t_1}{r^4 c} + \frac{\cos^2\omega t_1}{r^4 c^2}\right.$$

$$\left. - \frac{\omega \sin\omega t_1 \cos\omega t_1}{r^3 c^2} - \frac{\omega \cos\omega t_1 \sin\omega t_1}{r^3 r^3} + \frac{\omega^2 \sin^2\omega t_1}{r^2 c^3}\right]$$

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$\sin 2\alpha = 2\sin\alpha\cos\alpha$

$\cos 2\alpha = \cos^2\alpha - \sin^2\alpha$

$= 2\cos^2\alpha - 1$

$2\cos^2\alpha = 1 + \cos 2\alpha$

$2\sin^2\alpha = 1 - \cos 2\alpha$

$$P_r = \left[\frac{I_m dl \sin\theta}{4\pi}\right]^2 \left[\frac{\sin\omega t_1 \cos\omega t_1}{\omega r^5} - \frac{2\omega \sin\omega t_1 \cos\omega t_1}{r^3 c^2} + \frac{\cos^2\omega t_1}{c r^4} - \frac{\sin^2\omega t_1}{c r^4} + \frac{\omega^2 \sin^2\omega t_1}{r^2 c^3}\right]$$

$$P_r = \left[\frac{I_m dl \sin\theta}{4\pi}\right]^2 \left[\frac{\omega^2}{c^3 r^2}\left[\frac{1-\cos 2\omega t_1}{2}\right] - \left[\frac{2\omega \sin\omega t_1 \cos\omega t_1}{c^2 r^3}\right] - \left[\frac{\omega(1-\cos 2\omega t_1)}{2c\omega r^4}\right] + \left[\frac{1+\cos 2\omega t_1}{2cr^4}\right] + \left[\frac{\sin\omega t_1 \cos\omega t_1}{\omega r^5}\right]\right]$$

Avg values of $\sin 2\omega t_1$, & $\cos 2\omega$ will be zero and also neglect high power of "r"

$$\therefore P_r = \frac{(I_m dl \sin\theta)^2}{16\pi^2 \varepsilon}\left[\frac{\omega^2}{2c^3 r^2}\right]$$

$\omega = 2\pi f$

$\lambda = c/f$

$c = f\lambda$

$\eta = \frac{1}{\varepsilon c}$

$$= \frac{(I_m dl \sin\theta)^2}{16\pi^2 \varepsilon c}\left[\frac{4\pi^2 f^2}{2c^2 r^2}\right]$$

$$= \eta\frac{(I_m dl \sin\theta)^2}{32\pi^2 r^2 c^2}\left[4\pi^2 f^2\right]$$

$$= \eta\frac{(I_m dl \sin\theta)^2}{8\lambda^2 r^2}$$

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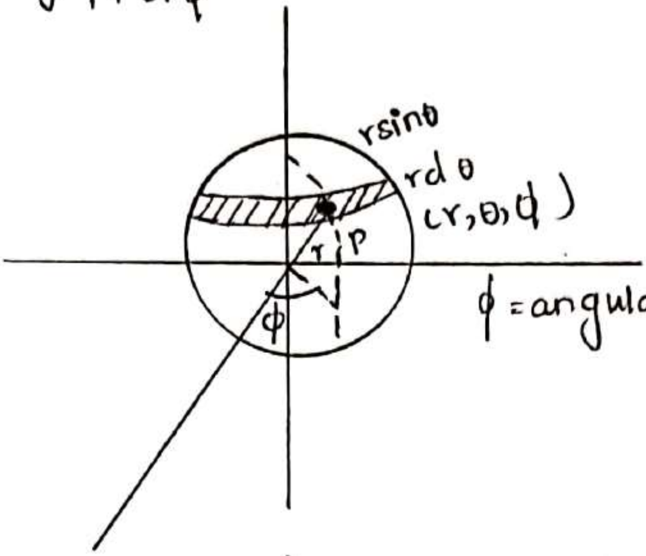
$$\boxed{P_r = \frac{377\,(I_m dl \sin\theta)^2}{8\lambda^2 r^2}\ \omega/m^2}$$

$$W = \oint P_r \, ds$$



ϕ = angular phase difference

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

$$ds = 2\pi [r\sin\theta] r \, d\theta$$

$$W = \oint \frac{377\,(I_m dl \sin\theta)^2}{8\lambda^2 \cancel{r^2}} [2\pi] (\cancel{r}\sin\theta)\, \cancel{r}\, d\theta$$

$$W = \eta \frac{I_m^2 \pi}{4} \left[\frac{dl}{\lambda}\right]^2 2 \int_0^{\pi} \sin^3\theta \, d\theta$$

$$W = \eta \pi \frac{I_m^2}{4} \left[\frac{dl}{\lambda}\right]^2 2 \int_0^{\pi} \sin^3\theta \, d\theta$$

$$\int_0^{\pi} f(x)\,dx = 2 \int_0^{\pi/2} f(x)\,dx$$

$$W = \eta \frac{\pi I_m^2}{4} \left[2 \int_0^{\pi/2} \sin^3\theta \, d\theta\right] \left(\frac{dl}{\lambda}\right)^2$$

Hilli's formulae:-

$$\int_0^{\pi/2} \sin^3\theta \, d\theta = \frac{3-1}{3} = 2/3$$

$$W = \eta \frac{\pi I_m^2}{4} \cdot 2 \cdot 2/3 \left(\frac{dl}{\lambda}\right)^2$$

$$W = \eta \frac{\pi I_m^2}{3} \left(\frac{dl}{\lambda}\right)^2$$

$$W = \frac{\cancel{120}^{40}\pi(\pi) I_m^2}{\cancel{3}} \left(\frac{dl}{\lambda}\right)^2$$

$$W = 40\pi^2 I_m^2 \left[\frac{dl}{\lambda}\right)^2$$

$$W = 40\pi^2 (\sqrt{2} I_{rms})^2 \times \left(\frac{dl}{\lambda}\right)^2$$

$$W = 80\pi^2 I_{rms}^2 \left(\frac{dl}{\lambda}\right)^2$$

$$W = (I_{rms})^2 R_r$$

Where $R_r = 80\pi^2 \left(\frac{dl}{\lambda}\right)^2$

R_r = Radiated Resistance

Quality factor:-

$$Q \cdot f = 2\pi \times \left[\frac{\text{Total Energy stored by the antenna}}{\text{Energy dissipated per cycle}}\right]$$

$$Q \cdot f \downarrow = \Omega_A \uparrow$$

∴ The relationship between Q·f and Bandwidth

$$\Delta\omega \Rightarrow B \cdot W = \frac{\omega_r}{Q}$$

ω_r = resonant frequency

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LOOP ANTENNA:-

It is a radiating coil of any convenient cross section of 1 or more turns carrying Rf current.

It is used for direction finding radio receivers, aircraft and UHF Transmitter.

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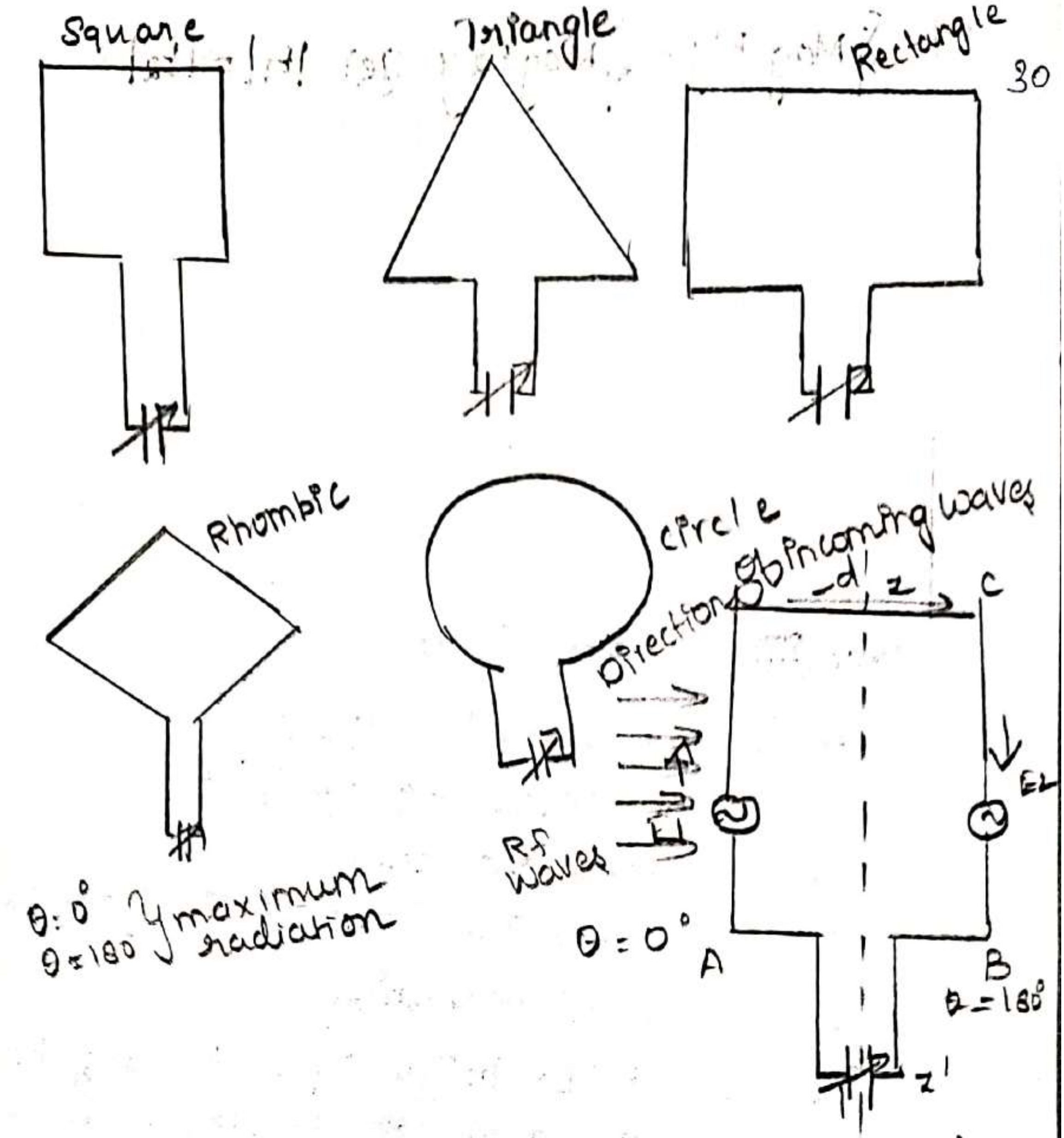


It has four arms, 2 horizontal arms and 2 vertical arms. So ABDC act as horizontal antennas.

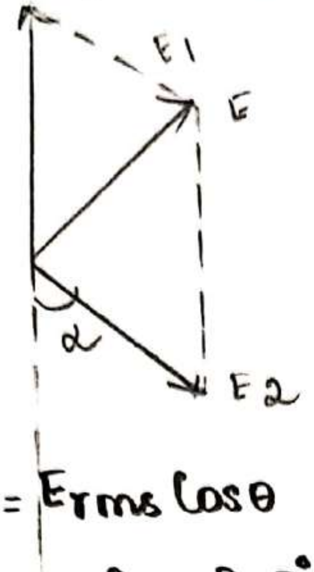
ADBC act as a vertical antennas.

Case(i) If the plane of the loop is perpendicular to the direction of incoming waves, the same voltage will be induced in each [E1 & E2] due to these voltages the vertical arms current flow in opposite direction.

Case(ii) If the plane of the loop is inline with the direction of the incoming waves then voltage induced at AD and BC is E1 and E2 respectively.

$\{Mag\ E_1\} = \{Mag\ E_2\}$ (ie) $|E_1| = |E_2|$

phase difference is α



Case (iii) $E_\theta = E_{rms} \cos\theta$

E_θ is max. when $\theta = 0°$ (or) $180°$

E_θ is mini when $\theta = 90°$ (or) $270°$

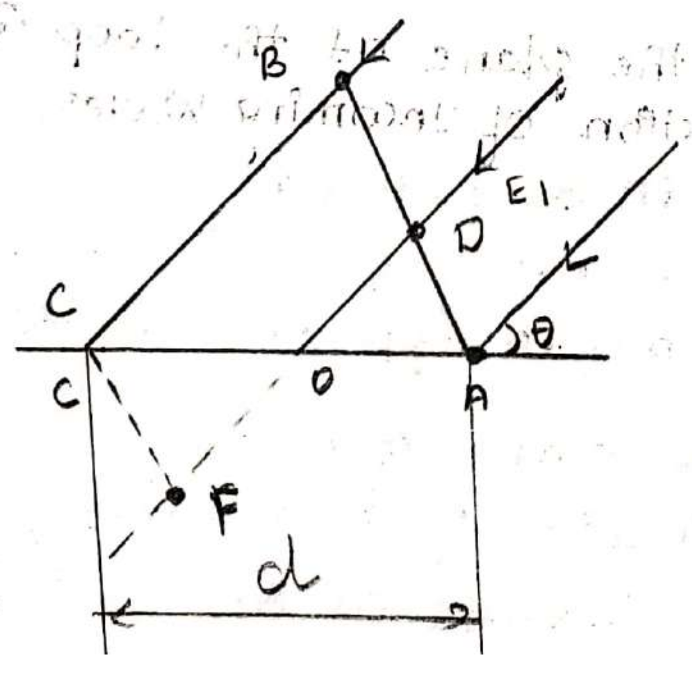
E_θ = max rms loop emf

θ = Angle b/w the plane of the loop antenna and direction of radiation.

E_θ depends on height h, d = spacing b/w

λ = wavelength, w = width, E = Electric field intensity.

EMF [I] Equation Of loop Antenna :-



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The Wavefront passes through AC then the Path difference is OD. 31

Wavefront is received at various points AOC.

Let At any instant Electric field at O

$$E = E_m \sin \omega t$$

with respect to O the D is leading by path difference α

At F its lagging by α

path diff, $OD = \frac{d}{2} \cos\theta$

phase diff, $\alpha = \frac{\cancel{2}\pi}{\lambda} \cdot \frac{d}{\cancel{2}} \cos\theta = \frac{\pi d \cos\theta}{\lambda}$

phase difference $= \frac{\pi d \cos\theta}{\lambda}$

at D, $E = E_m \sin(\omega t + \alpha)$ leading

at F, $E = E_m \sin(\omega t - \alpha)$ lagging

$E_1 = E_m h \sin(\omega t + \alpha)$ ... in AD $\rightarrow$ (1)

$E_2 = E_m h \sin(\omega t - \alpha)$ ... in BC $\rightarrow$ (2)

$e_\theta = E_1 - E_2$

$= E_m h \sin(\omega t + \alpha) - E_m h \sin(\omega t - \alpha)$

$= E_m h \left[\sin(\omega t + \alpha) - \sin(\omega t - \alpha)\right]$

$= 2 E_m h \cos \omega t \sin \alpha$

$\sin A \cos B + \cos A \sin B$

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

$e_\theta = 2 E_m h \cos \omega t \sin \alpha \rightarrow$ (2a)

Substitute $\alpha = \frac{\pi d \cos\theta}{\lambda}$

$e_\theta = 2 E_m h \cos(\omega t) \sin\left(\frac{\pi d \cos\theta}{\lambda}\right) \rightarrow$ (3)

Assume α = small value then d is much much smaller than λ, then $\sin\alpha = \alpha$

$\therefore e_\theta = 2 E_m h \cos\omega t \left(\frac{\pi d \cos\theta}{\lambda}\right)$

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$$e_\theta = \frac{2\pi d E_m h}{\lambda} \cos\omega t \cos\theta$$

$$e_\theta = \frac{2\pi h d \cos\theta}{\lambda} [E_m \cos\omega t] \rightarrow (4)$$

$hd = A$ (Area $= l \times b$)

$$e_\theta = \frac{2\pi A \cos\theta}{\lambda} [E_m \cos\omega t]$$

This is for single turn loop antenna

$$e_\theta = \frac{2\pi A N}{\lambda} \cos\theta [E_m \cos\omega t] \text{ for N turn}$$

loop antenna. $\rightarrow (5)$

rewritten as

$$e_\theta = \frac{2\pi A N}{\lambda} \cos\theta \left[E_m \sin(\omega t + \pi/2)\right]$$

$N=1$

$$e_\theta = \frac{2\pi A}{\lambda} \left[E_m \cos\omega t \sin(\theta + \pi/2)\right]$$

$$e_\theta = \frac{2\pi A}{\lambda} E_m \cos\omega t \sin(\theta + \pi/2)$$

Equation (5) is the general expression for the instantaneous value of emf at the centre of the loop.

$$V_m = \frac{2\pi A N}{\lambda} E_m \cos\theta \rightarrow (6)$$

$$\frac{V_m}{\sqrt{2}} = \frac{2\pi A N}{\lambda} \left[\frac{E_m}{\sqrt{2}}\right] \cos\theta$$

$$V_{rms} = \frac{2\pi A N}{\lambda} E_{rms} \cos\theta \rightarrow (7)$$

where V_{rms} = rms value of induced emf in the loop [in Volts]

E_{rms} = rms value of electric field strength of the wave in Volts.

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λ = Wavelength in meters

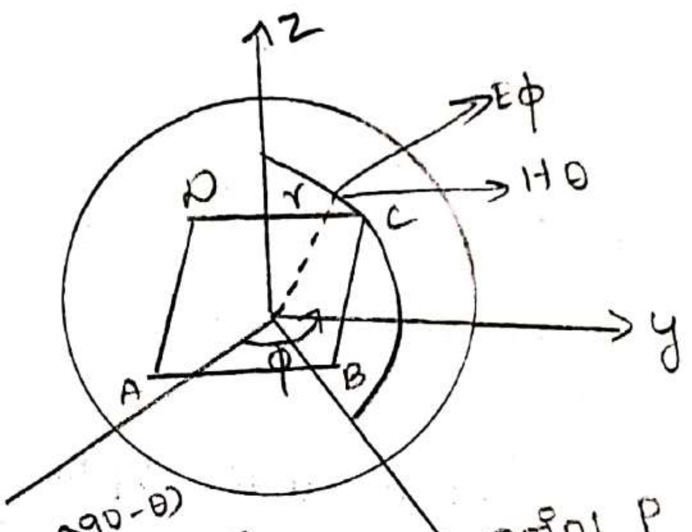
A = Area of the loop [m^2]

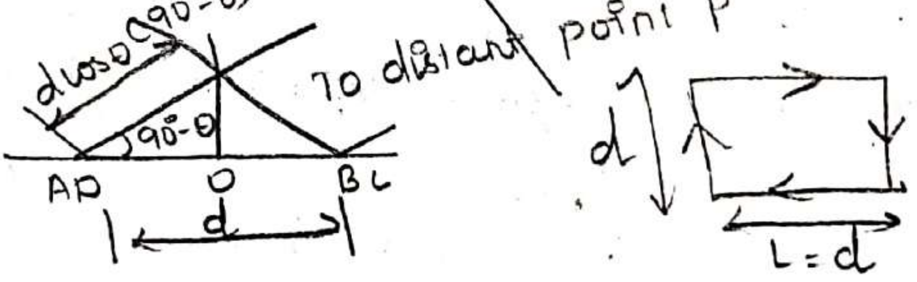
N = No. of. turns

θ = Angle b/w plane and direction of incoming waves.

$\frac{2\pi AN}{\lambda}$ = Effective length/height of an antenna

(ii) Transmitting the loop antenna :·





$2\pi r = c^2$

field pattern of the circular loop (or)

$d^2 = \pi a^2$

d → side length of the square loop.

$E\phi = \left\{ \begin{array}{l} \text{field amp} \\ \text{due to} \\ \text{dipole AD} \end{array} \right\} + \left\{ \begin{array}{l} \text{field component} \\ \text{due to dipole} \\ \text{Isc} \end{array} \right\}$

$E\phi = (-) E_0 e^{j\psi/2} + E_0 e^{-j\psi/2} \longrightarrow (9)$

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$E\phi = (-) 2j E_0 \sin(\psi/2)$

$\psi = \beta d \cos(90° - \theta)$

$\psi = \beta d \sin\theta \rightarrow (10)$

$E\phi = (-) 2j E_\theta \sin\left(\beta d \sin\theta/2\right) \rightarrow (11)$

(i) Efficiency is very low Used as transmitter

(ii) Current in phase throughout the loop

r is much smaller than d

from Equ (11) $E\phi$ = far field of two dipoles

$E_0 \Rightarrow$ Individual dipole.

The term j indicates the total field $E\phi$ is inphase Quadrature w.r.to individual field Component E_0.

fields of the Short Electric dipole.

Components	General Expression	far field
E_r	$\frac{IL\cos\theta}{2\pi\varepsilon_0}\left[\frac{1}{r^2} + \frac{1}{j\omega r^3}\right]$	0
E_θ	$\frac{IL\sin\theta}{4\pi\varepsilon_0}\left[\frac{j\omega}{c^2 r} + \frac{1}{cr^2} + \frac{1}{j\omega r^3}\right]$	$\frac{ILj\omega\sin\theta}{4\pi\varepsilon_0 c^2 r}$ (or) $\frac{j60\pi[I]\sin\theta}{rd}$
$H\phi$	$\frac{IL\sin\theta}{4\pi}\left[\frac{j\omega}{cr} + \frac{1}{r^2}\right]$	$\frac{ILj\omega\sin\theta}{4\pi cr}$ (or) $\frac{jI\sin\theta L}{2rd}$

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$$E_\theta = \frac{j60\pi[I]\sin\theta\, L}{r\lambda} \longrightarrow (12)$$

$$E_\theta = \frac{j60\pi[I]L}{r\lambda} \longrightarrow (13)$$

from Equ (12) θ is measured in X axis which is perpendicular to yz plane

Equ (13) becomes retarded current

$$I = I_m e^{j\omega(t - r/c)} \longrightarrow (6')$$

$r \ll \lambda$ / $d \ll \lambda$ then $\sin\psi/2 = \psi/2$

$$\psi = \beta d \sin\theta$$

Defn:- Retarded Current.

from Equ (11) we get

$$E_\phi = (-)j E_\theta \sin\left(\frac{\beta d \sin\theta}{2}\right)$$

$$= (-j)\left[\frac{j\,60\pi[I]L}{r\lambda}\right]\sin\theta$$

$$= (-)j\left[\frac{j60\pi[I]L}{r\lambda}\right]\sin\theta$$

$$E_\phi = \frac{60\pi[I]L\beta d\sin\theta}{r\lambda} \longrightarrow (14)$$

$\therefore L = d \quad d^2 = A \quad \beta = 2\pi/\lambda$

$$E_\phi = \frac{60\pi[I]d^2 \times 2\pi\sin\theta}{r\lambda^2}$$

$$E_\phi = \frac{120\pi^2[I]A\sin\theta}{r\lambda^2} \longrightarrow (15)$$

This is the instantaneous value of the E_ϕ component of the far field of a small loop area A

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from $E\phi \rightarrow H\theta$

$$\eta = \frac{E\phi}{H\theta}$$

$$\eta = 120\pi$$

$$\eta = \frac{E\phi}{H\theta}$$

$$120\pi = \frac{120\pi^2 [I] A \sin\theta}{H\theta}$$

$$H\theta = \frac{\pi [I] A \sin\theta}{r d^2}$$

Comparison of far field of small loop antenna and Short dipole antenna

field	Electric dipole	loop
EF	$E_\theta = j60\pi \frac{[I] \sin\theta L}{r d}$	$\frac{120\pi^2 [I] A \sin\theta}{r d^2}$
MF	$\frac{jI \sin\theta L}{2 r d}$	$\frac{\pi I A \sin\theta}{r d^2}$

Radiation Resistance of the loop antenna :-

$$P = I^2_{rms} . R_r$$

$$P = \left(\frac{I_m}{\sqrt{2}}\right)^2 R_r$$

$$P = \frac{1}{2} I^2 m . R_r$$

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

$$R_r = 31200 \left[\frac{NA}{\lambda}\right]^2 \Omega$$

$$R_r = 20\pi^2 [C/\lambda]^4 \Omega$$

$$R_r = 197 [C/\lambda]^4 \Omega$$

$$W = 0.682 [C/\lambda]$$

$$A_e = 0.0543 C \lambda$$

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Log Periodic Antenna

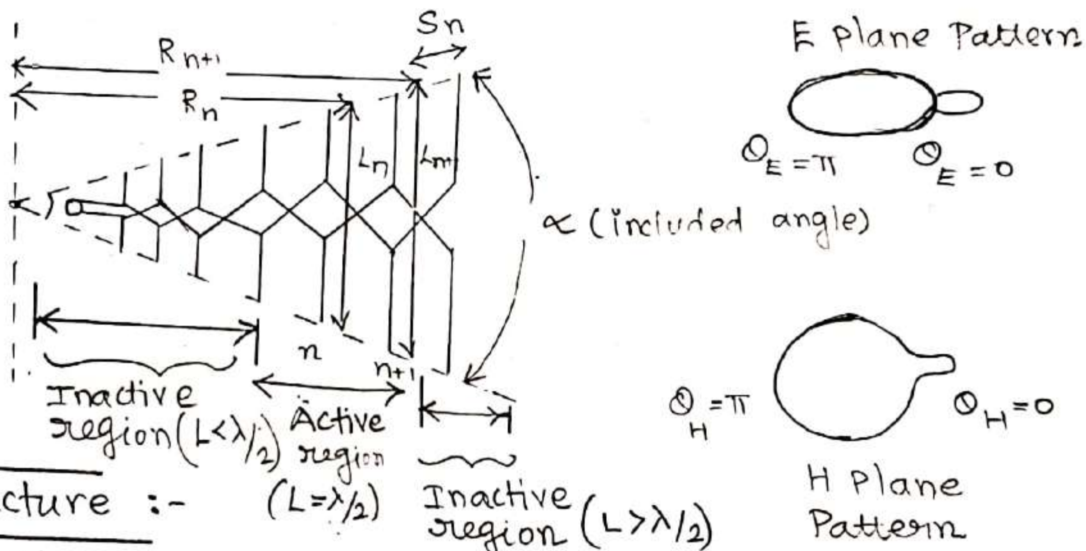
(Log Periodic Dipole Array Antenna)

[LPDA]

Frequency Independent Antennas.

* LPA are broad-band antennas.
* The electrical properties must repeat periodically with the log of the frequency.
* Freq independence can be obtained when the variation of th properties over one period(over all the periods) is small.

H-Plane View



Design & structure :-

The Design of LPA involves a basic geometric structu that is repeated with a changing size of the structure. The structure size changes with each repetition by a "constant scale factor". i.e the structure expands or contracts by c constant scale factor.

LPDA Structure :-

(1) It has no. of dipoles of diff lengths and spacing.

(2) It is fed by a 2wire balanced tx. line which is transposed betwee each adj pair of dipoles. It is usually fed at narrow en

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(3) All the dimensions increase in proportion to the distance from the origin.

(4) The dipole length also increases along the length of the antenna.

α = included angle

L_n = length of the n^{th} dipole antenna

R_n = distance of the n^{th} antenna from origin.

S_n = spacing between n^{th} antenna & $(n+1)^{th}$ antenna.

L, R, S are related as

$$\frac{R_1}{R_2} = \frac{R_2}{R_3} = \frac{R_3}{R_4} = \cdots = \frac{R_n}{R_{n+1}} = \tau$$

$$\frac{L_1}{L_2} = \frac{L_2}{L_3} = \frac{L_3}{L_4} = \cdots = \frac{L_n}{L_{n+1}} = \tau$$

i.e

$$\boxed{\frac{R_n}{R_{n+1}} = \frac{L_n}{L_{n+1}} = \frac{S_n}{S_{n+1}} = \tau}$$

τ is called as scale factor (or) design ratio (or) Peroiditi-city factor. usually $0 < \tau < 1$.

* usually the ends of the dipoles lie along 2 st lines. These st lines meet at angle α at one end and converge at other end.

* Typical value of $\alpha = 30^\circ$; $\tau = 0.7$.

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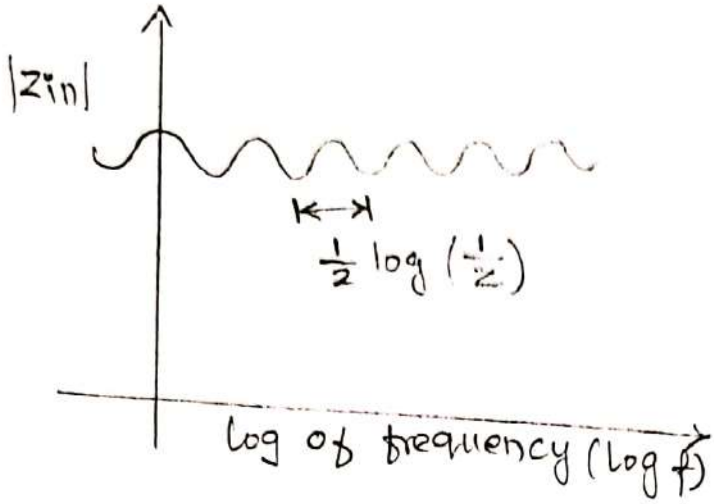
* In the plot of Z_{in} vs f, a repetitive variation can be observed. If the plot is made against $\log f$, then this variation will be periodic. (i.e) the i/p impedance Z_{in} will go through identical cycles of variations. It is shown below.



* all the electrical properties like radiation pattern, directive gain, side lobe level, beam width undergo simil periodic variation.

If impedance variation occurs at 2 frequecies f_1, f_2 then

$$\log \frac{f_2}{f_1} = \log \frac{1}{\tau}$$

(or)

$$\frac{f_2}{f_1} = \frac{1}{\tau}$$

i.e $\quad f_1 = \tau f_2 \qquad f_2 > f_1$

whatever properties a log periodic antenna is having at freq f_1, the same properties will be repeated at freq given by $(\tau^n f)$ or at $\frac{f}{\tau^n}$.

[assuming these freq are within cutoff limits of antenna]

* practically LPDA will have cut off freq due to limitations in size, spacing of conductors.

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Design of LPDA

1) **design ratio (τ) :-**

$$\frac{L_n}{L_{n+1}} = \frac{R_n}{R_{n+1}} = \frac{S_n}{S_{n+1}} = \frac{d_n}{d_{n+1}} = \frac{a_n}{a_{n+1}} = \tau \quad \ldots\ldots (A)$$

L_n : length of dipole antenna (n^{th})

R_n : distance of " " from the origin.

S_n : spacing between antenna n & $(n+1)$.

d_n : diameter of antenna (n^{th})

a_n : gap spacing at dipole centre of n^{th} antenna.

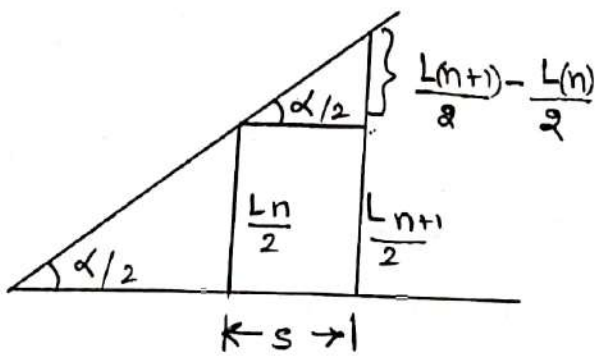
where $n = 1, 2, 3 \ldots\ldots$

2) **Spacing factor (σ) :-**

$$\sigma = \frac{R_{(n+1)} - R_n}{2L_n} = \frac{S_n}{2L_n} \quad \ldots\ldots (B)$$

i.e $\frac{L_{n+1}}{L_n} = \frac{S_{n+1}}{S_n} = k = \frac{1}{\tau}$ k is a constant.

[at any given frequency, only the fraction of antenna is used. (i.e antenna in active region



Section of LPDA

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from the above fig

$$\tan\left(\frac{\alpha}{2}\right) = \frac{\text{opp. side}}{\text{Adj. side}} = \frac{\frac{L_{n+1} - L_n}{2}}{S} = \frac{L_{(n+1)} - L_n}{2S} \quad \ldots (c)$$

from eqn(b),

$$\frac{L_{n+1}}{K} = L_n$$

for active region $L_{n+1} \simeq \lambda/2$

from eqn(c), $$\tan(\alpha/2) = \frac{L_{n+1} - \frac{L_{n+1}}{K}}{2S} = \frac{L_{n+1}\left[1 - \frac{1}{K}\right]}{2S}$$

$$= \frac{\frac{\lambda}{2}\left[1 - \frac{1}{K}\right]}{2S} = \frac{\left[1 - \frac{1}{K}\right]}{4S/\lambda}$$

$$\therefore \tan\left(\frac{\alpha}{2}\right) = \frac{1-\tau}{4\cdot(S/\lambda)}$$

$$\Rightarrow \left(\frac{S}{\lambda}\right) = \frac{1-\tau}{4\tan(\alpha/2)}$$

i.e $$\boxed{\sigma = \frac{1-\tau}{4\tan(\alpha/2)}}$$

(or) $$\tan \alpha/2 = \frac{1-\tau}{4\sigma}$$

$$\alpha/2 = \tan^{-1}\left[\frac{1-\tau}{4\sigma}\right]$$

$$\boxed{\alpha = 2\tan^{-1}\left[\frac{1-\tau}{4\sigma}\right]}$$

α : apex angle

K : Scale Factor

$S/\lambda(\sigma)$: spacing in wavelength

also $$\boxed{\tau = 1/K = \frac{S_n}{S_{n+1}} = \frac{L_n}{L_{n+1}}}$$

if out of (σ, α, τ) any 2 are specified, third canbe found.

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We know,

$$\frac{L_2}{L_1} = K \quad ; \quad \frac{L_3}{L_2} = K$$

$$\therefore \quad \frac{L_3}{L_1} = K.K = K^2$$

$$\therefore \quad \frac{L_n}{L_1} = K^{(n-1)} \Rightarrow \frac{L_{n+1}}{L_1} = K^n$$

$$\frac{L_{n+1}}{L_1} = F$$ (Frequency ratio or BW.)

Ex:- for optimum Design, for $n=4$

$$K = 1.19$$

$$\therefore \quad F = K^n = (1.19)^4 = 2.0053$$

$$\therefore \quad F \simeq 2$$

$\therefore$ no of elements $= n+1 = 4+1 = 5$.

Hence for 5 elements dipole array & $K = 1.19$, the BW is 2:1.

Analysis of LPDA

There are 3 regions exist for LPDA.

(i) Inactive transmission line region ($L < \lambda/2$)

(ii) Active region ($L \simeq \lambda/2$)

(iii) Inactive reflective region ($L > \lambda/2$)

* radiation from LPDA is always in backward direction.

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General characteristics

1) LPDA is fed by a balanced 2-wire tx.line. always excited from the shorter length side or high fre side.

2) Broadband will be with those LPDA which have small variation in periodicity properties.

3) Unidirectional LPDA - radiation is in backward dir. towards shorter element.

Bidirectional LPDA - maximum radiation is in Broad side direction.

4) Tx.line inactive region (between active and vertex) must have proper impedance with negligible radiation.

5) In active region, the currents magnitude and phase should be proper so that pt(3) is satisfied.

Typical value : $\lambda/4$ spacing, 90° phase (unidirectional)
0° phase (bidirectional)

6) In ~~inac~~ inactive reflective region, there should be rapid decay of current. (with in the reflective region).

Applications

(i) HF communication. No power is wasted in terminating resistance.

(ii) LPDA in TV reception. only one LPDA is enough upl UHF band.

(iii) If the cost of installation is not considered, then all round monitoring can be done. [one LPDA will cover all the higher frequency bands].

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Co-Ordinate System:

- Cartesian Coordinate system (x, y and z)
- Cylindrical Coordinate system (r, θ and z)
- Spherical Coordinate system (ρ, θ and φ)

Aperture Antenna (Horn Antenna)

Horn antennas are very popular at UHF (300 MHz-3 GHz) and higher frequencies (I've heard of horn antennas operating as high as 140 GHz). Horn antennas often have a directional radiation pattern with a high antenna gain, which can range up to 25 dB in some cases, with 10-20 dB being typical. Horn antennas have a wide impedance bandwidth, implying that the input impedance is slowly varying over a wide frequency range (which also implies low values for S11 or VSWR). The bandwidth for practical horn antennas can be on the order of 20:1 (for instance, operating from 1 GHz - 20 GHz), with a 10:1 bandwidth not being uncommon.

The gain of horn antennas often increases (and the beamwidth decreases) as the frequency of operation is increased. This is because the size of the horn aperture is always measured in wavelengths; at higher frequencies the horn antenna is "electrically larger"; this is because a higher frequency has a smaller wavelength. Since the horn antenna has a fixed physical size (say a square aperture of 20 cm across, for instance), the aperture is more wavelengths across at higher frequencies. And, a recurring theme in antenna theory is that larger antennas (in terms of wavelengths in size) have higher directivities.

Horn antennas have very little loss, so the directivity of a horn is roughly equal to its gain. Horn antennas are somewhat intuitive and relatively simple to manufacture. In addition, acoustic horn antennas are also used in transmitting sound waves (for example, with a megaphone). Horn antennas are also often used to feed a dish antenna, or as a "standard gain" antenna in measurements. Popular versions of the horn antenna include the E-plane horn, shown in Figure 1. This horn antenna is flared in the E-plane, giving the name. The horizontal dimension is constant at **w**.

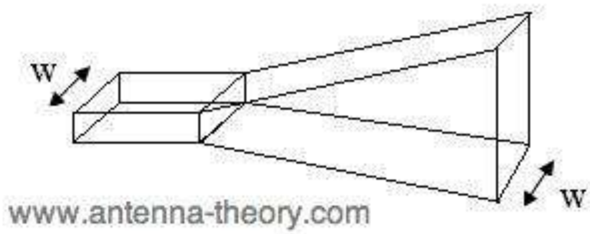

Figure 1. E-plane horn antenna.

Another example of a horn antenna is the H-plane horn, shown in Figure 2. This horn is flared in the H-plane, with a constant height for the waveguide and horn of *h*.

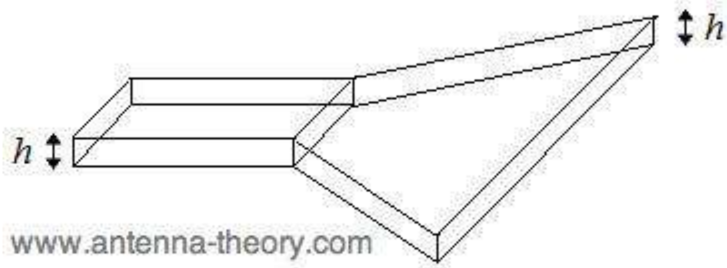

Figure 2. H-Plane horn antenna.

The most popular horn antenna is flared in both planes as shown in Figure 3. This is a pyramidal horn, and has a width *B* and height *A* at the end of the horn.

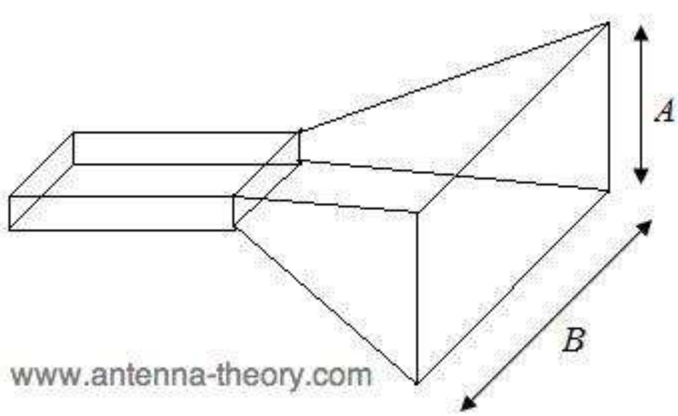

Figure 3. Pyramidal horn antenna.

Horn antennas are typically fed by a section of a waveguide, as shown in Figure 4. The waveguide itself is often fed with a short dipole, which is shown in red in Figure 4. A waveguide is simply a hollow, metal cavity (see the waveguide tutorial). Waveguides are used to guide electromagnetic energy from one place to another. The waveguide in Figure 4 is a rectangular waveguide of width b and height a, with $b>a$. The E-field distribution for the dominant mode is shown in the lower part of Figure 1.

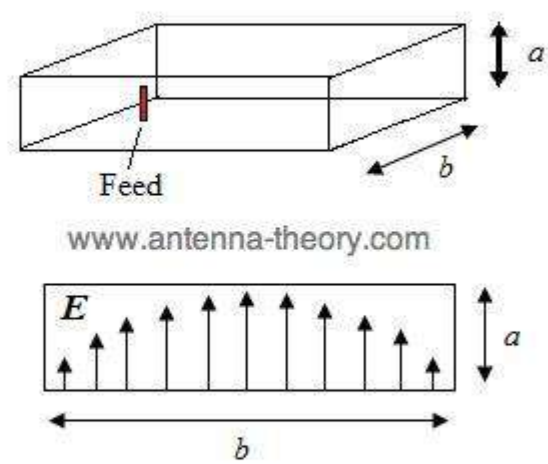

Figure 4. Waveguide used as a feed to horn antennas.

Fields and Geometrical Parameters for Horn Antennas

Antenna texts typically derive very complicated functions for the radiation patterns of horn antennas. To do this, first the E-field across the aperture of the horn antenna is assumed to be known, and the far-field radiation pattern is calculated using the radiation equations. While this is conceptually straight forward, the resulting field functions end up being extremely complex, and personally I don't feel add a whole lot of value. If you would like to see these derivations, pick up any antenna textbook that has a section on horn antennas. (Also, as a practicing antenna engineer, I can assure you that we never use radiation integrals to estimate patterns. We always go on previous experience, computer simulations and measurements.)

Instead of the traditional academic derivation approach, I'll state some results for the horn antenna and show some typical radiation patterns, and attempt to provide a feel for the design parameters of horn antennas. Since the pyramidal horn antenna is the most popular, we'll analyze that. The E-field distribution across the aperture of the horn antenna is what is responsible for the radiation.

The radiation pattern of a horn antenna will depend on B and A (the dimensions of the horn at the opening) and R (the length of the horn, which also affects the flare angles of the horn), along with b and a (the dimensions of the waveguide). These parameters are optimized in order to taylor the performance of the horn antenna, and are illustrated in the following Figures.

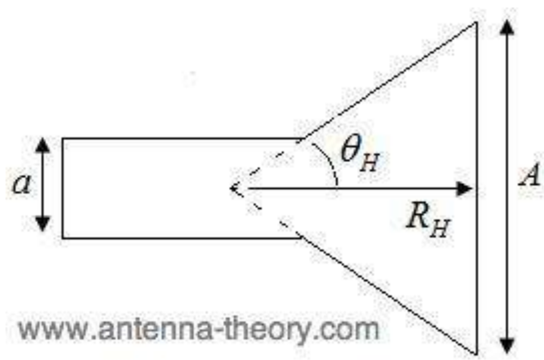

Figure 5. Cross section of waveguide, cut in the H-plane.

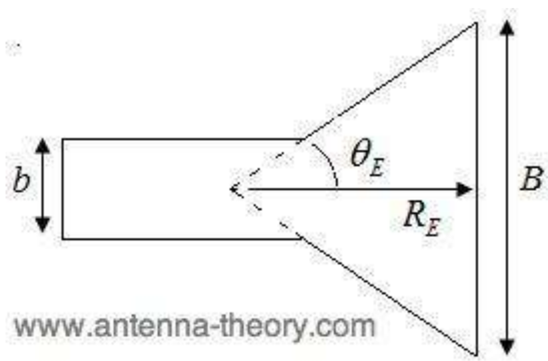

Figure 6. Cross section of waveguide, cut in the E-plane.

Observe that the flare angles (θ_E and θ_H) depend on the height, width and length of the horn antenna.

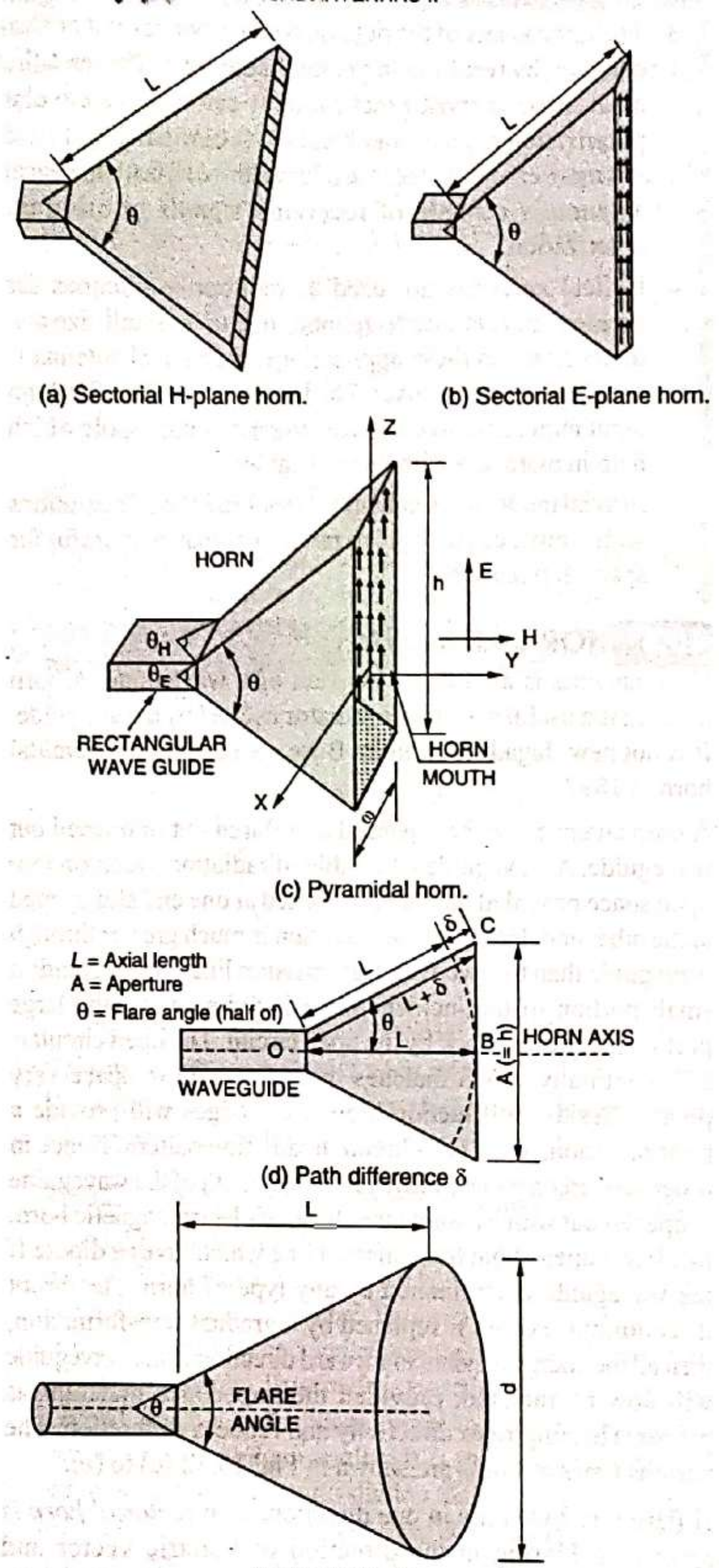

Fig. 10.32. (a to e) Important horn shapes.

However, this may be treated as transition region where the change over from the guided propagation to free space propagation occurs. Since the waveguide impedance and free space impedance are not equal, *hence to avoid standing wave ratio, flaring of walls of waveguide is done which besides matching of impedance, also provides concentrated radiation pattern i.e. greater directivity and narrower beamwidth. It is the flared structure that is given the name electromagnetic horn radiator.*

The function of the electromagnetic horn is to produce a uniform phase front with a larger aperture in comparison to waveguide and thus the directivity is greater. Although the principle of equality of path length is applicable to horn design but in different sense *i.e.* instead of specifying that the wave over the *plane of the horn mouth is in phase exactly, we allow that phase may deviate but by an amount less than specified amount.* From the geometry of the Fig. 10.32 (*d*), we have

$$\cos\theta = \frac{L}{L+\delta} \text{ and } \tan\theta = \frac{h/2}{L} \text{ or } \tan\theta \frac{h}{2L}$$

$$\theta = \tan^{-1}\left(\frac{h}{2L}\right) = \cos^{-1}\left(\frac{L}{L+\delta}\right) \quad ...(10.59)$$

where δ = Permissible phase angle variation expressed as fraction of 360°.

From right angled triangle *OBC* [Fig. 10.30(*d*)]

$$(L+\delta)^2 = L^2 + \left(\frac{h}{2}\right)^2 \text{ or } L^2+\delta^2+2L\delta = L^2+\frac{h^2}{4}$$

If δ is small, then δ^2 can be neglected

$$2L\delta = \frac{h^2}{4}$$

or

$$L = \frac{h^2}{8\delta} \quad ...(10.60)$$

Eqns. (10.59) and (10.60) give the design equations of the horn antenna. If flare angle (2θ) is very large, the wave front on the mouth of the horn will be curved rather than plane. This will result in non uniform phase distribution over the aperture, resulting increased beam width and decreased directivity, and *vice-versa* occurs directivity is proportional to the aperture size for a given aperture distribution. Thus there is optimum aperture angle given by Eqn. (10.59). The maximum directivity is achieved at the largest flare angle for which in δ are 0.25, 0.32, 0.40 for plane horn, conical horn and H-plane horn respectively.

As customary for E-plane horn, phase difference upto 72° (*i.e.* ± 36° variation) for δ less than 0.20 λ and for H-plane phase difference upto 135° for δ less than 0.375 λ are allowed. In practice 2θ varies from 40° to 15° which gives beamwidth 66°. Directivity 40 for $L = 6\lambda$ and beamwidth 23°, gain 120 for $L = 50\lambda$. Directivity with pyramidal or conical horn antenna increases as they have more than one flare angle. However, the directivity of parabolic antenna is more than the horn antenna. As there is no resonant element involved in the horn antennas hence they can be operated over a broad band of frequency.

Although derivation of exact relation for beam width of horn antenna is possible yet approximate formulae for the half power beamwidth of optimum flare horns are as follow [refer Fig. 10.32 (*c*)].

$$\theta_E = \frac{56\lambda}{h} \text{ degree} \quad ... [10.61\,(a)]$$

and

$$\theta_H = \frac{67\lambda}{\omega} \text{ degree} \quad ... [10.61\,(b)]$$

where θ_E and θ_H are HPBW in *E* and *H* directions. Thus the directivity is given by

$$D = \frac{7.5\,h\cdot\omega}{\lambda^2} = \frac{7.5\,A}{\lambda^2} \quad ... [(10.62)]$$

where $A = h \times \omega$
= area of horn mouth opening (aperture).

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and power gain

$$G_P = \frac{4.5\, h \cdot \omega}{\lambda^2} = \frac{4.5\, A}{\lambda^2} \quad \text{... [(10.63)]}$$

10.6.1. Uses of Horn Antenna

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

10.6.2. Application of Horn Antennas

1. Horn antennas are used as feed element for parabolic reflectors and lenses.
2. Most suitable antennas for various applications in microwave frequency range where moderate gains are sufficient enough.
3. Most widely used for measurement of different parameters in the laboratories like gain etc. It is used for calibration and gain measurement of other antennas and such horn antennas are known as 'Standard gain horn antennas'.
4. The horn is widely used as a feed element for parabolic dishes which are used for large radio astronomy, satellite tracking, and communication dishes found installed all over the world.
5. It is widely used for microwave frequencies (3 GHz and above) because of their moderate gain and low USWR.

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10.7. BABINET'S PRINCIPLE AND COMPLEMENTARY ANTENNAS

One may enquire whether there is any relation between wire antenna and aperture antenna, the same can be answered better by first introducing Babinet's Principle of optics. The Babinet's (Ba-bi-nay's) in optics states that "*when the field behind a screen with an opening is added to the field of a complementary structure, the sum is equal to the field when there is no screen*". Babinet principle in optics does not consider polarization, which is so vital in antenna theory. It deals primarily with absorbing screens. An extension of Babinet's principle, which induces polarization and the more practical conducting screens, was introduced by Booker. By introduction of Babinet's principle many of the problems of slot antennas can be reduced to situation

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involving complementary linear antennas for which solutions have already been obtained.

The Baninet's principle may be illustrated by considering the following example with three cases. Let a source and two imaginary planes be arranged as shown in Fig. 10.35 in which the first plane is a plane of screens S_1 and the plane is a plane of observation S_2. Now three cases arise.

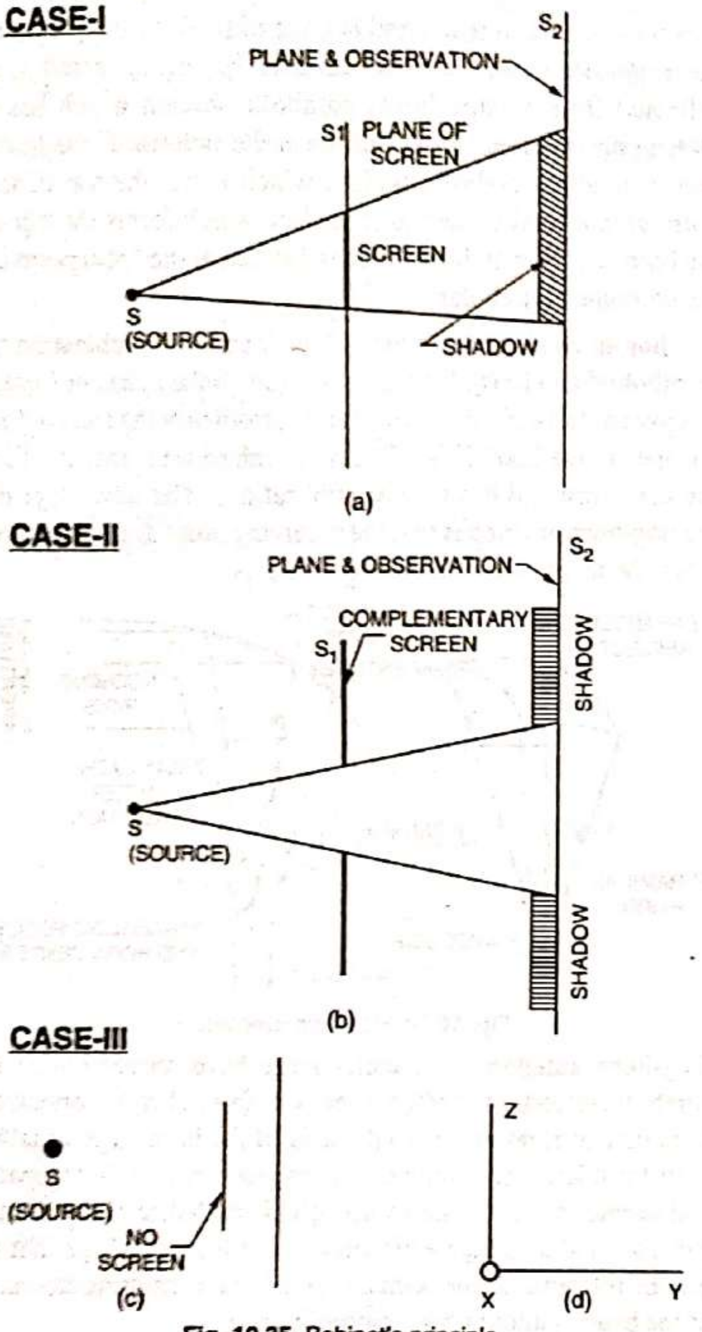

Fig. 10.35. Babinet's principle.

Case I. Let a perfectly absorbing screen be placed in plane S_1 then in plane, there is a region of shadow as shown. Let the field behind this screen be some function of f_1 (x, y, z) *i.e.* be replaced by its complementary screen and the field behind it be given by

$$F_1 = f_1(x\,y\,z) \qquad ...(10.64)$$

Case II. Let the first screen S_1 be replaced by its complementary screen and the field behind it be given by

$$F_2 = f_2(x\,y\,z) \qquad ...(10.65)$$

Case III. Let there is no screen present, then the field is given by

$$F_3 = f_3(x\,y\,z) \qquad ...(10.66)$$

Babinet's principle then states that at the same point $(x\,y\,z)$

$$F_3 = (x\,y\,z) = F_1\,(x\,y\,z) + F_2\,(x\,y\,z) \qquad ...(10.66)$$

or

$$F_3 = F_1 + F_2 \qquad ...(10.67)$$

The source may be a point as in the above example or a distribution of sources. The principle applies not only to point in the plane of observation S_2 as outlined in Fig. 10.34 but also to any point behind screen S_1. The principle is obvious enough for shadow (Case I), it is also true when diffraction is taken into account.

The correctness of this valid statement [Eqn. (10.67)] can be verified easily for the simple cases of complementary screens consisting of semi-infinite absorbing planes.

In electromagnetics at ratio frequencies, thin perfectly absorbing screens are not available, even approximately and one is concerned with *conducting screens and vector fields* for which polarization plays an important role. As such the simple statement of optics could not be expected to apply but an extension of the principle, valid for conducting screens and polarized fields has been formulated by H.G. Booker.

As an illustration of Booker's extension of Babinet's principle, let us consider the following three cases shown in Fig. 10.36. The source (s) in all the three cases is a short dipole, theoretically infinitesimal dipole.

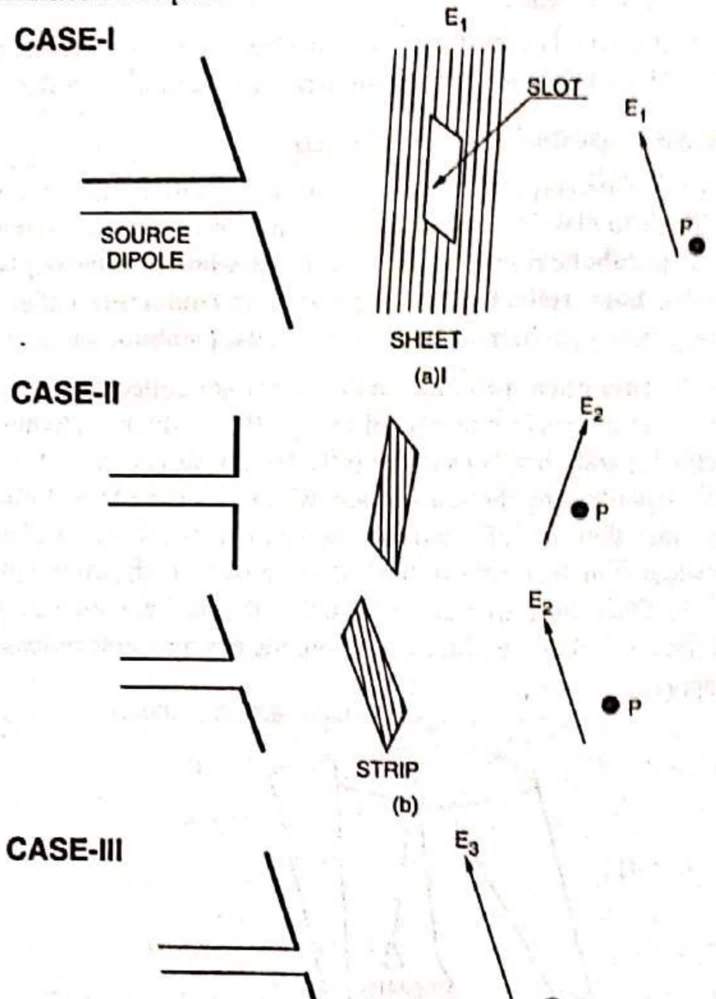

Fig. 10.36. Extension of Babinet principle for slot of infinite metal sheet and the complementary metal strip.

Case I. The dipole is horizontal and original screen is an infinite, perfectly conducting, plane, infinitesimally thin sheet with a

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vertical slot cut out. At a point P behind the screen the field is E_1.

Case II. In this case the original screen is replaced by the complementary screen consisting of a perfectly conducting, plane infinitesimally thin strip of the **same dimensions as the slot in the original screen.** Besides, the dipole is source and is turned vertical so that $\vec{E}$ and $\vec{H}$ are interchanged. At the same point P, behind the screen the field is E_2.

Alternatively, the dipole source is turned horizontal and so also the strip.

Case III. In this case, no screen is placed and the field at point P is E_3.

According to Baninet's principle

$$E_1 + E_2 = E_3$$

or

$$\frac{E_1}{E_3} + \frac{E_2}{E_3} = 1 \quad ...(10.68)$$

The principle may also be applied to points in front of the screens, In Case-I, a large amount of energy may be transmitted through the slot so that $E_1 \simeq E_3$. In such situation the complementary dipole (Case-II) acts like a reflector and E_2 is very small.

Using Booker's extension, it can be shown that if a screen and its complementary are immersed in a medium with an intrinsic impedance η and have terminal impedances of Z_S (screen) and Z_C (complementary) respectively, then the impedances are related by

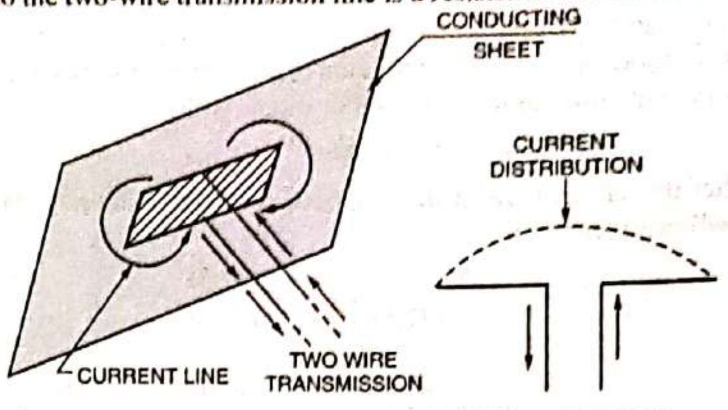
$$Z_S Z_C = \frac{\eta^2}{4} \quad ...(10.69)$$

In order to obtain the impedance Z_C of the complementary dipole in practical arrangement a gap must be introduced to represent the feed points.

10.8. SLOT ANTENNAS

The slot antenna, as its name suggests, is a simply an opening cut in a sheet of conductor which is energized in some appropriate manner, such as via a coaxial cable or waveguide. One simple type of slot antenna is a half wavelength long with narrow width and excited via a 50 ohm coaxial cable normally connected about 0.05λ from one end of the slot to achieve reasonable matching conditions. A horizontal slot so energized produces vertical polarization in the direction normal to the slot, and a vertical slot produces horizontal polarization. Radiation occurs from both sides of the conductive sheet but if the slot is "boxed" with internal dimension of depth $d = \lambda/4$, the radiation is outwards from the opening of the box. A single half wavelength slot in many ways resembles the half wave dipole in terms of gain and radiation except that there is a difference in polarization. In order to enhance the gain and directive properties of the basic slot antenna, it is common to have arrays of slots in a manner similar to the arrays of dipoles. Some VHF transmitters employ cylindrical arrays of slots to produce omni-directional radiation in the horizontal plane with horizontal polarization.

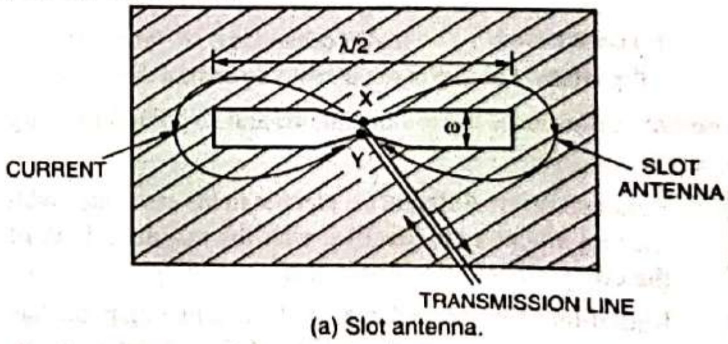
The slot antenna makes use of the fact that energy is radiated when a high frequency field exists across a narrow slot in a conducting plane. A typical slot antenna is shown in Fig. 10.37. Here the fields are excited by a two-wire transmission line. The electric field across the slot is maximum at the centre and tapers off towards the edges as indicated, while at sometime currents flow in the conducting plane in the general manner indicated when the slot is exactly half wavelength long, the electric field distribution is sinusoidal and the impedance offered by the slot to the two-wire transmission line is a resistance of 365 ohm.



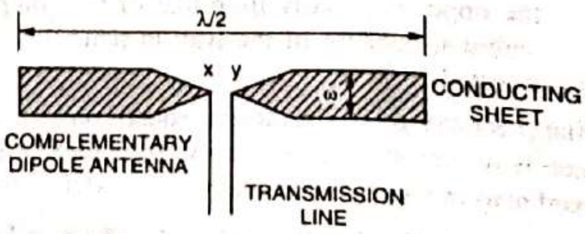
(a) Slot antenna, (b) Complementary wire antenna.

Fig. 10.37.

If a λ/2 slot is cut in a large metal sheet and a transmission line connected to the point XY as shown in Fig. 10.38 (*a*), the arrangement will radiate effectively due to currents flowing on the sheet. The analysis of *such a slot antenna is greatly facilitated by considering the slot's complementary antenna.* Therefore, the antenna which is complementary to the slot of Fig. 10.38 (*a*) is the dipole of Fig. 10.38 (*b*). The metal and air regions of the slot are interchanged for the dipole. According to the G.Booker's theory the pattern of the slot of Fig. 10.38 (*a*) is identical in shape to that of the dipole of Fig. 10.38 (*b*) **except that the electric field will be vertically polarized for the slot and horizontally polarized for the dipole.** Besides, the terminal impedance Z_s of the *slot* is related to the terminal impedance of *dipole* (Z_d) by intrinsic impedances η_0 of free space by the relation



(a) Slot antenna.



(b) Complementary dipole antenna.

Fig. 10.38. Slot and complementary dipole antenna of long λ/2 and width w, fed at point XY.

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$$Z_s Z_d = \frac{\eta_0^2}{4} \qquad \text{...(10.70)}$$

$$Z_s Z_d = \frac{(376.7)^2}{4} = \frac{141902.89}{4} = 35475.722$$

or $$Z_s \approx \frac{35476}{Z_d} \text{ Ohms} \qquad \text{...(10.71)}$$

Hence by knowing the properties of dipole antennas, the properties of the complementary slot antenna can be determined. For example, let the width of the dipole and slot of Fig. 10.38 be reduced to a very small fraction of a wavelength so that the dipole qualities as a thin λ/2 linear dipole with

$$Z_d = 73 + j\,(42.5) \text{ Ohms}$$

then the terminal impedance of the complementary slot antenna will be given by

$$Z_s = \frac{35476}{73 + j\,(42.5)} \times \frac{73 - j\,(42.5)}{73 - j\,(42.5)}$$

$$= \frac{35476}{(73)^2 + (42.5)^2}(73 - j\,42.5)$$

$$= 4.9719\,(73 - j\,42.5)$$

$$Z_s \simeq 363 - j\,211 \text{ Ohms}$$

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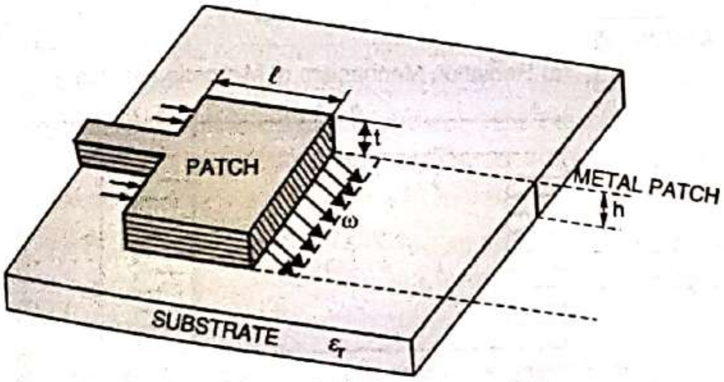
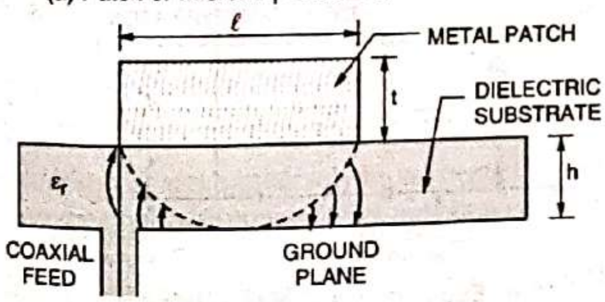
10.10. MICROSTRIP OR PATCH ANTENNAS

In spacecraft or aircraft applications, where size, weight, cost, performance, ease of installation, and aerodynamic profile are constraints, low profile antennas are required. In order to meet these specifications microstrip or patch antennas are used. These antennas can be flush-mounted to metal or other existing surfaces and they only require space for the feed line which is normally placed behind the ground plane. The major disadvantages of patch or microstrip antennas are their inefficiency and very narrow frequency bandwidth which is typically only a fraction of a percent or at the most a few percent.

Microstrip or patch antennas are popular for low profile applications at frequencies above 100 MHz (or $\lambda_0 < 3$ m). They usually consist of a rectangular metal patch on a dielectric-coated ground plane (circuit board). Microstrip antennas consist of a very thin metallic strip (patch) ($t << \lambda$) placed on a small fraction of wavelength ($h << \lambda$) above a ground plane. The strip (patch) and the ground plane are separated by a dielectric sheet referred to as the substrate, Fig. 10.49. The radiating element and the feed lines are normally photoetched on the dielectric substrate. The radiating patch may be square, circular, elliptical, rectangular or any shape. However, square, circular or rectangular are mostly preferred because of the ease of analysis and fabrication and their attractive radiation characteristics, especially low cross polarization radiation.. The feed line is also a conducting strip normally of smaller width. Coaxial line feeds where the inner conductor of the coaxial line is attached to the radiating patch are widely used. Linear and circular polarization can be achieved with microstrip or patch antennas and arrays of microstrip elements with single or multiple feeds may be used for greater directivity.



(a) Patch or Microstrip antenna.



(b) Side view of patch antenna with feed at the left edge.

Fig. 10.49.

As the thickness of the Microstrip is normally very small, the waves generated within the dielectric (substrate between the patch and the ground plane) undergo reflections to some extent when they arrive at the edge of the strip, resulting in radiation of only small fraction of the incident energy. Therefore, the antenna is considered to be very inefficient and it behaves more like a cavity rather than a radiator.

The patch antenna acts as a resonant $\lambda/2$ parallel plate microstrip transmission line with characteristic impedance equal to the reciprocal of the number n of ***parallel field cell transmission lines.*** Each field transmission line has a characteristic impedance Z_0 equal to intrinsic impedance of the medium *i.e.*

$$Z_0 = \eta_i = \sqrt{\frac{\mu}{\varepsilon}} = \sqrt{\frac{\mu_0}{\varepsilon_0}}\sqrt{\frac{\mu_r}{\varepsilon_r}}$$

$$[\because \mu = \mu_0 \mu_r, \quad \varepsilon = \varepsilon_0 \varepsilon_r]$$

$$Z_0 = 120\pi\sqrt{\frac{\mu_r}{\varepsilon_r}} \qquad \text{... [10.82 (a)]}$$

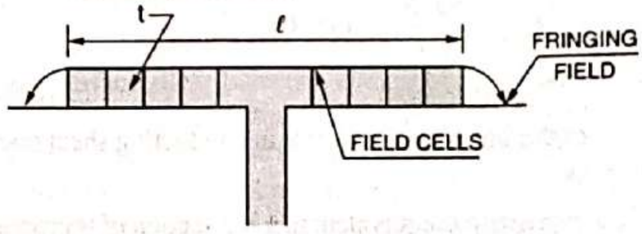


Fig. 10.50. Microstrip antenna with gap or slot divided into n-square field cells.

It is obvious from the patch (from left to right, Fig. 10.50), that the cross-section has 10 field cells transmission lines, hence for $\varepsilon_r = 2$ the characteristic impedance of patch antenna is given by

$$Z_c = \frac{Z_0}{\eta\sqrt{\varepsilon_r}} = \frac{376.7}{10\sqrt{2}} = 26.63 \text{ ohm} \qquad \text{...(10.83)}$$

Eqn. (10.83) can be written as

$$Z_c = \frac{Z_0 t}{\ell\sqrt{\varepsilon_r}} \qquad \left[\because n = \frac{\ell}{t}\right] \qquad \text{...(10.84)}$$

which is he general expression for Z_c.

In the Eqn. (10.84), fringing effects of the field at the edges has been neglected. As ω is typically even much larger than the t, the fringing effect is small for a patch. However, for a microstrip transmission line, where the ratio (ℓ/t) is smaller, the fringing effect can be accounted for by adding 2-cells, giving a more accurate formula for microstrip line impedance,

$$Z_c = \frac{Z_0 t}{\sqrt{\varepsilon_r}\,\ell} = \frac{Z_0}{\sqrt{\varepsilon_r}\,\ell/t} \qquad \text{...[10.84 (a)]}$$

$$Z_c = \frac{Z_0}{\sqrt{\varepsilon_r}[\ell/t + 2]}$$

The resonant length l of the patch is critical and typically a couple of percentage less than $\lambda/2$, where λ is the wavelength in the dielectric $\left(\lambda = \frac{\lambda_0}{\sqrt{\varepsilon_r}}\right)$. Radiation from the patch occurs as if from 2-slots [Fig. 10.49 (*b*)]. The impedance can be calculated for a case where dielectric constant is air ($\varepsilon_r = 1$). It

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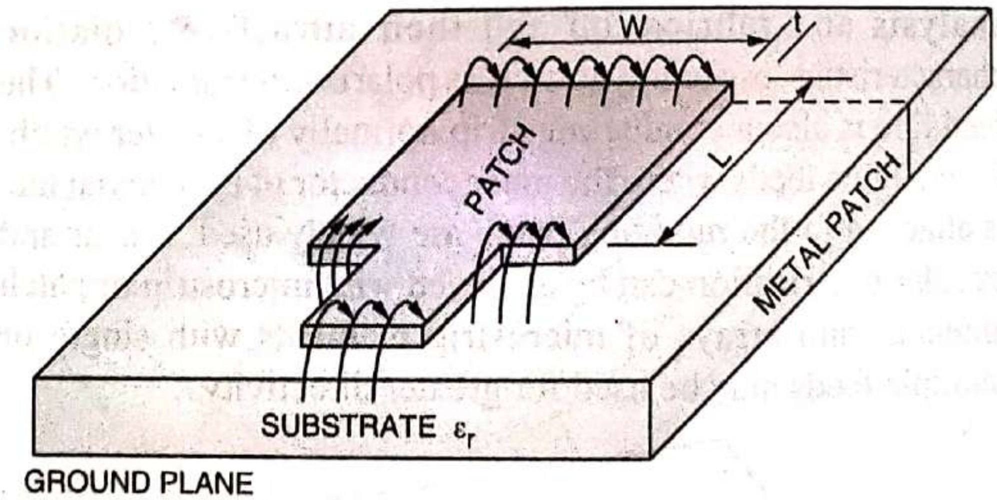
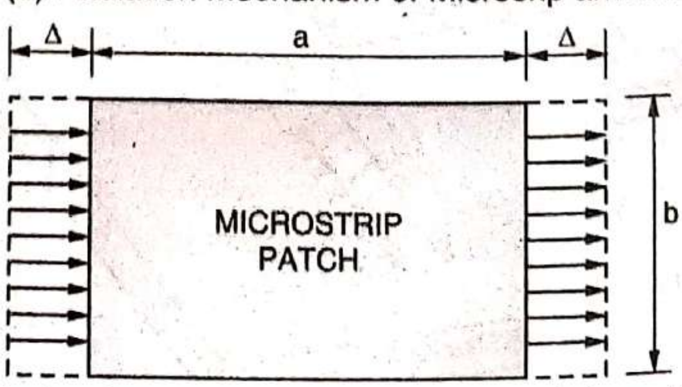
Discontinuities change the electric and magnetic field distributions. Therefore, these result in energy storage and sometimes radiation at the discontinuities. As long as the physical dimensions and relative dielectric constant (ε_r) of the line remain constant, virtually there is no radiation. **However, the discontinuity introduced by the rapid change in line width at the junction between the feed line and patch radiates. Not only this, the other end of the patch where the metallization abruptly ends also radiates.**

When the fields on a microstrip line encounter an abrupt change in width at the input to the patch, electric fields spread out. It creates fringing fields at this edge as shown in Fig. 10.49. After this transition the patch looks like another microstrip line. The fields propagate down this transmission line until the other edge is reached. At this point, the abrupt ending of the line again creates fringing fields as for the open end discontinuity. The fringing fields store energy. The fringing fields store energy. The edge appear as capacitors to ground as the changes in the electric field are greater than that for the magnetic field. As the patch is much wider than a typical microstrip line, the fringing fields also radiate, which is represented by conductance in shunt with the edge capacitance. This accounts for power lost due to radiation as shown in Fig. 10.51.



(a) Radiation Mechanism of Microsrip antenna



(b) Fringing fields at the input and output of the path

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Realization of a microstrip like antenna integrated with microstrip transmission line was developed in 1953 by **Deschamps.** Microstrip antenna design was patented by **Gutton and Baissinot** by 1955. Development of microstrip transmission line analysis and design continued in the mid to late 1960's by **Wheeler and Purcel *et al.* Denlinger in 1969 noted that rectangular and circular microstrip resonator could radiate efficiently.** Microstrip antenna concept atlast began to receive closer attention in the early 1970s when aerospace applications, such as space craft and missiles, produced the impetus for researchers to investigate the utility of conformal antenna designs.

The geometry of microstrip antenna is shown in Fig. 10.53 and Fig. 10.54. A conductive patch exists along the plane of the upper surface of a dielectric slab. **This area of conductor, which forms the radiating element, is generally rectangular or circular but it may be of any shape.** The dielectric substrate has ground plane on its bottom surface. The widespread use of printed circuits led to the idea of constructing radiating elements and interconnecting transmission lines using the same technology. **Thus, antenna made from patches of conducting material on a dielectric substrate above a ground plane is referred to as microstrip antenna. Microstrip antenna is also often sometimes referred to as Patch antenna.**

The patch is typically of rectangular or circular shape with dimensions of order of one-half wavelength. The radiating patch may also be square, diamond, triangle, ring, thin strip (dipole), circular, elliptical or any configuration (Fig. 10.53.) Microstrip dipoles are attractive because they possess inherently a large bandwidth and occupy less space, which make them attractive for arrays. Arrays of microstrip elements, with single or multiple feeds may also be used to introduced scanning capabilities and achieve greater directivities.

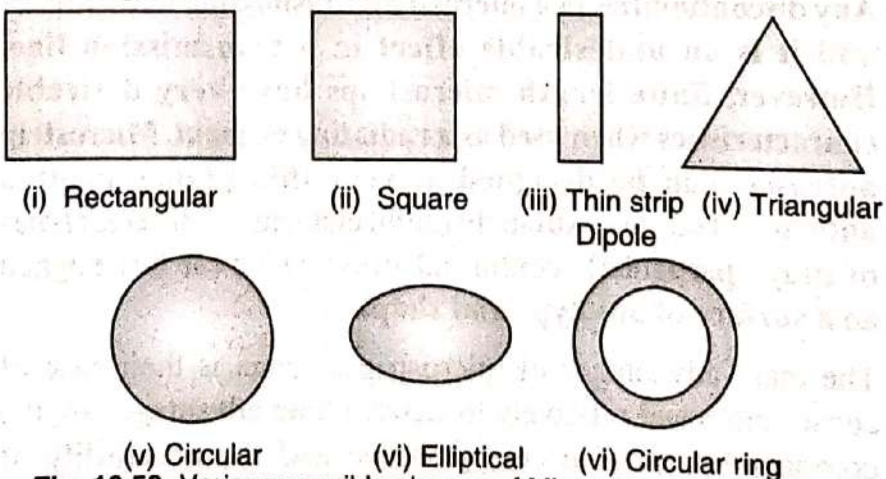

Fig. 10.53. Various possible shapes of Microstrip patch antenna.

The antenna may be fed with microstrip transmission line or a coaxial line as shown in Fig. 10.54. The feed is positioned away from the end by an amount that will give a good impedance match.

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10.10.3. Feeding Methods of Microstrip Patch Antennas

1. Microstrip antennas can be fed in a number of ways. These feeding methods can be classified as

   (*a*) Contacting feed

   (*b*) Non-contacting feed.

   In the former method, the RF power is directly **fed to radiating patch with the help of a microstrip or coaxial line.**

   In the latter method, electromagnetic coupling is done to transfer the power between the feedline and the radiating patch.

2. There are many configurations that can be used to feed microstrip antenna. The most popular feed techniques are:

   (*a*) Microstrip line
   (*b*) Co-axial probe } ...(contacting scheme)

   (*c*) Aperture coupling
   (*d*) Proximity coupling } ...(non-contacting scheme)

3. **Microstrip feed line** is also a conducting strip, normally of much smaller width as compared to the width of patch. Microstrip feed line is easy to fabricate, simple to match by controlling the inset position. This has the advantages that the feed can be etched on the same substrate to provide planar structure. However as the substrate thickness increases, surface waves and spurious feed radiation increases which for practical design limit the bandwidth (typically 2% to 5%). There are many versions of microstrip feeds

   (*a*) centre feed

   (*b*) offset feed

   (*c*) inset feed.

4. **Co-axial feed or probe feed** is a very common technique employed for feeding microstrip patch antenna. In this the inner conductor of the coax is attached to the radiation patch and the outer conductor is connected to the ground

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plane. It is also widely used. The position of the feed can be changed to control the input impedance. The inner conductor of the coaxial connector extends through dielectric and is soldered to the radiating patch.

There are many configurations that can be used to feed microstrip antenna. The most popular feeds are the

Contacting feed

1. Microstrip line feed,
2. Coaxial probe feed,

Non-contacting feed

3. Aperture coupling feed and
4. Proximity coupling feed.

Fig. 10.55 shows several popular feed mechanisms that can be utilized with microstrip antenna. One set of equivalent circuits for each one of the above facts have also been shown in Fig. 10.56. Each feed configuration has its own advantages and disadvantages.

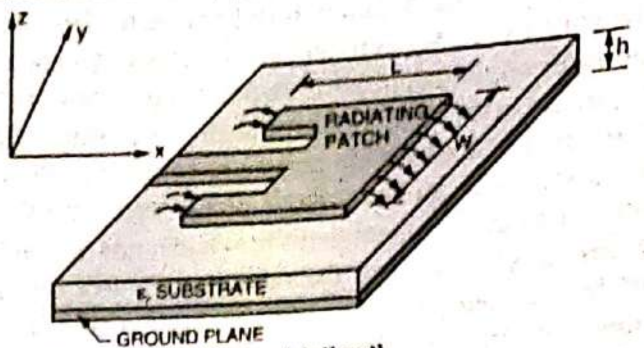
1. Microstrip feed line. [Fig. 10.55 (a). For impedance matching purposes, the offset microstrip line feed is the easiest to use as the offset depth controls the input impedance of the antenna. Moreover, this configuration is simple to fabricate and analyse. Microstrip feedline is also a conducting strip, usually of much smaller width compared to the patch. The microstrip feedline is easy to fabricate, simple to match by controlling the inset position and also simple to the model. However, as the substrate thickness increases surface waves and spurious feed radiation increase, which for practical design limit the bandwidth by 2 to 5%. There are many versions of microstrip feeds (*a*) centre feed (*b*) off-feed, (*c*) Inset feed. It is very common technique employed for feeding microstrip antenna.

2. Coaxial line or probe feed [Fig. 10.55 (*b*)], where the inner conductor of the coaxial is attached to the radiation patch while outer conductor is connected to the ground plane, are also widely used.

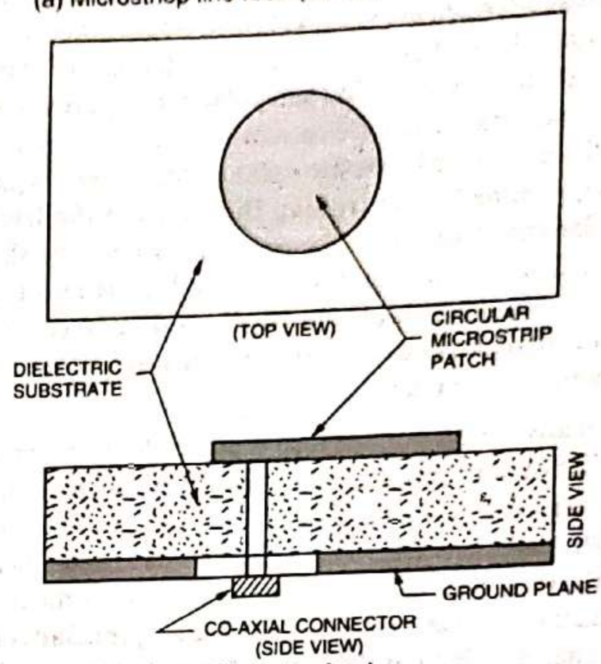
Coaxial probe feed is also easy to fabricate and match and it has low spurious radiation. However, it also has narrow bandwidth and it is more difficult to model, especially for thick substrate ($h > 0.02\ \lambda_0$). Both the **microstrip feed line** and **the probe** have inherent asymmetries which generate higher order modes leading to produce cross-polarized radiation. To overcome the problem, **non-contacting aperture coupling feeds** as shown in [Fig. 10.55 (*c*), (*d*)] have been used.

The **main advantage** of coaxial feed is that the feed can placed at any desired location inside the patch to match with its input impedance. Its **major disadvantage** is that ground plane coaxial provide a narrow band-width and difficult to model because a hole has to be drilled in the substrate.

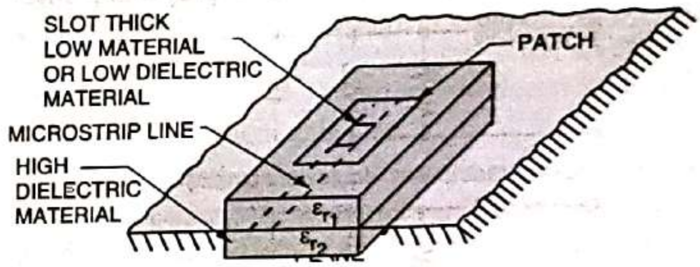
It is also called as electro magnertic coupling scheme. In this two dielectric substances are used such that the feed line is sandwiched between the two and the radiating patch is on the top of upper subatrate as shown in Fig. 10.55(c). The feed is shielded from the antenna by a conducting plane with a hole/slot ti transmit the energy to the antenna.



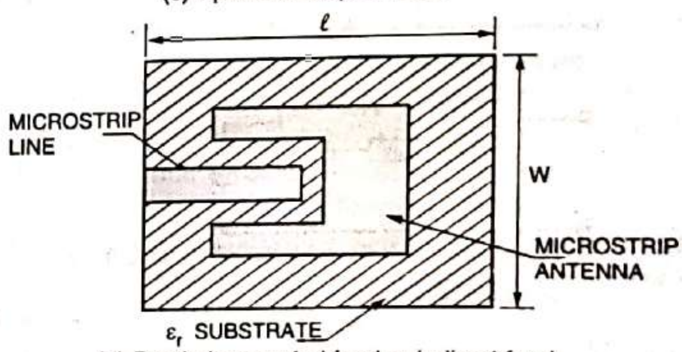
(a) Microstriop line feed (off-set).



(b) Probe-feed.



(c) Aperture-coupled feed.



(d) Proximity-coupled feed or indirect feed.

Fig. 10.55. Various feeds for microstrip antennas.

3. The aperture coupled feed [Fig. 10.55 (*c*)]. It is also called as electromagnetic coupling scheme. In this, two dielectric substrates are used such that the field line is sandwiched between the two and the radiating patch is on the top the upper substrate as shown in Fig. 10.55 (*c*). The feed circuitry is shielded from the antenna by a conducting plane with a slot/hole to transmit energy to the antenna. Aperture coupling of Fig. 10.55 (*c*) is the most difficult methods of all four to fabricate. It has also narrow banwidth. However, it is somewhat easier to model and

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0.10.4. Advantages of Microstrip Antenna

'he main advantage of microstrip antenna are

1. Low cost fabrication.
2. Can easily conform to a curved surface of a vehicle or product.
3. Many designs readily produce linear or circular polarization.
4. Considerable range of gain and pattern options (2.5 to 10 dB) available.
5. Antenna thickness (profile) is small.
6. Other microwave devices in microstrip may be integrated with a microstrip antenna with no extra fabrication steps (*v*|*z* branch line hybrid to produce circular polarization or corporate feed network for an array of microstrip antenna).
7. Microstrip antennas meet the prime needs *i.e.* small size, low weight and hence are easy to manufacture on mass scale with low manufacturing cost. These can be directly applied to metallic surface on aircraft, missile and do not disturb aerodynamic flow and thus have better aerodynamic properties. Thus, these antennas are replacing of old and bulky aerospace vehicles.

0.10.5. Disadvantage of Microstrip Antenna

he main disadvantages of microstrip antennas are :

1. Narrow bandwidth (5% to 10%, VSWR 2:1) is typically without special techniques.
2. Sensitivity to environmental factors like temperature and humidity.
3. Dielectric and conductor losses can be large for thin patches leading to poor antenna efficiency.
4. Low power handling capability.
5. Poor end-fire radiation characteristics and limited gain.

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Characteristics:-

Different dielectric substrates can be Used in the microstrip antennas.

The Value of dielectric constant Varies from $1 \leq \varepsilon_r \leq 13$.

Substrate Material	ε_r
1) Air	1
2) PTFE / glass	2.2
3) Roger's RT Duroid	2.26
4) FR-4	4.0 – 4.8
5) Alumina	9.6 - 10
6) Sapphire	9.4
7) GaAs Gallium Arsenoid	11 - 13
8) Silicon (Si)	12

It provides larger bandwidth, better efficiency.

The thin substrate are Used to small size of antenna.

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Definition :-

The antenna which is made up of metal patches placed on dielectric and fed by microstrip (or) Coplanar Transmission line is called microstrip antenna (or) patch antennas

Types of planar Transmission line:-

(i) Slot line

(ii) Strip line

(iii) Coplanar Wave Guides (cpw)

Construction:-

(i) Thickness of microstrip is very small compared to free space wavelength $(t \ll \lambda_0)$

(ii) The height of the substrate is very small $[h \ll \lambda_0]$, the typical value is $[0.003\lambda_0 \leq h \leq 0.05\lambda_0]$

(iii) The substrate in b/w the patch and the ground plane it is a dielectric sheet.

(iv) The Typical length of the patch is in b/w $\frac{\lambda_0}{3} < L < \frac{\lambda_0}{2}$.

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10.10.10. Limitations of Microstrip Antennas

1. The bandwidth of a square or circular patch antenna for a VSWR S can be given by

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$$\text{Bandwidth} = \frac{100(S-1)}{\sqrt{S}} \cdot \frac{8}{\varepsilon_r} \cdot \frac{h}{\lambda_0}\,\% \qquad \ldots (10.99)$$

This shows that the bandwidth decreases with increase of h *i.e.* **thinner antennas have lesser bandwidth.**

2. The feed structure of these antennas is usually printed on the substrate substance together with radiating elements. The feeder lines, therefore, introduces additional loss, thereby reducing the efficiency.
3. The mechanical tolerances of thin microstrip antenna normally place a limit on the precision with which the aperture phase and amplitude distribution can be controlled in manufacture.
4. Practical limitations on maximum gain (nearly 20 dB)
5. Poor end-fire radiation performance.
6. Low power handling capability.
7. Possibilities of excitation of surface waves.

Some of the above limitations may be overcome by

1. Using thick sustrate
2. Cutting slots in the metallic patch
3. Introducing parasitic patches either on the top of the main patch on the same layer
4. Using aperture coupled stacked patch antenna.

10.10.11. Applications of Microstrip Antennas

1. Microstrip antenna (MSA) are gaining popularity for use in wireless applications because of their low-profile structure.
2. They are extremely compatible for embedded antenna in handheld wireless devices like mobile phone. (cellular) and pagers.
3. Telemetry and communications antennas on missiles required to be thin and conformal and are usually microstrip antennas.
4. It is used in satellite communication because of their small size and low profile features.
5. It has widespread use in microwave and millimeter wave systems.
6. These are employed in airborne and spacecraft systems because of their low profile and conformal nature.
7. In phased arrays radars, where low profile antennas are needed and bandwidths less than a few percent are tolerable, microstrip antennas are quite popular.
8. A large number of commercial requirements are met by the use of microstrip and printed antennas. The most popular microstrip antenna is certainly the rectangular patch. **The Global Positioning System (GPS) has** become abiquitous in its applications.
9. GPS applications such as the asset tracking of vehicles and marine use have created a large demand for these antennas.
10. Satellite Digital Auto Radio Services (SDARS) have become a viable alternative to AM and FM commercial broadcasts in automobiles.

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10.21. ANTENNA WITH PARABOLIC REFLECTORS

10.21.1. Beam Formation by Parabolic Reflectors

A parabola may be defined *as the locus of a point which moves in such a way that its distance from the fixed point (called focus) plus its distance from a straight line (called directrix) is constant.* A parabola with focus F and vertex O is shown in Fig. 10.85. The Parabola is a two-dimensional plane curve.

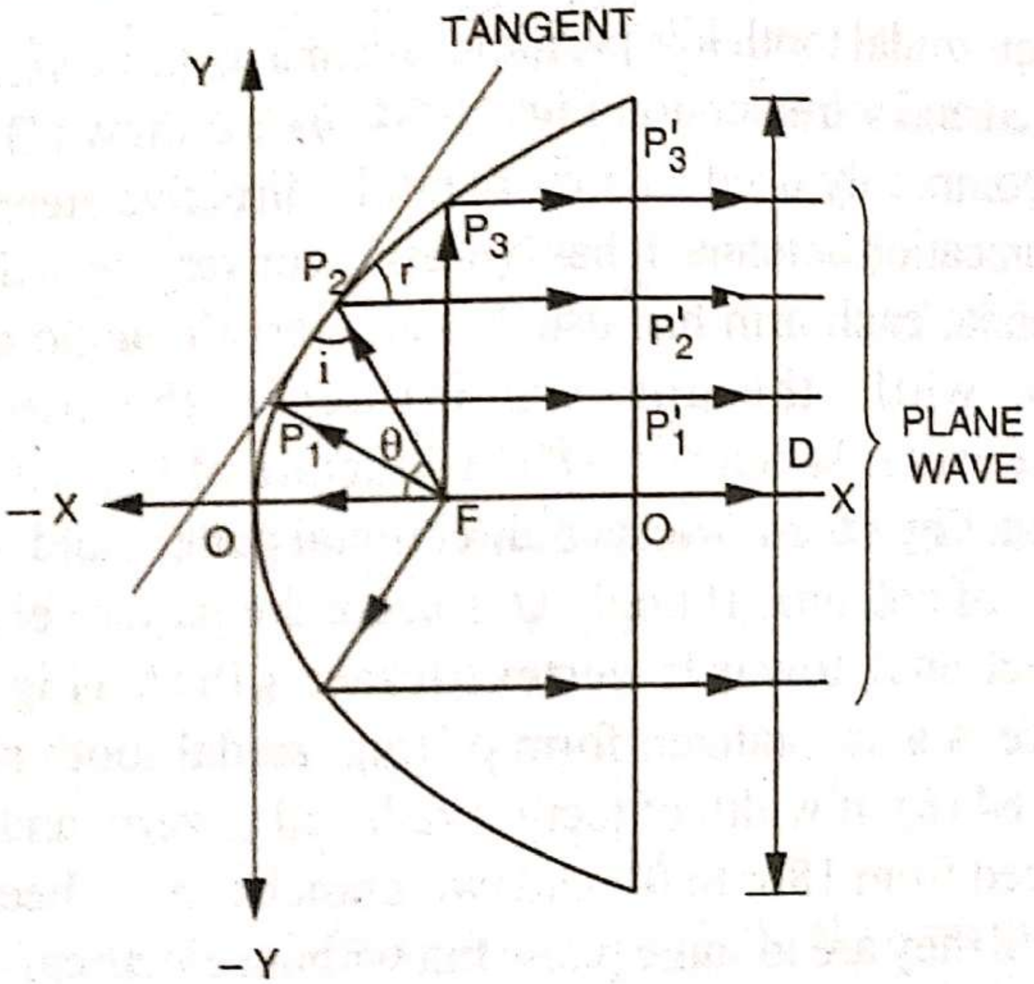

Fig. 10.85. Geometry of parabolic reflector.

OF = Focal length = f

K = A constant which depends on the shape of Parabola curve

F = Focus

O = Vertex

OO' = Axis of parabola.

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By definition of parabola, apparently,

$$FP_1 + P_1P_1' = FP_2 + P_2P_2' = FP_3 + P_3P_3'$$
$$= \text{constant (say, } K) \qquad ...(10.152)$$

The equation of Parabola curve interms of its coordinate is given by

$$Y^2 = 4fx \qquad ...[(10.153(a)]$$

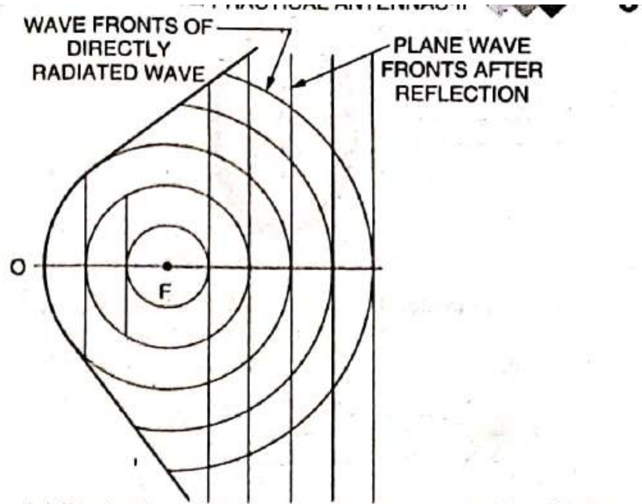
The open mouth (D) of the parabola is known as the **Aperture.** The ratio of focal length to Aperture size (*i.e.* f/D) known as "***f* over *D* ratio**" is an important characteristic of parabolic reflector and its value usually varies between 0.25 to 0.50.

Focussing or beam formation action of parabolic reflector can be understood by considering a source of radiation at the focus. Let a ray start from the focus (F) at an angle θ w.r.t. parabolic axis (OO'). The curve strikes at point P_2 on the parabola curve. Let a tangent is drawn at P_2 on the curve. According to law of reflection, the angle of incidence ($\angle i$) and angle reflection ($\angle r$) will be equal as shown. This results the reflected ray in the reflected ray being is parallel to the parabolic axis, regardless of the particular value of θ that may be considered. *In other words, all the waves originating from focus will be reflected parallel to the parabolic axis. This implies that all the waves thus, reaching at the aperture plane are in phase.* This shows that a wavefront–a surface of constant phase–is created in the aperture plane. Therefore, the rays are parallel to the parabolic axis, because rays are always perpendicular to a wavefront. *Since all the waves are in phase, so a very strong and concentrated beam of radiation is there along the parabolic axis.*

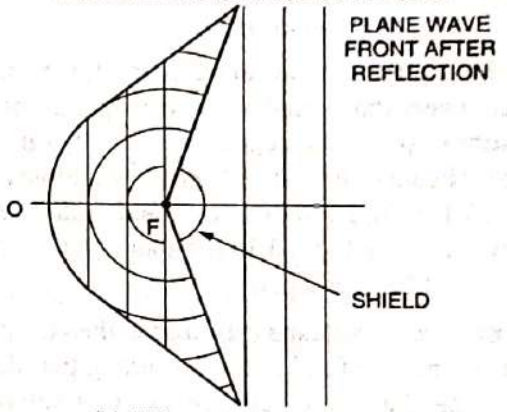
Alternatively, all the waves emanating from the source at focus and reflected by parabola are travelling the same distance (because distances are equal by Eqn. 10.152) in same time in reaching the directrix and hence they are in phase. The principle of equality of path length is maintained between all rays of two wavefronts. Putting in another way where there is path length difference between the two rays cancellation action will take place. *Hence the geometrical properties of parabola provide excellent microwave reflectors that lead to the production of concentrated beam of radiation.*

In fact, parabola converts a spherical wavefront coming from the focus into a plane wavefront at the mouth of the parabola as illustrated in Fig. 10.86. The part of the radiation from the focus which is not striking the parabolic curve as spherical wave appears as minor lobes. Obviously this is a waste of power. This is minimized by partially shielding the source as shown in Fig. 10.86 (b).

Further if a beam of parallel rays is incident on the parabolic surface, they will be focussed at a point *i.e.* Focus. This is in effect due to the principle of reciprocity theorem already discussed which says that properties of an antenna are independent whether it is for transmission or reception, the parabolic reflector is directional for reception case also as only rays coming perpendicular to directrix will be focussed at the focus and not others due to path length difference (Fig. 10.87). Parallel rays are known as *collimated.*



(a) Production of plane wavefront by parabolic reflector with omnidirectional source at Focus



(b) With partially shield source

Fig. 10.86.

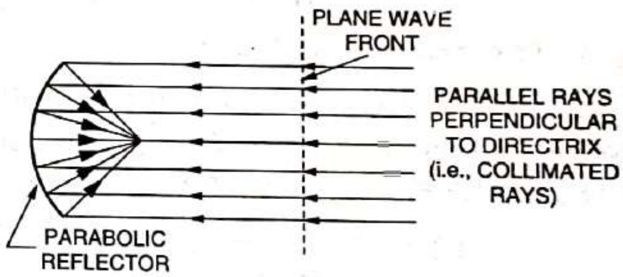


Fig. 10.87. Focussing by a parabolic reflector (Receiving Case.)

10.21.2. Paraboloidal Reflector or Microwave Dish

A parabola is a two-dimensional plane curve. A practical reflector is a three-dimensional curved surface. Therefore a practical reflector is formed by rotating a parabola about its axis (OO'). The surface so generated is known as ***Paraboloid*** which is often called ***microwave dish*** or ***Parabolic reflector*** (Fig. 10.88). Paraboloid produces a parallel beam of circular cross-section, because the mouth of the paraboloid is circular. If a third Cartesian coordinate z has its axis perpendicular to both x-axis and y-axis in Fig. 10.88, then equation of paraboloid will be

$$y^2 + z^2 = 4fx \qquad ... [10.153\ (b)]$$

The intersection of any plane perpendicular to x-axis with the paraboloid surface is a circle. In the conventional automobile, (*e.g.* motor-car headlight, or in search light), this beam forming property is utilized.

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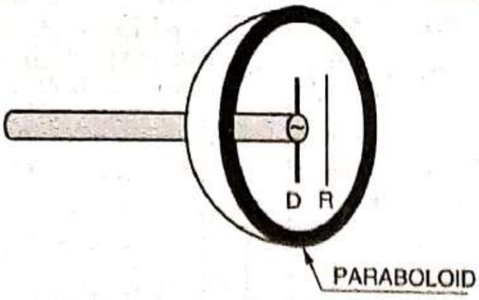
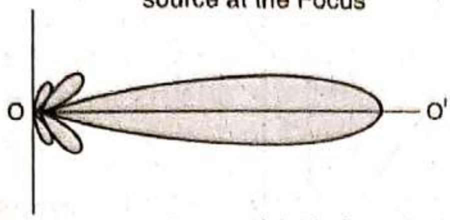


(a) Full paraboloidal reflector with dipole source at the Focus



(b) Radiation pattern of a paraboloid of aperture $D = 10\,\lambda$.

Fig. 10.88.

The radiation pattern on an antenna employing paraboloid reflector has a very sharp major lobe accompanied by a number of minor lobes which, of course, are smaller in size. The narrow major beam is in the direction of paraboloid axis shown in Fig. 10.88 (*b*). The three-dimensional pattern is a figure obtained by revolving Fig. 10.88 (*b*) about OO' and the actual shape would be like a fat cigar.

If the *feed* or *primary* antenna is isotropic, then the paraboloid will produce a beam of radiation. Assuming that the circular aperture is large, the Beamwidth between first null is given by

$$\text{BWFN} = \frac{140\lambda}{D}\text{ degree} \qquad \text{...[10.154 (a)]}$$

where λ = Free space wavelength, in m.

D = Diameter of aperture, in m *i.e.* mouth diameter.

The beamwidth between first nulls for a large uniformly illuminated rectangular aperture is given by

$$\text{BWFN} = \frac{115\lambda}{L}\text{ degree} \qquad \text{... [10.154 (b)]}$$

where L = Length of Aperture, in λ

Also width between Half-power points for a large circular aperture is given by

$$\text{HPBW} = \frac{58\lambda}{D}\text{ degree} \qquad \text{... [10.154 (c)]}$$

Further, the directivity D of a large uniformly illuminated aperture is

$$D = \frac{4\pi A}{\lambda^2} \qquad \text{... [10.154 (d)]}$$

and for a circular, aperture

$$D = \frac{4\pi}{\lambda^2}\left(\frac{\pi D^2}{4}\right) = \pi^2\left(\frac{D}{\lambda}\right)^2$$

$$D = 9.87\left(\frac{D}{\lambda}\right)^2 \qquad \text{... [10.154 (e)]}$$

where D = Diameter of the aperture, in λ.

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10.21.4. Primary and Secondary Pattern

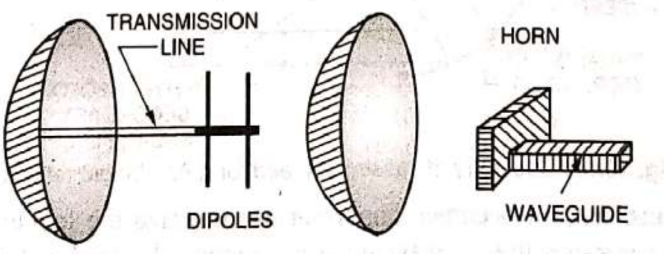
The antenna placed at the **focus** of a paraboloid is known as ***Feed radiator*** or ***primary radiator*** or simply ***feed*** and its radiation pattern is known as **primary pattern**. The parabolic reflector is known as **Secondary radiator** and the radiation pattern of entire antenna system (*e.g.* Reflector and primary radiator) is called as *Secondary pattern.* Sometimes *Antenna pattern* is used for secondary pattern and *Feed pattern* for *Primary pattern.*

10.21.5. Feed Systems

The entire Parabolic reflector antenna consists of two basic components *e.g.* the reflector and a source of primary radiation at the focus. The source is called the primary radiator or feed radiator or simply feed while the reflector, the secondary radiator. Now detailed design of feed is discussed.

An ideal *feed* would be that radiator which radiates towards reflector in such a way that it illuminates the entire surface of reflector and no or zero energy is radiated in any other direction. Of course, such an ideal radiator is not available in practice. Clearly an isotropic antenna as feed would not be a better choice. As far as the secondary radiator is concerned, the best choice is the Paraboloid which is inherited with compactness and simplicity. However, there are a number of choices for primary feed.

Similarly, a dipole antenna is also not very much suitable for the feed but occasionally used. The simplest and generally used is a dipole with parasitic reflector (*i.e.* Yagi-Uda) or a small plane reflector, which is fed with a coaxial line (Fig. 10.88). Typically the spacing between driven element and parasitic element is $0.125\ \lambda$ and for a plane reflector it may be around $0.4\ \lambda$. Besides end fire arrays of dipoles are also used in front of reflector as shown in Fig. 10.95 (*a*). The double diploes are so spaced and phased that end-fire pattern is produced which illuminates the paraboloid reflector. It may be noted, however, that feeding with a dipole involves changing from unbalanced system to a balanced system.



(a) Dipole end-fire feed. (b) Horn feed.

Fig. 10.95.

A most common feed radiation for paraboloid reflector antenna is a '*waveguide horn*' [Fig. 10.95 (*b*)].

The horn feed is waveguide feed. As shown horn antenna (*i.e.* feed antenna) is pointing the paraboloid and radiation pattern of horn antenna is mild, in the same direction. Thus, the direct radiation from the horn (*i.e.* feed) antenna is minimum. Further, if circular polarization is required then, conical horn antenna or helix, antenna can be used as feed at the focus of paraboloid. For getting maximised beam pattern along the parabolic axis,

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feed is placed at the focus. But if the feed is moved laterally from the focus *i.e.* perpendicular to axis, then beam deteriorate *i.e.*, limited beam motion can be obtained. On the other hand, if the feed is moved along the axis, then the pattern is broadened. Thus important position of feed is the focus and for the small reflector of short focal length the position of feed is as shown in Fig. 10.96.

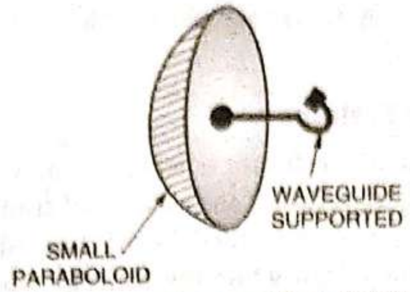

Fig. 10.96. Feed support for a small paraboloid reflector.

10.21.6. Cassegrain Feed

It is named after the name of 18th Century Astronomer and is illustrated in Fig. 10.97 in which the primary feed radiator is positioned around an opening near the vertex of the paraboloid instead of at focus. Cassegrain feed system employs a hyperboloid secondary reflector whose one of the foci coincides with the focus of paraboloid.

The feed radiator is aimed at the *secondary hyperboloid reflector or sub-reflector.* As such, the radiations emitted from feed radiator are reflected from cassegrain secondary reflector which illuminates the main Paraboloid reflector as if they had originated from the focus. Then the paraboloid reflector colliminates the rays (renders parallel) as usual.

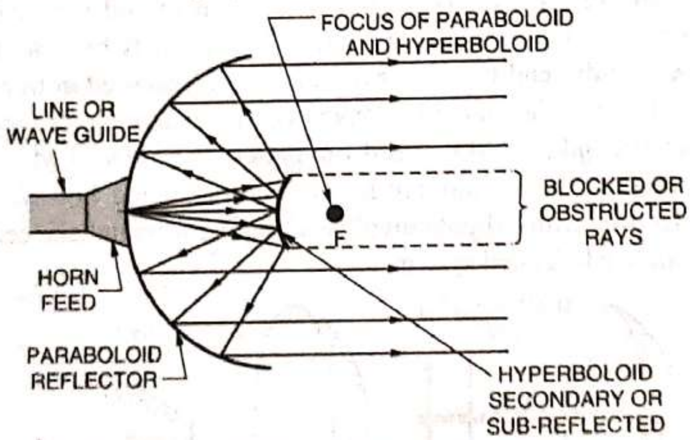

Fig. 10.97. Geometry of cassegrain feed for a paraboloid reflector.

Sometimes, it becomes important to minimize the length of transmission line or waveguide connecting the feed radiator with receiver or transmitter. This is needed specially to avoid losses. Although there could be a solution of this problem by placing the RF Amplifier stage of R_x near the focus which minimizes the losses on reception, but this is not practicable for transmitters, as the RF amplifier of a transmitter is bulky, heavy and having enough power so not possible to place at feed point. *Hence the practical solution in such cases is cassegrain feed when the transmission line or waveguide length between feed and transmitter and receiver, is required to be short.*

The disadvantage of the cassegrain feed is that some of the radiation from the Paraboloid reflector is obstructed. This is

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tolerable in greater dimension paraboloid but becomes problem with small dimension paraboloid. The dimension of secondary reflector depends on the distance between horn feed and sub-reflector, mouth of horn which in turn depends on frequency. This aperture blocking defect can be avoided by using an *off set reflector* which is applicable to focal point feed shown in Fig. 10.98. The other method is to use a polarization twisting scheme in which hyperboloid reflector is made of wire grating (transparent) instead of solid.

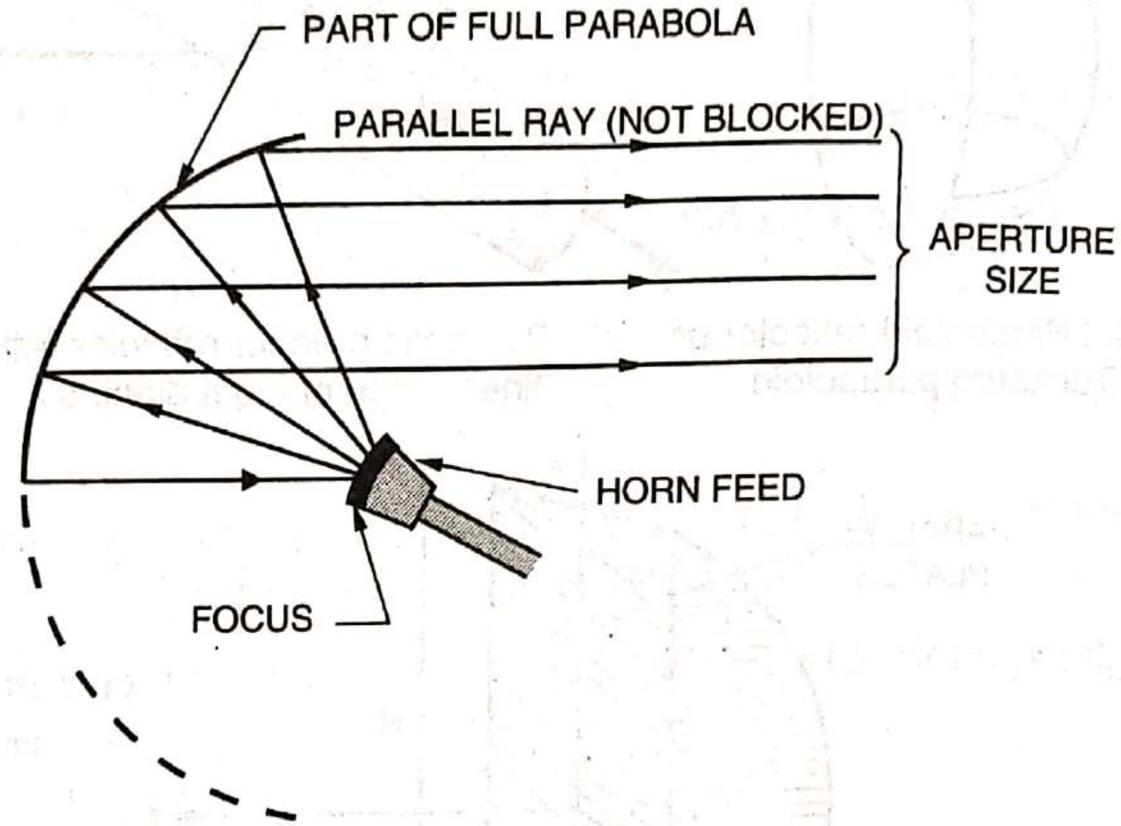

Fig. 10.98. Off-set Paraboloid reflector showing no blocking of rays.

10.21.7. Advantages of Cassegrain Feed

The following are the advantages of cassegrain feed arrangements in general :

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

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